3D Beam-Steering MIMO Antenna for On-Body IoT Applications

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Abstract

In this article, a compact 3D beam-steering MIMO antenna operating at 5.75 GHz is proposed for on-body Internet of Things (IoT) applications. The planar antenna comprises a central patch and two concentric annular rings. The theory of spherical modes is used to discuss the beam-steering concept, where the free space measured beam-scanning covers the entire xy-plane, while a 44° range is realized in xz and yz-planes. The on-body performance is tested using a multi-layer phantom, and the measured peak realized gain is 6.41 dBi. The antenna is characterized by compact size (0.77λ), low profile (0.03λ), specific absorption rate below the established limits, good efficiency and 3D beam-steering characteristics while operating in an on-body setup. Therefore, the proposed solution can enable advanced wireless applications like localization and physical layer security in emerging size-constrained on-body IoT devices.
Abstract—In this article, a compact 3D beam-steering MIMO antenna operating at 5.75 GHz is proposed for on-body Internet of Things (IoT) applications. The planar antenna comprises a central patch and two concentric annular rings. The theory of spherical modes is used to discuss the beam-steering concept, where the free space measured beam-scanning covers the entire $xy$-plane, while a $44^\circ$ range is realized in $xz$ and $yz$-planes. The on-body performance is tested using a multi-layer phantom, and the measured peak realized gain is 6.41 dBi. The antenna is characterized by compact size (0.77λ), low profile (0.03λ), specific absorption rate below the established limits, good efficiency and 3D beam-steering characteristics while operating in an on-body setup. Therefore, the proposed solution can enable advanced wireless applications like localization and physical layer security in emerging size-constrained on-body IoT devices.

Index Terms—On-body antennas, MIMO antennas, beamsteering antennas, planar antennas, spherical modes.

I. INTRODUCTION

Beam-steering antennas are an actively researched topic due to their ability to alter the antenna-generated radiation pattern. This feature enhances numerous aspects of modern wireless communication systems, e.g., spatial coverage, spectral efficiency, and interference mitigation [1]–[3].

Recently, different methods have been proposed to realize compact beam-steering antennas, e.g., in the azimuth plane [4]–[9], or in the elevation plane [10]–[14]. However, because of size and profile, integrating these solutions into on-body IoT systems is still a challenging problem. Also, only a few works have investigated compact beam-steering antennas covering both the azimuth and elevation planes. In [15], [16], stacked patches are used for full-scan in the azimuth plane and a $120^\circ$ coverage in the elevation plane. In [17], a highly pattern-reconfigurable Alford loop antenna is proposed for 3D full-space scanning. While these antennas offer a wide-scanning range, the structures are of relatively large diameter [17] (> 1.30λ, where λ is the wavelength at the center frequency);

or else, the designs are bulky and not planar, with profile > 0.2λ, which is still relatively high for on-body IoT devices [15], [16].

On-body beam-steering antennas have been proposed in [18]–[29]. Elevation plane beam-steering is discussed in [18]. A dual-mode pattern synthesis (omnidirectional and broadside) is achieved in [19]–[21]. In [22], the theory of characteristic modes is used to design a dual-port metal-frame smart watch antenna capable of changing between a bidirectional and an omnidirectional pattern in the $xz$ and $yz$-planes. Lastly, a dual-mode pattern synthesis is obtained in [28] (using tunable ring slots and p-i-n diodes) and in [29] (using rectangular patches loaded with rectangular slots and p-i-n diodes). However, the beam-steering performance of these works is still limited to only $yz$-plane [18], dual-mode patterns [19]–[21], [28], [29], or bidirectional and omnidirectional patterns in $xz$ and $yz$-planes [22]. Recently, in [30], we exploited the theory of spherical modes to design a stacked-patch beam-steering antenna for directional modulation applications. Although the diameter is around λ/2, it only allows for beam-steering in the $xy$-plane, is not planar, and has a 0.063λ profile, which limits its use for on-body IoT scenarios, e.g., integration with a smart watch.

In this work, we propose for the first time a compact and planar 3D beam-steering on-body antenna. The theory of spherical modes [31] is used to describe the antenna beam-steering in the $xy$-plane, and the beam-scanning is extended to cover the $xz$ and $yz$-planes through the excitation of broadside modes. A measured 360$^\circ$ $xy$-plane scan is realized for both free space and on-body operation. A 44$^\circ$ coverage is demonstrated in free space for $xz$ and $yz$-planes, reducing to around 32$^\circ$ for the on-body case. The simulated total efficiency is $\geq 73\%$, and the peak gain is 8.61 dBi in free space, decreasing to respectively 58$\%$ and 7.8 dBi with the phantom. This increased antenna gain allows improved signal strength, reduces battery use, and enables angle of arrival localization, to name just a few benefits. The remainder of this article is organized as follows. Section II discusses the beam-steering principle and antenna design. Section III presents the free space performance, and the on-body performance is discussed in Section IV. Finally, conclusions are presented in Section V.

II. BEAM-STEERING PRINCIPLE AND ANTENNA DESIGN

A. Spherical Modes Analysis

The electric field $\vec{E}(r, \theta, \phi)$ can be represented as a weighted sum of spherical modes functions $F_{smn}(r, \theta, \phi)$, and assuming the $e^{-i\omega t}$ time dependency this is expressed as [32]
\[ \vec{E}(r, \theta, \phi) = \frac{k}{\sqrt{\eta}} \sum_{c, m, n} Q_{c, m, n}\vec{F}_{c, m, n}(r, \theta, \phi) \]

where \( k = 2\pi / \lambda \) is the wavenumber, \( \lambda \) is the wavelength, \( \eta \) is the specific impedance of the medium, and \( Q_{c, m, n} \) are the spherical modes coefficients. The index \( s \) is related to polarization (used to distinguish between two functions); for transverse magnetic (TM) waves, \( s = 1 \) corresponds to the magnetic fields, and \( s = 2 \) represents the electric fields, while this is reversed for the transverse electric (TE) waves. The index \( n = 1, 2, \cdots, N \) denotes the order of the spherical mode, and \( m (|m| \leq n) \) denotes the azimuthal phase variations. The upper index \( c \) specifies a radial function, while \( c = 1, 2 \) are standing waves, \( c = 3 \) and \( c = 4 \) represent the outward and inward traveling waves, respectively. The spherical modes’ vector fields \( \vec{F}_{c, m, n}(r, \theta, \phi) \) are power-normalized dimensionless solutions of the vector wave equation and are defined in [32] and [33]. A general outgoing wave field can then be written as \( \vec{E}(r, \theta, \phi) = \frac{k}{\sqrt{\eta}} \sum_{c, m, n} Q_{c, m, n}\vec{F}_{c, m, n}(r, \theta, \phi) \).

At a large distance \((kr \to \infty)\), the far-field can be described solely in terms of the angular variables \((\theta, \phi)\) as

\[ \vec{K}_{c, m, n}(\theta, \phi) = \lim_{kr \to \infty} \left[ \sqrt{\frac{4\pi}{kr}} e^{ikr} \vec{F}_{c, m, n}(r, \theta, \phi) \right] \]

The far-field functions can then be defined explicitly as [32]

\[ \vec{K}_{1, m, n}(\theta, \phi) = \frac{2}{n(n+1)} \left[ -\frac{m}{|m|} \right]^m \frac{\sin \theta}{\sin \theta} e^{i\beta_n \theta} - \frac{d\vec{P}_{n}^{m|}(\cos \theta)}{d\theta} \vec{e}_{\phi} \]

\[ \vec{K}_{2, m, n}(\theta, \phi) = \frac{2}{n(n+1)} \left[ -\frac{m}{|m|} \right]^m \frac{\sin \theta}{\sin \theta} e^{i\beta_n \theta} + \frac{d\vec{P}_{n}^{m|}(\cos \theta)}{d\theta} \vec{e}_{\phi} \]

Note that the radius component of \( \vec{F}_{c, m, n}(r, \theta, \phi) \) now disappears, as it is intrinsic to the far-field region. In terms of \( \vec{K}_{c, m, n}(\theta, \phi) \) the \( \vec{E} \)-field at a large distance is then given by

\[ \vec{E}(r, \theta, \phi) = \frac{k}{\sqrt{\eta}} \sqrt{4\pi} e^{ikr} \sum_{c, m, n} Q_{c, m, n}\vec{K}_{c, m, n}(\theta, \phi) \]

**B. Beam-Steering Principle**

Let us assume that an antenna can be fully enclosed by a sphere of smallest radius \( r_0 \) (see Fig. 1), and radiates omnidirectional spherical modes of type \( s = 1 \) so that indexes \( n \) and \( m \) of each mode \((n \geq 2)\) are related as follows

\[ m = \pm n \]

The modes satisfying the above condition can be expressed as \( \vec{K}_{1, \pm N, N} \). To realize beam-steering in the whole azimuth plane, a MIMO system is designed to excite all omnidirectional spherical modes of order \( N, N - 1, \cdots, 1 \) that can be supported within the enclosing sphere of radius \( \leq r_0 \). Except for the fundamental mode \((n = 1 \text{ with } m = 0, \text{i.e., } \vec{K}_{1, 0, 0} = \vec{K}_{1, 0, 1})\), which has a constant phase across the entire \( xy \)-plane, the remaining modes have their phase changing in two opposite directions, \( \pm N, \pm(N - 1), \cdots \); whereas the “+” sign indicates clockwise rotation, and the “−” sign denotes counter-clockwise rotation. The beam-steering towards \( \phi_a \) direction is then realized using

\[ D(\theta, \phi_a) = w_1 \vec{K}_{1, 1, 0}(\theta, \phi_a) + \sum_{n=2}^{N} w_n \vec{K}_{1, \pm n, n}(\theta, \phi_a) \]

where \( D(\theta, \phi_a) \) is the maximum directivity in a given \( \theta \)-cut, \( w_n = |A_n|e^{i\beta_n} \Delta \beta_n \) term is a weighting factor applied at the port exciting the respective omnidirectional spherical mode, \(|A_n|\) and \( \Delta \beta_n \) are the magnitude and phase shift, respectively.

The fundamental mode \( \vec{K}_{1, 0, 1} \) has a constant phase across the entire \( xy \)-plane, hence it is used as the reference mode when computing the required phase shift

\[ \Delta \beta_n = \beta_{K_{1, 0, 1}}(\phi_a) - \beta_{K_{1, \pm n, n}}(\phi_a) \]

where \( \beta_{K_{1, 0, 1}}(\phi_a) \) and \( \beta_{K_{1, \pm n, n}}(\phi_a) \) is the phase value of \( \vec{K}_{1, 0, 1} \) and the phase-varying mode, respectively.

**C. Spherical Modes Based Planar Beam-Steering Antennas**

To implement the proposed principle in on-body devices, a centrally located patch and concentric annular rings are used to design a planar structure. The \( \vec{K}_{1, 0, 1} \) mode is excited using a central circular patch, while the phase-varying modes are excited using concentric annular rings. Fig. 2 shows a planar configuration, where each structure excites a total of \( l = 1, 2, 3, \cdots \) omnidirectional spherical modes. The structures are supported by a 1.52 mm thick RO4003C substrate \((c_r = 3.38, \tan \delta = 0.0027)\) and operate at the center frequency \( f_0 = 5.8 \text{ GHz} \). The metallic radiators and the ground plane are made of a 35 \( \mu \)m thick copper (5.8e7 S/m), and the shorting pins are also made of copper, but with 0.5 mm diameter.

Radiator A is a center-fed patch (using P1) with 14 mm diameter. The patch excites the single fundamental spherical mode \((l = 1)\). It includes four shorting pins \((n_{pins} = 4)\) located at \( v_1 = 2.75 \text{ mm} \) away from the center of the patch and rotated by 90° with respect to the disk center. Since the patch is out of resonance in the investigated frequency range, the integration of the shorting pins allows it to have resonance at the desired \( f_0 \) [34], [35]. This is highlighted in Fig. 3a, where \( n_{pins} \) varies from 0 to 6. It can be seen that within the investigated...
frequency range, the patch has non-zero resonant frequency for $n_{\text{pins}} = 2, 4$, and $6$, and good matching ($|S_{11}| < -20\, \text{dB}$) is achieved for $n_{\text{pins}} = 4$. It is also important to note that due to the pins’ inductance, the frequency shifts upwards for an increased $n_{\text{pins}}$ value. Moreover, frequency tuning and matching enhancement can be obtained by adjusting the pins’ distance, i.e., $v_1$ value. This is demonstrated in Fig. 3b, where $v_1$ is increased from 2.5 mm to 3.25 mm. It can be seen that the frequency increases with $v_1$, and at $v_1 = 2.75$ mm good matching is observed at the desired $f_0 = 5.8$ GHz.

Ring B in Fig. 2 excites the dual-phase varying modes $\vec{K}_{1, \pm 2.2}$ (P2), and $\vec{K}_{1, -2.2}$ (P3). It has an inner diameter of 15 mm and an outer diameter of 32 mm. The ports P2 and P3 are located 2 mm from the edges of the inner diameter of the ring and are rotated by $\alpha_2 = \pi/4$, computed from

$$\alpha_n = \begin{cases} \frac{\pi}{2} \left( \frac{1}{n} + u \right), & n = 2, 4, 6, \cdots; u = 0, 1, 2, \cdots \\ \frac{\pi}{6} \left( \frac{3}{n} + 2u \right), & n = 3, 5, 7, \cdots; u = 0, 1, 2, \cdots \end{cases}$$

where $n$ is the order of the excited mode, and $u$ denotes an integer that can be arbitrarily selected to obtain $\alpha_n$ that satisfies the condition to generate omnidirectional spherical mode of order $n$. It is introduced here as numerous possible $\alpha_n$ satisfy the required criteria. The ring includes shorting pins for frequency tuning and isolation enhancement, and the pin numbers need to take into account the ports’ position. This is necessary to ensure symmetry between the two excited modes of the same order. For this work, the total number of pins in a given annular ring is calculated using

$$n_{\text{pins}} = \frac{2\pi}{\alpha_{\text{pins}}} p$$

where $\alpha_n$ is the angle between the two feeds as calculated in (9), whereas $p = 1, 2, 3, \cdots$ is an integer that controls how many pins are located between the two feeds, while taking into account the desired $f_0$, $|S_{ii}| \leq -10\, \text{dB}$, and isolation $\geq 10\, \text{dB}$. Please note that (10) also ensures that the pins are located symmetrically with respect to the two ports exciting each ring. Due to the inductance of the pins, an increase in $f_0$ is observed for larger values of $p$. For ring B, $p = 2$ is used, and from (10) $n_{\text{pins}} = 16$. The pins of this configuration are located 1 mm from the edges of the inner diameter of the ring and are rotated with respect to the center by $\alpha_{\text{pins}} = \pi/8$ obtained using

$$\alpha_{\text{pins}} = \frac{\alpha_n}{p}$$

Fig. 4 shows the effect of adding shorting pins into the ring structures, demonstrated for ring B. Fig. 4a shows the case without pins, where it can be seen that while good isolation is obtained at $f_0$, $|S_{ii}|$ stays above $-2.5\, \text{dB}$. In contrast, when $n_{\text{pins}} = 16$ are added into the ring (see Fig. 4b), good isolation is obtained with an improvement of the matching ($|S_{ii}| < -14\, \text{dB}$), while due to the pins inductance, the $f_0$ is also shifted towards the desired value of 5.8 GHz.

The annular ring C (inner diameter 34 mm and outer diameter 46.8 mm) excites two modes with triple-phase variations ($\vec{K}_{1, \pm 3.3}$). The feed ports of these modes (P4 and P5) are located 1.75 mm from the ring inner diameter and are rotated by $\pi/6$ obtained from (9). A total of $n_{\text{pins}} = 12$, obtained using (10) for $p = 1$ are integrated 1 mm from the edges of the inner diameter and rotated by $\pi/6$, computed using (11). Lastly, Ring D excites the quad-phase varying modes ($\vec{K}_{1, \pm 4.4}$), with outer diameter = 61.3 mm. The ring is fed using P6 and P7 located 1.8 mm from the inner diameter = 49.1 mm, and rotated by $\alpha_3 = \pi/8$. It uses $n_{\text{pins}} = 16$ placed 1 mm from the inner diameter and rotated by $\alpha_{\text{pins}} = \pi/8$.  

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Fig. 2. Proposed antenna: (a) A excites $\vec{K}_{1,0.1}$; B excites $\vec{K}_{1, \pm 2.2}$ for bidirectional beam-steering; C excites $\vec{K}_{1, \pm 3.3}$ for unidirectional beam-steering; and D excites $\vec{K}_{1, \pm 4.4}$ for enhanced directivity.

Fig. 3. Parametric results of radiator A: (a) S-parameter results for different numbers of shorting pins; and (b) S-parameter results for different $v_1$ values.

Fig. 4. S-parameters of radiator B, port index $i, j = 2, 3$: (a) results of the annular ring without shorting pins; and (b) results including shorting pins.
TABLE II
ANTENNA PERFORMANCE COMPARISONS BETWEEN FIG. 2 STRUCTURES FOR $\phi_a = 90^\circ$, $\theta = 45^\circ$ BEAM-STEERING DIRECTION.

<table>
<thead>
<tr>
<th>$l$</th>
<th>Ant</th>
<th>Size ($\lambda \times \lambda \times \lambda$)</th>
<th>Isolation (dB)</th>
<th>Tot. Eff. (%)</th>
<th>$10\text{dB}$ IBW (MHz)</th>
<th>Directivity (dBi)</th>
<th>HPBW ($^\circ$)</th>
<th>SLL (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>A</td>
<td>0.27\lambda \times 0.27\lambda \times 0.038\lambda</td>
<td>-</td>
<td>-</td>
<td>87</td>
<td>-1 (O)</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>3</td>
<td>A+B</td>
<td>0.62\lambda \times 0.62\lambda \times 0.037\lambda</td>
<td>31.2</td>
<td>-63</td>
<td>34</td>
<td>3.95 (B)</td>
<td>88.7</td>
<td>-</td>
</tr>
<tr>
<td>5</td>
<td>A+B+C</td>
<td>0.93\lambda \times 0.93\lambda \times 0.036\lambda</td>
<td>23.2</td>
<td>55.8</td>
<td>32</td>
<td>7.08 (U)</td>
<td>95.4</td>
<td>-4.8</td>
</tr>
<tr>
<td>7</td>
<td>A+B+C+D</td>
<td>1.18\lambda \times 1.18\lambda \times 0.036\lambda</td>
<td>23</td>
<td>44.6</td>
<td>27</td>
<td>9.93 (U)</td>
<td>71.3</td>
<td>-7.3</td>
</tr>
</tbody>
</table>

$\lambda$ is the total number of excited modes, O-Omnidirectional, B-Bidirectional, and U-Unidirectional patterns.

The required $\Delta \beta_n$ to steer the main beam towards $\phi_a = 90^\circ$ are shown in Table I and are computed using (8), i.e. no phase shift is added to P1 (exciting the $K_{1,0,1}$ mode) and the phase shifts of the remaining ports is obtained by subtracting the phase value of P1 (at $\phi_a = 90^\circ$ direction) with the respective port phase value (also at $\phi_a = 90^\circ$ angle). The generated patterns are shown in Fig. 6, and the beam characteristics are outlined in Table II. It can be concluded that:

- A desired level of directivity can be realized by increasing the number of modes excited in the structure, e.g., for the configuration in Fig. 2, a directivity $\geq 7$ dBi is achieved by exciting at least $l = 5$ modes (using A, B, and C).
- A and B combination provides the first type of beam-steering, a bidirectional one, where a second main beam is always located at $\phi_a + 180^\circ$ (see Fig. 6).
- The smallest diameter to realize a unidirectional beam-steering is obtained using $n = 1, 2, 3$ modes (i.e., $l = 5$).
- Because higher-order modes have increased phase variations that require a much larger diameter, they present decreased bandwidth and total efficiency.
- Lastly, at the cost of a larger antenna diameter, an increase in the number of modes allows the generation of a much more selective beam pattern. This is seen from $l = 7$ modes, which present the lowest half-power beamwidth (HPBW), highest directivity, and lowest side-lobe level (SLL) compared to the other tested unidirectional patterns.

D. 360° Azimuth plane Beam-Steering

Without loss of generality, we analyze the beam-steering capabilities by steering the beam towards $\phi_a = 90^\circ$. All the ports exciting the phase-varying modes in the same ring are fed in quadrature. Fig. 5 shows the phase of the radiation pattern of the final design shown in Fig. 2, i.e. including all structures A, B, C, and D that generate respectively the fundamental mode $K_{1,0,1}$, $K_{1,\pm2,2}$ modes, the $K_{1,\pm3,3}$ modes and the $K_{1,\pm4,4}$.

**Fig. 5.** Simulated phase of the radiation pattern for the final design (shown in Fig. 2) including structures A, B, C, and D.

**Fig. 6.** Proposed beam-steering performance of the design shown in Fig. 2 for $\phi_a = 90^\circ$, $\theta = 45^\circ$ direction: (a) $\theta = 45^\circ$ cut; and (b) $\phi = 90^\circ$ cut.

E. 3D Beam-Steering Antenna

To realize beam-steering across the azimuth and elevation planes, the first step is to select the type of beam-steering desired in the azimuth plane (e.g., bidirectional, unidirectional). Next, the number of modes required is selected according to the principle discussed in Section II-D. For the antenna proposed in this work, the lowest combination capable of beam-scanning in the azimuth plane is investigated, i.e., radiator A and B. It is understood that solely the omnidirectional modes of type $K_{1,\pm n,n}$ cannot effectively control the beam outside the azimuth plane. Therefore, two orthogonal broadside radiating modes are excited using an annular ring, i.e., the modes generating the main beam along the $z$-axis, with two modes corresponding to two orthogonal polarization. A single broadside mode will allow the structure to cover $\theta = 0^\circ$ without any phase control. To steer the beam towards other $\theta$ directions, e.g., within $xz$- and $yz$-planes, the broadside modes are combined with the phase-varying $K_{1,\pm n,n}$ modes.

**Fig. 7a** shows an antenna proposed for 3D beam-steering. The antenna is $h = 1.59$ mm thick, and uses the RO4003C...
substrate. For azimuth scanning, the antenna excites $\vec{K}_{0,1}$ (using the center-fed circular patch), and $\vec{K}_{1,\pm 2}$ modes (using the outer-most ring fed using P4 and P5); while two broadside modes are used to cover other planes, e.g. $xz$ and $yz$-planes. These modes are excited using the first ring fed using P2 and P3 (rotated by 90°) and include 8 shorting pins that are placed $v_2 = 1$ mm from the inner diameter and rotated by 45° with respect to the center. Please note that this function and the generated patterns differ from the ring B in Fig. 2.

The proposed antenna includes a total of 28 shorting pins, and all the pins have a diameter of 0.5 mm. During the prototype stage, the pins will consist of copper wires of 0.5 mm diameter, which typically have a tolerance of ±0.005 mm. To highlight the antenna robustness with respect to the shorting pins dimensions, we conducted parametric studies where the tolerance value was doubled to ±0.01 mm, i.e., the diameter was varied from 0.49 mm to 0.5 mm and 0.51 mm. The results are shown in Fig. 8. As expected, due to the inductance of the pins, the $f_0$ of each radiator shifts upwards as the diameter of the pins increases. However, it can also be noticed that only a slight frequency shift occurs, where $f_0$ is around 5.73, 5.75, and 5.76 GHz, for respectively 0.49, 0.5, and 0.51 mm diameter. Additionally, the −10 dB impedance bandwidth changes from 42.3, to 43, and, 41 MHz, when the diameter is increased from 0.49 to 0.51 mm, while the isolation stays around 24 dB in all the three cases. The results demonstrate that while slight frequency shifts may occur due to the tolerance of the pins, the isolation is little affected, and only slight changes in the impedance bandwidth are observed. The final diameter of the pins is therefore 0.5 mm, and the antenna final dimensions are 40.4 mm × 40.4 mm × 1.59 mm, or correspondingly $0.77\lambda \times 0.77\lambda \times 0.03\lambda$ for $f_0 = 5.75$ GHz.

### III. Free Space Performance

Fig. 7b shows the image of the manufactured prototype. The antenna is probe fed using a 50 Ohm semi-rigid coaxial cable, the Taoglas CAB.058 cable [36]. The connectors’ inner conductor has a diameter of 0.287 mm, while the outer conductor has a diameter of 1.193 mm. The semi-rigid coaxial cable is used for antenna measurements. For integration within an on-body device (e.g., a smart watch), the antenna is intended to be placed on top of the device so that it can be directly probe-fed by its circuitry. The simulated and measured S-parameters are shown in Fig. 9. The $f_0$ in both cases is near 5.75 GHz. An isolation > 24 dB and a 42.2 MHz −10 dB Impedance...
BandWidth (IBW) are obtained in simulations; decreasing to 16 dB and 31 MHz in measurements, respectively. Such discrepancies are most likely due to manufacturing tolerances.

Fig. 10 shows the simulated and measured radiation patterns, omnidirectional spherical modes (Fig. 10a), and broadside modes (Fig. 10b). Overall, a good agreement is demonstrated, and the required omnidirectional patterns are seen in Fig. 10a, while the broadside patterns have their main beam at $\theta = 0^\circ$. Fig. 11a shows the simulated total efficiency for each port of the proposed antenna, where at $f_0$, the total efficiency is 88%, 82.5%, 82.7%, 73%, and 73%, for P1, P2, P3, P4, and P5, respectively. The lowest efficiency and bandwidth limitations are observed for the dual-phase varying modes $\vec{K}_{1,\pm 2,2}$ (excited using P4 and P5). These characteristics agree well with the performance highlighted in Section II-D, where a bidirectional beam-steering performance is realized as only the fundamental and the dual-phase varying modes are used.

The simulated peak realized gain at 5.75 GHz is 2.1 dBi (for P1), 8.58 dBi (P2), 8.61 dBi (P3), 4.25 dBi (P4), and 4.3 dBi (P5). The measured values change to 1.87 dBi (for P1), 7.21 dBi (P2), 7.15 dBi (P3), 3.51 dBi (P4), and 3.43 dBi (P5). It can be seen that the realized gain slightly deteriorates in the measured cases (especially for the ports exciting the dual-phase varying modes), where a discrepancy of up to 0.87 dBi is observed. Such variations are most likely due to substrate material tolerance and other manufacturing inaccuracies.

Lastly, the Envelope Correlation Coefficient (ECC) is calculated from the complex 3D radiation patterns using [37]

$$\rho_{ECC} = \frac{\iint_{4\pi} \vec{E}_i \cdot \vec{E}_j^* d\Omega}{\sqrt{\iint_{4\pi} \vec{E}_i \cdot \vec{E}_i^* d\Omega \iint_{4\pi} \vec{E}_j \cdot \vec{E}_j^* d\Omega}}$$  (12)

where $\vec{E}_i$ and $\vec{E}_j$ are the far-field radiation patterns from ports $i, j = 1, 2, 3, 4,$ and, 5. The ECC stays below 0.001 within the entire bandwidth and is $\leq 3 \times 10^{-4}$ at $f_0 = 5.75$ GHz.

A. Free Space Azimuth Plane Beam-Steering

The phase characteristics of the excited omnidirectional modes are shown in Fig. 11b for the simulated and measured cases in the azimuth plane. In both scenarios, the patterns have the desired dual-phase variations, where the phase changes in two opposing directions as required for the $\vec{K}_{1,\pm 2,2}$ modes (please see P4 and P5), while P1 has a constant phase (as it excites the $\vec{K}_{1,0,1}$ mode). The small discrepancies seen in the measured case may be due to the manufacturing tolerances.

To test the antenna beam-steering capabilities, the main beam was steered towards four different directions separated by 90° to cover the entire plane. Fig. 12 shows the beam-steering performance, where Fig. 12a depicts the beam-steering for $0^\circ/180^\circ$ and $90^\circ/270^\circ$ directions. Fig. 12b highlights the beam-steering for $45^\circ/225^\circ$ and $135^\circ/315^\circ$ directions. The required phase shifts to steer the beams toward the desired directions are computed based on the phase values shown in Fig. 11b and using (8). The calculated phase shifts for the measured beam-steering results are detailed in Table III. A good agreement is realized between the two cases, with the generated main beams covering the desired directions. The few discrepancies seen in the measured cases (beamwidth, shouldering, and dips) may be due to the small phase dips in measurements and other manufacturing tolerances.

B. Free Space Elevation Plane Beam-Steering

Fig. 13a shows the $xz$-plane beam-steering ($\phi = 0^\circ$), while the $yz$-plane ($\phi = 90^\circ$) beam-steering is given in Fig. 13b. This performance is obtained by exciting one phase-varying omnidirectional spherical mode (P4 or P5) and one broadside radiating mode (P2 or P3). For completeness, Table IV outlines the excitation used to generate the elevation plane beam-steering. It can be seen that a slight discrepancy (of up to $2^\circ$) occurs in the direction of the steered beams between...
TABLE IV  
EXCITATIONS FOR ELEVATION PLANE BEAM-STEERING

| Beam direction | \( |A_2| \) | \( \Delta \beta_2 \) | \( |A_4| \) | \( \Delta \beta_4 \) | \( |A_5| \) | \( \Delta \beta_5 \) |
|----------------|-----------|----------------|-----------|----------------|-----------|----------------|
| \( \phi = 0^\circ, \theta = 22^\circ \) | 1 | 0° | - | - | 1 | 0° |
| \( \phi = 0^\circ, \theta = -22^\circ \) | 1 | 0° | - | - | 1 | 180° |
| \( \phi = 90^\circ, \theta = 20^\circ \) | 1 | 0° | 1 | 0° | - | - |
| \( \phi = 90^\circ, \theta = -22^\circ \) | 1 | 0° | 1 | 180° | - | - |

IV. ON-BODY PERFORMANCE

To investigate the antenna performance for on-body applications, a multi-layer phantom is used (see Fig. 14). In the simulation, the phantom comprises three layers: skin layer (thickness of 1.3 mm), fat layer (10.5 mm thick), and muscle layer (20 mm thick), where the dimensions are approximated from [25]. The antenna is first placed in direct contact with the multi-layer phantom, i.e., the gap between the antenna ground plane and the phantom is 0 mm. For measurements, a pork trunk of size 110 mm \( \times \) 70 mm \( \times \) 31 mm is used, and the measurement setup is shown in Fig. 16a.

Fig. 15 shows the S-parameters results for the on-body setup. For both cases, simulations, and measurements, it is seen that the center frequency \( f_0 = 5.75 \text{GHz} \) shifts to lower values compared to the free space case, i.e., 5.73 GHz (on-body simulated), and 5.71 GHz (on-body measured). Due to the phantom, in simulations, the isolation deteriorates to 21 dB, and IBW changes to 67 MHz (compared to 24 dB and 42.2 MHz in free space). For the measured case the isolation and IBW change from 16 dB and 31 MHz in free space to 13.2 dB and 29 MHz with the phantom. Table V shows the total efficiency of each antenna port when the gap between the antenna ground plane and the phantom is increased from 0 to 3 mm, along with the free space case values. It can be seen that the total efficiency increases as the gap increases, and the best values are seen for the free space scenario. Additionally, these results show that the most significant efficiency drop due to the phantom occurs for ports P4 and P5. This may be explained by the dual-phase change of the modes excited by these ports \( (K_{1,0,1} \pm 2.2) \) modes, which require a much larger diameter to support these variations. In contrast, the highest efficiency is seen for the fundamental mode \( (K_{1,0,1}) \) mode excited using

![Fig. 14. Proposed on-body setup using a three-layer phantom.](image)

![Fig. 15. On-body S-parameters results (solid lines: simulations, dashed lines: measurements), port index \( i, j = 1, 2, 3, 4, \) and, 5.](image)

![Fig. 13. Normalized radiation patterns of the elevation plane beam-steering, solid lines (simulations), and dashed lines (measurements): (a) \( \text{xz-plane results}; \) and (b) \( \text{yz-plane results}. \) ](image)
P1), which has a constant phase across the azimuth plane. Nevertheless, the total efficiency of the ports exciting the dual-phase varying modes is still above 58% for the worst investigated case, i.e., with gap = 0 mm.

A. Specific Absorption Rate (SAR)

To investigate the electromagnetic (EM) field exposure, the SAR is computed following the FCC guidelines [38]. The setup is shown in Fig. 17, and for the wrist-worn condition, the antenna is placed in direct contact with a block filled with human hand tissue (\(\epsilon_r = 21.3\), \(\delta = 0.51\) S/m, and \(\rho = 1000\) kg/m\(^3\)), averaged 10 g, with an input power of 10 mW. As shown in Fig. 17, the peak SAR values for each port are 0.0073 W/kg (for P1), 0.0024 W/kg (P2), 0.0032 W/kg (P3), 0.023 W/kg (P4 and P5). Therefore, the antenna peak SAR value is 0.023 W/kg, which is well below the 4 W/kg (FCC) [38] and (ICNIRP) [39] SAR limits.

The results also show that the omnidirectional modes (P1, P4, and P5) have the highest SAR values for this setup compared to the broadside radiating modes. This can be explained by the fact that when the antenna is mounted on the phantom, the omnidirectional modes have strong \(\vec{E}\)-fields along the phantom length and its depth (−z direction), which results in more EM absorption by the human tissue. In contrast, for P2 and P3, since their main beam is pointing towards +z (away from the phantoms’ depth and length), the EM absorption by the tissue is consequently smaller.

B. On-Body Beam-Steering

The on-body measurement setup is shown in Fig. 16a. The measured patterns of the omnidirectional modes are shown in Fig. 16b, and the respective phase properties are shown in Fig. 16c. The measured broadside patterns are shown in Fig. 16d. Due to the phantom, the radiation patterns of the omnidirectional spherical modes exhibit shouldering and dips, especially for P4 and P5. This can also be explained by the SAR analysis, where higher SAR values are seen for the omnidirectional modes. In general, the required phase properties are still satisfied. It should also be mentioned that the reflections from the antenna holder and other materials used for the on-body measurement may also explain the above-described pattern dips, shoulders, and phase asymmetries. Lastly, the main beam in the broadside radiating modes (P2 and P3) is not significantly affected, as it radiates towards +z. However, some impact on the patterns is seen around the (−z) direction, especially for the P3 excited mode. This is because P3 is an x-polarized mode, and the length of the pork trunk is larger along the x-direction than the y-direction. The simulated peak realized gain for the on-body scenario at \(f_0\) is 1.4 dBi (for P1), 7.76 dBi (P2), 7.8 dBi (P3), 2.78 dBi (P4), and 2.91 dBi (P5). These values decrease to 0.81 dBi (for P1), 6.41 dBi (P2), 6.32 dBi (P3), 1.53 dBi (P4), and 1.6 dBi (P5) for the measured setup. It can be seen that a 1.3 dBi realized gain deterioration occurs between the simulated and measured cases. This is most likely due to the manufacturing tolerances and the differences in dielectric properties between the phantom used in simulations and the actual pork trunk that was used for on-body measurements.

Fig. 18 shows the azimuth plane on-body beam-steering for both the simulated and measured cases. The main beam is steered in four directions, separated by 90° to cover the entire plane, with Fig. 18a showing the beam-steering performance for 0°/180° and 90°/270° directions. In comparison, Fig. 18b highlights the performance for 45°/225° and 135°/315° directions. It can be seen that in both cases, simulated and measured, the antenna can direct the beam in all the desired directions, demonstrating good beam-steering characteristics. The small dips in the measured 0°/180° pattern and the beamwidth discrepancies are due to asymmetries of the pork.
TABLE VI
COMPARISONS WITH PREVIOUSLY PUBLISHED ON-BODY ANTENNAS

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Freq. (GHz)</th>
<th>Size ((\lambda \times \lambda))</th>
<th>Profile ((\lambda))</th>
<th>Peak Gain (dBi)</th>
<th>Structure</th>
<th>Elev. Steer</th>
<th>Range</th>
<th>Azim. Steer</th>
<th>Range</th>
</tr>
</thead>
<tbody>
<tr>
<td>[18]</td>
<td>6.0</td>
<td>1.2 × 0.6</td>
<td>0.03</td>
<td>6.69</td>
<td>Planar</td>
<td>Yes</td>
<td>yz (–31(^\circ), 0(^\circ), 30(^\circ))</td>
<td>No</td>
<td>NA</td>
</tr>
<tr>
<td>[23]</td>
<td>2.35/5.8</td>
<td>0.54 × 0.54(^*)</td>
<td>0.012</td>
<td>5.4</td>
<td>Planar</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>[19]</td>
<td>2.4</td>
<td>0.8 × 0.8</td>
<td>0.027</td>
<td>3.9</td>
<td>Planar</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>[6]</td>
<td>2.45/5.8</td>
<td>0.18 × 0.18</td>
<td>0.04</td>
<td>2.74</td>
<td>Non-planar</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>[20]</td>
<td>2.45</td>
<td>0.55 × 0.55</td>
<td>0.051</td>
<td>7.15 sim.</td>
<td>Stacked</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>[21]</td>
<td>2.45</td>
<td>0.39 × 0.39</td>
<td>0.026</td>
<td>3.83</td>
<td>Planar</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>[22]</td>
<td>2.44</td>
<td>0.34 × 0.34</td>
<td>0.056</td>
<td>2</td>
<td>Non-planar</td>
<td>BD,O</td>
<td>NA</td>
<td>No</td>
<td>NA</td>
</tr>
<tr>
<td>[28]</td>
<td>2.45</td>
<td>0.83 × 0.83</td>
<td>0.012</td>
<td>6</td>
<td>Planar</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>[29]</td>
<td>2.4</td>
<td>0.45 × 0.45</td>
<td>0.021</td>
<td>3.86</td>
<td>Planar</td>
<td>B#</td>
<td>NA</td>
<td>O#</td>
<td>NA</td>
</tr>
<tr>
<td>Prop.</td>
<td>5.71</td>
<td>0.77 × 0.77</td>
<td>0.63</td>
<td>6.41</td>
<td>Planar</td>
<td>Yes</td>
<td>xz(32(^\circ), 63(33(^\circ))</td>
<td>Yes</td>
<td>360(^\circ)</td>
</tr>
</tbody>
</table>

\(^*\)Excluding ground plane size, \# dual-pattern synthesis (O-Omnidirectional, B-Broadside, BD-Bidirectional); NA (Not applicable).

phantom, manufacturing tolerances, and the effects of the antenna holder. For completeness, the excitations for each beam direction are shown in Table III.

The simulated and measured results for the elevation plane beam-steering are shown in Fig. 19, where the \(xz\)-plane beam-steering performance is shown in Fig. 19a, while the \(yz\)-plane results are shown in Fig. 19b, both are obtained using similar excitations as those outlined in Table IV. It can be seen that good beam-steering characteristics are realized in both planes. A slight increase in the beamwidth is seen for the measured cases, and a 1\(^\circ\) discrepancy in the main beam direction is observed in the \(yz\)-plane. Such discrepancies may be the result of the combined effects of the phantom, antenna holder, and other manufacturing inaccuracies. The scanning range in the elevation plane for the on-body setup is lower compared to the free space case, changing from 44\(^\circ\) to 32\(^\circ\) (for the \(xz\)-plane) and from 44\(^\circ\) to 33\(^\circ\) (for the \(yz\)-plane). It is interesting to point out that this range can be further enhanced with our proposed beam-steering solution, mainly by adding a ring to excite the omnidirectional spherical modes with higher phase variations.

Finally, Table VI compares the proposed antenna with previously published on-body antennas. It is seen that, unlike previous works, the design is capable of continuous beam-steering over the entire azimuthal plane with good scanning in the elevation plane. Therefore, it represents a significant breakthrough to enable advanced wireless applications in on-body IoT devices.

V. Conclusion

A compact antenna (0.77\(\lambda\) × 0.77\(\lambda\) × 0.03\(\lambda\)) is proposed for 3D beam-steering in on-body IoT applications. By exploiting the theory of spherical modes, advanced beam-steering is demonstrated in the \(xy\)-plane while using planar structures. The method is shown to support beam-steering in the \(xz\) and \(yz\)-planes by exciting two orthogonal broadside modes. For on-body scenarios, the antenna has a 360\(^\circ\) scanning range in the \(xy\)-plane, while a scanning range of 32\(^\circ\) and 33\(^\circ\) is realized in the \(xz\) and \(yz\)-planes, respectively. This performance is achieved with a total efficiency > 58\%, and measured peak realized gain of up to 6.41 dBi. The antenna is manufactured using a single printed layer. The 3D beam-steering characteristics are achieved without externally controlled switches, supporting the generation of multiple simultaneous patterns at the same frequency, i.e. digital beamforming. Overall, the design is compact, low-profile, planar, suitable for packaging with emerging on-body IoT devices, and compatible with many advanced wireless applications like MIMO, angle of arrival-based localization, and physical layer security.
REFERENCES


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