Antenna-on-Package Design: Achieving Near-Isotropic Radiation Pattern and Wide CP Coverage Simultaneously

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Abstract—Internet of Things (IoT) applications require orientation insensitive wireless devices to maintain stable and reliable communication as well as the localization capability using the global positioning system (GPS). For those reasons, antennas providing near-isotropic radiation pattern with wide circular polarization (CP) coverage are high in demand for IoT applications. In this work, we derive conditions for 3-D antenna on package (AoP), through a combination of three orthogonal dipoles model and particle swarm optimization (PSO) technique to achieve near-isotropic radiation pattern and wide CP coverage simultaneously. Design guidelines have been provided, and a practical AoP has been designed on a 3-D cube package by following this guideline. A prototype has been fabricated and the measured results demonstrate an isotropy of $\sim 93\%$ and CP coverage of $\sim 18\%$, which matches well with the predicted results from the theoretical analysis as well as the full-wave electromagnetic simulations.

Index Terms—Circularly polarized antenna, near-isotropic radiation pattern, orientation insensitive antenna.

I. INTRODUCTION

INTERNET of Things (IoT) will bring connectivity to the next level, where billions of smart devices to be designed for making decisions for a large number of applications in an automated fashion. These huge numbers of IoT devices will depend on high speed and uninterrupted wireless communication for their smooth functioning. Thus, a new generation of antennas is required which are space-efficient, low cost, and orientation insensitive in terms of stable performance. Low cost and efficient space can be achieved through the antenna-on-package (AoP) concept. The package of the sensors or circuits is used to realize the antennas in addition to providing the required protection from the environment [1]–[5]. Since the package accommodates the circuits, the battery, and so

on, typically, it is in the shape of a rectangular box, which facilitates the realization of 3-D antennas. As the IoT devices can be placed or dispersed arbitrarily, the antenna performance must not degrade due to the orientation or position of the package. Thus, it is important to design antennas with near-isotropic radiation patterns. Furthermore, for circularly polarized (CP) antennas, the CP coverage must be enhanced as much as possible.

In literature, many papers report antenna designs with near-isotropic radiation pattern [1]–[18]. Near-isotropic radiation pattern is achieved based on various mechanisms, such as folded dipoles [1]–[9], split-ring resonators [10], folded slots [11], [12], and a combination of electric dipoles with slots or loops [13]–[16]. However, these papers only focus on optimizing the radiation pattern coverage and not CP coverage. On the other hand, there are papers focusing only on enhancing the CP coverage, by bending the dipoles [17], [18], double-layer circular patch [19] or with hybrid perturbation method [20], through dielectric resonance [21], by cavity-backed reflector [22], or through conical log spiral [23], [24]. Although some of them have very large CP coverage [25], [26], their radiation patterns have deep nulls and are thus not suitable for isotropic coverage. There are a handful of papers that have attempted to achieve isotropy in radiation pattern and CP at the same time. However, their goals for CP are limited to one direction only, such as boresight, and not for wide-angle coverage [27]–[29]. Also, these figures of merit are not well discussed in the articles, for example, papers [27]–[29] do not have the CP coverage information, and papers [28], [29] do not have isotropy integral information. Moreover, the reported antennas are planar structures and are thus not suitable for AoP applications. It is worth mentioning here, to the best of our knowledge, that there is no report which investigates conditions for simultaneous isotropy and wide CP coverage in radiation pattern for 3-D AoP designs.

In this article, we present a three-dipole model, in which the dipoles are orthogonally placed on a 3-D package. The excitation phases of the dipoles are optimized through the particle swarm optimization (PSO) algorithm to obtain near-isotropic radiation pattern as well as simultaneous enhanced CP coverage. Design guidelines have been provided, and a design example has been shown based on the guidelines. The prototype has been fabricated and measured to verify the

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proposed concept. There is a decent agreement between the measured and the predicted results from the derived theoretical model as well as the EM simulations.

II. THEORETICAL MODEL

A. Isotropy Definitions

Isotropy in the radiation pattern is defined differently in the literature. The smaller gain variation is, the more near-isotropic antenna is. The antenna can be considered to be isotropic or near-isotropic when the gain variation is smaller than 7 dB [3]. Sometimes a major issue with this method is that the feeding cable can cause a minimum which can affect the overall gain variation and thus foul the value of isotropy, while, in most practical antenna designs, the minimum gain is set as the bottom limit $G_{\min}$ to support reliable communication. An alternate way is to calculate the integral isotropy factor by finding the ratio of the surface where the gain deviation exceeds the threshold gain level, TH_{iso} , over a complete sphere surrounding the antenna [30]. The sum of the radiation pattern beams solid angles whose gain between the peak gain (PG) and TH_{iso} below the PG relative to the closed spherical surface surrounding the antenna is a measure of the radiation pattern isotropy. This is reflected mathematically by an integral parameter I^{th-iso} as follows:

$$I^{th-iso} = 100\% \times \iint_{S_1} \frac{\sin \theta d\theta d\varphi}{4\pi}. \quad (1.a)$$

The surface S_1 is determined by

$$S_1 \text{ such that } PG - TH_{iso} \leq Gain(\theta, \varphi) \leq PG.$$

Similarly, the sum of the radiation pattern beam solid angles that satisfies axial ratio (AR) in the range of $1 \leq AR(\theta, \varphi) \leq \sqrt{2}$ [31], [32] relative to closed spherical surface surrounding the antenna is a measure of the radiation pattern CP. A CP coverage factor is introduced as in a normalized integral form I^{CP} , to estimate the CP pattern isotropy. This is reflected mathematically by the following equation:

$$I^{CP} = 100\% \times \iint_{S_2} \frac{\sin \theta d\theta d\varphi}{4\pi}. \quad (1.b)$$

The surface S_2 is determined by

$$S_2 \text{ such that } 1 \leq AR(\theta, \varphi) \leq \sqrt{2}.$$

B. Model of Three Infinitesimal Electric Dipoles

We have previously shown how to achieve near-isotropic AoP design with the help of three orthogonally placed infinitesimal electric dipoles model [5]. Some preliminary work about achieving CP coverage has also been shown in [31] and [33]. In this work, we present the three printed electric dipoles model to achieve simultaneous near isotropy and wide CP coverage by finding the right phase conditions through PSO. Based on the calculated conditions, a practical antenna design with an appropriate feed network is also conducted in this work (shown in Section III).

Fig. 1 shows the three orthogonally placed infinitesimal electric dipoles located arbitrarily in the Cartesian coordinate

Fig. 1. Configuration of X-, Y-, and Z-oriented electrical dipoles.

system (X, Y, and Z). The current densities $\mathbf{J}_x$, $\mathbf{J}_y$, and $\mathbf{J}_z$ of the electrical dipoles oriented along the X, Y, and Z axes and located at points $P_x(X_x, Y_x, Z_x)$, $P_y(X_y, Y_y, Z_y)$, and $P_z(X_z, Y_z, Z_z)$, respectively, are given as follows:

$$\mathbf{J}_x(X, Y, Z) = m_x \delta(X - X_x) \delta(Y - Y_x) \delta(Z - Z_x) \mathbf{x}_0 \quad (2.a)$$

$$\mathbf{J}_y(X, Y, Z) = m_y \delta(X - X_y) \delta(Y - Y_y) \delta(Z - Z_y) \mathbf{y}_0 \quad (2.b)$$

$$\mathbf{J}_z(X, Y, Z) = m_z \delta(X - X_z) \delta(Y - Y_z) \delta(Z - Z_z) \mathbf{z}_0 \quad (2.c)$$

where m_x , m_y , and m_z are the moments of the dipoles along X, Y, and Z axes, respectively, $\delta(\cdot)$ is the Dirac delta-function, $\mathbf{x}_0$, $\mathbf{y}_0$, and $\mathbf{z}_0$ are the unit vectors along the X, Y, and Z axes. The X-oriented dipole has the elevational E_θ and azimuthal E_φ components of the radiation pattern, as shown in the following equations:

$$E_\theta(\theta, \varphi) = -\frac{i\eta k m_x}{4\pi} \cos \theta \cos \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} \quad (3.a)$$

$$E_\varphi(\theta, \varphi) = \frac{i\eta k m_x}{4\pi} \sin \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))}. \quad (3.b)$$

The components of the radiation pattern of the Y-oriented dipole are given in the following equations:

$$E_\theta(\theta, \varphi) = -\frac{i\eta k m_y}{4\pi} \cos \theta \sin \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))} \quad (4.a)$$

$$E_\varphi(\theta, \varphi) = -\frac{i\eta k m_y}{4\pi} \cos \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))}. \quad (4.b)$$

The radiation pattern of the Z-oriented dipole has only a meridional component as in the following equation:

$$E_\theta(\theta, \varphi) = \frac{i\eta k m_z}{4\pi} \sin \theta e^{ikr'_z \cos(\alpha_z(\theta, \varphi))} \quad (4.c)$$

where the phase error coming from the j -oriented dipole positions in spherical coordinates is expressed in [31] through spherical components as follows:

$$r'_j = \sqrt{X_j^2 + Y_j^2 + Z_j^2} \quad (6.a)$$

$$\cos(\alpha_j(\theta, \varphi)) = \sin \theta \sin \theta'_j \cos(\varphi - \varphi'_j) + \cos \theta \cos \theta'_j \quad (6.b)$$

$$\theta'_j = \arccos\left(\frac{Z_j}{r'_j}\right), \quad \varphi'_j = \arctan\left(\frac{Y_j}{X_j}\right). \quad (6.c)$$

We assume that the three orthogonal dipoles have equal moments and choose the moment in such a way to normalize the elevational and azimuthal components, i.e., $m_x = m_y = m_z = i4\pi/(\eta k)$. The X- and Y-oriented dipoles have a phase shift ψ_1 between them, while the X- and Z-oriented dipoles

Fig. 2. RHCP far-field zone sphere coverage for the model of three dipoles.

Fig. 3. LHCP far-field zone sphere coverage for the model of three dipoles.

Fig. 4. Isotropy of the model of three dipoles.

of 28.4% of CP coverage can be achieved for either RHCP or LHCP. In Fig. 4, the gain variations for various combinations of ψ_2 and ψ_1 are moderate (within ~ 12 dBs), with the smallest variation of 3 dBs for some combinations such as $\psi_2 = 60^\circ$ and $\psi_1 = 120^\circ$, or $\psi_2 = 240^\circ$ and $\psi_1 = 120^\circ$. A sharp peak in gain variation is seen when $\psi_1 = 0^\circ$ and $\psi_2 = 0^\circ$, 180° , or 360° , as shown in Fig. 4. This happens because the three dipoles are almost in phase. For the best CP coverage cases (either RHCP or LHCP) and the minimum gain variation of 3 dB, the level of cross-polarization is only 1%. This analysis with the small electric dipoles is a good starting point. However, these dipoles are not practically realizable antenna combination due to the difficulty in developing a feeding system that matches the transmission feed line as well as due to dipoles' low radiation efficiency. This means that the model of the three small dipoles has to be extended to dipoles with practically realizable lengths. Also, the manual sweep method is not time efficient particularly for the dipoles which have finite lengths and are placed in arbitrary locations.

C. Model of Three Half-Wavelength Dipoles

As mentioned in the previous section, dipoles with decent finite lengths must be incorporated in the model for practical antenna design. For this work, it is preferred that matching network (stubs, etc.) is not employed, because the metallic lines of the matching network may influence the radiated fields. To simplify the matching problem, half-wavelength dipoles can be utilized here, because theoretically, the real part of the impedance for a half-wavelength dipole is 73Ω which is relatively easy to match with the typical 50Ω transmission systems. However, for a half-wavelength dipole, contrary to the small dipole, uniform current distribution assumption can no longer be used, rather a sinusoidal distribution must be incorporated in the model.

Then, the current densities for the half-wavelength dipoles along the X, Y, and Z axes are given by the following equations:

$$\mathbf{J}_x(X, Y, Z) = m_x \sin \left[k \left(\frac{L}{2} - |X_x| \right) \right] \delta(Y - Y_x) \delta(Z - Z_x) \mathbf{x}_0 \quad (8.a)$$

have a phase shift of ψ_2 between them. The normalized meridional and azimuthal components of radiation pattern then become

$$E_\theta^\Sigma(\theta, \varphi) = \cos \theta \cos \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} + \cos \theta \sin \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi)) + i\psi_1} - \sin \theta e^{ikr'_z \cos(\alpha_z(\theta, \varphi)) + i\psi_2} \quad (7.a)$$

$$E_\varphi^\Sigma(\theta, \varphi) = -\sin \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} + \cos \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi)) + i\psi_1}. \quad (7.b)$$

The CP coverage of the far-field zone, as defined previously in part A, and the power radiation pattern gain variations are calculated through MATLAB by manually sweeping the phase difference values ψ_1 and ψ_2 . The RHCP and LHCP far-field zone sphere coverage and the gain deviation are illustrated in Figs. 2–4. To simplify the problem and reduce the calculation load, the three dipoles are located at the origin for various phase shifts ψ_1 and ψ_2 . The three-dipole antenna's polarization varies from fully linear to circular polarized. As shown in Fig. 2, the RHCP coverage shows a maximum when ψ_2 is 300° and ψ_1 is equal to 60° . Though other combinations are also possible such as $\psi_2 = 280^\circ$ and $\psi_1 = 45^\circ$, however, the CP coverage percentage is slightly lower. Similarly, for LHCP, the maximum appears for $\psi_2 = 60^\circ$ and $\psi_1 = 120^\circ$, as shown in Fig. 3. From Figs. 2 and 3, we can see that a maximum

$$\mathbf{J}_y(X, Y, Z) = m_y \delta(X - X_y) \sin \left[k \left(\frac{L}{2} - |Y_y| \right) \right] \delta(Z - Z_y) \mathbf{y}_0 \quad (8.b)$$

$$\mathbf{J}_z(X, Y, Z) = m_z \delta(X - X_z) \delta(Y - Y_z) \sin \left[k \left(\frac{L}{2} - |Z_z| \right) \right] \mathbf{z}_0 \quad (8.c)$$

where L is the total length of the half-wavelength dipole. For the half-wavelength dipoles placed along the three axes in the cartesian coordinate system, the normalized fields are given in the following equations:

$$E_{x\theta}(\theta, \varphi) = \frac{2 \cos(\frac{kL}{2} \sin \theta \cos \phi)}{1 - \sin^2 \theta \cos^2 \phi} \cos \theta \cos \phi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} \quad (9.a)$$

$$E_{x\phi}(\theta, \varphi) = -\frac{2 \cos(\frac{kL}{2} \sin \theta \cos \phi)}{1 - \sin^2 \theta \cos^2 \phi} \sin \phi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} \quad (9.b)$$

$$E_{y\theta}(\theta, \varphi) = \frac{2 \cos(\frac{kL}{2} \sin \theta \sin \phi)}{1 - \sin^2 \theta \sin^2 \phi} \cos \theta \sin \phi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))} \quad (9.c)$$

$$E_{y\phi}(\theta, \varphi) = \frac{2 \cos(\frac{kL}{2} \sin \theta \sin \phi)}{1 - \sin^2 \theta \sin^2 \phi} \cos \theta \sin \phi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))} \quad (9.d)$$

$$E_{z\theta}(\theta, \varphi) = -\frac{\cos[\frac{\pi}{2} \cos \theta]}{\sin \theta} e^{ikr'_z \cos(\alpha_z(\theta, \varphi))} \quad (9.e)$$

Finding the right phase conditions for the half-wavelength dipoles is not easy by manually sweeping the phase differences between the dipoles; thus, an automated optimization routine is required for this purpose.

D. Particle Swarm Optimization

PSO is a population-based evolutionary optimization algorithm [34]. This optimization method employs a large number of swarm members to record, share, and compare the individual solutions and then iteratively improve the candidate solution. It can achieve the optimization target for a relatively smaller number of optimization parameters and avoids stagnation at the local optimum point without requiring an initial estimation at the start. The PSO flowchart, pertinent to this work, is shown in Fig. 5. The algorithm begins by defining the parameters, the cost function, and the optimization boundary. The three parameters, isotropy threshold TH_{iso} , its near isotropy coverage limit I^{th-iso} , and the minimum gain G_{min} , are set by designers for their specific environments and requirements. Other than those three customized parameters, the excitation phases of the three dipoles are defined as the particle positions. Another two important vectors Pbest and Gbest are defined to store the intermediate information to generate the new velocity of the particles for the next movements. The former keeps the personal best position and lowest cost value of every single particle, and the latter stores the global best position and lowest cost among all the particles. The velocity of each particle movement is defined as the particle movement step, proportional to the difference between its position and the global best position. The objective cost function F is defined as the reverse function of the CP coverage determined by the antenna radiation model. Then,

Fig. 5. PSO flowchart. * $I^{Pth-iso} > I^{th-iso}$, and $G_P > G_{min}$.

those parameters for a swarm of particles are initialized. The initial positions of all the particles are randomly selected, and the initial velocity is zero. The two vectors, Pbest and Gbest, are set to have empty position values and infinite large cost values. In the next step, a time-domain iteration nested with a space domain iteration is performed. The first particle position and cost function are filled into its Pbest as the space domain iteration begins. The first stop criterion is checked whether the antenna is near-isotropic or not, depending on the isotropy requirement setup by the user. The near-isotropic antenna standard is characterized by G_{min} , TH_{iso} , and I^{th-iso} . $I^{Pth-iso}$ is the temporary value that represents the particle isotropy integral parameter I^{th-iso} , and similarly, G_P is the temporary vector that records the minimum gain across the whole sphere. If $I^{Pth-iso} > I^{th-iso}$, and $G_P > G_{min}$, the radiation pattern is near-isotropic. Otherwise, another space domain iteration is triggered with Particle Index increasing by 1. Once the condition of near isotropy is met, then it moves forward to the next step where the antenna CP coverage is calculated. After that Pbest and Gbest are updated accordingly. Once the space domain iterations are completed, meaning that all the particles' current positions and cost values are calculated, the program runs to the next time-domain iteration. When the number of time-domain iterations reaches the set limit, the required excitation phase conditions for 3 dipoles model and the corresponding CP and isotropy coverage are obtained. In

Fig. 6. Configuration of the three off-origin half-wavelength dipoles on the package. The parameters $L = W = H = 95$ mm are the dimensions of the cubic package for antenna at 1.57 GHz operating frequency.

MATLAB, the computation time could be further reduced by a parallel computation that realizes the space iteration through matrix operation.

For practical antenna design, the three half-wavelength dipoles may be placed away from each other to minimize the coupling between them, as shown in Fig. 6. This is a complicated optimization problem as compared with the small dipoles at the origin because the antennas are electrically large and have been strategically placed on the package in an orthogonal fashion. As can be seen in Figs. 2 and 3, multiple local peaks may appear in the CP coverage optimization; thus, PSO is a good option for this problem because it can easily reach the global maximum to provide phase conditions for simultaneously achieving near isotropy and wide CP coverage.

As shown in Fig. 6, the dipoles are placed on the three orthogonal edges of the cubic package with air as the medium. The parameters W_d and L_d represent the width and total length of the dipoles, respectively. The centers of the half-wavelength dipoles are located at $(W/2, 0, 0)$, $(W, L/2, H)$, and $(0, L, H/2)$ along the X, Y, and Z, respectively. The target of the PSO algorithm is to figure out the phase conditions to maximize the CP coverage while maintaining the radiation pattern isotropy higher than the isotropy limit I^{th-iso} , as defined in (1.a). The optimization procedure minimizes a defined cost function that is a reverse function of the CP coverage factor in (1.b) as given by the following equation:

$$F = \frac{1}{I^{CP}}. \quad (10)$$

The solution is searched within the search-space boundary for good near isotropy that satisfies $I^{Pth-iso} > I^{th-iso}$, and $G_P > G_{min}$. For example, we set $TH_{iso} = 7$ dB, $I^{th-iso} = 90\%$, and $G_{min} = -10$ dBs. The optimized phase difference between the X and Y dipoles is 117° , whereas the phase difference between the X and Z dipoles is 236° , with all three ports fed with equal power level. The theoretical model in Section II-C calculated by MATLAB predicts a 98% isotropy and 21.8% CP coverage. For verification, the optimized phase difference values from the PSO have been applied to a model in Fig. 6 operating at 1.57 GHz utilized in the full-wave EM simulator, Ansys HFSS. The EM simulation results are

Fig. 7. LHCP gain of (a) theoretical model with PSO optimization and (b) HFSS simulation model. RHCP gain of (c) theoretical model with PSO optimization and (d) HFSS simulation model.

close to the predicted values from the theoretical model as can be seen from Figs. 7 and 8, except for some very minor differences. In the example, the simulated LHCP gain and RHCP radiation patterns of the HFSS model, fed with the calculated phase conditions, are compared with the theoretical model in Fig. 7. Both simulated RHCP and LHCP gains are showing good consistent with the theoretically predicted ones. Also, the simulated AR radiation pattern with the theoretically calculated result is shown in Fig. 8. Note that for angles where $0 \text{ dB} < \text{AR} < 3 \text{ dB}$, they will be accounted for CP coverage. As seen from Fig. 8 that the simulated HFSS AR radiation pattern is very similar to theoretical results, which means a good consistency of CP coverage is also obtained between the HFSS model and the theoretical model. This comparison is a good validation of the theoretical model. The 3-D package has been kept hollow from inside to accommodate the driving electronics, which matches the concept of AoP. Though not shown here, the antenna has been simulated with a metal cube (representative of the driving electronics, for example, a PCB with a typical dimension around $(\lambda/8) * (\lambda/8) * (\lambda/8)$ placed either at the corner or the center of the package, λ is the wavelength of the operation frequency) and the performance of the AoP remains almost stable (around 22% CP coverage). Having said that, if a larger metallic structure such as a battery needs to be placed inside the package, then the antenna may need to be optimized a bit, as we have shown in our previous works [3]–[5].

E. Design Guidelines

To make the design process systematic, a design guideline containing basic steps to design a near-isotropic antenna with wide CP polarization simultaneously is given as follows.

First, get the total field expressions of the dipoles groups following the steps.

- 1) Choose any corner of the package as the Cartesian coordinate system origin. The choice of the origin position does not make any difference to the final results.

Fig. 8. CP coverage for three half-wavelength dipoles. (a) Theoretical model with PSO optimization. (b) HFSS simulation model.

- 2) Locate the three half-wavelength dipoles along with the X, Y, and Z directions, respectively, on the surface of the package. Note down the center positions of the three dipoles referred to the origin as r'_j ($j = x, y, \text{ or } z$) in (6.a).
- 3) Use the dipole center positions in (6.a)–(6.c) to calculate the spatial phase errors ($kr'_j \cos(\alpha_j(\theta, \varphi))$).
- 4) Derive the total E-field expressions for the three dipoles through (9.a)–(9.e) using the spatial phase errors calculated in step 3.

Second, derive the excitation phase conditions for each dipole with the PSO technique for the user's requirements. There are four steps to accomplish this procedure as follows.

- 1) The isotropy threshold TH_{iso} , its near isotropy coverage limit I^{th-iso} , and the minimum gain G_{min} can be defined by the readers for their requirements. In our example as mentioned, we choose $TH_{iso} = 7$ dB as isotropy threshold, coverage limit $I^{th-iso} = 90\%$, and $G_{min} = -10$ dB.
- 2) At this point, an optimization method can be used. We have used the PSO technique and the algorithm flow that has been provided in the flowchart (Fig. 5). Utilize the expressions derived in step 4, along with the reader-defined isotropy threshold TH_{iso} , near isotropy coverage limit I^{th-iso} , and minimum gain G_{min} from step 5 in the first decision box in Fig. 5.
- 3) Run the PSO code to get the excitation phase conditions for three half-wavelength dipoles which lead to the maximum CP coverage under the condition defined by the readers.
- 4) Set the excitation phases of the dipoles in the HFSS model using PSO optimization results. Verify the radiation property performance calculated in the PSO with HFSS simulation results.

Third, use the phase conditions derived from the theoretical model to a practical design and the detailed steps are listed as follows.

- 1) Design a feeding network that ensures the required phases (calculated in step 7) reach the dipoles with equal magnitudes.
- 2) Final optimization can now be done for the complete AoP (dipoles with feeding network) in an EM simulator. Adjust the feeding network to optimize the complete system performance, if needed.

A design example is shown by following these guidelines, and an AoP has been designed, simulated, fabricated, and measured in Section III.

III. PRACTICAL ANTENNA DESIGN

In this section, a practical AoP design example is provided based on the design guideline provided in Section II-E to prove the feasibility of our design guidelines.

To simplify the design process and get a better match with the previous simulation, the package dimensions are set to be $L = W = H = 95$ mm, which is slightly longer than a half-wavelength dipole. Low-loss Rogers laminates with a dielectric constant $\epsilon_r = 3.3$, loss tangent $\tan\delta = 0.0015$, and thickness $t = 0.762$ mm have been used for the package fabrications. Following the first step, the origin is selected as the same as the one shown in the PSO example (Fig. 6). This can be changed to another corner and it does not affect the final results. Following step 2, the three half-wavelength dipoles, X, Y, and Z, are located along the three edges of the package, as shown in Fig. 6, and their centers are noted as (47.5 mm, 0, 0), (95, 47.5, and 95 mm), and (0, 95, 47.5 mm), respectively. Then, following step 3, the spatial error of each dipole is calculated by (6.a)–(6.c), and following step 4, the total E-field is expressed using (9.a)–(9.e) with the spatial errors calculated from step 3.

In step 5, the near isotropy requirements are set up by the user for their requirements, and here in our design example, we have set $TH_{iso} = 7$ dB and $I^{th-iso} = 90\%$ to have a better match with the example we showed in the last section. Next, following step 6, the PSO codes are updated with the parameters setup in steps 4 and 5. The readers can apply their optimization programs for the calculation, and here we choose the PSO algorithm shown in Fig. 5 for the phase conditions calculations. Following step 7, the PSO code is executed to get the excitation phase conditions for minimum cost function, which means the optimized CP coverage is obtained. Here, in our example, the phase conditions are $\psi_1 = 117^\circ$ and $\psi_2 = 236^\circ$ with an overall CP coverage of 21.8%. In step 8, the obtained phase conditions are implemented in the HFSS model. For our example, the simulation results in HFSS match the theoretical model prediction, as shown in Figs. 7 and 8.

After getting a good match between the theoretical model and HFSS simulations, the next phase is to design a practical AoP with the appropriate feeding network to feed the three orthogonal dipoles. As shown in step 8, each dipole is fed through an independent lumped port in HFSS simulation. In step 9, a feeding network which provides the phase conditions ($\psi_1 = 117^\circ$ and $\psi_2 = 236^\circ$) is required to realize the final AoP. It is difficult to realize this feed arrangement for practical antenna design as this requires either an individual excitation cable or an individual driving circuit for each dipole. Other than the practical use, for testability as well, three half-wavelength dipoles fed by a single SMA connector are less complex and more convenient than three coaxial lines with three baluns to feed the dipoles individually. Thus, for practical antenna design, it is better to utilize a single driving network. A balun is also required to transit single-ended signal from coaxial cable to differential mode to feed dipole antennas. In step 9, to feed the three orthogonal dipoles on the cubic package with the correct phases, a phase shifter with three parallel output lines has been designed, as shown in Fig. 9. Though the feed network has to be implemented on the 3-D

Fig. 9. Planar phase shifter. (a) Top view, (b) Bottom view.

Fig. 10. Three half-wavelength dipoles with feeding network. Antenna width = $W_d = 2$ mm, antenna length = $L_d = 82$ mm, feeding lines width = $W_f = 1.2$ mm, tapered section width = from 1.2 to 0.3 mm, additional phase shift length for Z-axis dipole = $L_f = 47$ mm.

package, it is initially designed on a planar surface such that when folded, all the squares shown in Fig. 9 become various faces of the cubic package. Each square in Fig. 9 has the same dimensions 95 mm as that of an individual side of the package. To feed the differential antennas, parallel lines have been designed such that one part of the line is on the top side of the substrate, whereas the other part is on the bottom side, as shown in Fig. 9. This arrangement also helps in minimizing the effect of these lines on dipoles' radiation. The parallel line is designed to be as short as possible to further reduce the effects on the radiating field. In Fig. 10, the red parallel line feeds the X-axis dipole, the green one feeds the Y-axis dipole, and the blue one feeds the Z-axis dipole. The width of the parallel lines determines their characteristic impedance and their lengths determine the output port phases. The difference in lengths between the parallel lines provides the phase differences between the output ports, which are the key factors affecting the antenna performance. The tapered part of the parallel lines helps in impedance matching with the dipoles. Once the right phases, according to the theoretical method, have been achieved, the individual substrates are folded to form the cubic package. Simulation results confirm that the folding of the feed network has a negligible effect on its planned performance. The feed network connects to a Bazooka balun at the intersection point of the three parallel lines to transform the differential signals to the single-ended mode for the coaxial cable, as shown in Fig. 10. The length of the Bazooka balun is designed to be a quarter wavelength long at 1.57 GHz, which is $l_b = 28.9$ mm.

After step 9, the three dipoles connected with the feeding network are shown in Fig. 10, where the two dipole's

Fig. 11. Simulated antenna radiation pattern in contour plot.

Fig. 12. Antenna CP coverage at target frequency.

arms are placed on either side of the substrate, consistent with the parallel line layout mentioned above. Despite the precautions are taken with the feed network, the finite width and lines of the feedline do affect the antenna radiation and thus the CP coverage of the antenna with the feeding network is not as good as has been predicted from the theoretical model and initial simulations (without the feed network). Therefore, in step 10, the dimensions of the antennas and phase shifters are optimized slightly to get good CP coverage. The re-optimized phase difference between the X and Y dipoles is 88° , whereas the phase difference between the X and Z dipoles is 276° . The final dimensions for the antenna and the feed network are given in the caption of Fig. 10.

The simulated results for the radiation pattern as well as CP coverage are shown in Figs. 11 and 12. It can be calculated from the simulation result that a 7 dB near isotropy coverage of 92% is achieved which is somewhat lower from the initial prediction of 98% but still acceptable given that the initial prediction did not include the effects of the feeding network. Furthermore, a CP coverage of 21% is achieved which is marginally below the initial prediction of 21.8%. The antenna has a gain of 3 dBi at 1.575 GHz.

IV. FABRICATION AND MEASUREMENTS

The antenna prototype has been created by fabricating six individual faces of the cube with the required metallization and then assembling them in the form of a cube, as shown in Fig. 13. Each board is fabricated individually by LPKF using the low-loss Rogers boards, as shown in Fig. 13(a). The metal traces on the boards are soldered at appropriate junction points to strengthen the metal connection. The bazooka balun is fabricated by realizing the sleeve through the copper tape on the coaxial line. The balun is then soldered to the parallel traces on the bottom plate. Finally, all the faces of the cube

Fig. 13. (a) Fabricated six roger board with metallic patterns for antenna and phase shifter. (b) Glued antenna on package. (c) Antenna on package with bazooka balun.

Fig. 14. Measured and simulated S-parameters of the total antenna system.

are attached with the help of superglue to form the cube. The complete prototype is shown in Fig. 13(b) and (c).

The input reflection coefficient of the antenna has been measured through Agilent VNA Keysight N9912A and the results are shown in Fig. 14. As can be seen from the measurement results, the antenna is matched to 50Ω from 1.34 to 1.81 GHz, which is quite similar to the simulated bandwidth. Overall, there is a good match between the simulation and measurement results. However, there is a slight shift in the measured S_{11} curve (approximately 50 MHz) around the center frequency, which could be due to the fabrication tolerances. The other difference is that it is very well-matched resonance at 1.52 GHz, which does exist but is not that prominent in the simulations. It is worth mentioning here that the narrowest parts of the parallel strip line are 0.3 mm and they have been connected to the bazooka balun through soldering which may have introduced some additional differences. Moreover, the

Fig. 15. (a) Radiation pattern measurement setup in Satimo Chamber. (b) 3-D radiation pattern.

Fig. 16. Measured CP coverage.

TABLE I
COMPARISON BETWEEN SIMULATED AND MEASURED RESULTS

	7dB isotropy	CP coverage
Simulation	94.01%	17.3%
Measurement	92.86%	17.2%

bazooka balun has been manually fabricated and it appears that the shorted stub length is not as accurate as has been simulated.

To obtain the radiation parameters, the proposed antenna is measured in Satimo anechoic chamber, as shown in Fig. 15(a). The best isotropy and CP coverage performance is observed at 1.52 GHz instead of the center frequency of 1.57 GHz. This 50 MHz shift is consistent with the impedance measurements, as described above. The measured 3-D radiation pattern at 1.52 GHz is shown in Fig. 15(b). The measured 7 dB isotropy at 1.52 GHz is 92.86%, which is quite close to the simulation results, as can be seen in Table I. The antenna's measured CP coverage is shown in Fig. 16, with a CP coverage of 17.2% at 1.52 GHz. This value is also quite close to the simulated value at 1.52 GHz; however, it is somewhat lower than the 21% CP coverage predicted by simulations at 1.57 GHz. As explained above, this may be due to the 50 MHz frequency shift observed in measurements. Nonetheless, in simulations, the performance is quite similar at 1.52 GHz, which validates the proposed theory and design concept.

It is well known that antennas' radiation performance gets affected by nearby objects and the proposed antenna is not exempted from that. However, if the antenna system is smart enough to sense the environment, then the phases of the signals fed to the individual elements can be adjusted to reconfigure the radiation pattern accordingly.

V. CONCLUSION

A combination of three orthogonal dipoles' model and PSO technique has been utilized in this work to design 3-D AoP with near-isotropic radiation pattern and wide circular polarization coverage. Appropriate phase conditions to achieve maximum CP coverage and decent isotropy from the theoretical model are then utilized to demonstrate a practical AoP. A design guideline is provided for readers to design AoP which can achieve maximum CP and isotropy coverage simultaneously. A design example has been presented based on the design guidelines. The prototype shows a decent performance and matches well with the predicted performance from the theoretical analysis. The methodology presented in this work can be utilized to design and optimize antennas on 3-D packages for future IoT applications. Also, the radiating structure can be replaced by other suitable antenna types and thus more advanced functions, such as different orbital angular momentum beams, can be realized [35].

REFERENCES

- [1] C. M. Kruesi, R. J. Vyas, and M. M. Tentzeris, "Design and development of a novel 3-D cubic antenna for wireless sensor networks (WSNs) and RFID applications," *IEEE Trans. Antennas Propag.*, vol. 57, no. 10, pp. 3293–3299, Oct. 2009.
- [2] I. T. Nassar and T. M. Weller, "Development of novel 3-D cube antennas for compact wireless sensor nodes," *IEEE Trans. Antennas Propag.*, vol. 60, no. 2, pp. 1059–1065, Feb. 2012.
- [3] M. F. Farooqui, C. Claudel, and A. Shamim, "An inkjet-printed buoyant 3-D lagrangian sensor for real-time flood monitoring," *IEEE Trans. Antennas Propag.*, vol. 62, no. 6, pp. 3354–3359, Jun. 2014.
- [4] M. F. Farooqui, M. A. Karimi, K. N. Salama, and A. Shamim, "3D-printed disposable wireless sensors with integrated microelectronics for large area environmental monitoring," *Adv. Mater. Technol.*, vol. 2, no. 8, 2017, Art. no. 1700051.
- [5] Z. Su, K. Klionovski, R. M. Bilal, and A. Shamim, "A dual band additively manufactured 3-D antenna on package with near-isotropic radiation pattern," *IEEE Trans. Antennas Propag.*, vol. 66, no. 7, pp. 3295–3305, Jul. 2018.
- [6] C. Cho, H. Choo, and I. Park, "Printed symmetric inverted-F antenna with a quasi-isotropic radiation pattern," *Microw. Opt. Technol. Lett.*, vol. 50, no. 4, pp. 927–930, 2008.
- [7] H.-K. Ryu, G. Jung, D.-K. Ju, S. Lim, and J.-M. Woo, "An electrically small spherical UHF RFID tag antenna with quasi-isotropic patterns for wireless sensor networks," *IEEE Antennas Wireless Propag. Lett.*, vol. 9, pp. 60–62, 2010.
- [8] G. Pan, Y. Li, Z. Zhang, and Z. Feng, "Isotropic radiation from a compact planar antenna using two crossed dipoles," *IEEE Antennas Wireless Propag. Lett.*, vol. 11, pp. 1338–1341, 2012.
- [9] C. Deng, Y. Li, Z. Zhang, and Z. Feng, "A wideband isotropic radiated planar antenna using sequential rotated L-shaped monopoles," *IEEE Trans. Antennas Propag.*, vol. 62, no. 3, pp. 1461–1464, Mar. 2014.
- [10] J.-H. Kim and S. Nam, "A compact quasi-isotropic antenna based on folded split-ring resonators," *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 294–297, 2017.
- [11] C. Deng, Y. Li, Z. Zhang, and Z. Feng, "Design of a three-dimensional folded slot antenna with quasi-isotropic radiation pattern," in *Proc. IEEE Int. Symp. Antennas Propag. USNC/URSI Nat. Radio Sci. Meeting*, Jul. 2015, pp. 588–589.
- [12] S. Shamsinejad, F. De Flaviis, and P. Mousavi, "Microstrip-fed 3-D folded slot antenna on cubic structure," *IEEE Antennas Wireless Propag. Lett.*, vol. 15, pp. 1081–1084, 2016.
- [13] S. Long, "A combination of linear and slot antennas for quasi-isotropic coverage," *IEEE Trans. Antennas Propag.*, vol. 23, no. 4, pp. 572–576, Jul. 1975.
- [14] C. Cho, H. Choo, and I. Park, "Broadband RFID tag antenna with quasi-isotropic radiation pattern," *Electron. Lett.*, vol. 41, no. 20, pp. 1091–1092, Sep. 2005.
- [15] L. Liang and S. V. Hum, "A low-profile antenna with quasi-isotropic pattern for UHF RFID applications," *IEEE Antennas Wireless Propag. Lett.*, vol. 12, pp. 210–213, 2013.
- [16] C.-H. Wu and T.-G. Ma, "Miniaturized self-oscillating active integrated antenna with quasi-isotropic radiation," *IEEE Trans. Antennas Propag.*, vol. 62, no. 2, pp. 933–936, Feb. 2014.
- [17] B. Wang, S. W. Cheung, W. Wang, M. Li, and T. I. Yuk, "A broadband and wide beamwidth dual circularly polarized antenna using crossed bent dipoles," in *Proc. 11th Eur. Conf. Antennas Propag. (EUCAP)*, Mar. 2017, pp. 1936–1939.
- [18] X.-Y. He, L. Chang, and L.-L. Chen, "A multifunction broad-beam antenna with dual bands and dual circular-polarizations," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Mar. 2016, pp. 1–4.
- [19] Y. Jie, L. Chun-lan, H. Yong-jin, J. Jian-xin, and J. Wei-wei, "Design and analysis of a broad band wide beam circular polarization microstrip antenna," in *Proc. Int. Conf. Microw. Millim. Wave Technol. (ICMMT)*, May 2012, pp. 1–4.
- [20] D. Fan, Z. Wang, B. Huang, and W. Zhang, "Design of broad beam circular-polarized microstrip antenna," in *Proc. Int. Symp. Antennas Propag.*, vol. 1, Nanjing, Oct. 2013, pp. 439–441.
- [21] H. Nguyen-Bich, H. Chu-Ba, and C. Dao-Ngoc, "All GPS bands dielectric resonator antenna with circular polarization for GPS application," in *Proc. Int. Conf. Adv. Technol. Commun.*, Oct. 2012, pp. 65–68.
- [22] K. Saurav, D. Sarkar, A. Singh, and K. V. Srivastava, "Multiband circularly polarized cavity-backed crossed dipole antenna," *IEEE Trans. Antennas Propag.*, vol. 63, no. 10, pp. 4286–4296, Oct. 2015.
- [23] J. Cardoso and A. Safaai-Jazi, "A spherical helical antenna," in *Proc. IEEE Antennas Propag. Soc. Int. Symp.*, Jun./Jul. 1993, pp. 1558–1561.
- [24] A. Gu, S. Yang, and Z. Nie, "Analysis and design of miniaturized ultra-wideband conical log spiral antennas," in *Proc. Cross Strait Quad-Regional Radio Sci. Wireless Technol. Conf.*, Jul. 2013, pp. 191–194.
- [25] G. H. Brown and O. M. Woodward, "Circularly-polarized omnidirectional antenna," *RCA Rev.*, vol. 8, no. 2, pp. 259–269, Jun. 1947.
- [26] V. Galindo and K. Green, "A near-isotropic circularly polarized antenna for space vehicles," *IEEE Trans. Antennas Propag.*, vol. 13, no. 6, pp. 872–877, Nov. 1965.
- [27] M. Huchard, C. Delaveaud, and S. Tedjini, "Miniature antenna for circularly polarized quasi isotropic coverage," in *Proc. 2nd Eur. Conf. Antennas Propag. (EuCAP)*, Nov. 2007, pp. 1–5.
- [28] H. H. Tran, S. X. Ta, and I. Park, "A compact circularly polarized crossed-dipole antenna for an RFID tag," *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 674–677, 2015.
- [29] S. Choi and K. Sarabandi, "A W-Shaped antenna with spatial polarization variation for direction finding," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 12, pp. 2429–2433, Dec. 2018.
- [30] D. Bugnolo, "A quasi-'Isotropic' antenna in the microwave spectrum," *IRE Trans. Antennas Propag.*, vol. 10, no. 4, pp. 377–383, Jul. 1962.
- [31] K. Klionovski, M. B. Arboleda, Z. Su, and A. Shamim, "On synthesis of orientation insensitive antennas," in *Proc. IEEE Indian Conf. Antennas Propagation (InCAP)*, Dec. 2019, pp. 1–4.
- [32] C. Balanis, *Antenna Theory: Analysis and Design*, 4th ed. Hoboken, NJ, USA: Wiley, 2016.
- [33] Z. Su, K. Klionovski, H. Liao, Y. Chen, A. Elsherbeni, and A. Shamim, "3D antenna in package design: Maximizing radiation pattern isotropy and CP coverage," in *Proc. IEEE Int. Symp. Antennas Propag. USNC/URSI Nat. Radio Sci. Meeting*, to be published.
- [34] D. Wang, D. Tan, and L. Liu, "Particle swarm optimization algorithm: An overview," *Soft Comput.*, vol. 22, pp. 387–408, Jan. 2018, doi: 10.1007/s00500-016-2474-6.
- [35] A. F. Morabito, L. Di Donato, and T. Isernia, "Orbital angular momentum antennas: Understanding actual possibilities through the aperture antennas theory," *IEEE Antennas Propag. Mag.*, vol. 60, no. 2, pp. 59–67, Apr. 2018.

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AUTHOR QUERIES

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Antenna-on-Package Design: Achieving Near-Isotropic Radiation Pattern and Wide CP Coverage Simultaneously

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Abstract—Internet of Things (IoT) applications require orientation insensitive wireless devices to maintain stable and reliable communication as well as the localization capability using the global positioning system (GPS). For those reasons, antennas providing near-isotropic radiation pattern with wide circular polarization (CP) coverage are high in demand for IoT applications. In this work, we derive conditions for 3-D antenna on package (AoP), through a combination of three orthogonal dipoles model and particle swarm optimization (PSO) technique to achieve near-isotropic radiation pattern and wide CP coverage simultaneously. Design guidelines have been provided, and a practical AoP has been designed on a 3-D cube package by following this guideline. A prototype has been fabricated and the measured results demonstrate an isotropy of $\sim 93\%$ and CP coverage of $\sim 18\%$, which matches well with the predicted results from the theoretical analysis as well as the full-wave electromagnetic simulations.

Index Terms—Circularly polarized antenna, near-isotropic radiation pattern, orientation insensitive antenna.

I. INTRODUCTION

INTERNET of Things (IoT) will bring connectivity to the next level, where billions of smart devices to be designed for making decisions for a large number of applications in an automated fashion. These huge numbers of IoT devices will depend on high speed and uninterrupted wireless communication for their smooth functioning. Thus, a new generation of antennas is required which are space-efficient, low cost, and orientation insensitive in terms of stable performance. Low cost and efficient space can be achieved through the antenna-on-package (AoP) concept. The package of the sensors or circuits is used to realize the antennas in addition to providing the required protection from the environment [1]–[5]. Since the package accommodates the circuits, the battery, and so

on, typically, it is in the shape of a rectangular box, which facilitates the realization of 3-D antennas. As the IoT devices can be placed or dispersed arbitrarily, the antenna performance must not degrade due to the orientation or position of the package. Thus, it is important to design antennas with near-isotropic radiation patterns. Furthermore, for circularly polarized (CP) antennas, the CP coverage must be enhanced as much as possible.

In literature, many papers report antenna designs with near-isotropic radiation pattern [1]–[18]. Near-isotropic radiation pattern is achieved based on various mechanisms, such as folded dipoles [1]–[9], split-ring resonators [10], folded slots [11], [12], and a combination of electric dipoles with slots or loops [13]–[16]. However, these papers only focus on optimizing the radiation pattern coverage and not CP coverage. On the other hand, there are papers focusing only on enhancing the CP coverage, by bending the dipoles [17], [18], double-layer circular patch [19] or with hybrid perturbation method [20], through dielectric resonance [21], by cavity-backed reflector [22], or through conical log spiral [23], [24]. Although some of them have very large CP coverage [25], [26], their radiation patterns have deep nulls and are thus not suitable for isotropic coverage. There are a handful of papers that have attempted to achieve isotropy in radiation pattern and CP at the same time. However, their goals for CP are limited to one direction only, such as boresight, and not for wide-angle coverage [27]–[29]. Also, these figures of merit are not well discussed in the articles, for example, papers [27]–[29] do not have the CP coverage information, and papers [28], [29] do not have isotropy integral information. Moreover, the reported antennas are planar structures and are thus not suitable for AoP applications. It is worth mentioning here, to the best of our knowledge, that there is no report which investigates conditions for simultaneous isotropy and wide CP coverage in radiation pattern for 3-D AoP designs.

In this article, we present a three-dipole model, in which the dipoles are orthogonally placed on a 3-D package. The excitation phases of the dipoles are optimized through the particle swarm optimization (PSO) algorithm to obtain near-isotropic radiation pattern as well as simultaneous enhanced CP coverage. Design guidelines have been provided, and a design example has been shown based on the guidelines. The prototype has been fabricated and measured to verify the

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proposed concept. There is a decent agreement between the measured and the predicted results from the derived theoretical model as well as the EM simulations.

II. THEORETICAL MODEL

A. Isotropy Definitions

Isotropy in the radiation pattern is defined differently in the literature. The smaller gain variation is, the more near-isotropic antenna is. The antenna can be considered to be isotropic or near-isotropic when the gain variation is smaller than 7 dB [3]. Sometimes a major issue with this method is that the feeding cable can cause a minimum which can affect the overall gain variation and thus foul the value of isotropy, while, in most practical antenna designs, the minimum gain is set as the bottom limit $G_{\min}$ to support reliable communication. An alternate way is to calculate the integral isotropy factor by finding the ratio of the surface where the gain deviation exceeds the threshold gain level, TH_{iso} , over a complete sphere surrounding the antenna [30]. The sum of the radiation pattern beams solid angles whose gain between the peak gain (PG) and TH_{iso} below the PG relative to the closed spherical surface surrounding the antenna is a measure of the radiation pattern isotropy. This is reflected mathematically by an integral parameter I^{th-iso} as follows:

$$I^{th-iso} = 100\% \times \iint_{S_1} \frac{\sin \theta d\theta d\varphi}{4\pi}. \quad (1.a)$$

The surface S_1 is determined by

$$S_1 \text{ such that } PG - TH_{iso} \leq Gain(\theta, \varphi) \leq PG.$$

Similarly, the sum of the radiation pattern beam solid angles that satisfies axial ratio (AR) in the range of $1 \leq AR(\theta, \varphi) \leq \sqrt{2}$ [31], [32] relative to closed spherical surface surrounding the antenna is a measure of the radiation pattern CP. A CP coverage factor is introduced as in a normalized integral form I^{CP} , to estimate the CP pattern isotropy. This is reflected mathematically by the following equation:

$$I^{CP} = 100\% \times \iint_{S_2} \frac{\sin \theta d\theta d\varphi}{4\pi}. \quad (1.b)$$

The surface S_2 is determined by

$$S_2 \text{ such that } 1 \leq AR(\theta, \varphi) \leq \sqrt{2}.$$

B. Model of Three Infinitesimal Electric Dipoles

We have previously shown how to achieve near-isotropic AoP design with the help of three orthogonally placed infinitesimal electric dipoles model [5]. Some preliminary work about achieving CP coverage has also been shown in [31] and [33]. In this work, we present the three printed electric dipoles model to achieve simultaneous near isotropy and wide CP coverage by finding the right phase conditions through PSO. Based on the calculated conditions, a practical antenna design with an appropriate feed network is also conducted in this work (shown in Section III).

Fig. 1 shows the three orthogonally placed infinitesimal electric dipoles located arbitrarily in the Cartesian coordinate

Fig. 1. Configuration of X-, Y-, and Z-oriented electrical dipoles.

system (X, Y, and Z). The current densities $\mathbf{J}_x$, $\mathbf{J}_y$, and $\mathbf{J}_z$ of the electrical dipoles oriented along the X, Y, and Z axes and located at points $P_x(X_x, Y_x, Z_x)$, $P_y(X_y, Y_y, Z_y)$, and $P_z(X_z, Y_z, Z_z)$, respectively, are given as follows:

$$\mathbf{J}_x(X, Y, Z) = m_x \delta(X - X_x) \delta(Y - Y_x) \delta(Z - Z_x) \mathbf{x}_0 \quad (2.a)$$

$$\mathbf{J}_y(X, Y, Z) = m_y \delta(X - X_y) \delta(Y - Y_y) \delta(Z - Z_y) \mathbf{y}_0 \quad (2.b)$$

$$\mathbf{J}_z(X, Y, Z) = m_z \delta(X - X_z) \delta(Y - Y_z) \delta(Z - Z_z) \mathbf{z}_0 \quad (2.c)$$

where m_x , m_y , and m_z are the moments of the dipoles along X, Y, and Z axes, respectively, $\delta(\cdot)$ is the Dirac delta-function, $\mathbf{x}_0$, $\mathbf{y}_0$, and $\mathbf{z}_0$ are the unit vectors along the X, Y, and Z axes. The X-oriented dipole has the elevational E_θ and azimuthal E_φ components of the radiation pattern, as shown in the following equations:

$$E_\theta(\theta, \varphi) = -\frac{i\eta k m_x}{4\pi} \cos \theta \cos \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} \quad (3.a)$$

$$E_\varphi(\theta, \varphi) = \frac{i\eta k m_x}{4\pi} \sin \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))}. \quad (3.b)$$

The components of the radiation pattern of the Y-oriented dipole are given in the following equations:

$$E_\theta(\theta, \varphi) = -\frac{i\eta k m_y}{4\pi} \cos \theta \sin \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))} \quad (4.a)$$

$$E_\varphi(\theta, \varphi) = -\frac{i\eta k m_y}{4\pi} \cos \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))}. \quad (4.b)$$

The radiation pattern of the Z-oriented dipole has only a meridional component as in the following equation:

$$E_\theta(\theta, \varphi) = \frac{i\eta k m_z}{4\pi} \sin \theta e^{ikr'_z \cos(\alpha_z(\theta, \varphi))} \quad (4.c)$$

where the phase error coming from the j -oriented dipole positions in spherical coordinates is expressed in [31] through spherical components as follows:

$$r'_j = \sqrt{X_j^2 + Y_j^2 + Z_j^2} \quad (6.a)$$

$$\cos(\alpha_j(\theta, \varphi)) = \sin \theta \sin \theta'_j \cos(\varphi - \varphi'_j) + \cos \theta \cos \theta'_j \quad (6.b)$$

$$\theta'_j = \arccos\left(\frac{Z_j}{r'_j}\right), \quad \varphi'_j = \arctan\left(\frac{Y_j}{X_j}\right). \quad (6.c)$$

We assume that the three orthogonal dipoles have equal moments and choose the moment in such a way to normalize the elevational and azimuthal components, i.e., $m_x = m_y = m_z = i4\pi/(\eta k)$. The X- and Y-oriented dipoles have a phase shift ψ_1 between them, while the X- and Z-oriented dipoles

Fig. 2. RHCP far-field zone sphere coverage for the model of three dipoles.

Fig. 3. LHCP far-field zone sphere coverage for the model of three dipoles.

have a phase shift of ψ_2 between them. The normalized meridional and azimuthal components of radiation pattern then become

$$E_{\theta}^{\Sigma}(\theta, \varphi) = \cos \theta \cos \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} + \cos \theta \sin \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi)) + i\psi_1} - \sin \theta e^{ikr'_z \cos(\alpha_z(\theta, \varphi)) + i\psi_2} \quad (7.a)$$

$$E_{\varphi}^{\Sigma}(\theta, \varphi) = -\sin \varphi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} + \cos \varphi e^{ikr'_y \cos(\alpha_y(\theta, \varphi)) + i\psi_1}. \quad (7.b)$$

The CP coverage of the far-field zone, as defined previously in part A, and the power radiation pattern gain variations are calculated through MATLAB by manually sweeping the phase difference values ψ_1 and ψ_2 . The RHCP and LHCP far-field zone sphere coverage and the gain deviation are illustrated in Figs. 2–4. To simplify the problem and reduce the calculation load, the three dipoles are located at the origin for various phase shifts ψ_1 and ψ_2 . The three-dipole antenna's polarization varies from fully linear to circular polarized. As shown in Fig. 2, the RHCP coverage shows a maximum when ψ_2 is 300° and ψ_1 is equal to 60°. Though other combinations are also possible such as $\psi_2 = 280^\circ$ and $\psi_1 = 45^\circ$, however, the CP coverage percentage is slightly lower. Similarly, for LHCP, the maximum appears for $\psi_2 = 60^\circ$ and $\psi_1 = 120^\circ$, as shown in Fig. 3. From Figs. 2 and 3, we can see that a maximum

Fig. 4. Isotropy of the model of three dipoles.

of 28.4% of CP coverage can be achieved for either RHCP or LHCP. In Fig. 4, the gain variations for various combinations of ψ_2 and ψ_1 are moderate (within ~12 dBs), with the smallest variation of 3 dBs for some combinations such as $\psi_2 = 60^\circ$ and $\psi_1 = 120^\circ$, or $\psi_2 = 240^\circ$ and $\psi_1 = 120^\circ$. A sharp peak in gain variation is seen when $\psi_1 = 0^\circ$ and $\psi_2 = 0^\circ$, 180°, or 360°, as shown in Fig. 4. This happens because the three dipoles are almost in phase. For the best CP coverage cases (either RHCP or LHCP) and the minimum gain variation of 3 dB, the level of cross-polarization is only 1%. This analysis with the small electric dipoles is a good starting point. However, these dipoles are not practically realizable antenna combination due to the difficulty in developing a feeding system that matches the transmission feed line as well as due to dipoles' low radiation efficiency. This means that the model of the three small dipoles has to be extended to dipoles with practically realizable lengths. Also, the manual sweep method is not time efficient particularly for the dipoles which have finite lengths and are placed in arbitrary locations.

C. Model of Three Half-Wavelength Dipoles

As mentioned in the previous section, dipoles with decent finite lengths must be incorporated in the model for practical antenna design. For this work, it is preferred that matching network (stubs, etc.) is not employed, because the metallic lines of the matching network may influence the radiated fields. To simplify the matching problem, half-wavelength dipoles can be utilized here, because theoretically, the real part of the impedance for a half-wavelength dipole is 73 Ω which is relatively easy to match with the typical 50 Ω transmission systems. However, for a half-wavelength dipole, contrary to the small dipole, uniform current distribution assumption can no longer be used, rather a sinusoidal distribution must be incorporated in the model.

Then, the current densities for the half-wavelength dipoles along the X, Y, and Z axes are given by the following equations:

$$\mathbf{J}_x(X, Y, Z) = m_x \sin \left[k \left(\frac{L}{2} - |X_x| \right) \right] \delta(Y - Y_x) \delta(Z - Z_x) \mathbf{x}_0 \quad (8.a)$$

$$\mathbf{J}_y(X, Y, Z) = m_y \delta(X - X_y) \sin \left[k \left(\frac{L}{2} - |Y_y| \right) \right] \delta(Z - Z_y) \mathbf{y}_0 \quad (8.b)$$

$$\mathbf{J}_z(X, Y, Z) = m_z \delta(X - X_z) \delta(Y - Y_z) \sin \left[k \left(\frac{L}{2} - |Z_z| \right) \right] \mathbf{z}_0 \quad (8.c)$$

where L is the total length of the half-wavelength dipole. For the half-wavelength dipoles placed along the three axes in the cartesian coordinate system, the normalized fields are given in the following equations:

$$E_{x\theta}(\theta, \varphi) = \frac{2 \cos\left(\frac{kL}{2} \sin \theta \cos \phi\right)}{1 - \sin^2 \theta \cos^2 \phi} \cos \theta \cos \phi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} \quad (9.a)$$

$$E_{x\varphi}(\theta, \varphi) = -\frac{2 \cos\left(\frac{kL}{2} \sin \theta \cos \phi\right)}{1 - \sin^2 \theta \cos^2 \phi} \sin \phi e^{ikr'_x \cos(\alpha_x(\theta, \varphi))} \quad (9.b)$$

$$E_{y\theta}(\theta, \varphi) = \frac{2 \cos\left(\frac{kL}{2} \sin \theta \sin \phi\right)}{1 - \sin^2 \theta \sin^2 \phi} \cos \theta \sin \phi e^{ikr'_y \cos(\alpha_y(\theta, \varphi))} \quad (9.c)$$

$$E_{y\varphi}(\theta, \varphi) = \frac{2 \cos\left(\frac{kL}{2} \sin \theta \sin \phi\right)}{1 - \sin^2 \theta \sin^2 \phi} \cos \theta e^{ikr'_y \cos(\alpha_y(\theta, \varphi))} \quad (9.d)$$

$$E_{z\theta}(\theta, \varphi) = -\frac{\cos\left[\frac{\pi}{2} \cos \theta\right]}{\sin \theta} e^{ikr'_z \cos(\alpha_z(\theta, \varphi))}. \quad (9.e)$$

Finding the right phase conditions for the half-wavelength dipoles is not easy by manually sweeping the phase differences between the dipoles; thus, an automated optimization routine is required for this purpose.

D. Particle Swarm Optimization

PSO is a population-based evolutionary optimization algorithm [34]. This optimization method employs a large number of swarm members to record, share, and compare the individual solutions and then iteratively improve the candidate solution. It can achieve the optimization target for a relatively smaller number of optimization parameters and avoids stagnation at the local optimum point without requiring an initial estimation at the start. The PSO flowchart, pertinent to this work, is shown in Fig. 5. The algorithm begins by defining the parameters, the cost function, and the optimization boundary. The three parameters, isotropy threshold TH_{iso} , its near isotropy coverage limit I^{th-iso} , and the minimum gain G_{min} , are set by designers for their specific environments and requirements. Other than those three customized parameters, the excitation phases of the three dipoles are defined as the particle positions. Another two important vectors Pbest and Gbest are defined to store the intermediate information to generate the new velocity of the particles for the next movements. The former keeps the personal best position and lowest cost value of every single particle, and the latter stores the global best position and lowest cost among all the particles. The velocity of each particle movement is defined as the particle movement step, proportional to the difference between its position and the global best position. The objective cost function F is defined as the reverse function of the CP coverage determined by the antenna radiation model. Then,

Fig. 5. PSO flowchart. * $I^{Pth-iso} > I^{th-iso}$, and $G_P > G_{min}$.

those parameters for a swarm of particles are initialized. The initial positions of all the particles are randomly selected, and the initial velocity is zero. The two vectors, Pbest and Gbest, are set to have empty position values and infinite large cost values. In the next step, a time-domain iteration nested with a space domain iteration is performed. The first particle position and cost function are filled into its Pbest as the space domain iteration begins. The first stop criterion is checked whether the antenna is near-isotropic or not, depending on the isotropy requirement setup by the user. The near-isotropic antenna standard is characterized by G_{min} , TH_{iso} , and I^{th-iso} . $I^{Pth-iso}$ is the temporary value that represents the particle isotropy integral parameter I^{th-iso} , and similarly, G_P is the temporary vector that records the minimum gain across the whole sphere. If $I^{Pth-iso} > I^{th-iso}$, and $G_P > G_{min}$, the radiation pattern is near-isotropic. Otherwise, another space domain iteration is triggered with Particle Index increasing by 1. Once the condition of near isotropy is met, then it moves forward to the next step where the antenna CP coverage is calculated. After that Pbest and Gbest are updated accordingly. Once the space domain iterations are completed, meaning that all the particles' current positions and cost values are calculated, the program runs to the next time-domain iteration. When the number of time-domain iterations reaches the set limit, the required excitation phase conditions for 3 dipoles model and the corresponding CP and isotropy coverage are obtained. In

Fig. 6. Configuration of the three off-origin half-wavelength dipoles on the package. The parameters $L = W = H = 95$ mm are the dimensions of the cubic package for antenna at 1.57 GHz operating frequency.

MATLAB, the computation time could be further reduced by a parallel computation that realizes the space iteration through matrix operation.

For practical antenna design, the three half-wavelength dipoles may be placed away from each other to minimize the coupling between them, as shown in Fig. 6. This is a complicated optimization problem as compared with the small dipoles at the origin because the antennas are electrically large and have been strategically placed on the package in an orthogonal fashion. As can be seen in Figs. 2 and 3, multiple local peaks may appear in the CP coverage optimization; thus, PSO is a good option for this problem because it can easily reach the global maximum to provide phase conditions for simultaneously achieving near isotropy and wide CP coverage.

As shown in Fig. 6, the dipoles are placed on the three orthogonal edges of the cubic package with air as the medium. The parameters W_d and L_d represent the width and total length of the dipoles, respectively. The centers of the half-wavelength dipoles are located at $(W/2, 0, 0)$, $(0, L/2, 0)$, and $(0, 0, H/2)$ along the X, Y, and Z, respectively. The target of the PSO algorithm is to figure out the phase conditions to maximize the CP coverage while maintaining the radiation pattern isotropy higher than the isotropy limit I^{th-iso} , as defined in (1.a). The optimization procedure minimizes a defined cost function that is a reverse function of the CP coverage factor in (1.b) as given by the following equation:

$$F = \frac{1}{I^{CP}}. \quad (10)$$

The solution is searched within the search-space boundary for good near isotropy that satisfies $I^{Pth-iso} > I^{th-iso}$, and $G_P > G_{min}$. For example, we set $TH_{iso} = 7$ dB, $I^{th-iso} = 90\%$, and $G_{min} = -10$ dBs. The optimized phase difference between the X and Y dipoles is 117° , whereas the phase difference between the X and Z dipoles is 236° , with all three ports fed with equal power level. The theoretical model in Section II-C calculated by MATLAB predicts a 98% isotropy and 21.8% CP coverage. For verification, the optimized phase difference values from the PSO have been applied to a model in Fig. 6 operating at 1.57 GHz utilized in the full-wave EM simulator, Ansys HFSS. The EM simulation results are

Fig. 7. LHCP gain of (a) theoretical model with PSO optimization and (b) HFSS simulation model. RHCP gain of (c) theoretical model with PSO optimization and (d) HFSS simulation model.

close to the predicted values from the theoretical model as can be seen from Figs. 7 and 8, except for some very minor differences. In the example, the simulated LHCP gain and RHCP radiation patterns of the HFSS model, fed with the calculated phase conditions, are compared with the theoretical model in Fig. 7. Both simulated RHCP and LHCP gains are showing good consistent with the theoretically predicted ones. Also, the simulated AR radiation pattern with the theoretically calculated result is shown in Fig. 8. Note that for angles where $0 \text{ dB} < \text{AR} < 3 \text{ dB}$, they will be accounted for CP coverage. As seen from Fig. 8 that the simulated HFSS AR radiation pattern is very similar to theoretical results, which means a good consistency of CP coverage is also obtained between the HFSS model and the theoretical model. This comparison is a good validation of the theoretical model. The 3-D package has been kept hollow from inside to accommodate the driving electronics, which matches the concept of AoP. Though not shown here, the antenna has been simulated with a metal cube (representative of the driving electronics, for example, a PCB with a typical dimension around $(\lambda/8) * (\lambda/8) * (\lambda/8)$ placed either at the corner or the center of the package, λ is the wavelength of the operation frequency) and the performance of the AoP remains almost stable (around 22% CP coverage). Having said that, if a larger metallic structure such as a battery needs to be placed inside the package, then the antenna may need to be optimized a bit, as we have shown in our previous works [3]–[5].

E. Design Guidelines

To make the design process systematic, a design guideline containing basic steps to design a near-isotropic antenna with wide CP polarization simultaneously is given as follows.

First, get the total field expressions of the dipoles groups following the steps.

- 1) Choose any corner of the package as the Cartesian coordinate system origin. The choice of the origin position does not make any difference to the final results.

Fig. 8. CP coverage for three half-wavelength dipoles. (a) Theoretical model with PSO optimization. (b) HFSS simulation model.

- 2) Locate the three half-wavelength dipoles along with the X, Y, and Z directions, respectively, on the surface of the package. Note down the center positions of the three dipoles referred to the origin as r'_j ($j = x, y, \text{ or } z$) in (6.a).
- 3) Use the dipole center positions in (6.a)–(6.c) to calculate the spatial phase errors ($kr'_j \cos(\alpha_j(\theta, \varphi))$).
- 4) Derive the total E-field expressions for the three dipoles through (9.a)–(9.e) using the spatial phase errors calculated in step 3.

Second, derive the excitation phase conditions for each dipole with the PSO technique for the user's requirements. There are four steps to accomplish this procedure as follows.

- 1) The isotropy threshold TH_{iso} , its near isotropy coverage limit I^{th-iso} , and the minimum gain G_{min} can be defined by the readers for their requirements. In our example as mentioned, we choose $TH_{iso} = 7$ dB as isotropy threshold, coverage limit $I^{th-iso} = 90\%$, and $G_{min} = -10$ dB.
- 2) At this point, an optimization method can be used. We have used the PSO technique and the algorithm flow that has been provided in the flowchart (Fig. 5). Utilize the expressions derived in step 4, along with the reader-defined isotropy threshold TH_{iso} , near isotropy coverage limit I^{th-iso} , and minimum gain G_{min} from step 5 in the first decision box in Fig. 5.
- 3) Run the PSO code to get the excitation phase conditions for three half-wavelength dipoles which lead to the maximum CP coverage under the condition defined by the readers.
- 4) Set the excitation phases of the dipoles in the HFSS model using PSO optimization results. Verify the radiation property performance calculated in the PSO with HFSS simulation results.

Third, use the phase conditions derived from the theoretical model to a practical design and the detailed steps are listed as follows.

- 1) Design a feeding network that ensures the required phases (calculated in step 7) reach the dipoles with equal magnitudes.
- 2) Final optimization can now be done for the complete AoP (dipoles with feeding network) in an EM simulator. Adjust the feeding network to optimize the complete system performance, if needed.

A design example is shown by following these guidelines, and an AoP has been designed, simulated, fabricated, and measured in Section III.

III. PRACTICAL ANTENNA DESIGN

In this section, a practical AoP design example is provided based on the design guideline provided in Section II-E to prove the feasibility of our design guidelines.

To simplify the design process and get a better match with the previous simulation, the package dimensions are set to be $L = W = H = 95$ mm, which is slightly longer than a half-wavelength dipole. Low-loss Rogers laminates with a dielectric constant $\epsilon_r = 3.3$, loss tangent $\tan\delta = 0.0015$, and thickness $t = 0.762$ mm have been used for the package fabrications. Following the first step, the origin is selected as the same as the one shown in the PSO example (Fig. 6). This can be changed to another corner and it does not affect the final results. Following step 2, the three half-wavelength dipoles, X, Y, and Z, are located along the three edges of the package, as shown in Fig. 6, and their centers are noted as (47.5 mm, 0, 0), (95, 47.5, and 95 mm), and (0, 95, 47.5 mm), respectively. Then, following step 3, the spatial error of each dipole is calculated by (6.a)–(6.c), and following step 4, the total E-field is expressed using (9.a)–(9.e) with the spatial errors calculated from step 3.

In step 5, the near isotropy requirements are set up by the user for their requirements, and here in our design example, we have set $TH_{iso} = 7$ dB and $I^{th-iso} = 90\%$ to have a better match with the example we showed in the last section. Next, following step 6, the PSO codes are updated with the parameters setup in steps 4 and 5. The readers can apply their optimization programs for the calculation, and here we choose the PSO algorithm shown in Fig. 5 for the phase conditions calculations. Following step 7, the PSO code is executed to get the excitation phase conditions for minimum cost function, which means the optimized CP coverage is obtained. Here, in our example, the phase conditions are $\psi_1 = 117^\circ$ and $\psi_2 = 236^\circ$ with an overall CP coverage of 21.8%. In step 8, the obtained phase conditions are implemented in the HFSS model. For our example, the simulation results in HFSS match the theoretical model prediction, as shown in Figs. 7 and 8.

After getting a good match between the theoretical model and HFSS simulations, the next phase is to design a practical AoP with the appropriate feeding network to feed the three orthogonal dipoles. As shown in step 8, each dipole is fed through an independent lumped port in HFSS simulation. In step 9, a feeding network which provides the phase conditions ($\psi_1 = 117^\circ$ and $\psi_2 = 236^\circ$) is required to realize the final AoP. It is difficult to realize this feed arrangement for practical antenna design as this requires either an individual excitation cable or an individual driving circuit for each dipole. Other than the practical use, for testability as well, three half-wavelength dipoles fed by a single SMA connector are less complex and more convenient than three coaxial lines with three baluns to feed the dipoles individually. Thus, for practical antenna design, it is better to utilize a single driving network. A balun is also required to transit single-ended signal from coaxial cable to differential mode to feed dipole antennas. In step 9, to feed the three orthogonal dipoles on the cubic package with the correct phases, a phase shifter with three parallel output lines has been designed, as shown in Fig. 9. Though the feed network has to be implemented on the 3-D

Fig. 9. Planar phase shifter. (a) Top view, (b) Bottom view.

Fig. 10. Three half-wavelength dipoles with feeding network. Antenna width = $W_d = 2$ mm, antenna length = $L_d = 82$ mm, feeding lines width = $W_f = 1.2$ mm, tapered section width = from 1.2 to 0.3 mm, additional phase shift length for Z-axis dipole = $L_f = 47$ mm.

package, it is initially designed on a planar surface such that when folded, all the squares shown in Fig. 9 become various faces of the cubic package. Each square in Fig. 9 has the same dimensions 95 mm as that of an individual side of the package. To feed the differential antennas, parallel lines have been designed such that one part of the line is on the top side of the substrate, whereas the other part is on the bottom side, as shown in Fig. 9. This arrangement also helps in minimizing the effect of these lines on dipoles' radiation. The parallel line is designed to be as short as possible to further reduce the effects on the radiating field. In Fig. 10, the red parallel line feeds the X-axis dipole, the green one feeds the Y-axis dipole, and the blue one feeds the Z-axis dipole. The width of the parallel lines determines their characteristic impedance and their lengths determine the output port phases. The difference in lengths between the parallel lines provides the phase differences between the output ports, which are the key factors affecting the antenna performance. The tapered part of the parallel lines helps in impedance matching with the dipoles. Once the right phases, according to the theoretical method, have been achieved, the individual substrates are folded to form the cubic package. Simulation results confirm that the folding of the feed network has a negligible effect on its planned performance. The feed network connects to a Bazooka balun at the intersection point of the three parallel lines to transform the differential signals to the single-ended mode for the coaxial cable, as shown in Fig. 10. The length of the Bazooka balun is designed to be a quarter wavelength long at 1.57 GHz, which is $l_b = 28.9$ mm.

After step 9, the three dipoles connected with the feeding network are shown in Fig. 10, where the two dipole's

Fig. 11. Simulated antenna radiation pattern in contour plot.

Fig. 12. Antenna CP coverage at target frequency.

arms are placed on either side of the substrate, consistent with the parallel line layout mentioned above. Despite the precautions are taken with the feed network, the finite width and lines of the feedline do affect the antenna radiation and thus the CP coverage of the antenna with the feeding network is not as good as has been predicted from the theoretical model and initial simulations (without the feed network). Therefore, in step 10, the dimensions of the antennas and phase shifters are optimized slightly to get good CP coverage. The re-optimized phase difference between the X and Y dipoles is 88° , whereas the phase difference between the X and Z dipoles is 276° . The final dimensions for the antenna and the feed network are given in the caption of Fig. 10.

The simulated results for the radiation pattern as well as CP coverage are shown in Figs. 11 and 12. It can be calculated from the simulation result that a 7 dB near isotropy coverage of 92% is achieved which is somewhat lower than the initial prediction of 98% but still acceptable given that the initial prediction did not include the effects of the feeding network. Furthermore, a CP coverage of 21% is achieved which is marginally below the initial prediction of 21.8%. The antenna has a gain of 3 dBi at 1.575 GHz.

IV. FABRICATION AND MEASUREMENTS

The antenna prototype has been created by fabricating six individual faces of the cube with the required metallization and then assembling them in the form of a cube, as shown in Fig. 13. Each board is fabricated individually by LPKF using the low-loss Rogers boards, as shown in Fig. 13(a). The metal traces on the boards are soldered at appropriate junction points to strengthen the metal connection. The bazooka balun is fabricated by realizing the sleeve through the copper tape on the coaxial line. The balun is then soldered to the parallel traces on the bottom plate. Finally, all the faces of the cube

Fig. 13. (a) Fabricated six roger board with metallic patterns for antenna and phase shifter. (b) Glued antenna on package. (c) Antenna on package with bazooka balun.

Fig. 14. Measured and simulated S-parameters of the total antenna system.

are attached with the help of superglue to form the cube. The complete prototype is shown in Fig. 13(b) and (c).

The input reflection coefficient of the antenna has been measured through Agilent VNA Keysight N9912A and the results are shown in Fig. 14. As can be seen from the measurement results, the antenna is matched to 50Ω from 1.34 to 1.81 GHz, which is quite similar to the simulated bandwidth. Overall, there is a good match between the simulation and measurement results. However, there is a slight shift in the measured S_{11} curve (approximately 50 MHz) around the center frequency, which could be due to the fabrication tolerances. The other difference is that it is very well-matched resonance at 1.52 GHz, which does exist but is not that prominent in the simulations. It is worth mentioning here that the narrowest parts of the parallel strip line are 0.3 mm and they have been connected to the bazooka balun through soldering which may have introduced some additional differences. Moreover, the

Fig. 15. (a) Radiation pattern measurement setup in Satimo Chamber. (b) 3-D radiation pattern.

Fig. 16. Measured CP coverage.

TABLE I
COMPARISON BETWEEN SIMULATED AND MEASURED RESULTS

	7dB isotropy	CP coverage
Simulation	94.01%	17.3%
Measurement	92.86%	17.2%

bazooka balun has been manually fabricated and it appears that the shorted stub length is not as accurate as has been simulated.

To obtain the radiation parameters, the proposed antenna is measured in Satimo anechoic chamber, as shown in Fig. 15(a). The best isotropy and CP coverage performance is observed at 1.52 GHz instead of the center frequency of 1.57 GHz. This 50 MHz shift is consistent with the impedance measurements, as described above. The measured 3-D radiation pattern at 1.52 GHz is shown in Fig. 15(b). The measured 7 dB isotropy at 1.52 GHz is 92.86%, which is quite close to the simulation results, as can be seen in Table I. The antenna's measured CP coverage is shown in Fig. 16, with a CP coverage of 17.2% at 1.52 GHz. This value is also quite close to the simulated value at 1.52 GHz; however, it is somewhat lower than the 21% CP coverage predicted by simulations at 1.57 GHz. As explained above, this may be due to the 50 MHz frequency shift observed in measurements. Nonetheless, in simulations, the performance is quite similar at 1.52 GHz, which validates the proposed theory and design concept.

It is well known that antennas' radiation performance gets affected by nearby objects and the proposed antenna is not exempted from that. However, if the antenna system is smart enough to sense the environment, then the phases of the signals fed to the individual elements can be adjusted to reconfigure the radiation pattern accordingly.

V. CONCLUSION

A combination of three orthogonal dipoles' model and PSO technique has been utilized in this work to design 3-D AoP with near-isotropic radiation pattern and wide circular polarization coverage. Appropriate phase conditions to achieve maximum CP coverage and decent isotropy from the theoretical model are then utilized to demonstrate a practical AoP. A design guideline is provided for readers to design AoP which can achieve maximum CP and isotropy coverage simultaneously. A design example has been presented based on the design guidelines. The prototype shows a decent performance and matches well with the predicted performance from the theoretical analysis. The methodology presented in this work can be utilized to design and optimize antennas on 3-D packages for future IoT applications. Also, the radiating structure can be replaced by other suitable antenna types and thus more advanced functions, such as different orbital angular momentum beams, can be realized [35].

REFERENCES

- [1] C. M. Kruesi, R. J. Vyas, and M. M. Tentzeris, "Design and development of a novel 3-D cubic antenna for wireless sensor networks (WSNs) and RFID applications," *IEEE Trans. Antennas Propag.*, vol. 57, no. 10, pp. 3293–3299, Oct. 2009.
- [2] I. T. Nassar and T. M. Weller, "Development of novel 3-D cube antennas for compact wireless sensor nodes," *IEEE Trans. Antennas Propag.*, vol. 60, no. 2, pp. 1059–1065, Feb. 2012.
- [3] M. F. Farooqui, C. Claudel, and A. Shamim, "An inkjet-printed buoyant 3-D lagrangian sensor for real-time flood monitoring," *IEEE Trans. Antennas Propag.*, vol. 62, no. 6, pp. 3354–3359, Jun. 2014.
- [4] M. F. Farooqui, M. A. Karimi, K. N. Salama, and A. Shamim, "3D-printed disposable wireless sensors with integrated microelectronics for large area environmental monitoring," *Adv. Mater. Technol.*, vol. 2, no. 8, 2017, Art. no. 1700051.
- [5] Z. Su, K. Klionovski, R. M. Bilal, and A. Shamim, "A dual band additively manufactured 3-D antenna on package with near-isotropic radiation pattern," *IEEE Trans. Antennas Propag.*, vol. 66, no. 7, pp. 3295–3305, Jul. 2018.
- [6] C. Cho, H. Choo, and I. Park, "Printed symmetric inverted-F antenna with a quasi-isotropic radiation pattern," *Microw. Opt. Technol. Lett.*, vol. 50, no. 4, pp. 927–930, 2008.
- [7] H.-K. Ryu, G. Jung, D.-K. Ju, S. Lim, and J.-M. Woo, "An electrically small spherical UHF RFID tag antenna with quasi-isotropic patterns for wireless sensor networks," *IEEE Antennas Wireless Propag. Lett.*, vol. 9, pp. 60–62, 2010.
- [8] G. Pan, Y. Li, Z. Zhang, and Z. Feng, "Isotropic radiation from a compact planar antenna using two crossed dipoles," *IEEE Antennas Wireless Propag. Lett.*, vol. 11, pp. 1338–1341, 2012.
- [9] C. Deng, Y. Li, Z. Zhang, and Z. Feng, "A wideband isotropic radiated planar antenna using sequential rotated L-Shaped monopoles," *IEEE Trans. Antennas Propag.*, vol. 62, no. 3, pp. 1461–1464, Mar. 2014.
- [10] J.-H. Kim and S. Nam, "A compact quasi-isotropic antenna based on folded split-ring resonators," *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 294–297, 2017.
- [11] C. Deng, Y. Li, Z. Zhang, and Z. Feng, "Design of a three-dimensional folded slot antenna with quasi-isotropic radiation pattern," in *Proc. IEEE Int. Symp. Antennas Propag. USNC/URSI Nat. Radio Sci. Meeting*, Jul. 2015, pp. 588–589.
- [12] S. Shamsinejad, F. De Flaviis, and P. Mousavi, "Microstrip-fed 3-D folded slot antenna on cubic structure," *IEEE Antennas Wireless Propag. Lett.*, vol. 15, pp. 1081–1084, 2016.
- [13] S. Long, "A combination of linear and slot antennas for quasi-isotropic coverage," *IEEE Trans. Antennas Propag.*, vol. 23, no. 4, pp. 572–576, Jul. 1975.
- [14] C. Cho, H. Choo, and I. Park, "Broadband RFID tag antenna with quasi-isotropic radiation pattern," *Electron. Lett.*, vol. 41, no. 20, pp. 1091–1092, Sep. 2005.
- [15] L. Liang and S. V. Hum, "A low-profile antenna with quasi-isotropic pattern for UHF RFID applications," *IEEE Antennas Wireless Propag. Lett.*, vol. 12, pp. 210–213, 2013.
- [16] C.-H. Wu and T.-G. Ma, "Miniaturized self-oscillating active integrated antenna with quasi-isotropic radiation," *IEEE Trans. Antennas Propag.*, vol. 62, no. 2, pp. 933–936, Feb. 2014.
- [17] B. Wang, S. W. Cheung, W. Wang, M. Li, and T. I. Yuk, "A broadband and wide beamwidth dual circularly polarized antenna using crossed bent dipoles," in *Proc. 11th Eur. Conf. Antennas Propag. (EUCAP)*, Mar. 2017, pp. 1936–1939.
- [18] X.-Y. He, L. Chang, and L.-L. Chen, "A multifunction broad-beam antenna with dual bands and dual circular-polarizations," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Mar. 2016, pp. 1–4.
- [19] Y. Jie, L. Chun-lan, H. Yong-jin, J. Jian-xin, and J. Wei-wei, "Design and analysis of a broad band wide beam circular polarization microstrip antenna," in *Proc. Int. Conf. Microw. Millim. Wave Technol. (ICMMT)*, May 2012, pp. 1–4.
- [20] D. Fan, Z. Wang, B. Huang, and W. Zhang, "Design of broad beam circular-polarized microstrip antenna," in *Proc. Int. Symp. Antennas Propag.*, vol. 1, Nanjing, Oct. 2013, pp. 439–441.
- [21] H. Nguyen-Bich, H. Chu-Ba, and C. Dao-Ngoc, "All GPS bands dielectric resonator antenna with circular polarization for GPS application," in *Proc. Int. Conf. Adv. Technol. Commun.*, Oct. 2012, pp. 65–68.
- [22] K. Saurav, D. Sarkar, A. Singh, and K. V. Srivastava, "Multiband circularly polarized cavity-backed crossed dipole antenna," *IEEE Trans. Antennas Propag.*, vol. 63, no. 10, pp. 4286–4296, Oct. 2015.
- [23] J. Cardoso and A. Safaai-Jazi, "A spherical helical antenna," in *Proc. IEEE Antennas Propag. Soc. Int. Symp.*, Jun./Jul. 1993, pp. 1558–1561.
- [24] A. Gu, S. Yang, and Z. Nie, "Analysis and design of miniaturized ultra-wideband conical log spiral antennas," in *Proc. Cross Strait Quad-Regional Radio Sci. Wireless Technol. Conf.*, Jul. 2013, pp. 191–194.
- [25] G. H. Brown and O. M. Woodward, "Circularly-polarized omnidirectional antenna," *RCA Rev.*, vol. 8, no. 2, pp. 259–269, Jun. 1947.
- [26] V. Galindo and K. Green, "A near-isotropic circularly polarized antenna for space vehicles," *IEEE Trans. Antennas Propag.*, vol. 13, no. 6, pp. 872–877, Nov. 1965.
- [27] M. Huchard, C. Delaveaud, and S. Tedjini, "Miniature antenna for circularly polarized quasi isotropic coverage," in *Proc. 2nd Eur. Conf. Antennas Propag. (EuCAP)*, Nov. 2007, pp. 1–5.
- [28] H. H. Tran, S. X. Ta, and I. Park, "A compact circularly polarized crossed-dipole antenna for an RFID tag," *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 674–677, 2015.
- [29] S. Choi and K. Sarabandi, "A W-Shaped antenna with spatial polarization variation for direction finding," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 12, pp. 2429–2433, Dec. 2018.
- [30] D. Bugnolo, "A quasi-'Isotropic' antenna in the microwave spectrum," *IRE Trans. Antennas Propag.*, vol. 10, no. 4, pp. 377–383, Jul. 1962.
- [31] K. Klionovski, M. B. Arboleda, Z. Su, and A. Shamim, "On synthesis of orientation insensitive antennas," in *Proc. IEEE Indian Conf. Antennas Propagation (InCAP)*, Dec. 2019, pp. 1–4.
- [32] C. Balanis, *Antenna Theory: Analysis and Design*, 4th ed. Hoboken, NJ, USA: Wiley, 2016.
- [33] Z. Su, K. Klionovski, H. Liao, Y. Chen, A. Elsherbeni, and A. Shamim, "3D antenna in package design: Maximizing radiation pattern isotropy and CP coverage," in *Proc. IEEE Int. Symp. Antennas Propag. USNC/URSI Nat. Radio Sci. Meeting*, to be published.
- [34] D. Wang, D. Tan, and L. Liu, "Particle swarm optimization algorithm: An overview," *Soft Comput.*, vol. 22, pp. 387–408, Jan. 2018, doi: 10.1007/s00500-016-2474-6.
- [35] A. F. Morabito, L. Di Donato, and T. Isernia, "Orbital angular momentum antennas: Understanding actual possibilities through the aperture antennas theory," *IEEE Antennas Propag. Mag.*, vol. 60, no. 2, pp. 59–67, Apr. 2018.

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