

Design and Parametric Sensitivity Analysis of a Six-Disc Short-Backfire Disc-on-Rod Antenna for 2.4 GHz Applications

Ashikur Mollik, Md Emadur Rahman Ekra*,
Md Nadim Hosain, Md Daud Ibrahim

School of Electronics and Information Engineering,
Nanjing University of Information Science and Technology, Nanjing, China

E-mail: ashikurmollik@nuist.edu.cn; mdemandekra@nuist.edu.cn;
mdnadimhosain@nuist.edu.cn; ibrahimmddaud@nuist.edu.cn

*Corresponding author details: Md Emadur Rahman Ekra; mdemandekra@nuist.edu.cn

ABSTRACT

This study designs and analyzes a copper disc-on-rod endfire antenna for the 2.4 GHz industrial, scientific, and medical (ISM) band. Six coaxial copper discs of graded radius are mounted on a continuous copper rod of 2 mm radius, and a cylindrical copper cavity at the feed end serves as a surface-wave launcher, impedance-matching element, and rear shield. Although the structure resembles a Yagi-Uda array, this paper shows that it belongs to the disc-on-rod class of surface-wave radiators and analyzes it with slow-wave and travelling-wave theory. This paper models the complete geometry using the finite element method in Ansys HFSS and sweeps twelve-dimensional parameters individually. The final design resonates at 2.400 GHz with a simulated reflection coefficient of -39.08 dB and a -10 dB band extending from approximately 2.31 to 2.58 GHz. The far field shows a single forward lobe with rear radiation suppressed by more than 20 dB. Converting the simulated far-field data to absolute units gives a peak realized gain of 10.67 dBi, which lies within the envelope of analytical endfire estimates for the same electrical length and corresponds to 55% of the Harrington normal-gain bound. The normalized disc size and pitch of the uniform guiding section fall within the design window established empirically in the classical disc-on-rod literature. The sensitivity analysis shows that the launcher region governs both the resonant frequency and the quality of the match and requires sub-millimetre accuracy, whereas the guiding section tolerates errors of about 1 mm and the conductor thickness has no electromagnetic effect. All results are simulation-based; this study states the limitations and outlines a validation protocol.

Keywords: disc-on-rod antenna; surface-wave antenna; slow-wave structure; 2.4 GHz ISM band; finite element method

INTRODUCTION

The 2.4 GHz industrial, scientific and medical (ISM) band carries wireless local area network, Bluetooth and low-rate telemetry traffic and remains the most heavily used unlicensed band for short-range links. Two requirements follow for an antenna intended for it. The impedance match must hold across the band rather than at a single frequency, because the designer does not control channel assignment, and angular selectivity is as valuable as gain, because a narrow beam rejects interferers that an omnidirectional radiator accepts. Consumer equipment in this band typically uses omnidirectional sleeve dipoles of 2 to 5 dBi, which suit mobile terminals but not fixed links between two known locations. For fixed links, the established directional solutions are the Yagi-Uda array, printed quasi-Yagi structures, the biquad, and the parabolic reflector [1], [2].

A less common alternative is the metallic disc-on-rod antenna, in which conducting discs mounted on

a common conducting rod guide a bound surface wave and radiate it from the truncated end. Ehrenspeck and Poehler showed that an endfire array longer than about one wavelength behaves as a surface-wave structure whose gain is governed by the phase velocity of the guided wave, which depends in turn on the height, diameter and spacing of the directors [3]. Terminating such a structure with a reflector gives the backfire antenna [4], [5]. Short backfire antennas of one wavelength or less reach 12 to 16 dBi, and long backfire antennas reach 17 to 25 dBi [6]. Kirov, Hristov and Feick documented an empirical design procedure for long backfire disc-on-rod antennas, validated on prototypes two to four wavelengths long, and reported a four-wavelength design for the 2.4 to 2.5 GHz band with a measured gain of approximately 24 dBi [6]; a companion study examined the reflector profile and the feed placement [7]. Patent literature supplies complementary design data. For structures 3 to 12 wavelengths long, near-optimum gain is obtained

when $(D - d)$ lies between $0.23\lambda_0$ and $0.27\lambda_0$ and the disc pitch lies between $0.18\lambda_0$ and $0.23\lambda_0$, where D is the disc diameter, d the rod diameter and λ_0 the free-space wavelength [8]. These works address electrically long structures with two reflectors; the compact single-reflector regime lies outside their scope. The structure suits low-cost construction because it is formed entirely from copper sheet and rod and has no dielectric substrate, and hence no dielectric loss.

Despite these advantages, the disc-on-rod geometry is sparsely represented in recent work for this band. Most published 2.4 GHz antenna studies concern microstrip patches on dielectric substrates [9] or printed Yagi variants [10], [11], which are compact but incur substrate loss and constrain the element separation along the direction of propagation. The closest analogue to the present work is the study of Alcala Garrido et al. [12], who simulated a nine-disc structure described as a disc-Yagi alongside a biquad and a printed Yagi and reported a return loss of -44.6 dB at 2.4068 GHz for the disc structure. That study reported no gain figure, included no sensitivity analysis, and did not interpret the structure in surface-wave terms. A second gap is methodological. Low-cost antenna studies in this band commonly report a single optimized dimension set without showing how sensitive it is to fabrication error, although a geometry that meets specification only within a fraction of a millimetre is of limited use to builders who rely on economical construction.

In this study, we design a six-disc copper structure with a graded radius profile and a cavity-backed launcher for the 2.4 GHz band and analyze it by the finite element method. We make four contributions. First, we classify the structure as a short-backfire disc-on-rod surface-wave radiator rather than a Yagi-Uda array and interpret its operation through the phase velocity of the guided wave. Second, we sweep twelve-dimensional parameters individually to identify which dimensions govern the response, and we treat this sensitivity analysis as a principal result. Third, we convert the far-field data, which the solver reports in field units, to absolute gain and compare the result with analytical estimates and a fundamental bound. Fourth, we show that the normalized proportions of the optimized guiding section fall within the design window established empirically for this antenna class [8].

THEORETICAL BACKGROUND

Slow-Wave Guidance on a Disc-Loaded Rod

A thin conducting rod in free space supports no bound propagating mode. Periodic reactive loading changes this, in the sense established by Goubau for modified conductor surfaces [13]. The transverse discs add shunt capacitance, the intervening rod sections add series inductance, and the composite behaves as a periodic slow-wave transmission structure [14]. We characterize the guided wave by the phase-velocity ratio.

$$p = \frac{\beta}{k_0} = \frac{c}{v_p} \quad (1)$$

Where β is the guided phase constant, k_0 the free-space wavenumber, c the speed of light, and v_p the phase velocity. A bound surface wave requires $p > 1$. Such a wave cannot radiate from a uniform structure, so radiation occurs only at the two discontinuities, the launcher and the truncation, and the antenna behaves as an aperture radiator rather than as an array of independently excited elements. The field of the bound mode decays away from the axis over a transverse decay length.

$$\rho_e = \frac{\lambda_0}{2\pi\sqrt{p^2-1}} \quad (2)$$

Equation (2) links electrical behavior to physical size, because a disc loads the wave effectively only when its radius is comparable with ρ_e . A larger disc diameter increases the shunt capacitance per period, and therefore p . A larger pitch increases the series inductance per period but reduces the number of periods per wavelength, so the two effects partly cancel, and the response is less sensitive to pitch than to diameter. Because the structure is air-filled, both quantities are purely geometric and the design scales with frequency.

Because the structure is periodic with pitch S , its field can be expanded in Floquet space harmonics with phase constants $\beta_n = \beta_0 + 2\pi n/S$. The antenna radiates a fixed endfire beam, rather than a frequency-scanned leaky-wave beam, only if the first backward harmonic remains slow, which requires

$$S(p+1) < \lambda_0 \quad (3)$$

The first stopband (Bragg condition) opens at $\beta S = \pi$.

Endfire Directivity and Fundamental Bounds

For a travelling-wave distribution of length L , an ordinary endfire radiator ($p = 1$) has a directivity of approximately [1].

$$D \approx \frac{4L}{\lambda_0} \quad (4)$$

Hansen and Woodyard showed that the directivity of a long uniform endfire array is maximized when the total phase delay along the structure exceeds the free-space value by π radians [15], which corresponds to

$$p_{\text{opt}} = 1 + \frac{\lambda_0}{2L} \quad (5)$$

and raises the directivity to approximately

$$D \approx \frac{7.28L}{\lambda_0} \quad (6)$$

The additional directivity comes at the cost of higher stored reactive energy, a narrower impedance bandwidth and greater sensitivity to dimensional error.

For an antenna contained within a sphere of radius a , Harrington gave the maximum directivity attainable without super gain as [16].

$$D_{\max} = (ka)^2 + 2ka \quad (7)$$

Where $k = 2\pi/\lambda_0$. The effective aperture corresponding to a gain G is

$$A_{\text{eff}} = \frac{G\lambda_0^2}{4\pi} \quad (8)$$

Gain from the Solver Far-Field Output

Ansyz HFSS reports the far field by default as $r|E|$, the product of range and electric-field magnitude, normalized to 1 W of accepted power and expressed in decibels relative to 1 V. This quantity is not gain. In the far zone, the radiation intensity is $U = |rE|^2/(2\eta_0)$, where $\eta_0 = 376.73 \Omega$ is the free-space wave impedance. Dividing by the intensity $1/(4\pi)$ W/sr of an isotropic radiator that accepts the same power gives the gain in the IEEE sense [17],

$$G = \frac{4\pi|rE|^2}{2\eta_0} = \frac{|rE|^2}{59.96} \quad (9)$$

so that, for the peak-amplitude field convention,

$$G \text{ (dBi)} = \text{dB}(rE_{\text{total}}) - 17.78 \quad (10)$$

If the solver reports root-mean-square fields, the constant becomes 29.98 and the offset 14.77 dB, so the convention in use must be confirmed before (10) is applied.

ANTENNA DESIGN

Geometry

Figure 1 defines the antenna and the coordinate system. The structure axis is z , the endfire beam points along $+z$, away from the cavity, and the polar angle θ is measured from that axis. Figure 2 gives the dimensions. The antenna comprises a cylindrical copper cavity of 45 mm radius and 15 mm depth, a straight copper rod of 2 mm radius and 160 mm length that passes through the center of the cavity, and six circular copper discs of 0.25 mm thickness mounted normal to the rod. The disc radius decreases from 34.5 mm at the first disc to 18 mm at the last three. A coaxial connector on the rear face of the cavity excites the base of the rod. Table 1 lists the dimensions in millimetres and in free-space wavelengths at 2.400 GHz ($\lambda_0 = 124.91$ mm), to which all electrical lengths in this paper refer.

TABLE 1: Dimensions of the Proposed Antenna in mm and in Free Space. Wavelengths at 2.400 GHz ($\lambda_0 = 124.91$ mm)

Element	r (mm)	$2r/\lambda_0$	z (mm)	Δz (mm)	$\Delta z/\lambda_0$
Cavity	45.0	0.721	0.0	–	–
Disc 1	34.5	0.552	4.2	4.2	0.034
Disc 2	26.5	0.424	13.2	9.0	0.072
Disc 3	19.5	0.312	45.2	32.0	0.256
Disc 4	18.0	0.288	73.9	28.7	0.230
Disc 5	18.0	0.288	102.6	28.7	0.230
Disc 6	18.0	0.288	131.3	28.7	0.230
Rod	2.0	0.032	–	–	–

r : radius; z : axial position; Δz : spacing to the preceding element. Cavity depth 15.0 mm ($0.120\lambda_0$); disc thickness 0.25 mm; rod length 160.0 mm ($1.281\lambda_0$); structure length L , cavity to disc 6, 131.3 mm ($1.051\lambda_0$).

We selected copper for all conducting parts. The model uses a conductivity $\sigma = 5.80 \times 10^7$ S/m and a relative permeability $\mu_r = 1.0$, for which the skin depth at 2.4 GHz is $\delta = 1.35 \mu\text{m}$ and the surface resistance is $R_s = 12.8 \text{ m}\Omega/\text{sq}$ [1]. The 0.25 mm sheet

is therefore about 185 skin depths thick, so its thickness is set by mechanical stiffness and stock availability rather than by electromagnetic considerations. No dielectric is present, so ohmic dissipation in the copper is the only loss mechanism.

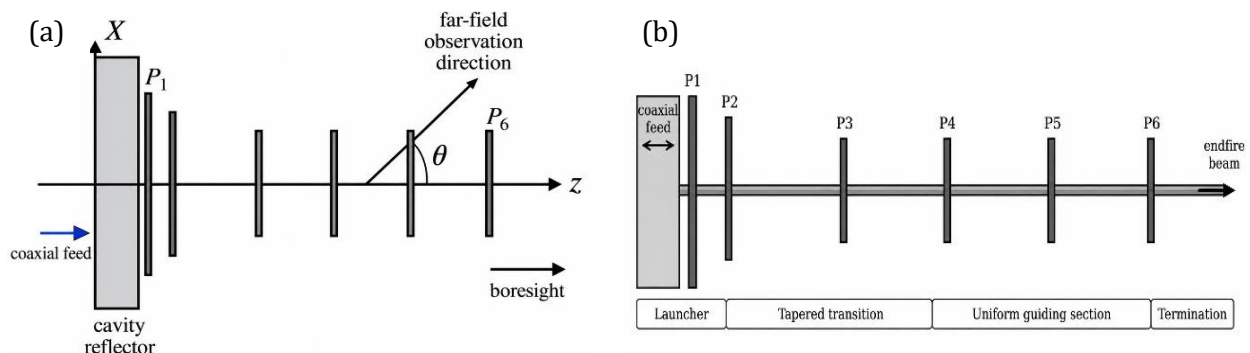


FIGURE 1: Geometry of the Proposed Antenna: (a) Coordinate System; (b) Functional Regions. Discs 1 To 6 Are Labelled P1 to P6.

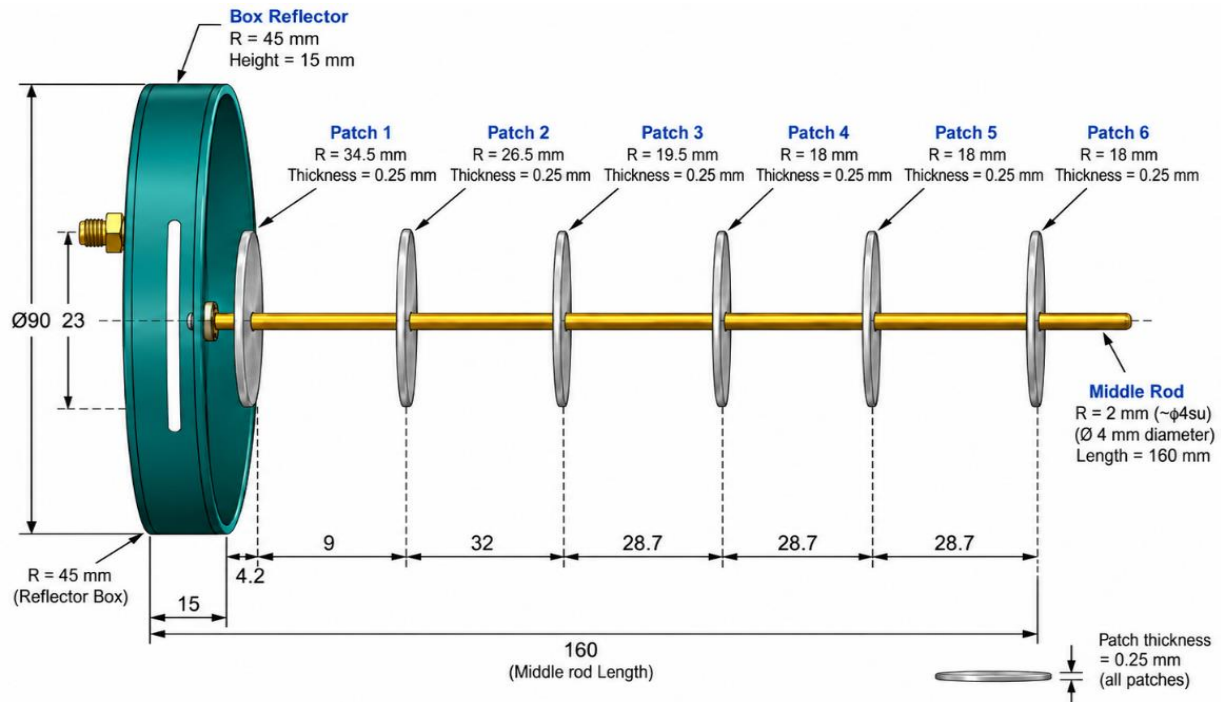


FIGURE 2: Dimensioned Schematic of the Six-Disc Cavity-Backed Disc-on-Rod Antenna (Dimensions in mm). The Drawing Labels the Discs “PATCH 1” to “PATCH 6” and the Cavity “Box Reflector”.

Functional Regions

The structure is divided into four functional regions (Figure 1b). The cavity and disc 1 form the launcher, which converts the coaxial mode into a bound surface wave and sets the input impedance. Disc 1 has a diameter of $0.552\lambda_0$ and stands only $0.034\lambda_0$ from the cavity. Discs 2 and 3, of progressively smaller radius, form a tapered transition. Discs 4 to 6, of equal diameter ($0.288\lambda_0$) and equal pitch ($0.230\lambda_0$), form the uniform guiding section in which the phase delay required for endfire operation accumulates. The 28.7 mm of rod beyond disc 6 forms the termination from which the guided wave radiates. The length from the cavity to the last disc, $L = 131.3 \text{ mm} = 1.051\lambda_0$, is the electrical aperture length that enters (4) to (6).

Taking the disc radius as an estimate of ρ_e in (2) gives $p \approx 1.15$ for disc 1 (34.5 mm) and $p \approx 1.49$ for the guiding-section discs (18 mm), so the taper slows the guided wave progressively between the launcher and the aperture (Figure 3). For $L = 1.051\lambda_0$, (5) gives a Hansen-Woodyard optimum of $p_{\text{opt}} = 1.476$, which the uniform guiding section approaches closely. Applying (3) with $p = 1.476$ limits the pitch to $0.404\lambda_0$, or 50.5 mm. The chosen pitch of 28.7 mm is smaller by a factor of 1.76, so no space becomes fast, and the antenna radiates a fixed endfire beam across the band. The Bragg edge lies at $p = \lambda_0/(2S) = 2.177$, well above the operating value. Figure 4 places the light line, the Bragg edge, and the operating point on a dispersion diagram.

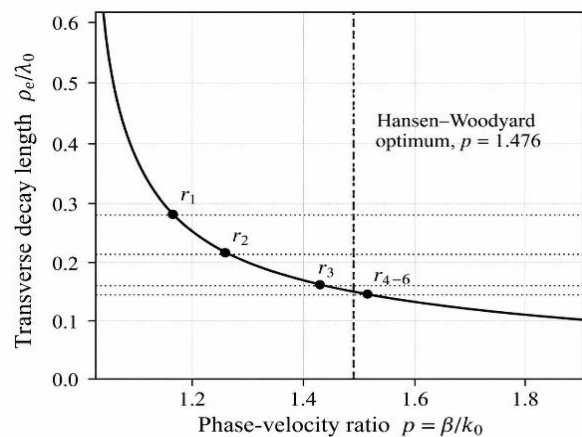


FIGURE 3: Transverse Decay Length of the Bound Surface Wave Versus Phase-Velocity Ratio, With the Four-Disc Radii Marked.

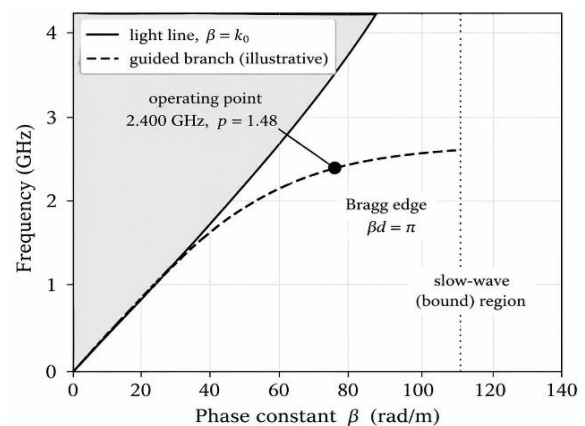


FIGURE 4: Dispersion Diagram of the Disc-Loaded Rod Showing the Light Line, an Illustrative Guided Branch and the Operating Point at 2.400 GHz. The Drawing Denotes the Pitch By d .

Relation to the Classical Design Window

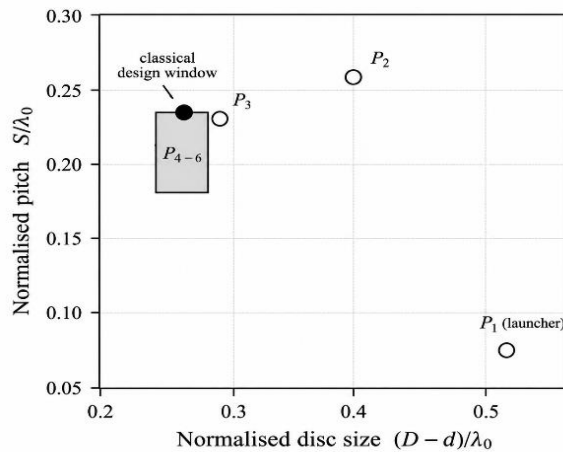


FIGURE 5: Normalized Disc Size and Pitch of the Four-Disc Groups Relative to the Classical Design Window [8].

We obtained the geometry by numerical parametric search, not from a published design procedure, so its relation to the classical disc-on-rod data is an independent check. For the uniform guiding section, $D = 36$ mm and $d = 4$ mm give $(D - d) = 32$ mm, or $0.256\lambda_0$, and the pitch is $0.230\lambda_0$. Both values fall within the window of [8], with the pitch at its upper edge (Figure 5). The corresponding values of $(D - d)/\lambda_0$ and S/λ_0 are 0.520 and 0.072 for disc 1, 0.392 and 0.256 for disc 2, and 0.280 and 0.230 for disc 3. The launcher and transition discs therefore lie outside the window, with disc 3 meeting the pitch condition only, which is expected because they perform a different function. The window was established for structures 3 to 12 wavelengths long, whereas the present structure is about one wavelength long. We regard this agreement, reached by an independent optimization, as the strongest single piece of evidence for the disc-on-rod classification.

NUMERICAL METHOD

Model and Solver Settings

We performed all simulations in Ansys HFSS, a three-dimensional frequency-domain solver based on the finite element method [18]. Geometry motivated this choice. It combines a 0.25 mm sheet thickness with a 160 mm axial extent, a 4.2 mm launcher gap, and a coaxial feed transition; a ratio of largest to smallest feature above 600; and a conformal tetrahedral mesh handles this range without the staircase error that a Cartesian grid introduces at the disc edges and the curved cavity wall.

We modelled the cavity, the rod, and the six discs as copper solids surrounded by air (Figure 6). A single wave port of $50\ \Omega$ reference impedance excites a feed pin of 0.5 mm radius that joins the base of the rod. We placed the port reference plane close to the point at which the feed enters the copper structure, so that the computed reflection coefficient represents the antenna input directly. An air box with radiation boundaries on its outer faces, placed a fraction of a wavelength from the structure, encloses the antenna. We used a driven-mode solution with adaptive mesh refinement at 2.4 GHz, which refines the mesh preferentially around the thin discs, the inter-disc gaps, and the 4.2 mm gap between the cavity and disc 1, and we stopped the refinement when the change in S_{11} between successive adaptive passes became negligible. The frequency sweep covered 2.0 to 3.0 GHz, and the far-field setup was defined at 2.4 GHz. We kept identical settings for the final design and for every parametric sweep so that all results are directly comparable.

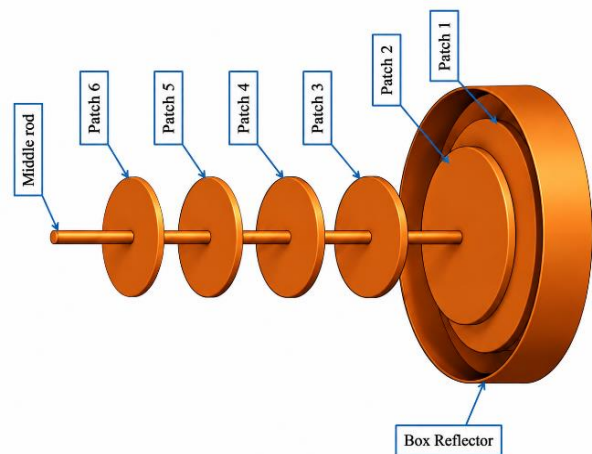


FIGURE 6: HFSS Model of the Six-Disc Antenna with the Cavity Reflector.

Parametric Sweeps

We varied twelve-dimensional parameters one at a time about the final geometry, holding all other dimensions fixed (Table 2). We re-solved each perturbed geometry and reduced its reflection coefficient to two figures of merit: the shift of the resonant frequency from 2.400 GHz, and the change in depth and width of the -10 dB matched band. The 0.5 mm step used for the linear dimensions equals $0.004\lambda_0$ and represents the accuracy attainable in workshop construction.

TABLE 2: Parametric Sweep Ranges.

Parameter	Symbol	Range (mm)	Step (mm)
Disc 5 to disc 6 spacing	d_{56}	27.0–30.0	0.5
Disc 4 to disc 5 spacing	d_{45}	27.0–30.0	0.5
Disc 3 to disc 4 spacing	d_{34}	27.0–30.0	0.5
Disc 2 to disc 3 spacing	d_{23}	30.5–33.5	0.5
Disc 1 to disc 2 spacing	d_{12}	13.5–16.5	0.5
Cavity to disc 1 spacing	d_{r1}	4.5–7.5	0.5

Parameter	Symbol	Range (mm)	Step (mm)
Cavity size	b_r	43.5–46.5	0.5
Disc 1 radius	r_1	33.0–36.0	0.5
Disc 2 radius	r_2	25.5–28.0	0.5
Disc 3 radius	r_3	18.0–21.0	0.5
Discs 4–6 radius	$r_4 = r_5 = r_6$	16.5–19.5	0.5
Disc thickness	t	0.22–0.28	0.01

RESULTS

Reflection Coefficient and Bandwidth

Figure 7 shows the simulated reflection coefficient of the final geometry from 2.0 to 3.0 GHz. A single deep resonance occurs at 2.400 GHz, where $S_{11} = -39.08$ dB. This corresponds to a reflection coefficient magnitude $|r| = 0.0111$, a reflected power of 0.0124%, a voltage standing wave ratio (VSWR) of 1.022, and a mismatch loss of 5.4×10^{-4} dB. The depth of the minimum is not in itself a measure of performance: improving the match from -20 dB to -39 dB increases the delivered power by only 0.043 dB. The bandwidth matters more. The response crosses -10 dB at approximately 2.31 GHz and remains below that level to approximately 2.58 GHz, with a secondary shoulder near 2.6 GHz, so the matched band covers the 2.4 GHz band with margin at both edges.

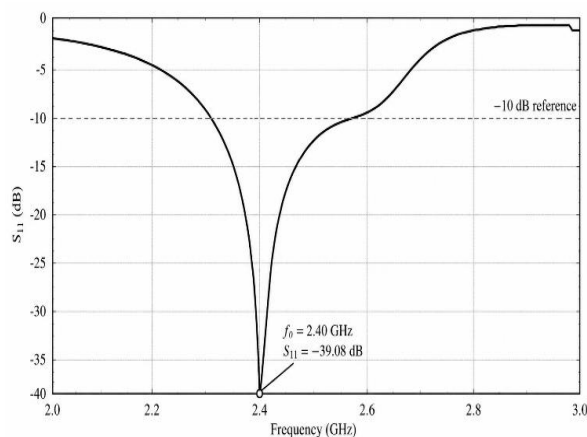


FIGURE 7: Simulated Reflection Coefficient (S_{11}) of the Proposed Antenna.

Three factors contribute to this bandwidth. First, the antenna is electrically large, with $ka = 3.73$ for the enclosing sphere defined below, so the Chu bound on the radiation quality factor [19] is far from binding. Second, a disc is a wide loading element, and the quality factor of a resonant element falls as its transverse dimension grows relative to its length. Third, the graded radius profile gives the regions of the structure different reactance slopes with frequency, which broadens the composite response in the manner of a tapered transmission-line transformer. A response that is both wide and deep can also arise from dissipation. The only loss mechanism here is ohmic, and a surface resistance of $12.8 \text{ m}\Omega/\text{sq}$ cannot account for a broadened response. Confirming this directly requires radiation efficiency, which we did not extract from the model.

Radiation Pattern

Figures 8 and 9 show the far field at 2.400 GHz as an elevation cut and as a three-dimensional surface, both in the field units of the solver. The pattern consists of a single dominant lobe along $+z$, a shallow collar of sidelobe structure, and a small residual lobe in the rear hemisphere. The sidelobe structure appears as a broad shoulder between approximately 50° and 80° from boresight rather than as a sequence of discrete lobes, which is characteristic of a continuous, tapered aperture illumination and not of an array of discretely excited elements.

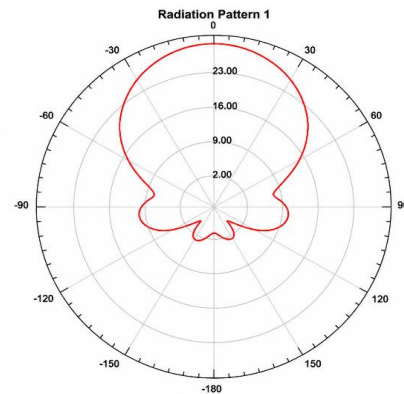


FIGURE 8: Simulated Far-Field Elevation Pattern at 2.400 GHz in the $\varphi = 0^\circ$ Plane, in $r|E|$ Units (dB).

The rear response in Figure 8 remains close to the plot origin, which indicates suppression of rear radiation by more than 20 dB relative to the forward peak. Two mechanisms contribute. The cavity intercepts the rearward component of the field launched at the feed, and the guided wave carries energy forward so that the truncation radiates mainly into the forward hemisphere. The small residual lobe on the axis behind the cavity is consistent with diffraction around the cavity rim, which has a diameter of $0.721\lambda_0$ and carries no quarter-wave choke.

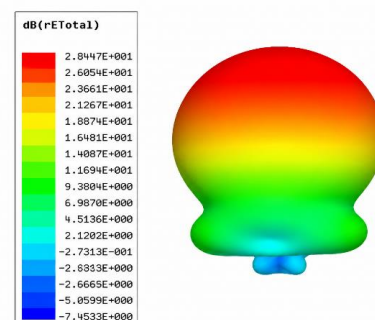


FIGURE 9: Simulated Three-Dimensional Far Field at 2.400 GHz in $r|E|$ Units. The Peak of 28.447 dB corresponds to a Realised Gain of 10.67 dBi by (10).

Peak Gain and Comparison with Theory

The color scale of Figure 9 peaks at 28.447 dB in field units. Applying (10) gives a peak realized gain of $G = 28.447 - 17.78 = 10.67$ dBi, which is the principal far-field result of this study. Table 3 compares this value with estimates that depend only on the structure length $L = 1.051\lambda_0$. Ordinary endfire theory (4) gives 6.2 dBi, and the Hansen-Woodyard result (6) gives 8.8 dBi. The empirical relation between boom length and gain for one-wavelength endfire structures gives approximately 11 dBi [2], [20], and a short backfire reflector adds 1 to 3 dB to the unterminated endfire structure [4], [5]. The simulated value lies within this envelope and close to its upper end, which is the expected outcome for a structure that approaches the Hansen-Woodyard phase condition and carries a rear reflector.

TABLE 3: Independent Estimates of Peak Endfire Gain for $L = 1.051\lambda_0$.

Basis	Relation	Value (dBi)
Ordinary endfire	(4)	6.2
Hansen-Woodyard optimum	(6)	8.8
Empirical one-wavelength endfire structure	[2], [20]	≈ 11.0
Harrington normal gain bound, $ka = 3.73$	(7)	13.3
This work, from the simulated far field	(10)	10.67

The bound (7) provides a stricter test. The enclosing sphere has a radius $a = 74.2$ mm, obtained from the half-length of the structure and the radius of the largest disc, which gives $ka = 3.73$ and a maximum normal-gain directivity of 21.4, or 13.3 dBi. The simulated gain lies 2.6 dB below this bound, at 55% of the limiting directivity, so the design uses its electrical volume efficiently without entering the super gain regime, in which bandwidth and dimensional tolerance would both deteriorate. From (8), the effective aperture is 0.0146 m^2 , equivalent to a circular aperture of 68.1 mm radius. This is approximately twice the 34.5 mm radius of the largest disc, which is normal for surface-wave endfire radiators: the field at the truncation extends to the decay length of (2), so the antenna presents a virtual aperture wider than any disc.

Parametric Sensitivity

Table 4 summarizes the outcome of the twelve sweeps, ranked by influence on the resonant frequency and on the quality of the match, and Figure 10 presents the same ranking as an influence map.

• Guiding Section

The three spacings between discs 3 and 6 and the common radius of discs 4 to 6 exhibit essentially similar behavior in terms of their influence on the overall response. Varying any of these parameters across its respective sweep range shifts the resonant frequency only slightly and primarily affects the depth and sharpness of the minimum; in every case, the response consistently remains below -10 dB across the entire 2.4 GHz band. The guiding section therefore tolerates dimensional errors of approximately 1 mm, or $0.008\lambda_0$. The influence increases mildly towards the feed: the disc 3 to disc 4 spacing produces slightly greater spread than the disc 5 to disc 6 spacing. Disc 2 to disc 3 spacing marks the transition to the tighter behavior of the launcher region.

• Launcher Region

The cavity-to-disc 1 spacing, the disc 1 to disc 2 spacing, and the radius of disc 1 each change both the resonant frequency and the quality of the match markedly. Reducing the cavity-to-disc 1 gap lowers the resonant frequency and improves the match up to a point beyond which the trend reverses; increasing the gap has the opposite effect. Increasing the radius of disc 1 lowers the resonant frequency, and the 33 to 36 mm sweep moves the resonance across a substantial part of the examined band. We selected 34.5 mm to place the minimum at 2.400 GHz. Part of this sensitivity follows from the small nominal values. With launcher spacings of 9 mm and 4.2 mm, a 1 mm error is a relative change of 11% and 24%, respectively, whereas the same error in the 28.7 mm guiding-section pitch is only 3.5%.

• Transition Discs, Cavity Size and Conductor Thickness

The radii of discs 2 and 3 have a weaker but still significant influence, mainly on the depth and width of the minimum rather than on its center frequency, which is consistent with their role as a taper between the launcher and the guiding section. Varying the cavity size between 43.5 and 46.5 mm changes the reflection coefficient only modestly and does not disturb the 2.4 GHz resonance. The thickness sweep from 0.22 to 0.28 mm produces curves that are essentially indistinguishable, as expected for a sheet that is well over one hundred skin depths thick. Sheet thickness tolerance, the parameter least under the control of an economical builder, therefore, has no bearing on performance. Of the twelve dimensions examined, only three require better than millimetre accuracy, and all three lie in the launcher region.

TABLE 4: Summary of Parametric Sensitivity.

Parameter	Frequency influence	Match-depth influence	Tolerance implied
<i>Launcher</i>			
Disc 1 radius r_1	Strong	Strong	Sub-millimetre
Cavity to disc 1 spacing d_{r1}	Strong	Strong	Sub-millimetre
Disc 1 to disc 2 spacing d_{12}	Moderate–strong	Strong	Tight
<i>Transition</i>			
Disc 2 radius r_2	Moderate	Strong	Tight
Disc 3 radius r_3	Moderate	Moderate	Moderate
Disc 2 to disc 3 spacing d_{23}	Weak–moderate	Moderate	Moderate
<i>Guiding section</i>			
Discs 4–6 radius	Weak	Moderate	Relaxed
Spacings d_{34}, d_{45}, d_{56}	Weak	Weak–moderate	Relaxed, ≈ 1 mm
<i>Structure-wide</i>			
Cavity size b_r	Weak	Weak	Relaxed
Disc thickness t	Negligible	Negligible	None

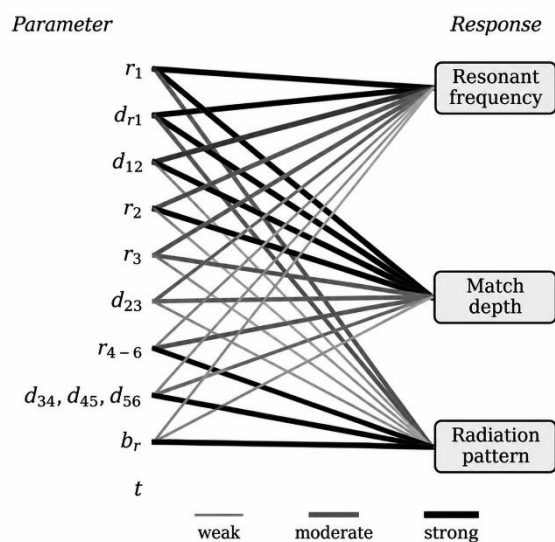


FIGURE 10: Parameter–Response Influence Map of The Sensitivity Analysis. The Sweeps Monitored S_{11} Only; The Links to the Radiation Pattern Follow from the Surface-Wave Interpretation.

DISCUSSION

Physical Interpretation and Classification

The results support a consistent physical account. The cavity and disc 1 launch a bound surface wave; the tapered transition slows it progressively, from $p \approx 1.15$ at disc 1 to $p \approx 1.49$ in the guiding section according to (2), and the uniform guiding section holds p close to the Hansen-Woodyard optimum for the structure length. The accumulated phase delay produces an endfire beam whose gain lies within the envelope predicted by travelling-wave theory; the truncated rod radiates the guided wave forward, and the cavity suppresses the rearward component. The sweeps confirm that each region responds mainly to its own dimensions. Because the launcher sets the input impedance and the guiding section only accumulates phase, a perturbation near the feed alters both the resonant frequency and the match, whereas a perturbation far from the feed has little effect on either. Surface-wave theory predicts

this asymmetry, and the influence of spacing on the input impedance indeed decays with its distance from the launcher.

Three observations distinguish the structure from a Yagi-Uda array. First, the parasitic elements of a Yagi-Uda array are electrically isolated and carry currents induced only by mutual coupling, whereas all six discs here are galvanically bonded to a continuous rod that is connected to the feed. Second, a Yagi-Uda reflector stands $0.15\lambda_0$ to $0.25\lambda_0$ behind the driven element [2], [20]. The cavity here stands $0.034\lambda_0$ from disc 1, between four and seven times closer, which is incompatible with that mechanism but consistent with the role of a launcher. Third, Yagi-Uda directors are close to half a wavelength long and therefore near resonance, whereas the standard estimate for an isolated circular disc radiator in air places the fundamental resonance of the 34.5 mm first disc well below 1 GHz, so none of the discs is individually resonant at 2.4 GHz.

Comparison with Published Designs

Table 5 compares the design with reported 2.4 GHz directional antennas, using only quantities that the cited sources report. The simulated gain is comparable with that of optimized six-element Yagi-Uda arrays of similar electrical length [2], [20]. Unlike the closest published disc structure [12], the present design is accompanied by an absolute gain figure and a systematic sensitivity analysis. The air-dielectric floating patch of Fernandez et al. [21], which also removes the substrate to reduce loss and cost, reports a measured gain of 9.63 dBi with a fractional bandwidth above 10%; the present structure reaches a higher simulated gain from a three-dimensional topology, at the cost of a larger physical envelope. The design does not approach the gain of long backfire structures [6], which are two to four times longer. These comparisons therefore distinguish reported measurements from simulated predictions while retaining the different antenna dimensions and configurations as important context.

TABLE 5: Comparison with Reported 2.4 GHz Directional Antennas.

Design	Method	Structure length	Reported gain	Reported match
Nine-disc structure [12]	CST, simulation	not stated	not reported	−44.6 dB at 2.4068 GHz
Optimized six-element Yagi-Uda [2], [20]	Method of moments	$\approx 1\lambda_0$	10.5 dBi	not stated
Printed Yagi, coupled director [AUTHOR INPUT REQUIRED: source]	Simulation and measurement	planar	high gain, wideband	48% fractional bandwidth
Bi-Yagi array [10]	Simulation and measurement	138.6 mm	6.2 to 9.34 dBi	1.79–2.56 GHz
Long backfire disc-on-rod [6]	Design procedure and measurement	$4\lambda_0$	≈ 24 dBi	not stated
Air-dielectric floating patch [21]	Simulation and measurement	planar (patch)	9.63 dBi	> 10% fractional bandwidth
This work	HFSS, simulation	$1.051\lambda_0$	10.67 dBi	−39.08 dB at 2.400 GHz

Practical Implications

The sensitivity results translate directly into construction guidance. The radius of disc 1 and its spacing from the cavity should be held to better than 0.5 mm. These are also the dimensions most easily adjusted after assembly, so the mechanical design should allow disc 1 to slide along the rod during tuning and be secured afterwards. The three guiding-section spacings can be set with ordinary workshop accuracy, and any available sheet thickness can be used. Because part of the launcher sensitivity follows from its small nominal spacings, a variant with larger launcher spacings, obtained for example with a deeper cavity, should be inherently more tolerant. Solder joints between the discs and the rod lie in the region of highest surface current and should be small and well wetted, and the rod should be continuous, because a discontinuity in the guiding structure would scatter the surface wave and degrade both the match and the front-to-back ratio.

Overall, these recommendations prioritize precise launcher adjustment, practical fabrication tolerances, and continuous conductive connections to preserve the intended antenna performance.

The analysis also supports a compact design procedure: (i) choose L from the required gain using (4) to (6), allowing 1 to 3 dB for the rear reflector; (ii) choose the disc diameter and pitch of the guiding section within the window of [8] and verify (3); (iii) choose a first disc of about $0.55\lambda_0$ diameter and place it a small fraction of a wavelength in front of a cavity of about $0.7\lambda_0$ diameter; (iv) grade the intermediate discs so that p increases progressively; and (v) sweep the first disc radius and the cavity to disc 1 spacing numerically to place the S_{11} minimum at the required frequency. Because the structure contains no dielectric, it scales with frequency. Retargeting the design to 2.45 GHz, for example, requires a scale factor of $2.400/2.450 = 0.9796$ applied to all dimensions except the rod radius and the sheet thickness, which are set by mechanical considerations (Table 6).

TABLE 6: Dimensions of the Present Design Rescaled to a Centre Frequency Of 2.45 GHz.

Dimension	At 2.400 GHz (mm)	Rescaled to 2.450 GHz (mm)
Disc 1 radius	34.50	33.80
Disc 2 radius	26.50	25.96
Disc 3 radius	19.50	19.10
Discs 4–6 radius	18.00	17.63
Cavity radius	45.00	44.08
Cavity depth	15.00	14.69
Cavity to disc 1 spacing	4.20	4.11
Disc 1 to disc 2 spacing	9.00	8.82
Disc 2 to disc 3 spacing	32.00	31.35
Disc 3–4, 4–5, 5–6 spacing	28.70	28.11
Rod length	160.00	156.73
Rod radius (unscaled)	2.00	2.00
Disc thickness (unscaled)	0.25	0.25

In a link, the free-space range at a fixed receiver sensitivity scales as $\sqrt{G}$ [22]. Replacing a 2 dBi sleeve dipole at one end of a link with the present antenna would therefore increase the free-space range by a factor of 2.7, and using the antenna at both ends would increase it by a factor of 7.4. Table 7 evaluates a representative free-space link budget. In an obstructed environment with path-loss exponent n , the range scales as $G^{1/n}$, so that at $n = 3.5$ the same 8.7 dB gain advantage yields a factor of 1.75 rather than 2.7. The narrow beam must also be aimed at and its polarization aligned with that of the far end; a 30° misalignment costs 1.25 dB. The gain does not increase the permitted radiated power. Under the 20 dBm equivalent isotropically radiated power (EIRP) limit that applies to general use of the 2.4 GHz band in China and in Europe [23], the conducted transmitter power must be reduced to approximately 9.3 dBm. The benefit is therefore realized as receiving sensitivity, rejection of interference outside the main lobe, and link margin at a compliant radiated power.

TABLE 7: Free-Space Link Budget at 2.44 GHz for 20 dBm EIRP with the proposed antenna at the receiving end.

Range (m)	Path loss (dB)	Received power (dBm)	Supported 802.11n rate
10	60.2	-29.5	MCS7
100	80.2	-49.5	MCS7
300	89.8	-59.1	MCS7
500	94.2	-63.5	MCS7 (marginal)
1000	100.2	-69.5	MCS4 to MCS5

Limitations and Proposed Validation

All quantitative results come from full-wave simulation; we have not measured the reflection coefficient, the gain, or the radiation pattern. Four further limitations apply. First, the excitation is a wave port on a bare feed pin. A model that represents the coaxial line explicitly, with its outer conductor bonded to the cavity, would define the return-current path more rigorously, and connector and solder interfaces will add a small loss in any physical realization. Second, we did not extract the radiation efficiency, the total efficiency, the input impedance as a function of frequency, the orthogonal-plane pattern cut, or the cross-polarization level. The radiation efficiency is expected to exceed 0.97 for an all-copper structure in air, but this is unconfirmed, and the origin of the wide matched band therefore remains incompletely established. Third, the parametric study monitors S11 alone, although the guiding-section dimensions act mainly on the accumulated guided phase and hence on the far-field gain, so a complete optimization should monitor gain and pattern in parallel with the match. Fourth, all simulations assume free space.

Validation should follow established antenna measurement practice [24]. The largest dimension

of the antenna is approximately 0.16 m, for which the far-field criteria $2D^2/\lambda$, $5D$ and 1.6λ evaluate to 0.41 m, 0.80 m and 0.20 m, so a separation of at least 1 m satisfies all three. The reflection coefficient should be measured with a vector network analyzer calibrated at the connector reference plane, with a ferrite choke or a quarter-wave sleeve balun fitted to suppress current on the cable shield. The gain can be obtained by the three-antenna method or by comparison with a calibrated standard; both principal planes should be recorded with co-polar and cross-polar responses, and the radiation efficiency can be obtained from the ratio of gain to directivity or by the Wheeler cap method.

Three extensions follow. A Monte Carlo tolerance study based on the ranking of Table 4 would quantify the manufacturing yield of the design. Extending the guiding section from 1.05 to approximately 3 wavelengths, which requires eight additional discs at the same pitch, and adding a large terminating reflector to form a long backfire antenna should raise the gain towards the 18 to 20 dBi reported for that configuration [6], [7]. A quarter-wave choke on the cavity rim is a further candidate refinement.

CONCLUSIONS

We designed and numerically analyzed a six-disc copper endfire radiator for the 2.4 GHz band, built around a continuous rod and a cavity-backed launcher. The galvanic bonding of every disc to a common rod, the 0.034λ spacing between the cavity and the first disc, and the absence of individual disc resonance at the design frequency place the structure in the short-backfire disc-on-rod family, and we analyzed it with surface-wave and travelling-wave theory. Full-wave simulation gives a reflection coefficient of -39.08 dB at 2.400 GHz and a -10 dB band from approximately 2.31 to 2.58 GHz. Converting the field-unit output of the solver gives a peak realized gain of 10.67 dBi, which lies within the envelope of analytical endfire estimates for a one-wavelength aperture and equals 55% of the Harrington normal-gain bound, short of the super gain regime. The disc size and pitch of the uniform guiding section, obtained by numerical optimization, fall inside the empirical design window reported for this antenna class. The twelve-parameter sweep separates a launcher that must be held to sub-millimetre tolerance from a guiding section that tolerates about 1 mm of error, and it shows that the conductor thickness has no measurable electromagnetic role. These results rest entirely on simulation. The next steps are to measure a prototype, extract radiation efficiency, run a statistical tolerance study, and extend the guiding section towards a long-backfire configuration.

ACKNOWLEDGMENT

The author thanks the Department of Electronic Information Engineering, Nanjing University of Information Science and Technology, for access to the Ansys HFSS license and computing resources used in this study, the supervisor for guidance on antenna theory, on the interpretation of full-wave

simulation results and on the classification of the structure, and colleagues in the department for discussions during the parametric sweeps and for comments on early drafts.

REFERENCES

- [1] Balanis, C. A. (2016). *Antenna Theory: Analysis and Design* (4th ed.). Hoboken, USA: Wiley.
- [2] Kraus, J. D., & Marhefka, R. J. (2002). *Antennas for All Applications* (3rd ed.). New York, USA: McGraw-Hill.
- [3] Ehrenspeck, H. W., & Poehler, H. (1959). A new method for obtaining maximum gain from Yagi antennas. *IRE Transactions on Antennas and Propagation*, 7(4), 379–386.
- [4] Ehrenspeck, H. W. (1960). The backfire antenna, a new type of directional line source. *Proceedings of the IRE*, 48(1), 109–110.
- [5] Ehrenspeck, H. W. (1965). The backfire antenna: New results. *Proceedings of the IEEE*, 53(6), 639–641.
- [6] Kirov, G. S., Hristov, H. D., & Feick, R. (2009). Backfire disk-on-rod antenna for long-range point-to-point communications. *Microwave Journal*, 52(6).
- [7] Kirov, G. S., & Hristov, H. D. (2011). Study of backfire antennas. *Journal of Microwaves, Optoelectronics and Electromagnetic Applications*, 10(1), 1–12.
- [8] Bogner, R. D. (1981). Disc-on-rod end-fire microwave antenna (U.S. Patent No. 4,295,141). Washington, USA: U.S. Patent and Trademark Office.
- [9] Rana, M. S., Sen, B. K., Mamun, M. T.-A., Mahmud, M. S., & Rahman, M. M. (2023). A 2.45 GHz microstrip patch antenna design, simulation, and analysis for wireless applications. *Bulletin of Electrical Engineering and Informatics*, 12(4), 2173–2184.
- [10] Abdulwali, Z. S. A., Alqahtani, A. H., Aladadi, Y. T., Alkanhal, M. A. S., Al-Moliki, Y. M., Aljaloud, K., & Alresheedi, M. T. (2023). Miniaturized broadband Bi-Yagi antenna array for ambient RF energy harvesting. *Micromachines*, 14(12), 2181.
- [11] Constantinescu, C., Pacurar, C., Giurgiuman, A., Munteanu, C., Andreica, S., & Gliga, M. (2023). High-gain improved planar Yagi Uda antenna for 2.4 GHz applications and its influence on human tissues. *Applied Sciences*, 13(11), 6678.
- [12] Alcalá Garrido, H. A., Barrera Figueroa, V., & Rivero-Angeles, M. E. (2018). Design and simulation of antennas for energy harvesting systems in the WiFi band. In M. F. Mata-Rivera & R. Zagal-Flores (Eds.), *Telematics and Computing: WITCOM 2018 (Communications in Computer and Information Science, Vol. 944, pp. 45–55)*. Cham, Switzerland: Springer.
- [13] Goubau, G. (1950). Surface waves and their application to transmission lines. *Journal of Applied Physics*, 21(11), 1119–1128.
- [14] Collin, R. E., & Zucker, F. J. (Eds.). (1969). *Antenna Theory, Part 2*. New York, USA: McGraw-Hill.
- [15] Hansen, W. W., & Woodyard, J. R. (1938). A new principle in directional antenna design. *Proceedings of the IRE*, 26(3), 333–345.
- [16] Harrington, R. F. (1958). On the gain and beamwidth of directional antennas. *IRE Transactions on Antennas and Propagation*, 6(3), 219–225.
- [17] IEEE. (2013). *IEEE Standard for Definitions of Terms for Antennas (IEEE Std 145-2013)*. New York, USA: IEEE.
- [18] Jin, J.-M. (2014). *The Finite Element Method in Electromagnetics* (3rd ed.). Hoboken, USA: Wiley-IEEE Press.
- [19] Chu, L. J. (1948). Physical limitations of omnidirectional antennas. *Journal of Applied Physics*, 19(12), 1163–1175.
- [20] Viezbicke, P. P. (1976). *Yagi Antenna Design (NBS Technical Note 688)*. Boulder, USA: National Bureau of Standards.
- [21] Fernández, M. D., Herraiz, D., Herraiz, D., Alomainy, A., & Belenguer, A. (2022). Design of a wide-bandwidth, high-gain and easy-to-manufacture 2.4 GHz floating patch antenna fed with the through-wire technique. *Applied Sciences*, 12(24), 12925.
- [22] Friis, H. T. (1946). A note on a simple transmission formula. *Proceedings of the IRE*, 34(5), 254–256.
- [23] ETSI. (2019). *Wideband transmission systems; Data transmission equipment operating in the 2,4 GHz band; Harmonised standard for access to radio spectrum (ETSI EN 300 328 V2.2.2)*. Sophia Antipolis, France: ETSI.
- [24] IEEE. (2021). *IEEE Recommended Practice for Antenna Measurements (IEEE Std 149-2021)*. New York, USA: IEEE.