

Research Paper

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A wide axial ratio bandwidth circularly polarized filtering antenna using two quarter-wavelength resonators and a single feedline

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Abstract

This study proposes a single feedline technique that results in a wide axial ratio bandwidth (ARBw) of a circularly polarized filtering antenna. The antenna is designed to operate at 4.65 GHz. The design employed a rectangular radiator and two quarter-wavelength resonators. The two components are integrated using a single feedline and proximity coupling. The quarter-wavelength resonators excite the second mode polarization on the coupled rectangular radiator. The combination of width and length of the radiator, as well as the gap between the second resonator and radiator, equalized the two modes' e-fields' magnitude. The design requires no additional delay circuit, for the gap between the second resonator and radiator provides both selectivity and a 90° phase shift delay. The proposed design has a wide ARBw of 10.10% and an impedance bandwidth of 9.04%. The use of two resonators results in a gain curve resembling a bandpass filter response, with a maximum gain of 7.6 dBic. The antenna is fabricated and measured to validate the proposed method, and the measurement results align well with the simulation results.

Introduction

A circularly polarized (CP) antenna has the benefit of combating the multipath propagation [1], accommodating the object movement [2], and increasing the channel capacity rather than the linear polarization [3]. Many methods were proposed to obtain a CP patch antenna, such as using a single feedline with radiator perturbation [4, 5], helical shape antenna [6, 7], and two orthogonal elements [8, 9]. In the two orthogonal feedlines method, a 90° delay circuit is essential to ensure the generation of circular polarization. Conversely, in the single feedline approach, a perturbation radiator is employed to manipulate the surface current flow on the radiator, achieving circular polarization. It is known that two orthogonal feedline techniques can provide wider axial ratio bandwidth (ARBw) than the single feedline because, with the right composition, the two feedlines can create two oscillations on axial ratio, thus widening the ARBw [2, 10]. However, many single feedline techniques have been conducted to obtain a wide ARBw using parasitic elements and radiator perturbation [11–15].

In most cases, integrating an antenna and filter is primarily performed to add the selectivity function to an antenna [16–19]. However, some studies have proposed filtering antennas with an added circular polarization feature, achieved through two orthogonal feedlines [2, 10, 20], circular disk radiator [21], truncated radiators, and cross slots [22–25]. These methods typically involve a two-step design process, with the first step focusing on the filtering structure design and the second step on the generating circular polarization design. The most common method for designing a filtering antenna is by replacing the last stage resonator of the filter with a radiator [21, 26]. Several studies have explored these types of filtering antennas with circular polarization using two orthogonal feedlines [2, 10, 20, 27]. However, all CP filtering antennas that employ two feedlines were designed with both a filtering structure and an added delay circuit, making the design process more complicated due to the requirement of a two-step design approach. Moreover, employing two orthogonal feedlines further complicates the design of array antennas.

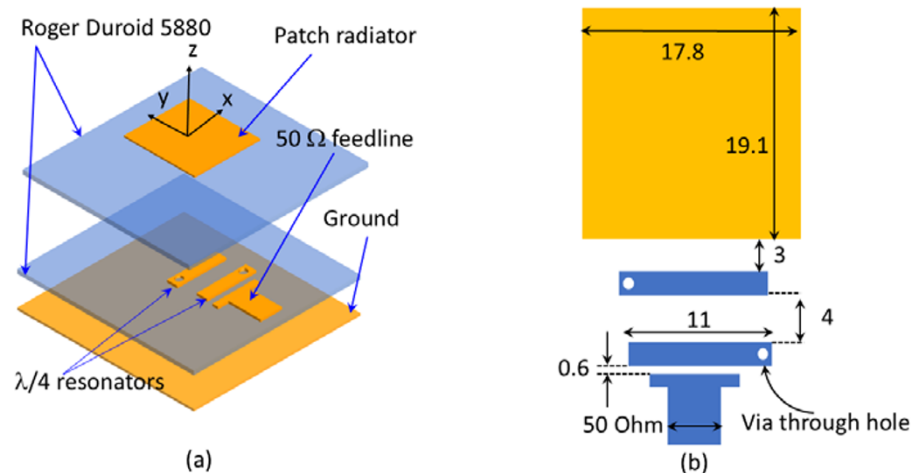


Figure 1. The configuration of the proposed CP filtering antenna (a) 3-dimensional view with transparent substrates and (b) top view with transparent substrates and dimensions (all units in mm).

Recently, a new method of combining an integrated antenna and filter component has resulted in a filtering antenna with a linear or circular polarization [28–31]. This approach enables the simultaneous achievement of selectivity and polarization features through an integrated step design. By applying this method, a second-order CP filtering antenna is designed using a single feedline [28]. However, its ARBW remains narrow even when designed with double-layer substrates. Here, we propose a novel third-order filtering antenna design that integrates circular polarization generation using only a single feedline. In contrast to previous work [28], which employed lower-order designs, our approach enhances selectivity, impedance, and ARBW by leveraging the inherent benefits of a higher-order filtering circuit. By codesigning the filtering circuit and the polarization generator, we eliminate the need for an additional delay circuit, thereby simplifying the overall architecture. The proposed antenna is optimized for operation at 4.65 GHz, well-suited for C-band wireless systems where polarization robustness and inherent out-of-band suppression are required. Potential applications include short- to medium-range radar systems, microwave point-to-point communication links, and RF energy harvesting, where compactness, single-feed simplicity, and integrated filtering characteristics offer practical advantages.

Antenna design

Antenna configuration

As shown in Fig. 1(a) and (b), the antenna consists of a rectangular radiator measuring 17.8 mm × 19.10 mm printed on the upper substrate, coupled with two $\lambda/4$ resonators with a thickness of 2.4 and 11 mm in length lying on the lower substrate. There is a 3 mm gap between the rectangular radiator and the second $\lambda/4$ resonator, which not only provides coupling between them but also a 90° phase delay. There is a 4 mm gap between the two $\lambda/4$ resonators. A 50-ohm feedline is coupled to the first resonator with a gap of 0.6 mm. The extraction of the radiator and the resonator is conducted as in Ref. [28], where the length of the radiator (L_p) determined the radiator resonance frequency, while the length (L_k), thickness of the resonator, and the position of a via through hole determined the resonator's resonance frequency [28, 29, 31]. The two $\lambda/4$ resonators with a via through-hole not only serve to shape the gain for enhanced selectivity but also act as a mechanism to induce second-mode polarization in the radiator.

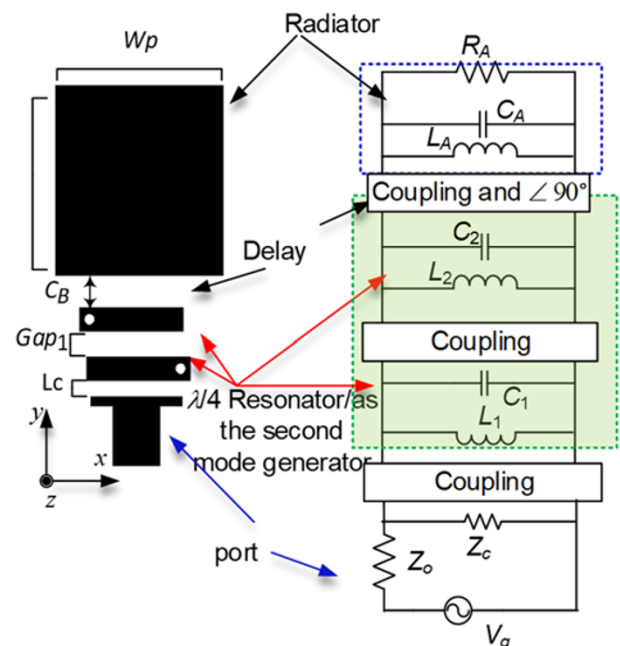


Figure 2. Equivalent circuit and design concept.

Equivalent circuit and design procedure

The equivalent circuit network of the CP filtering antenna is shown in Fig. 2, where two series resonators are described as L_1 , C_1 , and L_2 , C_2 for the first and second resonators, respectively. The short circuit holes on the left and right arms of the first and second resonators lead to a horizontal current flow along the transmission line. Upon reaching the rectangular radiator (L_A , C_A , and R_A), there arises an interaction between two electric field vectors that are orthogonal to each other. The CP filtering antenna is obtained when the two vectors' magnitudes are balanced, and the phase delay is $\pm 90^\circ$. The space between the second resonators and the radiator (C_B) functions both as the delay circuit and coupling. The radiator's dimensions are not square, dissimilar to the orthogonal feedline methods. W_p should be marginally shorter than L_p to balance the magnitudes of the orthogonal vectors. The space between the resonator (Gap_1) affects the IBW, selectivity response, and axial ratio at the center and lower frequency.

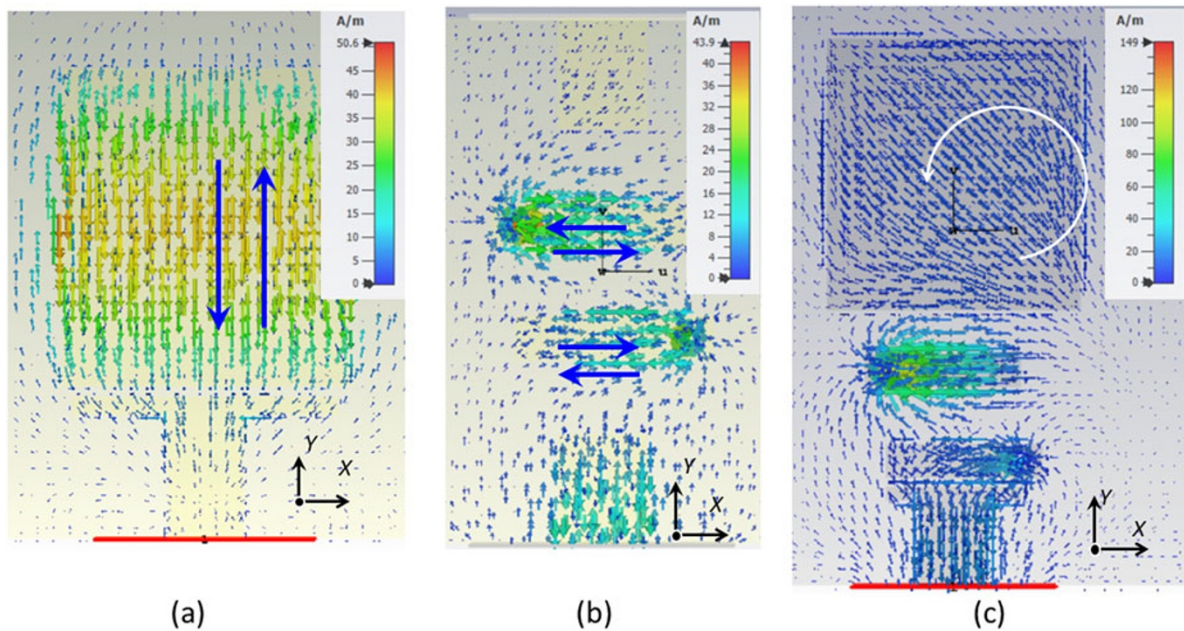


Figure 3. Concept of orthogonal resonators from the surface current of the (a) $\lambda/4$ resonator as the first resonator, (b) radiator as the second resonator, and (c) the interaction of the two vectors on the filtering antenna's radiator.

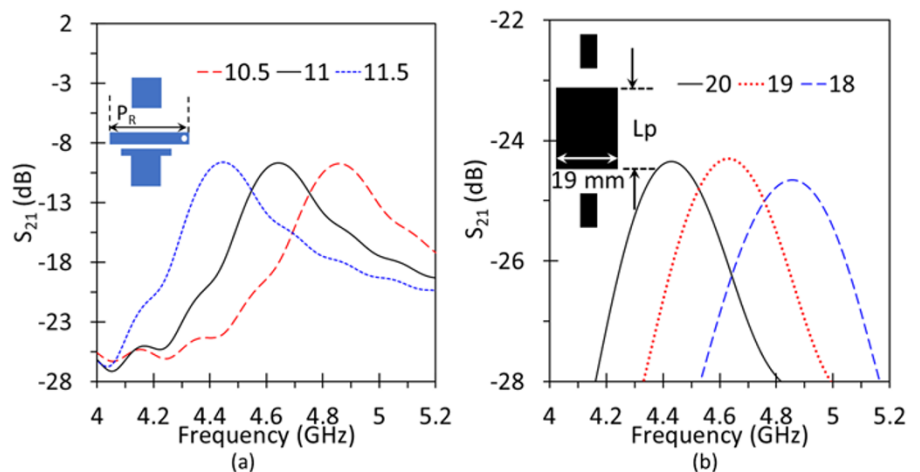


Figure 4. The resonance frequency simulation of (a) the radiator as a function of the variation in its length (L_p) with a fixed width ($W_p = 19$ mm) and (b) the resonator as a function of the variation in its length (P_R). All dimensional units are in millimeters.

The concept of the orthogonal resonators refers to the fundamental equation of the electromagnetic wave polarization vector in Eq. (1). This equation explains that the polarization vector is the resultant of the superposition of two distinct polarization vectors, each oriented orthogonally in the θ and ϕ directions. The production of circular polarization is subject to the equal magnitude of the θ and ϕ vectors as in Eq. (2) and a phase difference (δ_L) of approximately $\pm 90^\circ$ between these vectors. Consequently, the orthogonality of these vectors can be established through the integration of two resonators featuring orthogonal current orientations. In the context of this study, this is exemplified by the utilization of the two quarter-wavelength resonators in the θ direction as the first resonator, and a rectangular radiator in the ϕ direction as the second resonator.

$$\vec{E} = E_\theta \cos(\omega t) \hat{\theta} + E_\phi \cos(\omega t + \delta_L) \hat{\phi} \quad (1)$$

$$|E_\theta| = |E_\phi| \quad (2)$$

The orthogonality can be observed from the surface currents of both resonators, as seen in Fig. 3(a), where the horizontal surface current of two $\lambda/4$ resonators flows reversely along the x -direction from the open circuit to the short circuit (via through hole). On the contrary, Fig. 3(b) depicts the radiator surface current that flows along the vertical position or y direction when fed from the lower side of its length. Both circuits must operate at an identical frequency. The integration result of the two vector directions is shown on the radiator, forming the right-hand circular polarization (RHCP) as depicted in Fig. 3(c). The length of the $\lambda/4$ resonator (P_R) and the position of the via through-hole determine its resonant frequency as shown in Fig. 4(a). Two $\lambda/4$ resonators are used to create the two minimal axial ratios and multiple resonants in S_{11} . The addition of the resonator in filter design also aims to increase its selectivity. The resonant frequency of the rectangular radiator shown in Fig. 4(b), is affected essentially by its length (L_p) while its width (W_p) can be adjusted for balancing the magnitude of the two orthogonal vectors.

After obtaining the frequencies of both resonators, the integration is carried out using a single port. The determination of the gap distance between the $\lambda/4$ resonator and the port (L_c) is made with the consideration of being sufficiently close to provide a strong current to the feedline, thus setting the distance at 0.6 mm. So, there are four variables left to obtain the CP characteristics which are the size of the radiator, the space between the $\lambda/4$ resonators and the space between the radiator and $\lambda/4$ resonator. From the previous study, the space between the radiator and resonator functions both as the delay and matching circuit [28], then in this design, we start with 3.6 mm to iterate the response. The optimization goal was to achieve the axial ratio less than 3 dB over the bandwidth and IBW of -10 dB using the variations of W_p , L_p , C_B , and Gap_1 .

Analysis of coupled quarter-wavelength resonators

In Ref [16], a single quarter-wavelength resonator was employed to implement the filtering structure for the antenna, which provides acceptable filtering performance. However, the single-resonator approach inherently limits the operational bandwidth and selectivity. In this work, we extend the design by employing two coupled quarter-wavelength resonators as shown in Fig. 2. It is well established that a single quarter wavelength resonator typically yields a first-order response with a relatively narrow bandwidth and a moderate roll-off. In contrast, when two quarter-wavelength resonators are coupled together, a higher-order response is achieved. This coupled configuration introduces additional degrees of freedom, allowing for enhanced selectivity, a broader operational bandwidth, and steeper attenuation slopes in the transition bands. The coupling also enables tailoring the resonant modes, leading to improved out-of-band rejection and a more flexible design for achieving the desired filter characteristics.

To explain the presence of a higher-order mode in our proposed filter design, the coupled mode theory (CMT) framework is used to derive analytical expressions for the transmission S_{21} and reflection S_{11} coefficients. In this model, we consider two resonators (assumed identical) having a resonant angular frequency $A = \omega_0$ and are coupled to ports 1 and 2 with an external coupling rate $\sqrt{\gamma_e}$. They are also mutually coupled with a coupling coefficient κ . Assuming negligible intrinsic loss, the time-domain dynamics (with a harmonic excitation $s_{1+}(t) = s_{1+}e^{-j\omega t}$) lead to the steady-state equations for the resonator amplitudes a_1 and a_2 written as [32, 33],

$$a_2 = -\frac{j\kappa}{-j(\omega - \omega_0) + \frac{\gamma_e}{2}}a_1, \quad (3)$$

$$a_1 = \frac{\sqrt{\gamma_e}s_{1+} \left[-j(\omega - \omega_0) + \frac{\gamma_e}{2} \right]}{\left[-j(\omega - \omega_0) + \frac{\gamma_e}{2} \right]^2 + \kappa^2}a_1. \quad (4)$$

The output at port 2 is given by the standard input–output relation $s_{2-} = -\sqrt{\gamma_e}a_2$, which gives us,

$$s_{2-} = \sqrt{\gamma_e} \frac{j\kappa}{-j(\omega - \omega_0) + \frac{\gamma_e}{2}}a_1. \quad (5)$$

Substituting the expression for a_1 we obtain

$$s_{2-} = \gamma_e \frac{j\kappa s_{1+}}{\left[-j(\omega - \omega_0) + \frac{\gamma_e}{2} \right]^2 + \kappa^2}. \quad (6)$$

Thus, we can derive the transmission coefficient as,

$$S_{21}(\omega) = \frac{s_{2-}}{s_{1+}} = \frac{\gamma_e j\kappa}{\left[-j(\omega - \omega_0) + \frac{\gamma_e}{2} \right]^2 + \kappa^2}. \quad (7a)$$

Similarly, using the input–output relation at port 1, $s_{1-} = s_{1+} - \sqrt{\gamma_e}a_1$, we obtain the reflection coefficient

$$S_{11}(\omega) = 1 - \frac{\gamma_e \left[-j(\omega - \omega_0) + \frac{\gamma_e}{2} \right]}{\left[-j(\omega - \omega_0) + \frac{\gamma_e}{2} \right]^2 + \kappa^2}. \quad (7b)$$

To confirm the above CMT model in our proposed filter design, we simulate a microstrip filter using two quarter-wavelength resonators in CST Microwave Studio. Details of the filter geometrical parameters can be found in Fig. 5, which illustrates the coupled quarter-wavelength resonators analysis based on the coupled-mode theory (CMT) model. Figure 5(a) illustrates the geometrical parameters of the microstrip filter utilizing two quarter-wavelength resonators. Figure 5(b) shows the simulation setup in CST. Figure 5(c) and 5(d) presents the comparison between simulation results and analytical calculation results using CMT for Fig. 5(c) transmission coefficient (S_{21}) and Fig. 5(d) reflection coefficient (S_{11}). Figure 5(a) and the simulation setup is shown in Fig. 5(b). Figure 5(c) and (d) shows the transmission and reflection coefficients obtained by using the simulation and the analytical calculation (Eqs. (7a) and (7b)). In our analytical model, we employed a resonant frequency $f_0 = 4.65$ GHz, an external coupling rate of $\gamma_e = 1.7 \times 10^9$ per port, and a resonator coupling coefficient $\kappa = 1.25 \times 10^9$. The simulation results confirm that the two-resonator configuration achieves a broader bandwidth while maintaining improved selectivity, as evidenced by the enhanced transmission and reduced reflection over the frequency range of interest. The two coupled quarter-wavelength resonators yield a higher-order response with steeper roll-off compared to single-resonator designs, enabling better selectivity.

A 90° phase shifter between $\lambda/2$ and $\lambda/4$ resonators

In Analysis of coupled quarter-wavelength resonators resonant modes using higher order is proven to enhance the selectivity response and expand the bandwidth impedance of the filter using $\lambda/4$ resonator as demonstrated both through simulation and theoretical analysis. In this section, we investigate the $\lambda/4$ resonator's function as a passive phase shifter in the circuit. Experimentally and theoretically, the $\lambda/4$ resonator as a 180° passive phase shifter has been discussed in Refs. [29, 34], however, as explained in Equivalent circuit and design procedure, the circuit needs a $\pm 90^\circ$ phase difference between the orthogonal modes to obtain a circular polarization.

The design initiates with the radiator functioning as a $\lambda/2$ resonator; without the existence of the filter, the radiator has an initial vertical polarization. The radiator can be modeled as a terminated transmission line, where the radiator impedance corresponds to the energy radiated to the free space. The ABCD matrix ($M(l)$) for a transmission line of length l is

$$M(l) = \begin{bmatrix} \cos(\beta l) & jZ_0 \sin(\beta l) \\ \frac{j}{Z_0} \sin(\beta l) & \cos(\beta l) \end{bmatrix}, \quad (8)$$

where Z_0 is the characteristic impedance, β is propagation constant ($2\pi/\lambda$) and l is the resonator's length.

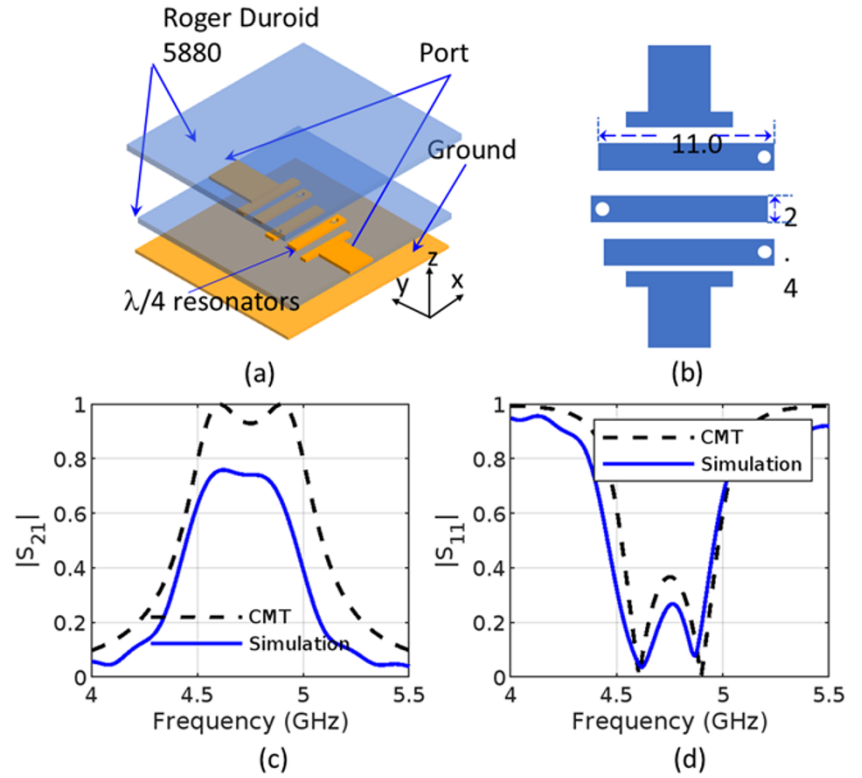


Figure 5. Coupled quarter-wavelength resonators analysis by the coupled-mode theory (CMT) model. (a) Geometrical parameters of the microstrip filter utilizing two quarter-wavelength resonators. (b) Simulation setup in CST. (c) and (d) Comparison between simulation results and analytical calculation results using CMT for (c) transmission coefficient (S_{21}) and (d) reflection coefficient (S_{11}).

The radiator performs as the $\lambda/2$ resonator, so we can substitute $l = \lambda/2$, then the ABCD matrices are

$$(M_{/2}) = \begin{bmatrix} -1 & 0 \\ 0 & -1 \end{bmatrix}. \quad (9a)$$

The second mode polarization arises when the $\lambda/4$ resonator with horizontal surface current coupled to the radiator, generates an orthogonal vector polarization relative to the initial radiator's vector polarization. Substituting the $l = \lambda/4$ in Eq. (8), this leads to

$$M (/4) = \begin{bmatrix} 0 & jZ_0 \\ \frac{j}{Z_0} & 0 \end{bmatrix}. \quad (9b)$$

Performing the multiplication:

$$M_{\text{total}} = M_{/2} \cdot M_{/4}.$$

$$M_{\text{total}} = \begin{bmatrix} 0 & -jZ_0 \\ -\frac{j}{Z_0} & 0 \end{bmatrix}. \quad (9c)$$

The δ_L between the orthogonal resonators can be obtained from the voltage along the x -axis generated by the $\lambda/4$ resonator is:

$$V_x = jZ_0 \cdot I_0. \quad (10a)$$

Voltage along the y -axis generated by the $\lambda/2$ resonator (radiator) is

$$V_y = -jZ_0 \cdot I_0 \quad (10b)$$

whereas:

$$\delta_L = \angle V_x - \angle V_y \quad (11)$$

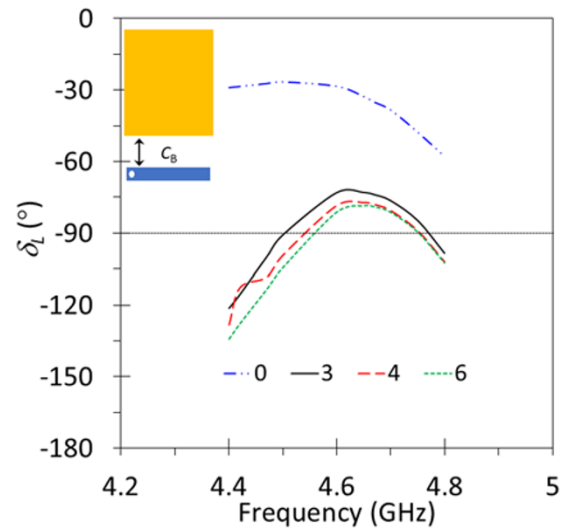


Figure 6. The δ_L response under variation of C_B (unit in mm).

The phase of the two voltages is $\angle jZ_0 = 90^\circ$ and $\angle -jZ_0 = -90^\circ$, results in δ_L of 180° , which corresponds to linear polarization. To obtain circular polarization, we need a 90° phase shifter by modifying the coupling between the orthogonal resonators. From the equivalent circuit, it can be understood that tuning the reactance through the C_B adjustment affect the coupling and δ_L . The response of the C_B variations and δ_L can be seen in Fig. 6, where with C_B of 0 mm or no distance between the radiator and resonator the δ_L is close to 0° . Whereas increasing C_B shifts the δ_L close to -90° and extending the C_B from 3 to 6 mm shows only a slight shift in δ_L . The

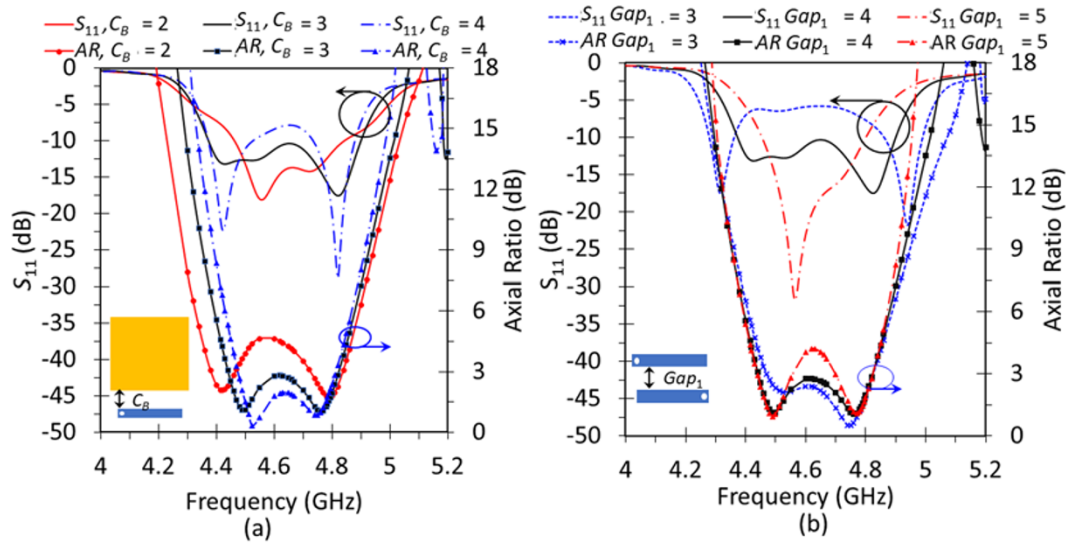


Figure 7. S_{11} and axial ratio response in different (a) C_B and (b) Gap_1 (all units in mm).

CB-tuned 90° phase shift simplifies CP generation without external circuits, advantageous for scalable array designs.

Results and discussion

Parametric study

The optimized results obtained from the iterative process, along with the corresponding responses of S_{11} and axial ratio under varying C_B , are shown in Fig. 7(a). It is observed that C_B exerts a notable impact on both impedance and AR values in the middle and lower band. Specifically, deviations from the 4 mm leads to a reduction in AR bandwidth, indicating a sensitivity to C_B alterations. Moreover, variations in S_{11} degrade with decreases or increases in C_B from the 4 mm.

The variation of Gap_1 to the S_{11} and AR is depicted in Fig. 7(b). The -10 dB IBW and S_{11} are significantly affected by the variation of Gap_1 . A wider Gap_1 improves S_{11} but degrades the bandwidth, and vice versa. AR is also considered in this optimization within the same parametric study, Gap_1 influences the AR in the middle and lower bands. Two minimum optimized ARs occur at 4.49 and 4.76 GHz, resulting in 380 MHz ARBW or 8.2%. The characteristic of the RHCP can be presented by the magnitude ratio of the orthogonal electrical field and its phase difference, as shown in Fig. 8. The ratio of E_θ and E_ϕ magnitude over the bandwidth 4.5–4.8 GHz is varied between the range 1.04 to 1.4 while the phase differences are between 43° and 106° . These values are close to the ideal value of 1° and 90° .

Measurement

To enhance the confirmation of both the proposed method and the simulation outcomes, we construct and assess a prototype of the CP patch antenna. The antenna is fabricated on two layers of Roger Duroid 5880, each with a thickness of 1.575 mm and a permittivity of 2.2, as shown in Fig. 9 (a)–(c). In this study, the fabricated antenna is measured using the Rohde Schwarz ZNB-40 vector network analyzer as shown in Fig. 9(d). The measured ARBW is depicted in Fig. 10(a), where the two minimum axial ratios are observed at 4.48 and 4.79 GHz. The presence of two distinct

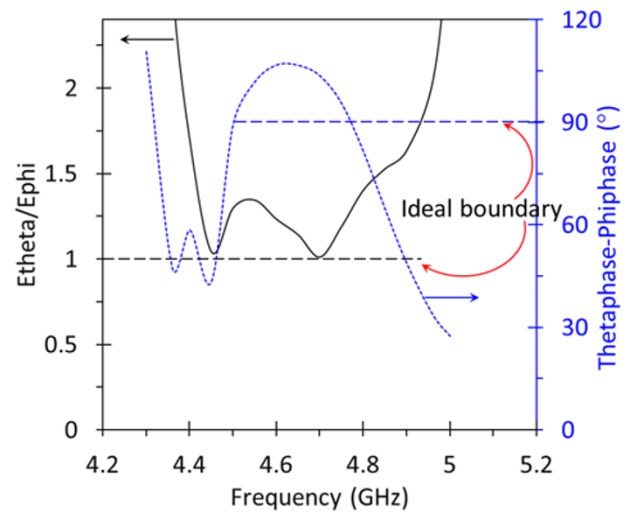


Figure 8. Magnitudes and phases of the E_θ and E_ϕ .

minimum ARs implies that the antenna achieves approximately three times the AR bandwidth of the second-order filter-based design in Ref. [28]. The AR bandwidth measurement is 10.1% span from 4.41 to 4.88 GHz or 1.9% wider than the simulation result. The reflection coefficients obtained through simulation and measurement are plotted in Fig. 10(b). The reflection coefficient result shows that the measured antenna covers a wide -10 dB IBW from 4.49 to 4.93 GHz, equivalent to a 9.04% fractional bandwidth, which is 2.16% narrower than the simulation result of 11.2%. The lower frequency shifts by 120 MHz from 4.37 to 4.49 GHz, while the upper frequency exhibits a slight shift of 20 MHz. The discrepancies observed in the ARBW and IBW results are attributed to fabrication imperfections, angular averaging, and off-boresight improvements during the AR measurement. The simulated gain response remains flat within a frequency band centered at 4.65 GHz and gradually decreases on both sides of this band, demonstrating desirable filtering characteristics that are better than the previous design in Ref. [28]. The measurement result exhibits a slightly

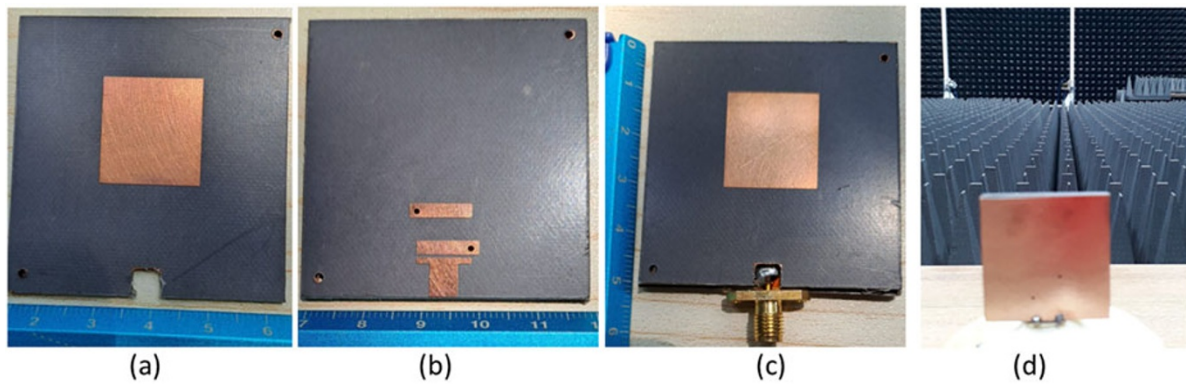


Figure 9. The fabricated antenna (a) radiator, (b) feedline, (c) assembled prototype, and (d) measurement.

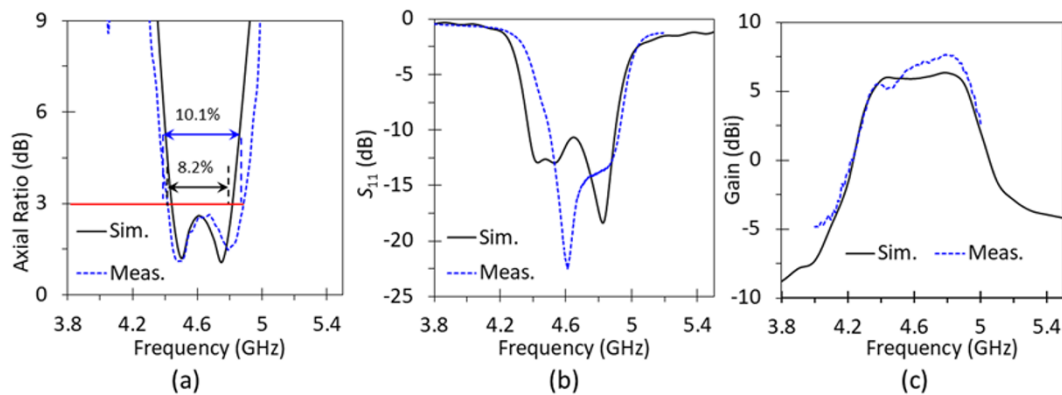


Figure 10. Simulated and measured results of (a) axial ratio, (b) S_{11} , and (c) gain of the proposed design.

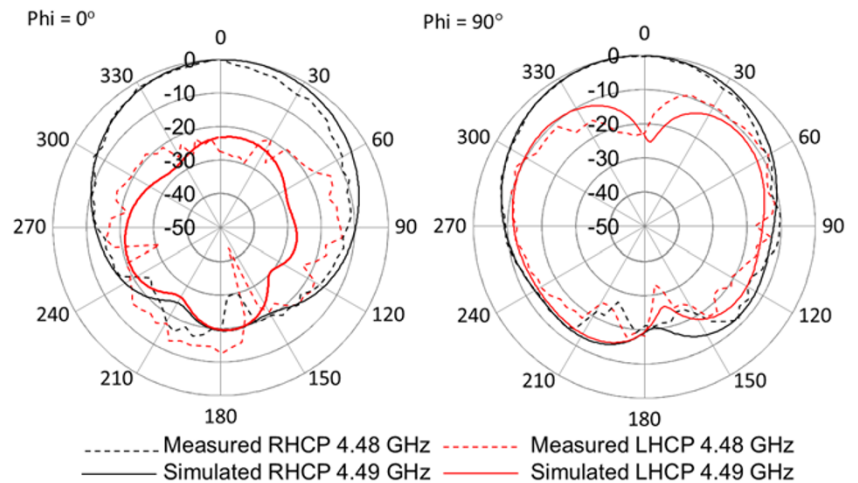


Figure 11. Normalized radiation pattern at 4.48 (simulation)/4.48 (measurement) GHz.

lower gain at lower frequencies compared to the simulation result, as shown in Fig. 10(c). This discrepancy is attributed to a higher measured reflection coefficient (S_{11}) in this region, indicating sub-optimal signal transmission. Conversely, at higher frequencies, the measured gain is rather higher than the simulated result. This is likely caused by an air gap between the stacked substrates above the second resonator, which may have formed during measurement and contributed to the gain increase. The maximum gain of simulation and measurement is 6.30 at 4.65 GHz and 7.60 dBi at 4.81 GHz, respectively.

The far-field patterns were measured at two frequency points (4.48 and 4.79 GHz), corresponding to two minima in the AR response. The normalized gain patterns for RHCP and left-hand circular polarization for both frequencies are depicted in Figs. 11 and 12. The cross-polarization gain is roughly 20 dB less than the co-polarization gain in the direction of the main beam. The radiation patterns reveal a coherent beam shape in co-polarization, underlining the stable performance of the radiation. The 3 dB beamwidths measured in the $\phi = 0^\circ$ at two different frequencies are 76.5° and 78.6° , respectively. Similarly, in the $\phi = 90^\circ$, the

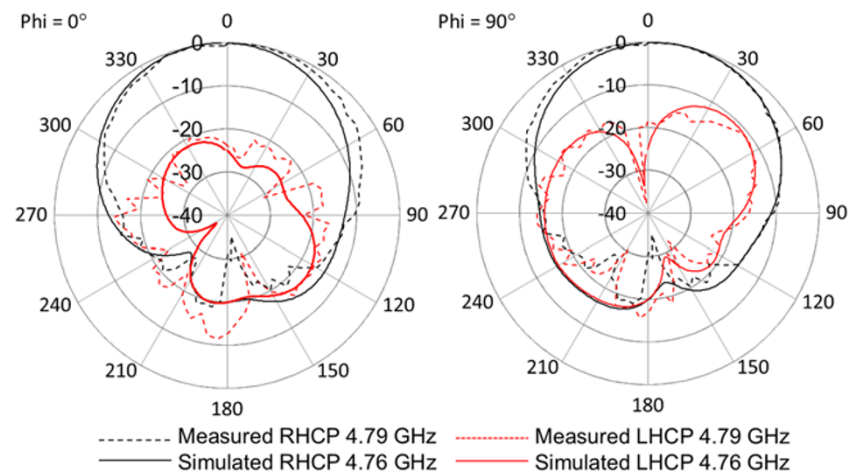


Figure 12. Normalized radiation pattern at 4.76 (simulation)/4.79 (measurement) GHz.

Table 1. Comparison of the proposed design with previous published CP filtering antenna

Ref.	Dimension (L × W × T) in λ_0	Integrated CP and filtering method	CP method	Phase shifter analysis	IBw/ARBw (%)	Substrate layer	Gain (dBic)
[10]	1.06 × 1.06 × 0.027	No	Orthogonal feedline	Yes	10.3/8.8	2	5.8
[15]	1.03 × 1.03 × 0.023	No	C-shaped aperture	No	28.2/20.3	1	5.25
[20]	0.944 × 0.944 × 0.041	No	Orthogonal feedline	Yes	9.3/6.2	2	7
[21]	0.2 × 0.2 × 0.042	No	Radiator perturbation	No	1.8/6 and 2.3/2.5	4	0/1.5
[23]	0.52 × 0.52 × 0.067	No	Truncated radiator/U shaped resonator	No	15.2/8.2	2 and 1 air gap layer	7
[25]	0.62 × 0.62 × 0.29	No	Truncated radiator	No	59.0/51.8	4 and 3 layers of air gap	8.54
[27]	0.726 × 0.726 × 0.34	No	Radiator perturbation	No	58.3/46.9	3 and 2 layers of air gap	7.6
[28]	0.69 × 0.69 × 0.048	Yes	Orthogonal resonators using second-order	No	4.7/3.2	2	6.2
This work	0.775 × 0.775 × 0.048	Yes	Orthogonal resonators using third-order	Yes	9.04/10.1	2	7.6

measured 3 dB beamwidths are 84° and 77.5° at the corresponding frequencies.

Table 1 shows the comparison of the proposed design with the previous studies in terms of the method and the ARBW results. It is known that orthogonal feedlines can achieve a wider ARBW than the single feedline filtering antenna. The two orthogonal feedlines were implemented in Ref. [10, 20] using two substrate layers, while our proposed design, utilizing a single feedline, achieves a wider ARBW compared to those two designs. The circular radiator with C-shaped perturbation in Ref. [15] achieves a wide ARBW; however, the CP generation and filtering functions were designed separately, and the perturbation results in a lower gain than our design. Reference [21] employs an omnidirectional antenna, which should theoretically have a wider bandwidth. However, despite emphasizing its dual-band performance, the ARBW, IBw, and gain results fall below those of our design. In Ref. [23], using coaxial probe and foam between the two substrates layer make the design

complicated, especially if we need to design an array antenna, though it has wider IBw, the ARBW is narrower than our design. The widest impedance and ARBWs in this comparison are reported in Refs. [25] and [27], the design using multiple layers and air gap layers to achieve a wider ARBW, however the IBw is wider than ARBW, while in our design, the ARBW is wider than the IBw. Additionally, the use of multiple substrate layers with air gaps in between results in a high-profile and complex structure. The proposed single-feed design achieves a competitive ARBW while maintaining significantly reduced structural complexity compared to multi-layer CP filtering antennas. Our proposed design demonstrates an ARBW that is nearly 300% wider and an IBw that is almost 200% wider than the previous study in Ref. [28], which also uses a single $\lambda/4$ resonator. However, the report lacked both a phase shifter and higher-order filter analysis. The improvement in the proposed design is achieved by incorporating an additional resonator.

Conclusion

In this study, a wide ARBw filtering antenna using orthogonal resonators has been designed and measured. The orthogonal resonators are modelled with a rectangular radiator and two quarter-wavelength resonators integrated through a third-order filter design. The CP filtering antenna is determined by three key factors: the ratio of length to width of the rectangular radiator, the gap between the radiator and the resonator, and the gap between the two resonators. The realized model results in an 10.1% ARBw and 9.04% IBw, complemented by a gain exhibiting filtering characteristics. The proposed design verifies that using the orthogonal resonators method with a higher order filter for integrating a filter and an antenna can effectively achieve a wider ARBw CP filtering antenna using only a single feed line. Future work will primarily focus on the application of the proposed CP filtering antenna in energy harvesting systems.

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Competing interests. The authors declare none.

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