



OPEN Design of a novel filtering antenna using waveguide with enhanced stopband attenuation

Lubina Iram¹, Nouf Abd Elmunim², Abdul Aziz Memon¹, Waleed Tariq Sethi³, Harris Sarfraz⁴, Bilal Tariq Malik⁵, Xue Jun Li⁴, Shahid Khan⁶ & Mariana Dalarsson⁷✉

This work details a novel type of waveguide filtering antenna (filtenna) with a two-post design for use in wireless satellite communications and radar systems. The proposed design incorporates the filtering and radiating capabilities in one small, rectangular, rigid system to overcome the constraint of conventional design methods which demand distinct filtering and radiating elements. The design uses a rectangular waveguide (WR-90) with two metallic posts, in two configurations inline and rotated offset which allows accurate control of the transmission zeros (TZs), high quality factor and stopband attenuation. This novel approach eliminates the requirement for the usual coupling irises to control bandwidth. The inline filtenna has a filtering bandwidth of 1.02 GHz at the X-band and a radiation bandwidth of 0.20 GHz. In contrast, the rotated-offset filtenna has been demonstrated at 9 GHz to 9.5 GHz, with a radiation bandwidth of 0.5 GHz. The simulated and measured results are in good agreement, with maximum gains of 7.05 dBi in the inline and 6.15 dBi in the rotated-offset orientations, and overall efficiency greater than 90 %. The inline filter has a reflection coefficient of less than -19 dB over the 10.6 GHz to 11.25 GHz range, which guarantees a good impedance match. The rotated-offset filtenna has wideband characteristics whose impedance is closely matched in the 8.92 GHz to 9.5 GHz operating band. The final design combines filtering and radiation capabilities in a compact form, increasing frequency selectivity and efficiency. These results emphasize the suitability of the suggested design for radar and satellite systems requiring small size and high-performance RF components.

Rapid technological advances in satellite, radar, and radio frequency (RF) systems have spurred the development of RF front-end components that improve efficiency, compactness, integration, and performance. Traditionally, antennas and filters utilize separate transmission (Tx) and reception (Rx) units, connected via a 50 Ω interconnect¹. However, this arrangement can increase circuit complexity, physical size, and insertion loss, particularly near band edges, leading to signal attenuation².

Waveguide filtennas are considered highly suitable for high-power applications compared with conventional microstrip based structures. In addition, waveguide structures offer lower losses and higher selectivity than their microstrip counterparts. Recent research has demonstrated the advantages of integrating filtering and antenna functions into a single unit, thereby simplifying the design, reducing the size, and improving overall performance. Multifunctional filtering antennas provide compact solutions for miniaturized wireless systems by addressing mutual-coupling issues and enhancing frequency selectivity^{3, 4}. Moreover, frequency-variant coupling (FVC) has been explored for designing compact waveguide filters with transmission zeros (TZs), particularly for narrowband applications with bandwidths below 10 %⁵⁻⁷. However, the direct connection of filters and antennas can increase the overall circuit size, which may lower radiation efficiency and gain, especially in broadband designs with multiple resonators⁸.

Various studies have also investigated different waveguide filtering antenna structures⁹. For instance, the sand casting method is used to produce an open ended waveguide antenna fed by a coaxial probe; the main

¹Department of Electrical Engineering, Sukkur IBA University, Sukkur 65200, Pakistan. ²Department of Electrical Engineering, College of Engineering, Princess Nourah bint Abdulrahman University, P.O. Box 84428, Riyadh 11671, Saudi Arabia. ³Ghulam Ishaq Khan Institute of Engineering Science and Technology, Topi, Swabi, Pakistan. ⁴Department of Electrical and Electronic Engineering, Auckland University of Technology, Auckland 1010, New Zealand. ⁵Department of Electrical Engineering, COMSATS University Islamabad, Islamabad, Pakistan. ⁶Department of Electronics, Telecommunication and Informatics, Gdansk University of Technology, Gdansk, Poland. ⁷School of Electrical Engineering and Computer Science, KTH Royal Institute of Technology, SE 100-44, Stockholm, Sweden. ✉email: mardal@kth.se

focus is to develop an antenna that not only meets the filtering needs but also has a single design that reduces assembly complexity¹⁰. An advanced X-band waveguide filtering antenna array makes use of a non-uniform feed structure to improve its performance, emphasizing a design methodology aimed at improved impedance matching and consistent radiation characteristics throughout the array. The non-uniform feeding technique allows improved control over the amplitude and phase distribution, leading to increased directivity and gain². A comprehensive classification of various filtenna structures, highlighting their design approaches and operational principles, has also been detailed in the literature¹¹. A novel waveguide antenna incorporating inline bandpass filtering, two TZs and an all-resonator design has been reported to provide improved filtering operation without compromising the effective radiation properties¹². In addition, a small wideband waveguide filtering antenna with a TZ has been designed to enhance filtering characteristics, focusing on a wide operational frequency and small dimensions for contemporary communication systems¹³. The effective bandwidth can be further increased by placing metal posts into the cavity resonator to tune the resonant frequency¹⁴.

More recent innovations in filtering antenna technology, such as resonant iris mode, dual-mode cavities, and cascaded resonators, have become much more efficient, small, and selective, resulting in improved bandwidth, gain, and filtering capacity^{15, 16}. Additionally, a patch antenna with a filtering feed structure, applying cross-coupling and mixed electric/magnetic coupling, has been proposed to eliminate cross-polarization¹⁷. Another study examined a full-metal cavity antenna with slots and flexible metal posts to perturb modes and selection frequency¹⁸. The majority of current published designs are overly complex, oversized, and have limited performance. FVC based waveguide filtering antennas have been studied in high-power applications, where lightweight, compact designs are provided while still ensuring strong performance¹⁹. In²⁰, the authors proposed a filtering antenna using coaxial cavity resonators to provide a tunable bandwidth and resonant frequency and a tunable filter/antenna integration with bandwidth control was demonstrated in²¹, while in²² resonance windows are used to enhance the bandwidth. Highly selective filtering slot antennas based on stub-loaded resonators have also been reported²³. However, there is very little research on slot antennas in conjunction with waveguide bandpass filters, which limits further understanding of their potential advantages. Advanced filtering techniques, such as incorporating slot lines and shorting pins on microstrip patches, have been proposed. Another approach involves using different material properties to fine-tune resonant frequencies by injecting materials into the slot and patch lines^{24, 25}. Beyond conventional cavity- and post-based approaches, several planar and miniaturization techniques have also been explored for compact, highly selective filters and filtering antennas. For instance, stepped impedance resonators (SIRs) and fractal geometries have been employed to reduce size and suppress spurious responses while sharpening the frequency response^{26–28}. Substrate integrated waveguide (SIW) technology has likewise been widely adopted at millimetre-wave frequencies, where slot- and cavity-based resonators provide filtering selectivity within a single dielectric layer, as in the E-shaped slot filtenna of²⁹; a comprehensive review of SIW filter miniaturisation techniques is given in³⁰. Within the same SIW framework, defected-ground resonators have been used to place transmission zeros and compress the cavity footprint in dual-band designs³¹, while composite-mode coplanar structures have been applied where a large frequency ratio between bands must be maintained without harmonic leakage between them³². Although effective in planar technologies, such approaches generally exhibit higher conductor losses and lower power handling capability than the metallic waveguide implementation adopted in this work, which motivates the post-based cavity filtenna proposed here.

This work presents a compact dual-post filtering antenna that employs post rotation and post offset movements. It uses WR-90 as a coupling element and, instead of conventional coupling irises, half-length cylindrical posts, which provide better bandwidth control and a third-order bandpass response with four TZs. The simulation results show that control over the TZs and varying the post-rotation and offset allow fine control of the TZs on either side of the passband. This results in improved performance in the stopband attenuation and a more compact design. The main novel aspect of this research is that the typical coupling irises have been replaced by rotated and displaced half posts, which not only serve the purpose of coupling, resonating, and radiation but also independently control four TZs in a single WR-90 cavity, thereby resulting in the bandpass filtenna of third order. Throughout this work, the antenna performance is reported in terms of the realized gain expressed in dBi, following the convention adopted for waveguide filtennas in²⁰.

The key contributions of the proposed design are as follows.

- The design integrates simultaneous filtering and radiation characteristics in a single WR-90 cavity, reducing system complexity and overall size.
- Half length rotated and offset posts are used as the coupling mechanism in place of conventional coupling irises, providing improved frequency selectivity and stopband attenuation through precise TZ control, in contrast to the inverter and cross coupling based waveguide approaches in^{5, 6, 33}.
- Detailed equivalent circuit models for both the inline and rotated-offset configurations are developed to provide theoretical insight into the coupling and TZ behaviour.
- The prototype is fabricated and tested to obtain the measured responses of the filtenna using a vector network analyzer (VNA), and the radiation patterns using an anechoic chamber.

The rest of the paper is structured as follows. Section 2 gives an overview of the proposed design. Section 3 presents the development of the design and its working principles. The parametric study is outlined in Section 4. The simulated and measured results are provided in Section 5, the benchmarking against recent works is presented in Section 6, and the work is summarized in Section 7.

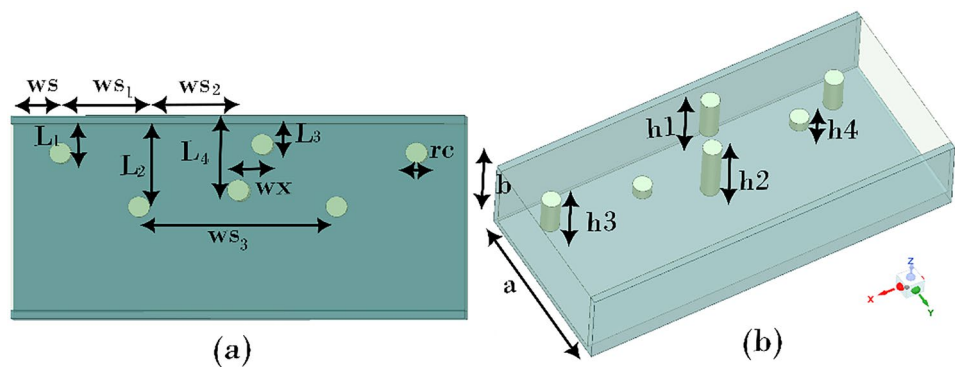


Fig. 1. Overview of the proposed dual-post filtenna: (a) top view showing the inter-post spacings W_s , W_{s1} , W_{s2} , and W_{s3} , the longitudinal post positions L_1 – L_4 , the post radius R_c , and the aperture width W_x ; and (b) 3D view showing the four cylindrical post heights h_1 – h_4 inside the WR-90 rectangular waveguide of width (a) and height (b).

Parameter	Dim. (mm)	Parameter	Dim. (mm)
a	22.86	b	10.16
h_1	6.847	L_1	3.01
h_2	9	L_2	11.43
h_3	5.87	L_3	3.23
h_4	0.98	L_4	4.35
R_c	1.25	$W_s + W_{s1} + W_{s2}$	5.03 + 15.7 + 13.11
W_x	3.3	W_{s3}	20.9

Table 1. Optimized dimensions of the proposed ROSDP filtenna design parameters.

Proposed design overview

This section presents an overview of the proposed filtenna design, focusing on its technical development and the basic principles behind its design and performance. Fig. 1(a) and (b) show the top and 3D views of the proposed design, respectively. W_s is the distance between the waveguide terminal and the pole. W_{s1} and W_{s2} are the distances between the first and second poles, and the second and third poles, respectively. W_{s3} is the distance between two posts, while L_1 , L_2 , L_3 , and L_4 are the distances of the posts from the waveguide wall. R_c is the radius of both the posts and poles. The design process utilizes EM theory and formulas to calculate the initial size and layout of the components in the WR-90 rectangular waveguide. h_1 , h_2 , h_3 , and h_4 are the heights of the cylindrical posts, while a and b are the width and height of the waveguide, respectively. The optimal values of the design parameters are shown in Table 1, which comprises the height, length, distance, and offset of the aperture. This configuration of the design parameters provides good filtering and radiation characteristics for the filtenna.

Design methodology

Design equation, working principle, and evolution steps

The mathematical formulation is used to justify the filtenna design described in this paper, giving rise to the initial dimensions^{12,34}. The design progresses by following the physical hierarchy of the structure. It starts from the WR-90 waveguide basics, at which the TE_{10} is the dominating mode and dictates the operation bandwidth and the cutoff properties for each new stage. Then a post-coupling stage is introduced since the half-length posts affect the TE_{10} and generate an evanescent coupling of the electromagnetic fields needed to transfer energy in the absence of irises for coupling. This coupling arrangement is arranged in one stage of inverter, which changes the local field disturbance into a phase relationship. Increasing the number and spacing of the posts turns the inverter into a cavity filter, since the bounded field regions between the posts act as resonators that establish the passband and the TZs. Finally, radiating posts are added to the top surface so that the aperture couples energy to free space, integrating the antenna and completing the filtenna in a single structure. Cylindrical posts were adopted throughout this work for reasons of both electromagnetic behaviour and realisability. For an inductive obstacle in a rectangular waveguide, the transverse dimension of the post governs the shunt susceptance and hence the coupling strength, whereas the longitudinal dimension governs the effective electrical length of the discontinuity and hence the frequency dispersion of the scattered field. A circular post has a single radius and therefore couples these two effects, while an elliptical cross section separates them through its axial ratio and introduces an orientation angle as a further variable, raising the number of independent parameters per post from two to four. Non-circular geometries have been shown to exploit this freedom, as in the modified T-shaped post

inverter of³³ and the cross-shaped frequency dependent coupling structure of⁶. Two considerations nevertheless favour the circular geometry here. First, the rotated-offset configuration already employs post rotation and lateral offset as independent variables, so that the orientation of an elliptical post would be largely redundant with them; the four transmission zeros of the present design are obtained without recourse to any further degree of freedom. Second, circular posts are realised by drilling to a standard tolerance, whereas elliptical cross sections require milling or electrical discharge machining. As the measured band-edge and zero shifts reported in Section 6 indicate, the TZ positions of this topology are already sensitive to post dimensions at the level of tens of micrometres, and a geometry demanding tighter tolerances would reduce the reproducibility of the fabricated response. A parametric study of elliptical and other non circular cross sections, carried out under an equal-area constraint so that the effect of shape is isolated from that of blockage, therefore remains a well defined direction for subsequent work.

Fig. 2(a) represents the first stage, which consists of two half-length cylindrical posts of a perfect electric conductor (PEC) placed in the WR-90 waveguide. These posts, whose heights are h_1 and h_2 , serve as coupling elements. The coupling posts are placed at multiples of $\lambda_g/4$ to achieve the strongest possible coupling, and their locations along the waveguide, L_1 and L_2 , are calculated with respect to a reference plane and carefully selected and control the phase of the transmitted signal. The posts distort the primary TE_{10} mode, forming evanescent fields that facilitate the inverter functionality. This interaction is the foundation of the TZ generation of the ROSDP configuration. Following the mode matching formulation, the reflection coefficient Γ of a single post depends on its transverse position x_p through a $\sin^2(\pi x_p/a)$ dependence and on its longitudinal position z_p through the phase term $e^{-j2k_z z_p}$, where a is the WR-90 waveguide width (22.86 mm), k_0 is the free space wavenumber, and k_z is the propagation constant of the TE_{10} mode. The coupling is therefore strongest when the post is near the waveguide centre ($x_p = a/2$) and is tuned by its longitudinal placement. The second stage, illustrated in Fig. 2(b), improves the inverter by adding a traditional rectangular window, located symmetrically at a distance d_3 to the coupling window externally. This coupling window controls the bandwidth of the waveguide. Fig. 2(c) shows the filter with two additional posts as a substitute for windows, which further enhance the bandwidth while making the filter more selective by adding more TZs. The cavity resonates because the PEC boundary conditions

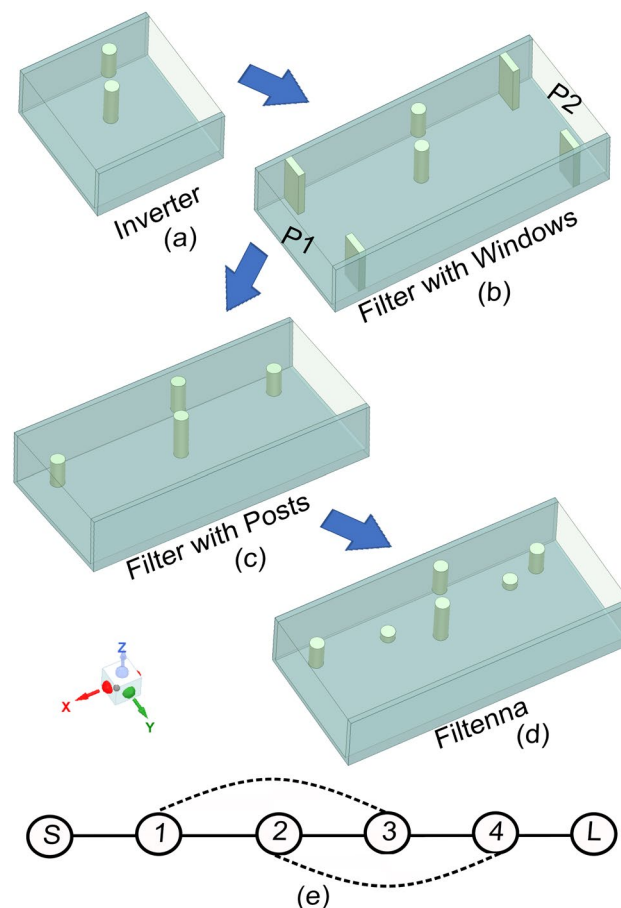


Fig. 2. Evolution of the proposed structure through its four design stages: (a) the single-section inverter formed by two half-length coupling posts; (b) the filter obtained by adding a rectangular coupling window; (c) the filter in which the windows are replaced by additional posts to sharpen selectivity and add TZs; (d) the complete filtenna with the two radiating posts on the top wall; and (e) the corresponding source-load coupling-routing topology.

stimulate higher-order TE modes and inhibit TM modes. The resonant frequency of the dominant TE₁₀₁ cavity mode is determined by^{12,14}

$$f_{101} = \frac{c}{2} \sqrt{\left(\frac{1}{a}\right)^2 + \left(\frac{1}{d}\right)^2}, \quad (1)$$

where $d = W_{s3} = 20.9$ mm is the separation between the coupling posts that bound the cavity. With $a = 22.86$ mm this yields 9.72 GHz. The intrusion of the offset post of height h_3 loads the cavity capacitively through the residual gap $(b - h_3) = 4.29$ mm and lowers the resonance towards the design value of 9.3 GHz; this shift is accounted for in the full-wave optimisation rather than in the closed-form estimate. The separation d_3 controls the coupling bandwidth through the coupling coefficient k^5 . The coupling scales with the square of the post-intrusion ratio (h_3/b) and with the ratio of the cutoff to the operating wavenumber, so that an approximate design estimate, calibrated against the full wave parametric simulations, is

$$k \approx \left(\frac{h_3}{b}\right)^2 \frac{(\pi/a)^2}{k_0^2} = \left(\frac{h_3}{b}\right)^2 \left(\frac{f_c}{f_0}\right)^2, \quad (2)$$

where $(\pi/a)/k_0 = f_c/f_0$ is the normalized cutoff. This provides an initial value of the coupling that is subsequently refined so that the filter stage matches the cavity resonance to the desired response.

In the last step, filtering and radiation are incorporated by adding two PEC posts of height h_4 to the top surface of the waveguide, as illustrated in Fig. 2(d). At longitudinal offsets L_4 , these posts adjust the effective length of the waveguide aperture (width W_x). The filtenna balances the radiative admittance of the apertures with the reactive components of the cavities to achieve impedance matching and efficient radiation. The design method integrates analytical modeling and full-wave simulations to develop a compact filtenna with mode control and low loss. In addition, electromagnetic simulations and analytical models guide the design procedure to optimize the position of the posts and reduce losses due to mode conversion. The WR-90 waveguide supports the TE₁₀ mode up to a certain frequency given by¹²

$$f_c = \frac{c}{2a}. \quad (3)$$

The initial set of parameters, such as, waveguide height b , aperture length, and post parameters are obtained from Eqs. (1)–(3) and then refined through a number of simulations and parametric analyses to ensure outstanding performance in the X band. The details are presented in Table 1, which provides a good filtenna design with proven filtering and radiating capabilities.

In this section, IDP and ROSDP structures are considered individually: Fig. 3(a) presents IDP and Fig. 3(b) ROSDP. Eigenmode analysis is used in the preliminary design phase to evaluate the variations in resonant frequencies with respect to the position of posts. A series of parametric simulations is carried out in order to optimize the band width and obtain required TZs; see Fig. 3. The coupling process takes place due to the introduction of dual cylindrical metallic posts within the WR-90 waveguide illustrated in Fig. 2(d). These posts serve as field perturbation elements that introduce controlled coupling.

The scattering parameters over the given frequency range, shown in Fig. 3, provide key insights into system performance. In particular, S_{11} represents the reflection coefficient that gives the amount of reflected power at port 1, while S_{21} is the transmission coefficient, representing the amount of power transferred from port 1 to port 2. The response of the IDP and ROSDP inverters is illustrated in Fig. 3(a) and (b), respectively. The deep nulls in the S_{21} graph indicate the TZs as the result of mode coupling generated by the post offset. The S_{11} minima show that the impedance matching is good around the passband center.

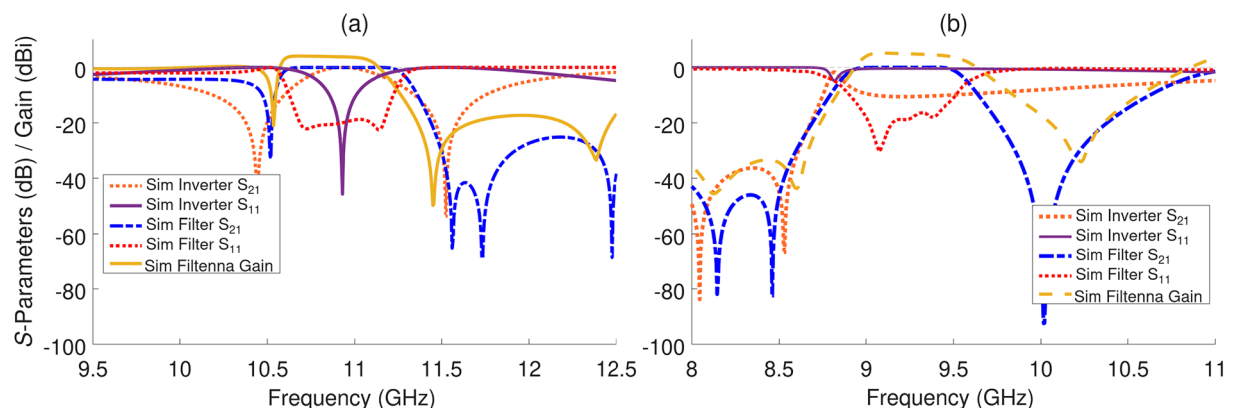


Fig. 3. Simulated response at the three evolution stages (inverter, filter, filtenna) for (a) the *inline dual-post* (IDP) configuration and (b) the *rotated-offset dual-post* (ROSDP) configuration.

The transformation of the inverter into a cavity filter is achieved by increasing the number of posts and adjusting the distance between them to create bounded regions of fields in the WR-90 waveguide. They serve as resonant cavities, with the distance adjusted according to the half-wavelength resonance, which allows for transmission poles creation in Fig. 3(a). The inner cavity posts give better frequency selectivity and reject the out-of-band signals. The response of the ROSDP filter configuration, which involves rotated and offset posts, is demonstrated in Fig. 3(b). In each panel of Fig. 3, the dotted red curves denote the reflection coefficient $|S_{11}|$ and the solid/dashed blue curves denote the transmission coefficient $|S_{21}|$, both read on the left S-parameters (dB) axis, whereas the solid yellow curve is the realized filtenna gain read on the right Gain (dBi) axis. The latter is an active (radiated) quantity rather than a passive scattering parameter, which is why it exceeds 0 dB within the passband. The stopband rejection is improved by the introduction of asymmetric coupling due to the rotated-offset posts, which introduces additional TZs on either side of the passband. These half cylindrical posts improve mode coupling and enable precise positioning of the TZs.

The next design step integrates filtering and radiating functionalities, resulting in a compact filtenna structure, as shown in Fig. 2(d). Here, one rectangular port serves as the excitation input, while the other functions as a radiating output. To fine-tune performance, two additional posts are introduced, enabling precise control over the overall response and limiting the cavity length, as illustrated in Fig. 5.

Quality-factor analysis of the proposed design

The Q factor is the ratio of energy stored in the resonator to the energy lost and is one of the key parameters in the performance of the filtenna¹². The external quality factor Q_e can be computed using the driven modal analysis technique based on the frequency shift introduced by the post insertion at the new offset position. Parametric simulations have been carried out for different post heights and offsets in order to obtain optimized Q_e with the desired bandwidth and zeros, as shown in Fig. 4(a) and (b). The Q-factor can be obtained from Eqs. (4) and (5)^{20,22}:

$$Q_e = \left(\frac{f_0}{\Delta f} \right) \frac{1}{1 - |S_{21}(f_0)|^2}, \quad (4)$$

$$Q_r = \left(\frac{1}{Q_l} - \frac{1}{Q_e} \right)^{-1}. \quad (5)$$

Here Q_l is the loaded quality factor of the resonator, f_0 is the center frequency, Δf is the -3 dB bandwidth, and $S_{21}(f_0)$ is the transmission coefficient at f_0 .

The radiation quality factor Q_r is calculated using the values of $S_{21}(f_0)$, Δf , and different center frequencies f_0 . Two major parameters for design in the proposed waveguide filter, which affect both Q_e and Q_r , are the height of the offset post (h_3) and the lateral offset distance (L_1). Fig. 4(b) depicts the variations of Q_e and Q_r in two offset cases ($L_1 = 2.22$ mm and $L_1 = 2.25$ mm) at 9.3 GHz for different values of h_3 . Both Q_e and Q_r become smaller as h_3 becomes larger, indicating better coupling and enhanced radiation efficiency. An important indicator for tuning the radiation bandwidth and matching is the rapid reduction in Q_r with the increase of the post height, indicating larger losses due to the radiation.

Parametric analysis

A parametric sweep was performed to investigate the effects of structural dimension variations on the TZs and pole locations for the inline and rotated-offset configurations. This includes the impact of offset height in the inline configuration, as well as the effects of offset distance and post height in the ROSDP.

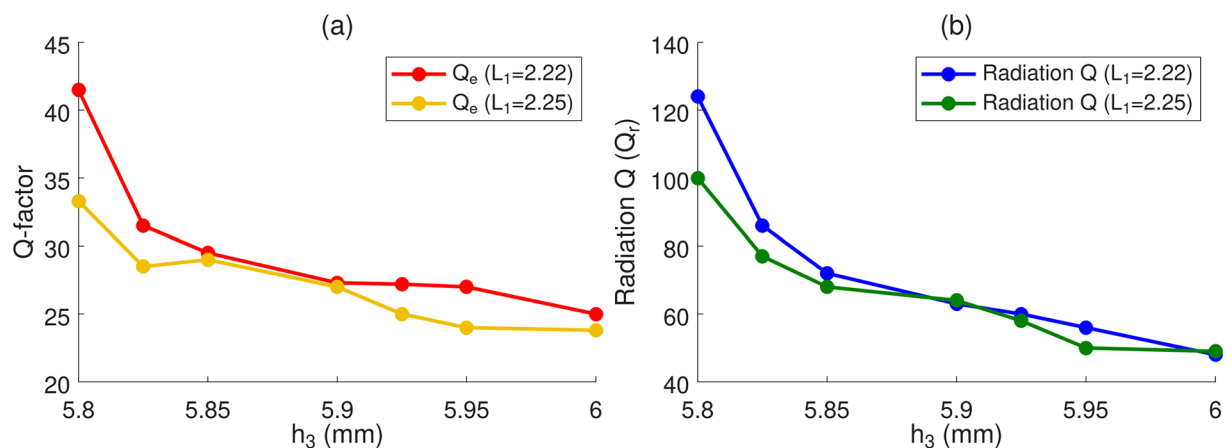


Fig. 4. Variation of (a) the external quality factor Q_e and (b) the radiation quality factor Q_r with the offset-post height h_3 , for two lateral offsets $L_1 = 2.22$ mm and $L_1 = 2.25$ mm at 9.3 GHz. Both factors decrease as h_3 increases, indicating stronger coupling and enhanced radiation.

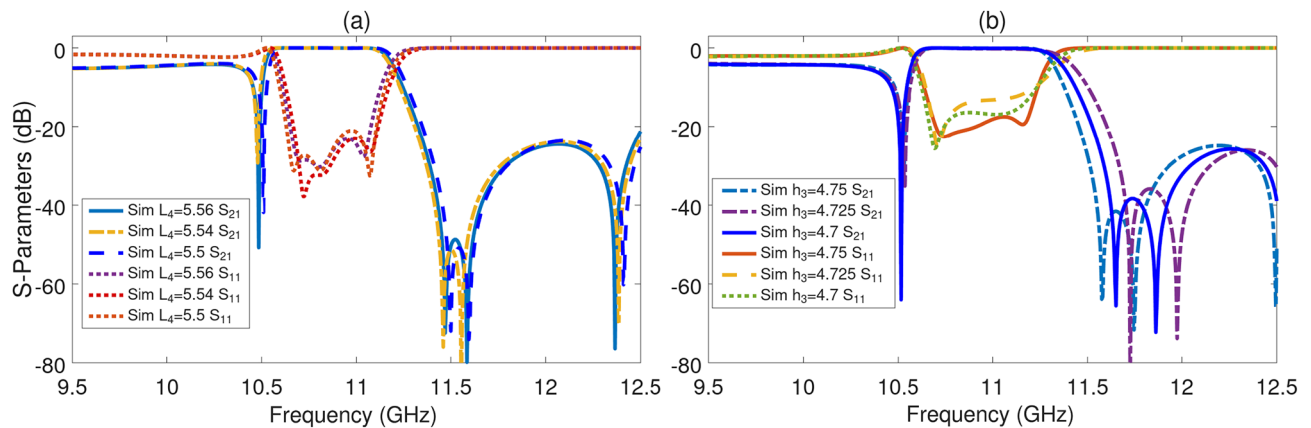


Fig. 5. Parametric study of the *inline* (IDP) configuration: (a) effect of the post offset distance L_4 and (b) effect of the post height h_3 on the simulated S -parameters. Varying L_4 mainly shifts the pole frequencies and bandwidth, whereas h_3 shifts the transmission zero toward higher frequencies.

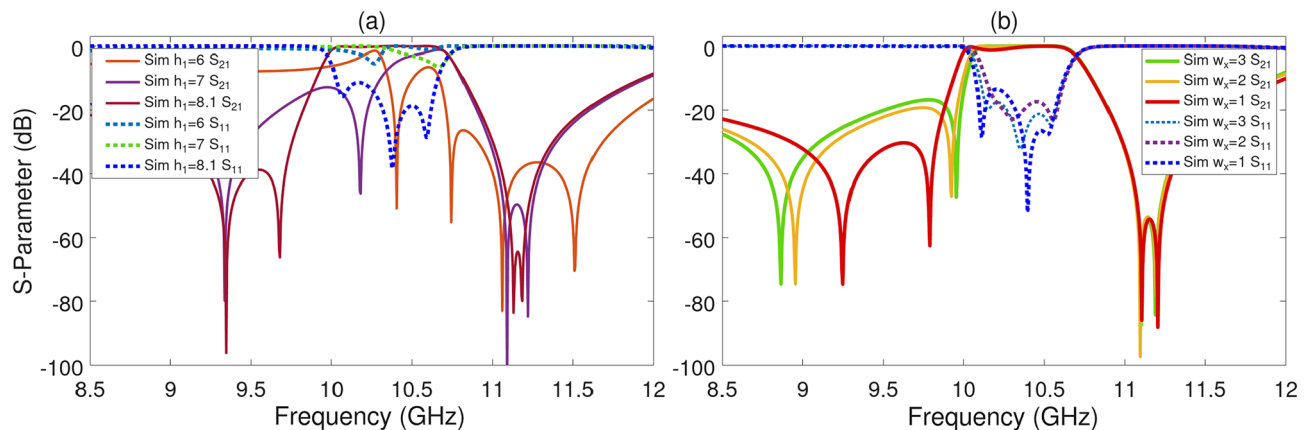


Fig. 6. Parametric study of the *rotated-offset* (ROSDP) configuration: (a) effect of the post height h_1 and (b) effect of the transmission-line section length W_x on the simulated S -parameters. Both parameters relocate the TZs and thereby control the stopband selectivity.

Effect of height on the inline configuration

The sweeps of this sub-section were carried out during the design of the inline (IDP) configuration, whose passband is centred near 10.9 GHz, so that the values of L_4 and h_3 used here differ from the optimized ROSDP dimensions listed in Table 1. The trends identified, namely the control of the pole frequencies by L_4 and of the transmission-zero position by h_3 , are common to both configurations and guided the selection of the final ROSDP dimensions. The analysis of the S -parameters reveals a significant influence of varying the post height and the offset distance on the filters frequency response. As shown in Fig. 5(a), simultaneously varying the post offset distance L_4 has a minimal impact on the TZ positions but primarily influences the bandwidth and pole frequencies. As shown in Fig. 5(b), decreasing the height h_3 progressively shifts the TZ toward higher frequencies, underscoring the designs sensitivity to this parameter. Adjustments to the offset-post height h_3 therefore shift the response across the spectrum, particularly affecting the upper band edge and the response bandwidth. This detailed parametric study underscores the critical need for precise control over these dimensions to effectively optimize the devices frequency response.

Effect of offset distance and post height on ROSDP

The effect of the offset distance and the post height on the return loss and stopband performance of the ROSDP is illustrated in Fig. 6. The effect of the transmission line segment length W_x is studied by using the parametric sweep technique that reveals that when the offset distance increases, the TZs move to the low-frequency side but increase the bandwidth as well. Beyond a critical gap value, the TZs shift toward higher frequencies. The interaction between h_1 and W_x is evident in Fig. 6(a) and (b), where controlling these parameters clearly shifts the TZ to the upper side, enhancing frequency selectivity and stopband performance by optimizing the relative positioning of poles and TZs.

Equivalent-circuit model

As explained in⁵, the dual-post structure can be represented using a nonlinear frequency-variant coupling (NFVC) inverter, in which the strength of coupling changes with frequency. The inverter is characterized by shunt connected series resonators, and the location of the transmission pole relative to the TZs is determined by the sign and magnitude of the mutual inductance (L_M). When posts are placed on the same broad wall, a negative L_M is obtained; when on opposite broad walls, a positive L_M is obtained. Mutual coupling naturally occurs when posts are closely spaced. The addition of a transmission line section between the posts, represented by a characteristic impedance (Z_L) and electrical length θ , improves the ability to control the placement of L_M and the TZ.

The lumped-element modeling of the IDP and ROSDP filter configurations is given in Fig. 7(a)–(c). It takes into account the placement of the posts, the material, and the excitation of the waveguide-fed system. Each metal post is modeled as a series of inductive–capacitive (LC) resonators connected in parallel^{22,35}. A straight section of waveguide surrounded by discontinuities such as posts acts as a resonator and can be represented using lumped inductance (L) and capacitance (C) elements in a similar way. In Fig. 7(a), the posts serve as external coupling components, represented by series LC networks at the input and output, which help regulate the bandwidth. In Fig. 7(b), the placement of the TZ is influenced by L_M , introduced by the interaction between the rotated and offset posts. An increase in positive L_M shifts the TZs above the passband, improving upper-stopband attenuation. Conversely, a decrease in negative L_M shifts the TZs below the passband, resulting in better lower-stopband suppression and a steeper roll-off. L_M is adjusted by changing the spacing and position of the posts along the waveguide, as evidenced by the changing notches in the S_{21} response. The S-parameter response is accurately predicted by both models, demonstrating strong agreement between EM simulations and circuit based simulations.

The lumped model is more than just a simple illustration in which each shunt series-LC branch resonates at $f = 1/(2\pi\sqrt{LC})$, and these resonances map directly onto the observed response. The element values quoted in Fig. 7 are not obtained by fitting the complete circuit to the full-wave response. They are extracted stage by stage from the simulated S-parameters of the structures in Fig. 2, as follows. The scattering parameters of each stage are first referred to the post plane using the TE₁₀ propagation constant⁵,

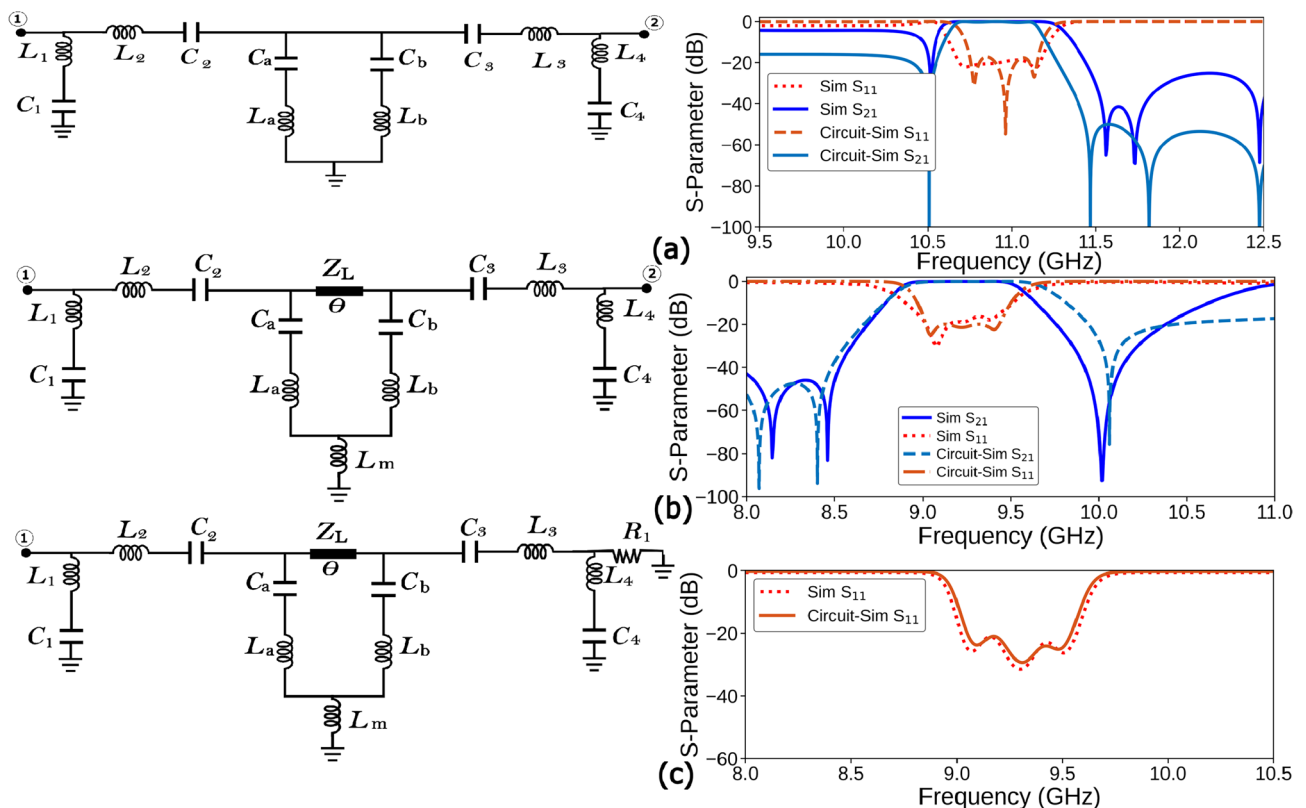


Fig. 7. Lumped circuit model (a) Inline (IDP) configuration: $L_1 = 1.538$ nH, $C_1 = 0.116$ pF, $L_2 = 1.929$ nH, $C_2 = 0.128$ pF, $C_a = 0.137$ pF, $L_a = 1.639$ nH, $C_b = 0.247$ pF, $L_b = 0.784$ nH, $C_3 = 1.300$ pF, $L_3 = 0.744$ nH, $C_4 = 0.174$ pF, $L_4 = 0.913$ nH. (b) ROSDP configuration: $L_1 = 3.109$ nH, $C_1 = 0.116$ pF, $L_2 = 3$ nH, $C_2 = 0.087$ pF, $C_a = 0.194$ pF, $L_a = 1.329$ nH, $Z_L = 3.48 \Omega$, $\theta = 18^\circ$, $C_b = 1.107$ pF, $L_b = 0.359$ nH, $L_M = -0.021$ nH, $C_3 = 0.155$ pF, $L_3 = 2.277$ nH, $C_4 = 0.318$ pF, $L_4 = 0.464$ nH. (c) Filtenna values: $L_1 = 3.109$ nH, $C_1 = 0.120$ pF, $L_2 = 3$ nH, $C_2 = 0.087$ pF, $C_a = 0.190$ pF, $L_a = 1.333$ nH, $Z_L = 2.98 \Omega$, $\theta = 16.8^\circ$, $C_b = 1.115$ pF, $L_b = 0.365$ nH, $L_M = -0.017$ nH, $C_3 = 0.154$ pF, $L_3 = 2.273$ nH, $C_4 = 0.296$ pF, $L_4 = 0.488$ nH, $R = 40 \Omega$.

$$S'_{ij} = S_{ij} e^{+j2k_z \ell}, \quad k_z = \sqrt{k_0^2 - (\pi/a)^2}, \quad (6)$$

where ℓ is the offset of the reference plane. A single post acts as a shunt discontinuity, for which the normalised admittance and the corresponding branch reactance are³³

$$\frac{Y}{Y_0} = 2 \left(\frac{1}{S'_{21}} - 1 \right), \quad X(\omega) = \text{Im} \left\{ \frac{1}{Y} \right\} = \omega L - \frac{1}{\omega C}. \quad (7)$$

The series resonance ω_0 of the branch is the zero crossing of $X(\omega)$, and since $dX/d\omega|_{\omega_0} = 2L$, the two elements follow without further adjustment as¹⁶

$$L = \frac{1}{2} \frac{dX}{d\omega} \Big|_{\omega_0}, \quad C = \frac{1}{\omega_0^2 L}. \quad (8)$$

The frequency-variant coupling is then obtained from the two-post structure as $L_M(\omega) = \text{Im}\{Z_{12}(\omega)\}/\omega$. A value of L_M that varies across the operating band, rather than a constant, is the defining signature of the nonlinear frequency-variant inverter and is what permits an N th-order structure to support $N + 1$ transmission zeros. The inter-post section is characterised over the same band by $\theta = -\angle S'_{21}$ together with the characteristic impedance Z_L of the de-embedded two-port, and the loss resistance follows from the unloaded quality factor of the lossy model as $R = \omega_0 L/Q_u$, consistent with (9).

The procedure is applied in the order of the evolution stages of Fig. 2: elements identified at one stage are held fixed while those introduced at the next are determined, so that each parameter is associated with a specific physical discontinuity rather than with a global fit. The values reported in Fig. 7 are the outcome of this sequence followed by a final refinement, and the resulting circuit and electromagnetic responses are overlaid in the same figure. For the inline configuration, the input/output branches (L_1 – C_1 and L_4 – C_4) are asynchronously tuned to place the TZs; together with the coupling elements, they produce the upper TZ at 12.475 GHz, in agreement with the full-wave electromagnetic simulation. The passband centre ($f_0 \approx 10.9$ GHz) corresponds to the geometric mean of the series- and shunt-arm resonances, confirming a generalized Chebyshev (elliptic) rather than an all-pole synchronous response. The remaining zeros are set jointly by the central C_a – L_a and C_b – L_b legs together with the mutual inductance L_M and the transmission-line section (Z_L , θ). This asynchronous tuning that precisely allows for the four TZs to be independently located around the passband. In order to take into consideration the losses involved, a resistance R is added to the filtenna circuit of Fig. 7(c) in order to model the conductor resistance, dielectric loss, and radiation loss. The magnitude of R follows from the unloaded quality factor of the lossy full-wave model as $R = \omega_0 L/Q_u$. Resistance R influences the quality factor of the resonators as^{5, 36}

$$Q = \frac{\omega_0 L}{R}. \quad (9)$$

This process ensures that the modeled insertion loss accurately reflects real-world losses due to conductor, dielectric, and radiation effects. R influences the Q -factor of the resonators, affecting insertion loss and bandwidth control.

It should be noted that, in the classical coupling-matrix formulation, the number of finite TZ of an N th-order filter is limited to N , since the zeros arise only from destructive interference between non-adjacent resonators. The nonlinear frequency-variant coupling employed here removes this restriction: because the coupling itself is dispersive, each NFVC inverter contributes an additional zero, so that the present third-order structure supports $N + 1 = 4$ TZs, consistent with the FVC filters reported in^{5, 33}. Of these four zeros, three fall within the frequency window of Fig. 10, while the fourth lies outside the plotted range.

Filtenna performance evaluation using full-wave simulation and measurements

This section validates the functionality of the proposed filtenna design through analysis of full-wave simulation and experimental results. The main performance indicators investigated in the operating range are reflection coefficients, current distributions, gain, efficiency, and radiation patterns.

Experimental setup

To validate the simulated results, the filtenna was fabricated and the ROSDP configuration was experimentally measured. The designed device was fabricated using a high-precision CNC milling machine to accurately produce the aluminum cavity and coupling posts. The prototype of this design is given in Fig. 8(a) and (b) along with labels of its components that help to compare it with simulation model in Fig. 2. The filtering part is the central cavity with two posts and antenna part is mounted on output port as shown in Fig. 8(c) and (d). Therefore both functions are done together i.e. filtering and radiation. The S parameter measurements (reflection and transmission coefficients) were done using Rohde & Schwarz ZVL VNA. The gain of antenna was measured in an anechoic chamber where there was no interference. Fig. 8(b) shows the prototype of ROSDP setup.

Scattering parameters

This part details the simulated and experimental scattering parameters of the IDP and ROSDP filtenna designs. The experiment was carried out by attaching the filtenna prototype to a pre-calibrated VNA. Fig. 9 illustrates the simulated reflection coefficient S_{11} , which is less than -19 dB in the 10.6 GHz to 11.25 GHz range, indicating

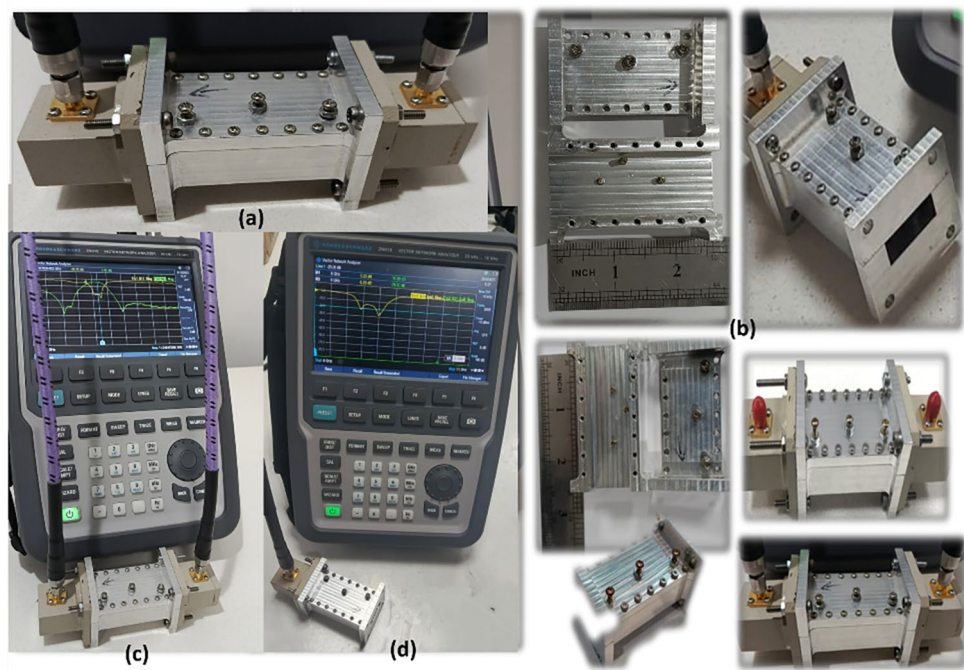


Fig. 8. Prototype of the fabricated ROSDP filtenna, fabricated from aluminum using high-accuracy CNC milling technology: **(a)** assembled ROSDP filter; **(b)** top view and bottom with dimensions and SMA connector; **(c)** S-parameter measurements for filter; and **(d)** the filtenna measurement using the same VNA.

Performance metric	Simulated	Measured
Operating band (GHz)	8.95–9.49	8.925–9.5
Bandwidth (MHz)	540	575
In-band return loss, $ S_{11} $ (dB)	−41.4	−29.8
at frequency (GHz)	9.40	9.08
Upper TZ, f_{TZ} (GHz)	10.015	9.98
TZ depth, $ S_{21} $ (dB)	−92.5	−44.6
Peak realized gain (dBi)	6.15	7.25
XPD at boresight (dB)	>38	17.5–23.3
XPD over main lobe, $\pm 60^\circ$ (dB)	>36	12.5–18.6
Upper-band roll-off (dB/GHz)	≈ 100	—
External / radiation Q (Q_e/Q_r)	30 / 73	—
Low-profile height (λ_0)	0.341	0.341

Table 2. Quantitative comparison of simulated and measured performance of the ROSDP filtenna.

excellent impedance matching and low signal reflection. S_{21} , the transmission/mutual coupling parameter, is a measure of insertion loss and port isolation. The minimal attenuation and efficient signal transmission are established by the low insertion loss, which ensures high port-to-port isolation and effective signal filtering. The simulated gain is constant at about 7.05 dBi across the operational band, indicating consistent radiation performance.

These results verify the effectiveness of the IDP filtenna in rejecting out-of-band signals attenuating frequencies lower than 10.5 GHz and higher than 11.5 GHz to 12.5 GHz as well as maintaining significant radiation in the 10.6 GHz to 11.25 GHz range. In general, Fig. 9 demonstrates that the filtenna has a 650 MHz passband, constant gain, and high filtering, making it an ideal candidate for high-performance communication systems capable of dealing with possible interferences. Fig. 10 represents the scatter parameter simulations and measurements for the proposed ROSDP system, thus verifying the efficiency of the proposed solution. The measured passband spans 8.925 GHz to 9.5 GHz, in close agreement with the simulated span of 8.95 GHz to 9.49 GHz (an edge shift of only about 25 MHz). Within the passband, the simulated return loss reaches −41.4 dB (at 9.4 GHz) and the measured $|S_{11}|$ reaches −29.8 dB (at 9.08 GHz), both confirming a good impedance match. The upper-band TZ, simulated at 10.015 GHz with a depth of −92.5 dB, appears in measurement at 9.98 GHz

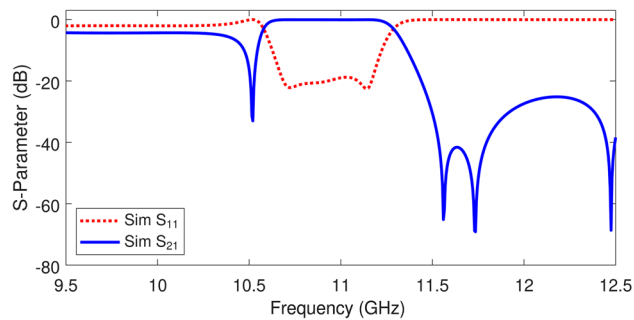


Fig. 9. Simulated response of the inline (IDP) filtenna, showing the reflection coefficient $|S_{11}|$, the transmission coefficient $|S_{21}|$ across the X-band, with the passband centred in the 10.6 GHz to 11.25 GHz range.

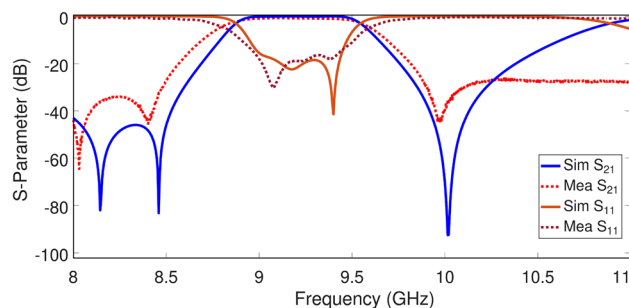


Fig. 10. S-parameters of simulated and measured ROSDP filter and filtenna, comparing the magnitude of S_{11} and S_{21} in the operating frequency band; the measured passband (8.925 GHz to 9.5 GHz) and the upper TZ match the simulation very well.

, a shift of only 34 MHz. These small deviations are consistent with fabrication tolerances and the SMA-to-waveguide transition. The origin of this difference can be quantified directly. Expressed as a linear reflection coefficient, the simulated in-band value of -41.4 dB corresponds to $|T| = 0.0085$ and the measured -29.8 dB to $|T| = 0.0324$, so that the two differ by only 0.024 in absolute magnitude. Because reflections combine vectorially rather than additively, a residual reflection of magnitude 0.024, that is -32.4 dB, at any point in the measurement path is by itself sufficient to account for the entire discrepancy. A commercial SMA-to-waveguide transition is typically matched to between -25 and -30 dB across an X-band octave, so the return loss of the assembled filtenna is bounded by the transitions rather than by the cavity response, however well the latter is matched.

This also accounts for the fact that the difference appears almost entirely in the depth of the reflection null and not in the passband edges or the transmission-zero frequency, which shift by only 25 MHz and 34 MHz respectively. A dissipative mechanism such as finite conductor conductivity broadens the response and raises the insertion loss; it does not selectively fill a reflection null, since that null arises from the cancellation of two reflected waves rather than from absorption. Finite conductivity is therefore a secondary contributor here, the dominant terms being the port transitions together with the machining tolerance of the posts, both of which perturb the cancellation condition directly.

The performance of the proposed filtenna is verified through comparison of measured and simulated scattering parameters shown in Fig. 10. There is a high level of agreement between these curves which implies that the impedance matching is achieved in the 9 GHz to 9.5 GHz band with the value of S_{11} being about -20 dB. The high return loss means the low reflection and high efficiency of the power transfer; hence the impedance of the filtenna is well matched in the required band. Furthermore, there is also a high level of agreement between measured and simulated scattering parameter S_{21} in the above range of frequencies. The low values of the scattering parameter S_{21} between 9 GHz to 9.5 GHz imply extremely low mutual coupling, providing the isolation of ports. This is an important property that helps to avoid inter-port interferences and provides stable filtering and radiating properties.

Surface current distribution

The surface current distributions of the IDP and ROSDP configurations exhibit distinct features related to their geometry and element arrangement. In the inline configuration of Fig. 11(a), the surface current is more linearly oriented and strongly concentrated around the excitation region and the closely spaced posts. This localized, high-density current indicates strong mutual coupling between the posts. Although it increases the local field strength, it also reduces the polarization purity and the isolation that are important for suppressing interference. In contrast, as illustrated in Fig. 11(b), the ROSDP structure results in an orthogonal and well-distributed

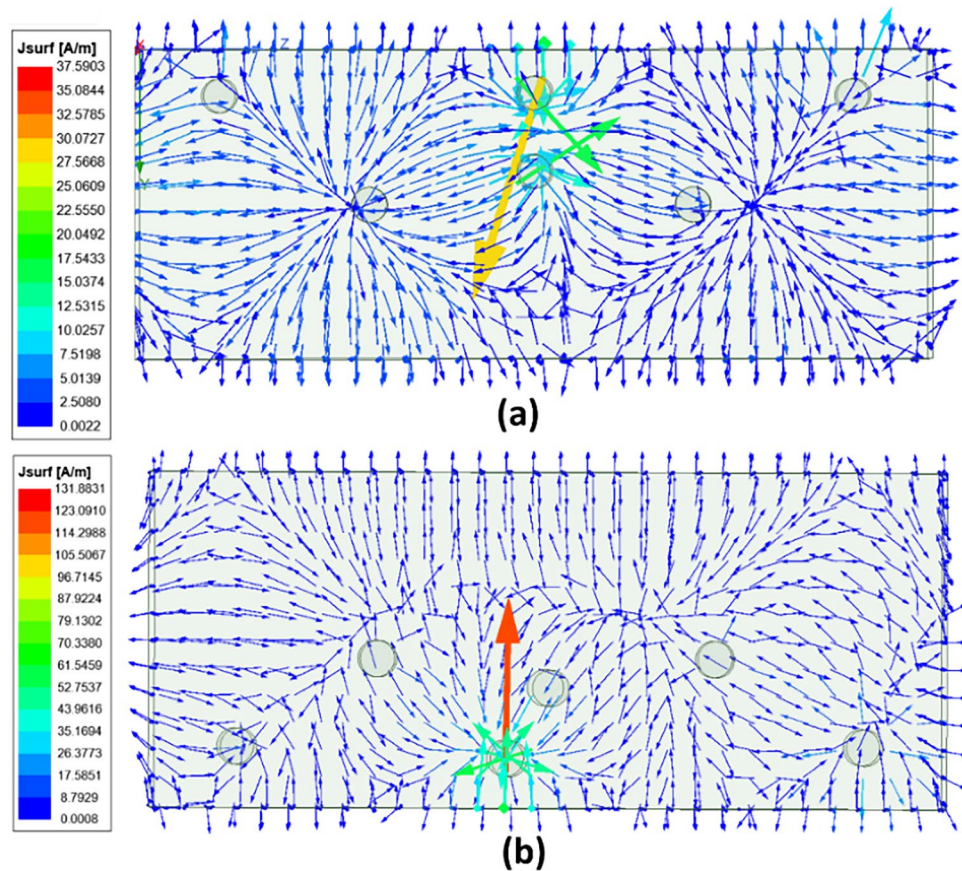


Fig. 11. Simulated surface-current distribution on the waveguide walls and posts at the resonant frequency for (a) the inline (IDP) configuration, where the current is strongly localized around the closely spaced posts, and (b) the rotated-offset (ROSDP) configuration, where the current is more uniformly distributed, indicating reduced inter-post coupling and improved polarization purity.

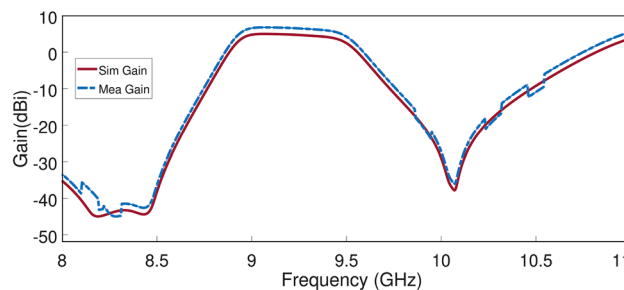


Fig. 12. Simulated and measured realized gain of the ROSDP filtenna, obtained by the gain-comparison (substitution) method; the third-order bandpass filtering response passes the 8.925 GHz to 9.5 GHz band while rejecting out-of-band signals.

surface-current pattern. This is due to the rotational symmetry of the rotated-offset structure, which leads to a more uniform current distribution, better polarization purity, and reduced coupling between the posts. In addition, this is further supported by the low intensity of currents in the surrounding areas. Therefore, the performance and efficiency of the ROSDP structure are superior to those of the inline structure.

Gain

It is shown in Fig. 12, the filtenna exhibits a third-order bandpass filtering response in its realized gain, transmitting signals within the operating band while attenuating signals outside it. The simulated gain is stable and corresponds to the expected values at the operational frequency interval. It provides efficient filtering of unwanted signals in the interval from 8.1 GHz to 8.5 GHz, where there is no radiation; then the filtenna starts transmitting signals in the interval from 8.925 to 9.5 GHz; the bandwidth equals 575 MHz with stable gain; in

this case, the minimum value of the S-parameter equals -20 dB. Then, beyond 10 GHz, the filtenna becomes a filter again and efficiently filters the signals; the roll-off rate of the filter has the characteristic of the third-order filter. The realized gain of the fabricated filtenna was measured with the help of the gain comparison (two-antennas substitution) method in the anechoic chamber. First of all, the standard gain horn with known gain G_{ref} was used as the receiving antenna to establish the reference power level, and then it was replaced by the antenna under test (AUT), while the transmitting antenna and the distance between the antennas were the same as before. The condition for far-field separation R was met, namely, $R \geq 2D^2/\lambda_0$ fulfilled (D is the size of the largest aperture). The gain of the AUT was then computed from the measured received-power ratio as^{20,35}

$$G_{\text{AUT}}(\text{dBi}) = G_{\text{ref}}(\text{dBi}) + 10 \log_{10} \left(\frac{P_{\text{AUT}}}{P_{\text{ref}}} \right), \quad (10)$$

where P_{AUT} and P_{ref} are the powers received by the AUT and the reference horn, respectively. The measurement was repeated across the operating band, and mismatch correction was applied using the measurement $|S_{11}|$ so that the reported values correspond to realized gain. The close agreement between the measured and simulated curves in Fig. 12 therefore reflects the accuracy of this calibrated substitution procedure rather than an idealized measurement environment.

Radiation patterns

The measured and simulated far-field patterns shown in Fig. 13 are compared in order to evaluate the performance of the filtenna in terms of its directivity, beam width, and radiation characteristics. Radiation patterns have been measured for both the E-plane (XZ-plane), Fig. 13(a)–(d), and the H-plane (YZ-plane), Fig. 13(e)–(h), at 9.07 GHz, 9.23 GHz, 9.38 GHz and 9.97 GHz. In general, the measurements agree well with the simulations, with a high level of efficiency over the entire frequency range where the simulated passband radiation efficiencies exceed 90%. The measured realized gain is 6.15 dBi at 9.23 GHz, confirming effective in-band radiation with minimal loss. Within each panel of Fig. 13 the co- and cross-polarization curves are referred to a common

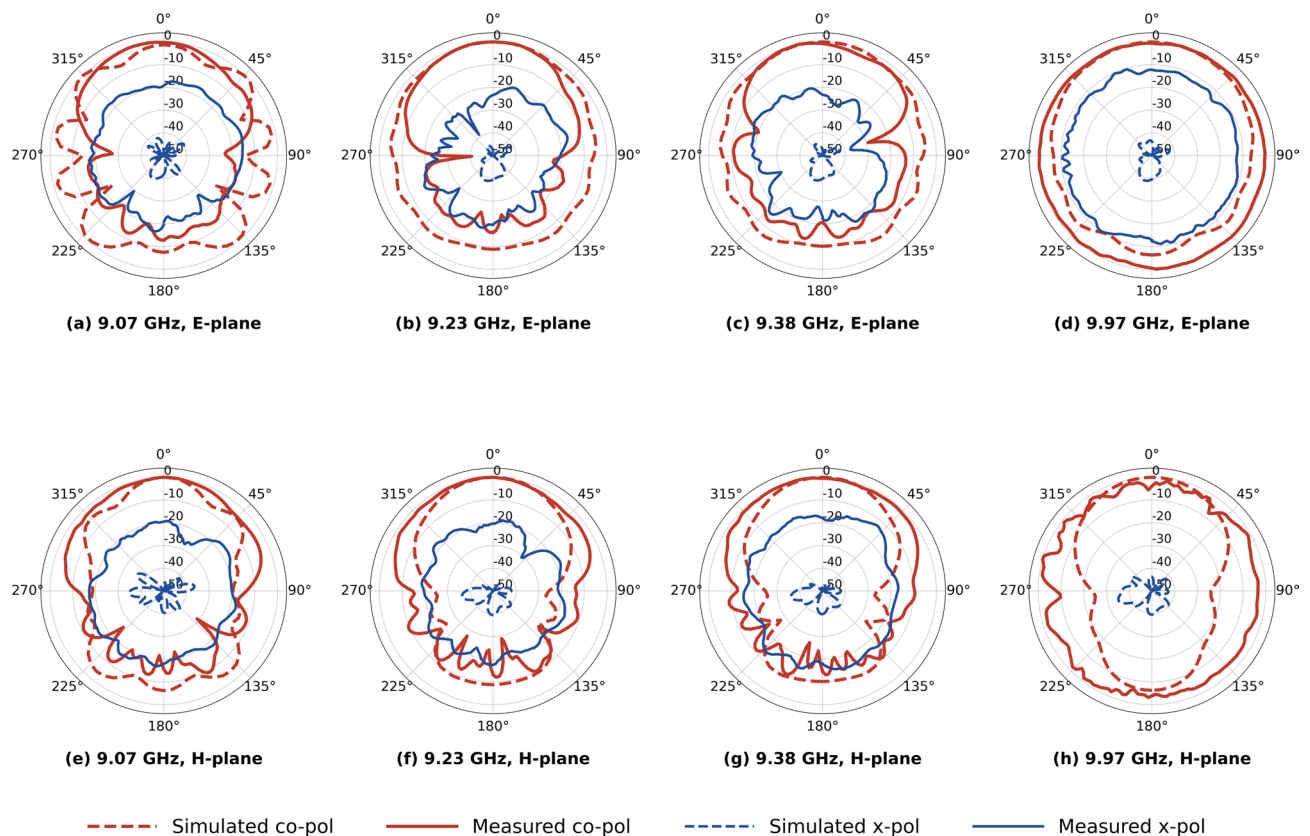


Fig. 13. Measured and simulated co- and cross-polarization radiation patterns of the ROSDP filtenna: (a)–(d) E-plane (XZ-plane) and (e)–(h) H-plane (YZ-plane) at 9.07, 9.23, 9.38, and 9.97 GHz, respectively. Co-polarization curves are shown in red and cross-polarization in blue, with dashed lines for simulation and solid lines for measurement. Within each panel all four curves are referred to the same co-polarization peak, so the cross-polarization level is read directly as the cross-polarization discrimination. The 9.97 GHz panels lie at the upper transmission zero, where the radiated level falls more than 30 dB below band centre; they are included to show out-of-band rejection rather than polarization performance.

reference, namely the co-polarization peak of that panel, so that the cross-polar level can be read directly as the cross-polar discrimination.

The side-lobe levels remained below -5 dB across the operational frequencies. The measured cross-polarization discrimination at boresight ranges from 17.5 dB to 23.3 dB across the passband and remains better than 12.5 dB over the main lobe ($\pm 60^\circ$), confirming that the rotated-offset arrangement preserves a well-defined linear polarization despite the asymmetry of the posts. The full-wave model predicts a discrimination above 36 dB; the difference is attributed to probe alignment in the chamber, the finite cross-polarization isolation of the measurement setup, and polarization leakage at the SMA-to-waveguide transition, none of which are represented in the simulation. Despite discrepancies in the side-lobe and back-lobe regions, the main-beam direction, -3 dB beamwidth, and boresight gain in the measured results agree fairly well with those of the simulation for the entire operating band. The broad-side radiation is maintained at all the four frequencies, confirming the stable in-band radiation characteristics. Large deviations from boresight have no effect on the main radiation characteristics of the filtenna. The deviation observed between the measured and simulated radiation patterns, especially in the H-plane pattern and near the nulls and side lobes, is mainly due to fabrication tolerances, the SMA-to-waveguide transition, misalignment of the anechoic-chamber setup, and ground and mounting effects that are neglected in the simulation. The same effects account for the larger part of the gap between the measured and simulated cross-polarization levels, since a cross-polarization component some 35 dB below the main beam is comparable with the residual cross-polarization response of the chamber itself. The input match has less effect on the far-field radiation compared with the above-mentioned aspects. This is the reason that S_{11} and realized gain of Figs. 10 and 12 agrees more closely with the simulations than does the radiation pattern.

Power handling and localised field stress

High power-handling capability is one of the motivations for a metallic waveguide realisation, and the peak power limit of the proposed structure is therefore estimated analytically. For the TE₁₀ mode in an air filled rectangular waveguide the transmitted power is related to the peak transverse electric field by³⁷

$$P = \frac{abE_0^2}{4Z_{TE}}, \quad Z_{TE} = \frac{\eta_0}{\sqrt{1 - (f_c/f_0)^2}}, \quad (11)$$

where η_0 is the free-space impedance. With $a = 22.86$ mm, $b = 10.16$ mm, $f_c = 6.56$ GHz and $f_0 = 9.3$ GHz, the wave impedance is $Z_{TE} = 532 \Omega$. Taking the breakdown strength of dry air at standard conditions as $E_{br} = 3 \text{ MV m}^{-1}$ gives a theoretical peak-power limit of approximately 0.98 MW for the unloaded WR-90 section.

The metallic posts raise the field locally and therefore reduce this limit. The dominant enhancement occurs at the tip of the offset post, where the field concentrates in the reduced clearance between the post and the opposite broad wall, and is further increased by the edge curvature of the post tip. For the offset post of the present design ($h_3 = 5.87$ mm, leaving a clearance of 4.29 mm), the combined enhancement is of order four. Since the transmitted power scales with the square of the field, the breakdown threshold falls by more than an order of magnitude, to the region of 50 kW to 100 kW peak. Applying the derating factor of three conventionally used to account for surface roughness, humidity, and assembly tolerance, the structure is expected to sustain of the order of 20 kW peak. Being air-filled and free of dielectric loading, its average power capability is limited by conduction cooling of the aluminium body rather than by breakdown.

These levels place the proposed filtenna within the requirements of the radar and satellite front-ends identified in Section 5, and well above what a planar equivalent based on stepped impedance or fractal resonators would sustain, which supports the choice of a metallic waveguide implementation. The values quoted are analytical estimates; a definitive breakdown threshold would require dedicated high power testing.

Benchmarking

The proposed design is compared with recent waveguide-based filtering antennas in terms of operating frequency, bandwidth, gain, selectivity, and physical size. Table 3 summarizes the performance of the proposed filtering antenna and its advantages relative to the comparison designs. The design simplifies the structure by employing rotated and opposite posts, making the structure compact and high performance. The major benefits are high selectivity and effective filtering, enabling efficient rejection of out-of-band signals.

The design is highly efficient, and its compact, low profile configuration is suitable for space-constrained applications. This design is simple compared with high-performance designs that rely on multi-mode or multi-cavity structures, and hence it is applicable to both practical and advanced communication systems^{12–17, 20, 22, 33, 36}. As shown in the comparison table, the primary benefit of the proposed design is the combination of filtering and radiating properties with an overall efficiency over 90 % and a peak efficiency of 92 %. The design also offers a compact construction suitable for space-limited applications while providing a moderate gain of 7.05 dBi and 6.15 dBi. Its flexibility is offered by its ability to operate in two configurations, the inline and the rotated-offset dual-post, at their respective centre frequencies in the X-band. This design is simpler to implement, with inline, rotated, and shifted post waveguides compared with complex multi-mode or multi-cavity designs.

The advantages of the proposed filtenna are best assessed together with the compromises they entail. First, integrating the filtering and radiating functions within a single WR-90 cavity removes the interconnection, matching network, and insertion loss required by a cascaded filter and antenna arrangement and reduces the overall footprint and assembly complexity; the cost of this integration is that the same rotated and offset posts couple, resonate, and radiate simultaneously, so the filtering and radiation responses are not fully independent and cannot be tuned in complete isolation. Second, the metallic-waveguide realization provides a high unloaded quality factor, owing to its low conductor loss and negligible dielectric loss, which sustains the sharp roll-off

	λ_0	Elec. Dim.	Phys. Dim.	F_0	FBW	FBW	Technique	Rad.	No.	Eff.	Gain
Refs	(mm)	(λ_0)	(mm ³)	(GHz)	Pass	Stop		elem.	TZs	(%)	(dBi)
²	30.00	$0.762 \times 0.339 \times 0.833$	2286401625	10	2% (9.91–10.1)	4% (9.4–9.8); 2% (10.4–10.6)	4x2 wg. aperture array; three-layer	8 (array)	NG	>90	15.58
¹³	32.26	0.710315×1.29	22864016415	9.3	23% (8.23–10.37)	3.23% (11.2–11.5)	Inductive singlet (FVC)	1	1	>90	3.53
²⁰	32.43	1.23108142711	3992644879	9.25	2.16% (9.15–9.35)	2.16% (9.6–9.8)	Coaxial cavity, tuning screws	1	NG	>75	>6.5
³⁸	30.00	NG	NG	10.0	4% (9.8–10.2)	NG	Triple-mode cavity; grid-slotted patch	1	NG	NG	>8.5
³⁵	30.00	2.116357×0.40	63484064242	10	11% (9.5–10.5)	6% (11.4–12)	Waveguide planar antenna array	Array	NG	<70	20.32
This Work	32.26	21500700341	6035245141	9.3	5.8% (8.95–9.49)	4.3% (8.2–8.6); 2.15% (9.9–10.1)	Rotated and shifted posts	1	4	>90	6.15

Table 3. Comparison of the proposed filtenna with recent waveguide-based filtering antennas in terms of frequency band, bandwidth, gain, selectivity, and physical size. Significant values are in bold.

(100 dB GHz⁻¹ to 140 dB GHz⁻¹) and the deep TZ (−92.5 dB) without lossy loading elements; the same high-Q resonant mechanism, however, inherently limits the attainable bandwidth, so the passband fractional bandwidth of 5.8 % remains well below wideband designs such as¹³ (23 %) and the design is unsuited to ultra-wideband operation. Third, the geometry-based control of four TZs keeps the filter order and element count low, but the zero positions depend sensitively on the post rotation angle and lateral offset, so small machining and assembly tolerances shift the band edges and zeros; this is reflected in the measured results, where the band edge shifts by only about 25 MHz and the upper TZ by 34 MHz, while the in-band return loss changes from −41.4 dB (simulated) to −29.8 dB (measured). In last, the single radiating aperture results in reasonable peak gain (7.05 and 6.15 dBi) compared to the multi-element arrays of² and³⁵ (15.58 and 20.32 dBi). Since the gain is limited by the physical aperture, the narrow 0.341 λ_0 aperture, which contributes to the compactness of the design, limits the possible gain at the same time, since the two goals are in conflict with each other. In summary, the design purposely sacrifices peak gain and bandwidth in favor of high selectivity, a compact low-profile design, and integration of filter and antenna, making it more suited for compact front ends with a high requirement for out-of-band rejection rather than to wideband or high-gain applications^{19, 33, 39}.

Beyond the tabulated comparison, the selectivity of the proposed design can be quantified directly from the simulated $|S_{21}|$ response. For the ROSDP filtenna, the upper-band roll-off is approximately 100 dB GHz⁻¹ (from −3 dB at 9.65 GHz to −20 dB at 9.82 GHz), with the adjacent TZ reaching −92.5 dB at 10.0 GHz; the lower-band roll-off is approximately 60 dB GHz⁻¹. The inline configuration is even sharper, with an upper-band roll-off of about 140 dB GHz⁻¹ and a TZ located immediately below the passband. At the design point ($h_3 = 5.87$ mm, $L_1 = 2.22$ mm, $f_0 = 9.3$ GHz), the external and radiation quality factors are $Q_e \approx 30$ and $Q_r \approx 73$, respectively. This level of selectivity is achieved using only two cylindrical posts in a single machined WR-90 cavity, whereas² and³⁵ employ multi-element aperture and planar arrays and²⁰ uses a coaxial cavity with tuning screws of 2.711 λ_0 length. It should be noted that the higher-gain designs in² and³⁵ are multi-element arrays, whereas the proposed work is a single radiating element. The proposed structure maintains a low-profile height of 0.341 λ_0 . The designs of² and¹³ share the same WR-90 guide height and are therefore bounded by the same limit, so the height of the present work is set by the waveguide standard rather than by the post topology; the advantage over the cavity-based designs of²⁰ and³⁵ is by contrast a factor of 2.4 and 1.2, respectively. The rotated and offset-post approach thus delivers high selectivity within a compact low-profile volume using the fewest coupling and radiating elements.

Conclusion

This paper introduced a compact dual-post waveguide filtenna based on a WR-90 rectangular waveguide with two metallic posts arranged in inline and rotated-offset formats. The design offers a high quality factor and high stopband rejection with accurate control of TZs. This novel method eliminates the need for traditional coupling irises to control bandwidth. The design achieves maximum gains of 7.05 and 6.15 dBi and efficiencies above 90 % over the frequency ranges of 10.6 GHz to 11.25 GHz (inline) and 9 GHz to 9.5 GHz (ROSDP). The close agreement between simulated and measured results validates the design and demonstrates its potential for radar and satellite communication applications. Its compact size and high performance make it suitable for advanced wireless communication platforms, addressing challenges in modern RF front-end components. The present single-element design provides a moderate gain; extending the concept to a multi-element array to increase gain, while preserving the compact filtering behavior, is identified as a limitation of the current work and a direction for future investigation. Further development will involve working with advanced materials and using optimization techniques to minimize structural complexity and improve performance, thereby expanding the operational scope of emerging RF technologies.

Data availability

“The data used and/or analyzed during the current study may be available from the corresponding author upon reasonable request under applicable policies.”

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Author contributions

Lubna Iram: Visualization, Formal analysis, writing original draft. Nouf Abd Elmunim and Abdul Aziz Memon and Mariana Dalarsson: Writing, review & editing, Resources, Conceptualization. Mariana Dalarsson and Shahid Khan: Validation, Supervision, Conceptualization and Funding acquisition. Haris Sarfraz : Visualization, Methodology, Investigation. Xue Jun Li: Validation, Resources, Investigation. Bilal Tariq Malik: Writing original draft, Supervision, Conceptualization. Shahid Khan: Writing, review & editing, Resources, Project administration.

Declarations

Competing interests

There are no conflicts of interest to declare.

Additional information

Correspondence and requests for materials should be addressed to M.D.

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