

Mushroom Resonator Loaded Substrate Integrated Waveguide Based Dual-Band Bandpass Filter with Independent Passband Tunability

Soumit Samadder Chaudhury^{1,*}, Seema Awasthi², and Rajat Kumar Singh³

¹*Department of Electronics and Communication Engineering, Faculty of Science and Technology (ICFAI Tech.)
ICFAI Foundation for Higher Education, Hyderabad, India*

²*Department of Electrical Engineering, Indian Institute of Technology Kanpur, Kanpur, India*

³*Department of Electronics and Communication Engineering, Indian Institute of Information Technology Allahabad
Allahabad, India*

ABSTRACT: This paper presents a compact dual-band bandpass filter, based on substrate integrated waveguide (SIW) technology, highlighting independent tuning abilities over center frequencies. Rectangular mushroom resonators (RMRs) are inserted inside an SIW cavity to realize the filtering structure, where two transmission bands are generated, which are centered at 2.2 GHz and 4.2 GHz, respectively, by simultaneously employing TE_{101} and TE_{201} resonant modes. Unlike conventional tunable dual-band filters, which generally exhibit interdependent frequency shift between the operating bands, the proposed design employs an independent modal-tuning mechanism via selective perturbation of the resonant current paths. In the proposed methodology, radial and linear perturbation slots are strategically introduced at locations derived from modal electric-field and surface-current analyses to selectively modify the effective electrical length of separate resonant modes. One mode can be shifted without noticeable influence on the other by properly orienting these perturbation elements and thereby enabling isolated control of each passband while preserving the original cavity geometry. To obtain practical dual-band bandpass filtering, cascaded RMR-loaded SIW resonators are configured into a highly selective filter topology. Metallic radial perturbation shafts are incorporated into the resonator structure to facilitate dynamic adjustment of slot exposure for real-time and post-fabrication tuning. Experimental verification confirms independent tuning of the lower passband from 2.0 GHz to 2.2 GHz, corresponding to around 9.5% frequency shift, while maintaining the upper band unchanged. Similarly, the upper passband is independently shifted from 4.0 GHz to 4.1 GHz, yielding close to 2.5% tuning range without affecting the lower band. The filter has insertion loss below 2 dB across the full tuning range while showing negligible compromise in passband selectivity. With its compact structure, low-loss characteristics, and independently reconfigurable capability, the proposed design is well suited for adaptive multiband microwave and wireless front-end implementations.

1. INTRODUCTION

Independent control over passbands offered by bandpass filters of multiple passbands has substantially influenced transponder systems in microwave communication. In contemporary microwave engineering, reconfigurable multiband bandpass filters are expected not only to provide independent passband tunability but also to maintain high Q-factor, low passband insertion loss, enhanced selectivity, compact form factor, and ease of integration with planar circuitry. In response to these demanding specifications, considerable progress has been made in advancing Substrate Integrated Waveguide (SIW) technology, paving the way for implementing high-performance filter designs with compact filters [1–3]. Extensive prior work has exploited inherent advantages of SIW technology to realize multiband bandpass filters, with various techniques proposed for regulating the resulting passbands. However, these approaches generally have limited flexibility in achieving truly independent control due to

inherent modal coupling and structural constraints [4–6]. Composite right/left-handed (CRLH) structures have been used to enhance selectivity and out-of-band response while enabling compact multiband bandpass filter designs [7–9]. Mushroom resonators, as typical CRLH structures, have been used to enable passband tuning in SIW cavity-based multiband bandpass filters [10]. The present approach surpasses existing methodologies by facilitating precise, decoupled control of passband characteristics through a systematically engineered compact resonant architecture, resulting in better selectivity and improved overall performance for next-generation reconfigurable microwave systems.

Various research groups have carried out extensive investigation and systematic engineering regarding passband tuning in bandpass filters. Much of the literature has reported tunable center frequency control over SIW-based bandpass filter as significant progress. In certain cases, exploiting the electric-field intensity of the SIW cavity and employing lumped capacitor and varactor-type tuning elements establish the control over the resonant frequencies of individual passbands [11]. The usual

* Corresponding author: Soumit Samadder Chaudhury (soumit@ifheindia.org).

types of tuning elements implemented to tune the passbands in multiple bandpass filters are electrical, mechanical, or radio frequency microelectromechanical systems (RF-MEMSs). Microwave filter designers, researchers, and microwave engineers choose electrical tuning elements above all due to their ease in handling the controlling voltage to control passband center frequencies. The CFs of passbands are controlled, as reported, by multiple types of electrical tuning elements, such as stub capacitors, lumped capacitors, surface-mount capacitors, and piezoelectric actuators [12–17]. The complexity and operational limitations of those implemented electric tuning elements are critical to those tunable bandpass filters [12–17]. Although incorporating electrical or electronic tuning elements helps gain noticeable control over the CFs of passbands, they add significant insertion losses in passbands [12, 14, 17]. RF-MEMS devices are also considered impactful and advantageous members of passband tuning elements along with other conventional electrical tuning elements as far as center frequencies and bandwidth tuning of SIW based bandpass filters are concerned [18, 19]. Mechanical tuning elements have also been used for the development of passband control [20–22]. Metallic perturbation elements are formed by inserting multiple tuning screws into the auxiliary vias of SIW cavity resonators to control the passbands while maintaining low insertion loss in both common and differential modes [22]. Passband control has been demonstrated predominantly for single-band bandpass filters, whereas comparatively fewer studies have explored dual- and multi-band configurations with tunable passband controlling ability [12, 14–16, 22, 23]. Introducing complementary split ring resonators (CSRRs) and perturbation slots within the resonator cavity became an interesting modification in the entire structure as far as achieving control over the passband is concerned, although exact tuning techniques were not defined in such reports [23].

Extensive research on tunable bandpass filters has led to developing numerous techniques that provide operational frequency flexibility, bandwidth reconfigurability, and compact filter architectures suitable for modern wireless communication systems. Although CMOS-based programmable filters show excellent bandwidth reconfigurability and rapid operational frequency tuning, their application in high Q microwave environments is often restricted by intricate circuits and relatively low power-handling capacity [24]. Dual-band filters using hybrid resonator configurations with varactor-based tuning have demonstrated independent adjustment of operating frequencies, enabling large frequency separation and broad stopband characteristics. However, these benefits are often achieved at the expense of increased structural complexity and reduced resonator quality factor at higher operating frequencies [25]. The realization of band-reconfigurable filtering responses through mode-switching techniques typically requires multiple tuning elements, which facilitate transitions between single-band and multiband operation. Nevertheless, the resulting architectures often have increased implementation complexity and limited compatibility with compact planar platforms [26]. SIW filters incorporating RF-MEMS technology offer high quality factors and substantial size reduction; however, their practical implementation is often constrained by fabrication complexity and

limited tuning flexibility [27]. Continuous frequency tuning and improved stopband attenuation can be achieved through varactor-integrated SIW filters. However, inherent coupling between the tuning mechanism and other filter performance metrics may introduce operational trade-offs and compromise system stability [28]. Wide-range frequency reconfigurability and high selectivity have been achieved via coaxial resonator and liquid crystal-based filter designs. Despite these advantages, their practical implementation is often complicated by intricate biasing schemes and constrained tuning responsiveness [29, 30]. Graphene-assisted filter designs have demonstrated tunable attenuation and enhanced functional versatility. Nevertheless, these techniques generally cannot achieve independent reconfiguration of multiple passband center frequencies [31, 32]. Liquid-metal-based resonator technologies offer an attractive route toward continuous frequency tuning while maintaining low insertion loss, but challenges associated with response speed and implementation complexity continue to limit their widespread adoption [33]. By effectively suppressing unwanted reflections, absorptive and quasi-reflectionless filters contribute to improved system-level performance. Despite this advantage, most reported implementations provide tuning capability only within a single operating band [34, 35]. A Substrate Integrated Suspended Line (SISL)-based quad-operation-mode bandpass filter has been developed to provide simultaneous control over center frequency, bandwidth, and external quality factor. This functionality is achieved by combining varactor-integrated resonators with separate electric and magnetic coupling mechanisms [36]. The discrete switching nature of PIN diode makes established passband filters specifically suitable for applications with fixed operating modes but suffers from high insertion loss [37]. Availability of commercial components with wide capacitance tuning ranges and relative ease of integration with planar microwave circuits are the consequences of matured varactor technology [38]. Multiple passbands with tuning ranges spanning from modest 10–15% fractional bandwidths to impressive multi-octave coverage has been proposed too, but with significant insertion loss [39]. Absolute bandwidth design, which maintains fixed channel bandwidth across the tuning range, is essential for applications with standardized channel allocations, such as cellular base stations and satellite communications [40]. However, such implementations require more complex tuning-element configurations, typically involving simultaneous adjustment of resonator frequencies and coupling coefficients, which increases design complexity and control-circuit requirements. Multiple passband filters with applications requiring continuous frequency agility and wide tuning ranges, such as cognitive radio and spectrum-sensing systems with varactor-based designs, demonstrate clear advantages despite their performance trade-offs. The emergence of variable bandwidth designs that offer independent control of both center frequency and bandwidth represents the advanced flexibility of passbands, though at the cost of significantly increased complexity [41]. While recent developments have considerably advanced the state of tunable microwave filters, developing a compact SIW architecture capable of independently reconfiguring multiple passbands without compromising selectivity, insertion loss, or modal isolation re-

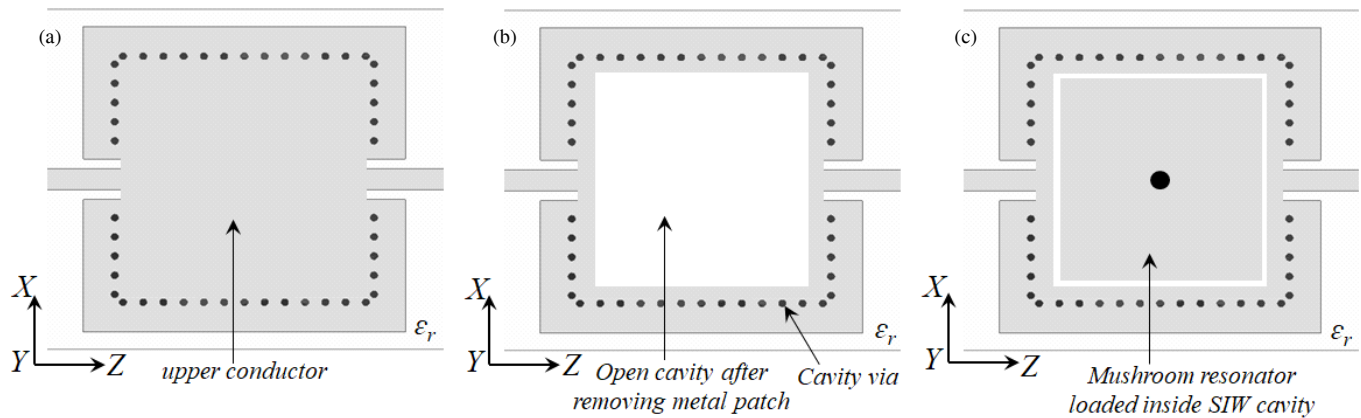


FIGURE 1. (a) A substrate integrated waveguide (SIW) cavity, (b) a SIW cavity with upper conductor portion etched out, and (c) schematic of SIW cavity loaded with a mushroom resonator.

mains an open research challenge. The proposed work is directed toward addressing this critical gap.

To take care of these ongoing issues in multiple-passband tuning, this article proposes a dual-band SIW bandpass filter based on rectangular mushroom resonators, and independently controlling fundamental and higher-order resonant modes is achieved through a systematic perturbation strategy. Selectively positioned perturbation slots are introduced to alter the effective electrical path of each mode individually after exploiting the distinct electric-field distributions and thereafter the surface current distributions of TE_{101} and TE_{201} modes, which enabled true decoupled tuning of the corresponding passband center frequencies. Structural simplicity with minimal passband interactions is ensured in this proposed method, which is a quite different approach from existing ones, fundamentally based on multiple coupled tuning elements or hybrid resonator combinations. A post-fabrication tuning approach based on radial metallic perturbation elements is further introduced to realize continuous, real-time frequency tuning without structural modification of the cavity. As a result, the filter delivers independent dual-passband tuning with minimal insertion loss, maintained bandwidth, and enhanced selectivity, offering a compact and effective solution for reconfigurable microwave filtering.

The proposed work explores the development of independently tunable dual-band bandpass filters using rectangular mushroom resonators (RMRs) embedded in substrate-integrated waveguide structures, with emphasis on modal control and frequency reconfigurability. This research article has the following key contributions: (1) By employing strategically positioned arc-shaped and linear perturbation slots, independent modal tuning is realized in RMR-loaded SIW cavities, with tuning ranges of nearly 12.5% and 7% for the first and second excited modes, respectively, without cross-coupling effects. (2) The development of a bandpass filter architecture of dual passbands with enhanced selectivity, by using cascaded RMRs with extended coupling slots that generate transmission zeros. (3) A novel post-fabrication tuning mechanism is implemented using radial metallic shafts and allows adjusting both passbands independently with tuning ranges of 9.5% and 2.5%, respectively. (4) Displaying

insertion loss values of about 1.0 dB and 1.8 dB across the 1st and 2nd passbands, respectively, measured results authenticate the proposed approach and thereby establishes state-of-the-art performance for dual-band filters in 2–5 GHz frequency region with mechanical tuning ability.

2. THEORY DESIGN

An approach for independently manipulating the excited modes of a single mushroom resonator embedded in an SIW cavity is used in this overall concept of CF passband tuning. This strategy facilitates autonomous passband tuning in mushroom-loaded filters intrinsically characterized by high Q-factor, enhanced selectivity, and excellent out-of-band rejection. The mushroom resonator comprises two parallel metallic plates, but separated by a dielectric substrate and connected by a via. It is loaded into an SIW cavity by etching a selected region of the SIW upper conductor. The total structure is based on SIW, and Ansys HFSS is used to model and simulate a prototype of the proposed structure. A part of the SIW's upper conductor is etched out to make an exposed cavity, as shown in Figure 1(a) and Figure 1(b). Figure 1(c) shows the schematic of the entire structure after loading a mushroom resonator into SIW cavity. Parametric dimensions of the mushroom resonator are extracted by carrying out the simulator-driven eigenmode analysis.

2.1. Rectangular Mushroom Resonator

The fundamental architectural units of the proposed dual-band bandpass filter structure are rectangular mushroom resonators (RMRs). A rectangular metallic patch fabricated on a dielectric substrate and connected to a continuous metallic ground plane through a central cylindrical via are the constituents of a mushroom resonator. A composite resonant structure that allows multiple electromagnetic modes to be excited, which is studied and analyzed for distinct field distributions at excited resonance frequencies, is created by this mushroom configuration. Rogers RT/Duroid 5880 substrate, a low-loss microwave laminate with relative permittivity $\epsilon_r = 2.2$, loss tangent $\tan \delta = 0.0009$, and thickness $h = 0.787$ mm, is used to implement the RMR. A balance between the electrical per-

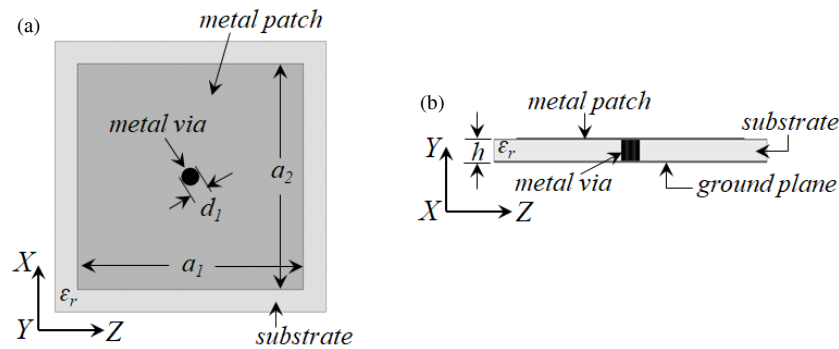


FIGURE 2. (a) Top view of the mushroom resonator and (b) cross-sectional view of the mushroom resonator.

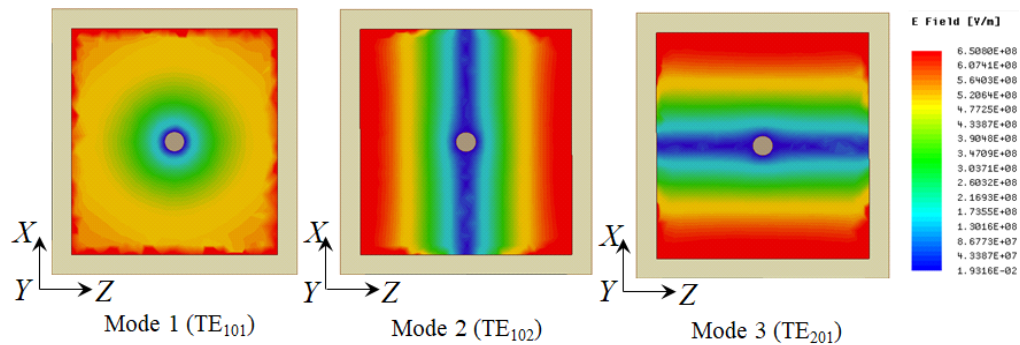


FIGURE 3. Electric fields distributions of mode 1, mode 2, and mode 3 of the rectangular mushroom resonator.

formance and mechanical stability for operational frequencies is primarily provided by this chosen substrate layer. Standard copper cladding of thickness $35\ \mu\text{m}$, which maintains compatibility with the fabrication process of conventional printed circuit board and provides low resistive losses, is used to implement the top patch and bottom ground plane of metallic layers. The rectangle-shaped mushroom structure is viewed from the top end as shown in Figure 2(a) and from the cross-sectional end as shown in Figure 2(b).

Careful optimization of the RMR's geometric dimensions is carried out to excite desired resonance modes at target frequencies. The rectangular patch dimensions are defined as $a_1 = 21\ \text{mm}$ along the X -axis and $a_2 = 23\ \text{mm}$ along the Y -axis, creating a slightly asymmetric structure that facilitates mode separation. The central via has diameter $d_1 = 2\ \text{mm}$ and provides a low-inductance connection between the metal patch and the ground plane. These dimensions are selected through eigenmode analysis to position fundamental and second-order modes at frequencies suitable for dual-band operation in the $2\ \text{GHz}$ – $5\ \text{GHz}$ range. Eigenmode analysis of the RMR structure reveals the dominant resonance mode within the frequency range of interest. Figure 3 shows electric-field distributions of the fundamental mode, i.e., mode 1 (TE_{101}), and the next higher-order modes, i.e., modes 2 and 3.

The fundamental mode, designated as TE_{101} , shows a resonant frequency of approximately $2.4\ \text{GHz}$. This mode is characterized by a half-wavelength variation of the electric field along the x -axis (a_1 dimension) with maximum field intensity concentrated near the edges of the patch perpendicular to the x -

axis. The surface-current distribution for the TE_{101} mode flows radially to the patch's perimeter, with maximum current density at the patch edges parallel to the y -axis. The next higher-order resonant modes, designated as TE_{102} and TE_{201} , occur at approximately 4.3 and $4.5\ \text{GHz}$, respectively, whose electric-field distributions are presented in Figure 3. These modes exhibit a full-wavelength variation of the electric field along the Z -axis and x -axis, respectively, creating two regions of maximum field intensity separated by a null at the patch center. Surface-current distributions for TE_{102} and TE_{201} modes are more complex than the fundamental mode, featuring multiple current loops with maximum density occurring at specific locations along the Y -axis. The spatial separation of maximum current density regions between the TE_{101} and the higher-order TE_{102} and TE_{201} modes provide the physical basis for independent mode control through strategically positioned perturbation elements. The transmission line theory can be used to anticipate the resonant frequencies of these mushroom resonators' excited modes. An individual mushroom resonator structure is modelled as a short-circuited transmission line with additional capacitive loading from the patch edges. The expression of resonant frequency for TE_{10n} modes is given by

$$f_{\text{resonance}} = \frac{nc}{2a_1\sqrt{\epsilon_{\text{eff}}}} \quad (1)$$

where n represents the mode number (1 for TE_{101} , 2 for TE_{102} , and likewise); c represents the speed of light; ϵ_{eff} is the effective permittivity accounting for fringing fields; and a_1 is the patch dimension along the direction of field variation. The effective

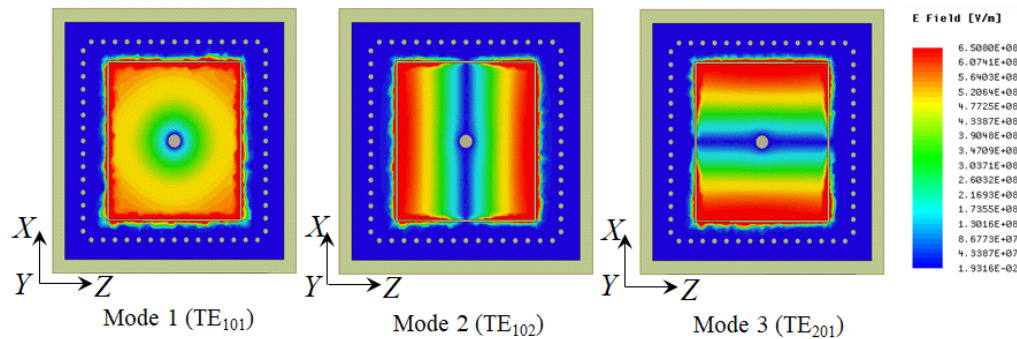


FIGURE 4. Electric-field distributions of mode 1, mode 2, and mode 3 of a rectangular mushroom resonator loaded into an SIW cavity.

permittivity is lower than the substrate permittivity due to the air above the patch, typically ranging from 1.6 to 1.8 for the chosen substrate parameters. A separate eigenmode analysis is carried out in order to study the excited modes after loading the mushroom resonator into the open SIW cavity. Electric field distribution is studied for the excited modes of the SIW cavity loaded with mushroom resonators and shown in Figure 4. It is evident from Figure 3 and Figure 4 that the electric field of mode 1 is perturbed by vias of the mushroom resonators placed at the center of the cavity, and the maximum field intensity is pushed towards the mushroom boundary. However, field distributions remain undisturbed in cases of modes 2 and 3 because the maximum field intensity is over the resonator's sides. The quality factor of the mushroom resonator is determined primarily by conductor losses in metallic layers, dielectric losses in the substrate, and radiation losses from the patch's open edges. For the Rogers RT/Duroid 5880 substrate with its low loss tangent, dielectric losses are minimal. Conductor losses are minimized via using standard copper cladding with adequate thickness. Radiation losses are inherently low for the mushroom structure due to the short-circuit boundary condition imposed by the central via, which suppresses radiation modes.

2.2. Rectangular Mushroom Resonator (RMR)-Loaded SIW Cavity

A hybrid structure composed of a rectangular mushroom resonator integrated into an SIW cavity provides the benefits of both technologies, i.e., the mushroom resonator and the SIW cavity. The SIW cavity provides electromagnetic shielding, suppresses spurious modes, and facilitates integration with planar transmission lines, while the loaded mushroom resonator introduces multiple controllable resonance modes within a compact footprint. The SIW cavity is formed by embedding two parallel rows of metallic vias in the substrate, creating sidewalls that confine electromagnetic fields in a manner analogous to conventional rectangular waveguides. The SIW cavity via diameter is selected as 0.8 mm with center-to-center spacing of 1.5 mm, satisfying the design rule that via spacing should be less than twice the via diameter to minimize leakage. The SIW width is designed to support desired propagation modes while maintaining single-mode operation below the cutoff frequency of higher-order modes. To accommodate the mushroom resonator, a rectangular aperture is etched in the upper con-

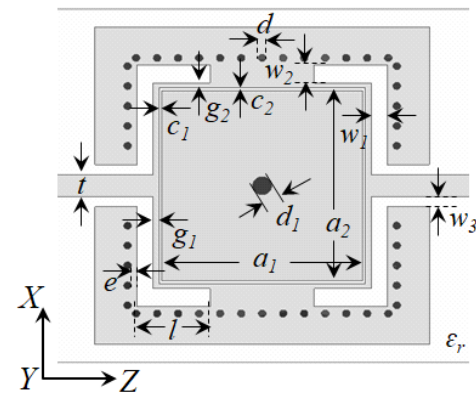


FIGURE 5. Schematic of mushroom resonator loaded SIW cavity.

ductor of the SIW cavity, creating an open region into which the mushroom patch is loaded. The aperture dimensions are slightly larger than the mushroom patch dimensions to prevent direct contact between the patch edges and the SIW sidewalls, maintaining the intended electromagnetic coupling characteristics. The mushroom via is positioned at the aperture's center, providing symmetric coupling to SIW cavity modes. As shown in Figure 5, a rectangular mushroom resonator is loaded inside a rectangular SIW cavity, and the loaded resonator is excited through capacitive coupling slots using a microstrip line of 50 Ω characteristic impedance. This resonator with its fundamental mode (TE_{101}) excited at 2.4 GHz and 2nd mode excited at 4.3 GHz has parametric values as follows: $a_1 = 21$ mm, $a_2 = 23$ mm, $d_1 = 2.0$ mm, $c_1 = 0.3$ mm, $c_2 = 0.3$ mm, $g_1 = 0.66$ mm, $g_2 = 0.53$ mm, $w_1 = 1.65$ mm, $w_2 = 2.0$ mm, $e = 0.59$ mm, $d = 0.8$ mm, $t = 2.42$ mm, $w_3 = 1.12$ mm and $l = 7.65$ mm. Microstrip lines are positioned to couple to the electric-field distribution of desired resonance modes, with the coupling strength controlled by the overlap length and gap between the microstrip and the mushroom patch. This coupling configuration provides external quality-factor control, enabling the adjustment of the filter bandwidth without affecting resonant frequencies. When a single mushroom resonator is loaded into the SIW cavity and excited through the microstrip coupling structure, the transmission response, as shown in Figure 6, shows two distinct resonance peaks corresponding to TE_{101} and TE_{102} modes of the mushroom resonator. Sharp transmission peaks at 2.4 GHz and 4.35 GHz, with insertion loss be-

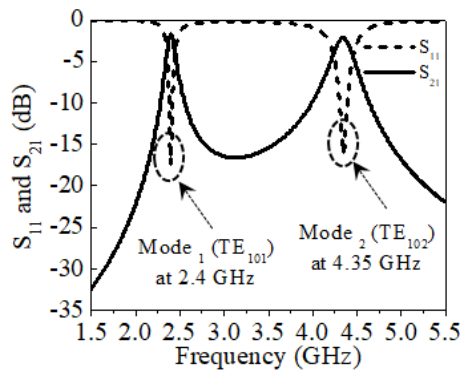


FIGURE 6. Simulated S_{11} and S_{21} of mushroom resonator-loaded SIW cavity.

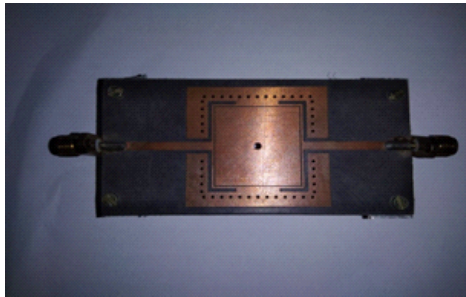


FIGURE 7. Fabricated mushroom resonator-loaded SIW cavity.

low 1 dB at both frequencies, are revealed after a full-wave electromagnetic simulation. The slight frequency shift compared to the isolated mushroom resonator, i.e., from 4.3 GHz to 4.35 GHz for the second mode, results from electromagnetic coupling between the mushroom structure and the SIW cavity walls. Figure 7 shows the fabricated sample of a mushroom resonator-loaded single cavity with parametric values, as mentioned above. Figure 8 shows the comparison between simulated and measured frequency responses of S_{21} of the mushroom resonator-loaded SIW cavity, where the measured response agrees fairly well with the simulated one.

The electric-field distributions, as shown in Figure 9 at the two resonant frequencies, confirm the mode assignments and provide insight into the coupling mechanisms. At 2.4 GHz (TE_{101} mode), the electric field shows maximum intensity near boundaries of the mushroom patch perpendicular to the x -axis, with field lines extending vertically between the patch and the ground plane. The field distribution is pushed toward the edges of the mushroom structure due to the short-circuit boundary condition imposed by the central via. This field configuration creates strong coupling to the microstrip feed lines when they are positioned near the patch edges, as shown in Figure 9 (Mode 1). At 4.35 GHz (TE_{102} mode), the electric-field distribution becomes more complex, with maximum intensity at two locations along the x -axis of the mushroom patch, separated by a field null at the patch center, as shown in Figure 9 (Mode 2). The field lines show a full-wavelength variation along the x -axis, creating two distinct regions of high field intensity along z -axis. This spatial field distribution is critical for achieving independent control of the second mode through perturbation

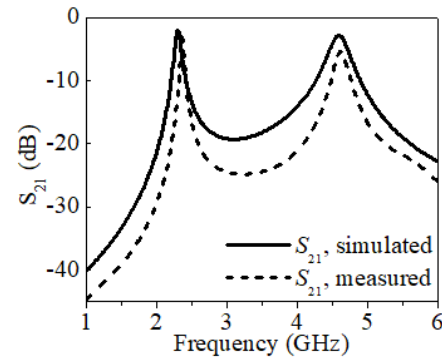


FIGURE 8. Simulated and measured S_{21} of mushroom resonator-loaded SIW cavity.

elements positioned at the patch areas of maximum field intensity.

The magnetic-field distributions at the two resonant frequencies reveal complementary patterns that are essential for understanding the surface-current distributions and designing effective perturbation strategies. Figure 10 shows surface-current distributions of an SIW cavity loaded with a mushroom resonator at modes 1 and 2.

At the TE_{101} resonance mode, magnetic field lines form closed loops around the perimeter of the patch, with maximum intensity occurring at the patch edges parallel to the Y -axis. So, in Mode 1 a strong concentration of surface current is observed around the mushroom via, corresponding to a minimum in the electric field distribution and conversely, the mushroom edges exhibit reduced current density while supporting the highest electric-field intensity as it has been illustrated in Figure 10 (Mode 1). At the TE_{102} resonance mode, the magnetic field pattern exhibits multiple loops corresponding to the full-wavelength field variation, with maximum intensity at specific locations along the x -axis. Therefore, the mode 2 current distribution shows a pronounced concentration of surface current along the central axis of the mushroom structure in the x -axis. This region coincides with near-zero electric-field intensity, indicating a current-dominant resonant behavior.

2.3. Independent Control over Fundamental and Higher-Order Modes Using Perturbation Slots

The proposed modal-selective perturbation scheme, realized through strategically positioned slots, represents the key innovation enabling independent dual-band tunability. Mode-selective perturbation is implemented by strategically introducing slots at regions corresponding to the maximum surface current density of the targeted resonant mode, while simultaneously avoiding high-current regions associated with other modes. This deliberate placement ensures selective modification of the desired mode without inducing significant interaction with adjacent modes. The perturbation slots elongate the surface current path, increasing the electrical length and producing a controlled reduction in the resonance frequency.

For controlling the first mode (TE_{101}), a pair of arc-shaped slots is etched into the mushroom patch. The arc slots are characterized by thickness t_1 , radius of curvature r , and angular

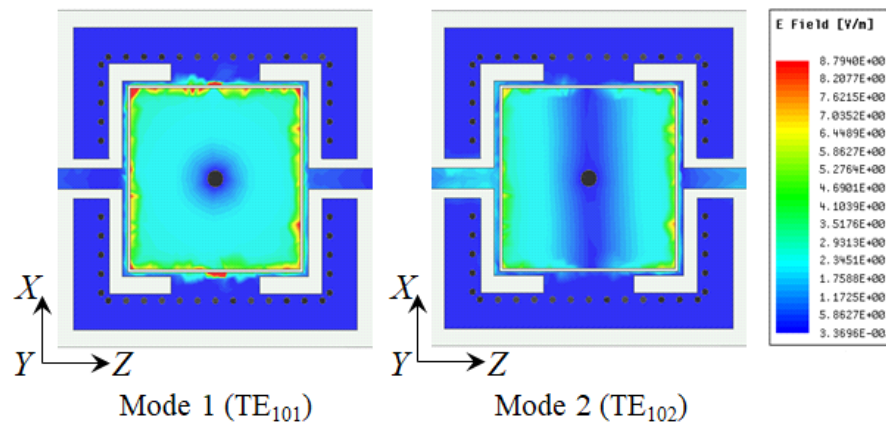


FIGURE 9. Electric-field distributions of Mode 1 (TE_{101}) and Mode 2 (TE_{102}) of mushroom resonator-loaded SIW cavity.

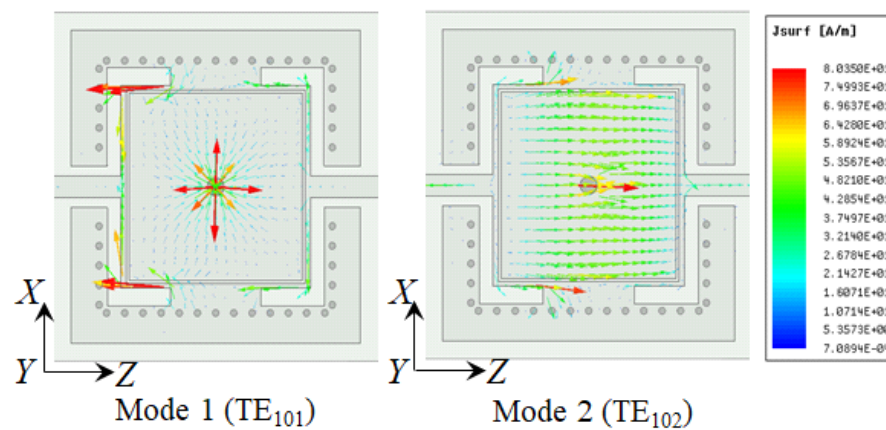


FIGURE 10. Surface-current distributions of Mode 1 (TE_{101}) and Mode 2 (TE_{102}) of mushroom resonator loaded SIW cavity.

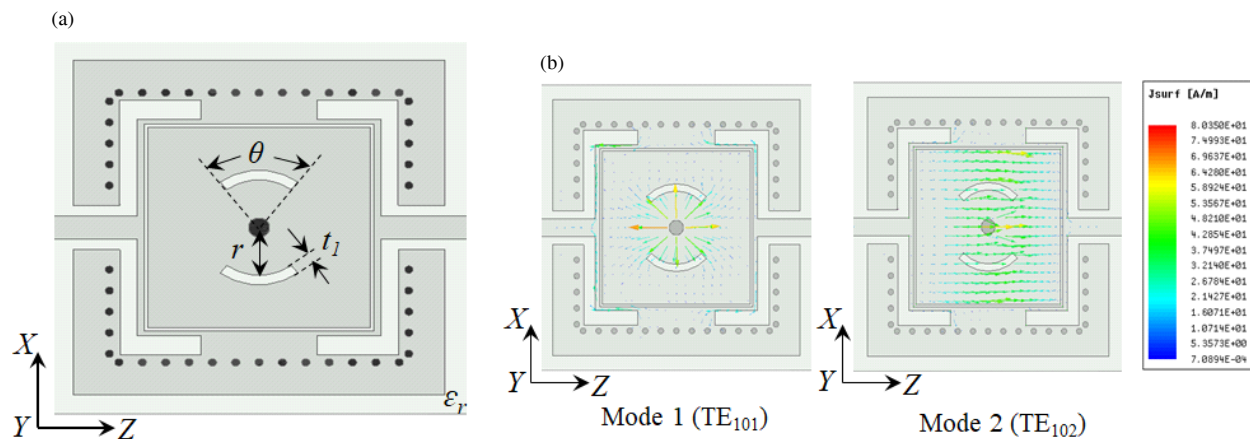


FIGURE 11. (a) Schematic, (b) surface current distribution of mode 1 and mode 2 of the mushroom resonator-loaded SIW cavity with perturbation slots of parameters, $t_1 = 1$ mm, $r = 5$ mm and $\theta = 30^\circ$, to control mode 1.

extent θ as shown in Figure 11. As shown in Figure 11(a), the slots are positioned at a radial distance of 5 mm from the mushroom center, corresponding to the location of maximum surface-current density for the TE_{101} mode. The arc shape is chosen to follow the natural current flow pattern of the fundamental mode, maximizing the perturbation effect while minimizing impact on higher-order modes. Owing to the nearly

transverse alignment of the perturbation slot with respect to the mode 1 surface current direction, the current is redistributed around the slot boundaries, extending the effective current path, as shown in Figure 11(b). As the arc slot angle θ increases from 60° to 120° , the surface currents associated with the TE_{101} mode are forced to follow progressively longer paths around the slot boundaries. This increased path length reduces the reso-

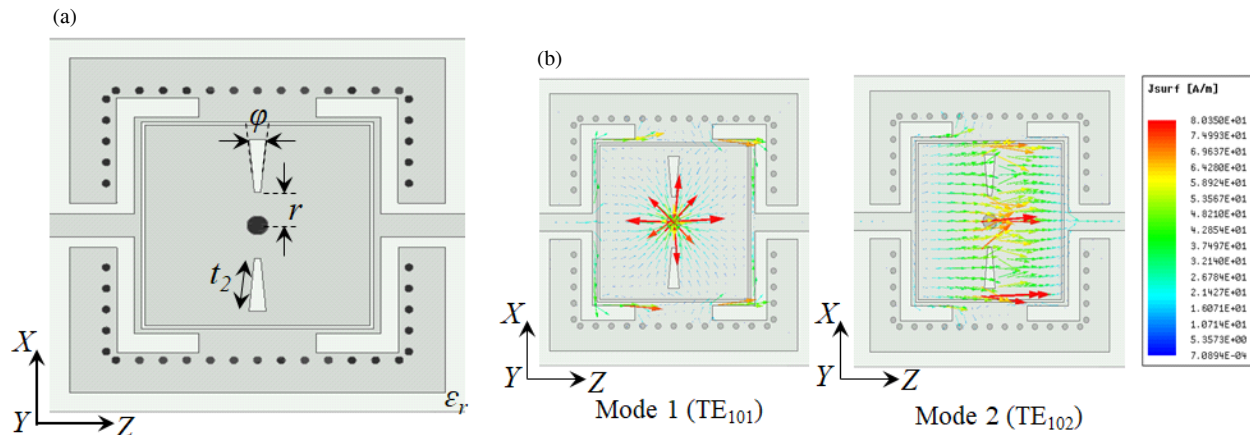


FIGURE 12. (a) Schematic, (b) surface current distribution of mode 1 and mode 2 of the mushroom resonator-loaded SIW cavity with perturbation slots of parameters, $t_2 = 5.5$ mm, $r = 3.5$ mm and $\varphi = 10^\circ$, to control mode 2.

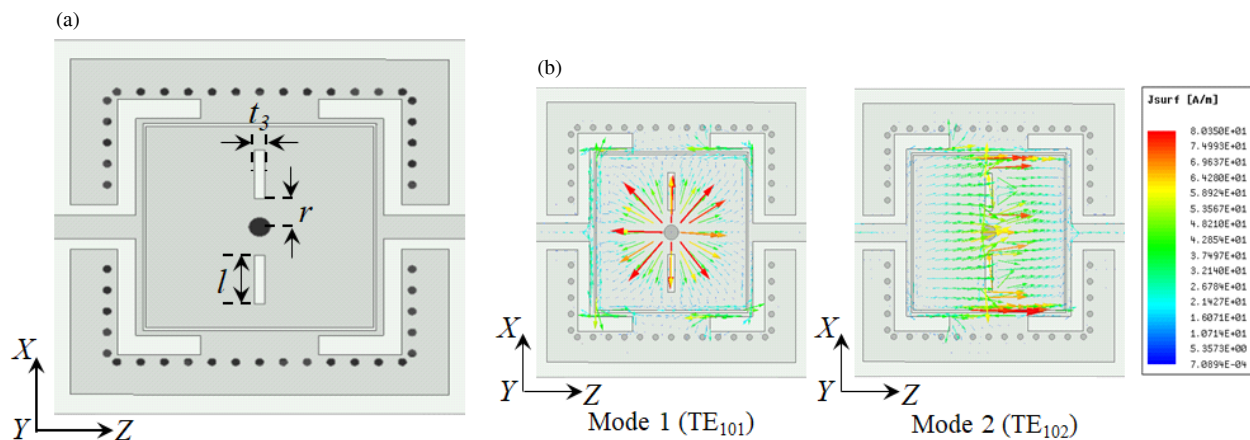


FIGURE 13. (a) Schematic and (b) surface current distribution of mode 1 and mode 2 of the mushroom resonator-loaded SIW cavity with perturbation slots of parameters, $l = 6$ mm, $t_3 = 1$ mm, and $r = 3.5$ mm, to control mode 2.

nance frequency from 2.4 GHz to 2.1 GHz, representing a shifting range of 12.5%. Critically, the TE_{102} mode remains virtually undisturbed throughout this tuning range, with frequency variation less than 0.5%. This independence results from the arc slots positioned at locations of relatively low current density for the TE_{102} mode, where the current distribution exhibits a null due to the full-wavelength field variation. The dominant surface current distribution of mode 2 is oriented nearly parallel to the perturbation slots etched on the metallic patch. Owing to this alignment, the slots exert minimal influence on the effective current path of the TE_{102} mode, as illustrated in Figure 11(b). Therefore, adjusting the perturbation angle θ selectively alters the resonance frequency of mode 1, whereas the resonant characteristics of mode 2 remain largely unaffected. This behaviour establishes a practical approach for independently tuning mode 1 while preserving the original cavity geometry.

For independently controlling the second mode (TE_{102}), two alternative perturbation strategies are investigated. The first approach employs arc-shaped slots with different orientations and dimensions compared to those used for the first mode. Figure 12(a) shows the schematic of the mushroom-loaded SIW

cavity with perturbation slots introduced to independently control mode 2. These two perturbation slots are aligned along X axis with a flaring angle φ with respect to the mushroom center. Such a location of the perturbation slots directs the current path of mode 2 to diverge from its actual path and cover an increased path length, as shown in Figure 12(b). This increase in surface current path reduces the resonant frequency of mode 2 (TE_{102}). These slots have thickness $t_2 = 5.5$ mm and radius $r = 3.5$ mm, and are positioned to intersect the regions of maximum current density for the TE_{102} mode along the x -axis of the mushroom patch. As the slot angle φ increases from 0° to 30° , the resonance frequency of the TE_{102} mode decreases from 4.55 GHz to 4.25 GHz, achieving a 7% tuning range. The TE_{101} mode remains unaffected because these slots are positioned at locations of low current density for the fundamental mode.

The second approach to control the TE_{102} mode uses a pair of linear slots positioned at a distance $r = 3.5$ mm from the mushroom center along the x -axis as shown in Figure 13(a). These linear slots have thickness $t_3 = 1$ mm and variable length l . Increasing the slot length l from 0 mm to 6 mm reduces the resonance frequency of the TE_{102} mode from 4.65 GHz to

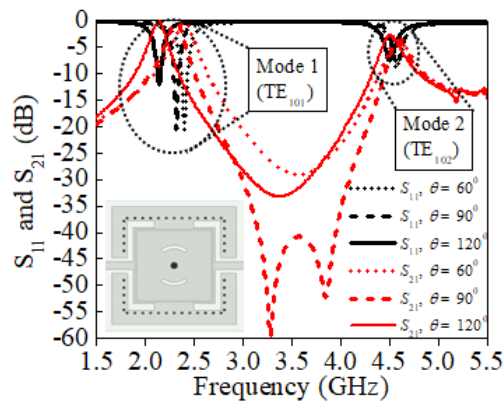


FIGURE 14. Simulated S_{11} and S_{21} of mushroom resonator-loaded SIW cavity with perturbation slots to control mode 1, with $t_1 = 1$ mm and $r = 5$ mm.

4.35 GHz, providing a tuning range of 4.5%. The linear slot configuration offers simpler mechanical implementation compared to arc slots, though with somewhat reduced tuning range.

Again, the TE_{101} mode remains undisturbed due to strategically positioning slots at locations of low current density for the fundamental mode (TE_{101}). These linear slots are parallel to the direction of the surface current density along the X axis in mode 1; conversely, in the case of mode 2, the slot is transverse to the direction of the current density, as shown in Figure 13(b). The independence of mode control is quantitatively verified through parametric electromagnetic simulations. Figure 14 shows the frequency responses of S_{11} and S_{21} when the slot angle, θ of 1 mm thick perturbation slots, placed at $r = 5$ mm is increased at uniform steps from 60° to 120° . Figure 14 clearly shows that an increase in perturbation slot angle, θ , from 60° to 120° decreases the resonance frequency of mode 1 (TE_{101}) from 2.4 GHz to 2.1 GHz, keeping mode 2 (TE_{102}) almost undisturbed. A minimum insertion loss of less than 1 dB is observed during this shift in the resonant frequency of mode 1. Naturally, independent control over the fundamental mode of resonance is attained when the resonant frequency of mode 1 is shifted by 12.5% after changing θ from 60° to 120° . The fabricated samples of such a structure, which controls mode 1, with a set of three perturbation slots having perturbation slot angle, $\theta = 60^\circ$, 90° and 120° , with slot thickness $t_1 = 2$ mm and $r = 5$ mm, are shown in Figure 15. Figure 16 shows the measured frequency response of S_{21} of these fabricated structures. For $\theta = 0^\circ$, we consider the fabricated structure with no perturbation slot, as shown in Figure 7. The measured results in Figure 16 show the lower shift in resonance frequency of mode 1 as the perturbation slot angle, θ , increases. As per the analysis of mode 2, it is expected to remain undisturbed, but the shift observed here can be attributed to fabrication discrepancies. Practical fabrication limitations make discrepancies between the simulated and measured responses almost unavoidable. The minor shift in the resonant frequency of mode 2, which is expected to remain unaffected by variations in the perturbation slot angle, can be primarily attributed to PCB manufacturing tolerances, including deviations in the dimensions of the perturbation slots, mushroom patch, and via locations. Additional dimensional inaccuracies arising during PCB milling

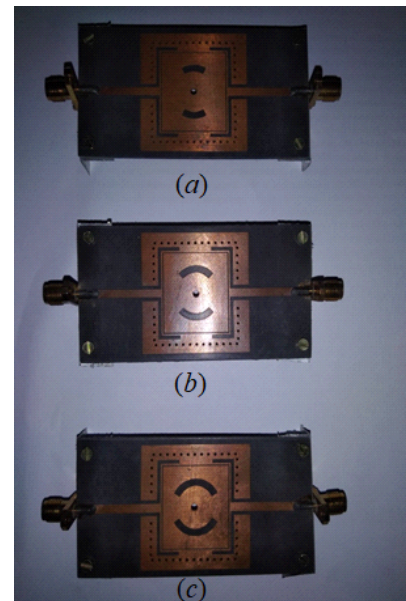


FIGURE 15. Fabricated mushroom resonators with radial perturbation slots loaded into an SIW cavity to vary mode 1 with $t_1 = 2$ mm and $r = 5$ mm, (a) $\theta = 60^\circ$, (b) $\theta = 90^\circ$, and (c) $\theta = 120^\circ$.

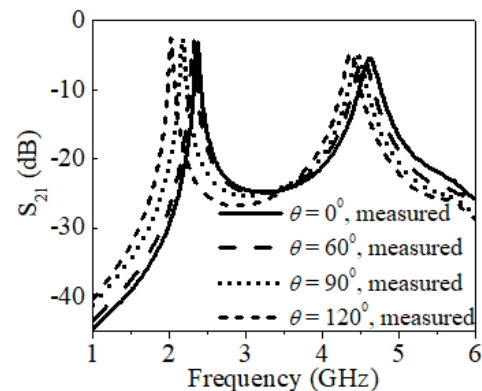


FIGURE 16. Measured S_{21} of mushroom resonator-loaded SIW cavity with perturbation slots to control mode 1, with $t_1 = 2$ mm and $r = 5$ mm.

and metallization, together with variations in substrate parameters such as dielectric constant and substrate thickness from their nominal values, further contribute to deviations in the resonant characteristics. Moreover, assembly-related errors, including slight misalignment of the SMA connectors and soldering imperfections, introduce additional parasitic effects that are not accounted for in the ideal electromagnetic model. The combined effect of these non-idealities leads to slight frequency aberrations while preserving the anticipated modal behaviour of the proposed filter. Figure 17 shows the frequency response of S_{11} and S_{21} for the first approach for controlling mode 2, when the slot angle, φ , of 5.5 mm thick perturbation slots, placed at $r = 3.5$ mm, is increased at uniform steps from 0° to 30° . Resonance frequency of mode 1 (TE_{101}) remains nearly fixed as the increment in φ from 0° to 30° hardly changes the current path of TE_{101} . However, the resonance frequency of mode 2 (TE_{102}) decreases from 4.55 GHz to 4.25 GHz due to an in-

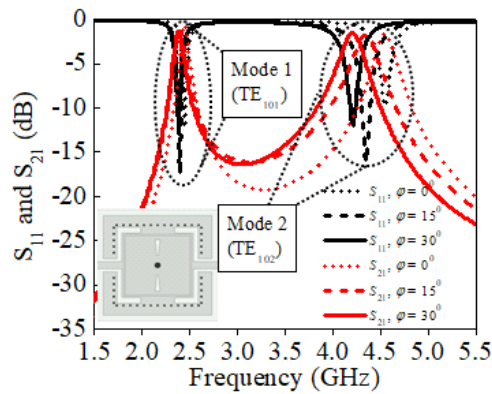


FIGURE 17. Simulated S_{11} and S_{21} of mushroom resonator loaded SIW cavity with perturbation slots to control mode 2, with $t_2 = 5.5$ mm and $r = 3.5$ mm.

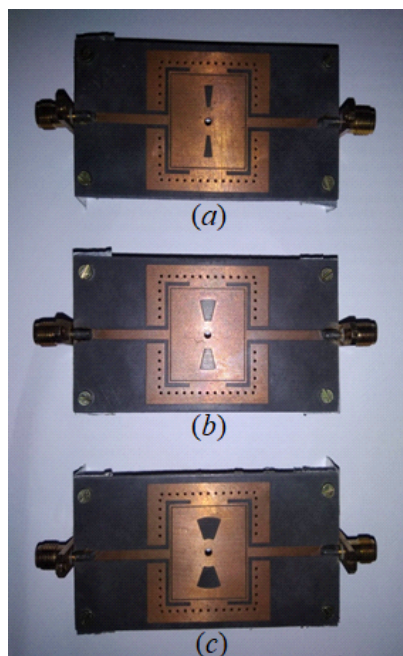


FIGURE 18. Fabricated mushroom resonators with radial perturbation slots loaded to SIW cavity to vary mode 2 with $t_2 = 5.5$ mm and $r = 3.5$ mm, (a) $\varphi = 15^\circ$, (b) $\varphi = 30^\circ$, and (c) $\varphi = 45^\circ$.

crease in the current path as the slot angle, φ , increases from 0° to 30° . So, the resonance frequency of 2^{nd} mode (TE_{102}) with minimum insertion loss less than 2 dB is separately controlled within a range of 7% when φ increases from 0° to 30° . Figure 18 shows a set of three such structures with flaring angle of $\varphi = 15^\circ, 30^\circ$, and 45° , and slot thickness $t_2 = 5.5$ mm and $r = 3.5$ mm, which are fabricated to verify the fact practically. Figure 19 shows measured results of S_{21} frequency responses of these fabricated samples. The lower shift in resonance frequency of mode 2, keeping mode 1 undisturbed, as the flaring angle of the perturbation slot, φ , increases, is observed in Figure 19.

The frequency response of S_{11} and S_{21} for the 2^{nd} approach for controlling mode 2 is illustrated in Figure 20, where the slot length, l of 1 mm thick linear perturbation slots, placed

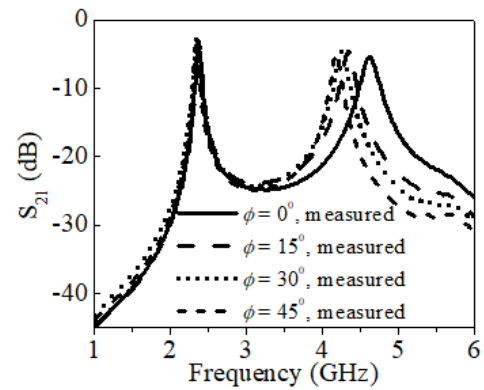


FIGURE 19. Measured S_{21} of mushroom resonator loaded SIW cavity with perturbation slots to control mode 2, with $t_2 = 5.5$ mm and $r = 3.5$ mm.

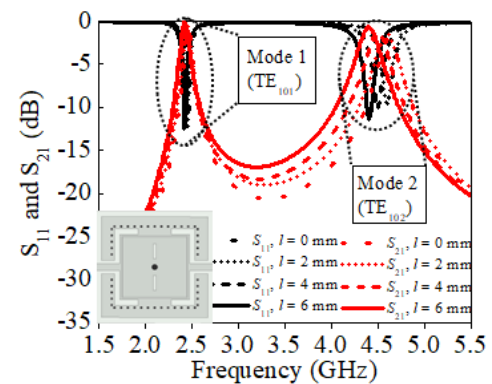


FIGURE 20. Simulated S_{11} and S_{21} of mushroom resonator-loaded SIW cavity with perturbation slots to control mode 2, with $t_3 = 1$ mm and $r = 3.5$ mm.

at $r = 3.5$ mm is increased from 0 mm to 6 mm. This figure shows that the resonance frequency of mode 2 decreases from 4.65 GHz to 4.35 GHz when the length of linear slot is increased by 6 mm. This change in the length of linear slot shifts the resonance frequency of mode 2 by 4.5% without disturbing the resonance frequency of mode 1, as shown in Figure 20.

When the TE_{101} mode is tuned across its full range by varying the arc slot angle θ , the TE_{102} mode frequency variation is less than 15 MHz (0.35% relative variation). Conversely, when the TE_{102} mode is tuned by varying either the arc slot angle φ or the linear slot length l , the TE_{101} mode frequency variation is less than 10 MHz (0.42% relative variation). These marginal cross-coupling effects confirm that the perturbation slots enable truly independent control of the two resonance modes. The physical mechanism underlying this individuality in mode control can be understood via examination of the surface current distributions. For the TE_{101} mode, surface currents flow primarily in closed loops around the patch's perimeter, with maximum density at the edges parallel to the y -axis. The arc slots designed for TE_{101} control are positioned to intersect these current paths, forcing them to bypass the slot boundaries. However, the TE_{102} mode current distribution exhibits a different pattern, with currents flowing in multiple loops along the x -axis and exhibiting a null at the radial distance where the TE_{101} con-

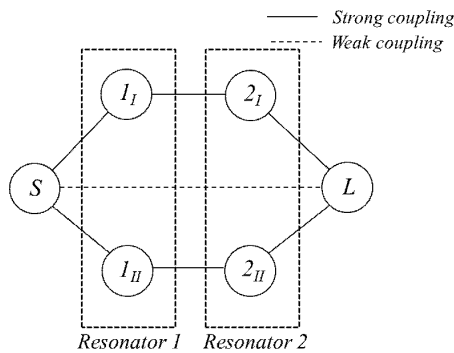


FIGURE 21. Coupling topology of the dual-bandpass filter.

tol slots are positioned. Consequently, the TE_{101} mode control slots have marginal effect on the TE_{102} mode currents. Similarly, the perturbation elements designed for TE_{102} mode control are positioned at locations of maximum current density for the second mode but low current density for the fundamental mode (TE_{101}). This spatial orthogonality of the current distributions for different modes provides the physical basis for independent mode control. A careful selection of perturbation element positions, shapes, and dimensions is critical to achieving mode-control independence while maintaining adequate tuning range for each mode.

2.4. Dual-Band Bandpass Filter Design

This section describes a method to separately reconfigure the passband's CF and BW of a bandpass filter based on SIW loaded with rectangular mushroom resonators (RMRs). The dual-band bandpass filter is constructed by cascading two rectangular mushroom resonators within a single SIW cavity. This configuration creates a second-order filter response for each passband, providing steeper skirt selectivity and improved out-of-band rejection compared to single-resonator designs. Figure 21 shows the coupling topology of the proposed dual-band bandpass filter used to extract the coupling coefficient in order to design the filter. The subscripts I and II denote the first and second passbands, which are dominated by the TE_{101} and TE_{201} modes, respectively. The operating specifications of the first and second passbands are: center frequencies at 2.4 GHz and 4.35 GHz, respectively, 3 dB fractional bandwidth (FBW) of 12.9% and 10.8%, respectively, and 15 dB return loss in both passbands. According to the adapted coupling topology, strong direct couplings exist between the source and Resonator 1 and between Resonator 2 and the load for two separate passbands. The resonators directly couple between them for individual passbands, but there is no coupling between two different modes. Based on specified design requirements, the proposed dual-band bandpass filter is developed through the following sequential design steps.

i) As the first step, the specified center frequency (CF) and FBW are used to determine the required filter quality factor (Q).

ii) Based on the CF, FBW, and calculated Q, a fairly equivalent resonating structure is assumed according to the considered resonator coupling topology.

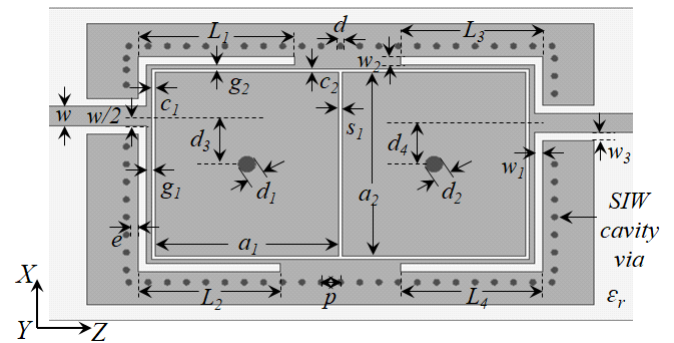


FIGURE 22. Schematic of the RMRs-loaded SIW-based bandpass filter.

iii) Thereafter, depending on this anticipated coupling topology, coupling coefficients are roughly calculated on the basis of the respective coupling mechanism among Source (S), 1st Resonator (1), 2nd Resonator (2), and Load (L). The split frequencies used in the calculation are considered around the CF based on the bandwidth (BW), as mentioned in filter design specification.

iv) Lastly, as mentioned, the optimization method is employed in this design; the roughly extracted coupling coefficients are used as references to optimize coupling parameters in the filter structure using weakly capacitive coupled microstrip feed lines between Source (S) and the 1st Resonator (1), the 1st Resonator (1) and 2nd Resonator (2), and the 2nd Resonator (2) and Load (L).

Finally, full-wave optimization is carried out to achieve the desired bandpass response. The resulting coupling topology confirms direct coupling between the excited modes along the signal path from the source (S) through resonators 1 and 2 to the load (L).

The two RMRs are positioned along the signal propagation direction with a controlled spacing that determines the inter-resonator coupling strength. Figure 22 shows the schematic of the RMRs-loaded SIW-based bandpass filter. The filter structure, as modelled in Ansys Electronic Desktop 2020, is implemented on a double-sided copper-clad Rogers RT/Duroid 5880 substrate with thickness 0.787 mm and relative permittivity $\epsilon_r = 2.2$, maintaining consistency with single-resonator designs. The SIW cavity is formed using via arrays with diameter 1.0 mm and spacing 1.5 mm, providing effective electromagnetic shielding while maintaining compatibility with standard PCB fabrication processes. Input and output coupling is achieved through $50\ \Omega$ microstrip transmission lines that are capacitively coupled to the first and second mushroom resonators, respectively. The mushroom resonator is made of a metallic patch on one surface of a dielectric substrate and shorted by a metallic via to a ground plane attached to the other surface of the dielectric substrate, as shown in Figure 2(a) and Figure 2(b). An eigenmode analysis is carried out to determine RMR's TE_{101} mode excitation at nearly 2.3 GHz. Parametric values of the loaded mushroom resonators in the SIW cavity are fixed as follows: $a_1 = 23.0$ mm, $a_2 = 21.0$ mm, $d_1 = 2.0$ mm, $h = 0.78$ mm, and $\epsilon_r = 2.2$. Two such RMRs are cascaded and loaded to an SIW cavity after etching a rectangular portion from the upper conductor of the SIW cavity. A capacitively coupled

TABLE 1. Optimized schematic parameters of the RMR-loaded bandpass filter.

Parameters	L_1	L_2	L_3	L_4	g_1	g_2	c_1	c_2	w	s_1	d	d_1	d_2	d_3	d_4	w_1	w_2	w_3	a_1	a_2	e	p
Values (mm)	18.32	16.2	16.2	16.2	0.59	0.58	0.3	0.3	2.42	0.2	0.8	2	2	6	4.52	0.9	0.98	1.01	23	21	1	2.2

feeding technique transitioned from a microstrip transmission line of $50\ \Omega$ characteristic impedance, as shown in Figure 22, is employed to excite the RMR-loaded SIW cavity. Table 1 lists optimized schematic parameters of the RMR-loaded bandpass filter shown in Figure 22.

Due to loading two cascaded RMRs in the SIW cavity, a 2nd-order passband response with two poles is observed in the frequency response of S -parameters, as shown in Figure 23. A critical innovation in the filter design is the implementation of extended coupling slot lines between the two cascaded RMRs. These coupling slots, designated as L_1 , L_2 , L_3 , and L_4 , are etched in the ground plane between the two mushroom vias. The slots create controlled capacitive coupling between the resonators, with coupling strength determined by the slot dimensions and positions. The extended slot configuration introduces a transmission zero (TZ) at the lower stopband, significantly enhancing the passbands' lower-edge selectivity. The transmission zero generated by the extended coupling slot lines appears at approximately 1.75 GHz, below the lower edge of the first passband. This transmission zero results from interference between the direct coupling path through coupling slots and the indirect path through resonant modes of the mushroom structures. By carefully designing the slot dimensions and positions, the transmission zero can be positioned to provide maximum improvement in lower-stopband rejection while maintaining low insertion loss in the passband.

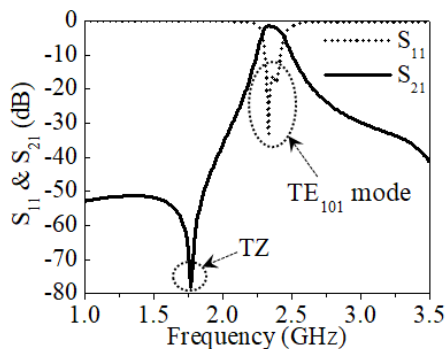


FIGURE 23. Simulated S -parameters of the RMRs-loaded SIW-based bandpass filter.

Figures 24(a) and 24(b) show electric-field distributions and surface-current distributions, respectively, of the RMR-loaded SIW-based bandpass filter at the excited fundamental mode (TE_{101}). The electric-field intensity maximizes at the mushroom periphery, and the surface-current distribution shows its maximum intensity near the mushroom vias and decreases radially towards the periphery.

2.5. Single Passband Control

Independent control over the characteristics of a single passband is demonstrated using perturbation slots and coupling gap adjustment. This capability is essential for applications requiring dynamic reconfiguration of filter parameters to accommodate changing system requirements or environmental conditions.

Center-frequency control of a single passband is achieved by introducing arc-shaped perturbation slots in mushroom resonators, as shown in Figure 25(a). The slots are characterized by thickness t , radius r , and angular extent φ . As the slot angle φ increases, it is clearly shown in Figure 25(b) that surface currents are forced to follow longer paths around slot boundaries, effectively increasing the electrical length of the resonant structure and reducing the resonant frequency. This perturbation mechanism affects the excited mode (TE_{101} in this case) while leaving other modes relatively undisturbed. Parametric analysis reveals that varying the arc slot angle φ from 0° to 90° shifts the passband's center frequency by 14.5%, from approximately 2.1 GHz to 2.35 GHz, as shown in Figure 26. Throughout this tuning range, the 3 dB fractional bandwidth remains nearly constant at approximately 3.5%, demonstrating that the perturbation slots primarily affect the resonance frequency rather than the coupling characteristics. The minimum insertion loss across the tuning range is 0.7 dB, occurring at mid-range slot angles where the current path perturbation is balanced between minimal and excessive deviations. The physical mechanism underlying center-frequency control can be understood via examination of the surface current distributions at different slot angles. At $\varphi = 0^\circ$ (no perturbation), currents flow in the shortest possible paths around the mushroom perimeter, resulting in the highest resonant frequency. As φ increases, the slots force currents to detour around the slot boundaries, increasing the effective current path length. This increased path length is equivalent to increasing the inductance of the resonant structure, which reduces the resonant frequency according to the standard LC resonator formula

$$f = \frac{1}{2\pi\sqrt{LC}} \quad (2)$$

Bandwidth control is achieved independently of center frequency through adjusting the coupling gap, s , between the two cascaded mushroom resonators. The coupling gap directly affects the inter-resonator coupling coefficient, k , which determines the frequency splitting between the two transmission poles of the second-order filter response. According to coupled resonator filter theory, the 3 dB fractional bandwidth is approximately proportional to the coupling coefficient (k) for weakly coupled resonators. The internal electrical coupling between the loaded mushroom resonators across the linear slot

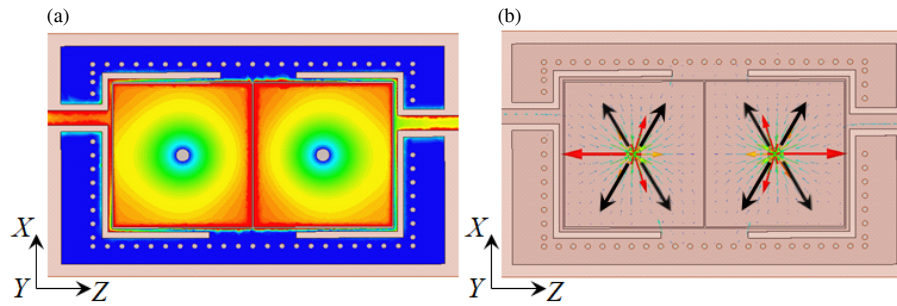


FIGURE 24. (a) Electric-field distributions and (b) Surface current distributions.

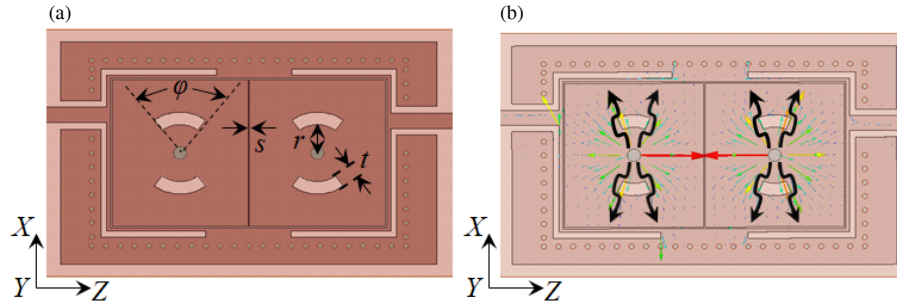


FIGURE 25. (a) Bandpass filter loaded with perturbation slot of $t = 2$ mm, $r = 5.5$ mm and angle, φ , and (b) surface-current distributions after loading perturbation slots.

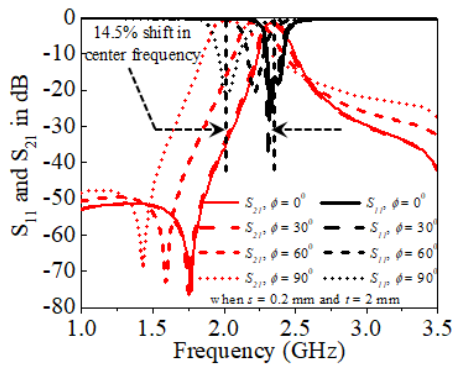


FIGURE 26. Shift in the center frequency by varying the perturbation slot angle (φ).

gap, s , forms the 2^{nd} order dual-pole passband. To understand the effect of a change in k_{101} on passband bandwidth, Equation (3) shows the modified expression of the coupling coefficient (k_{101}) at the TE_{101} mode.

$$k_{101} = \frac{f_2 - f_1}{(f_1 + f_2) - 2 \frac{f_1 f_2}{(f_1 + f_2)}} \quad (3)$$

where f_1 and f_2 are split 1^{st} and 2^{nd} resonances of the cascaded mushroom resonators loaded into an SIW cavity at the TE_{101} mode. As the coupling gap, s , increases from 0.1 mm to 0.5 mm, the coupling coefficient k_{101} decreases, bringing the split resonance frequencies f_1 and f_2 closer together and narrowing the passband bandwidth, as illustrated in Figure 27 and Figure 28. Specifically, increasing the split gap, s , from 0.1 mm

to 0.5 mm reduces the 3 dB bandwidth by 2.9%, from approximately 90 MHz to 87.4 MHz, while the center frequency remains virtually unchanged (variation less than 0.3%). The minimum insertion loss for bandwidth control is 0.8 dB, slightly higher than for center frequency control due to the reduced coupling strength at larger gap values.

The independence of center frequency and bandwidth control is verified via cross-parametric simulations. When the center frequency is tuned across its full range by varying φ , the bandwidth variation is less than 2% for a fixed coupling gap. Conversely, when the bandwidth is adjusted by varying s , the center frequency variation is less than 0.5% for fixed slot angle. This independence enables flexible filter design where center frequency and bandwidth can be optimized separately to meet specific application requirements.

2.6. Post-Tuning of Dual Passbands

The ultimate goal of the filter design is to achieve independent post-fabrication tuning of both passbands simultaneously. This capability is realized through a novel mechanical tuning mechanism based on radial metallic shafts that can be rotated to vary the effective perturbation of each resonance mode.

As illustrated in Figure 29(a), the electric-field intensity increases towards the loaded mushroom peripheries inside the SIW cavity at the TE_{101} mode. At the TE_{201} mode, two half-wavelength electric fields attaining their maximum value along the X axis and one half-wavelength electric field attaining its maximum value along the Z axis are observed, as shown in Figure 29(b). At the TE_{101} mode, the intensity of the surface current distribution over the metal patch of loaded mushroom

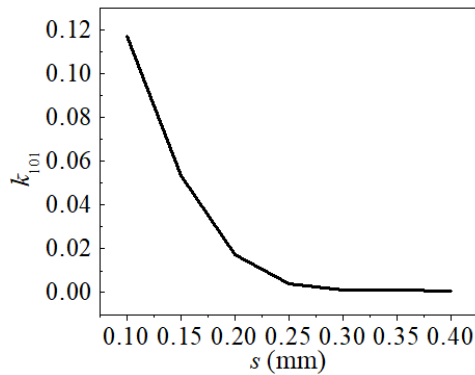


FIGURE 27. Coupling coefficient (k_{101}) vs slot gap (s).

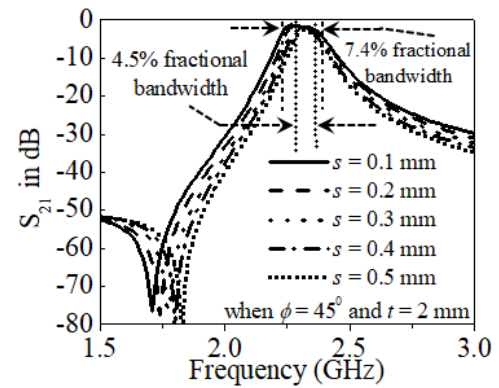


FIGURE 28. Change in the passband bandwidth by varying the coupling slot gap, (s).

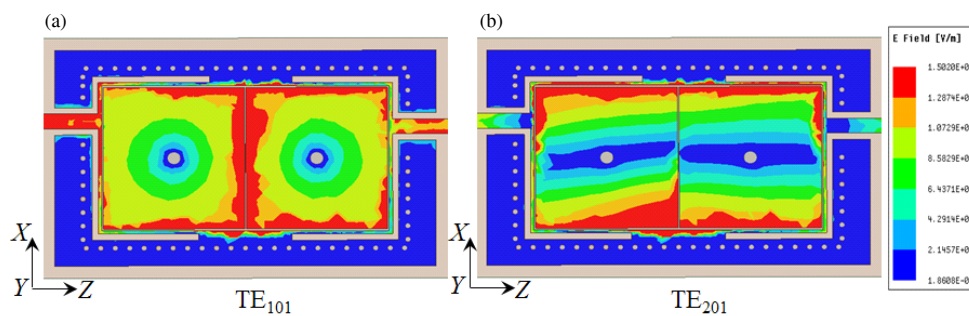


FIGURE 29. Electric field distributions of the dual-band bandpass filter structure at (a) TE_{101} mode and (b) TE_{201} mode.

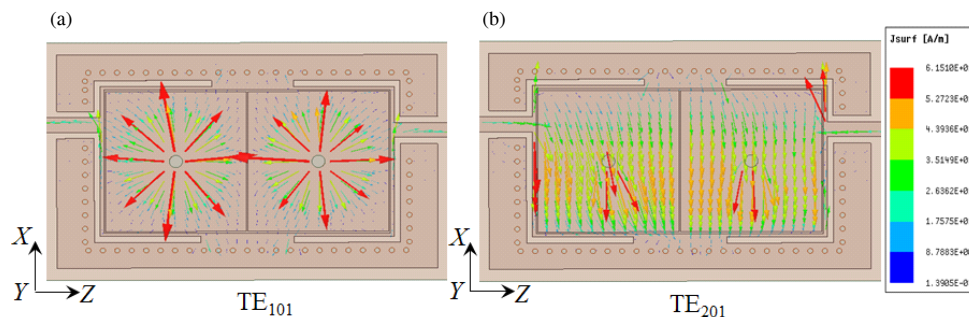


FIGURE 30. Surface current distributions of the dual-band bandpass filter structure at (a) TE_{101} mode and (b) TE_{201} mode.

resonators is maximum near the mushroom via, and it minimizes away from the via, i.e., near the mushroom periphery, as shown in Figure 30(a). In the case of TE_{201} mode, surface-current distributions attain maximum intensity at the region where the electric-field distribution is minimum, as presented in Figure 30(b). These particular orientations of surface-current distributions at two different resonant modes are used to propose the idea of independently tuning passbands.

Specific perturbation slots are introduced into the metal patch of loaded mushroom resonators with typical orientations in order to separately disturb excited modes which are responsible for generating passbands. A pair of radial slots of thickness t_1 , which make an angle φ w.r.t the center of the mushroom via, are placed at a distance r_1 from the mushroom via center along the X axis in both the RMRs, as presented in Figure 31.

Another pair of linear slots with thickness t_2 and length l is placed at a distance r_2 from the mushroom via center along the Z axis in both RMRs, as presented in Figure 31. A careful observation of surface-current distributions at TE_{101} mode, shown in Figure 30(a), helps realize that the pair of radial slots of thickness t_1 is placed nearly in the region of maximum intensity of surface current and transverse to the direction of surface current vector. Similarly, the surface-current distributions at the TE_{201} mode shown in Figure 30(b) helps realize that the pair of linear slots of thickness t_2 is placed in the region of maximum intensity of the surface current and is transverse to the direction of the surface-current vector at the TE_{201} mode.

After introducing these perturbation slots to control excited modes, surface-current distributions at TE_{101} and TE_{201} modes are studied again. Figure 32(a) and Figure 32(b) show the

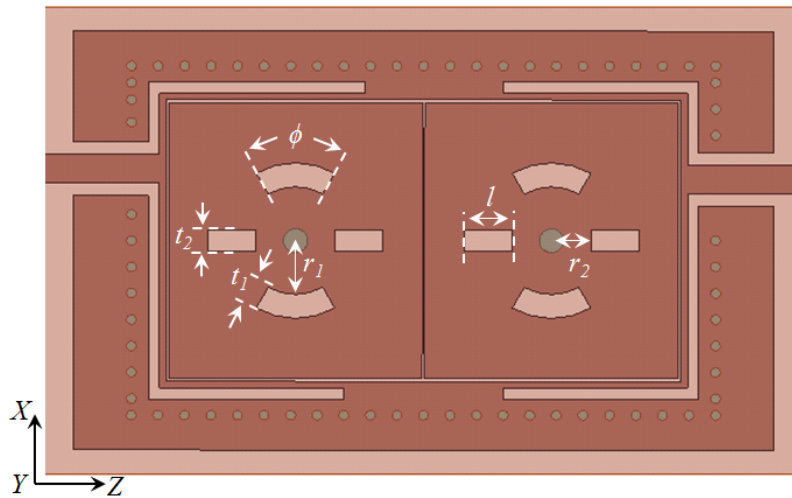


FIGURE 31. The RMRs loaded SIW-based dual-band bandpass filter structure with pairs of perturbation slots loaded to each RMR for independent control over the excited modes with slot parameters, $t_1 = t_2 = 2$ mm, $r_1 = 4.5$ mm, $r_2 = 3$ mm, $l = 4.5$ mm and $\varphi = 60^\circ$ and other parametric values of the structure are as listed in Table 1.

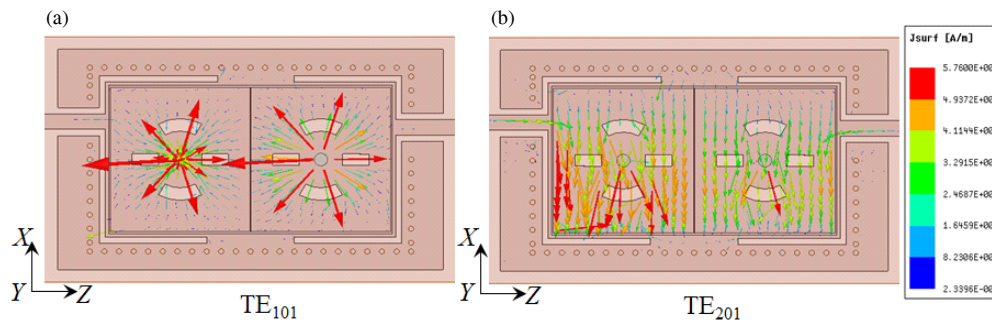


FIGURE 32. Surface current distributions of the dual-band bandpass filter structure at (a) TE_{101} mode and (b) TE_{201} mode after loading the mode perturbation slots.

surface-current distributions after loading those perturbation slots into the metal patch of the RMRs. At the fundamental mode, i.e., the TE_{101} mode, the surface current emerges from the center of the mushroom via with maximum intensity near the vias. The direction of these surface current vectors at the TE_{101} mode is normal to the orientation of radial perturbation slots, as explained earlier. Figure 32(a) clearly illustrates that radial perturbation slots positioned in the region of maximum intensity at the TE_{101} mode force the surface current to diverge from its original path and cover an elongated path length. Therefore, more the radial slot angle, φ increases more the path length is covered by the surface current at the TE_{101} mode. The linear perturbation slot of thickness t_2 and length l is oriented parallel to the surface-current vector at the TE_{101} mode. Consequently, keeping the slot thickness t_2 constant, a change in the slot length, l , hardly makes any deviation in the current path length at the TE_{101} mode. So it is quite understandable that a change in radial slot angle, φ , will disturb the surface current distributions at the TE_{101} mode, which can change the resonant frequency of the excited mode, but a change in slot length l of the linear perturbation slot will hardly affect surface-current distributions at the TE_{101} mode or its corresponding resonance frequency. A detailed observation of the surface current dis-

tributions at the TE_{201} mode shown in Figure 32(b) illustrates that the linear perturbation slots of thickness t_2 and length l are placed exactly in the region of maximum intensity of surface-current distributions at the TE_{201} mode, and the radial perturbation slots of thickness t_1 with an angle φ w.r.t the center of mushroom vias are placed away from the maximum intensity of surface-current distributions at the TE_{201} mode. Therefore, it is not difficult to understand that linear slots transversely oriented to surface-current vectors at the TE_{201} mode will obstruct the current path length when their length l is increased. As a consequence, the TE_{201} mode will be disturbed, resulting in a change in the corresponding resonance frequency. The radial slots making an angle φ , placed in regions of weak intensity of surface current at the TE_{201} mode, are unable to significantly diverge the surface currents at the TE_{201} mode until the radial angle φ is made too large. The technique of diverging the surface currents by changing the parametric values of perturbation slots loaded into the RMRs helps realize independent control over excited modes, which are responsible for generating dual passbands in the dual-band bandpass filter. Thus, it is evident that the radial perturbation slots are useful to control the TE_{101} mode without disturbing TE_{201} mode, and the linear perturba-

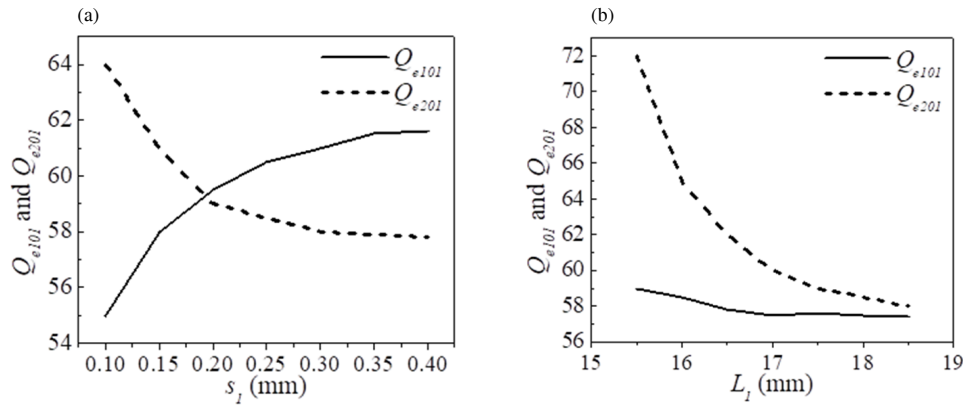


FIGURE 33. Variation of External Quality factors (Q_{e101} and Q_{e201}), (a) with variation in split gap (s_1) and (b) with variation in coupling slotline (L_1).

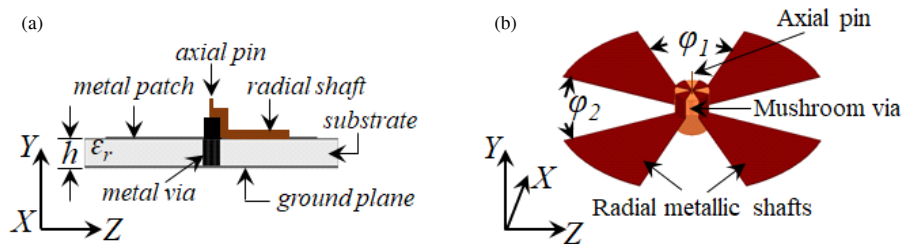


FIGURE 34. The tuning element implemented into the dual-band bandpass filter (a) with a single radial shaft shown in a 2-dimensional view and (b) with four radial shafts, presenting the entire tuning element shown in a 3-dimensional view.

tion slots are useful to control the TE_{201} mode without disturbing the TE_{101} mode.

The signal-coupling path from the source to the load through SIW-integrated mushroom resonators is a critical aspect of the bandpass filter design. Accordingly, once the perturbation slots are verified for effective passband tuning, the external quality factor (Q_e) of the complete dual-band filter is analysed to evaluate the source-resonator coupling characteristics. Q_{e101} and Q_{e201} are external quality factors of the entire filter structure at TE_{101} and TE_{201} modes, respectively. The slot gap s_1 between two cascaded rectangular mushroom resonators control the coupling strength between the individual excited modes during the generation of respective passbands. So, variations of Q_{e101} and Q_{e201} w.r.t slot gap s_1 are observed and presented in Figure 33(a). Here, we found that as the gap between the mushroom resonators, that is, s_1 increases, the electric coupling strength between those resonators decreases, which brings the passband reflection zeros close to each other and increases Q_{e101} initially. However, as the gap increases further, the energy transfer is restricted due to a lack of necessary coupling, and thus, Q_{e101} remains constant thereafter. In the case of Q_{e201} , it initially decreases as the gap s_1 increases, and after a certain value of s_1 , Q_{e201} becomes constant because there is no further change in coupling strength. Coupling slotlines, L_1 , L_2 , L_3 , and L_4 , shown in the schematic of the proposed filter, are used to excite the two modes of this dual mushroom structure. The role of one slotline L_1 in controlling Q_e is realized and explained in this study. Figure 33(b) shows the variation in external quality factors, Q_{e101} and Q_{e201} , for change in L_1 . Both Q_{e101} and Q_{e201} decrease due to increase in L_1 , but

rate of Q_{e201} decrease is clearly prominent compared to that of Q_{e101} which is almost uniform after a certain value of L_1 . Both studies of external quality factor (Q_e) show that the values of Q_{e101} and Q_{e201} at those optimised values of $s_1 = 0.2$ mm and $L_1 = 18.32$ mm come nearly at 58 or 59.

Figure 34(a) shows the cross-sectional view of a single radial shaft acting as a tuning element implemented in a single mushroom resonator. The radial shaft is like a metal patch having a thickness equal to that of a metal cladding, with a given shape of a fraction of the entire circular patch, and is made to rest on the patch of the mushroom resonator, as shown in Figure 34(a). The radial shaft is connected to an axial pin fitted at the top of the extension of the mushroom via along the Y axis, as shown in Figure 34(a). As the radial shaft rests on the metal patch of the mushroom resonator, it is easy to move the metallic radial shaft by rotating the shaft connector about the axial pin either clockwise or anti-clockwise. To increase or decrease the radial angle φ of the radial perturbation slot, the arc length should be increased or decreased at both ends, for which two radial shafts are needed. These radial shafts either cover or expose the radial perturbation slot by decreasing or increasing the angle φ_1 in between. Two additional radial shafts are required as well to vary the other radial perturbation slot placed radially opposite to the former. Therefore, four such radial metallic shafts are employed in each mushroom resonator loaded to the SIW cavity to control the TE_{101} mode and tune the 1st passband. Figure 34(b) shows the entire tuning element composed of four radial metallic shafts resting on the metal patch of the mushroom resonator. These four radial shafts are simultaneously used to vary the length l of the pair of linear slots loaded

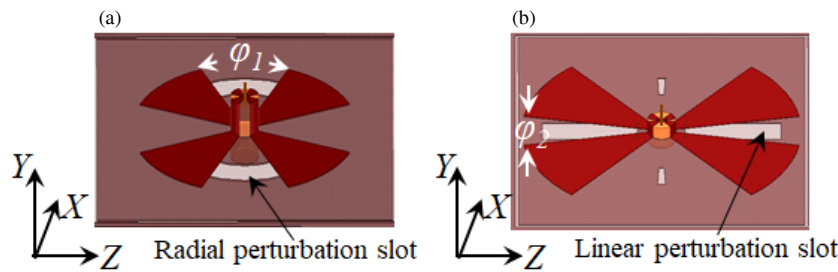


FIGURE 35. Installation of radial metallic shafts in mushroom resonators to control (a) 1st passband by covering the pair of radial perturbation slots and (b) 2nd passband by covering the pair of linear perturbation slots.

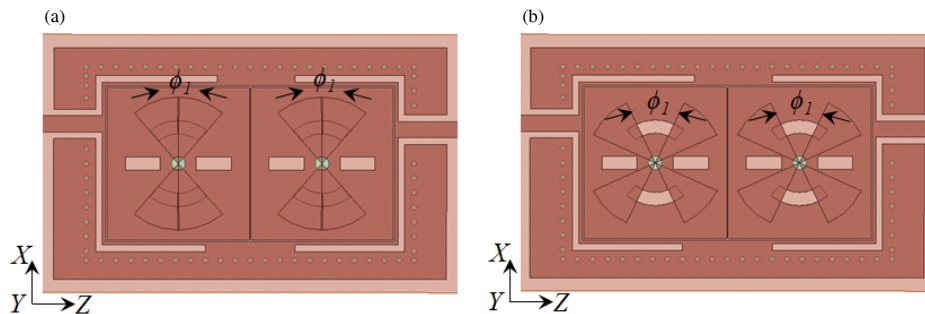


FIGURE 36. Installation of radial metallic shafts in the dual-band bandpass filter (a) $\phi_1 = 20.5^\circ$ and (b) $\phi_1 = 60^\circ$.

to mushroom resonators. If the angle between the radial shafts along the Z-axis, i.e., ϕ_2 , is decreased, the corresponding pair of shafts comes closer to each other and consequently covers the linear perturbation slot placed along Z-axis in the mushroom resonator. On the contrary, linear perturbation slots are exposed when the angle ϕ_2 in between the radial shafts is increased. Therefore, the simultaneous use of these four radial metallic shafts is quite convincing to control the TE₂₀₁ mode and tune the 2nd passband of the dual-band bandpass filter. Figures 35(a) and (b) show the installation of radial metallic shafts in mushroom resonators to control the 1st passband and 2nd passband, respectively.

3. RESULTS AND DISCUSSION

3.1. Tuning of 1st Passband

As explained earlier, the variation of angle φ made by the pair of radial perturbation slots w.r.t the center of mushroom vias is employed to tune the 1st passband. This variation in angle φ_1 is made possible by using the angular separation φ_1 between radial metallic shafts along the X axis. As the radial metallic shafts rotate about the axial pin and come closer to each other, the angle φ_1 decreases, and a cover of metal patch is formed over the radial perturbation slot by decreasing the arc length from both ends of the radial perturbation slot. This variation in angle φ_1 directly covers or exposes the pair of radial perturbation slots and consequently governs the deviation in surface-current path length. This leads to the establishment of control over the 1st passband. Figure 36 shows the structure of this dual-band bandpass filter after implementing the metallic radial shafts into the loaded RMRs as explained in the previous section. Figure 36(a) shows the angle φ_1 in between the two radial

shafts aligned along the X axis, with a value of 20.5° . Such a position of the entire set of radial shafts covers almost the whole arc length of the radial perturbation slots, as presented in Figure 36(a). An increase in angle φ_1 to a value of 60° partially exposes the pairs of radial perturbation slots, as shown in Figure 36(b). These angular positions of the radial metallic shafts are considered keeping in mind not to disturb the linear slot which is responsible for the tuning of 2nd passband. Figure 37 shows surface-current distributions over the upper conductor of the filter structure at TE₁₀₁ and TE₂₀₁ modes for $\varphi_1 = 20.5^\circ$, and Figure 38 shows those for $\varphi_1 = 60^\circ$. Figure 37(a) shows that the path length of current distributions at the TE₁₀₁ mode are completely undeviating over the radial perturbation slots as those slots are almost covered by the radial metallic shafts. Linear perturbation slots are observed not to deviate the surface-current path length at the TE₁₀₁ mode, clearly observed in Figure 37(a). Figure 37(b) shows a deviation in the surface current path length at the TE₂₀₁ mode across the linear perturbation slots when angle $\varphi_1 = 20.5^\circ$. The increase in angle φ_1 further from 20.5° to 60° exposes the pairs of radial perturbation slots partially, as shown in Figure 38. These partial exposures of radial perturbation slots restrict the path of the surface current partially at the TE₁₀₁ mode, as shown in Figure 38(a). The more the value of φ_1 increases, the more the perturbation slot increases, resulting in a larger deviation in surface current at the TE₁₀₁ mode. The linear perturbation slot aligned to the surface-current vector and its dimension being constant hardly disturbs the TE₁₀₁ mode.

At the TE₂₀₁ mode, the partial exposure of these radial perturbation slots does not make any significant deviation in the surface-current path as these slots are positioned in regions of surface-current distributions with lower intensity, which is clearly illustrated in Figure 38(b), whereas the linear perturba-

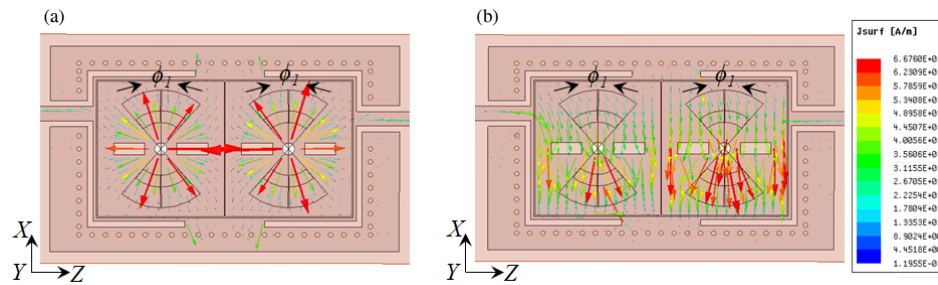


FIGURE 37. Surface current distributions over the upper conductor of the dual-band bandpass filter (a) at TE₁₀₁ mode when $\varphi_1 = 20.5^\circ$ and (b) at TE₂₀₁ mode when $\varphi_1 = 20.5^\circ$.

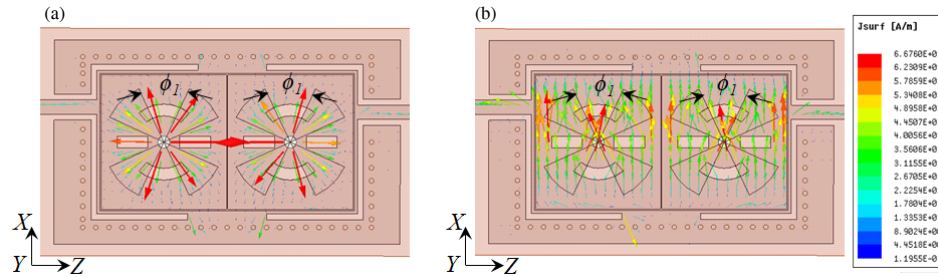


FIGURE 38. Surface current distributions over the upper conductor of the dual-band bandpass filter (a) at TE₁₀₁ mode when $\varphi_1 = 60^\circ$ and (b) at TE₂₀₁ mode when $\varphi_1 = 60^\circ$.

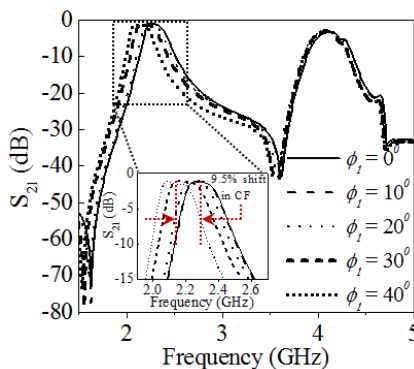


FIGURE 39. Shift in the 1st passband, keeping the 2nd passband undisturbed.

tion slots being transverse to the surface-current vector extensively deviate the current paths at the TE₂₀₁ mode, as shown in Figure 38(b). During the variation of angle φ_1 from the lowest attainable angle to the value for which radial perturbation slots are completely exposed, the dimensions of linear perturbation slots are kept unchanged, due to which the TE₁₀₁ mode is perturbed, and consequently the control over the CF of the 1st passband is established. As no change takes place in the deviation of surface current at the TE₂₀₁ mode, the 2nd passband remains unchanged. Figure 39 shows the S_{21} frequency response characteristics of this dual-band bandpass filter with tuning over CF of the 1st passband. It is clearly illustrated that the increase in angle φ_1 exposes the radial perturbation slot more, and consequently, the surface-current path length elongates due to deviation in path length, which results in the decrease of resonance frequency at the TE₁₀₁ mode. Thus, the CF of 1st passband shifts to lower value as φ_1 increases. It is observed in Figure 39

that a change in angle φ_1 continuously by 40° shifts the CF of 1st passband independently in real time by 9.5% keeping 2nd passband undisturbed. The dimensions of perturbation slots are the same as those in the caption of Figure 31.

3.2. Tuning of the 2nd Passband

As explained earlier, the linear perturbation slots of thickness t_2 and length l are positioned transverse to the direction of the surface-current vector, and naturally, variations in slot dimensions are used to control the TE₂₀₁ mode. The variation in angular separation φ_2 in between the radial metallic shafts along the Z-axis is employed to vary the thickness t_2 of the linear perturbation slot along the Z-axis. As the radial metallic shafts rotate about the axial pin and come closer to each other, the angle φ_2 decreases, and an inclined cover of metal patches is formed over the linear perturbation slots. Figure 40(a) and Figure 40(b) show the positions of these radial metallic shafts with angle $\varphi_2 = 0^\circ$ and 38° which respectively cover linear perturbation slots completely and partially, as illustrated in these figures. The two extreme values of angle φ_2 mentioned above are fixed such that, within this range, the linear perturbation slots will be exposed partially from complete cover, that too without disturbing the dimensions of the radial perturbation slots. A further increase in φ_2 from 38° to a higher value is also possible to completely expose the pairs of linear perturbation slots, as some margin remains for the radial metallic shafts to rotate before the dimensions of the pairs of radial perturbation slots are disturbed. Surface-current distributions at TE₁₀₁ and TE₂₀₁ modes for $\varphi_2 = 0^\circ$ and 38° are studied and presented in Figure 41 and Figure 42. When $\varphi_2 = 0^\circ$, the radial metallic shafts completely cover the linear perturbation slots, for which

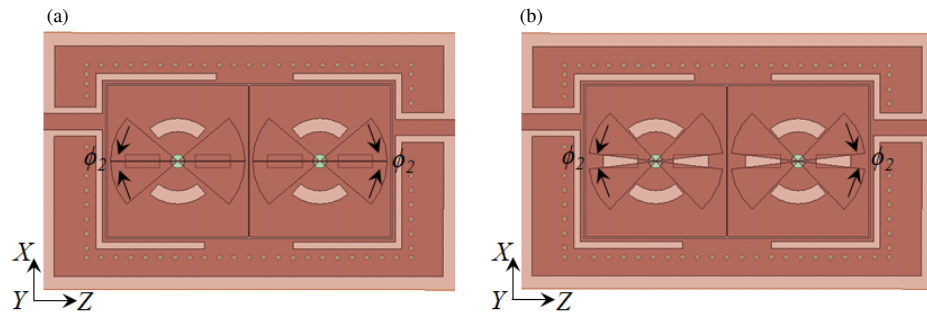


FIGURE 40. Installation of radial metallic shafts in the dual-band bandpass filter (a) $\varphi_2 = 0^\circ$ and (b) $\varphi_2 = 38^\circ$.

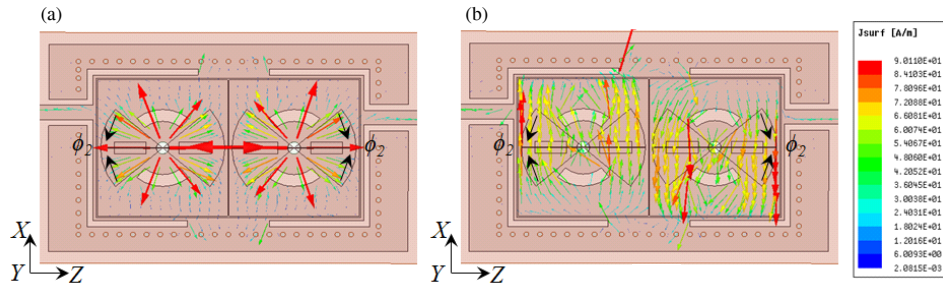


FIGURE 41. Surface current distributions over the upper conductor of the dual-band bandpass filter (a) at TE_{101} mode when $\varphi_2 = 0^\circ$ and (b) at TE_{201} mode when $\varphi_2 = 0^\circ$.

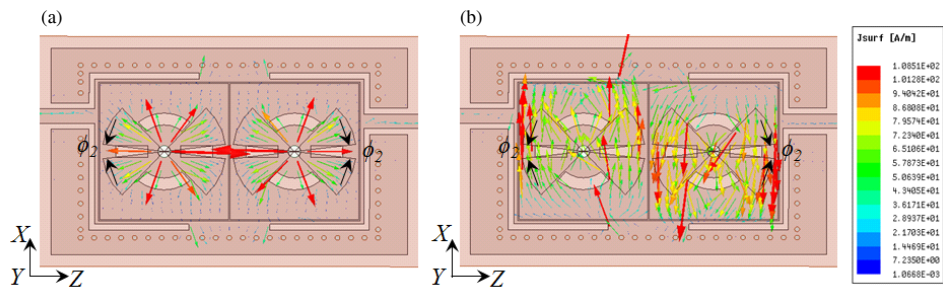


FIGURE 42. Surface current distributions over the upper conductor of the dual-band bandpass filter (a) at TE_{101} mode when $\varphi_2 = 38^\circ$ and (b) at TE_{201} mode when $\varphi_2 = 38^\circ$.

the surface current distributions at the TE_{101} mode remain unperturbed over the region of the linear perturbation slot but face a deviation across the region of the radial perturbation slot, as shown in Figure 41(a). Figure 41(b) shows surface-current distributions at the TE_{201} mode, where the surface current remains undeviated over the region of the linear perturbation slot, but a constant deviation of the surface current occurs across the radial perturbation slot. An increase in φ_2 to a value of 38° after rotating metallic radial shafts exposes the linear perturbation slots partially, but they are aligned with current distributions, and TE_{101} modes are hardly disturbed, as illustrated in Figure 42(a). As φ_2 increases and attains a value of 38° after rotating the metallic radial shafts, the linear perturbation slots are partially exposed and deviate the surface current over the upper conductor of the filter structure at the TE_{201} mode, as shown in Figure 42(b). For both values of φ_2 , the radial perturbation slot offers a constant deviation in the surface current at TE_{101} and TE_{201} modes because the dimensions of the radial perturbation slots are constant. During the variation of angle φ_2 from

0° to the value for which the linear perturbation slots are completely exposed, the dimensions of the radial perturbation slots are kept unchanged, due to which the TE_{201} mode is perturbed, and consequently, the control over the CF of the 2^{nd} passband is established. As no change takes place in the deviation of the surface current at the TE_{101} mode, the 1^{st} passband remains unchanged. Figure 43 shows S_{21} frequency response characteristics of this dual-band bandpass filter with tuning over CF of 2^{nd} passband. It clearly illustrates that the increase in angle φ_2 exposes the linear perturbation slot more in an inclined fashion, and consequently the surface current path length elongates due to deviation in current path length, which results in the decrease of resonance frequency at the TE_{201} mode. Thus, the CF of the 2^{nd} passband shifts to lower value as φ_2 increases. It is observed in Figure 43 that a continuous change in angle φ_2 by 40° shifts the CF of the 2^{nd} passband independently in real time by around 2.5%, keeping the 1^{st} passband undisturbed. The dimensions of the perturbation slots are the same as the initial ones. Therefore, this entire process of rotating installed

TABLE 2. Comparison with other referred tunable bandpass filters.

References	Technology and Independent tuning (Yes/No)	Biasing network requirement (Yes/No)	Continuous and, Mechanical tuning (Yes/No, Yes/No)	CF (GHz) (f_1/f_2)	Order 1 st /2 nd band	FBW (%)	Tuning in CF (%) of 1 st passband	Tuning in CF (%) of 2 nd passband	TZs	IL (dB) 1 st /2 nd /3 rd band	Size $\lambda_g^* \lambda_g$
12	SIW, No	Yes	No, No	1.3	2	—	26%	—	0	7	—
13	SIW, No	Yes	Yes, No	2.14	2	—	60.8%	—	2	3.6	0.19 * 0.11
14	SIEvM Cavity, No	Yes	Yes, No	0.8	2	4	52.4%	—	0	2.05	—
15	HMSIW, Yes	Yes	Yes, No	2.26/4.59	2/2	—	36.4%	32.7%	2	2.33/3.38	0.19 * 0.15
16	Microstrip and SIW, No	Yes	Yes, No	1.2	2	—	17.3%	—	2	—	—
17	SIW, No.	No	Yes, Yes	22.8	2	2.1	10.5%	—	0	3.05	—
18	SIW, No	Yes	No, No	1.4	2	3.7	28%	—	0	4	1.19 * 1.7
19	HMSIW, No	Yes	No, No	1.4	2	—	28%	—	1	2.3	1.0 * 0.7
20	SIW, No	No	Yes, Yes	10.75	1	—	4.56% (1 st method), 5.1% (2 nd method)	—	0	—	—
21	SIW, No	No	Yes, Yes	10	4	8	10%	—	0	—	4.65 * 15.78
22	SIW, No	No	Yes, Yes	7.62	2	—	4.2%	—	2	1.3	1.06 * 1.06
23	SIW, Yes (Mode control)	No	Yes, No	7.71/9.64	2/2	5.4/8.1	6.6%	5.5%	0	1.9/1.65	0.58 * 0.58
24	CMOS Chip-LC Resonator	Yes	No, No	1.5	2	1.5 - 67	—	—	0	—	—
25	Varactor-loaded microstrip and dielectric-loaded cavity, Yes	Yes	Yes, Yes	3.52/26.5	2/2	9/2.8	9.1%	7.9%	4	3.0/3.5	1.16 * 1.01
26 (Design 1)	Waveguide, Yes	No	Yes, Yes	15.1/16.62	2/2	2.42/1.73	7.92%	6.08%	1	—	—
26 (Design 2)	Waveguide, Yes	No	Yes, Yes	14.81/16.69	3/2	2.5/2.58	6.77%	5.16%	1	2.0/2.0	—
29	Coaxial cavity resonator, No	No	Yes, Yes	0.65	4	—	25.8%	—	2	1.4	—
30	Stub loaded resonator and liquid crystal, No	Yes	Yes, No	10.04	3	29.33	11.21%	—	2	2.77	—
32	Varactor diodes + Graphene sheet, No	Yes	Yes, No	0.85	3	—	82.87%	—	0	2.70	—

33	3-D printed liquid metal helical resonator, No	No	Yes, Yes	1.3	1	—	22.14%	—	0	2.1	—
34	SIW, No	Yes	Yes, No	2.3	2	—	34.57%	—	0	1.77	—
35	SIC, No	No	Yes, Yes	3.0	2	6.0	22.38%	—	0	2.5	—
36	SISL, No	Yes	No, No	1.0	2	—	75%	—	2	2.5	0.09 * 0.07
37	Single switchable J inverter, Yes	Yes	No, No	0.7/1.0	2/2	0.15	30%	25%	3	3	—
38	Capacitor Loaded Coupled Lines, Yes	Yes	No, No	2/3.79	2/2	22.2/8.5	43.8%	—	4	1.3/1.28	1.13 * 0.2
39	LPF/HPF Diplexer, No	Yes	No, No	3.5	3	42	74%	—	3	5.7	—
40	Quadrature coupler, Yes	Yes	No, No	1.2/1.7	2/2	35.3/34.6	37%	27.1%	4	4.3/6.1	—
41	SISL, Yes	No	Yes, Yes	6.9/8	4/4	—	—	—	7	0.54	0.61 * 0.51
Proposed work	SIW, Yes	No	Yes, Yes	2.2/4.2	2/2	12.5/10.5	9.5%,	2.65%,	3	1.0/1.8	0.31 * 0.5

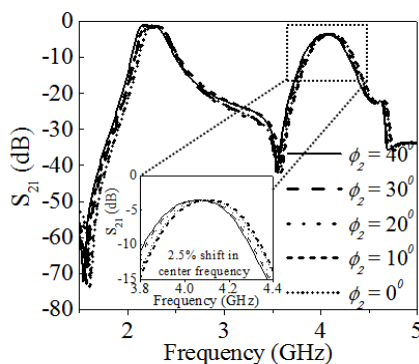


FIGURE 43. Shift in the 2nd passband, keeping the 1st passband undisturbed.

radial metallic shafts separately over both pairs of perturbation slots establishes the independent tuning technique of dual passbands in a dual-band bandpass filter. In practical implementation, factors such as contact resistance, mechanical clearance, friction, and wear may bring small variations in tuning characteristics of the proposed mechanism. However, these effects have not been separately quantified through dedicated contact-resistance, repeated-cycle, or long-term stability measurements in the proposed design. Therefore, the results are considered primarily a tentative validation of the proposed radial-shaft-based tuning concept and its electromagnetic tuning capability.

The proposed technique of independent tuning of center frequencies of this dual-band bandpass filter is comprehensively compared with those significantly tunable state-of-the-art bandpass filters reported by other research groups and is highlighted in Table 2. The comparison includes key performance metrics such as frequency range, tuning range percentage, number of bands, independent tuning capability, passband insertion loss, and fractional bandwidth. In addition, some signif-

icant features, such as the used technology, independent passband tunability, requirement of any biasing network, possibility of continuous tuning of passband CF, and usage of mechanical tuning elements, are also included in this comparative study, as listed in Table 2. Quite considerable and serious insertion losses are observed at passband CFs in those referred tunable filters where biasing networks are used as part of tuning elements [12–15, 18, 19, 25, 30, 32, 36, 37, 39–41]. Pure mechanical elements are introduced in the filter structure to deal with the exposed dimensions of the perturbation slots in order to independently tune both CFs of the passbands, which is unique in itself and novel compared to conventional methods using electrical, electronic, or RF-MEMS tuning elements. As an added advantage, the possibility of continuous tuning is established in both passbands, which is hindered if electrical or electronic tuning elements are employed [12, 18, 19, 24, 36–40]. In comparison to other referred works, center frequencies of both passbands in the proposed bandpass filter are independently tunable. The independent tunability of passband CFs is achieved around 2.2 GHz for the 1st passband and around 4.2 GHz for the 2nd passband for this compact filter structure compared to other referred works with operational frequencies at higher values and larger size [21, 23, 25]. More flexibility is achieved, as far as CF tuning is concerned, for the proposed filter, compared to other referred tunable bandpass filters [20, 22, 23, 25, 26 (both designs)]. Both passbands can be tuned independently in the proposed filter structure compared to [12–14, 16–22, 29, 30, 32–36] where single-passband tuning is only reported. The minimum insertion losses suffered by the 1st and 2nd passbands of the proposed filter are close to 1 dB and 1.8 dB, respectively, which is quite low compared to the referred tunable filters [12–21, 23–26, 30, 32–36]. Both passbands show relatively wider Fractional Bandwidth (FBW) at comparatively lower operating frequencies in this proposed

CF tunability technique than other reported filters that work at higher operating frequencies [14, 17, 18, 21, 23–26, 30, 35]. A compact area of $0.31 \times 0.5\lambda_g^2$ is covered as the footprint by the proposed structure in comparison to other referred works which occupy significantly more area with poor passband selectivity due to fewer transmission zeros (TZs) through the passband response [18, 19, 21–23, 25].

4. CONCLUSION

This work presents a detailed study on independently tunable dual-band bandpass filters using rectangular mushroom resonator-loaded substrate integrated waveguide (SIW) structures. The proposed approach establishes that carefully engineered perturbation slots can selectively manipulate multiple resonant modes independently, enabling controllable dual-band tuning within a compact SIW platform. Unlike conventional techniques relying on complex electronic biasing circuitry or mechanically intensive tuning mechanisms, the proposed method provides a simpler and more efficient solution for achieving reconfigurable dual-band filter operation. One of the principal contributions of this work is the realization of mode-selective perturbation using arc-shaped and linear slots strategically introduced at regions of maximum surface current concentration for the intended resonant modes, while minimizing interaction with modes showing high current density in other regions. A 12.5% frequency tuning for the TE_{101} mode and 7% tuning for the TE_{102} mode are achieved using this approach with minimal mutual interference. The independent tuning behavior originates from the spatial orthogonality of surface current distributions associated with different resonant modes, which allows individual modes to be selectively perturbed with minimal mutual interaction. Furthermore, the proposed dual-band bandpass filter configuration, realized using cascaded rectangular mushroom resonators and extended coupling-slot structures, achieves second-order filtering characteristics for each passband while improving selectivity through the introduction of transmission zeros. Incorporating a transmission zero in the lower stopband provides more than 40 dB attenuation, resulting in a sharp passband-to-stopband transition and improved filtering selectivity. The presented topology further demonstrates that compact planar structures can effectively support high-performance multiband filtering while preserving compatibility with standard PCB manufacturing techniques. By employing radial metallic perturbation shafts, the proposed post-fabrication tuning scheme enables independent adjustment of the lower and upper passbands with measured tuning ranges of 9.5% and 2.5%, respectively. The filter consistently shows average insertion losses close to 1.4 dB and return losses exceeding 15 dB over all tuning conditions, indicating robust impedance matching and stable filtering characteristics. The achieved performance highlights the practicality of the proposed technique for compact and independently tunable dual-band SIW filter implementations. Based on this proposed tunability technique over CFs, the achieved frequency coverage incorporates critical bands for contemporary wireless applications, including Wi-Fi (2.4 GHz), Long Term Evaluation (LTE; 2.3–2.7 GHz), and satellite communications

(4.2–4.4 GHz). The achieved operating range, together with the capability for independent dual-band frequency reconfiguration, makes the proposed filter well suited for wireless platforms requiring simultaneous support for multiple communication standards. Such functionality is particularly advantageous in multi-service communication environments where individual frequency bands must be optimized independently according to system requirements. In mobile and wireless terminals, the proposed filter can facilitate concurrent operation across multiple frequency bands while maintaining tailored filtering characteristics for each band. For satellite and space communication systems, independently tunable passbands offer the flexibility to accommodate frequency allocation changes and compensate for frequency offsets arising from operational conditions. Likewise, in cognitive, adaptive radio networks, dynamically adjusting passband frequencies enables efficient spectrum utilization by mitigating interference and accessing available spectral resources in real time.

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