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A Compact High-Selectivity Bandpass Filter Based on Coaxial Substrate Integrated Waveguide Technology for Ku-band Satellite Communications

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Abstract– This paper presents the design and simulation of a compact Ku-band bandpass filter based on Coaxial Substrate Integrated Waveguide (CSIW) technology for Very Small Aperture Terminal (VSAT) satellite communication systems. The proposed filter employs a shared-cavity cascaded-singlet configuration consisting of electric- and magnetic-coupling structures to generate asymmetric transmission zeros (TZs) below and above the passband, respectively. Owing to the slow-wave effect of the CSIW resonator and the shared-cavity implementation, the proposed structure achieves significant size reduction compared with conventional SIW filters. A key contribution of this work is the independent control of the TZs flanking the passband, allowing for flexible selectivity tuning without disturbing the passband response. The results demonstrate a filter with an insertion loss of less than 1.12 dB and return loss better than 15 dB within the passband, while successfully generating two TZs at 12.67 GHz and 15.55 GHz. The proposed filter therefore provides a compact, low-loss, and highly selective solution for Ku-band VSAT BUC front-end applications.

Keywords– Bandpass filter, CSIW, TZ, singlet, cascaded structure.

1 INTRODUCTION

In recent years, the demand for Ku-band satellite communication systems has increased significantly, particularly in VSAT applications for military, maritime, and civilian purposes. To ensure transmission quality, the BUC in VSAT systems requires bandpass filters that possess low insertion loss, high return loss (good impedance matching), and high selectivity, while maintaining robust out-of-band rejection and high power-handling capabilities. These requirements become even more stringent within the 13.75–14.5 GHz frequency range to comply with current standards for both commercial and military satellite connectivity.

Various filter design solutions have been extensively researched and implemented, including waveguide filters, microstrip filters, and substrate integrated waveguide (SIW) filters. Although conventional waveguide filters [1, 2] offer low loss and high selectivity, they are bulky and difficult to integrate into compact systems. Conversely, microstrip filters [3, 4] are compact and easily integrable but suffer from high losses at high frequencies such as the Ku-band. In this context, SIW filters have emerged as a favorable compromise due to their low loss, ease of fabrication using standard printed circuit board (PCB) processes, and compatibility with high-frequency systems. However, traditional SIW filters [5, 6] still face limitations regard-

ing their ability to generate TZs for enhanced out-of-band rejection, as well as their relatively large overall circuit dimensions.

Unlike conventional SIW filters that operate in the TE_{101} mode with a physical size approximately half of the guided wavelength ($\lambda_g/2$), the proposed CSIW structure utilizes a center metallic via to create a quasi-TEM response [7]. This configuration introduces a significant loading capacitance (C_l), which effectively reduces the guided wavelength based on the slow-wave effect

$$\lambda_g = \frac{v_p}{f} = \frac{1}{f\sqrt{LC_l}}. \quad (1)$$

Consequently, the resonance frequency can be maintained within a much smaller physical footprint compared to standard SIW cavities, providing a fundamental advantage in miniaturization for Ku-band satellite terminals.

To address these limitations, this paper proposes a novel filter structure based on CSIW technology. This structure significantly reduces the filter's footprint, making it highly suitable for system integration, while simultaneously enabling the generation of two TZs to enhance selectivity. The remainder of this paper is organized as follows: Section II discusses the principle of TZ generation and the proposed filter structure. The specific design and simulation results are presented in Section III. Finally, Section IV concludes the paper.

2 PRINCIPLE AND PROPOSED STRUCTURE

2.1 Singlet

The singlet structure [8, 9] consists of a single resonator coupled to both the source and the load, together with a direct coupling path between the source and the load. The normalized coupling matrix of a singlet is given by

$$M = \begin{bmatrix} 0 & m_{sr} & m_{sl} \\ m_{sr} & 0 & m_{rl} \\ m_{sl} & m_{rl} & 0 \end{bmatrix}. \quad (2)$$

In the commonly encountered symmetric case

$$m_{sr} = m_{rl}. \quad (3)$$

The singlet is capable of generating a TZ by introducing a direct source-to-load (S-L) bypass path in parallel with the main path from the source through the resonator to the load (S-R-L). Due to these two parallel signal paths, a TZ is formed at the output when the signals between the paths exhibit a 180° phase difference (phase cancellation). By enforcing the cancellation condition of the transmission amplitude, the TZ location in the normalized frequency domain Ω can be expressed as

$$\Omega_{TZ} = \frac{m_{sr}m_{rl}}{m_{sl}}. \quad (4)$$

For a symmetric singlet, this reduces to

$$\Omega_{TZ} = \frac{m_{sr}^2}{m_{sl}}. \quad (5)$$

This expression indicates that the TZ position is governed by the ratio between the resonant coupling and the direct source-load coupling, the sign of m_{sl} determines whether the TZ appears above or below the passband, increasing $|m_{sl}|$ shifts the TZ closer to the passband. Therefore, the singlet topology provides an effective mechanism for controlling TZs through coupling engineering.

The characteristics of the TZ depend on the nature of the coupling between the source and the load:

- **Electric coupling:** Generates a TZ below the resonant frequency;
- **Magnetic coupling:** Generates a TZ above the resonant frequency.

To analytically reproduce the frequency responses shown in Figure 1, the singlet structures can be described using the coupling-matrix representation. For a center frequency $f_0 = 14.125$ GHz, two different positions of the TZ are selected in order to illustrate electric and magnetic coupling behaviors.

When the TZ is located below the passband (e.g., $f_{TZ} = 13$ GHz), the singlet is modeled by the following normalized coupling matrix

$$M = \begin{bmatrix} 0 & 0.85 & -0.54 \\ 0.85 & 0 & 0.85 \\ -0.54 & 0.85 & 0 \end{bmatrix}. \quad (6)$$

In this case, the negative sign of the direct coupling term m_{sl} indicates a capacitive (electric) cross-coupling.

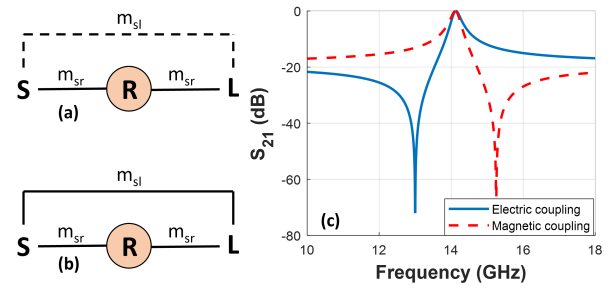


Figure 1. Schematic of the singlet with: (a) Electric coupling, (b) Magnetic coupling, (c) Frequency responses.

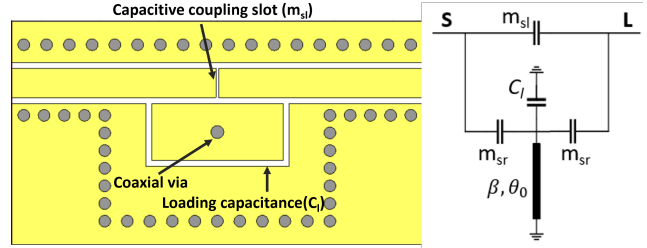


Figure 2. Electric coupling CSIW singlet and equivalent circuit.

The destructive interference between the main resonant path and the direct path produces a TZ on the lower side of the passband.

When the TZ is placed above the passband (e.g., $f_{TZ} = 15.25$ GHz), the coupling matrix becomes

$$M = \begin{bmatrix} 0 & 0.85 & 0.54 \\ 0.85 & 0 & 0.85 \\ 0.54 & 0.85 & 0 \end{bmatrix}. \quad (7)$$

Here, the positive direct coupling coefficient corresponds to inductive (magnetic) coupling. Changing the sign of the cross-coupling shifts the TZ to the upper side of the passband.

2.1.1 Electric coupling: Figure 2 illustrates the electric coupling singlet, it consists of two paths: the main path and the direct path. The main path incorporates a CSIW resonant cavity comprising a coaxial via (center via) and a loading capacitance (C_l), which is coupled to the source/load through capacitive coupling slots (m_{sr}) following the S-R-L path. The direct S-L path is implemented via a capacitive coupling slot (m_{sl}) between the source and the load, creating a signal path in parallel with the main path. These capacitive coupling slots are modeled as equivalent capacitors.

For an input signal below the resonant frequency f_0 (i.e., below the passband), the phase shift induced by the resonant cavity approaches $+90^\circ$. Electric coupling, represented by series capacitors, introduces an additional $+90^\circ$ phase shift to the input signal. Consequently, the total phase shift of the S-R-L main path with m_{sr} capacitive coupling is $+90^\circ + 90^\circ + 90^\circ = +270^\circ$. Meanwhile, the direct S-L path exhibits a phase shift of $+90^\circ$. At the output, the signals combined from these two branches exhibit a 180° phase difference, thereby generating a TZ below the resonant frequency.

To validate the proposed mechanism, an equivalent circuit model is implemented in ADS, as shown in

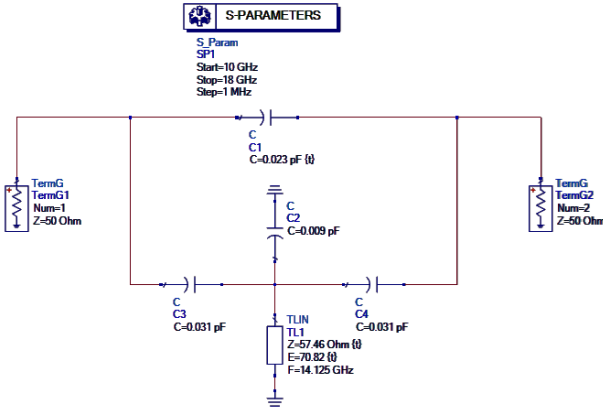


Figure 3. Equivalent circuit model of the proposed singlet in ADS.

Figure 3. The capacitive slot is modeled as a lumped capacitor, while the CSIW resonator is represented by a transmission-line section loaded with a shunt capacitance. To establish the equivalent-circuit model, the loading capacitance C_2 was first estimated using a parallel-plate capacitive-loading approximation of the CSIW resonator, resulting in an initial value of approximately 0.013 pF. The transmission-line section TL_1 was then determined to realize the desired resonant frequency f_0 . Afterward, the remaining coupling capacitors were adjusted to obtain the desired TZ locations and passband response. Finally, all parameters were fine-tuned in ADS to achieve good agreement with the CST full-wave simulation results.

Figure 4 compares the S-parameters obtained from CST full-wave simulation (a) and ADS equivalent-circuit modeling (b). It can be observed that both approaches exhibit a resonant frequency at 14.125 GHz and a TZ around 12.54 GHz. The overall response trends are consistent, confirming that the equivalent circuit successfully captures the essential coupling behavior of the CSIW singlet.

Figure 5 illustrates the simulated E-field distribution at the TZ frequency, f_{1TZ} . It is observed that a significant electric field intensity exists within the structure, particularly along the capacitive coupling slots and inside the resonant cavity. However, the field is effectively canceled out at the output port. This phenomenon confirms that energy is still stored and transmitted through both the main path and the direct path; nevertheless, due to their out-of-phase relationship, they undergo destructive interference at the output, resulting in the formation of a TZ at this specific frequency.

2.1.2 Magnetic coupling: Figure 6 illustrates the magnetic coupling singlet. Similar to the electric coupling structure, this configuration consists of two paths: the main path and the direct path. The main path features a CSIW resonant cavity with a coaxial via and loading capacitance (C_l), which is magnetically coupled to the source/load through the coaxial via and inductive coupling vias. The direct S-L path transmits signals from the source to the load via two inductive coupling vias (m_{sl}), creating a parallel signal path.

For an input signal above the resonant frequency f_0 (i.e., above the passband), the phase shift induced by

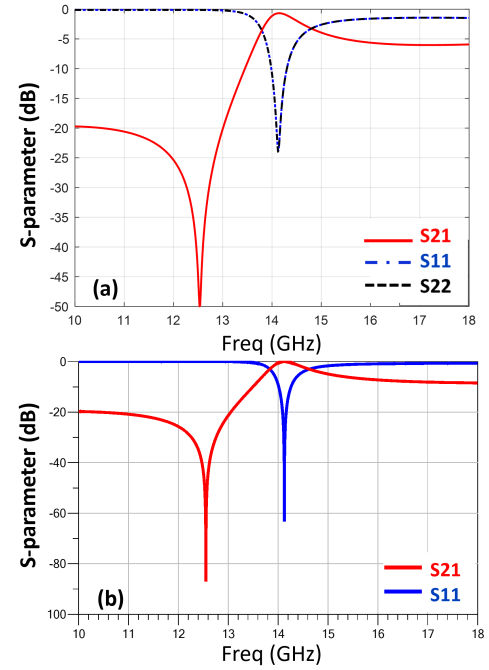


Figure 4. Comparison between full-wave CST simulation and ADS equivalent-circuit results of the proposed singlet: (a) CST-simulated S-parameters; (b) ADS-simulated S-parameters.

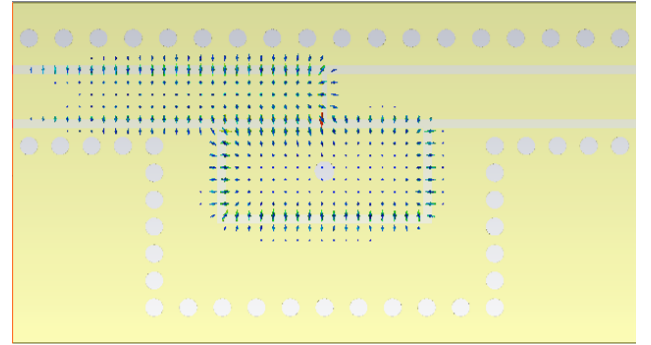


Figure 5. E-field distribution of the electric coupling CSIW singlet at $f_{1TZ} = 12.54$ GHz.

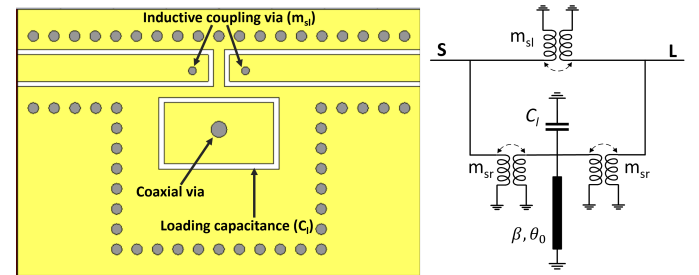


Figure 6. Magnetic coupling CSIW singlet and equivalent circuit.

the resonant cavity is -90° . Magnetic coupling, represented by series inductors, introduces a -90° phase shift to the input signal. Consequently, the total phase shift of the S-R-L main branch with m_{sl} magnetic coupling is $-90^\circ - 90^\circ - 90^\circ = -270^\circ$. Meanwhile, the direct S-L path exhibits a phase shift of -90° . At the output, the signals from both branches are combined with a 180° phase difference, thereby generating a TZ above the resonant frequency.

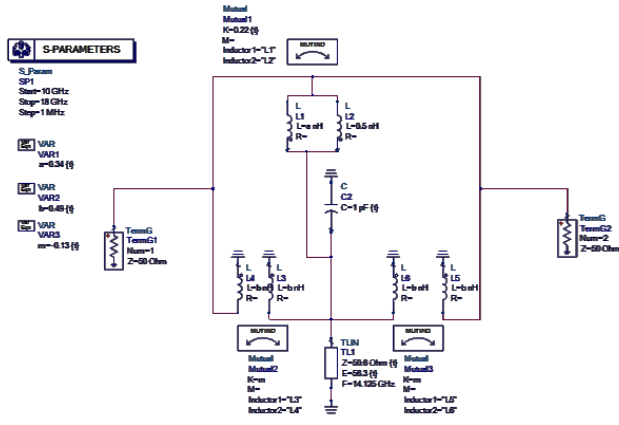


Figure 7. Equivalent circuit model of the magnetic coupling CSIW singlet in ADS.

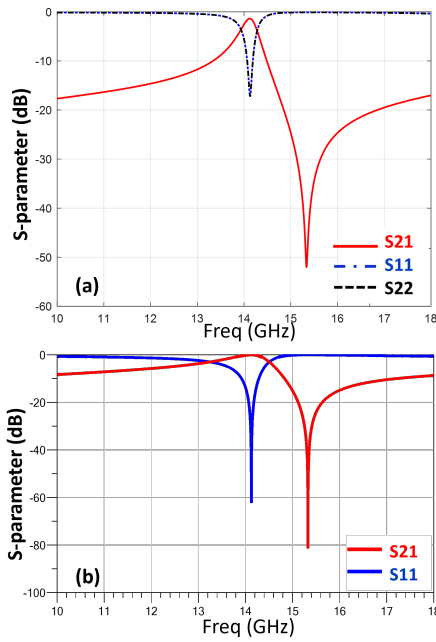


Figure 8. Comparison between full-wave CST simulation and ADS equivalent-circuit results of the magnetic coupling CSIW singlet: (a) CST; (b) ADS.

To validate the proposed magnetic coupling mechanism, an equivalent circuit model is implemented in ADS, as shown in Figure 7. The CSIW cavity is modeled by a transmission-line section with a shunt capacitance, while the magnetic source-load coupling is represented by mutually coupled inductors. The initial value of C_2 was first estimated from the equivalent capacitive loading of the CSIW resonator (approximately 0.02 pF), followed by the determination of TL_1 for the desired resonant frequency f_0 . The inductive coupling parameters were then adjusted for the desired TZ response. Finally, all parameters were fine-tuned in ADS to achieve good agreement with the CST full-wave simulation results.

Figure 8 compares the S-parameters of the magnetic coupling singlet obtained from CST full-wave simulation (a) and ADS equivalent-circuit modeling (b). Both results exhibit a resonant frequency around 14.125 GHz and a TZ located above the passband at 15.33 GHz.

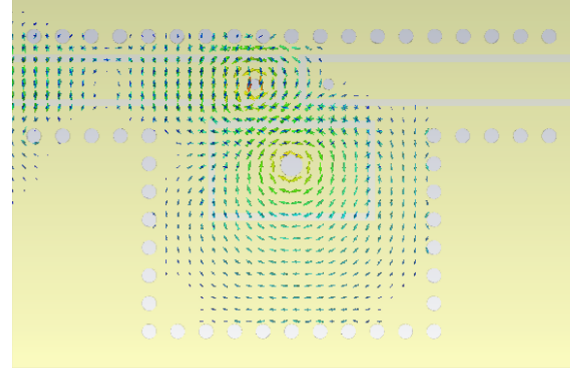


Figure 9. H-field distribution of the magnetic coupling CSIW singlet at $f_{2TZ} = 15.33$ GHz.

The overall response characteristics show good agreement, validating the proposed magnetic source-load coupling mechanism.

Figure 9 presents the simulated H-field at f_{2TZ} , where a clear cancellation effect is observed at the output port. This occurs because the magnetic field vectors from the inductive coupling via and the central coaxial via are in direct opposition. Despite the energy being maintained within the individual branches, their reciprocal interference at the output leads to the emergence of the TZ.

2.2 Cascaded structure

The previous sections investigated the singlets as well as the characteristics of each type, including electric and magnetic coupling. To enhance selectivity and simultaneously create two asymmetric TZs around the passband, the structure proposed in this paper utilizes a method of cascading two singlets, consisting of one electric coupling singlet and one magnetic coupling singlet. The filter schematic is shown in Figure 10.

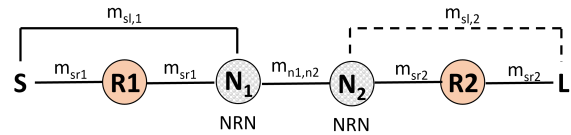


Figure 10. Filter schematic.

In the filter schematic, the signal from the source port is fed through the first singlet, then passes through two non-resonating nodes (NRNs) implemented by $\lambda/4$ transmission line sections, before reaching the second singlet and finally the load port. These NRNs act as intermediates, ensuring controlled signal propagation without introducing unwanted parasitic resonances.

3 FILTER DESIGN

3.1 Design

The filter is designed on a Rogers RO4003C laminate, with a relative permittivity of $\epsilon_r = 3.55$, a thickness of $h = 0.813$ mm, and a loss tangent of $\tan \delta = 0.0027$. The filter structure is illustrated in Figure 11. The filter is

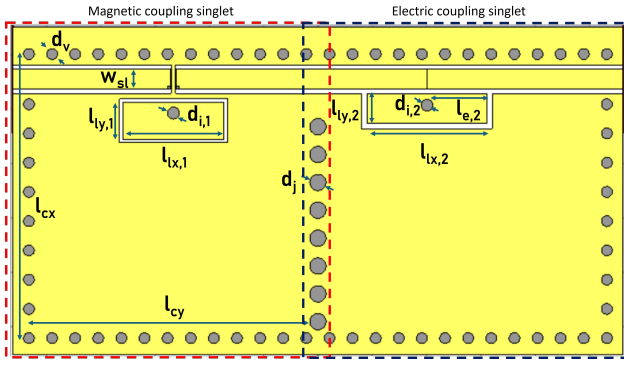


Figure 11. Layout of the cascaded filter with TZs on both sides of the passband.

realized by cascading a magnetic and an electric CSIW singlet within a shared cavity. Each singlet comprises a CSIW resonator coupled to the ports and a direct source-load bypass path that generates a TZ. The intermediate via row connects the two sections, forming a compact dual-TZ cascaded filter. Key dimensions of the filter are summarized in Table I.

Table I
MAIN DESIGN PARAMETERS OF THE CASCADED FILTER (UNIT: MM)

Parameter	Value (mm)	Parameter	Value (mm)
w_{sl}	0.36	d_v	0.2
$l_{e,2}$	1.1	$d_{i,1}$	0.22
$l_{x,1}$	1.862	$d_{i,2}$	0.22
$l_{x,2}$	2.2	d_j	0.3
$l_{y,2}$	0.545	l_{cx}	5.3
$l_{y,1}$	0.675	l_{cy}	5.2

A key contribution of this work is the integrated shared-cavity configuration. Instead of utilizing discrete, separated resonators as in many conventional cascaded SIW topologies, the proposed design integrates both the electric-coupling and magnetic-coupling singlets within a single physical footprint. By sharing the center via fences, the redundant space between resonators is eliminated, resulting in approximately 89% smaller footprint compared with the SIW filter reported in [5] under similar Ku-band operation.

3.2 Parametric analysis of the cascaded filter

The singlet and the cascaded filter were simulated using CST Microwave Studio 2024 with the frequency domain solver. Boundary conditions were set to Perfect Electric Conductor (PEC) for the bottom plane and Open for the top plane.

Figure 12 (a)-(c) demonstrate the tuning behavior of the cascaded singlet structure. The resonator dimension, l_{cx} , mainly determines the passband location, whereas the two bypass couplings independently control the lower and upper TZs. The lower TZ is primarily governed by the electric bypass coupling (g_{sl2}), while the upper TZ is independently tuned via the magnetic bypass coupling (s_{sl1}) [9]. Notably, shifting these TZs does not significantly disturb the center frequency or

the passband impedance matching. This flexibility allows for a highly selective “fine-tuning” of the filter’s rejection characteristics without requiring a complete redesign of the core structure.

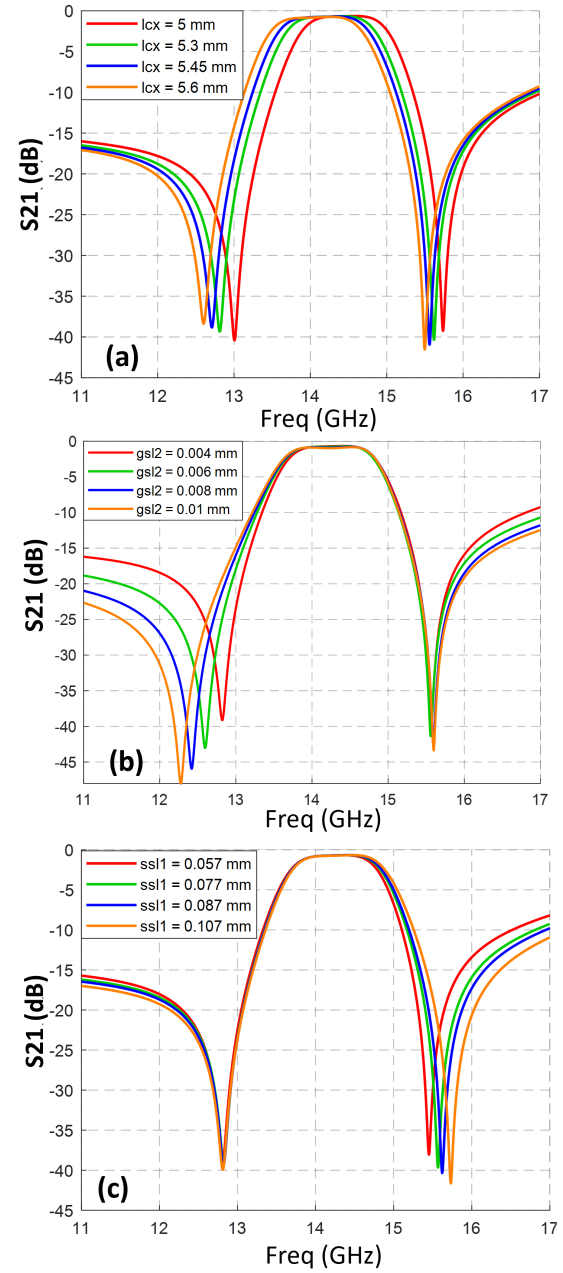


Figure 12. Parametric study of the cascaded filter demonstrating independent control of the center frequency and TZs: (a) tuning of f_0 via l_{cx} ; (b) lower TZ tuning via g_{sl2} ; (c) upper TZ tuning via s_{sl1} .

3.3 Final optimized result

Based on the above analysis, the final dimensions are chosen to achieve the desired center frequency and TZs. The corresponding simulated S-parameters are presented in Figure 13.

The simulation results indicate that the cascaded filter achieves a bandpass response covering the 13.75–14.5 GHz range, with an insertion loss maintained below 1.12 dB across the entire passband. The

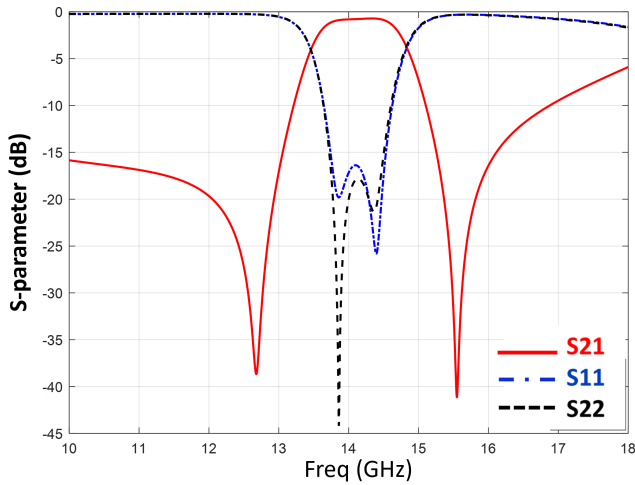


Figure 13. Simulated S-parameters of the cascaded filter.

return loss at the input and output ports is consistently greater than 15 dB, demonstrating excellent impedance matching. Furthermore, two distinct TZs are formed at 12.67 GHz and 15.55 GHz. These TZs significantly increase selectivity of the filter, thereby enhancing out-of-band rejection and ensuring effective suppression of adjacent interference.

Regarding power handling, the filter is intended for the pre-amplifier stage of a VSAT BUC, where it processes small signals before final amplification. Given that typical VSAT BUCs operate at 8 W, the CSIW structure offers a massive safety margin, as similar SIW technologies can withstand up to 450 W [10]. This confirms the design's reliability against thermal and breakdown risks in the target application. Table II provides a performance comparison between the proposed filter structure and various SIW-based filters reported in previous studies. It is evident that the proposed design achieves a significantly more compact footprint while maintaining competitive performance in terms of bandwidth, selectivity, and insertion loss.

Table II
COMPARISON WITH OTHER PUBLISHED WORKS

Ref.		[9]	[5]	[6]	This work
f_0 (GHz)		7.5	13.34	14.1	14.125
FBW (%)		5	5.8	17.02	5.3
IL (dB) @BW		< 2.8	1.8 (@ f_0)	< 2	< 1.4
RL (dB) @BW		> 15	19.4 (@ f_0)	> 10	> 13
30-dB Rejection	Lower ($f_0 - \Delta f_0$ %)	13.33	7.9	NI	9.67
	Upper ($f_0 + \Delta f_0$ %)	13.33	5.9	NI	8.81
Size (λ_0^2)		0.245 $\times 0.49$	0.863 $\times 1.614$	0.635 $\times 3.287$	0.283 $\times 0.527$
Substrate		RO 4003C	PTFE ZYF255DA	RO 3003	RO 4003C
Technology		SIW	SIW	SIW	CSIW

Note: Ref.: Reference; FBW: Fractional Bandwidth; RL: Return Loss; IL: Insertion Loss; NI: No Information.

4 CONCLUSION

This paper presented a design method for a Ku-band bandpass filter based on the CSIW structure. By strategically cascading electric and magnetic coupling structures within a shared-cavity configuration, the proposed filter achieves two controllable and asymmetric TZs at 12.67 GHz and 15.55 GHz. This approach not only provides high selectivity and high out-of-band rejection but also results in a significant footprint reduction of approximately 89% compared to conventional SIW designs. Simulation results demonstrate that the proposed filter achieves an insertion loss maintained below 1.12 dB within the 13.75–14.5 GHz range. The reflection coefficients at both input and output ports remain below –15 dB, indicating excellent impedance matching. Beyond its electrical performance, the CSIW structure offers a compact footprint and ease of fabrication using standard PCB processes, making it a cost-effective solution for mass production. Future work will focus on the fabrication and measurement of the prototype using high-frequency characterization tools to fully verify the slow-wave effects and practical performance.

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