

S-Band Fully Canonical Dual-Mode BaTiO₃ Ceramic Filter

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Abstract—This work presents the design and fabrication of a narrowband, fully canonical, fourth-order monolithic filter with fourth transmission zeros (TZs) operating in the S-band. The device is based on dielectric resonators made of barium titanate (BaTiO₃) ceramics. The TE₁₁₀, TE₁₂₀, and TE₂₁₀ modes are employed to generate the poles and TZs. The resonant frequencies are tuned by modifying the chamfer lengths at opposite corners of the square resonator. The coupling between cavities is implemented through an iris structure strategically arranged to fully exploit the capabilities of the degenerate dual modes. The experimental results show excellent agreement with the simulations.

Index Terms—Barium titanate (BaTiO₃), dual-mode, electroplating, narrowband, resonator, sputtering.

I. INTRODUCTION

THE need to develop compact and efficient devices is critical in 5G/6G base stations and satellite applications, where system miniaturization and weight reduction are driven by decreasing the size of individual components. Dielectric-loaded cavity filters enable significant volume reduction; however, they typically require metallic housings and, in some cases, low-permittivity supporting materials [1], [2], [3], [4]. To further reduce volume and mass, metallized ceramic dielectric filters have emerged as a promising solution, offering miniaturization, low insertion loss, and high quality factor. Several works [5], [6], [7] report fully dielectric-loaded dual-mode resonators implementing filters with two poles and one transmission zero (TZ), while other studies explore metallized ceramic dielectric waveguide resonators in monoblock configurations [8], [9].

In this article, a fourth-order filter with four transmission zeros (TZs) is proposed, obtained by concatenating basic building blocks providing two poles and two TZs. The resonator geometry is fabricated using high-pressure abrasive waterjet cutting, enabling high-dimensional accuracy without external tuning elements. The metallization process combines gold sputtering to form a seed layer with subsequent copper electroplating to increase thickness, ensuring high conductivity and enhanced resonator quality factor.

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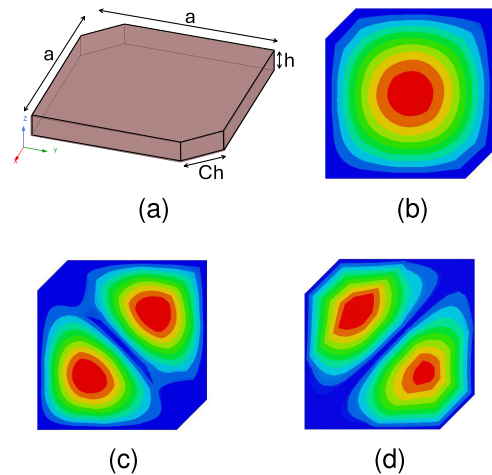


Fig. 1. Dual-mode resonator. (a) Cavity dimensions. (b) Electric field distribution TE₁₁₀ mode. (c) TE₁₂₀ mode. (d) TE₂₁₀ mode.

II. DUAL-MODE RESONATOR ANALYSIS

For functional validation, this work proposes a dual-TE-mode filter based on a monolithic ceramic structure composed of barium titanate (BaTiO₃), which exhibits a relative permittivity of $\epsilon_r = 37$ and a loss tangent of $\tan \delta = 1.5 \times 10^{-4}$. The geometry consists of a fully metallized square prism with a surface whose conductivity was measured experimentally as $\sigma = 4.1 \times 10^7$ S/m, shown in Fig. 1(a), with dimensions $a = 23$ mm, $ch = 4.5$ mm, and a total height of $h = 3$ mm. This structure, resembling a waveguide section, supports the fundamental TE₁₁₀ mode, as shown in Fig. 1(b), along with the higher-order degenerate modes TE₁₂₀ and TE₂₁₀, shown in Fig. 1(c) and (d), respectively. These modes share the same resonant frequency but exhibit orthogonal field distributions due to the structure's symmetry [10], enabling dual-mode operation within a single physical resonator.

To achieve mode splitting, perturbations are introduced in the operating conditions. In this work, controlled geometric modifications are implemented to fine-tune the resonant frequencies of the degenerate modes—an essential aspect in dual-mode filter design—by selecting a chamfer that separates the TE₁₂₀ and TE₂₁₀ modes. Notably, these modes are independent of the z -direction, implying that the height h does not affect their resonant frequency.

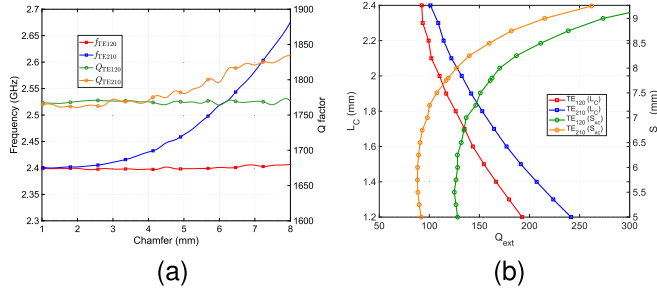


Fig. 2. (a) Degenerate TE_{120} and TE_{210} mode splitting and quality factor for different chamfer values. (b) External quality factors for different lengths of L_c and positions of Y_{Coax} .

Fig. 2(a) illustrates the frequency splitting effect and the unloaded quality factor Q for the TE_{120} and TE_{210} modes, obtained through electromagnetic simulation for different values of the chamfer dimension ch , while keeping a constant. The results confirm that the perturbation is enabling mode separation while preserving the orthogonality between the TE_{120} and TE_{210} fields. This mechanism allows precise control over both the resonant frequencies of the degenerate modes and the filter bandwidth.

According to rectangular waveguide theory [11], starting from a square geometry with side a that includes a chamfer (ch), the cross section is no longer strictly square, and the analytical cutoff expression for a rectangular waveguide serves only as a reference estimate. To compute the actual cutoff frequencies, we numerically solve the 2-D eigenvalue problem on the cross section $\Omega(a, ch)$, formulated from (1) considering the Laplacian operator in (2), to obtain the eigenvalues α_i corresponding to the transverse electric field eigenmodes, which are then converted into frequencies via (3)

$$\nabla_t^2 \varphi(x, y) + \alpha_i(a, ch) \varphi(x, y) = 0, \quad \varphi|_{\partial\Omega(a, ch)} = 0 \quad (1)$$

$$\nabla_t^2 = \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} \quad (2)$$

$$f_{c,i} = \frac{c}{2\pi \sqrt{\epsilon_r}} \sqrt{\alpha_i(a, ch)}. \quad (3)$$

The quality factor remains practically constant with this perturbation mechanism. For the dimensions selected in this design, the resonant frequencies of the first three modes are being $f_{TE110} = 1.40$ GHz, $f_{TE120} = 2.203$ GHz, and $f_{TE210} = 2.233$ GHz, and the unloaded quality factors obtained from simulations are being $Q_{TE110} = 1387$, $Q_{TE120} = 1730$, and $Q_{TE210} = 1746$. The two resonance frequencies of the degenerate modes are being used to form the passband, while the fundamental mode, which is resonating out of band, is being employed to create a direct reactive path between the source and the load, which is necessary to achieve a fully canonical transfer function.

III. DUAL-MODE MONOBLOCK FILTER

An interesting feature of this configuration is that the resonator can provide two TZs, located in both the upper and lower stopbands. As shown in Fig. 3(a), the two resonant modes are excited at the input/output (I/O) ports by means

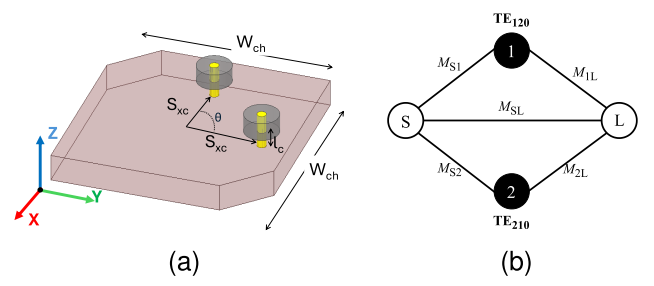


Fig. 3. (a) Filter passband. (b) Starting topology.

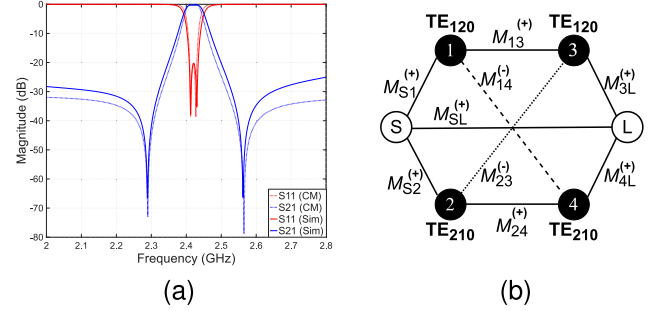


Fig. 4. (a) EM simulation and coupling-matrix model (CM_1) of a dual-mode TE cavity with two real-frequency TZs. (b) Fourth-order filter topology.

of a simple coaxial feeding probe. The input/output coupling structure includes an inner conductor, which is in contact with the dielectric walls, and an outer conductor, which belongs to the SMA connector and is in contact with the ground plane; these elements determine the port impedance.

The starting topology of the dual-mode filter is shown in Fig. 3(b). The coupling coefficients satisfy $M_{S1} = M_{L1}$ and $-M_{S2} = M_{L2}$, due to the fact that the electric field direction of the TE_{120} mode is identical at the input (S) and output (L) ports, whereas for the TE_{210} mode, the opposite is true. In addition, there is a direct source-load coupling M_{SL} , achieved through a reactive energy path generated by the TE_{110} mode, which resonates out of band. This enables the introduction of an additional transmission zero, thus achieving a fully canonical transfer function with two TZs, one on each side of the band.

By using the group delay method described in [11], the necessary link is established between the physical parameters through the values of the external quality factor Q_{Ext} and the normalized impedance of the coupling matrix model, using the (4) and (5). In this case, these values are calculated as a function of the lengths L_c , as well as S_{xc} , as illustrated in Fig. 2(b). These two parameters provide two degrees of freedom that allow for fine-tuning the required coupling. Consequently, since the external quality factor is not completely independent for each mode, as both are excited through the same physical coupling structure, the adjustment of the external coupling is addressed as a tradeoff between L_c and S_{xc} , selecting an optimal combination that simultaneously satisfies the required Q_{Ext} values for both modes in accordance

with the coupling matrix synthesis

$$\tau_{\max} = \tau(\omega_0) = \frac{4Q_e}{\omega_0} \quad (4)$$

$$Q_e = \frac{f_0}{BW \times M_{Si}^2}. \quad (5)$$

Considering the electric field distributions shown in Fig. 1(b)–(d), and following the approach in [12], we simultaneously excite the first three modes by placing the coaxial feeding connectors orthogonally in the xy -plane, with a separation angle of $\theta = 90^\circ$. In this design, we achieve the desired external coupling through an optimal combination of probe lengths and their positioning on one of the structure's faces. Additionally, the normalized resonant frequencies of the TE₁₂₀ and TE₂₁₀ modes—which define the filter's operational bandwidth—closely match the eigenvalues of the transverse coupling matrix obtained via the synthesis method. Fig. 4(a) presents a comparison between the transfer-function response associated with the coupling matrix CM₁ and the corresponding full-wave electromagnetic simulation results. The filter was synthesized following the coupling-matrix synthesis method proposed in [13], using the normalized TZs $\Omega_{TZ} = [-13.3668, 14.0921]$ as input parameters, from which the filter coupling matrix was generated. For this second-order design, the resulting response is characterized by a center frequency $f_0 = 2.42$ GHz, a bandwidth $BW = 20$ MHz, corresponding to a fractional bandwidth $FBW = 0.84\%$, and a return loss of $RL = 20.2$ dB.

$$\mathbf{CM}_1 = \begin{bmatrix} 0 & 0.86504 & 0.87006 & 0.01364 \\ 0.86504 & 1.684999 & 0 & -0.86504 \\ 0.87006 & 0 & -1.683545 & 0.87006 \\ 0.01364 & -0.86504 & 0.87006 & 0 \end{bmatrix}$$

$$\mathbf{CM}_2 = \begin{bmatrix} 0 & 0.28667 & 1.09832 & 0 & 0 & 0.00053 \\ 0.28667 & 0 & 0 & 0.22659 & -1.08486 & 0 \\ 1.09832 & 0 & 0 & -1.08486 & 0.51085 & 0 \\ 0 & 0.22659 & -1.08486 & 0 & 0 & 0.28667 \\ 0 & -1.08486 & 0.51085 & 0 & 0 & 1.09832 \\ 0.00053 & 0 & 0 & 0.28667 & 1.09832 & 0 \end{bmatrix}$$

The implementation of the topology shown in Fig. 4(b), where the coupling paths can be clearly observed, enabling the realization of a fourth-order filter with Four TZs through the concatenation of basic blocks employing coupling irises [14]. In this case, the input–output and intracavity couplings can be simplified as $M_{S1} = M_{3L}$, $M_{S2} = M_{4L}$, $M_{14} = M_{23}$, and $M_{13} \neq M_{24}$.

An essential part of the design relies on the couplings between resonators M_{24} , M_{14} , M_{23} , and M_{13} , which, owing to the dual-mode nature of the resonators can be implemented in a clever manner without shifting the iris, but rather through an appropriate placement of the feeding ports and the blind holes, without the need to rotate or displace the resonators, which remain aligned with the symmetry axis.

Analogous to the initial topology, a direct source–load coupling M_{SL} also exists, generated by a reactive path associated with the nonresonant TE₁₁₀ mode located outside the passband. Both ports excite this mode, which couples through the central iris and establishes the direct bypass associated with M_{SL} .

The simultaneous presence of this source–load path and the intracavity cross-couplings naturally leads to the appearance

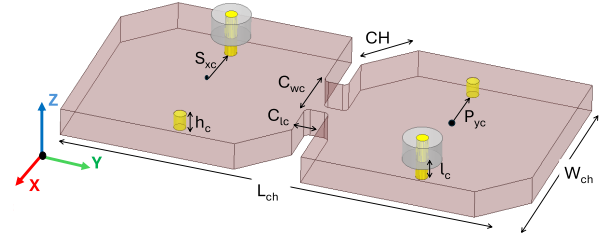


Fig. 5. Electromagnetic 3-D design of the fourth-order filter.

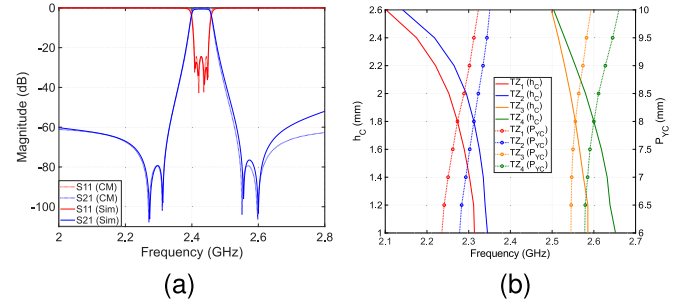


Fig. 6. Fourth-order filter. (a) Comparison between the EM simulation and the coupling-matrix model (CM₂). (b) Transmission-zero positioning for different lengths of h_c and positions of P_{Yc} .

of two TZs on each side of the passband. This occurs because the structure is homologous to two *box sections* in cascade, but with source–load coupling [15]. The internal cross-couplings generate the pair of zeros located closest to the passband, whereas the TE₁₁₀ mode provides the bypass that enables the outer pair of TZs.

The TZs appear on both sides of the passband. The full 3-D electromagnetic model is designed as shown in Fig. 5 and optimized using the following dimensions: $C_H = 4.5$ mm, $C_{wc} = 5.5$ mm, $C_{LC} = 2.5$ mm, $L_{ch} = 48$ mm, $W_{ch} = 22.8$ mm, $h_c = 1.8$ mm, $l_c = 2.3$ mm, $S_{xc} = 7.5$ mm, $P_{xc} = 8$ mm. All dimensions are measured orthogonally from the center of each resonator, considering the symmetric positions of both the port and the blind hole. The blind hole h_c , aligned with the feeding connectors, modifies the electromagnetic field distribution and enables modal coupling. By properly adjusting its position and depth, it is possible to finely control the interaction between the TE₁₂₀ and TE₂₁₀ modes, allowing precise tuning of the transmission-zero locations. As shown in Fig. 6(b), TZs TZ1 and TZ2 appear below the passband, whereas TZ3 and TZ4 are located above it. It is worth noting that these zeros are approximately symmetrically distributed around the center frequency, contributing to the filter selectivity and providing balanced out-of-band rejection on both sides of the passband. Although different physical parameters mainly control the transmission zero locations and the bandwidth, such control is not fully independent due to the intrinsically coupled nature of the structure.

Finally, Fig. 6(a) shows a comparison between the transfer-function response obtained from the coupling-matrix model CM₂ and the results of the full 3-D electromagnetic simulation. In this case, a fourth-order filter was synthesized with a return loss of $RL = 24.4$ dB, using the normalized

TABLE I
COMPARISON WITH PRIOR STUDIES

References	Operative Modes	Fully Dielectric Loaded Resonator	Fully Canonical Filtering Response	Source-Load Coupling (MSL)	CF (GHz)	Insertion Loss (dB)	FBW (%)	Filter Order	Number of TZs	Dimensions (mm)	V (mm ³)	V_N (V/ λ_g^3)
[3]	TM110, TM120, TM210	No	No	Yes	4.25	1.3	1.5	3	3	15 × 26 × 19	7410	3.46
[7]	quasi-TM120, quasi-TM210	Yes	No	No	3.5	1	5.7	4	2	36 × 18 × 5	3240	0.50
[5]	TE102, TE201	Yes	No	No	3.44	0.3	2.85	4	2	50 × 22 × 8.5	9350	1.18
[6]	TM120, TM210	Yes	No	No	3.55	0.45	1.69	4	2	41.6 × 20.8 × 9.6	8306	1.36
This Work	TE110, TE120, TE210	Yes	Yes	Yes	2.42	0.86	1.98	4	4	48 × 22.8 × 3	3283	0.39

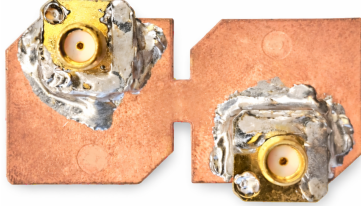


Fig. 7. Fourth-order filter fabricated prototype.

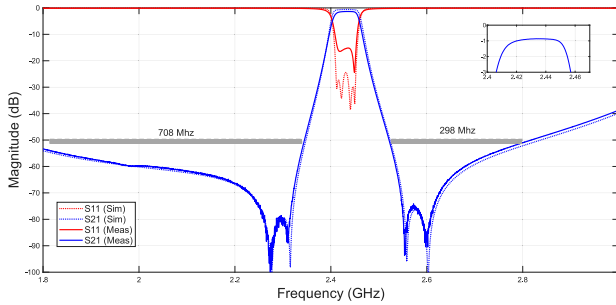


Fig. 8. Comparison between the measurement and the electromagnetic simulation.

TZs $\Omega_{TZ} = [-7.7810, -5.7079, 5.7201, 7.7898]$ as input parameters for the synthesis procedure.

Photographs of the fabricated fourth-order filter are shown in Fig. 7. The input and output couplings are fixed and implemented using 50-ohm coaxial connectors. The measurement results, compared with the simulations, are presented in Fig. 8, showing good agreement between them. The filter exhibits an insertion loss $IL = 0.86$ dB, a center frequency $f_0 = 2.42$ GHz, a bandwidth $BW = 48$ MHz, and a fractional bandwidth $FBW = 1.98\%$, making it suitable for narrowband applications. In addition, the filter provides out-of-band rejection greater than -50 dB at frequency ranges above 1 GHz. The final dimensions of the filter are $48 \times 22.8 \times 3$ mm. Small variations observed in the response can be attributed to inherent tolerances in the fabrication and assembly processes, as well as the physical properties of the materials used. Table I summarizes the comparison between this work and previous relevant works.

IV. CONCLUSION

A fourth-order prototype based on an ultra-compact waveguide section with dual-TE-mode dielectric cavities was designed, fabricated, and measured. This approach enables

fully canonical responses with TZs on both sides of the passband, as each cavity provides two poles and two zeros. A high-quality metallization process ensured low conductor losses and validated the proposed design. The resulting filter exhibits high selectivity, confirming the effectiveness of the implemented topology.

REFERENCES

- [1] L. Pelliccia, F. Cacciamani, C. Tomassoni, and R. Sorrentino, "Ultra-compact filters using TM dual-mode dielectric-loaded cavities with asymmetric transmission zeros," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Jun. 2012, pp. 1–3.
- [2] L. Pelliccia et al., "On-board miniaturized L-band input filter based on TM-mode dielectric resonator for MSS applications," in *Proc. 47th Eur. Microw. Conf. (EuMC)*, Oct. 2017, pp. 981–984.
- [3] V. Nocella, L. Pelliccia, C. Tomassoni, and R. Sorrentino, "Miniaturized dual-band waveguide filter using TM dielectric-loaded dual-mode cavities," *IEEE Microw. Wireless Compon. Lett.*, vol. 26, no. 5, pp. 310–312, May 2016.
- [4] R. Zhang and R. R. Mansour, "Low-cost dielectric-resonator filters with improved spurious performance," *IEEE Trans. Microw. Theory Techn.*, vol. 55, no. 10, pp. 2168–2175, Oct. 2007.
- [5] Y. Xie, F.-C. Chen, and Q.-X. Chu, "Inline box-like dielectric filters with asymmetric and symmetric responses," *IEEE Trans. Microw. Theory Techn.*, vol. 71, no. 6, pp. 2522–2531, Jun. 2023.
- [6] Y. Xie and F.-C. Chen, "Silver-plated dielectric waveguide filters using dual-ridge dual-mode resonator," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Nov. 2021, pp. 169–171.
- [7] Z. Huang, Y. Cheng, and Y. Zhang, "Dual-mode dielectric waveguide filters with controllable transmission zeros," *IEEE Microw. Wireless Compon. Lett.*, vol. 31, no. 5, pp. 449–452, May 2021.
- [8] L.-X. Lu, Y.-K. Zhou, W. Qin, W.-W. Yang, and J.-X. Chen, "A novel and compact dual-orthogonal-ridged dielectric waveguide resonator and its applications to bandpass filters," *IEEE Trans. Microw. Theory Techn.*, vol. 73, no. 3, pp. 1671–1679, Mar. 2025.
- [9] Z.-L. Liu, F. Liu, J. Li, X. Liu, J.-Y. Zhu, and L.-A. Bian, "A monoblock dual-mode dielectric waveguide filter with symmetric transmission zeros," in *Proc. Int. Appl. Comput. Electromagn. Soc. Symp. (ACES-China)*, Aug. 2024, pp. 1–3.
- [10] J. Uher, J. Bornemann, and U. Rosenberg, *Waveguide Components for Antenna Feed Systems: Theory and CAD*. Norwood, MA, USA: Artech House, 1993.
- [11] R. J. Cameron, C. M. Kudsia, and R. R. Mansour, *Microwave Filters for Communication Systems: Fundamentals, Design, and Applications*. Hoboken, NJ, USA: Wiley, 2007.
- [12] S. Bastioli, C. Tomassoni, and R. Sorrentino, "TM dual-mode pseudoelliptic filters using nonresonating modes," in *IEEE MTT-S Int. Microw. Symp. Dig.*, May 2010, pp. 880–883.
- [13] R. J. Cameron, "General coupling matrix synthesis methods for Chebyshev filtering functions," *IEEE Trans. Microw. Theory Techn.*, vol. 47, no. 4, pp. 433–442, Apr. 1999.
- [14] M. Guglielmi, P. Jarry, E. Kerherve, O. Roquebrun, and D. Schmitt, "A new family of all-inductive dual-mode filters," *IEEE Trans. Microw. Theory Techn.*, vol. 49, no. 10, pp. 1764–1769, Oct. 2001.
- [15] R. J. Cameron, A. R. Harish, and C. J. Radcliffe, "Synthesis of advanced microwave filters without diagonal cross-couplings," in *IEEE MTT-S Int. Microw. Symp. Dig.*, Jun. 2002, pp. 1437–1440.