Design of New Compact Multi-Layer Quint-Band Bandpass Filter

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ABSTRACT In this paper, we propose the new compact multilayer quint-passband bandpass filter. The multilayer substrate technique is used for further miniaturizing the circuit size. The filter has quint passband at 1.8, 2.5, 3.3, 3.8 and 4.5 GHz. Coupled resonators such as the stepped impedance resonators (SIRs), stub-loaded uniform impedance resonators (UIRs), and quarter-wavelength uniform impedance resonators are applied to produce the quint-passband of the filter simultaneously. The stub-loaded uniform impedance resonators and stepped impedance resonators can easily obtain a filter with closed passbands. Using a multilayer substrate technique, the filter can provide multipath propagation to improve the frequency response and achieve a compact circuit size. The proposed quint-passband bandpass filter shows a simple configuration, a practical design method, and a small circuit size.

INDEX TERMS Quint-passband, bandpass filter, cross-coupling, stepped impedance resonator, multilayer.

I. INTRODUCTION

Planar multiband bandpass filters (BPFs) with compact size, low insertion losses, high passband selectivity, and fixable frequencies have become the critical component in front-end radio frequency (RF) applications used in rapidly developing multiservice and multifunction wireless communication systems.

Some methods demonstrating the new multiband bandpass filters have been reported [1]–[6]. In [1], the defected ground structure (DGS) was used to build a quint-passband bandpass filter. In [2], a multiplexer that used a direct-feed filter in each channel selection and matching circuit was proposed; the circuit size of the filter could be further improved. In [3], the multiple stub-loaded ring resonators were used for forming a directly coupled multiband bandpass filter; the insertion loss and passband selectivity required improvement. In [4] and [5], the compact triple- and quad-band bandpass filters using the multilayer substrate technique were proposed; the integrated circuit packaging with the defected ground structure of the filter could cause the high insertion loss. In [6], the filter possesses a multilayered substrate structure, thereby providing multipath propagation to enhance the filter performance and compact circuit size. However, using a via hole to connect top- and bottom-layer substrate was complex and hard to align with the resonators located on top-layer and bottom-layer substrate. The concepts of the abovementioned references have inspired the new design presented in this paper and prompted us to consider how to improve the performance of multiband filters.

In traditional multiband filter design, the passband is generated by a pair of coupled resonators; the quad-band filter needs ten resonators. We propose a quint-passband filter that uses a multilayered substrate technique to have the advantage of both small size and good passband performance. The filter features use a multilayer technique to reduce the circuit size and provide a very closed quint-passband bandpass filter with high in-band isolation and adequate passband selectivity. Each pair of coupled stub-loaded uniform impedance resonators (UIRs) and stepped impedance resonators (SIRs) was
designed to have multi resonant paths and resonant modes that can be determined near or far away to achieve high design freedom. Furthermore, using a via-hole grounding structure can effectively control the 5th passband performance and have less current distribution on the ground plane than the defected ground structures. In addition, the via-hole grounding structure has the advantage of a simple fabrication process and miniaturizing the circuit size. In addition, each passband of the filter has symmetric transmission zeros for improving the passband selectivity. The cross-coupling resonator pairs can control the locations of every transmission zeroes. The measured results are in close agreement with the full-wave electromagnetic (EM) simulation results [7].

### TABLE 1. The design parameter of the quint-band bandpass filter (all are in mm).

| Parameter | Value |
|-----------|-------|
| L₁        | 39    |
| L₂        | 13    |
| L₃        | 13    |
| L₄        | 2.8   |
| L₅        | 10    |
| W₁        | 0.7   |
| W₂        | 0.2   |
| W₃        | 0.2   |
| W₄        | 0.5   |
| W₅        | 1.3   |
| W₆        | 1     |
| g₁        | 0.5   |
| g₂        | 0.3   |
| g₃        | 0.8   |
| g₄        | 0.3   |
| g₅        | 0.5   |
| g₆        | 0.3   |
| r         | 0.5   |

### II. FILTER DESIGN

Fig. 1(a) and (b) illustrate a three-dimensional view and the top view of the proposed multilayer quint-band BPF. The filter was designed and fabricated on a Duroid 5880 substrate with a thickness of $h = 0.787$ mm, a dielectric constant of $\varepsilon_r = 2.2$, and a loss tangent of $\delta = 0.0009$. The filter consists of coupled SIRs, quarter-wavelength UIRs, and the source-load lines on the first substrate layer, generating 1.8, 3.3, and 4.5 GHz, and consists of stub-loaded UIRs generating 2.5 and 3.8 GHz on the second substrate layer. The frequency response of each passband and index of resonator pairs of the proposed filter is shown in Fig. 1(c). The design parameter of the quint-band bandpass filter is summarized in Table 1.
The filter features use a multilayer technique to reduce the circuit size and provide a closed quint-passband BPF with high in-band isolation and adequate passband selectivity. As demonstrated in this study, the even and odd modes of the stub-loaded UIRs and SIRs can be controlled individually, resulting in the quint-band filter having very close passbands. In addition, the multilayered filter can reduce the circuit size and generate multi-propagation paths due to the extra transmission zeros near the passband edges [1], [6].

Fig. 2(a) illustrates the transmission line model of the SIR. The SIR has two characteristic impedances constructed by cascading a $2\theta_1$-long high-impedance section $Z_1$ connected to a $2\theta_2$-long low-impedance section $Z_2$. The impedance ratio $k$ is defined as $k = Z_2/Z_1$. The impedance ratio $k$ and physical length $\alpha_1$ of the SIRs are varied to adjust the fundamental resonance $f_0$ and secondary resonance $f_1$ over a wide frequency range. When $\alpha_2$ is proportional to the physical length of a SIR, the length ratio $\alpha$ of the SIR is defined as $\alpha = 2\theta_2/\theta_1$, where $\theta_1 = 2(\theta_1 + \theta_2)$. Accordingly, $k \cdot \cot(\alpha_1 \cdot \theta_1/2) = \tan[(1 - \alpha_1 \cdot \theta_1)/2]$ [8]. Depending on $k$ and $\alpha_1$, the solutions can be obtained for $\theta_1$. Fig. 2(b) depicts the transmission line model of the stub-loaded UIR. The stub-loaded UIR is composed of a conventional half-wavelength UIR (2(Z_3, \theta_3)) for the odd mode at 2.5 GHz and a stub-loaded UIR ([(Z_3, \theta_3), (Z_4, \theta_4), (Z_5, \theta_5)]) for the even mode at 3.8 GHz. To simplify the design, the parameters of the stub-loaded UIR can flexibly change the arrangements of every resonant mode. The stub-loaded UIR can be analyzed for the even and odd modes along the symmetric plane. The resonant modes of the resonator can be derived by setting $Y_{i\text{ne}} = Y_{i\text{no}} = 0$ and is expressed as

$$Y_{i\text{ne}} = \frac{Z_2(\cot\theta_1^{-} - K \tan\theta_2^{-}) + (Z_5 \cot\theta_5^{-}) (K + \cot\theta_1 \tan\theta_2^{-})}{Z_2 Z_5 \cot\theta_5^{-} (\cot\theta_1 - K \tan\theta_2^{-})} = 0$$ (1a)

$$Y_{i\text{no}} = \frac{1}{jZ_2 \tan\theta_1^{-} + K \tan\theta_2^{-}} = 0$$ (1b)

where $Z_s = -jZ_5(Z_4 \cot\theta_1^{-} - Z_5 \tan\theta_5^{-})/[Z_5 + Z_4 \cot\theta_1^{-} \tan\theta_5^{-}]$ and $\theta_5 = \theta_4 + \theta_3$. Fig. 2(c) depicts the relationships of the normalized $f_1/f_0$ versus the length ratio $\alpha_1$ with $k$ and versus $\alpha_2$ with $r$ in the stub-loaded UIR [Fig. 2(c), upper side]. The even mode ($f_2/f_0$) can be designed to have a good response and close to the odd mode ($f_1/f_0$) when $r = 0.5$ and $\alpha_2$ is approximately 0.1 or 0.9. The even modes of the proposed stub-loaded UIRs can be tuned within a wide frequency range without affecting the odd modes. Thus, this design of multiband filters with close passbands also clearly has high isolation between the passbands. In the traditional SIR (Fig. 2(c) bottom side), SIR is varied to adjust the fundamental resonance $f_0$ and secondary resonance $f_1$ in a wide frequency range by changing the impedance ratio $k$ and length ratio $\alpha_1$ of SIR.

For the filter design, the center frequencies and fractional bandwidths are $f_1 = 1.8$ GHz, $f_2 = 2.5$ GHz, $f_3 = 3.3$ GHz, $f_4 = 3.8$ GHz and $f_5 = 4.5$ GHz with $\Delta_1 = 11\%$, $\Delta_2 = 15\%$, $\Delta_3 = 5\%$, $\Delta_4 = 5\%$ and $\Delta_5 = 5\%$, respectively. The

![FIGURE 3. Relations between the 3-dB fractional bandwidth ($\Delta_1$ to $\Delta_5$) of each passband and the dimension of each resonator pair.](image)

![FIGURE 4. Each transmission zero location between the traditional and the proposed interdigital source-load lines.](image)
lumped circuit element values of the low-pass prototype filter were found to be $g_0 = 1$, $g_1 = 0.94982$, $g_2 = 1.35473$, $J_1 = -0.12333$ and $J_2 = 1.018$. The required coupling coefficients and external quality factors were found to be $M_{12}^1 = M_{21}^1 = 0.104$ and $M_{12}^2 = M_{21}^2 = 0.042$ and $Q_{e1} = 8.6$ and $Q_{e2} = 6.3$ and $Q_{e4} = 18.9$ for the second and fourth passbands (at 1.8 GHz and 3.3 GHz); $M_{13}^3 = M_{32}^3 = 0.145$ and $M_{13}^4 = M_{32}^4 = 0.047$ and $Q_{e2} = 6.3$ and $Q_{e4} = 18.9$ for the second and fourth passband (at 2.5 GHz and 3.8 GHz); $M_{56} = M_{65} = 0.05$ and $Q_{e5} = 18.9$ for the Fifth passband (at 4.5 GHz).

Fig. 3 shows the relations between the 3-dB fractional bandwidth FBW ($\Delta f_{0.5}$) of each passband and the dimension of each resonator pair. The FBW design relations for each resonator pair are based on the simulated quality factor. There are two degrees of freedom in designing the source-load lines. Additionally, the external quality factor ($Q_e$) can be extracted from the simulated frequency response as

$$Q_e = \frac{f_0}{\delta_{3-dB}}$$ (2)

where $f_0$ and $\delta_{3-dB}$ express the center frequency and the 3-dB fractional bandwidth (FBW) of the passband.

Fig. 4 shows each transmission zero location between the traditional and the proposed interdigital source-load lines. The interdigital source-load coupling results in the cross-coupling effects of the multipath propagations between the resonators, enhancing the strength of the transmission zero. Fig. 5 shows the relations between transmission zero location and corresponded image admittance (Im $[Y_{21}]$) based on the proposed interdigital source-load lines. Each transmission zero location of the filter can be obtained by the set of ABCD matrix and founded the equation (3) – (8) as below.

$$[\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Total}} = [\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Source}} \times [\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Resonator1,2}} \times [\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Resonator3,4}} \times [\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Resonator5,6}} \times [\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Load}}$$ (3)

where

$$[\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Source}} = \begin{bmatrix} \cos \theta_S & jZ_S \sin \theta_S \\ jY_1 \sin \theta_S & \cos \theta_S \end{bmatrix}$$ (4)

$$[\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Resonator1,2}} = \begin{bmatrix} \frac{Z_3 - Z_2 \cot \theta_3 \tan \theta_1}{Z_2 + Z_2 \cot \theta_2 \tan \theta_1} & 1 \\ 0 & 1 \end{bmatrix}$$ (5)

$$[\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Resonator3,4}} = \begin{bmatrix} \frac{Z_5 - Z_2 \cot \theta_3 \tan \theta_1}{Z_2 + Z_2 \cot \theta_2 \tan \theta_1} & 1 \\ 0 & 1 \end{bmatrix}$$ (6)

$$[\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Resonator5,6}} = \begin{bmatrix} \frac{jZ_3 \tan \theta_6}{Z_6 + jZ_6 \sin \theta_6} & 0 \\ 0 & 1 \end{bmatrix}$$ (7)

$$[\begin{bmatrix} A & B \\ C & D \end{bmatrix}]_{\text{Load}} = \begin{bmatrix} \cos \theta_L & jZ_2 \sin \theta_L \\ jY_1 \sin \theta_L & \cos \theta_L \end{bmatrix}$$ (8)

When $Y_{in} = 1/Z_{in} = 0$ and $S_{21} = 0$, the transmission zeros can be generated. The $S_{21}$ parameters are given by

$$S_{21} = \frac{-2Y_{21}Y_0}{(Y_0 + Y_{11})(Y_0 + Y_{22}) - Y_{12}Y_{21}} = 0$$ (9)

$$Z_1 + Z_2 \cot \theta_2 \tan \theta_1 \times Z_3 + (Z_4 + Z_5 \cot \theta_4 + \theta_5) \tan \theta_3 \times \cot \theta_6 - jZ_3 \sin \theta_3 \times \cos \theta_6 + \cos \theta_3 \times jZ_4 \sin \theta_4 \times Z_2 + Z_2 \cot \theta_2 \tan \theta_1 \times Z_3 + (Z_4 + Z_5 \cot \theta_4 + \theta_5) \tan \theta_3 \times \cot \theta_6 + \cos \theta_3 \times \cos \theta_6 \times - jZ_3 (Z_4 + Z_5 \cot \theta_4 + \theta_5) - Z_3 \tan \theta_3 \times Z_1 + Z_2 \cot \theta_2 \tan \theta_1 \times \cot \theta_6 + \cos \theta_3 \times \cos \theta_6 \times jZ_4 \times Z_1 + Z_2 \cot \theta_2 \tan \theta_1 \times Z_3 + (Z_4 + Z_5 \cot \theta_4 + \theta_5) \tan \theta_3$$ (10)
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FIGURE 7. Relations of S-parameters $|S_{21}|$ and $\text{Im} \ Y_{21}$ based on the coupling strengths of the interdigital source-load line for controlling the transmission zero locations at each passband of 1.8 GHz, 2.5 GHz, 3.3 GHz, 3.8 GHz, and 4.5 GHz.

where $Y_0$ is the characteristic admittance of the filter structure. By transforming the ABCD matrix where $B$ is the ABCD matrix

$$
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} \Rightarrow \begin{bmatrix}
Y_{11} & Y_{12} \\
Y_{21} & Y_{22}
\end{bmatrix}
$$

where $\frac{-1}{B}$ equal to (10), as shown at the bottom of the previous page, parameter, the filter structure is symmetric; therefore, $Y_{11} = Y_{22}$ and $Y_{12} = Y_{21}$. Thus, we found that $Y_{21} = 0$, which provides the conditions of the transmission zeros.

Fig. 6 illustrates the simulated coupling coefficients between the resonators. The coupling coefficients ($M_{ij}$) can be obtained from the two resonant modes as

$$
M_{ij} = \frac{(f_H^2 - f_L^2)}{(f_H^2 + f_L^2)}
$$

Through full-wave EM simulation [7], the coupling spacing gaps $g_1$, $g_4$, and $g_5$ can be tuned to satisfy the coupling degree between the adjacent resonators. Fig. 7 shows the relations of S-parameters $|S_{21}|$ and $\text{Im} \ Y_{21}$ based on the coupling strengths of the interdigital source-load lines for controlling the transmission zero locations at each passband.

When $L_4$ changed from 0 to 2.8 mm, the transmission zeros of $T_{Z1}$, $T_{Z3}$, $T_{Z5}$, and $T_{Z7}$ increase by 0.207 GHz on average; meanwhile, the transmission zeros of $T_{Z2}$, $T_{Z4}$, $T_{Z6}$, and $T_{Z8}$ are decreased by 0.243 GHz on average. Thus, the interdigital source-load coupling strength results in the cross-coupling effects by multipath propagations between the resonators. The transmission zero locations at skirts of each passband are associated with the coupling strength of the interdigital source-load line; therefore, the passband selectivity can be controlled to achieve the filter’s good frequency response. Fig. 8 shows the current distribution of the filter. We observed that the proposed resonators generate the first to fifth passbands and that no interactions induced between the resonators interfere with the performance of each passband.

### III. RESULTS

The quint-band bandpass filter was successfully fabricated and measured results using a network analyzer (R&S), as shown in Fig. 9(a) and (b). The overall circuit size is $22.3 \times 36.5$ mm$^2$ (approximately $0.32\lambda_g \times 0.19\lambda_g$, where $\lambda_g$ is the guided wavelength at the center frequency of the first passband). The center frequencies of the filter measured at 1.8, 2.5, 3.3, 3.8, and 4.5 GHz, and the 3 dB FBWs of the
TABLE 2. Comparisons table. \( \lambda_g \) is the guided wavelength of the 1st center passband frequency; RL, Return loss; IL, Insertion loss; TZs, Transmission zeros.

|                | Number of Passbands | Number of Resonators | Substrate height (mm) / \( \varepsilon_r \) | Passbands (GHz) | RL (dB) | IL (dB) | TZs | Circuit Size (mm²) \( (\lambda_g \times \lambda_g) \) |
|----------------|---------------------|----------------------|---------------------------------|-----------------|--------|--------|-----|-----------------|
| Ref. [1]       | 5                   | 6                    | PCB 0.787 / 2.2                 | 1.5 / 2.5 / 3.5 / 4.5 / 5.8 | 35 / 40 / 35 / 35 / 35 | 1.5 / 2.8 / 0.9 / 1.2 / 2.5 | 10  | 898 (0.24 * 0.17) |
| Ref. [2]       | 5                   | 10                   | PCB 0.505 / 3.38                | 0.6 / 0.9 / 1.2 / 1.5 / 1.8 | 25 / 25 / 25 / 30 / 30 / 35 | 2.8 / 2.9 / 2.9 / 2.6 / 2.3 | 7   | 2549 (0.52 * 0.04) |
| Ref. [3]       | 5                   | 10                   | PCB 0.508 / 3.65                | 1.45 / 1.6 / 1.75 / 2.1 / 2.24 | 45 / 40 / 20 / 30 / 25 / 25 | 2.8 / 2.7 / 2.9 / 2.7 / 2.7 | 8   | 18554 (1.56 * 1) |
| Ref. [4]       | 5                   | 12                   | PCB 0.508 / 2.2                 | 0.63 / 1.3 / 2.03 / 2.7 / 3.4 | 20.6 / 15.9 / 12.3 / 20 / 22.8 | 0.47 / 1.14 / 1.8 / 1.39 / 1.2 | 5   | 7964 (0.04 * 0.18) |
| Ref. [9]       | 4                   | 4                    | PCB 0.8 / 2.2                   | 2.54 / 3.46 / 4.5 / 5.2 | -14.2 | 1.58 / 1.78 / 2.23 / 2.45 | 8   | 1748 (0.59 * 0.42) |
| Ref. [10]      | 4                   | 2                    | PCB 0.54 / 2.5                  | 2.46 / 3.41 / 15.1 / 15.3 / 25.7 / 14.3 | 0.68 / 0.73 / 0.81 / 1.23 | 4   | 952 (0.59 * 0.26) |
| Ref. [11]      | 4                   | 4                    | PCB 0.508 / 3.55                | 1.57 / 2.4 / 3.5 / 5.2 | -15   | 5.18 / 2.79 / 2.82 / 5.19 | 7   | 844 (0.32 * 0.16) |
| This work      | 5                   | 6                    | PCB 0.787 / 2.2                 | 1.8 / 2.5 / 3.3 / 3.8 / 4.5 | 27 / 29 / 31 / 26 / 34 | 0.7 / 0.28 / 0.8 / 0.6 / 0.9 | 8   | 813 (0.32 * 0.19) |

FIGURE 8. Current distribution of the quint-band bandpass filter.

FIGURE 9. (a) Photograph and (b) measured results of the fabricated filter.

filter are 11%, 15%, 5%, 5%, and 5%. The corresponding insertion losses are 0.7, 0.28, 0.8, 0.6 and 0.9 dB, and the return losses of each channel are 27, 29, 31, 26 and 34 dB. The transmission zeros at the skirts of each passband are improving the selectivity of the quint-band bandpass filter. In addition, by tuning the structure dimensions of the proposed resonators, the transmission zeros can be designed and generated at the skirts of each passband because the cross-coupling effects occurred between the multipath propagations in the coupled resonators. The comparison of the frequency response of the quint-band bandpass filter and previous works was summarized in Table 2. It noted that the proposed quint-band bandpass filter has good use for providing high passband selectivity and compact size.

IV. CONCLUSION

In this paper, a compact quint-passband filter using a multilayered structure has been presented successfully. The multilayer substrate technique is used for further miniaturizing the circuit size. The filter has quint passbands at 1.8, 2.5, 3.3,
3.8, and 4.5 GHz and having symmetric transmission zeros at the skirts of each passband. Each pair of coupled resonators, such as the stepped impedance resonators, stub-loaded uniform impedance resonators, and quarter-wavelength uniform impedance resonators, are applied to produce the filter’s quint-passband simultaneously. The design procedure is comprehensively discussed, and the resonant responses of the filter are analyzed. The proposed filter is helpful for applications in multiband and multiservice wireless communication systems.

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