Synthesis Design of Equal-Ripple and Quasi-Elliptic Wideband BPFs With Independently Reconfigurable Lower Passband Edge

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ABSTRACT This paper presents a new synthesis design of equal-ripple and quasi-elliptic wideband bandpass filters (WB BPFs) with independently reconfigurable lower passband edge. The synthesis design starts from a 5th-order LC circuit. By using optimum multipole concept, a T-shaped resonator for the design of Chebyshev BPF without redundant elements can be directly derived, which is formed by a pair of parallel-coupled lines and short-ended stub. To sharpen the selectivity and maintain equal-ripple response, a pair of transmission zeros are introduced through replacing the short-ended stub by its multi-section counterparts. Thus, two prototypes, whose equivalent LC circuits can be directly derived, are presented for the synthesis of equal-ripple and quasi-elliptic BPFs. To realize the desired reconfigurable response, a capacitor is used. By controlling the capacitance value, the lower passband edge can be independently tunable in continue manners while the upper one is fixed. For validation, two prototypes with fixed and reconfigurable bandwidth are fabricated and measured. As the measured and simulated results agree well with each other, a high-efficiency design of reconfigurable-bandwidth WB BPFs is validated.

INDEX TERMS Synthesis design, reconfigurable bandwidth, wideband bandpass filter, equal-ripple and quasi-elliptic, H-shaped resonator.

I. INTRODUCTION

Due to the desired features, including minimum variation in group delay, low cost, compact size, low profile, and easy integration with other devices, the microwave planar bandpass filters (BPFs) with equal-ripple response are extensively used in the modern telecommunication systems. To provide a valid circuit schematic and design method, many techniques have been presented [1]–[4]. Although the desired equal-ripple response are existed in the reported works, their skirt selectivity is poor, which will lead to low use efficiency of the electromagnetic spectrum. To deal with this issue, the traditional method is to increase the filter order [5]. However, this design approach not only enlarges the circuit size, but also increases the design complexity and cost.

As an alternative method, the skirt selectivity also can be improved by introducing transmission zeros (TZs) near the lower/upper passband edges. One of the effective methods
is to replace single-section circuited stubs by their multi-section counterparts [6]. Besides, TZs also can be generated by exploiting two transmission paths [7]–[8], introducing I/O cross couplings [9], and using shunt short/open circuits at designated TZ frequencies [10]–[11]. In addition, other methods for introducing TZs include using the bandgap structure [12] and intrinsic zeros of a special multi-mode resonator [13]–[15]. Although most of these above BPFs can own equal-ripple and quasi-elliptic responses, they still could not satisfy the requirements of modern transceivers, which need filters owning different reconfigurabilities to reduce the system size and design complexity.

The reconfigurable BPFs have been studied over several decades. Previous works are mainly focused on the center-frequency tunability [16]–[20]. But for the reconfigurable-bandwidth BPFs, much less efforts have been devoted [21]–[24], which are analyzed by odd-even-mode method and designed by cut-and-try method. Considering that the cut-and-try method will not only lead to low design efficiency, but also cannot provide the intrinsic design mechanism, many energy and time will be wasted to design a desired reconfigurable BPF. To tackle this, it is very significant to explore different topologies for the synthesis design of BPFs with reconfigurable bandwidth.

In this paper, a new H-shaped resonator is derived for the synthesis design of equal-ripple and quasi-elliptic wideband (WB) BPFs with independently reconfigurable lower pass-band edge. The proposed topology starts from a 5th-order Chebyshev LC circuit. By using the transfer function, a T-shaped resonator with its parameters is then directly derived. To sharpen skirt selectivity and maintain the equal-ripple response, two TZs are introduced near the lower and upper passband edges by replacing the short-ended stub with its multi-section counterparts. Thus, the passive filtering topology can be derived with LC circuit. Later, a capacitor is used as the tuning element for the design of WB BPFs with independently tunable lower passband edge. To verify this, two prototypes are fabricated and measured. To the best knowledge of authors, such synthesis design of equal-ripple and quasi-elliptic WB BPFs with reconfigurable bandwidth has never been reported before.

Except for the introduction in Section I, the derivation procedure from 5th-order Chebyshev LC circuit to the proposed topology is presented in Section II. In Section III, two prototypes with fixed and reconfigurable bandwidth are implemented. In section IV, a conclusion is summarized.

II. BASIC THEORY

In this section, the derivation procedure from a 5th-order Chebyshev LC circuit to the proposed topology is presented. To better understand this, some analysis steps are provided in this section, which are illustrated as follows:

A. TRANSFORMATIONS FROM LUMPED CIRCUITS TO QUARTER-WAVELENGTH RESONATORS

In 1960s, M.C. Horton proposed the initial optimum multipole concept in first time for the design of Chebyshev BPFs only formed by quarter-wavelength resonators and without redundant elements [25]. To successfully derive such BPFs from a lumped-element LC model, i.e., the one in Fig. 1(a), some transformations from the lumped circuits to quarter-wavelength resonators should be proved at first. Based on a characteristic function $F$ in [26], the squared $S_{21}$-magnitude of parallel coupled-line (PCL) in Fig. 1(b) can be expressed as

$$|S_{21}|^2 = \frac{1}{1 + |F|^2}$$

(1)

where $F = (B_b - C_b)/2$, $B_b$ and $C_b$ present the entries of the normalized ABCD matrix of the PCL. When the normalized impedance is $Z_0$, $B_b$ and $C_b$ can be expressed as

$$B_b = j \frac{B_B}{2B_b Z_0} \sin \theta$$

$$C_b = j \frac{2Z_0 \sin \theta}{B_B}$$

(2a, 2b)

with

$$A_B = Z_{oe}' + Z_{oo}' \quad B_B = Z_{oe}' - Z_{oo}'$$

(3)

Therefore, the function $F$ of a PCL can be derived as

$$F = j \left( \frac{4Z_0^2 - A_B^2 \cos^2 \theta}{4B_B Z_0} \frac{\cos^2 \theta}{\sin \theta} - \frac{4Z_0^2 - B_B^2}{4B_B Z_0} \frac{1}{\sin \theta} \right)$$

(4)

Based on the theories in [26] and [27], the squared $S_{21}$-magnitude of 2nd-order lumped Chebyshev BPF, which is formed by two LC components with three $J$ inverters, can be expressed as

$$|S_{21}|^2 = \frac{1}{1 + \varepsilon^2 \cos^2(\phi + \xi)}$$

(5)

with

$$\cos \phi = \cos \theta / \cos \theta_c \quad \cos \xi = \tan \theta_c / \tan \theta$$

(6)

where $\varepsilon$ is the specified equal-ripple constant while $\theta_c$ is the phase at lower cutoff frequency $f_c$. After some algebraic operations, the following equation can be determined as

$$\varepsilon \cos(\theta + \xi) = \varepsilon (1 + \sin \theta_c) \cos^2 \theta \frac{\cos^2 \theta_c}{\sin \theta} - \varepsilon \frac{1}{\cos^2 \theta_c} \sin \theta$$

(7)

Apparently, the coefficients of $\cos^2 \theta / \sin \theta$ and $1 / \sin \theta$ of the lumped circuit and distributed PCL will be equivalent under the condition that:

$$\frac{4Z_0^2 - B_B^2}{4B_B Z_0} = \varepsilon (1 + \sin \theta_c)$$

(8a)

$$\frac{4Z_0^2 - A_B^2}{4B_B Z_0} = \varepsilon \cos^2 \theta_c$$

(8b)
Based on the above analysis, the transformation from two LC components with three J inverters to quarter-wavelength PCL can be proved. For the transformation from a shunt LC component to a short-ended quarter-wavelength stub, which is illustrated in Fig. 1(c), it has been proved in [1] and will not be repeated. Thus, the 5th-order distributed equal-ripple BPFs, whose topology is only a T-shaped resonator formed by a pair of PCLs and short-ended stub, can be derived by optimum multipole concept, as shown in Fig. 1(d).

B. TRADITIONAL CHEBYSHEV BPFs

Apparently, the topology in Fig. 1(d) is symmetrical. Thus, its squared magnitude also can be expressed by (1), where $F = (B_d - C_d)/2$, $B_d$ and $C_d$ present the entries of the normalized ABCD matrix of topology in Fig. 1(d). When the normalized impedance is $Z_0$, the parameters of $B_d$ and $C_d$ can be written as

$$B_d = \frac{2A_D B_b}{B_D} \cos \theta + \frac{2Z_0 B^2_B}{jZ_0^2 \tan \theta} \quad (9a)$$

$$C_d = \frac{2A_D C_b}{B_D} \cos \theta + \frac{2A^2_D Z_0}{jB^2_D Z_0^2 \sin \theta} \quad (9b)$$

with

$$A_D = Z^d_{oe} + Z^d_{ee}, \quad B_D = Z^d_{oe} - Z^d_{ee} \quad (10)$$

Thus, the function $F$ of the topology in Fig. 1(d) can be derived as

$$F = j \left( k_1 \cos \frac{\theta}{\sin 3 \theta} + k_2 \cos \frac{3 \theta}{\sin 3 \theta} + k_3 \cos \frac{3 \theta}{\sin 3 \theta} \right) \quad (11)$$

with

$$k_1 = \frac{A^2_D}{2Z_0 B^2_D} - \frac{2A_D Z_0}{B^2_D} + \frac{A^4_D}{4Z_0 Z^T_{oe} B^2_D} - \frac{A^2_D Z_0}{Z^T_{oe} B^2_D} \quad (12a)$$

$$k_2 = \frac{4A_D Z_0}{B^2_D} - \frac{A_D}{2Z_0} - \frac{A^3_D Z_0}{2Z_0 B^2_D} - \frac{A^2_D}{2Z_0 Z^T_{oe}} + \frac{A^2_D Z_0}{Z^T_{oe} B^2_D} \quad (12b)$$

$$k_3 = \frac{A_D}{2Z_0} - \frac{2A_D Z_0}{B^2_D} + \frac{B^2_D}{4Z_0 Z^T_{oe}} \quad (12c)$$

For the 5th-order Chebyshev LC circuit illustrated in Fig. 1(a), its squared magnitude of transmission coefficient can be expressed as

$$|S_{21}|^2 = \frac{1}{1 + \epsilon^2 \cos^2 (2\phi + 3\xi)} \quad (13)$$

with

$$\cos \phi = \cos \theta / \cos \theta_c, \quad \cos \xi = \tan \theta_c / \tan \theta \quad (14)$$

where $\epsilon$ is the specified equal-ripple constant and $\theta_c$ is the phase at lower cutoff frequency $f_c$. Due to the frequency-distribution characteristic of Chebyshev BPF, the upper cut-off phase is located at $(180^\circ - \theta_c)$. Hence,
TABLE 1. Characteristic impedances respect to different fractional bandwidth and ripple constant.

| FBW (%) | 100 | 100 | 100 | 77.8 | 55.6 |
|-------|-----|-----|-----|------|------|
| \( L_A \) (dB) | 0.02 | 0.06 | 0.10 | 0.10 | 0.10 |
| \( \theta_c \) | 45° | 45° | 45° | 55° | 55° |
| \( \varepsilon \) | 0.068 | 0.118 | 0.153 | 0.153 | 0.153 |
| \( Z_{oL} \) (Ω) | 51.8 | 51.6 | 52.4 | 34.0 | 21.2 |
| \( Z_{oe} \) (Ω) | 25.0 | 31.5 | 35.5 | 55.0 | 91.4 |
| \( Z_{oc} \) (Ω) | 114.0 | 126.2 | 134.3 | 151.0 | 185.2 |

the relationship between lower cutoff phase \( \theta_c \) and fractional bandwidth (FBW) can be defined as

\[
\theta_c = 90^\circ \times \frac{2 - \text{FBW}}{2}.
\] (15)

After some algebraic operations, an explicit expression can be derived as

\[
\cos (2\phi + 3\xi) = k_1 \cos^2 \theta + k_2 \cos^3 \theta + k_3 \cos^4 \theta
\]

with

\[
k_1 = 2\varepsilon \frac{3 \sin^2 \theta_c + \sin^2 \theta + 3 \sin \theta_c + 1}{\cos \theta_c}
\] (17a)

\[
k_2 = -\varepsilon \frac{3 \sin^2 \theta_c + 6 \sin^2 \theta + 9 \sin \theta_c + 4}{\cos^3 \theta_c}
\] (17b)

\[
k_3 = \varepsilon \frac{3 \sin \theta_c + 2}{\cos \theta_c}
\] (17c)

Compared with (10) and (15), it is obvious that \( Z_{oL}^d \), \( Z_{oe}^d \), and \( Z_{oc}^d \) can be directly determined once the ripple constant and lower cutoff phase are given. As such, the squared \( S_{21} \) magnitude can be pre-specified with the help of (1), (10), and (11).

For a BPF based on topology in Fig. 1(d) owing different bandwidth and ripple constant, its characteristic impedances are summarized in Table 1. Obviously, the parameters of \( Z_{oo}^d \), \( Z_{oe}^d \), and \( Z_{oc}^d \) will be larger with the increase of bandwidth, while the ratio between \( Z_{oo}^d \) and \( Z_{oe}^d \) changes smaller. As the ripple level decreases, the parameters of \( Z_{oo}^d \) and \( Z_{oe}^d \) become smaller, and their ratio changes larger. To demonstrate this vividly, the frequency responses of BPF based on the topology in Fig. 1(d) with fixed ripple level \( L_A = 0.10 \) dB and different lower cutoff phases \( \theta_c = 45^\circ, 55^\circ, 65^\circ \) are plotted in Fig. 2(a), while the ones with different ripple levels \( L_A = 0.02 \) dB, \( 0.06 \) dB, \( 0.10 \) dB) and fixed lower cutoff phase \( \theta_c = 45^\circ \) are illustrated in Fig. 2(b). Here, the relationship between FBW and lower cutoff phase is shown in (14), and the fixed ripple level \( L_A \) is related to the given ripple constant by

\[
L_A = 10 \log \left(1 + \varepsilon^2\right).
\] (18)

Seen from the inset of Fig 2(a), the upper cutoff phase is exactly located at 135°, 125°, 115° as predicted, and the ripple levels are all restricted to 0.10 dB. When the lower cutoff phase is smaller, the BPF topology owns sharper out-of-band roll-off skirt, as shown in the main figure. In Fig. 2(b), the frequency responses of the topology with fixed lower cutoff phase and different ripple levels are given. As predicted, these three topologies own different magnitudes of transmission coefficient, and the one with larger ripple level owns narrower bandwidth. Under these conditions, the transmission pole phases are located at the same phases, and the ripple levels are restricted to 0.02, 0.06, and 0.10 dB as pre-specified. In summary, the topology in Fig. 1(d) can be used for designing equal-ripple BPFs with arbitrary bandwidth and ripple constant.

C. EQUAL-RIPPLE AND QUASI-ELLIPTIC BPFs

To increase the utilization efficiency of electromagnetic spectrum, the filter skirt selectivity should be improved by introducing TZs near the lower and upper passband edges, which are realized through replacing the short-ended stub by its multi-section counterparts, as shown in Fig. 1(e) and (f). To ensure the locations of TZs, the rigorous ABCD-matrix and S-parameters theories are used. For the topology
in Fig. 1(e), its overall ABCD-matrix can be expressed as

\[ T_O = T_{PCL} \ast T_T \ast T_{PCL} \] (19)

where

\[ T_{PCL} = \begin{bmatrix} A_E \cos \theta & B_E \sin \theta & j2B_E \sin \theta & A_E \cos \theta \\ B_E \sin \theta & -A_E \cos \theta & j2A_E \cos \theta & -B_E \sin \theta \\ j2A_E \tan \theta & -A_E^2 - B_T^2 - B_T^2 \tan^2 \theta & j2A_T \tan \theta & -A_T^2 - B_T^2 - B_T^2 \tan^2 \theta \end{bmatrix} \] (20a)

\[ T_T = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \] (20b)

where

\[ A_E = Z_{oe}^e + Z_{oo}^e, \quad B_E = Z_{oe}^e - Z_{oo}^e \] (21a)

\[ A_T = Z_{oe}^T + Z_{oo}^T, \quad B_T = Z_{oe}^T - Z_{oo}^T \] (21b)

For the transmission coefficient \( S_{21} \), it can be written as

\[ S_{21} = \frac{2}{A_e + B_e} \] (22)

where \( A_e, B_e, C_e, D_e \) are the entries of the overall ABCD matrix of the topology in Fig. 1(e), and \( Z_0 \) is the normalized impedance. Under the condition that \( S_{21} = 0 \), the introduced TZs can be determined and expressed as

\[ f_{z_{1l}} = \arctan \sqrt{\frac{A_T^2 - B_T^2}{B_T}} \] (23a)

\[ f_{z_{1l}} = 2f_0 - \arctan \sqrt{\frac{A_T^2 - B_T^2}{B_T}} \] (23b)

According to the theoretical analysis mentioned in [28], the quality factor of the topology shown in Fig. 1(d) can be approximately expressed as

\[ Q = \frac{1}{F_{BW3-dB}} \approx \left(1 - \frac{f_{z_{1l}}}{f_0}\right) \left(\frac{B_T^2}{A_T^2} + \frac{2B_T^2}{\pi Z_0^2}\right) \] (24)

where \( Z_0 \) is the normalized impedance, and \( F_{BW3-dB} \) is the 3-dB FBW of the topology in Fig. 1(e). As the parameters of \( Z_{oe}^e \) and \( Z_{oo}^e \) have been determined in Part A, the initial values of \( Z_{oe}^T \) and \( Z_{oo}^T \) can be determined when \( F_{BW3-dB} \) and \( f_{z_{1l}} \) are pre-specified. Hence, the initial parameters of the topology in Fig. 1(e) can be directly determined when the specifications including equal-ripple level, locations of \( f_{z_{1l}} \), and bandwidth (equal-ripple and 3-dB ones), are given.

It cannot be denied that the filter bandwidth based on the initial parameters is slightly different from the desired one. Besides, the filter based on the initial parameters also do not own the equal-ripple response anymore in the passband. Fortunately, these problems can be tackled by optimizations, which can be finished by the commercial software, such as, MATLAB. During the optimization procedure, two design skills are used:

1) The filter bandwidth is mainly controlled by \( Z_{oe}^e \) and \( Z_{oe}^T \). Based on (23), increase (decrease) these two parameters can enlarge (reduce) the bandwidth.

2) The ripple level is mainly determined by \( Z_{oe}^e \) and \( Z_{oe}^T \). As these two parameters increase (decrease), the ripple level will be enlarged (reduced).

For this design procedure for BPFs with equal-ripple and quasi-elliptic response, it can be summarized as a flowchart, as shown in Fig. 3. Here we give a filter example, whose specifications can be summarized as: ripple level of 0.10 dB, 0.10-dB FBW of 71.8%, 3-dB FBW of 75.3%, and TZs at 0.59 and 1.41 \( f_0 \). With the help of the flowchart, the characteristic impedances under Case A, B, C can be determined and summarized in Table 2. Seen from it, the initial values of \( Z_{oe}^e, Z_{oe}^T \) are almost same with the final ones. Besides, the corresponding frequency responses are shown in Fig. 4. Seen from the main figure of Fig. 4, it is obvious that the \( S_{21} \)-magitudes of BPFs with initial and optimized parameters are similar, and the locations of the introduced TZs are located at the same places. Meanwhile, there are also some slight discrepancies under Case B and C. For example, BPF with initial parameters does not own equal-ripple response. To tackle this, some optimizations are used. After some optimizations, the desired equal-ripple response can be realized, as shown in the inset of Fig. 4. As such, the topology in Fig. 1(e) can be used for the synthesis design of equal-ripple and quasi-elliptic BPFs, which will be experimentally verified in Part A of Section III.

For a terminated quarter-wavelength T-shaped resonator with \( Z_1 \) and \( Z_2 \), its corresponding input admittance can be
The characteristic impedance respect of the proposed topology under different work states.

|          | $Z_{in}$ (Ω) | $Z_{in}$ (Ω) | $Z_{in}$ (Ω) | $Z_{in}$ (Ω) |
|----------|--------------|--------------|--------------|--------------|
| Case A   | 62.7         | 158.1        | 0            | 30           |
| Case B   | 62.7         | 158.1        | 60.4         | 240.9        |
| Case C   | 57.0         | 169.9        | 60.3         | 240.5        |

Case A: Characteristic impedances of a traditional Chebyshev BPF with equal-ripple level of 0.10 dB and equal-ripple FBW of 71.8%.
Case B: The initial Characteristic impedances of the equal-ripple and quasi-elliptic BPF based on (11), (15), (16), (18), (23), (24).
Case C: The final Characteristic impedances of the equal-ripple and quasi-elliptic BPF.

expressed as

$$Y_{in}^T = j \left[ \frac{2Z_1 + Z_2}{Z_1Z_2} \right] \tan \theta \left[ 1 - 2 \left( \frac{Z_1}{Z_2} \right) \tan^2 \theta \right].$$  \hspace{1cm} (25)

For a terminated quarter-wavelength PCL with $Z_{in}^T$ and $Z_{in}^T$, its corresponding input admittance can be expressed as

$$Y_{in}^{PCL} = j \left[ \frac{2A_T}{A_T^2 - B_T^2} \right] \tan \theta \left[ 1 - \frac{B_T^2}{A_T^2} \right] \tan^2 \theta. $$  \hspace{1cm} (26)

Apparently, a T-shaped resonator with open terminations can be treated as the equivalent structure of a PCL when the following relationships are satisfied

$$\frac{2A_T}{A_T^2 - B_T^2} = \frac{2Z_1 + Z_2}{Z_1Z_2},$$  \hspace{1cm} (27a)

$$\frac{B_T^2}{A_T^2 - B_T^2} = \frac{Z_1}{Z_2}. $$  \hspace{1cm} (27b)

Hence, the topology in Fig. 1 (f) also can be used for the synthesis design of equal-ripple and quasi-elliptic BPFs. In addition, these two topologies also own the same equivalent LC circuit, which can be directly derived as follows: from Fig. 1(b), the equivalent circuit of a terminated PCL can be derived as the one plotted in Fig. 5(a). Based on the theory in [1], the input admittance of a short-ended \( J \) inverter \( Y_{in} \) can be written as

$$Y_{in} = D_J / B_J$$  \hspace{1cm} (28)

with

$$D_J = 0, \quad B_J = j / J$$  \hspace{1cm} (29)

After some algebraic operations, it is obvious that \( Y_{in} \) is zero. Thus, the equivalent circuit of the terminated PCL can be simplified as the one in Fig. 5(b). Considering that the equivalent circuit of a series PCL has been proved in Fig. 1(b), the final equivalent circuit of these two topologies can be directly determined and plotted in Fig. 5(c).

D. RECONFIGURABLE-BANDWIDTH BPFS

Compared with the topology in Fig. 1(e), it is apparent that the topology shown in Fig. 1(f) is more suitably used as the passive filtering structure for the design of reconfigurable filter. To realize the desired reconfigurable performances, a capacitor is inserted into the middle of H-shaped resonator, as illustrated in Fig. 1(g). Under this condition, the input admittance of the T-shaped resonator with a capacitor can be expressed as

$$Y_{in,C}^T = \frac{1}{Z_1 Z_{in1} + jZ_1 \tan \theta / 2}$$  \hspace{1cm} (30)

where

$$Z_{in1} = \frac{Z_1}{j} \left( \frac{Z_2 - 2Z_0 \tan \theta / 2 \tan \theta}{2Z_0 \tan \theta + Z_2 \tan \theta / 2} + \frac{1}{\omega C_V} \right). $$  \hspace{1cm} (31)

When the capacitance value of the inserted capacitor \( C_V \) is large enough, i.e., \( C_V = 128 \) pF, the input admittance of T-shaped resonator with capacitor \( YT_{in,C} \) can be simplified as the one without capacitor \( YT_{in} \). Under this case, the effect of the inserted capacitor on the filter responses is insignificant. However, the conditions will be different with the decrease

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of $C_V$. The relationships among the 3-dB lower/upper passband edge, TZs for sharp selectivity, and $C_V$ are illustrated in Fig. 6. Apparently, the lower passband edge will move upwards with the decrease of $C_V$, while the upper one are maintained unaltered. Besides, TZ near the lower passband edge also move upwards with the decrease of $C_V$, which can be ensure the excellent skirt selectivity. Thus, the proposed topology in Fig. 1(g) can be used to design WB BPFs with independently reconfigurable lower passband edge.

### III. FABRICATION AND MEASUREMENT

To verify the proposed synthesis design approach, two prototypes with fixed and reconfigurable bandwidth are presented in this section, whose topology are shown in Fig. 1(e) and (g), respectively. For the fixed-bandwidth filter, its corresponding FBW is 81.8%. For the reconfigurable-bandwidth filter, its FBW can be varied from 43.4% to 79.5%.

#### A. FILTER I

In this design, an equal-ripple and quasi-elliptic WB BPF is synthesis designed. The desired filter specifications can be summarized as: ripple level of 0.10 dB, 0.10-dB FBW of 71.8%, 3-dB FBW of 75.3%, and TZs at 0.59 and 1.41 $f_0$, and $f_0 = 4.05$ GHz. It is final parameters can be determined by using the flowchart illustrated in Fig. 3. In addition, the parameters of the filter under Case A, B, C are summarized Table 2, and the corresponding theoretical $S$-parameters are shown in Fig. 4. The final parameters can be summarized as: $Z'_{eo} = 169.9$ Ω, $Z'_{oo} = 57.0$ Ω, $Z'_{oe} = 240.5$ Ω, and $Z'_{oo} = 60.3$ Ω. After obtaining the final parameters, the filter layout can be determined with the help of commercial electromagnetic software. As the presented filter is manufactured on Rogers RO4003B with dielectric constant of 3.38, loss tangent of 0.022, and thickness of 0.813 mm, the final layout with corresponding dimensions can be summarized in Fig. 7.

In this design, there is a design skill used: to ensure the filter symmetry, the terminated PCL is replaced by a parallel-three-coupled-line (PTCL) [29]. The circuit size is 32.5 mm $\times$ 20.9 mm (0.80 $\lambda_g \times 0.51 \lambda_g$, where $\lambda_g$ is the wavelength in the substrate at the center frequency).

#### FIGURE 6. 3-dB lower and upper passband edges with two TZs for sharp skirt selectivity are presented respect to different $C_V$ (the parameters are: $Z'_{eo} = 176.7$ Ω, $Z'_{oo} = 63.2$ Ω, $Z'_{oe} = 74.8$ Ω, $Z'_{oo} = 86.0$ Ω).

#### A. FILTER I

In this design, an equal-ripple and quasi-elliptic WB BPF is synthesis designed. The desired filter specifications can be summarized as: ripple level of 0.10 dB, 0.10-dB FBW of 71.8%, 3-dB FBW of 75.3%, and TZs at 0.59 and 1.41 $f_0$, and $f_0 = 4.05$ GHz. It is final parameters can be determined by using the flowchart illustrated in Fig. 3. In addition, the parameters of the filter under Case A, B, C are summarized Table 2, and the corresponding theoretical $S$-parameters are shown in Fig. 4. The final parameters can be summarized as: $Z'_{eo} = 169.9$ Ω, $Z'_{oo} = 57.0$ Ω, $Z'_{oe} = 240.5$ Ω, and $Z'_{oo} = 60.3$ Ω. After obtaining the final parameters, the filter layout can be determined with the help of commercial electromagnetic software. As the presented filter is manufactured on Rogers RO4003B with dielectric constant of 3.38, loss tangent of 0.022, and thickness of 0.813 mm, the final layout with corresponding dimensions can be summarized in Fig. 7. In this design, there is a design skill used: to ensure the filter symmetry, the terminated PCL is replaced by a parallel-three-coupled-line (PTCL) [29]. The circuit size is 32.5 mm $\times$ 20.9 mm (0.80 $\lambda_g \times 0.51 \lambda_g$, where $\lambda_g$ is the wavelength in the substrate at the center frequency).

In Fig. 8(a), the measured $S$-parameters are compared with the simulated and synthesis ones. It is obvious that these three groups of results are in good agreement with each other, especially for the simulated and synthesis ones. When the insertion loss is smaller than 3 dB, the measured passband range is from 2.37 to 5.65 GHz. Therefore, the measured center frequency is 4.01 GHz, and the 3-dB FBW is 81.8%. At the center frequency, the insertion loss is 0.52 dB. The return loss is more than 13.9 dB within the main passband. Two TZs are observed at 0 and 7.80 GHz for the wide stopband, and five transmission poles (TPs) are found at 2.51, 2.71, 3.91, 5.11, and 5.55 GHz for the broad pass-band. Furthermore, this filter achieves a wide stopband with more than 16.0-dB attenuation from direct current (D.C.) to 1.29 GHz and 5.92 to 9.0 GHz. Two TZs located at 2.26 and 6.11 GHz for sharp skirt selectivity, and the attenuation slope in the lower and upper transition bands are 147 and 81 dB/GHz, respectively. The group delay is quite flat with-in the passband and varied

#### FIGURE 8. Measured results compared with the simulated and synthesis ones. (a) $S$-parameters. (b) Group delay.
from 0.25 - 0.54 ns, showing a good linearity. In Fig. 9, the photograph of the fabricated filter is shown. To highlight the advantages of the presented filter, some comparisons with previous works are listed in Table 3. As the presented filter is only constructed by four unknown parameters, whose initial values can be directly determined once the filter specifications are given, the biggest advantage of this synthesis work is high-efficiency design of equal-ripple and quasi-elliptic BPFs.

### B. FILTER II

The specifications of the desired reconfigurable-bandwidth BPF with widest bandwidth are summarized as: ripple level of 0.18 dB, 0.18-dB FBW of 75.8%, 3-dB FBW of 80.0%, and TZs at 0.52 and 1.48 $f_0$. In addition, the 3-dB narrowest FBW is 61.8%. According to (11), (15), (16), (18), the parameters of $Z_{g\omega}$ and $Z_{goe}$ can be determined directly. By using (23), (24), and (27), the initial values of $Z_1$ and $Z_2$ also can be derived. In addition, the capacitance values of $C_V$ can be determined with the assistance of Fig. 4. After some optimizations, the final parameters can be determined. Similar with Filter I, the design procedure for Filter II also can be concluded as a flowchart, as shown in Fig. 10. With the help of this design procedure, the final parameter can be summarized as: $Z_{goe} = 176.7 \Omega$, $Z_{g\omega} = 63.2 \Omega$, $Z_1 = 74.8 \Omega$, $Z_2 = 86.0 \Omega$, $C_V = 128$ pF (widest bandwidth), and $C_V = 1.0$ pF (narrow bandwidth). As the presented filter is manufactured on Rogers RO4003B with dielectric constant of 3.38, loss tangent of 0.022, and thickness of 0.813 mm, the final filter layout and dimensions can be summarized in Fig. 11. The circuit size is 29.0 mm $\times$ 19.0 mm ($0.76 \lambda_g \times 0.50 \lambda_g$, where $\lambda_g$ is the wavelength in the substrate at the center frequency).

When the capacitance value $C_V$ is large enough, such as, $C_V = 128$ pF, the simulated and theoretical $S$-parameters
TABLE 4. Comparison with previous works.

|       | CF1 (GHz) | TPs/TZs | Number of tuning elements | Design parameters1 | Number of tuning states | FBW tuning ranges (%) | Circuit size (kg x lg) | Design Method |
|-------|-----------|---------|---------------------------|--------------------|------------------------|----------------------|------------------------|---------------|
| [21]  | 5.70      | 3/2     | 4                         | 10                 | 3                      | 34.8-56.5            | 1.03 x 1.03          | Cut-and-try   |
| [22]  | 1.00      | 2/0     | 6                         | 4                  | multiple               | 7.0-18.0             | 0.10 x 0.03          | Cut-and-try   |
| [23]  | 1.00      | 3/2     | 11                        | 7                  | 2                      | 9.5-22.5             | N/A                    | Cut-and-try   |
| [24]  | 2.40      | 6/2     | 6                         | 7                  | 3                      | 37.0-92.0            | 0.65 x 0.65           | Cut-and-try   |
| Filter II | 3.09  | 5/1     | 1                         | 4                  | multiple               | 43.4-79.5            | 0.72 x 0.50           | Synthesis Design |

CF1: the center frequency of BPFs with widest bandwidth; Design parameters1: the design parameters of the passive filtering structure.

The simulated and measured $S$-parameters are compared in Fig. 13. It is apparent that these two groups of results are in good agreement, except for the ones near the upper passband edge, which is caused by the non-ideal tuning element. Under the condition that the impedance matching is better than 10.0 dB, the measured passband is from 4.16 to 6.04 GHz for $C_V = 1.0$ pF, and from 2.61 to 6.05 GHz for $C_V = 128$ pF, respectively. In addition, the TZ near the lower passband edge is shifted from 3.08 to 2.20 GHz when the capacitance value $C_V$ increases from 1.0 to 128 pF, while the one near the upper passband edge is kept fixed at about 6.41 GHz. If the center frequency is assumed as 4.33 GHz, the corresponding FBW can be varied from 43.4% to 79.5%. Within the passband, the average insertion loss is 1.05 dB for $C_V = 1.0$ pF, and 0.95 dB for $C_V = 128$ pF, respectively. Furthermore, the measured group variation is smaller than 0.48 ns for these two cases, indicating high linearity. The photograph of the presented reconfigurable-bandwidth filter is shown in Fig. 14. To highlight the merits of the proposed synthesis design, some comparisons with previous works are listed in Table 4. It is apparent that the biggest merit of this work is high-efficiency design of reconfigurable-bandwidth WB BPFs.

IV. CONCLUSION

A new synthesis design of equal-ripple and quasi-elliptic WB BPFs with independently tunable lower passband edge is presented in this paper. The lumped equivalent circuits of T- and H-shaped resonators for the design of equal-ripple and quasi-elliptic BPFs are derived directly in first-time. In addition, the initial parameters of all the proposed topologies can be determined directly. For verification, two prototypes with fixed and reconfigurable bandwidth are fabricated and measured. As the proposed topology owns minimum design parameters, whose initial values can be directly determined after knowing the specifications, i.e., equal-ripple level and bandwidth, it can be anticipated that this synthesis design can be beneficial to the high-efficiency design of reconfigurable-bandwidth WB BPFs.

![Photograph of filter II.](image)

**FIGURE 14.** Photograph of filter II.
REFERENCES

[1] J.-S. Hong and M. J. Lancaster, *Microstrip Filters for RF/Microwave Applications*. New York, NY, USA: Wiley, 2001.

[2] L. Zhu, S. Sun, and R. Li, *Microwave Bandpass Filters for Wideband Communications*. Hoboken, NJ, USA: Wiley, 2012.

[3] C.-C. Huang, W.-T. Fang, and Y.-S. Lin, “Miniaturization of broadband stub bandpass filters using bridged-T coils,” *IEEE Access*, vol. 6, pp. 20164–20173, 2018.

[4] X.-K. Bi, X. Zhang, W.-S. Wong, T. Yuan, and S.-H. Guo, “Design of equal-ripple dual-wideband bandpass filter with minimum design parameters based on cross-shaped resonator,” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, to be published, doi: 10.1109/TCSII.2019.2951781.

[5] R. Li, S. Sun, and L. Zhu, “Synthesis design of ultra-wideband bandpass filters with composite series and shunt stubs,” *IEEE Trans. Microw. Theory Techn.*, vol. 57, no. 3, pp. 684–692, Mar. 2009.

[6] C.-J. Chen, “A coupled-line coupling structure for the design of quasi-elliptic bandpass filters,” *IEEE Trans. Microw. Theory Techn.*, vol. 66, no. 4, pp. 1921–1925, Apr. 2018.

[7] A. Saghir, A. Quddious, S. Arain, P. Vryonides, and S. Nikolaou, “Single/Dual-BPF using coupled-line stepped impedance resonators (CLSIR),” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, vol. 66, no. 9, pp. 1497–1501, Sep. 2019.

[8] X. Jin, X. Huang, D. Chen, and C. Cheng, “Response diversity of stub-loaded ring bandpass filter based on commensurate line element: Single- and dual-band applications,” *IEEE Access*, vol. 7, pp. 25681–25689, 2019.

[9] H. Shamam and J.-S. Hong, “Input and output cross-coupled wideband bandpass filter,” *IEEE Trans. Microw. Theory Techn.*, vol. 55, no. 12, pp. 2562–2568, Dec. 2007.

[10] J. Xu, F. Xiao, Y. Cao, Y. Zhang, and X. Tang, “Compact microstrip filter with third-order quasi-elliptic bandpass response,” *IEEE Access*, vol. 6, pp. 63375–63381, 2018.

[11] Z. Li and K.-L. Wu, “Direct synthesis and design of wideband bandpass filters with composite series and shunt resonators,” *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 10, pp. 3789–3800, Oct. 2017.

[12] J. Garcia-Garcia, J. Bonache, and F. Martin, “Application of electromagnetic bandgaps to the design of ultra-wideband bandpass filters with good Out-of-Band performance,” *IEEE Trans. Microw. Theory Techn.*, vol. 54, no. 12, pp. 4136–4140, Dec. 2006.

[13] X.-K. Bi, T. Cheng, P. Cheong, S.-K. Ho, and K.-W. Tam, “Wide-band bandpass filter with reconfigurable bandwidth and fixed notch bands based on terminated cross-shaped resonator,” *IET Microw. Antennas Propag.*, vol. 13, no. 6, pp. 796–803, 2019.

[14] X.-K. Bi, T. Cheng, P. Cheong, S.-K. Ho, and K.-W. Tam, “Design of dual-band bandpass filters with fixed and reconfigurable bandwidths based on terminated cross-shaped resonators,” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, vol. 66, no. 3, pp. 317–321, Mar. 2019.

[15] X.-K. Bi, X. Zhang, G.-L. Huang, and T. Yuan, “Compact microstrip NWB/DWB BPFs with controllable isolation bandwidth for interference rejection,” *IEEE Access*, vol. 7, pp. 49169–49176, 2019.

[16] D. Tian, Q. Feng, and Q. Xiang, “Synthesis applied 4th-order constant absolute bandwidth frequency-agile bandpass filter with cross-coupling,” *IEEE Access*, vol. 6, pp. 72287–72294, 2018.

[17] M. Jung and B.-W. Min, “A widely tunable compact bandpass filter based on a switched varactor-tuned resonator,” *IEEE Access*, vol. 7, pp. 95178–95185, 2019.

[18] A. Zakharov, S. Rozenko, and M. Ilenchenko, “Varactor-tuned microstrip bandpass filter with loop hairpin and combline resonators,” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, vol. 66, no. 6, pp. 953–957, Jun. 2019.

[19] A. Iqbal, A.-W. Ahmad, A. Smida, and N.-K. Mallat, “Tunable SIW bandpass filters with improved upper stopband performance,” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, to be published, doi: 10.1109/TCSII.2019.2936495.

[20] A. Zakharov, S. Rozenko, S. Litvintsev, and M. Ilenchenko, “Hairpin resonators in varactor-tuned microstrip bandpass filters,” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, to be published, doi: 10.1109/TCSII.2019.2953247.

[21] T. Cheng and K.-W. Tam, “A wideband bandpass filter with re-configurable bandwidth based on cross-shaped resonator,” *IEEE Microw. Compon. Lett.*, vol. 27, no. 10, pp. 909–911, Oct. 2017.

[22] G. Zhang, Y. Xu, and X. Wang, “Compact tunable bandpass filter with wide tuning range of centre frequency and bandwidth using short coupled lines,” *IEEE Access*, vol. 6, pp. 2962–2969, 2018.

[23] R. Gómez-García, J.-M. Muñoz-Ferreras, J. Jiménez-Campillo, F. Braña-Roncati, and P. Martín-Iglesias, “High-order planar bandpass filters with electronically-reconfigurable passband width and flatness based on adaptive multi-resonator cascades,” *IEEE Access*, vol. 7, pp. 11010–11019, 2019.

[24] S. Arain, P. Vryonides, A. Quiddous, and S. Nikolaou, “Reconfigurable BPF with constant centre frequency and wide tuning range of bandwidth,” *IEEE Trans. Circuits Syst. II, Exp. Briefs*, to be published, doi: 10.1109/TCSII.2019.2938741.

[25] M. C. Horton and R. J. Wenzel, “General theory and design of optimum quarter-wave TEM filters,” *IEEE Trans. Microw. Theory Techn.*, vol. MTT-13, no. 3, pp. 316–327, May 1965.

[26] R. Levy and L. F. Lind, “Synthesis of symmetrical branch-guide directional couplers,” *IEEE Trans. Microw. Theory Techn.*, vol. MTT-16, no. 2, pp. 80–89, Feb. 1968.

[27] H. J. Carlin and W. Kohler, “Direct synthesis of band-pass transmission line structures,” *IEEE Trans. Microw. Theory Techn.*, vol. MTT-13, no. 3, pp. 283–297, May 1965.

[28] Y.-C. Chio, J.-T. Kuo, and E. Cheng, “Bandwidth quasi-Chebyshev bandpass filters with multimode stepped-impedance resonators (SIIRs),” *IEEE Trans. Microw. Theory Techn.*, vol. 54, no. 8, pp. 3352–3358, Aug. 2006.

[29] J. Oda, C.-P. Chen, K. Kamata, T. Kato, N. Kato, T. Anada, and S. Takeda, “An iterative synthesis scheme for wideband filter based on parallel-coupled three-line,” in *Proc. Eur. Microw. Conf. (EuMC)*, 2013, pp. 889–892.
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