A Highly Independent Multiband Bandpass Filter Using a Multi-Coupled Line Stub-SIR With Folding Structure

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ABSTRACT The main problem in designing a multiband bandpass filter (BPF) is making each passband response highly independent, where each bandwidth of multiband BPF can be controlled and adjusted separately. To overcome this problem, this paper proposes a highly independent multiband BPF based on a multicoupled line stub-SIR with a folding structure. The proposed multiband BPF is constructed as a multicoupled line to generate a highly independent inter-passband. Moreover, the multiband performance is produced separately and independently by using three sets of resonators: resonator A₁/A₂ (R₁A and R₂A), resonator B₁/B₂ (R₁B and R₂B), and resonator C₁/C₂ (R₁C and R₂C). The three passband frequencies can be independently arranged and designed. To miniaturize the multiband BPF, a folding structure is also proposed. As a result, the multiband BPF has a compact size that is reduced by over 61.29% compared to previous structures. The even-odd excitation model and the equivalent circuit model are used to analyze the multiband BPF structure. This BPF is designed for GPS applications at 1.57 GHz, WCDMA (3G) at 1.8 GHz, WLAN (WiFi) at 2.4 GHz, LTE (4G) at 2.6 GHz, and 5G communication at 3.5 GHz. To evaluate and validate the proposed structure of the multiband BPF, the circuits are fabricated and tested. The simulated and measured results of the multiband BPF show good agreement. In conclusion, the proposed multiband BPF structure has a highly independent inter passband response and a compact size.

INDEX TERMS Compact, folding structure, highly independent, multiband BPF.

I. INTRODUCTION

In modern wireless communications, multiband bandpass filters (BPFs) are essential components in a variety of wireless technologies with simultaneous multistandard applications [1], [2], such as global positioning system (GPS), wireless local area network (WLAN/WiFi), global system for mobile communications (GSM/2G), wideband code division multiplexing access (WCDMA/3G), long term evolution (LTE/4G), [2], [3] and 5G communications [4]. As an important subsystem of multiband wireless transceivers, multiband BPFs with excellent performance and the capability to support wireless technology by concurrently reducing interference and noise at multiple frequencies are in high demand [5], [6].

The fundamental part of multiband BPF design lies in the resonator structure. Recently, researchers have explored a multiband BPF base on a crossed resonator [7]. In [7], the crossed resonator successfully produced multiresonant frequencies. Furthermore, to reduce the multiband BPF size, the assembled resonator was proposed by [8], and the meander coupled-line resonator was investigated by [9]. As a result, the size of multiband BPFs is more compact. However, each passband response is not independent. Furthermore, the lumped element structure was suggested by [10]. However, the addition of lumped elements makes the filter structure more complex. Other multiband BPFs have been generated from substrate integrated waveguide (SIW) resonators [11], [12] and high temperature superconducting...
resonators [13]. However, for these methods, manufacturing is difficult, and the inter-passband have a low independence. The promising method of multiband BPFs is based on step impedance resonators (SIRs) such as the stub-loaded SIR [14], [15], quarter-wavelength SIR [16], asymmetric SIR [17], [18], coupled-line SIR [19], multimode SIR [20], and trisection SIR [21]. These resonators have achieved a good insertion loss $|S_{21}|$ and reflection coefficient $|S_{11}|$, and they have a low-isolation interband. However, these resonator structures have drawbacks such as a large size, complex structure, and non independent structure.

The dual-band independent BPFs was proposed by [22], and the short stub-loaded SIR and tri-section SIR resonator were used. However, these BPFs showed low independency at the upper bands and still worked at the dual passband. A tunable dual-band bandpass filter was proposed by [23], [24], but this BPF has drawbacks such as requiring active components and a power supply. A coplanar waveguide (CPW) bandpass was proposed by [25], but this BPF has a complex structure. An independent multiband BPF was proposed by [26]. To produce the multiband BPF, separated electric and magnetic coupling (SEMC) was proposed. However, this BPF has low independence at the first passband. Another independent multiband BPF was proposed by [27], but the insertion loss values are poor. The open-loop resonator was proposed by [28], but this filter also has low independence at the first passband. A dual composite right- and left-handed (D-CRLH) resonator was proposed by [29], but the resonator structure is quite complex, hard to extract, with non independent response.

This paper proposes a highly independent multiband bandpass filter using a multicoupled line stub-SIR with a folding structure, as shown in Fig 1(a) to 1(d). Fig 1(a), 1(b), 1(c), and 1(d) illustrate a single-band BPF, an independent dual-band BPF, an independent multiband BPF, and a miniaturized independent multiband BPF, respectively. The proposed multiband BPF is constructed with a multicoupled line to generate a highly independent inter passband. Moreover, the multiband performance is produced separately and independently by using three sets of resonators: resonator $A_1/A_2$ ($R_{A1}$ and $R_{A2}$), resonator $B_1/B_2$ ($R_{B1}$ and $R_{B2}$), and resonator $C_1/C_2$ ($R_{C1}$ and $R_{C2}$). Resonators A and B have an L-shaped structure. These resonators are placed at the upper and lower feeding lines, respectively. Furthermore, resonators A and B are designed for the first passband ($f_{CA}$) and the second passband ($f_{CB}$), respectively. The third resonator, resonator C, is a stub-step impedance resonator (SIR) embedded at the center of resonator B ($R_{B1}$ and $R_{B2}$) to generate the third passband ($f_{CC}$), which can be clearly seen in Fig 1(c). Moreover, to miniaturize the multiband BPF, a folding structure is also proposed. This method can reduce the size of a multiband BPF by more than half, as shown in Fig 1(d). The three passband frequency responses can be tuned independently, as shown in Fig 2(a) to 2(c).

Here, high independence means that each passband response does not depend on another passband, so the filter response can be controlled and adjusted separately and individually. Fig 2(a) shows that the first passband ($f_{CA}$) can be varied by tuning the dimensions of resonator $A_1/A_2$ ($R_{A1}$ and $R_{A2}$). The $f_{CB}$ and $f_{CC}$ can be adjusted by varying the dimensions of resonator $B_1/B_2$ ($R_{B1}$ and $R_{B2}$) and resonator $C_1/C_2$ ($R_{C1}$ and $R_{C2}$), respectively.

The significant contributions of this paper are as follows.

1) Most previous studies have focused on multiband BPF design [8]–[30] without considering the independence of inter-passbands. As novelty, we focus on highly independent multiband BPF. This structure is making each passband response highly independent, where each band of multiband BPF can be controlled and adjusted separately.

2) The important feature of multicoupled line stub-SIR with a folding resonator structure is its highly independent performance with a simple structure and it can be easily analyzed and manufactured.

3) The three passband frequencies could be tuned independently and separately, as shown by the independent response of the insertion loss $|S_{21}|$ and the reflection coefficients $|S_{11}|$ at each passband and by the different direction of the surface current flows.

4) The folding structure method successfully reduces the multiband BPF dimensions by over 61.29 %, which makes the proposed multiband BPF very compact. The multiband BPF has a size of 0.32 $\lambda_G \times 0.31 \lambda_G$. After the folding structure was applied, the multiband BPF size became 0.32 $\lambda_G \times 0.12 \lambda_G$, where $\lambda_G$ is the wavelength of the fundamental frequency.

5) The proposed multiband BPF has performance advantages such as an excellent insertion loss $|S_{21}|$, a reflection coefficient $|S_{11}|$ with good transmission zeros and
an isolation interband. The validity of the performance is shown by the excellent agreement between the simulated and measured results.

6) Finally, the proposed multiband BPF structure can be applied for 5G communication at 3.5 GHz.

To optimize the structure of the multiband BPF, the Momentum Advance Design Systems (ADS) simulation was used.

Furthermore, the multiband BPF is fabricated on an RT/Duroid 5880 substrate with a permittivity of 2.2 and a thickness of 1.575 mm.

To validate the proposed method, the multiband BPF has been tested. The structure of this paper is as follows. The first section gives a brief overview of multiband BPF design and research. The second section describes and analyzes the structure of coupled-line resonators, single-band BPFs, independent dual-band BPFs, and independent multiband BPFs. The third section focuses on the implementation and validation of the resonator structure and its miniaturization strategy. Finally, section 4 concludes this research.

II. PROPOSED COUPLED-LINE RESONATOR STRUCTURE AND THE DESIGN OF SINGLE-BAND, DUAL- BAND, AND MULTIBAND BPFS

A. COUPLING STRUCTURE AND EXTERNAL QUALITY FACTOR

The proposed single coupled-line resonator structure is shown in Fig 3(a). The input-coupled line is directly connected to the 50-Ω port 1 (P_IN), and the other end of the resonator is connected to the 50-Ω port 2 output (P_OUT). This structure is used to investigate the external quality factor (Qe), coupling coefficient (M), and |S_21| values under weak coupling with different gaps (s_1). Fig 3(b) shows the simulated results of odd-mode and even-mode frequency of |S_21| with varied gaps (s_1). The resonant frequencies (f_C) value are determined by odd-mode resonant frequency (f_odd-mode) and even-mode resonant frequency (f_even-mode). Usually, the f_odd-mode is higher than f_even-mode and the resonant frequencies (f_C) is in the middle of the f_odd-mode and f_even-mode. It can be seen that the f_odd-mode and f_even-mode excitations are shifted.
by different gap values \((s_1)\). However, the resonant frequencies \((f_c)\) remains stable, indicating that the gap variations do not affect the \(f_c\) values.

Furthermore, the coupling coefficient \((M)\) and the external quality factor \((Qe)\) should be determined. The coupling coefficient \((M)\) of the resonator structure can be calculated by [31], [32]

\[
M = f_o^2 - f_e^2 \over f_o^2 + f_e^2
\]

where \(f_o\) and \(f_e\) are the odd-mode and even-mode frequency excitations, respectively. The simulation results of \(M\) under different gaps \((s_1)\) is shown in Fig 4. Moreover, the external quality factor \((Qe)\) can be calculated by [31], [32]

\[
Qe = f_c \over BW
\]

where \(f_c\) and \(BW\) are the resonant frequency and the bandwidth, respectively.

**B. DESIGN OF SINGLE-BAND BPF**

The implementation of the coupled line as a single-band BPF is shown in Fig 5(a), and the coupling structure is shown in Fig 5(b). In the BPF topology, \(W_f, W_0, W_1,\) and \(W_2\) represent the width of the resonator, and \(l_f, l_0, l_1,\) and \(l_2\) represent the length of the resonator. Furthermore, the resonator structure consists of several transmission lines with different characteristic impedances \(Z_N\) (for \(N = f, 0, 1, 2\)) and corresponding electrical lengths \(\theta_N\) (for \(N = f, 0, 1, 2\)).

The coupling structure for the single-band BPF with resonators \(R_{A1}\) and \(R_{A2}\) as the main resonator is illustrated in Fig 5(b). This BPF is symmetrical on the \(T-T'\) plane. \(M_{S1}\) and \(M_{2L}\) represent the external quality factor and the electric coupling of input-output port, respectively. Furthermore, \(M_{MN}\) are the coupling matrix value between two resonator as shown in Table 1. The value coupling matrix was obtained using optimization technique [12]. The nonlinear least-squares and the minimum-maximum algorithm were used.

To investigate the BPF characteristics, the structure extractions should be provided as shown in Fig 6(a)-6(c). The equivalent circuit model for resonator \(A1\) \((R_{A1})\), the equivalent circuit model for the even-mode resonator, and the equivalent circuit model for the odd-mode resonator are shown in Fig 6.
To meet the resonant conditions \( Z_{IN} = \infty \) or \( Y_{IN} = 0 \), the denominator is set as follows \([5]\), \([6]\), \([31]\):

\[
Z_{IN-RA1 - odd\ mode} = \frac{1}{Z_{IN-RA1 - odd\ mode} + j \omega \theta A_{CRA1}}
\]

(3)

Moreover, Eq (4) can be deduced as

\[
Z_{IN-RA1 - odd\ mode} = \frac{j Z_1 \cot \left( \frac{\theta_1 + \theta_2}{2} \right)}{1 - 2 Z_1 \omega C_{RA1} \tan \left( \frac{\theta_1 + \theta_2}{2} \right)}
\]

(5)

To meet the resonant conditions \( Z_{IN} = \infty \) or \( Y_{IN} = 0 \), the denominator is set as follows \([5]\), \([6]\), \([31]\):

\[
1 - 2 Z_1 \omega C_{RA1} \tan \left( \frac{\theta_1 + \theta_2}{2} \right) = 0
\]

(6)

The resonant frequency can be calculated as follows:

\[
\omega_{CA} = \frac{1}{2 Z_1 C_{RA1} \tan \left( \frac{\theta_1 + \theta_2}{2} \right)}
\]

(7)

or the value of \( f_{CA} \) can be determined as

\[
f_{CA} = \frac{1}{\pi 4 Z_1 C_{RA1} \tan \left( \frac{\theta_1 + \theta_2}{2} \right)}
\]

(8)

where \( Z_1 \) is the characteristic impedance, \( C_{RA1} \) is the capacitance, \( \theta_1 + \theta_2 \) are the electrical lengths, and \( Z_1 = Z_2 \). The variation of \( Z_{IN-RA1 - odd\ mode} \) versus \( (\theta_1 + \theta_2) \) and \( C_{RA1} \) is shown in Fig 7.

The effect of \( \theta_1 + \theta_2 \) and \( C_{RA1} \) to \( Z_{IN-RA1 - odd\ mode} \) is shown in Fig 7. We can see that \( C_{RA1} \) has a lower impact on the impedance value of \( Z_{IN-RA1 - odd\ mode} \). However, the value of \( Z_{IN-RA1 - odd\ mode} \) has largely determined by the length of \( \theta_1 + \theta_2 \). For the value of \( \theta_1 + \theta_2 \) lower than \( 80^\circ \) and upper than \( 90^\circ \), \( Z_{IN-RA1 - odd\ mode} \) is more capacitive. Moreover, we can see that the value of \( \theta_1 + \theta_2 \) between \( 80^\circ \) and \( 90^\circ \), making \( Z_{IN-RA1 - odd\ mode} \) more inductive. Furthermore, the resonance occurs when the imaginary part of the \( Z_{IN-RA1 - odd\ mode} \) is zero.
The second section is focused on the dual-band BPF. The dual-band BPF response is obtained by adding the resonator B1/B2 (R_B1 and R_B2) as shown in Fig 8(a). The resonator B1/B2 (R_B1 and R_B2) is placed at the upper side of the feeding line. R_B1 and R_B2 are L-shaped and have a back-to-back coupled structure. Consequently, the network coupling structure is magnetic coupling (M_2) [40]. The detailed structure of the proposed dual-band BPF is shown in Fig 8(a), and the BPF is symmetrical on the T-T’ plane. In the dual-band BPF topology, W_0, W_f, W_1, W_2, W_3, and W_4 represent the width of the resonator and correspond to the characteristic impedances Z_N (for N = f, 0, 1, 2, 3, 4). Furthermore, the l_0, l_f, l_1, l_2, l_3, and l_4 values represent the length of the resonator, with corresponding electrical lengths θ_N (for N = f, 0, 1, 2, 3, 4).

As shown in Fig 8(b), the coupling structure of the dual-band BPF is expanded from the coupling structure of the single-band BPF in Fig 5(b). It can be seen that the additional coupling at the upper feeding line will change the coupling structure there. The complete coupling matrix of dual-band BPF structure is shown in Table 2.

The terminal 50-Ω port is connected to P_IN and P_OUT as the input-output port. To evaluate the resonant frequency (f_CB), the equivalent circuit model for resonator B1 (R_B1), as shown in Fig 9(a), should be provided. To simplify analysis and design, the values Z_3 and Z_4 are equal, but θ_1 ≠ θ_2. Because resonator R_B1 is L-shaped, the capacitance (C_{RB1}) value should be included in calculations [33]. The value coupling matrix was obtained using optimization technique [12]. The nonlinear least-squares and the minimum-maximum algorithm were used.

![Equivalent circuit model for resonator B1](image)

**FIGURE 9.** (a) Equivalent circuit model for resonator B1. (b) Equivalent circuit model for the even-mode resonator B1. (c) Equivalent circuit model for the odd-mode resonator B1.

**TABLE 2.** Coupling matrix of dual-band BPF structure.

| M_N  | M = S | M = 1 | M = 2 | M = 3 | M = 4 | M = L |
|------|------|------|------|------|------|------|
| N = S | 0    | 0.299| 0    | 0    | 0    | 0.103|
| N = 1 | 0.299| 0.380| 0.707| 0    | 0    | 0    |
| N = 2 | 0    | 0.707| 1.030| 0    | 0    | 0.932|
| N = 3 | -0.264| 0    | 0    | -26.271| -51.382| 0    |
| N = 4 | 0    | 0    | 0    | -51.382| -103.010| -0.699|
| N = L | 0.103| 0.932| 0    | 0    | -0.699| 0    |

Fig 9(a), 9(b), and 9(c) show the equivalent circuit models for R_B1, the even-mode resonator R_B1, and the odd-mode resonator R_B1, respectively. Z_IN−RB1−even mode can be determined as follows:

\[
Z_{IN−RB1−even mode} = -jZ_3 \cot \left( \frac{\theta_3 + \theta_4}{2} \right)
\]

Moreover, the Z_IN−RB1−odd mode has a shunt structure, so it can be calculated as

\[
Z_{IN−RB1−odd mode} = \frac{1}{-jZ_3 \tan \left( \frac{\theta_3 + \theta_4}{2} \right) + j2\omega CB C_{RB1}}
\]

Furthermore, Eq (10) can be deduced as

\[
Z_{IN−RB1−odd mode} = \frac{1}{1 - 2Z_3 \omega CB C_{RB1} \tan \left( \frac{\theta_3 + \theta_4}{2} \right)}
\]

To meet the resonant condition Z_IN = ∞ or Y_IN = 0, the denominator is set as follows:

\[
1 - 2Z_3 \omega CB C_{RB1} \tan \left( \frac{\theta_3 + \theta_4}{2} \right) = 0
\]

The resonant frequency can be calculated as follows:

\[
\omega CB = \frac{1}{2Z_3 C_{RB1} \tan \left( \frac{\theta_3 + \theta_4}{2} \right)}
\]

or the value of f_CB can be determined as

\[
f_CB = \frac{1}{\pi 4Z_3 C_{RB1} \tan \left( \frac{\theta_3 + \theta_4}{2} \right)}
\]

where Z_3 is the characteristic impedance, C_{RB1} is the capacitance, θ_2 and θ_3 are the electrical lengths, and Z_3 = Z_4.

**D. DESIGN OF MULTIBAND BPF**

The design of a highly independent multiband BPF and its miniaturization strategy are investigated in this section. Fig 10(a) and 10(b) show the proposed filter topology of an independent multiband BPF structure and the coupling structure, respectively. Furthermore, Table 3 shows the coupling matrix of multi-band BPF structure. The value coupling matrix was obtained using optimization technique [12]. The nonlinear least-squares and the minimum-maximum algorithm were used.

To produce the third passband (f_{CC}), the additional stub-stepped impedance resonators (Stub-SIR), namely, resonator...
$C1/C2$ ($R_{C1}$ and $R_{C2}$), are embedded in resonator B ($R_{B1}$ and $R_{B2}$), as shown in Fig 10(a). Resonator $C1/C2$ has the SIR structure composed of two transmission lines such as ($W_5$, $l_5$, $Z_5$, $\theta_5$) and ($W_6$, $l_6$, $Z_6$, $\theta_6$). In the multiband BPF topology, $W_f$, $W_6$, $W_1$, $W_2$, $W_3$, $W_4$, $W_5$, and $W_6$ represent the width of the resonator and correspond to the characteristic impedances $Z_N$ (for $N = f$, 0, 1, 2, 3, 4, 5, 6). Furthermore, $l_1$, $l_0$, $l_1$, $l_2$, $l_3$, $l_4$, $l_5$, and $l_6$ represent the length of the resonator with corresponding electrical lengths $\theta_N$ (for $N = f$, 0, 1, 2, 3, 4, 5, 6).

The multiband BPF coupling structure is shown in Fig 10(b). It can be seen that the additional Stub-SIR has changed the coupling structure at the upper resonator ($R_{B1}$ and $R_{B2}$). Furthermore, the structured resonators are illustrated as $R_{An}$ (for $n = l$, 2), $R_{Bn}$ (for $n = l$, 2), and $R_{Cn}$ (for $n = l$, 2) for resonator $A1/A2$, resonator $B1/B2$, and resonator $C1/C2$, respectively.

To evaluate the resonant frequency, the equivalent circuit models for resonators $B1$ and $C1$ ($R_{B1}$ and $R_{C1}$) should be investigated. They are shown in Fig 11(a). Fig 11(a), 11(b) and 11(c) show the equivalent circuit models for resonators $B1$ and $C1$ ($R_{B1}$ and $R_{C1}$), the even-mode resonator ($R_{B1}$ and $R_{C1}$), and the odd-mode resonator ($R_{B1}$ and $R_{C1}$), respectively. The even-mode and odd-mode resonant frequencies can be determined from the impedance condition $Z_{IN} = \infty$ or the admittance condition $Y_{IN} = 0$ [5, 6, 31].

First, $Z_{IN6}$ can be calculated as follows:

$$Z_{IN6} = -j2Z_6 \cot \theta_6$$  (15)

$Z_{IN5}$ can be determined as follows:

$$Z_{IN5} = \frac{jZ_5^2 \tan \theta_5 - j2Z_5Z_6 \cot \theta_6}{Z_5 + 2Z_6 \tan \theta_5 \cot \theta_6}$$  (16)

Then, the impedance $Z_{IN-RB1-RC1-even mode}$ can be determined as follows:

To meet the resonant condition $Z_{IN} = \infty$ or $Y_{IN} = 0$, the denominator (17), as shown at the bottom of the next page, is calculated as follows [5, 6, 31]:

$$2Z_3Z_5 + 2Z_3Z_6 \tan \theta_5 \cot \theta_6 - 4Z_5^2 \tan \theta_5 \times \tan \theta_5 + 2Z_5Z_6 \tan \theta_5 \cot \theta_6 = 0$$  (18)

Eq. (19) can be simplify as follows:

$$2K_1 + 2K_1K_2 \tan \theta_6 \cot \theta_6 - 4 \tan \theta_3 \tan \theta_5 + 4K_2 \tan \theta_5 \cot \theta_6 = 0$$  (19)

$$K_1 (2 + 2K_2 \tan \theta_6 \cot \theta_6) - 4 \tan \theta_3 (\tan \theta_5 + K_2 \cot \theta_6) = 0$$  (20)
TABLE 3. Coupling matrix of multi-band BPF structure.

| M₂N  | M = S | M = 1 | M = 2 | M = 3 | M = 4 | M = 5 | M = 6 | M = 7 | M = 8 | M = L |
|------|------|------|------|------|------|------|------|------|------|------|
| N = S | 0    | 6.058| 0    | 0    | 0    | 0    | 0.906| 0    | 5.547|
| N = 1 | 6.058| 83.428| 61.748| 70.393| 0    | 0    | -2.073| 0    | 0    |
| N = 2 | 0    | 61.748| 61.748| 103.026| 0    | 0    | 0    | 0    | 0    |
| N = 3 | 0    | 70.393| 103.026| 0    | 0    | 0    | 0    | 0    | 0    |
| N = 4 | 0    | 0    | 0    | 0    | 0    | -7.265| 0.047| 0    | 0    |
| N = 5 | 0    | 0    | 0    | 0    | -7.265| 1.788| 0    | 0    |
| N = 6 | 0    | 0    | 0    | 0    | 0.047| 1.1788| 0.0714| 0    | -2.004|
| N = 7 | 0    | 0    | 0    | 0    | 0    | 0    | 0.001| 0.207|
| N = 8 | 0    | 0    | 0    | 0    | 0    | 0    | 0.001| -0.369|
| N = L | 5.547| 0    | 0    | 0    | 0    | -2.004| 0    | -0.369|

FIGURE 11. (a) Equivalent circuit model for upper resonator (B₁ and C₁), (b) Equivalent circuit model of upper resonator (B₁ and C₁) for the even-mode, (c) Equivalent circuit model of upper resonator (B₁ and C₁) for the odd-mode.

where

$$K_1 = \frac{Z_3}{Z_5}$$

(22)

$$K_2 = \frac{Z_6}{Z_5}$$

(23)

Moreover, \(Z_{IN-RB1-RC1-odd\ mode}\) has a shunt structure as shown in Fig 11(c), so it can be calculated as

$$\frac{1}{Z_{IN-RB1-RC1-odd\ mode}} = \frac{1}{-jZ_3 \tan \theta_3} + j2\omega_3 C_{RB1}$$

(24)

Furthermore, Eq (27) can be deduced as

$$Z_{IN-RB1-RC1-odd\ mode} = \frac{jZ_3 \tan \theta_3}{1 - 2Z_3\omega_3 C_{RB1} \tan \theta_3}$$

(25)

To meet the resonant condition \(Z_N = \infty\) or \(Y_N = 0\), the denominator is set as follows [5], [6], [31]:

$$1 - 2Z_3\omega_3 C_{RB1} \tan \theta_3 = 0$$

(26)

where \(Z_N\) (for \(N = 3, 5, 6\)) and \(\theta_N\) (for \(N = 3, 5, 6\)) are the characteristic and electrical lengths, respectively.
III. IMPLEMENTATION OF HIGHLY INDEPENDENT SINGLE-BAND, DUAL-BAND, AND MULTIBAND BPFS

A. IMPLEMENTATION OF SINGLE-BAND BPF STRUCTURE

The parametric simulation results are provided in Fig 12(a) and 12(b), which show the $|S_{21}|$ characteristics with varied lengths ($l_2$) and the $|S_{11}|$ characteristics with varied lengths ($l_2$), respectively. It is clearly seen from Fig 12(a) that the frequency center can be adjusted by varying the length ($l_2$). This figure also shows that the optimum value of $l_2 = 13$ mm.

Moreover, the magnitude of $|S_{11}|$ must be set to lower than $-10$ dB to make sure that the reflection coefficient is low. These simulation results show that the value of the length $l_2$ corresponds with $\theta_2$ and has a positive contribution to $f_{CA}$. This interesting finding confirms the usefulness of Eq (8) and shows that $f_{CA}$ is affected by $\theta_2$ when $\theta_1$ is set to a constant value.

Fig 13 (a) and 13 (b) show the equivalent circuit model of a highly independent single-band BPF and a photograph of a fabricated single-band BPF, respectively. The equivalent circuit model of the single-band BPF is symmetrical on the $T-T'$ plane.

The resonator structure consists of several passive components such as capacitor $C_{An}$ (for $n = 1, 2, 3, 4, 5, 6, 7$) and inductor $L_{An}$ (for $n = 1, 2, 3, 4, 5$). Because resonators $R_{A1}$ and $R_{A2}$ are L-shaped and the back-to-back structure is coupled, the coupled network is a magnetic coupling structure ($M_1$) [33]. The corner part of the resonator becomes more capacitive and is thus replaced as a capacitor. Fig 13(a) shows the detailed equivalent circuit.

The simulation and optimization results are obtained with the ADS software. The overall dimensions of the single-band BPF are $W_0 = 3$ mm, $W_f = 1$ mm, $W_1 = 1$ mm, $W_2 = 1$ mm, $l_0 = 5$ mm, $l_f = 2$ mm, $l_1 = 23$ mm, $l_2 = 13$ mm, $s_1 = 0.5$ mm and $s_2 = 0.5$ mm. Furthermore, the values of the equivalent circuit models are $C_{A1} = 15$ pF, $C_{A2} = 1.5$ $\mu$F, $C_{A3} = 15$ pF, $C_{A4} = 0.1$ pF, $C_{A5} = 150$ pF, $C_{A6} = 75$ pF, $C_{A7} = 75$ pF, $L_{A1} = 0.01$ pH, $L_{A2} = 300$ pH, $L_{A3} = 0.01$ pH, $L_{A4} = 89$ pH, and $L_{A5} = 14$ pH. The single-band BPF is fabricated on an RT/Duroid 5880 substrate and
is soldered to a 50-Ω input/output port, as shown in Fig 13(b).

To evaluate the performance, the device under test (DUT) should be connected to a vector network analyzer (VNA). The R & S ZNB VNA was used to measure the BPF. Fig 14(a), 14(b), and 14(c) show the simulated, measured, equivalent circuit model, and coupling matrix model results of $|S_{21}|$ and $|S_{11}|$, with a bandwidth at $|S_{21}|$ of $-3$ dB. Fig 14(d) shows the surface current flow at $f_{CA}$.

The simulated/measured results of the single-band BPF are 3.25 GHz/3.27 GHz, 8.09 %/9.14 %, $-0.35$ dB/$-0.8$ dB, $-17.0$ dB/$-19.6$ dB for the $f_{CA}$, fractional bandwidth (FBW), $|S_{21}|$, and $|S_{11}|$, respectively. The single-band BPF
design strategies are validated by very good agreement between the measured and simulated results.

**B. IMPLEMENTATION OF INDEPENDENT DUAL-BAND BPF STRUCTURE**

This section discusses the optimization and implementation of an independent dual-band BPF. The dual-band BPF results are provided in Fig 15(a) and 15(b), which show the magnitude of the $|S_{21}|$ characteristics with varied lengths ($l_2$) and the $|S_{11}|$ characteristics with varied lengths ($l_2$), respectively. Fig 15(a) shows that varying the value of length ($l_2$) can tune the $|S_{21}|$ passband of $f_{CA}$ while the frequency $f_{CB}$ remains constant/fixed. This result indicates that the proposed dual-band BPF has highly independent responses and illustrates that the insertion loss value of the passband is higher than $-3$ dB.

Fig 15(b) clearly shows that varying the length ($l_2$) can tune the value $|S_{11}|$ at passbands of frequency $f_{CA}$, whereas
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FIGURE 22. Simulation results of (a) $|S_{21}|$ characteristics with varied length $l_6$ and (b) $|S_{11}|$ characteristics with varied length $l_6$.

this change has only a slight influence on the passband $f_{CB}$. It is also observed that the value $|S_{11}|$ is better than $-10$ dB, which indicates that the dual-band BPF has a low reflection coefficient. Therefore, we can adjust the length of $l_2$ to tune the $f_{CA}$ passband. Increasing $l_2$ will make the $f_{CA}$ shift to a lower frequency, while decreasing $l_2$ will make the $f_{CA}$ shift to a higher frequency.

Fig 16(a) and 16(b) illustrate the magnitude of the $|S_{21}|$ characteristics for varying lengths ($l_4$) and the $|S_{11}|$ characteristics for varying lengths ($l_4$), respectively. It is clearly seen that by varying the length ($l_4$), the frequency of $f_{CB}$ will shift independently. However, $f_{CA}$ will be stable. It is also shown that the values of $|S_{11}|$ and $|S_{21}|$ can be optimized to better than $-10$ dB and $-3$ dB, respectively. The passband $f_{CB}$ can be controlled by the length ($l_4$). This finding confirms the usefulness of Eq (14) and shows that $f_{CA}$ is affected by $\theta_4$ when $\theta_3$ is set as a constant value, with $\theta_4$ corresponding to $l_4$. Therefore, this result shows that the two passband frequencies could be tuned independently and separately and demonstrates the successful design of a highly independent dual-band BPF.

Fig 17(a) and 17(b) show the equivalent circuit model of the independent dual-band BPF and a photograph of the fabricated dual-band BPF, respectively. The value of the magnetic coupling of the structures is indicated by $M$ and is shown at the upper and lower resonators. The dimensions of the dual-band BPF and the values of the equivalent circuit model after optimization using the ADS software are shown in Table 4.

As shown in Fig 17(a), the equivalent circuit model consists of upper and lower feeding lines. Due to the high magnetic field, the coupling inter resonator is drawn as magnetic coupling. However, the equivalent circuit model at the center of the dual-band BPF is shown to be a capacitive circuit. Moreover, the photograph of the fabricated dual-band BPF shown in Fig 17(b) illustrates the complete dual-band BPF with 50-Ω port termination.

To verify the proposed BPF design, the fabricated results of the dual-band BPF must be tested. Fig 18 (a), 18(b), and 18(c) show the simulated, measured, equivalent circuit model, and coupling matrix model results of $|S_{21}|$, $|S_{11}|$, $|S_{21}|$ for a bandwidth $f_{CB}$ and $f_{CA}$ of $-3$ dB, respectively.

The results of simulated and measured dual-band BPFs are shown in Table 5.

Fig 19 (a) and 19(b) show the surface current flow at passband $f_{CA}$ and passband $f_{CB}$, respectively. The surface current distribution at the $f_{CA}$ band has flowed at the lower part of the feeding structure or it has flowed at the lower resonator. However, the surface current distribution at the $f_{CB}$ band has flowed at the upper resonator. These results indicate

| $N$ | $W_N$ (mm) | $l_N$ (mm) | $s_N$ (mm) |
|-----|------------|------------|------------|
| 1   | 1          | 2          | -          |
| 0   | 3          | 5          | -          |
| 1   | 1          | 23         | 0.5        |
| 2   | 1          | 13         | 0.5        |
| 3   | 1          | 23         | 0.5        |
| 4   | 1          | 44         | 0.5        |
| 5   | 1          | 5          | -          |
| 6   | 7          | 4.75       | -          |

| $f_{CB}$ | $f_{CA}$ |
|----------|----------|
| 1         | 15       |
| 2         | 265      |
| 3         | 1        |
| 4         | 450      |
| 5         | 1.1      |
| 6         | 16       |
| 7         | 225      |
| 8         | 120      |
| 9         | 60       |
| 10        | 0.1      |

TABLE 6. Dimensions of the multiband BPF and values of the equivalent circuit model.
that the resonator is working independently and separately. The dual-band BPF design strategies are validated by good agreement between the measured and simulated results.

### C. IMPLEMENTATION OF INDEPENDENT MULTIBAND BPF STRUCTURE

The parametric simulation results of the highly independent multiband BPF are shown in Fig 20(a) – 22(b). In our BPF design scenario, the value of the passbands are as follows: $f_{CA} > f_{CC} > f_{CB}$. The promising findings show that by varying the length ($l_2$), the passband $f_{CA}$ can be tuned independently, as shown in Fig 20(a) and 20(b). By increasing the length ($l_2$), the passband $f_{CA}$ will shift to a lower frequency. Varying the length ($l_2$) will not change the $f_{CB}$ passband or the $f_{CC}$ passband. This result demonstrates that the passband $f_{CA}$ is more independent than the $f_{CB}$ and $f_{CC}$ passbands.

Fig 21(a) and 21(b) illustrate the $|S_{21}|$ characteristics with varied length ($l_4$) and the $|S_{11}|$ characteristics with varied length ($l_4$), respectively. There is a significant correlation between the length ($l_4$) and the $f_{CB}$ passband. Fig 21(a) shows that by increasing the length ($l_4$), the $f_{CB}$ passband will shift to a lower frequency. This change is not accompanied by a shift of the $f_{CA}$ passband or the $f_{CC}$ passband. Therefore,

| TABLE 7. Comparison of multiband BPF results. |
|-----------------------------------------------|
| Parameters | $f_{CA}$ | $f_{CC}$ | $f_{CB}$ |
| Sim | Meas | Sim | Meas | Sim | Meas |
| $f_c$ (GHz) | 1.70 | 1.65 | 2.55 | 2.55 | 3.25 | 3.25 |
| FBW (%) | 20.0 | 15.90 | 11.76 | 11.40 | 10.27 | 9.10 |
| $|S_{11}|$ dB | -0.58 | -0.39 | -0.69 | -0.92 | -0.71 | -0.80 |
| $|S_{21}|$ dB | -21.6 | -21.9 | -22.6 | -17.7 | -20.2 | -21.0 |
| Isolation interband (ISO) | ISO1 = -15.96 dB at 2.01 GHz (sim) | - | - |
| | ISO2 = -19.05 dB at 2.01 GHz (meas) | - | - |
| | ISO2 = -23.04 dB at 2.75 GHz (sim) | - | ISO2 = -24.95 dB at 2.75 GHz (meas) | - | - |

*note: in our design, $f_{CA} < f_{CC} < f_{CB}$, ISO1 = isolation between $f_{CA}$ and $f_{CC}$, ISO2 = isolation between $f_{CC}$ and $f_{CA}$.
the $f_{CB}$ passband is more independent than the $f_{CA}$ and $f_{CC}$ passbands. The insertion loss values are better than $-3$ dB, and the reflection coefficient is lower than $-10$ dB.

Fig 22(a) and 22(b) illustrate the extraction of the $|S_{21}|$ and $|S_{11}|$ characteristics with varied length ($l_6$) and show that by varying the length ($l_6$), the values of $|S_{21}|$ and $|S_{11}|$ at the $f_{CC}$ passband will shift. The $f_{CC}$ passband will be shifted to a lower frequency by increasing the length ($l_6$) or to a higher frequency by decreasing the length ($l_6$). However, these changes have no effect on the $f_{CA}$ passband or the $f_{CB}$ passband. These results indicate that the $f_{CC}$ passband is more independent than the $f_{CA}$ and $f_{CB}$ passbands. In summary, the passbands $f_{CA}$, $f_{CB}$, and $f_{CC}$ are separately and independently influenced by length ($l_2$), length ($l_4$), and length ($l_6$), respectively. They are working with high independence. This important finding is proof that the designed multiband bandpass filter with a multicoupled line stub-SIR and a folding structure is highly independent.

Fig 23. (a) and 23(b) show the equivalent circuit model of the independent multiband BPF and a photograph of the fabricated multiband BPF, respectively. The dimensions of multiband BPF and the values of the equivalent circuit after optimization using the ADS software are shown in Table 6.

This multiband BPF structure also has a back-to-back resonator, so the coupling inter resonator is drawn as magnetic coupling. The third passband ($f_{CC}$) is generated by the SIR-structure, and the equivalent circuit model is clearly illustrated in Fig 23(a). Finally, the photograph of the fabricated multiband BPF with a 50-$\Omega$ port termination has a size of $0.32\lambda_G \times 0.31\lambda_G$, were $\lambda_G$ is the wavelength of the fundamental frequency [34], [35] as shown in Fig 23(b).

After the multiband BPF has been fabricated, it should be tested by VNA for verification. Fig 24 (a), 24(b), and 24(c) show the simulated, measured, equivalent circuit model, and coupling matrix model results of $|S_{21}|$, $|S_{11}|$, $|S_{21}|$ for a bandwidth $f_{CB}$ of $-3$ dB, and $f_{CA}$ of $-3$ dB, respectively.

### TABLE 8. Coupling matrix of miniaturized independent multiband BPF.

| $M_{xx}$ | $M = S$ | $M = 1$ | $M = 2$ | $M = 3$ | $M = 4$ | $M = 5$ | $M = 6$ | $M = 7$ | $M = 8$ | $M = L$ |
|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| $N = S$  | 4.930  | 0      | 0      | 0      | 0      | 0      | 1.114  | 0      | 6.202  |
| $N = 1$  | 76.672 | 34.705 | 0      | 0      | 0      | -1.954 | 0      | 0      | 0      |
| $N = 2$  | 34.705 | 0      | 59.158 | 0      | 0      | 0      | 0      | 0      | 0      |
| $N = 3$  | 0      | 64.511 | 59.158 | 0      | 0      | 0      | 0      | 0      | 0      |
| $N = 4$  | 0      | 0      | 0      | -78.887| 0      | 0      | 8.937  | 0      | 0      |
| $N = 5$  | 0      | 0      | 0      | -78.887| 0      | 8.937  | 0      | 0      | 0      |
| $N = 6$  | -1.954 | 0      | 0      | 0.044  | 9.156  | 0      | 0      | -2.522 |
| $N = 7$  | 1.114  | 0      | 0      | 0      | 0      | 0.001  | 0.273  | 0      |
| $N = 8$  | 0      | 0      | 0      | 0      | 0      | 0.273  | 0.001  | 0.616  |
| $N = L$  | 6.202  | 0      | 0      | 0      | 0      | 0      | -2.522 | 0      | 0.616  |

### FIGURE 27. Coupling matrix.

### TABLE 9. Dimensions of miniaturized multiband BPF.

| $N$ | $W_x$ (mm) | $l_x$ (mm) | $s_x$ (mm) | $N$ | $W_x$ (mm) | $l_x$ (mm) |
|-----|------------|------------|------------|-----|------------|------------|
| 1   | 2          | 5          | -          | 5   | 1          | 5          |
| 6   | 7          | 4.75       | -          | 7   | 1          | 20         |
| 8   | 1          | 7          | -          | 9   | 1          | 5          |

### TABLE 10. Comparison result of miniaturized multiband BPF.

| Parameters | $f_{CA}$ | $f_{CC}$ | $f_{CB}$ |
|------------|----------|----------|----------|
| $f_c$ (GHz) | 1.75     | 1.75     | 2.55     |
| FBW (%)    | 22.84    | 21.20    | 5.49     |
| $|S_{21}|$ dB | -0.61    | -0.55    | -0.95    |
| $|S_{11}|$ dB | -19.21   | -18.22   | -12.77   |
| Trans. zeros | -12.77   | -12.29   | -23.49   |
| (Isolation | -23.47   | -23.78   | -23.49   |
| interband) | -12.77   | -12.29   | -23.49   |

*note: for our design $f_{CB} < f_{CC} < f_{CA}$, ISO$_{1} = \text{isolation between } f_{CB}\text{ and } f_{CC}$, ISO$_{2} = \text{isolation between } f_{CA}\text{ and } f_{CC}$
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FIGURE 28. Simulated, measured, and equivalent circuit model results (a) $|S_{21}|$, (b) $|S_{11}|$, (c) bandwidth $f_{CB}$, $f_{CC}$, and $f_{CA}$ at $|S_{21}| = -3$ dB.

Fig 25(a), 25(b) and 25(c) show the surface current flows at $f_{CB}$, $f_{CC}$, and $f_{CA}$, respectively. At the $f_{CB}$ passband, the current distribution has flowed at the center resonator of $R_B$. The surface current distribution at the $f_{CC}$ band has flowed at the SIR-structure. The $f_{CA}$ band has flowed at the lower part of the feeding structure. This result indicates that the resonator is working independently and separately. The simulated and measured multiband BPF results are shown in Table 7.

D. MINIATURIZATION OF INDEPENDENT MULTIBAND BPF STRUCTURE

This section discusses the proposed miniaturization strategy. To reduce the multiband BPF size, a folded structure was proposed, as shown in Fig 26(a) and 26(b). Furthermore, Fig 27 shows the coupling matrix of miniaturized independent multiband BPF and the value of the coupling matrix is shown in Table 8. The value coupling matrix was obtained using optimization technique [12]. The nonlinear least-squares and the minimum-maximum algorithm were used.

The dimensions of the compact multiband BPF after optimization using the ADS software are shown in Table 9.

The fabricated results are shown in Fig 26(b). After applying the folding structure, the multiband BPF size became $0.32 \lambda_G \times 0.12 \lambda_G$, where $\lambda_G$ is the wavelength of the fundamental frequency. This method successfully reduced the multiband BPF dimensions by over 61.29%. Fig 28 (a), 28(b), and 28(c) show the simulated, measured, and coupling matrix model results for $|S_{21}|$, $|S_{11}|$, and $|S_{21}|$ for a bandwidth $f_{CB}$, $f_{CC}$, and $f_{CA}$ of $-3$ dB, respectively. The results of the simulated and measured multiband BPFs are compared in Table 10.

Fig 29(a), 29(b) and 29(c) show the surface current flow at $f_{CB}$, $f_{CC}$, and $f_{CA}$, respectively. At the $f_{CB}$ passband, the current distribution has flowed at the center resonator of $R_B$. The surface current distribution at the $f_{CC}$ band has flowed at the SIR-structure. The $f_{CA}$ band has flowed at the lower part of the feeding structure. This result indicates that the resonator is working independently and separately. Finally, Table 11 illustrates the performance comparisons of some triple-band BPFs in terms of transmission zeros, isolation, size, and independent bands.

Based on these parameters, the proposed bandpass filter has the following advantages: 1) the multiband BPF has a highly independent inter passband, 2) the insertion loss $|S_{21}|$ (dB) value has excellent performance, 3) the isolation inter passband is good, 4) the proposed multiband BPF has a higher number of transmission zeros, 4) the proposed multiband BPF has a higher number of transmission zeros, 5) the filter has a compact size, and 6) the proposed BPF structure can be applied for 5G communication at 3.5 GHz. Furthermore, by using the folded structure, the multiband BPF has a

| Ref | Freq (GHz) | $|S_{21}|$ (dB) | $|S_{11}|$ (dB) | $|S_{21}|$ at $f_{CB}$, $f_{CC}$, and $f_{CA}$ of $-3$ dB | $T_s$ (ns) | Independent passband ($f/f_i$) | Miniaturization |
|-----|------------|----------------|----------------|---------------------------------|----------|----------------|----------------|
| [30] | 1.95/3.46/5.25 | $-1.10/-1.50/-1.50$ | $9.70/6.40/9.00$ | $S_{21}$ | 3 | no/yes/no | no |
| [13] | 1.57/2.40/3.45 | $-1.60/-1.50/-2.30$ | $5.20/4.60$ | $S_{11}$ | 5 | no/no/yes | no |
| [16] | 1.80/3.50/5.20 | $-1.60/-1.50/-2.30$ | $11.00/6.20/6.10$ | $S_{21}$ | 3 | no/yes/no | no |
| [22] | 2.40/3.50 | $-1.07/-1.05$ | $15.60/5.20$ | $S_{11}$ | 4 | yes/yes | no |
| [23] | 2.60 | $-1.08$ | $2.15$ | $S_{21}$ | 3 | 0.028 | yes/yes |
| [26] | 1.57/3.50/5.20 | $-1.20/-0.90/-1.60$ | $12.70/10.00/3.08$ | $S_{11}$ | 6 | no/yes/no | no |
| [27] | 2.41/3.65/5.29 | $-1.91/-1.42/-1.51$ | $6.20/12.2/11.8$ | $S_{21}$ | 5 | yes/no/no | no |
| [28] | 1.59/3.12/4.02 | $-1.11/-1.12/-1.11$ | $15.70/12.70/5.71$ | $S_{11}$ | 5 | no/yes/no | no |
| Our | 1.65/2.55/3.25 | $-0.39/-0.72/-0.80$ | $15.90/11.40/9.10$ | $S_{21}$ | 5 | yes/yes/yes | yes |
| Our | 1.75/2.55/3.55 | $-0.55/-1.36/-1.37$ | $21.20/4.70/8.40$ | $S_{21}$ | 6 | yes/yes/yes | yes |

TABLE 11. Performance comparison with other reported BPFs.
This paper describes the successful design and implementation of a highly independent and compact multiband BPF based on a multicoupled line stub-SIR with a folding structure. The multiband BPF is constructed with a multicoupled line to generate a highly independent inter passband. Moreover, the multiband performance is produced separately and independently by using three sets of resonators. To miniaturize the multiband BPF, a folding structure is also proposed. As a result, the proposed multiband BPF is very compact and reduces the size by over 61.29%. This BPF is designed for applications such as GPS at 1.57 GHz, WCDMA (3G) at 1.8 GHz, WLAN (WiFi) at 2.4 GHz, LTE (4G) at 2.6 GHz, and 5G communication at 3.5 GHz. The simulated and measured results of the multiband BPF show good agreement.

Finally, our proposed multiband BPF structure has a highly independent inter passband response and a compact size.

**REFERENCES**

[1] Y. Xie, F.-C. Chen, and Z. Li, “Design of dual-band bandpass filter with high isolation and wide stopband,” *IEEE Access*, vol. 5, pp. 25602–25608, 2017, doi: 10.1109/ACCESS.2017.2773502.

[2] M. AbdulRehman and S. Khalid, “Design of tri-band bandpass filter using symmetrical open stub loaded step impedance resonator,” *Electron. Lett.*, vol. 54, no. 19, pp. 1126–1128, Sep. 2018, doi: 10.1049/el.2018.5240.

[3] B. Liu and Y. Zhao, “COMPACT tri-band bandpass filter for WLAN and WIMAX using tri-section stepped-impedance resonators,” *Prog. Electromagn. Res. Lett.*, vol. 45, pp. 39–44, Feb. 2014.

[4] B. Halvarsson, A. Simonsson, A. Elgcrona, R. Chana, P. Machado, and H. Asplund, “5G NR testbed 3.5 GHz coverage results,” in *Proc. IEEE 87th Veh. Technol. Conf. (VTC Spring)*, Jun. 2018, pp. 1–5, doi: 10.1109/VTCSpring.2018.8417704.

[5] T. Firmansyah, S. Praptodinyo, R. Wiryadinata, S. Suhendar, S. Wardoyo, A. Alimuddin, C. Chairunissa, M. Alaydrus, and G. Wibisono, “Dual-wideband band pass filter using folded cross-stub stepped impedance resonator,” *Microw. Opt. Technol. Lett.*, vol. 59, no. 11, pp. 2929–2934, Nov. 2017, doi: 10.1002/mop.30848.

[6] T. Firmansyah, S. Praptodinyo, A. S. Pramudyo, C. Chairunissa, and M. Alaydrus, “Hepta-band bandpass filter based on folded cross-loaded stepped impedance resonator,” *Electron. Lett.*, vol. 53, no. 16, pp. 1119–1121, Aug. 2017, doi: 10.1049/el.2017.1121.

[7] Q.-X. Chu, F.-C. Chen, Z.-H. Tu, and H. Wang, “A novel crossed resonator and its applications to bandpass filters,” *IEEE Trans. Microw. Theory Techn.*, vol. 57, no. 7, pp. 1753–1759, Jul. 2009.

[8] F.-C. Chen and Q.-X. Chu, “Design of compact tri-band bandpass filters using assembled resonators,” *IEEE Trans. Microw. Theory Techn.*, vol. 57, no. 1, pp. 165–171, Jan. 2009.

[9] V. K. Killamsetty and B. Mukherjee, “Miniaturised highly selective bandpass filter with very wide stopband using meander coupled lines,” *Electron. Lett.*, vol. 53, no. 13, pp. 889–890, Jun. 2017, doi: 10.1049/el.2017.1270.

[10] J. Xu, H. Wan, J. Ding, and Y. Zhu, “Miniaturised tri-band lowpass–bandpass filter using lumped-element structure,” *Electron. Lett.*, vol. 55, no. 5, pp. 272–274, 2019, doi: 10.1049/el.2018.8054.

[11] X. Guo, L. Zhu, and W. Wu, “Design method for multiband filters with compact configuration in substrate integrated waveguide,” *IEEE Trans. Microw. Theory Techn.*, vol. 66, no. 6, pp. 3011–3018, Jun. 2018, doi: 10.1109/TMTT.2018.2830337.

[12] K. Zhou, C.-X. Zhou, H.-W. Xie, and W. Wu, “Synthesis design of SIW multiband bandpass filters based on dual-mode resonances and split-type Dual- and triple-band responses,” *IEEE Trans. Microw. Theory Techn.*, vol. 67, no. 1, pp. 151–161, Jan. 2019, doi: 10.1109/TMTT.2018.2874250.

[13] T. Unno and N. Sekiya, “Compact high-pole HTS triband bandpass filters using a new feeding structure,” *IEEE Trans. Appl. Supercond.*, vol. 28, no. 4, Jun. 2018, Art. no. 1500305, doi: 10.1109/TASC.2018.2797980.

[14] W.-Y. Chen, M.-H. Weng, and S.-J. Chang, “A new tri-band bandpass filter based on stub-loaded step impedance resonator,” *IEEE Microw. Wireless Compon. Lett.*, vol. 22, no. 4, pp. 179–181, Apr. 2012, doi: 10.1109/LMWC.2012.2187884.

[15] J. Xu, W. Wu, S. Member, and C. Miao, “Compact microstrip dual/tri-/quad-band bandpass filter using open stubs loaded shorted stepped-impedance resonator,” *IEEE Trans. Microw. Theory Techn.*, vol. 61, no. 9, pp. 3187–3199, Sep. 2013, doi: 10.1109/TMTT.2013.2273759.

[16] C.-H. Lee, C.-I.-G. Hsu, and H.-K. Jhuang, “Design of a new tri-band microstrip BPF using combined quarter-wavelength SIRs,” *IEEE Microw. Wireless Compon. Lett.*, vol. 16, no. 11, pp. 594–596, Nov. 2006.
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