Constant absolute bandwidth tunable asymmetric order dual-band BPF with reconfigurable bandwidth using mode control technique

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Abstract

Herein, a dual-band bandpass filter with bandwidth tuning and constant bandwidth centre frequency tuning is presented, where the first and second bands are, respectively, of two and three modes. Based on λ/2 resonator modal analysis, a varactor-ended stub is centrally loaded to a varactor-terminated resonator to control even modes, and two varactor-ended stubs are also loaded off-centred symmetrically to tune, mostly, odd modes. Using these varactors, the first passband is formed by pairing the first even and odd modes while the second passband using the next pair of even and odd modes. To make the second band bandwidth also constant, the first mode of another resonator kept parallel to the first resonator has been used making the second passband of three poles. S-matrix has been derived to analyse and coupling matrices have been synthesised to design the filter. A filter has been fabricated where the first band is tunable from 0.9 to 1.24 GHz (31.7%) with absolute bandwidth of 220 MHz while the second passband is tunable from 1.8 to 2.45 GHz (30.5%) with absolute bandwidth of 280 MHz. The first and second bands’ bandwidth can be tuned from 110 to 360 MHz and from 230 to 700 MHz.

1 | INTRODUCTION

RF tunable filters are still a very active topic of research to fulfil the requirement of multiband frequency selection in the modern communication era. Among all the topologies, direct taping resonator using varactor diode is commonly used due to their ease of integration and versatile characteristics.

In the past, some attempts have been made to design dual-passband filters [1–5], but all these research works focused on making either both passbands fixed or tunable the second passband only, keeping the first passband of the filters fixed. In Ref. [6], for the first time, tunable dual-band bandpass filter (BPF) is presented in which both the passbands can be tuned independently. But its circuit size is large, and also it requires a large number of biasing circuits. Two tunable passbands have also been demonstrated in Ref. [7] by controlling odd and even modes of the resonators using varactors. A lumped-element dual-resonance resonator is introduced in Ref. [8] to design a dual-band BPF. In Ref. [9], quasi-elliptic dual-band, BPF is presented, in which a good selectivity has been achieved due to the presence of transmission zeros. Substrate integrated waveguide (SIW)-based switchable dual BPF, which is continuously tunable in S and X bands, is designed in Ref. [10] using piezoelectric actuators for tuning operation. More recently, an independently controllable dual-band is presented in Ref. [11], but the passbands are of the first order only. Two stub-loaded stepped-impedance resonators have been used to design a dual-band tunable BPF [12]. However, this filter has a small tuning of the centre frequency of both the passbands. Though in all of these works, the bandwidth of the filter is not constant. A few works have also been reported to achieve constant absolute bandwidth (CABW) in both or either passbands while tuning the centre frequencies. For example, open-loop resonator-based tunable dual-band BPF is reported in Ref. [13]. But only the second passband has CABW. In Ref. [14], synchronous dual-mode resonators are proposed to design dual-band BPF in which the absolute bandwidth of the first passband is constant while the second passband is tunable with constant fractional bandwidth (CBFW). The tunable dual BPF in which both passbands are tunable with constant absolute bandwidth are...
presented in Refs. [15–17]. Two or more resonators with different topologies are used to obtain the dual-band tunable filter with CABW. But in all these works, the tunable bandwidth at a fixed frequency is neither discussed nor achieved.

In Ref. [18], two dual-mode resonators are proposed to achieve independent tunability in centre frequency and bandwidth. But the bandwidth changes while tuning the centre frequency. Moreover, the achieved change in fractional bandwidth (FBW) is minimal. An RF MEMS-based dual-band BPF has been presented to achieve reconfigurable bandwidth in Ref. [19]. Although the achieved reconfigurable bandwidth is very high, the centre frequency is not tunable. Very few works have been reported on more than two modes of tunable dual-band BPF. For example, a quadruple-mode resonator has been analysed [20], but the filter is not tunable. In Ref. [21], waveguide-based dual-mode tunable dual-band filters are presented. In Refs. [14] and [18], dual-mode resonators have been used to achieve tunable dual-band BPF. But in all of this multi-mode dual-band BPFs, both passbands are of order two.

As per the authors’ information, multi-mode-based asymmetric-order tunable dual-band BPF with bandwidth control is not yet reported. An asymmetric-order dual-band filter is presented where the first passband is of order two and the second passband of order three. Both passbands are tunable with CABW, and also bandwidth (BW) of both the passbands can be varied at any fixed centre frequency.

2 | THE PROPOSED FILTER

Figure 1 shows the layout of the proposed tunable dual-band BPF with variable BW. The layout consists of two paths—BB’ and GG’—from input A to output A’ port. The first path, BB’, is itself a tunable dual-band BPF, which consists of a resonator, BB’, loaded with two off-centred symmetrically located equal length stubs (CE, C’E’) and one stub, DF, at the centre. All the three stubs have also been terminated by varactors, C3 and C5. In addition, variable capacitor C4 at the point B (or B’) can be written as:

\[
T_1 = \begin{bmatrix} 1 & -j/2\pi f C_1 \\ 0 & 1 \end{bmatrix}
\]

(1)

\[
T_2 = \begin{bmatrix} 1 & 0 \\ j2\pi f C_4 & 1 \end{bmatrix}
\]

(2)

where \( f \) is the frequency of interest. The \( ABCD \) matrix for the transmission line section, BC (or B’C’) can be written as:

\[
T_3 = \begin{bmatrix} \cos \theta_1 & j\sin \theta_1/Y_o \\ jY_o \sin \theta_1 & \cos \theta_1 \end{bmatrix}
\]

(3)
where $\theta_1$ and $Y_o$ are the electrical length and characteristic admittance of section, $BC$ (or $B'C'$), respectively. To obtain the $ABCD$ matrix of loaded stub at the point $C$ (or $C'$) with capacitor $C_3$, the input admittance, $Y_{inCE}$ at the point $C$ (or $C'$) for this section is required, and it can be written as:

$$Y_{inCE} = Y_o \frac{j2\pi f C_3 + jY_o \tan \theta_3}{Y_o - 2\pi f C_3 \tan \theta_3}$$

where $\theta_3$ is the electrical length of section $CE$ (or $C'E'$). The input admittance, $Y_{inCE}$ can be converted into $ABCD$ matrix as:

$$T_4 = \begin{bmatrix} 1 & 0 \\ Y_{inCE} & 1 \end{bmatrix}$$

Similarly, the input admittance of stub loaded at a point, $D$ with capacitor $C_2$ can be written as:

$$Y_{inDF} = Y_o \frac{j2\pi f C_2 + jY_o \tan \theta_2}{Y_o - 2\pi f C_2 \tan \theta_2}$$

Now, this input admittance is converted into $ABCD$ matrix, $T_5$ as:

$$T_5 = \begin{bmatrix} 1 & 0 \\ Y_{inDF} & 1 \end{bmatrix}$$

The $ABCD$ matrix, $T_6$ of transmission line section, $CD$ (or $C'D'$) can be written as:

$$T_6 = \begin{bmatrix} \cos \theta_4 & jY_o \sin \theta_4 / \cos \theta_4 \\ jY_o \sin \theta_4 / \cos \theta_4 & \cos \theta_4 \end{bmatrix}$$

where $\theta_2$ and $\theta_4$ are the electrical lengths of section, $DF$ and $CD$ (or $C'D'$). So, the $ABCD$ matrix, $T$ of the first path filter $BB'$ can be obtained by multiplying the $ABCD$ matrices of each cascaded section and can be written as:

$$T = T_1T_2T_3T_4T_5T_6T_7T_8T_9T_10T_11T_{12} = \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix}$$

Now, the above $ABCD$ matrix, $T$, is converted into $S$-matrix, and the $S_{21}$ can be written as:

$$S_{21} = \frac{2Y_o}{T_{11}Y_o + T_{12}Y_o + T_{21}Y_o + T_{22}Y_o}$$

The magnitude of calculated $S_{21}$ is plotted in Figure 3 for the loosely coupled ($C_1 = 0.01\, \text{pF}$) feed line. This plot shows the mode frequencies and transmission zeros generated by the filter $BB'$ with zero values of capacitances $C_2$, $C_3$ and $C_4$. The plot of variation of the resonating modes with the varactors shown in various figures (e.g. Figures 8, 10–15) has been done using (10). Positions of the mode frequencies can be calculated, and their odd/even nature can be determined by carrying out the odd- and even-mode analysis presented in the next section. The RT Duriod 6010.2LM substrate with dielectric constant $(\epsilon_r) = 10.2$, and thickness $(b)$ of 0.635 mm has been used for all theoretical and practical implementation.

### 3.2. Even- and odd-mode analysis of the $BB'$ path filter

Even- and odd-mode analysis is used to calculate the mode frequencies of the $BB'$ part of the proposed BPF. Figure 4a shows the even mode equivalent circuit obtained by centrally open circuiting the filter $BB'$ shown in Figure 2. The even mode admittance (for Figure 4(a)) can be written as:

$$Y_{even} = Y_{Beven} + j\omega_o C_1$$

where $Y_{Beven}$ is the admittance looking from point $B$ towards point $C$, and can be written as:

$$Y_{Beven} = Y_o \frac{Y_{even} + jY_o \tan \theta_1}{Y_o + jY_{even} \tan \theta_1}$$

where $Y_o$ is the characteristic admittance of transmission line section, $BC$ (or $B'C'$), $CE$ (or $C'E'$) and $CD$ (or $C'D'$). However, $Y_{even}$ is the admittance at point $C$ for the parallel section $CE$ and $CDF$ along with their loading capacitor $C_3$ and $C_2$, respectively, and can written as:

$$Y_{even} = Y_o \frac{j\omega_o C_3 + jY_o \tan \theta_3}{Y_o + jY_{Deven} \tan \theta_4}$$

Now the admittance $Y_{Deven}$ of section $DF$ loaded with capacitor $C_2$ can be written as:

![Figure 3](image-url) Mode frequencies of the $BB'$ filter with zero values of capacitances $C_2$, $C_3$ and $C_4$.
\[ Y_{\text{even}} = Y' \left( j \omega_{oe} C_2 + j 2 Y' \tan \theta_2 \right) \]

where \( Y' \) is the characteristic admittance of transmission line section \( DF \) shown in Figure 4(a). Width of this line is half of \( DF \) line in Figure 2 that is \( \omega_1/2 \). The even mode frequency, \( \omega_{oe} \) can be calculated by equating the input admittance, \( Y_{\text{even}} \) of (11) to zero.

Similarly, the odd mode equivalent circuit shown in Figure 4(b) can be obtained by centrally short-circuiting the filter \( BB' \) shown in Figure 2. The odd mode input admittance can be written as:

\[ Y_{\text{odd}} = Y_{\text{Bodd}} + j \omega_{oo} C_4 \]

where \( Y_{\text{Bodd}} \) is the admittance looking from point \( B \) towards point \( C \), and can be written as:

\[ Y_{\text{Bodd}} = \frac{Y_0}{Y_0 + jY_{\text{Codd}} \tan \theta_1} \]

where the admittance \( Y_{\text{Codd}} \) can be calculated for the section \( CE \) and \( CD \) at point \( C \) as:

\[ Y_{\text{Codd}} = \frac{j \omega_{oo} C_3 + j Y_0 \tan \theta_3}{Y_0 - \omega_{oo} C_3 \tan \theta_3} - j Y_0 \cot \theta_4 \]

where \( \theta_1, \theta_2, \theta_3 \) and \( \theta_4 \) are the electrical lengths of corresponding section \( BC, DF, CE \) and \( CD \). Now, odd-mode resonating frequency, \( \omega_{oo} \) can be calculated by equating the \( Y_{\text{odd}} \) of (15) to zero.

### 3.3 Qualitative modal analysis

The passband of a filter can be formed by bringing two or more resonating modes of a single resonator sufficiently close to each other, and controlling of the modes leads to the tuning of the centre frequency and also the bandwidth of passband. Modes of a resonator can be controlled using variable capacitors or stubs at suitable locations on the resonator. Stubs and varactors, in general, change the wavelength of the voltage and current wave of various modes resulting in a change in the resonance frequency of the corresponding mode. Voltage waveforms for the first few modes on an open-ended \( \lambda/2 \) resonator (of length 15 mm) are shown in Figure 5. All the modes, either odd or even, have the maximum voltage amplitude at the resonator ends (\( B \) and \( B' \)). But at the centre (\( D \)), only even modes have maximum amplitude, and odd modes have zero amplitude. To achieve more and more tunability of the resonance frequency of a mode, a stub or a varactor should be connected at the location where the modal voltage is as large as possible. For example, as explained in subsequent sections, a varactor at the open ends of the resonator allows tuning of all the modes, while a stub/varactor at
the resonator centre has no effect on the resonance frequencies of odd modes.

### 3.4 Effect of the stubs and capacitors on mode frequencies

#### 3.4.1 Role of capacitor $C_4$ loaded at $\lambda/2$ resonator ends

Since all the even and odd modes have maximum voltage at the resonator open ends ($B$ and $B'$ in Figure 5), creating a finite impedance path between the ends and ground leads to maximum change in the modal waveform and hence in the resonance frequencies also. Detailed analysis of voltage and current waveforms is required to understand the effect of $C_4$, and for simplicity, the first odd and even mode waveforms have been chosen. Figures 6 and 7 show the effect of $C_4$ on voltage and current waveforms of the first odd and even mode, respectively. Initially, for $C_4 = 0 \, \text{pF}$, an infinite impedance exists between the resonator ends and ground, therefore, there is zero current and maximum voltage at $B$ and $B'$ shown by $Z_1Z_1'$ and $P_1P_1'$ curves in Figures 6 and 7. Note (in Figures 6 and 7) the fundamental wavelength (occurring for $C_4 = 0 \, \text{pF}$) of the first odd and first even modes is $\lambda_o = 2L$ and $\lambda_e = L$, respectively, and their amplitudes at both ends of the resonator are marked by $P_1P_1'$ and $Z_1Z_1'$. The resonance frequencies of these two modes are shown by $f_{e1}$ and $f_{o1}$ peaks in Figure 8.

When $C_4$ is increased to a small finite value, a finite impedance comes into existence between end $B$ (or $B'$) and ground for all the modes, leading to a finite current and reduction in the voltage at $B$ (or $B'$) as shown by the continuous portion of $Z_1Z_1'$ and $P_1P_1'$ curves in Figures 6 and 7. As can be clearly seen from Figures 6 and 7, now voltage and current of slightly longer wavelength satisfy the new boundary conditions implying that the resonance frequency of the modes must decrease (see Figure 8). As the varactor $C_4$ value is increased further, the current increases and voltage decreases at $B$ (or $B'$), requiring that the wavelength of current and voltage waveforms increases further to satisfy the modified boundary condition. This, in turn, decreases the resonance frequencies of the two modes. Finally, when $C_4 \to \infty$, voltage for both odd and even modes become zero at the resonator ends and current maximum. Since the first odd mode has zero voltage at the centre, so the voltage of this mode becomes zero over the entire resonator and current constant, equivalently making its wavelength infinite and frequency zero. The wavelength of the even mode also increases but from $\lambda_e = L$ to $2L$, as shown in Figure 7 and hence its resonance frequency decreases to the location of the first odd mode for $C_4 = 0 \, \text{pF}$ case, $f_{o1}$. The reduction of the resonance frequency with the increment in $C_4$ occurs not only for the first odd and even modes but for all the modes as their voltage is non-zero at the resonator ends. In fact, when $C_4$ is increased to infinity, the resonance frequency of all the modes reduces to that of their immediate lower mode (Figure 8).

#### 3.4.2 Role of centre stub $DF$ and $C_2$

From the previous section, it is observed that loading the resonator at its centre by a varactor or stub will affect only the even modes as all the odd modes have zero voltage at the centre. Figure 9(a) and 9(b) show the effect of a varactor or a small stub, which is equivalent to a capacitor, on the first even mode. As soon as $C_2$ value (or the stub length $L_2$) is increased to a small finite value, a finite current starts to flow between the resonator centre and ground, making the current waveform discontinuous ($Z_2Z_2'$) and reducing the voltage (to $P_2$). The new values of the voltage and current at the centre can be...
satisfied using sinusoid of longer wavelength resulting in the decrement in the resonance frequency, as shown in Figure 10 (due to $C_2$) and in Figure 11 (due to stub $DF$). As $C_2$ (or $l_2$) value is further increased, the wavelength of the mode increases even more and hence bringing down the resonance frequency (Figures 10 and 11). Finally for $C_2 \rightarrow \infty$ (or stub of length $\lambda_o/4$), the centre point $D$ gets grounded making voltage zero (curve $P_3$) and current maximum ($Z_2 Z_2'$). For this case the wavelength of the voltage and current has increased to $\lambda_c = 2L$ (from $\lambda_c = L$ for $C = 0$ pF case), so the resonance frequency reduces to that of the first odd mode. This is the maximum change in the frequency possible using $C_2$.

The stub $DF$, however, can further bring down the frequency as its length can be increased even more. As the length of the stub is increased beyond $\lambda_o/4$, its impedance becomes inductive and increases with length; voltage starts increasing but with positive value ($P_4$, $P_3$), and current starts decreasing ($Z_3$, $Z_4$). The wavelength, however, must continue to increase to satisfy the new impedance at centre $D$. For sufficiently long stub ($l_2 = \lambda_o/2$), its impedance is infinite and hence zero current and finite voltage but with infinite wavelength and zero resonance frequency. In general, the maximum reduction in the resonance frequency of an even mode that $C_2$ can introduce is from current frequency to that of the odd mode immediately on its lower side.

The stub, on the other hand, not only brings down the frequency even more; it creates transmission zeros (TZs) also, as shown in Figure 11. These transmission zeros ($f_{TZ}$) can be calculated by equating the input impedance of stub ($DF$) connected at $D$ to zero as:

$$f_{TZ} = \frac{(2k + 1)c}{4l_2 \sqrt{\varepsilon_r}} \quad k = 0, 1, 2, \ldots$$  \hspace{1cm} (18)

where $c$ is the speed of light, and $\varepsilon_r = 7.16$ is the effective dielectric constant of the substrate (RT Duroid 6010.2LM). For integer, $k = 0$ and stub length, $l_2 = 2$ mm, transmission zero can be calculated using the above equation and is found to be at 14 GHz, which is not too far from $T_{Z1}$ obtained from the simulation in Figure 12. Further increment of $l_2$ to 5 mm generates two TZs at 5.6 and 16.8 GHz for $k = 0$ and $k = 1$, respectively. In the simulated result in Figure 12, the first TZ ($T_{Z1}$) matches well with the calculated one; however, the second TZ ($T_{Z2}$ in Figure 12) is slightly off from the calculated value. The stub $DF$ terminated by varactor $C_2$ has been used in the proposed filter (Figure 2) to allow the tuning of the modes even after the filter is fabricated. The new locations of TZs, which also get tuned along with the modes, are given by:

$$f_{TZ} = \frac{1}{2\pi} \sqrt[4]{\frac{-3c^2C_2Z_0 + \sqrt{3c^6C_2^2Z_0^2(4\varepsilon_r l_2 + 3cC_2Z_0)}}{2Z_0C_2l_2^2 \varepsilon_r}}$$  \hspace{1cm} (19)
3.4.3 Effect of $C_3$-loaded symmetrical stubs $CE$ and $C'E'$

The stubs $CE$ and $C'E'$ loaded off-centre on the resonator $BB'$ affect the odd and even modes simultaneously, in the same way, as centre-loaded stub affects only even modes, namely mode wavelengths increase leading to reduction of the mode frequencies. Here, as both even and odd modes have non-zero values at $C$ (or $C'$), the increment of stub length $l_3$ increases the wavelength of current and voltage waveform of all modes, and hence the frequencies of all modes decrease. However, the modes with the larger voltage at the location of stubs $CE$ and $C'E'$ are affected more. Apart from changing the mode frequencies, stubs also generate extra transmission zeros. The transmission zeros frequency $f_{TZ}$ can be calculated using (19). For example, values at frequencies, stubs also generate extra transmission zeros. The more suitable modes using capacitors $C_2$, $C_3$ and $C_4$ to form both passbands quite well.

For $l_2 = l_4 = 5$ mm, Figure 13 shows $f_{e1}$ and $f_{o1}$ can be used to form the first passband, whereas $f_{e2}$ and $f_{o2}$ are used to form the second passband. However, $f_{o1}$ and $f_{e1}$ are not close enough to each other to form well developed passband. Hence, the capacitors $C_2$, $C_3$ and $C_4$ are used to control the specific modes to form both passbands quite well.

For $C_2 = C_3 = C_4 = 0$ pF and $C_1 = 0.01$ pF, the first four modes—two odd modes, $f_{e1}$ and $f_{o1}$ and two even modes, $f_{e2}$ and $f_{o2}$—are shown in Figure 14. When the value of $C_2$ increases from 0 to 1 pF, even modes, $f_{e1}$ and $f_{o2}$ shift to lower value at $f_{o1}$ and $f_{e2}$, respectively. First even mode $f_{e1}$ reaches near to $f_{o1}$ and transmission zero $T_{Z1}$ shifts to $T_{Z11}$. Second even mode $f_{e2}$ shifts away from $f_{o2}$ to lower value. The transmission zeros generated in Figure 14 can be calculated using (19). For example, for $C_2 = 1$ pF and stub (DF) length, $l_2 = 5$ mm, (19) predicts a transmission zero at 3.38 GHz which is there at 3.4 GHz in the simulation result shown in Figure 14. The variable capacitor $C_2$ cannot tune the odd modes which has already been explained.

If $C_2$ further increases to 5 pF, $f_{e1}$ and $f_{o1}$ shift to $f_{e12}$ and $f_{o12}$, respectively. But $f_{e12}$ value is lower than $f_{e1}$, and the transmission zero also shifts from higher to the lower side of the first passband. As we further increase the value of $C_2$ (taking up to 10 pF), first even-mode frequency continuously shifts to lower-frequency value. As shown in Figure 14, as soon as the first even-mode frequency shifts to a frequency lower than the first odd mode, the tuning of the second even-mode frequency almost stops. So, the capacitor, $C_2$, is used to tune the centre frequency as well as the bandwidth of the first passband.

For $C_2 = 10$ pF, $C_3 = C_4 = 0$ pF and $C_1 = 0.01$ pF, four required resonating mode frequencies, two odd mode and two even modes, are shown in Figure 15. One transmission zero, $T_{Z1}$ and $T_{Z2}$, is generated on the lower side of each of the first and second passbands. As $C_1$ and $C_4$ values increase to 0.3 and 5 pF, respectively, modes, $f_{e1}$ and $f_{o1}$ of first pass band finally shift to $f_{e12}$ and $f_{o12}$, respectively. Similarly modes, $f_{e2}$ and $f_{o2}$ of second pass band finally shift to $f_{e22}$ and $f_{o22}$, respectively. As shown in Figure 15, the modes, $f_{e12}$ and $f_{o12}$ are close enough to form first passband with higher side transmission zero, $T_{Z1}$. Similarly, modes, $f_{e22}$ and $f_{o22}$ are also close enough to form second passband with higher side transmission zero, $T_{Z2}$. So by placing varactor terminated stubs at specific positions on the resonator $BB'$, the varactors can be used to form the dual band filter by bringing two modes sufficiently close to each other with properties such as centre frequency and bandwidth tuning.
3.4.5 | Effect of capacitor $C_1$ on bandpass response of filter

The capacitors, $C_2$, $C_3$ and $C_4$ are used to tune the centre frequency and bandwidth of the first and second passbands as discussed earlier. Whenever the return loss and insertion loss of passband degrade during the tuning of centre frequency and bandwidth, it can be enhanced by changing the capacitor, $C_1$. As shown in Figure 16, for $C_2 = 10 \text{ pF}$, $C_3 = 0.3 \text{ pF}$, $C_4 = 5.5 \text{ pF}$ and $C_1 = 0.01 \text{ pF}$, loosely coupled first and second passbands with two poles each and a transmission zero to the right of each passband are present. As $C_1$ increases to $0.11 \text{ pF}$, the coupling between feedline and resonator enhances, which yields improved passbands, as shown in Figure 16. Further increment of $C_1$ from $0.11$ to $3 \text{ pF}$ gives more improved passbands with good return loss and insertion loss.

3.5 | Theoretical and simulated responses

3.5.1 | Centre frequency tunability

The simulated and theoretically calculated $S_{21}$ and $S_{11}$ of the first path BPF (Figure 2) is shown in Figure 17. The first passband can be tuned from $0.82$ to $1.17 \text{ GHz}$ with an absolute constant bandwidth of $250 \text{ MHz}$ by changing the capacitor values as indicated in the plot. As the S-matrix of Figure 2 structure is simulated (or theoretically calculated) under ideal conditions; therefore, the obtained insertion loss is low, and return loss is high. The second passband tunes from $2.05$ to $3.14 \text{ GHz}$ with absolute bandwidth varying from $80$ to $180 \text{ MHz}$. Each passband consists of a transmission zero to their respective higher side.

3.5.2 | Bandwidth tuning

The bandwidth tuning of the first passband at a centre frequency of $0.95 \text{ GHz}$ for fixed centre frequency ($f_c = 2.50 \text{ GHz}$) and absolute bandwidth (ABW = $110 \text{ MHz}$) of the second passband is shown in Figure 18. The bandwidth of the first passband is tuned from $130$ to $360 \text{ MHz}$ by changing the various capacitor values, as written in the plot. As plotted responses are based on simulation and theoretical calculation, the insertion loss is very low with good return loss. The second passband consists the resonating modes, $f_{e2}$ and $f_{o2}$. Since these modes are controlled by the variable capacitors, $C_3$ and $C_4$, so the bandwidth of the second passband is unchanged.

4 | CABW DUAL-BAND TUNABLE BPF WITH RECONFIGURABLE BANDWIDTH

The previous sections presented the second-order dual-band BPF with simultaneously tunable centre frequencies of both the bands. However, as shown in Figure 17, the first passband has CABW throughout the centre frequency tuning range
whereas bandwidth of the second passband continuously increases as centre frequency is changed from low to high value. The second passband is created using a pair of modes—the second even mode $f_{o2}$ and second odd mode $f_{o3}$, as shown in Figures 19 and 20. When its centre frequency is changed by tuning the two modes using varactors $C_2$, $C_3$, and $C_4$, because of the different tuning rates of the two modes, it is not possible to maintain the separation between $f_{o2}$ and $f_{o3}$ constant leading to change in the bandwidth.

To ensure the CABW of second passband, an extra resonator, $GG'$, of length $l_3$ coupled with input/output feedline using capacitor $C_{11}$ has been used, as shown in Figure 1. The second passband now consists of the first odd mode, $f_{o1}$ of $GG'$ and the previous $f_{o2}$ and $f_{o3}$ making it now a three-pole band. The mode, $f_{o1}$ of $GG'$ can be tuned only using $C_{11}$, thus giving an extra degree of freedom for ensuring CABW of the second band also. The odd resonant frequency, $f_{o1}$ can be calculated as:

$$Y_{in} = j2\pi f_{o1}C_5 - jY_0\cot(\beta_{o1}l_3) = 0$$

(20)

where $Y_{in}$ and $\beta_{o1}$ are the input admittance of the resonator and propagation constant, respectively.

### 4.1 Effect of resonator $GG'$ on the second band

For $C_{11} = 0\, \text{pF}$, no signal passes through resonator $GG'$ hence the proposed filter works similar to the first path filter shown in Figure 2. The response of proposed filter for $l_3 = 10\, \text{mm}$, $C_{11} = C_5 = 0\, \text{pF}$, $C_1 = 0.11\, \text{pF}$, $C_2 = 10\, \text{pF}$, $C_3 = 0.3\, \text{pF}$ and $C_4 = 5.5\, \text{pF}$ is shown in Figure 19 which will be same as that of the first path filter for the above capacitor values. When $C_{11}$ increases from 0 to 0.02 pF, the signal starts passing through resonator $GG'$ as well. After exciting resonator $GG'$ (with $C_{11} = 0.02\, \text{pF}$ but $C_5 = 0\, \text{pF}$) in proposed filter, the previous modes, $f_{o1}$, $f_{e1}$, $f_{o2}$ and $f_{e2}$ do not shift, whereas $f_{o3}$ now coincides with the first odd mode, $f_{o1}$, of resonator $GG'$, as shown in Figure 19. The transmission zero $T_{Z2}$ shifts lower to $T_{Z21}$ while $T_{Z1}$ remains fixed.

### 4.2 Effect of capacitor $C_5$ on $GG'$ mode frequencies

Since $C_5$ is loaded at $GG'$ ends, its effect on the mode frequencies of $GG'$ would be similar to that of $C_4$ on resonator $BB'$ as explained in Section 3.4.1, namely it decreases the frequencies of all modes of $GG'$ as $C_5$ is increased. As shown by red curve in Figure 21, when $C_5$ is increased from 0 to 0.5 pF, the first mode of $GG'$, $f_{o1}$, shifts to a lower value, $f_{o11}$, and there is slight shift in transmission zero $T_{Z22}$ also. It is to be noted that modes $f_{o2}$ and $f_{e2}$ form the second band of the first path filter, in which frequency of mode $f_{o2}$ increases at a higher rate than that of $f_{e2}$ leading to an increase in the second passband bandwidth as its centre is tuned to a higher frequency. Once the first mode of $GG'$ comes sufficiently close to the pair $f_{o2}$ and $f_{e2}$, as is the case now, it becomes part of the second band of the proposed filter—making it a third-order
passband. The first band, however, continues to remain of the second-order and unaffected by $GG'$. Further increment of $C_5$ from 0.5 to 1 pF shifts the mode $f_{lo1}$ and $T_{Z21}$ from higher side of the second band to its lower side at $f_{lo2}$ and $T_{Z22}$ (the blue curve in Figure 21). In the proposed filter, the first mode of $GG'$ is kept on the lower side of $f_o$ (i.e. at $f_{lo2}$) so that bandwidth of the second band is now the separation between modes $f_{lo2}$ and $f_{e2}$ and it can be kept constant by tuning $f_{lo2}$ also at the same rate as $f_{e2}$.

4.3 | Effect of capacitor $C_2$ on first band of the proposed filter

There is a transmission zero, $T_{Z1}$, due to stub $DF$ and $C_2$, immediately on the higher side of $f_{e1}$ of the first band of proposed filter in Figure 22. To make $S_{21}$ attenuate faster on lower side of the first passband as well, $T_{Z1}$ along with $f_{e1}$ is moved to lower side by increasing $C_2$ from 10 to 20 pF. Values of other capacitors are $C_1 = 0.11$ pF, $C_{11} = 0.02$ pF, $C_3 = 0.3$ pF, $C_4 = 5.5$ pF and $C_5 = 1$ pF. Finally, the response shown in Figure 22 consists of two-pole first passband and three-pole second passband with transmission zero on lower side of each of passbands.

4.4 | External quality factors ($Q_{ex}$)

The external quality factors of odd ($Q_{exo1}/Q_{eso2}/Q_{exolo}$) and even ($Q_{exe1}/Q_{exe2}$) modes for the first and second passbands due to the varactors can be derived from the input reflection coefficient of odd and even modes as [23]:

$$Q_{exol}, o_2, lo_1 = \frac{\pi f_{o1}, o_2, lo_1}{\tau_{S_{1lo1}, o_2, lo_1}} \ldots (21)$$

$$Q_{exe1}, e_2 = \frac{\pi f_{e1}, e_2}{\tau_{S_{1lo1}, e_2}} \ldots (22)$$

where $\tau_{S_{1lo1}}, \tau_{S_{1lo1}}, \tau_{S_{1lo2}} \ldots \tau_{S_{1lo1}}$ are the group delays of $S_{1lo1}$ at their respective resonating frequencies, $f_{o1}, f_{e1}, f_{o2}, f_{e2}$ and $f_{lo1}$. Figure 23 shows the extracted external quality factors of odd and even modes for the first and second passbands due to the varactors loaded at various positions on the first path filter. The variation of the external quality factors with the tuning frequency is primarily needed to keep the bandwidth of the first and second bands unchanged. The observed variation of the external quality factor of the bands with various varactors in the first path filter is found to be in accordance with the fact

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**FIGURE 22** Variation of modes with capacitor, $C_2$

**FIGURE 23** Change in quality factors with capacitor. (a) $C_1$. (b) $C_2$. (c) $C_3$. (d) $C_4$. 

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that the external quality factor of a band is inversely related to the separation between the modes of the band [23].

As shown in Figure 14, even mode \( f_{e1} \) first comes closer to the odd mode \( f_{o1} \) and then moves away from \( f_{o1} \) as \( C_2 \) is increased. This implies that the external quality factor of the first band should first increase with \( C_2 \) and then decrease [23], and this is what is observed, as shown in Figure 23(b). Because the second mode frequencies, \( f_{e2} \) and \( f_{e3} \), do not get much affected by variation in \( C_2 \) when it is sufficiently large shown in Figure 20, so the external quality factors of the second band also do not vary much, as shown in Figure 23(b) [23]. Similarly in accordance with fact mentioned in above paragraph and in Ref. [23], as seen in Figure 23(c), \( Q_{ex_1} \) and \( Q_{ex_2} \) increase with capacitor \( C_3 \), whereas \( Q_{ex_2} \) and \( Q_{ex_2} \) increase initially and then decrease. The variation of the external quality factors of odd and even modes for the first and second passbands, which is shown in Figure 23(d), increases with capacitor \( C_4 \). As shown in Figure 23, the external quality factor, \( Q_{ex_0} \), remains unaffected due to varactors \( C_{1s}, C_{2s}, C_{3}, \) and \( C_4 \).

Figure 24(a) shows the external quality factor of the first and second passbands versus coupling capacitor \( C_{1s} \) of resonator \( GG' \). As \( C_{1s} \) is increased, the coupling to \( GG' \) increases so the external quality factor of mode \( f_{01} \) should decrease [23] leading to decrements of the second passband quality factor, as shown in Figure 24(a). However, the quality factor of the other band remains unaffected as \( C_{1s} \) has no effect on the coupling to other modes of the first path filter \( BB' \). When the tuning capacitor \( C_5 \) is increased, mode \( f_{o5} \) first comes closer to modes \( f_{e5}, f_{e5} \), and then moves away from them, so the quality factor of the second passband first increases then decreases [23], as shown in Figure 24(b). The external quality factors, \( Q_{ex_1}, Q_{ex_2}, Q_{ex_0}, \) and \( Q_{ex_0} \) of both the passbands are unchanged due to their independency of capacitors, \( C_{1s} \) and \( C_5 \).

### 4.5 Filter design steps

The design procedure is as follows:

**STEP 1:** Coupling-matrix synthesis from the given specifications: At the very first step of the filter design, two separate coupling-matrices for the two passbands—based on the desired specifications that is given centre frequency, bandwidth, and return loss—must be determined. The extended general coupling-matrix for \( m \)th-order filters can be represented as [23]:

\[
M = \begin{bmatrix}
    m_{11} & m_{12} & \ldots & m_{1n} & m_{1l} \\
    m_{12} & m_{22} & \ldots & m_{2n} & m_{2l} \\
    \vdots & \vdots & \ddots & \vdots & \vdots \\
    m_{nn} & m_{nl} & \ldots & m_{nn} & m_{ml} \\
    m_{l1} & m_{l2} & \ldots & m_{ln} & m_{ll}
\end{bmatrix}
\]

where the suffixes, \( s, n \) and \( l \) denote the source, mode and load, respectively. So the coefficients of this matrix will be calculated as per desired requirements.

**STEP 2:** Calculation of desired resonating frequencies and external quality factors: The required mode frequencies and associated external quality factors are calculated using the synthesised coupling matrix as described in Ref. [23] and can be written as:

\[
Q_{ex} = \frac{1}{m_{nn} F_{BW}}
\]

\[
f_n = f_c \left( 1 - \frac{m_{nn} F_{BW}}{2} \right)
\]

where \( f_n \) is the mode frequency and \( f_c \) is the centre frequency of desired band. For this particular filter five resonating modes, \( f_{o1}, f_{e1}, f_{o2}, \) \( f_{e2} \) and \( f_{load} \) and associated Quality factors, \( Q_{ex_1}, Q_{ex_2}, Q_{ex_0}, \) and \( Q_{ex_0} \) respectively, can be calculated using above equations.

**STEP 3:** Finding different dimensions (\( l_1, l_2, l_3, l_4, l_5 \)) of the filter: For the fixed length of \( l_1 \) (as discussed in Figure 5) the dimensions \( l_2 \) and \( l_4 \) can be obtained using (11) and (15) while keeping the loaded capacitors, \( C_3, C_3, \) and \( C_4 \) in the middle of the available capacitor tuning range. For the simplicity of the calculation, \( l_3 \) is kept same as \( l_2 \). Later on, the dimension of the resonator \( GG' \), \( l_5 \) in the second path is determined to create resonance at \( f_{load} \) using (20) when the loaded capacitors \( C_5 \) are in the middle of the available tuning range.

**STEP 4:** Finding \( C_1 \) and \( C_{1s} \): Two coupling capacitors, \( C_1 \) and \( C_{1s} \), can be determined for the desired quality factors using (21) and (22) to satisfy \( Q_{ex_1} \) and \( Q_{ex_2} \) obtained from STEP 2.
STEP 5: Simulation in ADS and optimisation: After getting all the physical dimensions, the filter response is simulated in ADS. Later on, if needed, the dimensions can be optimised with the available capacitors to achieve the required specifications.

5 | FILTER DESIGN, FABRICATION AND MEASUREMENT

5.1 | Design

For a BPF with second-order first band with centre frequency of 1.06 GHz and FBW of 15% and third-order second band with centre frequency of 2.94 GHz and FBW of 6.4%, coupling matrices are calculated using (23), which will be used for finding filter response and also can be used for designing purpose, and can be written as:

\[
m_1 = \begin{bmatrix}
0 & 0.6374 & 0.5128 & 0 \\
0.6374 & 0.6420 & 0 & -0.6374 \\
0.5128 & 0 & -0.6319 & 0.5128 \\
0 & -0.6374 & 0.5128 & 0 
\end{bmatrix}
\]

\[(26)\]

\[
m_2 = \begin{bmatrix}
0 & 0.462 & 0.454 & 0.448 & 0 \\
0.462 & 0.963 & 0 & 0 & -0.462 \\
0.454 & 0 & -0.107 & 0 & 0.454 \\
0.448 & 0 & 0 & -1.070 & -0.448 \\
0 & -0.462 & 0.454 & -0.448 & 0 
\end{bmatrix}
\]

\[(27)\]

The response obtained using the coupling matrices given by (26), \(m_1\) for the first passband and \(m_2\) given by (27) for the second passband is plotted in Figure 25. Here, the simulated response and calculated response using (26) and (27) coincides quite well with each other. The calculated desired external quality factors \(Q_{e1}, Q_{e2}, Q_{e201}, Q_{e202}\) and \(Q_{e203}\) from (26) and (27) are 16.4, 25.3, 73.2, 75.8 and 77.8, respectively, which are shown concerning varactors in Figures 23 and 24.

5.2 | Fabrication

The proposed dual-band tunable filter having first and second passbands of two and three poles, respectively, along with bandwidth control mechanism is designed on RT/Duroid 6010.2LM substrate of thickness 0.635 mm and of dielectric constant 10.2. The physical layout of proposed BPF along with its biasing scheme is shown in Figure 26. The varactors shown in Figure 1 are realised by varactor diodes while fabricating the design. Each varactor diode is biased with DC block capacitance of 100 pF and RF block resistance of 10 kΩ. The voltages \(V_{2}, V_{3}\) and \(V_{4}\) are used to control the odd- and even-mode resonant frequencies \(f_{01}, f_{02}, f_{03}\) and \(f_{20}, f_{21}\), voltage \(V_{5}\) is used for controlling mode \(f_{11}\), and voltages \(V_{1}\) and \(V_{11}\) are used for control of external quality factors. The Skyworks silicon varactors SMV1236 (\(C_{j} = 3.8–26.75\) pF), SMV1237 (\(C_{j} = 9.47–71.82\) pF), SMV1235 (\(C_{j} = 2.38–18.22\) pF), SMV1233 (\(C_{j} = 0.84–5.08\) pF), SMV1231 (\(C_{j} = 0.46–2.35\) pF) and SMV1232 (\(C_{j} = 0.72–4.15\) pF) are used for \(C_{e1}, C_{e2}, C_{e3}, C_{e4}, C_{e5}, C_{e6}\) and \(C_{e7}\), respectively. A SPICE model for the SMV123x varactors is shown in Figure 27 [24]. The required model parameters can be obtained from the SMV123x varactors datasheet [24]. Considering the design steps presented in Section 4.5, the BPF has been developed and optimised utilising the electromagnetic solver Keysight Advanced Design System (ADS). Finally, the response of the fabricated BPF (shown in Figure 28) is measured using a Vector Network analyser (VNA) and shown in Figures 29 and 30. All the geometrical parameters denoted in Figure 1 are given in Table 1.
5.3 | Measurement

Figures 29 and 30 show the simulated and measured responses of the dual-band tunable BPF. In Figure 29, the first and second passbands are tuned from 0.9 to 1.24 GHz and 1.8 to 2.45 GHz, respectively, with return loss better than 10 dB. The 3-dB absolute bandwidths of 220 and 280 MHz for the first and second passbands are maintained across their respective frequency tuning ranges. Moreover, the measured insertion losses \( S_{(21)} \) for the first and second passbands vary from 2.3 to 2.95 and 4.23 to 6.4 dB, respectively. Therefore, to uphold the absolute bandwidth constant throughout the frequency tuning ranges, the varactor diodes must be appropriately biased. Also, the transmission zero to the higher side of the first passband and to the lower side of the second passband provides isolation better than 39 dB between the two passbands. Furthermore, transmission zero on the lower side of the first passband gives good out of band rejection.

The bandwidth tuning of the first and second passbands is shown in Figure 30. The 3-dB bandwidth of the first passband can be tuned from 110 to 360 MHz for a given centre frequency between 0.9 and 1.24 GHz, and the observed insertion loss varies from 1.8 to 3.7 dB. Whereas the second passband bandwidth can be tuned from 230 to 700 MHz for a centre frequency range from 1.8 to 2.45 GHz, and the measured insertion loss varies from 3.02 to 6.7 dB. The insertion loss is mainly due to the varactor diodes and conductor loss, which increases with a decrease in bandwidth at any centre frequency. The voltages applied for biasing the varactor diodes for centre frequency and bandwidth tuning are listed in Tables 2 and 3, respectively.

Figure 31 shows the measured IIP3 with 1 MHz frequency spaced two tones and P1dB of two passbands. The measured IIP3 for the first passband varies from 30 to 40 dBm, and for the second passband from 23 to 42 dBm. P1dB of the proposed filter varies from 20.78 to 27 dBm for the first passband, and from 17.1 to 32 dBm for the second passband. The comparison between the proposed dual-band filter with other high-performance dual-band tunable BPFs is presented in Table 4. The designed filter provides a wide tuning range for both the passband because the stub-loaded varactors used on the resonator give the wide tuning range. Compared with other filters, as per the authors’ knowledge, bandwidth tuning is available only in this proposed filter along with centre frequency tuning, which is desirable. Both the passbands have constant absolute bandwidth tuning over the entire tuning range of the centre frequency, whereas other filters do not have this property.

| FIGURE 28 | Fabricated dual-band tunable filter |
| FIGURE 29 | Measurement and simulation results of the dual-band tunable BPF with first and second passband centre frequency tuning |
| FIGURE 30 | Bandwidth tuning of the first and second passbands at centre frequencies. (a) 0.9 and 1.8 GHz, (b) 1.12 and 2.24 GHz |

| TABLE 1 | Physical dimensions of the proposed BPF (units: millimetres) |
|---|---|---|---|---|---|---|---|
| \( w \) | \( w_1 \) | \( l_1 \) | \( l_2 \) | \( l_3 \) | \( l_4 \) | \( l_5 \) |
| 0.59 | 1 | 3 | 5 | 5 | 4.5 | 10 |
6 | IMPROVEMENT IN SELECTIVITY AND INSERTION LOSS OF THE FILTER

6.1 | Improvement in selectivity of second passband

The selectivity on the higher side of the second passband is not as good as the lower side of the passband (shown in Figures 29 and 30) because the first odd mode $f_{lo1}$ of the resonator $GG'$ shifts the higher side transmission zero to the lower side to maintain the CABW and quality factor, as shown in Figure 21. To improve the second passband selectivity, the varactors $C_4$ are removed from the resonator $GG'$ ends, and another varactor $C_6$ is connected symmetrically between the points $H$ and $H'$, as shown in Figure 32. The effect of varactor $C_6$ on the modes and transmission zeros of the newly designed filter is shown in Figure 33. As discussed in Figure 21, varactor $C_5$ changes all mode frequencies of the resonator $GG'$. However, the varactor $C_6$ can only change the odd mode frequencies of the resonator $GG'$.

As shown in Figure 33, when $C_6$ is increased from 0 to 1.6 pF, the first odd mode of $GG'$, $f_{lo1}$, shifts to a lower value, $f_{lo12}$ and also two transmission zeros—one at the lower side and another at the higher side of the second passband—are generated. The higher side transmission zero is generated due to the parallel combination of varactor $C_6$ and section $HH'$. 

6.2 | Fabrication and measurement

By considering the effect of varactor $C_6$, another filter has been designed and fabricated, as depicted in Figure 34. Figures 35 and 36 show the measured and simulated frequency response of this new fabricated filter. The centre frequency tuning of both the passbands is shown in Figure 35 where the first band is tuned from 0.9 to 1.2 GHz, and the second passband is tuned from 1.9 to 2.45 GHz with return loss better than 10 dB. The 3-dB absolute bandwidth of 200 and 250 MHz for the first and second passbands is maintained over the entire frequency tuning ranges. The bandwidth tuning of the first and second passbands is shown in Figure 36. The 3-dB bandwidth of the first passband can be tuned from 115 to 360 MHz for a given centre frequency between 0.9 and 1.2 GHz, whereas the

| Centre Frequency Tuning, $f_{c1}/f_{c2}$ (GHz) | $V_1$ | $V_{11}$ | $V_2$ | $V_3$ | $V_4$ | $V_5$ |
|---------------------------------------------|------|---------|------|------|------|------|
| 0.90/1.80                                   | 0    |  1.3   |  3.7 |  4.1 |  1.2 |  2.4 |
| 1.00/2.00                                   | 1.6  |  2.5   |  4.8 |  5.1 |  2.4 |  3.1 |
| 1.12/2.24                                   | 4.2  |  2.6   |  8.1 |  8.7 |  3   | 10.3 |
| 1.24/2.45                                   | 7.1  |  5.9   | 12.9 | 10.6 |  7   | 15   |

| $f_{c1}/f_{c2}$ (GHz) | BW@ $f_{c1}$ (MHz) | BW@ $f_{c2}$ (MHz) | $V_1$ | $V_{11}$ | $V_2$ | $V_3$ | $V_4$ | $V_5$ |
|-----------------------|-------------------|-------------------|------|---------|------|------|------|------|
| 0.90/1.80             | 110               | 700               | 0    |  0      |  3.3 |  2.7 |  3.7 |  3.9 |
| 1.00/2.00             | 190               | 460               | 0    |  0.4    |  3.3 |  3.4 |  2   |  2.9 |
| 1.12/2.24             | 360               | 230               | 0    |  1.7    |  2.3 |  4.9 |  0.9 |  2.5 |
| 1.24/2.45             | 360               | 230               | 1    |  3.3    | 10.8 |  2.5 | 10.3 |

**TABLE 2** Bias voltages for centre frequency tuning (units: volts)

**TABLE 3** Bias voltages for bandwidth tuning at different centre frequencies (units: volts)

**FIGURE 31** Measured IIP3 and P1dB of proposed dual-band BPF

**FIGURE 32** Physical layout of dual-band tunable filter with improved selectivity

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The second passband bandwidth can be tuned from 160 to 500 MHz for a centre frequency range from 1.9 to 2.45 GHz. The voltages applied for biasing the varactor diodes of this filter for centre frequency, and bandwidth tuning are listed in Tables 5 and 6.

The measured insertion loss of first and second passbands varies from 0.6 to 2.5 and 1.9 to 3.5 for the centre frequency and bandwidth tuning of the filter. Here, the insertion loss is improved as compared with the first fabricated filter. Improvement in the insertion loss of this filter compared with the previous fabricated filter is achieved due to two specific reasons—first, the number of varactors has been reduced, and second, varactor diodes (Cv4) SMV1233 has been replaced with SMV1234 which has lower resistant value.

### Table 4: Comparisons of measured results with other dual-band BPFs

| References | BPF Order First/Second | Freq. Tuning Range First/Second (GHz) | Bandwidth Tuning (MHz) | Centre Freq. Tuning (%) | CABW | Insertion Loss (dB) | Size |
|------------|------------------------|---------------------------------------|------------------------|-------------------------|------|---------------------|------|
| [11]       | 1/1                    | 0.69–0.88/2.67–3.78                   | No                     | 24.2/31.3               | No/No | 0.78–1.83/1.76–2.02 | 0.95λg × 0.39λg |
| [14]       | 2/2                    | 0.76–1.78/1.6–2.63                    | No                     | 80/48.7                 | 79 ± 9/No | 2.2–4.22/2.4–4.78 | 1.31λg × 0.23λg |
| [8]        | 2/2                    | 0.617–0.817/1.38–2.02                 | No                     | 27.9/37.3               | No/No | 1.5–3.4/2.1–5.2  | 0.10λg × 0.25λg |
| [25]       | 2/2                    | 1.25–1.52/Fix                         | No                     | 19.4/No                 | No/No | 1.45–4.21/1.51  | 1.06λg × 0.36λg |
| [13]       | 2/2                    | 0.7–1.0/1.57–2.0                      | No                     | 0.25/0.24               | No/120 ± 8 | 0.7–1.4/2.74–3.93 | 0.77λg × 0.36λg |
| [18]       | 2/2                    | 1.48–1.8/2.4–2.88                     | 100–148/230–347        | 19.5/18.2               | No/No | 3.02–4.8/2.78–4.6 | 0.32λg × 0.33λg |
| Figure 28  | 2/3                    | 0.9–1.24/1.8–2.45                     | 110–360/230–700        | 31.77/30.58             | 220 ± 5/280 ± 5 | 1.8–3.7/3.02–6.7 | 0.86λg × 0.57λg |
| Figure 34  | 2/3                    | 0.9–1.2/1.9–2.45                      | 115–360/160–500        | 28.5/25.3               | 200 ± 5/250 ± 5 | 0.6–2.5/1.9–3.5  | 0.54λg × 0.48λg |

$\lambda_g$ is the guided wavelength for highest frequency of the second band.

**Figure 33** Variation of modes with capacitor, $C_6$

**Figure 34** Photograph of fabricated dual-band tunable filter with improved selectivity

**Figure 35** Measurement and simulation results of the dual-band tunable BPF with first and second passbands centre frequency tuning
Figure 37 shows the measured IIP3 and 1 dB compression point (P1dB) of two passbands. The measured IIP3 for the first passband varies from 26.3 to 32.6 dBm, and for the second passband from 19.2 to 22.2 dBm. About 1 dB compression point varies from 23 to 28 dBm, and 15 to 20 dBm for the first and second passbands, respectively. The new filter's size is $0.54 \lambda_g \times 0.48 \lambda_p$, which is smaller than the previous fabricated filter. The filter's vertical size has been reduced by bending the stub $CE$ (or $C'E'$), and the DC block capacitor at the input and output feed line has been brought closer to the resonator to reduce the filter's horizontal size. Similarly, the size of the previously fabricated filter can also be reduced.

**Figure 36** Bandwidth tuning of the first and second passbands at centre frequencies. (a) 1 and 2.1 GHz. (b) 1.1 and 2.3 GHz

**Figure 37** Measured IIP3 and P1dB of dual-band BPF
The dual-band tunable BPF with CABW throughout the centre frequency tuning range is presented. A new design approach—study of the modal waveform nature of a conventional half-wavelength transmission line—has been considered to decide the suitable locations of the stubs and varactors for bandwidth and centre frequency tuning. Based on the proposed stub and varactor-loaded resonator, dual-band BPF with a controllable bandwidth mechanism has been created, while external quality factor control ability has also been investigated. To analyse the circuit behaviour, S-parameter is extracted using ABCD matrices of different sections of the filter. The coupling matrix method is used to achieve the desired frequency response of the filter, and based on that the required filter is designed.

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