Electrically varactor-tuned bandpass filter with constant bandwidth and self-adaptive transmission zeros

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Abstract: This paper presents a wideband frequency-agile bandpass filter (BPF) with constant absolute bandwidth (ABW) and two self-adaptive transmission zeros (TZs) at lower and higher stopbands. A frequency-dependent source-to-load (S-L) coupling is incorporated in this design so that two symmetrical TZs close to the tunable passband are self-adaptive and the separations between the passband center frequency and two TZs can almost keep the same for achieving continuous high selectivity in the frequency-tuning process. A lumped capacitor \( C_1 \) and a varactor diode are loaded to the two open ends of the half-wavelength resonator. With the aid of the employed \( C_1 \), the desired external quality factor can be obtained in a wide frequency-tuning range. Meanwhile, properly choosing the specific coupling region can make the coupling coefficient close to the desired one. As a result, the constant ABW can be obtained as the passband frequency changes. For demonstration, the proposed BPF is designed and fabricated. The simulated and measured results are presented. The results exhibit the tuning range from 430 MHz to 720 MHz with 3-dB constant ABW of around 75 MHz and the insertion loss of 1.34–2.92 dB, and the distance from TZs to center frequency is nearly constant at 150 MHz.

1 Introduction

Reconfigurable/tunable microwave devices are highly demanded in modern wireless communication systems. To cater for this trend, microwave filter, as a key frequency-selective component in the system, should be reconfigurable or tunable, which has been a hot research topic in the academic and industrial circles. In the past decades, various kinds of tunable filters have been developed using radio-frequency (RF) microelectromechanical system [1–3], yttrium–iron–garnet [4], piezoelectric transducer [5], semiconductor diodes, and so on. In among, varactor-tuned bandpass filters (BPFs) are widely explored and applied due to their small size, rapid tuning speed, and low cost [6–20]. Accordingly, many efforts have been paid to make the tunable BPFs more attractive such as expanding the frequency-tuning range, improving the linearity of the BPFs and realising controllable bandwidth [16–20].

It is well known that the selectivity is a key performance of the filter. For a frequency-tuning filter, the performance of the filter usually varies as the passband frequency is changed including the passband width and selectivity. Recently, the designs with constant absolute bandwidth (ABW)/fractional bandwidth have drawn much attention of the microwave engineer. However, the selectivity has rarely mentioned in these previous designs. For example, the tunable filters in [21–23] have only one or none transmission zero (TZ). In [24, 25], two or more TZs in the stopbands are realised, but the passband selectivity is variable in the frequency-tuning process. To achieve the continuous high selectivity, which is highly desirable in the tunable filter, a tunable source-to-load (S-L) coupling realised by extra varactor diodes in [26] is used to generate two TZs, and the TZs are always close to the passband. However, too many varactor diodes are used, leading to the increase of IL (about 3.8–4.9 dB).

In this paper, a high-selectivity tunable BPF with two self-adaptive TZs and constant ABW is presented. A frequency-dependent S-L coupling is incorporated for the first time to realise two self-adaptive TZs symmetrically loaded on the two sides of the passband. The two TZs keep the same distance to the passband centre frequency, so that continuously high selectivity can be kept as the passband frequency changes, which is highly desirable in many practical applications. Meanwhile, both the desired external quality factor (\( Q_e \)) and the coupling coefficient (\( K_{12} \)) between resonators in a wide-frequency-tuning range can be obtained by choosing the proper parameters. As a result, the constant ABW of the proposed tunable filter, which could hardly be affected by the S-L coupling, can be obtained as the passband frequency varies. The demonstration tunable filter is designed and implemented, and the simulated and measured results are presented, showing good agreement.

2 Design considerations for the proposed tunable BPF

The schematic layout of the proposed second-order tunable BPF with its coupling mechanism is illustrated in Fig. 1. It consists of two coupled asymmetric half-wavelength (\( \pi/2 \)) resonators with varactor \( C_v \) and capacitor \( C_1 \) loaded at both ends, a pair of feeding lines through capacitor \( C_2 \) as shown in Fig. 2 and two shunt stubs for realising S-L coupling. The specifications of the proposed second-order tunable BPF with Chebyshev response are ABW = 75 MHz and two TZs located at \( f_0 \pm 150 \) MHz, where \( f_0 \) is the centre frequency of the tunable passband. The filter design depends on extracting \( K_{12} \) and \( Q_e \), and the synthesis of the self-adaptive TZs is based on the coupling matrix \( M \) shown below.
where $g_0$ and $g_1$ are the element values of Chebyshev lowpass prototype. To maintain constant ABW across the tuning range of $f_0$, $Q_e$ should be increased as $f_0$ shifts upward, and the desired variation slope $Q_e/f_0$ is $g_0/g_1$. Accordingly, the passband performance can be good in the frequency-tuning range.

According to the feeding scheme shown in Fig. 2, where the stubs for S–L coupling in Fig. 1 are neglected because their effect on $Q_e$ is slight. Accordingly, $Q_e$ can be obtained from the equation below

$$Q_e = \frac{a_0}{2\nu_0} \frac{\partial \mathrm{Im}[Y_{in}]}{\partial \omega} = a_0$$  \hspace{1cm} (5a)

where

$$Y_{in} = \frac{j\omega C Y_c}{j\omega C + Y_i}$$  \hspace{1cm} (5b)

$$Y_i = \frac{j\omega Y(C_1 + C_v)(1 - \tan \theta_1 \tan \theta_2)}{(Y - \omega C \tan \theta_1)(Y - \omega C \tan \theta_2)}$$  \hspace{1cm} (5c)

From (5), it can be found that $Q_e$ is mainly controlled by $C_v$, $Y$, $u = \theta_1/\theta_2$ and $C_1$, where $Y$ is the characteristic admittance of the resonator, which is determined by the width $W_1$ of the employed resonator. As a result, $Q_e$ can be the function of these parameters

$$Q_e = f(C_v, W_1, u, C_1)$$  \hspace{1cm} (6)

According to the parameters’ study shown in Fig. 3 (the variation of frequency in horizontal axis results from the change of $C_v$, i.e. the resonant frequency of the resonator in Fig. 2 is changed by tuning the value of $C_v$, where $C_{vmin} = 1\, \text{pF}$ and $C_{vmax} = 8\, \text{pF}$ are used), $C_2$ affects the value of $Q_e$ and has rare effect on the variation slope of $Q_e$. As shown in Fig. 3b, as $W_1$ increases, its effect on $Q_e$ becomes slight, especially when $W_1 \geq 0.5\, \text{mm}$. Both $u$ and $C_1$ have influence on the $Q_e$’s value and variation slope. It is obvious that the effect of $C_1$ is much more significant than that of $u$, as can be seen from Figs. 3c and d. Accordingly, the incorporation of $C_1$ is significant for making the value of $Q_e$ close to the desired one in a wide-frequency range, resulting in the broadening of frequency-tuning range of the passband with constant ABW, as compared with the design [24].

### 2.2 $K_{12}$

To design a second-order filter, the inter-stage coupling $K_{12}$ is given by

$$K_{12} = \frac{A}{f \sqrt{g_0 g_1}}$$  \hspace{1cm} (7)

In this design, a mixed electric and magnetic coupling scheme based on the coupling region with length $L_t$, combining the effect of $C_1$, is investigated to achieve the constant ABW in a wide-frequency range.

As can be found from the coupling structure in Fig. 1, $K_{12}$ contains two different couplings, i.e. electric coupling ($K_e$) and magnetic coupling ($K_m$). To obtain the two couplings, the normalised voltage and current at fundamental resonance are employed, as in Fig. 4. According to the transmission line theory, the normalised voltage and current can be expressed as

$$V_1(x) = \cos \beta_1 x$$  \hspace{1cm} (8a)

$$I_1(x) = \sin \beta_1 x$$  \hspace{1cm} (8b)
where $\beta_i$ is the propagation constant and $i = L$ or $H$. The electric and magnetic coupling coefficients $K_e$ and $K_m$ can be calculated by

\begin{align}
K_{e,i} &= p \times \int_{x_1}^{x_2} |V_{i}|^2 \, dx \quad (9a) \\
K_{m,i} &= p \times \int_{x_1}^{x_2} |I_{i}|^2 \, dx \quad (9b)
\end{align}

where $p$ is a positive constant. Subscripts $L$ and $H$ represent the lower and higher-frequency states, respectively.

According to (9a) and (9b), $K_{m}$ and $K_{e}$ can be seen as the areas between the respective curves and $x$-axis (from $x_1$ to $x_2$) in Fig. 4. It can be found that the area of $K_{m}$ is larger than that of $K_{e}$, i.e. the magnetic coupling is dominated. As the coupling region (from $x_1$ to $x_2$) is chosen, the variation trends of $K_{m}$ and $K_{e}$ can be obtained, namely [24]

\begin{align}
|K_{e,L}| < |K_{e,H}| \\
|K_{m,L}| > |K_{m,H}|
\end{align}

(10a)

(10b)

Thus $K_{12}$ can be obtained as

\begin{align}
|K_{12}| = |K_{m}| - |K_{e}|
\end{align}

(11)

From (10) and (11), we can find $K_{12}$ decreases as $f_0$ shifts upward, conforming to the condition of constant ABW.

In the proposed filter structure, $K_{12}$ mainly depends on the coupling region with length $L_1$. Thus, it is controlled by the parameters $L_1$ (coupling region), $d_1$ (coupling gap between the two resonators), $C_1$ and $W_1$. Moreover, then it can be expressed as

\begin{align}
K_{12} = f(L_1, d_1, W_1, C_1)
\end{align}

(12)

Fig. 5 shows these parameters’ study. As expected, $L_1$ has significant effect on the $K_{12}$’s value and variation slope. In this design, $d_1$ mainly controls the value of $K_{12}$ and rarely affects the variation slope of $K_{12}$. On the contrary, $W_1$ mainly controls the variation slope of $K_{12}$ when $W_1 \geq 0.5$ mm and has slight effect on the value of $K_{12}$. Meanwhile, $C_1$ has slight effect on $K_{12}$ including the value and variation slope, as compared with other parameters. As a result, $C_1$ possesses the ability of fine tuning $K_{12}$ once the dominate parameters such as $L_1$ and $d_1$ are determined, so that $K_{12}$ can be further close to the desired one in a wide-frequency range, as in Fig. 5d.

According to the above discussion of Sections 2.1 and 2.2, the separate controlling of $Q_e$ and $K_{12}$ can be obtained in the wide-frequency-tuning range. In this design, $d_1$ acts as a dominate parameter for controlling $Q_e$ while has slight effect on the value of $K_{12}$. Meanwhile, there is a common point that when $W_1 \geq 0.5$ mm, $C_1$ acts as a dominate parameter for controlling $Q_e$ while has slight effect on the value of $K_{12}$. Meanwhile, there is a common point that when $W_1 \geq 0.5$ mm,
it has only slight effect on both the value of $Q_e$ and the variation slope of $K_{12}$. Thus, $W_1 = 0.5$ mm is determined in the filter design.

There are independent parameters, i.e. $C_2$ and $u$ for $Q_e$ and $L_1$ and $d_1$ for $K_{12}$, which can be used to independently tune $Q_e$ and $K_{12}$ for meeting the requirement. As a result, the design procedure of the proposed tunable filter is simplified. Accordingly, by properly choosing their values, a wideband tunable BPF with constant ABW can be realised.

2.3 Implementation of self-adaptive TZs

It is well known that the S–L coupling $[M_{SL}$ in (1)] is used to realise the TZs for improving the selectivity, but how to control the TZs in the tunable BPF is still a challenging issue. In this work, two coupled shunt stubs are employed to realise electrically dominated S–L coupling, so that $M_{SL}$ is negative, while $M_{S1}/M_{2L}$ and $M_{12}$ have opposite sign.

To investigate the improvement of selectivity by the S–L coupling, the coupling matrix simulation software (MATLAB) is employed to calculate the corresponding frequency responses ($S$-parameters). By substituting the extracted $Q_e$ and $K_{12}$ (meeting the requirement of the constant ABW mentioned above) under different $f_0$ into (1), the desired elements ($M_{SL}$, $M_{12}$, and $M_{2L}$) in the coupling matrix $[M]$ can be obtained. Subsequently, the desired S–L coupling $M_{SL}$ in $[M]$ can be calculated using MATLAB under different $f_0$ for meeting the requirements of two TZs at $f_0 \pm 150$ MHz. Hence, the desired variation of $M_{SL}$ which can realise the self-adaptiveness of the TZs along the tunable passband can be obtained. Accordingly, the S–L coupling strength versus frequency should be investigated.

To extract $M_{SL}$, $C_2$ in Fig. 1 is set as a very small value (e.g. 0.01 pF), so that the input/output port is isolated from the resonator [28], resulting in a single transmission path ($M_{SL}$) for the proposed filter. By simulation, the S–L coupling level ($S_{21}$, dB) can be obtained as shown in Fig. 6. By converting $S_{21}$ in dB to a linear format and then substituting the linear $S_{21}$ at different frequencies in Fig. 6 into (13), $M_{SL}$ at different frequencies can be calculated

$$M_{SL} = \frac{1 - |S_{21}|^2}{|S_{21}|}$$

(13)

where $|S_{21}|$ is in linear format. The calculated coupling matrix $M_1$ at the initial state ($C_v = C_{vmin}$) is obtained as follows:

$$M_1 = \begin{bmatrix} S & 1 & 2 & L \\ S & 0 & 0.682 & 0 & -0.072 \\ 1 & 0.682 & 0 & 0.531 & 0 \\ 2 & 0 & 0.531 & 0 & 0.682 \\ L & -0.072 & 0 & 0.682 & 0 \end{bmatrix}$$

(14)

By studying the parameters of the coupled stubs, the variation of $M_{SL}$ is achieved, as in Fig. 7. $M_{SL}$ can be tuned by adjusting the coupling between the stubs (i.e. $L_2$ and $d_2$ in Fig. 1) to approach the desired values, while the tunable passband keeps almost the same. As a result, two self-adaptive TZs are built for enhancing the selectivity across the frequency-tuning range effectively.
The demonstrated filter is fabricated on the substrate of Rogers RO4003C with a dielectric constant $\varepsilon_r = 3.38$ and a thickness $h = 32$ mil. The varactor diodes JDV2S71E from Toshiba are used. The DC bias voltage $V$ for the varactor diode is $11-0\text{V}$, resulting in the corresponding value of $C_v$ changes from 1 to 8 pF. The simulation and experiment are conducted by using high-frequency structure simulator software and Agilent E8363C network analyser, respectively. The design procedures are as follows:

(i) Calculating the desired values of $Q_e$ and $K_{12}$ according to (4) and (7), respectively.

(ii) On the basis of (6) for $Q_e$ and (12) for $K_{12}$, the extracted values can be close to the desired ones by using correlative parameter studies.

(iii) Substituting the extracted $Q_e$ and $K_{12}$ under different frequencies into (1), and using the coupling matrix $[M]$ to calculate the desired values of $M_{SL}$ under different frequencies.

(iv) Studying the key parameters $L_5$ and $d_2$ of S–L coupling, the extracted values of $M_{SL}$ under different frequencies can be close to the desired ones. Accordingly, two TZs located at $f_0 \pm 150\text{MHz}$ can be obtained across the frequency-tuning range of the passband.

After optimisation, the physical parameters of the filter are determined as $L_1 = 44\text{ mm}$, $L_2 = 12.9\text{ mm}$, $L_3 = 13.1\text{ mm}$, $L_4 = 5\text{ mm}$, $L_5 = 17.7\text{ mm}$, $W_1 = 0.5\text{ mm}$, $W_2 = 0.3\text{ mm}$, $d_1 = 0.7\text{ mm}$, $d_2 = 0.15\text{ mm}$, $C_1 = 2.2\text{ pF}$, and $C_2 = 5\text{ pF}$. The circuit size is around $0.16\lambda_g \times 0.14\lambda_g$, where $\lambda_g$ is the guided wavelength at lowest-frequency passband. Fig. 8a shows the photograph of the fabricated tunable filter for verification.

In the tunable filter design, the fractional frequency-tuning range is used to evaluate the tunability of the filter. It is generally defined as

$$R_f = \frac{f_{0h} - f_{0l}}{(f_{0h} + f_{0l})/2}$$

where $f_0l$ and $f_0h$ mean the lowest and highest passband frequencies in the tuning range. Fig. 8b shows the simulated and measured results of the proposed filter. The passband frequency can be tuned from 0.72 to 0.43 GHz, corresponding to a fractional tuning range of 50.4%, as the bias voltage $V$ of the varactor diode varies from 11 to 0 V. The measured insertion loss varies from 1.34 to 2.92 dB and the 3 dB ABW is within 75 ± 4 MHz across the frequency-tuning range. Generally, the parasitic resistance $R$ of the varactor diode increases as $V$ is decreased. In this process of frequency decreasing, the unloaded $Q$ value of the employed resonator becomes worst. Accordingly, the two return loss (RL) poles become closer gradually, and even overlapped. As can be seen from Fig. 8b, the passband performance including RL and IL is acceptable at every tuning state. As expected, two symmetrical TZs in the lower and higher stopbands are self-adaptive as the passband frequency is tuned, and the distance from the centre frequency to TZs is always constant at 150 MHz across the entire tunable range, as shown in Fig. 9.

Fig. 10a plots the output power against the input power of the filter with a bias voltage of 2 V under two-tone test with spacing of...
1 MHz. Fig. 10b shows the measured input third-order intercept point (IIP3) in the frequency-tuning range. It can be seen that the filter exhibits an IIP3 of better than 22 dBm throughout the entire filter tuning range. The IIP3 of tunable filter highly depends on the non-linearity of the employed varactor diode. The IIP3 of the proposed design is higher than those of the previous designs with mention. In [26], a tunable S–L coupling realised by extra non-linearity of the employed varactor diode. The IIP3 of the

Table 1

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Implement technology</th>
<th>Tuning range</th>
<th>Filter order/ varactor number</th>
<th>IL, dB</th>
<th>ABW, MHz</th>
<th>IIP3, dBm</th>
<th>Numbers of TZs/self-adaptiveness</th>
<th>Size ($\lambda_g \times \lambda_g$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[21]</td>
<td>Microstrip</td>
<td>1.1–1.86 GHz (52.3%)</td>
<td>3/3</td>
<td>4.3–6.8</td>
<td>90 ± 8</td>
<td>10.8–15.8</td>
<td>1/no</td>
<td>0.2 × 0.22</td>
</tr>
<tr>
<td>[22]</td>
<td>Microstrip</td>
<td>0.95–1.55 GHz (48.0%)</td>
<td>2/3</td>
<td>2.4–2.8</td>
<td>nearly 120</td>
<td>×</td>
<td>1/no</td>
<td>0.28 × 0.23</td>
</tr>
<tr>
<td>[23]</td>
<td>Microstrip</td>
<td>0.92–1.03 GHz (11.3%)</td>
<td>2/6</td>
<td>0.16–1.41</td>
<td>23.8 ± 0.3</td>
<td>×</td>
<td>0/no</td>
<td>0.2 × 0.17</td>
</tr>
<tr>
<td>[24]</td>
<td>Microstrip</td>
<td>0.63–0.96 GHz (38.5%)</td>
<td>2/2</td>
<td>1.2–1.5</td>
<td>80 ± 3.5</td>
<td>13</td>
<td>2/no</td>
<td>0.12 × 0.12</td>
</tr>
<tr>
<td>[25]</td>
<td>Microstrip</td>
<td>0.78–1.36 GHz (54.2%)</td>
<td>2/2</td>
<td>1.6–2.0</td>
<td>60 ± 3</td>
<td>13</td>
<td>2/no</td>
<td>0.12 × 0.12</td>
</tr>
<tr>
<td>[26]</td>
<td>Microstrip</td>
<td>1.7–2.7 GHz (45.5%)</td>
<td>2/7</td>
<td>3.8–4.9</td>
<td>nearly 110</td>
<td>×</td>
<td>2/yes</td>
<td>0.12 × 0.07</td>
</tr>
<tr>
<td>[29]</td>
<td>CPW</td>
<td>1.87–2.37 GHz (23.6%)</td>
<td>2/2</td>
<td>1.48–1.58</td>
<td>410 ± 10</td>
<td>×</td>
<td>1/no</td>
<td>0.75 × 0.3</td>
</tr>
<tr>
<td>this work</td>
<td>Microstrip</td>
<td>0.43–0.72 GHz (50.4%)</td>
<td>2/2</td>
<td>1.34–2.92</td>
<td>75 ± 4</td>
<td>22–27.6</td>
<td>2/yes</td>
<td>0.16 × 0.14</td>
</tr>
</tbody>
</table>

The good performance makes the filter more attractive in the practical and industrial applications.

5 Acknowledgments

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6 References


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