Design of compact tri-band bandpass filter using stub-loaded quarter-wavelength SIRs

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Abstract A compact tri-band bandpass filter (BPF) with high selectivity and wide stopband performance is developed based on a modified stub-loaded quarter-wavelength stepped-impedance resonator (SL-SIR). The modified SL-SIR has flexibly controllable resonance modes under different schematics of the loaded stubs, from which, a tri-mode case is chosen to design a tri-band BPF with controllable centre frequencies and bandwidths. Furthermore, up to ten transmission zeros are created by introducing two source-load couplings simultaneously, which not only enhance significantly the selectivity of three passbands, but also widen greatly the stopband of the BPF. The tri-band BPF, operating at 1.27, 3.65, and 5.20 GHz, is designed and fabricated, and its measured responses are in good agreement with the simulated ones.

key words: Tri-band bandpass filter, multiple coupling paths, controllable centre frequencies and bandwidths, transmission zero, wide stopband

Classification: Microwave and millimeter wave devices, circuits, and hardware

1. Introduction

Multi-band bandpass filter (BPF) with high-selectivity [1–5], compact size [6–10] and controllable bandwidths [11–15] is an essential component in modern RF/microwave multi-service wireless communication systems. Up to now, many novel structures and design methods have been proposed in developing multi-band BPF [1–5]. Among which, multi-mode resonator is one of the most widely used structures in designing tri-band BPFs with compact size and high selectivity [16–29]. In [1], a short-circuited stub-loaded SIR and side feedlines are used to design a tri-band BPF with compact size. However, the centre frequency of the second passband is hard to control. Two-stubs-loaded [26] and three-stubs-loaded [27] SIRs are used to construct tri-band BPFs with independently controllable central frequencies, but the controllability of the passband bandwidths is not favorable. In [30], by using dual composite right-/left-handed resonators, a novel tri-band BPF is presented with independently controllable centre frequencies and bandwidths. However, the design method, especially, the proposed equivalent lumped element circuit, is too complicated to use.

In this letter, a modified SL-SIR is proposed and analyzed to design a compact tri-band BPF. With different schematics of the loaded stubs, the SL-SIR has flexibly controllable resonance modes to construct dual-band or tri-band BPFs.

As an example, the tri-mode resonator is chosen to design a tri-band BPF. The configuration of the BPF is devised to allow four geometrical parameters to realize flexible variation of the FBWs of the three passbands. Two separately changeable source-load coupling paths are devised to produce up to ten transmission zeroes, which not only enhance significantly the selectivity of three passbands, but also widen greatly the stopband of the BPF.

2. Filter Design

In this study, two cases of the loaded structure are separately changeable source-load coupling paths are designed and fabricated, and its measured responses are in good agreement with the simulated ones. The tri-band BPF, operating at 1.27, 3.65, and 5.20 GHz, is designed and fabricated, and its measured responses are in good agreement with the simulated ones.

Fig. 1(a) shows the transmission line model of the proposed SL-SIR, where $Z_l (i = 1, 2, t, s1, s2)$ and $\theta_l (i = 1, 2, t, s1, s2)$ represent the characteristic impedance and electrical length, respectively. The SL-SIR has an open stub with $Z_l$ and $\theta_l$ at its one end, and a short-circuited stub with $Z_s$ and $\theta_s$ at the other end, together with a loaded structure indicated by $Z_u$.

In this study, two cases of the loaded structure are considered:
Case 1: the loaded structure is one open stub with \((Z_{1l}, \theta_{1l})\), as shown by Fig. 1(b).

Case 2: the loaded structure consists of two shunt open stubs with \((Z_{1l}, \theta_{1l})\) and \((Z_{2l}, \theta_{2l})\), respectively, as shown in Fig. 1(c).

The simulated response \(S_{11}\) of the proposed SL-SIR under week coupling is given in Fig. 1(d), from which, it is clearly seen that in case 1, the proposed SL-SIR behaves as a dual-mode resonator, while in case 2, it performs as a tri-mode resonator.

For the tri-mode SL-SIR, which consists of one shorted-circuited stub and three open stubs, Fig. 2(a)-(c) provide the current distribution on the microstrip lines at resonance, \(f_1\), \(f_2\), and \(f_3\), respectively. It is seen from Fig. 2(a) that the resonance at \(f_1\) is occurred at path I as a \(\lambda/4\) SIR, which includes lines with electrical lengths \(\theta_1\), \(\theta_2\), and \(\theta_3\). The resonance at \(f_2\), as shown in Fig. 2(b), is happed at path II as a \(\lambda/2\) SIR, which consists of lines with electrical lengths \(\theta_1\), \(\theta_2\), and \(\theta_3\). The resonance at \(f_3\), as shown in Fig. 2(c), involves all the four stubs. Therefore, with appropriate choosing of the line lengths of the stubs, it is easy to design these three resonances at desired frequencies. More detailed design formulas are given below.

By referring to Fig. 2(a)-(c), we can derive the input admittances of the three paths as follows, with assumption that \(Z_2 = Z_3 = Z_{4l} = Z_{5l}\) and \(R_4 = Z_2/Z_1\) (The input admittance equation of path III is shown in the bottom of this page):

\[
\begin{align*}
Y_{Path I}^{in} &= \frac{j}{Z_1} \left[ 1 + R_s \tan \theta_1 \tan(\theta_1 + \theta_2) + R_s \tan \theta_2 \tan(\theta_1 + \theta_2) \right] - 1 \quad \text{for } f_1 \\
Y_{Path II}^{in} &= \frac{j}{Z_2} \left[ 1 + R_s \tan \theta_1 \cot(\theta_1 + \theta_2) + R_s \tan \theta_2 \cot(\theta_1 + \theta_2) \right] - 1 \quad \text{for } f_2 \\
Y_{Path III}^{in} &= \frac{j}{Z_3} \left[ 1 + R_s + R_s \left( \cot \theta_1 - \tan \theta_1 - \tan \theta_2 \right) \tan \theta_2 + \left( \tan \theta_1 + \tan \theta_2 - \cot \theta_1 + \tan \theta_2 \right) \cot \theta_2 \right] \quad \text{for } f_3 
\end{align*}
\]

The resonance condition for the three modes at \(f_1, f_2,\) and \(f_3\) are given by the following equations:

\[
\begin{align*}
R_s \tan \theta_1 \tan(\theta_1 + \theta_2) - 1 &= 0 \quad \text{for } f_1 \\
1 + R_s \tan \theta_1 \cot(\theta_1 + \theta_2) &= 0 \quad \text{for } f_2 \\
R_s + R_s \left( \cot \theta_1 - \tan \theta_1 - \tan \theta_2 \right) \tan \theta_2 + \left( \tan \theta_1 + \tan \theta_2 - \cot \theta_1 + \tan \theta_2 \right) \cot \theta_2 &= 0 \quad \text{for } f_3
\end{align*}
\]

For the tri-mode SL-SIR, \(R_3 + R_s \tan \theta_2 \left( \cot \theta_1 - \tan \theta_1 - \tan \theta_2 \right) + \cot \theta_1 \left( \tan \theta_2 - \cot \theta_1 + \tan \theta_1 + \tan \theta_2 \right) = 0\) for \(f_1\) (6)

From (4) and (5), it is seen that we can change the frequencies \(f_1\) and \(f_2\) independently by varying the electrical length \(\theta_1\) and \(\theta_3\). Furthermore, we can change the frequency \(f_3\) by varying the electrical length \(\theta_2\) without affecting the other two frequencies \(f_1\) and \(f_2\). So in the design of the tri-mode resonator, we determine first the impedance ratio \(R_3\) and the electrical lengths \(\theta_1\), \(\theta_2\), and \(\theta_3\) to get the resonance at \(f_1\). Next by choosing an appropriate length of \(\theta_1\), we get the resonance at \(f_2\). Finally, with appropriate selection of the length of \(\theta_2\), we obtain the third resonance at \(f_3\).
third passbands, respectively. The two resonators are folded to reduce the circuit size and at the same time, to get four geometrical parameters to control the coupling strength between them: the diameter \( d \) of the ground via hole commonly used by the two resonators, the gaps \( g_0, g_1 \) and \( g_2 \) between the other three open stubs of the two resonators. Fig. 4(a)-(d) show the variation of the 3-dB fractional bandwidths (FBWs) of the three passbands of the filter versus these four parameters. From Fig. 4(a), (b), and (d), it is seen that the FBW of the first, second, and third passband is significantly varied versus \( d, g_0 \), and \( g_2 \), respectively, while the FBWs of the other two passbands varied little. Fig. 4(c) shows that the FBWs of all these three passbands can be adjusted simultaneously by changing \( g_1 \). Therefore, the FBWs of the tri-band BPF can be independently controlled using these four parameters.

On the other hand, source-loaded coupling between the feed lines of the filter are made at two places to form two coupling paths, \( M_{SL-upper} \) and \( M_{SL-lower} \), as shown in Fig. 3. As will be shown later, by employing these two source-loaded couplings, the number of transmission zeros (TZs) is significantly increased, which results in greatly improved selectivity and stopband suppression of the BPF. The tri-band BPF is designed and optimized by the EM simulator Sonnet, and its final dimensions are given in Fig. 3.

### 3. Results and discussion

![Simulated and measured responses of the proposed tri-band BPF with an inset showing a photograph of the BPF.](image)

The tri-band BPF is designed on a substrate (Rogers RO4003C) with a dielectric constant of 3.38 and thickness of 0.813 mm. A photograph of the fabricated filter is shown by the inset of Fig. 5, and it shows that the filter is compact with a size of \( 0.06\lambda_g \times 0.15\lambda_g \), where \( \lambda_g \) is the guided wavelength at the center frequency \( f_1 \) (1.28 GHz) of the first passband. In Fig. 5, the simulated and measured results show a good agreement. The tri-band BPF operates at 1.28/3.65/5.20 GHz with respective FBWs of 7.26/9.50/5.90%, respectively. The measured minimum passband insertion losses are 1.10/1.32/1.25 dB and return losses are better than 16/21/24 dB, respectively. The overall stopband with rejection level larger than 15 dB is extended to 9.76 GHz (7.6f). In addition, ten TZs appearing at 1.19/1.47/2.28/3.98/4.72/5.50/7.58/8.14/8.55/9.63 GHz are produced, which improve significantly both the selectivity and the stopband performance of the BPF. Compared with the filter with seven TZs in [1], the increased three TZs appeared in the upper stopband of the current filter, enlarging the minimum rejection level from about 10 dB to more than 15 dB. Most of the TZs are generated by the simultaneous use of the upper and lower source-loaded couplings, \( M_{SL-upper} \) and \( M_{SL-lower} \), as shown in Fig. 3. In Fig. 6, the distributions of TZs with different feed configurations are compared. It is seen that the lower S-L coupling \( M_{SL-lower} \) creates \( f_6, f_8, f_9 \), \( f_6 \), \( f_8 \), and \( f_9 \), and the upper S-L coupling \( M_{SL-upper} \) generates \( f_2 \) and \( f_3 \), while the open stub with \( \theta_1 \) and \( \theta_2 \) produced \( f_1 \) and \( f_5 \), respectively. With the simultaneous use of both the upper and lower S-L couplings, we get finally up to ten TZs.

![Distribution of TZs with different feed configurations.](image)

### 4. Conclusion

A compact second-order tri-band BPF is developed by using our proposed SL-SIRs. Separate geometrical parameters are devised to implement independent control of the centre frequencies and FBWs of the three passbands. Two source-loaded couplings are employed to produce a significantly increased number of TZs and achieve thereby high selectivity and stopband suppression of filter. The proposed tri-band BPF and design approach are verified well by the measured results.

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