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Compact Tri-Band Bandpass Filter for WLAN and WiMAX Using Tri-Section Stepped-Impedance Resonators

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Abstract—In this paper, a compact tri-band microstrip bandpass filter (BPF) using tri-section stepped-impedance resonators (TSSIRs) with open stub line and zero-degree feed line is presented. Three tunable passbands are achieved by changing the ratio of impedance. By adding the open stub line, the frequency responses of tri-band at 2.4 GHz/3.5 GHz/5.2 GHz can be obtained precisely. With a proper length of the cone-shaped zero-degree feed structure, we can obtain transmission zeros in stopbands and better return loss in passbands. The proposed tri-band BPF has high selectivity and is simple in fabrication.

1. INTRODUCTION

With the development of modern wireless and mobile communication systems, high performance filters with compact size, low losses, and high selectivity are the goals. The wireless local area networks (WLAN) products operate at 2.4 GHz/5.2 GHz, and the world interoperability for microwave access (WiMAX) IEEE 802.16e operates from 3.3 to 3.9 GHz [1]. As increasing kinds of wireless services, the trend that multi-wireless services can be provided in a single wireless terminal also makes multi-band bandpass filter a necessity. There are many articles about dual-band BPFs that work at these frequencies. It is much more convenient to cover the three frequencies with one filter. As a result, tri-band bandpass filters at 2.4 GHz, 3.5 GHz, and 5.2 GHz [14–17] have been investigated extensively in recent years. The tri-band BPF for WLAN and WiMAX applications has become a hot topic.

Several approaches used to realize the tri-band filter have been reported. In [2], a tri-band bandpass filter is designed based on quarter-wavelength resonators with U-folded coupled-line. But its configuration is quite complicated. The most common tri-band filters are designed by cascaded SIRs, and the method is reported in [6]. Obviously, this way would yield a large dimension. Another typical way to design tri-band BPF is stub-loaded resonator (SLR) [12, 13]. Three desired frequencies can be realized by the combination of one set of half-wavelength resonators and one or two sets of SLRs. The tri-band filter can also be realized through partly combined two half-wavelength SIRs (PCSIRs) [3] with the enhanced stop band rejection and steep passband skirts. However, the measured insertion losses are nearly 3 dB, particularly at the second band. The square ring loaded resonator (SRLR) has been successfully proven in [14]. Though the high order tri-band is realized, it is difficult to tune the centre frequencies, and the selectivity needs to be improved. Utilizing tri-section SIR (TSSIR) to achieve the tri-band feature has been proposed in [4]. It is possible for tri-section SIR to locate the three passbands at any frequency with proper impedance ratios K_1 and K_2. The problem of the second band's performance is still to be improved.

In this letter, a tri-band BPF applied to WLAN and WiMAX is introduced. The impedance ratios K_1 and K_2 can be derived from the frequency ratios based on the theory of TSSIR. Thus, the three desired frequencies can be confirmed by K_1, K_2 and the length of the embedded open stub line. With a proper dimension of zero-degree feed structure, high selectivity and better return loss in passbands can be obtained.

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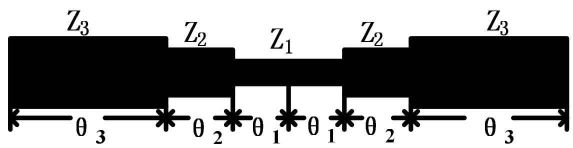

Figure 1. General structure of tri-section SIR.

2. PROPERTIES OF TRI-SECTION SIR

Figure 1 shows the basic structure of tri-section SIR (TSSIR). The SIR is symmetrical, and different characteristic impedances of three cascaded sections are Z_1, Z_2, Z_3. The corresponding electrical lengths are θ_1, θ_2, θ_3, respectively. For simplicity, it is preferable to have equal electrical lengths, $\theta_1 = \theta_2 = \theta_3 = \theta$. The condition for the fundamental resonance of a symmetrical tri-section SIR with the equal electrical lengths is given by [5, 10]

$$\theta = \tan^{-1}\left(\sqrt{\frac{K_1 K_2}{K_1 + K_2 + 1}}\right) \tag{1}$$

where $K_1 = Z_3/Z_2$ and $K_2 = Z_2/Z_1$. So the total electrical length of the resonator at the fundamental frequency can be derived as

$$\theta_{Tri} = 6\theta = 6\tan^{-1}\left(\sqrt{\frac{K_1 K_2}{K_1 + K_2 + 1}}\right) \tag{2}$$

The fundamental frequency, the first spurious resonant frequency, and second spurious resonant frequency are denoted as f_0, f_1, f_2. The f_1 locates at

$$f_1 = \frac{\theta_1}{\theta} f_0 \tag{3}$$

where

$$\theta_1 = \tan^{-1}\sqrt{\frac{1 + K_1 + K_1 K_2}{K_2}} \tag{4}$$

And the f_2 locates at

$$f_2 = \frac{\theta_2}{\theta} f_0 \tag{5}$$

$$\theta_2 = \frac{\pi}{2} \tag{6}$$

We can obtain that

$$\frac{f_1}{f_0} = \frac{\theta_1}{\theta} = \frac{\tan^{-1}\sqrt{\frac{1 + K_1 + K_1 K_2}{K_2}}}{\tan^{-1}\left(\sqrt{\frac{K_1 K_2}{K_1 + K_2 + 1}}\right)} \tag{7}$$

$$\frac{f_2}{f_0} = \frac{\theta_2}{\theta} = \frac{\pi}{2\tan^{-1}\left(\sqrt{\frac{K_1 K_2}{K_1 + K_2 + 1}}\right)} \tag{8}$$

From (7) and (8), tri-band bandpass filter can be achieved by adjusting the impedance ratio K_1 and K_2 properly. With these parameters properly determined, it is much more convenient to adjust the passbands at any desired frequency.

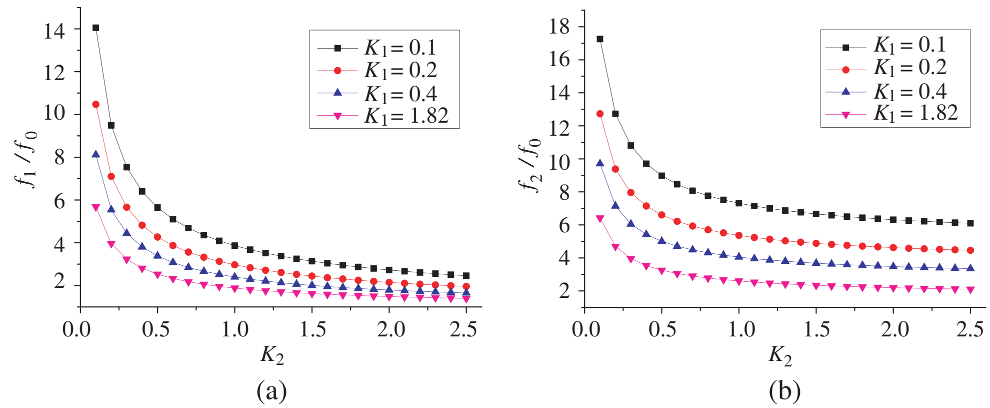

Figure 2. (a) f_1/f_0 for different K_1 and K_2 values. (b) f_2/f_0 for different K_1 and K_2 values.

Figures 2(a) and (b) show the frequency ratios as the function of the impedance ratios K_1 and K_2. For WLAN and WiMAX applications, frequencies 2.4 GHz, 3.5 GHz, and 5.2 GHz are chosen, so that $f_1/f_0 = 1.46$ and $f_2/f_0 = 2.17$. Then the impedance ratios can be determined from Fig. 2, where $K_1 = 1.82$, $K_2 = 2.13$.

3. THE DESIGN OF TRI-BAND BPF

According to K_1 and K_2, it can be obtained from (1) that $\theta = 41.5$ at 2.4 GHz. Letting $Z_3 = 95\,\Omega$, we can solve that $Z_2 = 52.2\,\Omega$, $Z_1 = 24.5\,\Omega$. The thickness of the substrate is 0.8 mm, with a relative dielectric constant of 2.65, and loss tangent of 0.003. With these electrical parameters, it is easy to obtain the physical parameters of the tri-section SIR. To create two extra transmission zeros in the stopband, the zero-degree feed structure [11] is applied to the design. The 0° feed structure is shown in Fig. 3, in which the difference between electrical delays of the upper path and lower path is 0°. The transmission matrices of the upper and lower paths can be simplified as

$$\tan\theta_1 \approx 1/Z_0\omega C_m \quad (9)$$

$$\tan\theta_2 \approx 1/Z_0\omega C_m \quad (10)$$

C_m is the coupling capacitance between the two paths and is very small. The transmission zeros occur at the frequencies when $\theta_1 \approx \pi/2$ or $\theta_2 \approx \pi/2$ [11]. One of the two resonators resonates at the frequency when θ_1 approaches $\pi/2$, and the other resonates at the frequency when θ_2 is about $\pi/2$. Since the two zeros are close to and on the opposite sides of the passband, the stopband rejection is increased significantly.

Owing to the extra created transmission zeros, the selectivity and performance of the filter are improved. Two open stubs are embedded in the filter to vary the resonant frequency precisely by changing the open stub length L_5.

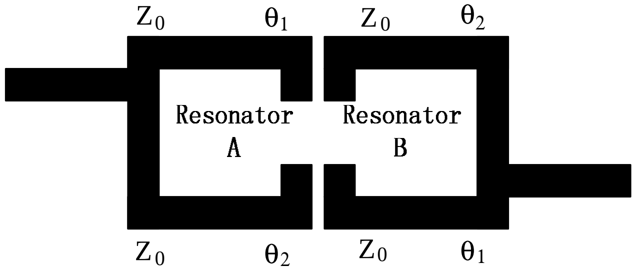

Figure 3. 0° feed structure.

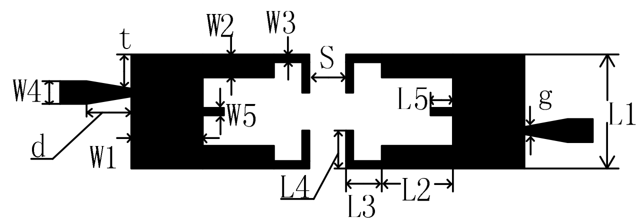

Figure 4. Layout of the proposed tri-band BPF.

Fig. 4 shows the layout of the proposed filter, composed of two TSSIR split rings, which use electric coupling. Electric coupling can achieve enough coupling volume with less design space, because it is inversely proportional to distance between two resonators.

The full wave simulation software HFSS has been taken to optimize the filter. The structure parameters of the tri-band BPF are chosen as follows: $W_1 = 5.28\,\text{mm}$, $W_2 = 2.03\,\text{mm}$, $W_3 = 0.67\,\text{mm}$, $W_4 = 2.18\,\text{mm}$, $W_5 = 1.6\,\text{mm}$, $L_1 = 18.77\,\text{mm}$, $L_2 = 9.24\,\text{mm}$, $L_3 = 5.42\,\text{mm}$, $L_4 = 5.34\,\text{mm}$, $g = 1\,\text{mm}$, $S = 0.72\,\text{mm}$, $t = 3.1\,\text{mm}$.

The lengths of the open stub L_5 and the cone-shaped structure of zero-degree feed line d influence the second and third passbands a lot.

All the other dimensions of this filter unchanged, the simulated frequency responses with different open stub lengths are presented in Fig. 5(a). Although L_5 varies from 1.6 mm to 7 mm, the first passband remain unchanged. While the central frequency of the second passband changes from 3.5 GHz to 3.3 GHz and the S_{21} of the third passband deteriorates rapidly. By using proper value of L_5, the central frequency of the second passband can be tuned to the desired frequency precisely with a good waveform around the third passband.

Letting $L_5 = 1.6\,\text{mm}$, the variation in d from 2 mm to 9 mm affects the return loss greatly in the second and third passbands. It relates to the conditions of the impedance matching at the three frequencies. The value of d has to be determined by optimization to guarantee the S_{11} of all the three passbands available. Fig. 5(b) shows the influenced S_{11} by the varied d at 3.5 GHz and 5.2 GHz.

According to the analysis above, L_5 and d can be determined, $L_5 = 1.6\,\text{mm}$ and $d = 9\,\text{mm}$. A

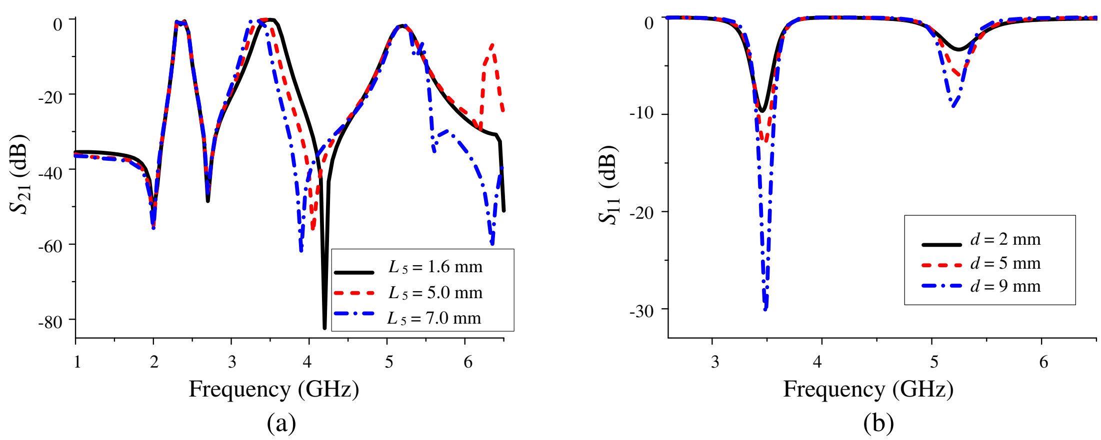

Figure 5. (a) Simulated frequency responses with different L_5. (b) Simulated frequency responses with different d.

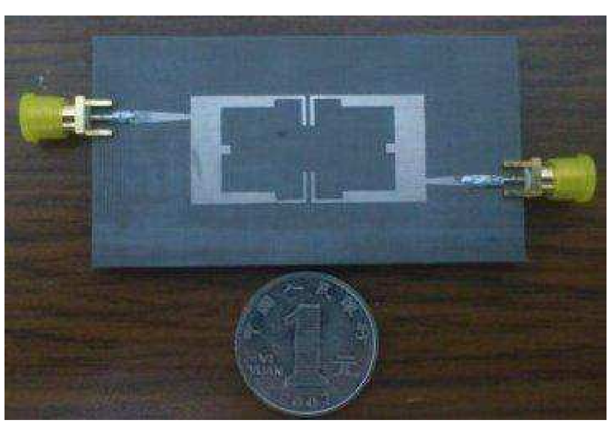

Figure 6. Photograph of the fabricated filter.

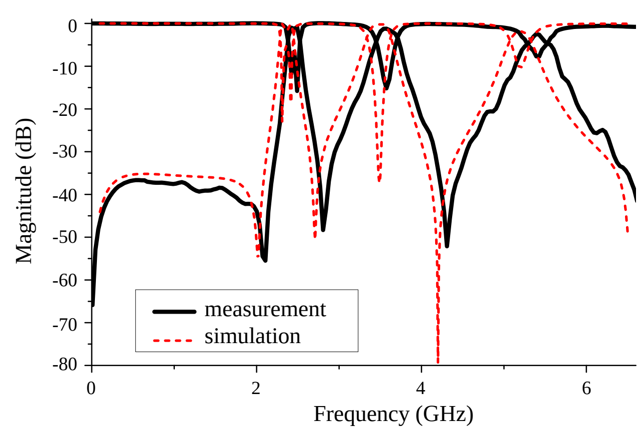

Figure 7. Simulated and measured results of the tri-band BPF.

photo of the fabricated filter is given in Fig. 6.

The full-wave simulated and measured results are plotted in Fig. 7. The measured three passbands are located at 2.42 GHz, 3.57 GHz, and 5.39 GHz with insertion losses of 0.99 dB, 1.24 dB, and 2.46 dB. The 3 dB fractional bandwidths are found to be 5.6%, 7.6%, and 5.8%, respectively. With an attenuation level of more than 40 dB, the three transmission zeros are realized at 2.11 GHz, 2.81 GHz, and 4.31 GHz, which achieve excellent band-to-band isolations. Owing to the accuracy of the fabrication, there are frequency shifts especially for the second and third passbands. The reason is that the fundamental and spurious resonant frequencies are highly sensitive to the widths of the lines. Compared with other reported tri-band filters [7–9], the IL and FBW of the proposed filter are comparative with other filters. It is noted that our filter performs much better with lower IL and larger FBW than others at the second passband. The performances of the four filters are listed in Table 1.

Table 1. Four tri-band filters for comparison. (IL: insertion loss, FBW: fractional bandwidth).

	1st/2nd/3rd Passband (GHz)	IL (dB)	FBW (%)	Application
Ref. [7]	2.45/3.5/5.25	2/2.4/1.7	2.5/1.7/5	WLAN/WiMAX/WLAN
Ref. [8]	1.8/2.7/3.3-4.8	2.2/2.1/1.3	3.9/2.6/40	GSM/WiMAX/UWB
Ref. [9]	1.95/3.46/5.25	1.5/1.2/1.6	9.7/6.4/9	WiMAX/WLAN
This work	2.42/3.57/5.39	0.99/1.24/2.46	5.6/7.6/5.8	WLAN/WiMAX/WLAN

4. CONCLUSIONS

A compact tri-band microstrip bandpass filter with high selectivity for WLAN and WiMAX applications is designed using tri-section stepped-impedance resonators. Three tunable passbands are generated with impedance ratios of the TSSIRs properly adjusted. Two open stub lines are added in the split rings to tune the second passband precisely. Two extra transmission zeros are introduced by zero-degree feed structure. The simulated and measured results exhibit a good performance around 2.4 GHz, 3.5 GHz and 5.2 GHz. The filter is simple in structure, easy to fabricate and applicable to WLAN and WiMAX.

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