DESIGN OF DUAL-BAND BANDPASS FILTER
FOR WLAN AND WIMAX APPLICATIONS

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Abstract

In this paper, dual-mode dual-band bandpass filters (BPFs) using short-stub loaded half-wavelength resonator are presented. Based on the lossless transmission line model analysis, it is found that two center frequencies can be easily adjusted to the desired value by tuning the length and width of short-stub and half-wavelength line. Furthermore, new structures of couple-line are introduced to satisfy desired external coupling degree for each band. In this way, the bandwidths of there filter are controllable, while taking good selectivity and improved upper-stopband performance. For demonstration purpose, a dual-band bandpass filter for WLAN (2.4GHz), WiMAX (3.5GHz) and two dual-band filters for WLAN (2.4/5.2GHz) applications are implemented with different structures. Three dual-band bandpass filters with very compact size are realized in the form of microstrip lines and its frequency responses are measured to validate this method.

1. Introduction

In recent years, the rapid development of the wireless communication applications are happening in the multiple band operations. As shown in the literatures, a dual-band bandpass filter (BPF) can be realized in various ways. Compact dual-band bandpass filters have been studied extensively as a key circuit block in dual-band wireless communication systems [1]–[10]. Stepped impedance resonator (SIR) can be used for achieving the fundamental resonance frequency which is related to the characteristic impedance and electrical length of the resonator [1]–[3], [5]. However, the size of this structure is not compact. In [4], a novel switchable bandpass filter with two-order frequency responses is presented, where dual-band bandpass and single bandpass characteristics can be conveniently switched by turning pin diodes to the on and off states. A dual band filter using dual-feeding structure and embedded resonators has been presented in [6]. Although the passband frequencies can be tuned to desirable values, it is difficult to control the bandwidths. To solve this problem, dual-band coupling and feed structures were proposed to fulfill the requirements of both passband frequencies and bandwidths [7]. By adjusting the length of the open stub, the central frequency

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of its second band can be changed without affecting the central frequency of the first band, a dual band filter has been proposed in [8]. Also dual-band bandpass filters using the novel embedded spiral resonators (ESRs), each ESR consisting of two sets of spiral-lines embedded in a microstrip rectangular open-loop is proposed in [9]. To conveniently control the center frequencies of one passband, while the another remains unchanged, source-load coupling is also employed to create four transmission zeros close to the passband edges, which has been presented in [10]. In [12], [13], two dissimilar ring resonators are employed. In these cases, the passbands with two poles are realized by the independent ring resonators. The dual-band filters with different shaped slots on the patch are proposed in [14], [15]. Recently, various defected grounded structures (DGS) have been presented in the filter designs [16]. However, the resonant frequencies of SIRs are dependent and the DGS structure should be insulated from other conductors of the ground plane, which complicates the filter fabrication.

Lately, the authors have presented an effective and simple method for dual-band filter design using short-stub loaded half-wavelength resonator [11], [17]. In this work, three compact microstrip dual-band bandpass filters for WLAN (2.4/5.2GHz) and WiMAX (3.5GHz) applications are presented. This paper is organized as follows: Section 2 describes the analysis of the proposed design concept. Dual-mode for dual-band filters have been presented. Section 3 provides the design of three microstrip dual-band bandpass filters for WLAN and WiMAX applications. The experimental data are presented and compared with the simulated results. Finally, Section 4 draws some brief conclusions.

2. Analysis dual-mode resonators

Figure 1(a) shows the conventional short stub loaded half-wavelength resonator. For even- and odd-mode excitation, the equivalent circuit are shown in Figures.1(b) and (c). In addition, $Z_1$, $L_1$ and $Z_2$, $L_2$ are the characteristic impedance and length of the half wavelength line and the short-circuited stub, respectively. The resonator is symmetrical and thus odd-and even-mode analysis can be used for characterization. Element values of $Z_1$, $L_1$ and $Z_2$, $L_2$ in the prototype filter are determined by its passband specifications using the well-known formulas in [4] and [10]. The input impedance for even-mode and odd-mode can be expressed as

$$Z_{ine} = jZ_1 \frac{Z_1 \tan(\beta L_1) + 2Z_2 \tan(\beta L_2)}{Z_1 - 2Z_2 \tan(\beta L_1) \tan(\beta L_2)}$$

(1)

$$Z_{ino} = -jZ_1 \cot(\beta L_1)$$

(2)

where $\beta$ is the propagation constant, and it is equal for even-mode and odd-mode. The resonance condition is

$$Z_{ine} \rightarrow \infty, Z_{ino} \rightarrow \infty$$

(3)
For the even-mode excitation, the equivalent circuit is shown in Figure 1(b). The fundamental resonant frequency $f_1$ can be determined as follows:

$$f_{\text{even}} = \frac{c}{4(L_1 + L_2)\sqrt{\varepsilon_{\text{eff}}}}$$  (4)

For odd-mode excitation, the equivalent circuit is shown in Figure 1(c), the odd-mode resonant frequency, namely $f_2$ can be determined as follows:

$$f_{\text{odd}} = \frac{c}{4L_1\sqrt{\varepsilon_{\text{eff}}}}$$  (5)

While $c$ is the light speed in free space, $\varepsilon_{\text{eff}}$ the effective dielectric constant of the substrate. From (4) and (5), when $L_1 > L_2 > 0$, we could obtain $f_{\text{even}}(f_1) < f_{\text{odd}}(f_2)$. Also, it means that the first resonance frequency, namely $f_1$ is controlled by the ratio of the characteristic impedance and electrical length of the half wavelength line and the short-circuited stub. Meanwhile, from (5), the second resonance frequency, namely $f_2$ can be dominated by adjusting the electrical length of the half wavelength line. To verify the proposed approach concept, all these filters are implemented using microstrip technology, and fabricated on a substrate with $\varepsilon_r = 9.2$ and $h = 0.8\,\text{mm}$. The design parameters obtained by Ansoft HFSS 10.0EM, and the measurement are accomplished using Agilent 8753ES network analyzer.

3. Design of dual-band bandpass filters

3.1. Dual-band for WLAN and WiMAX applications

Figure 2 shows the layout of proposed dual-band BPF utilizing short-stub loaded half-wavelength resonator in the midpoint with new structure. It consists of two identical half-
wavelength resonators and two $L$-shaped short-stubs. Both these two short-stub loaded half-wavelength resonators and two feed line are folded symmetrically. This coupling structure makes a very compact circuit. By tuning the length and width of short-stub and half-wavelength line, two center frequencies can be easily adjusted to the desired value. In this case, the odd-mode resonant frequency $f_{\text{odd}}$ is 3.5 GHz, and the even-mode resonant frequency $f_{\text{even}}$ is 2.4 GHz. The second resonance is caused by the half-wavelength resonator with length of $L$. The length $L$ is the overall length of the line, namely

$$L = L_1 + L_3 + L_4 + 2L_5 + 3W_3 + S_3$$

In this design, to obtain desired passband frequencies, we first determine $f_2$ by adjusting the length of the $L$ and then $f_1$ can be controlled simply by tuning the length of $(L_6 + L_7)$ and width $(W_2)$ of short-stub. As analyzed in [8] and [10], the voltage at the half-wavelength line center is zero at $f_1$. Hence, no signals at this frequency point can be delivered to the short-stub due to no effect on the coupling strength at $f_1$. Finally, a dual-band BPF with resonance frequencies at 2.4 GHz and 3.5 GHz and good selectivity is designed. The design parameters obtained by Ansoft HFSS 10.0 EM simulator are given as follows: $W_1 = 0.82\,\text{mm}, W_2 = 0.7\,\text{mm}, W_3 = 0.3\,\text{mm}, S_1 = 0.25\,\text{mm}, S_2 = 0.7\,\text{mm}, S_3 = 1\,\text{mm}, S_4 = 1\,\text{mm}, L_1 = 8.2\,\text{mm}, L_2 = 8.9\,\text{mm}, L_3 = 1.9\,\text{mm}, L_4 = 2.5\,\text{mm}, L_5 = 1.6\,\text{mm}, L_6 = 2.1\,\text{mm}, L_7 = 0.7\,\text{mm}, d = 0.5\,\text{mm}$. The fabricated dual-band BPF is compact with size of $0.1\lambda_g \times 0.22\lambda_g$, where $\lambda_g$ is the guided wavelength of the microstrip line at the center frequency of first band. Figure 3 illustrates the simulated frequency response, the resulting 3 dB fractional bandwidths (FBWs) for the two passbands centered at 2.4 GHz and 3.5 GHz are about $2.28 - 2.61$ GHz ($330MHz$) and $3.35 - 3.67$ GHz ($320MHz$), respectively. A prototype is fabricated to demonstrate the design strategies, as shown in Figure 4. For comparison, the measured results are also shown in Figure 3. The first passband is centered at 2.4 GHz, with the 3 dB
Figure 3. Measured and simulated results of the dual-band BPF with wide stop band.

Figure 4. Photograph of the fabricated dual-band BPF.

Bandwidth of 13.5% can meet the requirement of WLAN systems. The minimum insertion loss is 0.5 dB, while the return loss is 22 dB. The second passband is located at 3.5 GHz, meeting the application requirement of WiMAX. The 3 dB bandwidth is 9.1%, while the minimum insertion loss is 0.3 dB and the return loss within the passband is greater than 24.5 dB. The stopband rejection is better than 20 dB up to 9.77 GHz, namely $4f_1$, as shown in Figure 3.
3.2. Dual-band BPF for WLAN (2.4/5.2GHz) applications

Based on the above analysis in section 3.1. By tuning the length and width of short-stub and half-wavelength line, two center frequencies can be easily adjusted to the desired value. After an efficient optimisation process using EM simulator software, the design parameters obtained are given as follows: \( W_1 = 0.82\, \text{mm}, W_2 = 0.6\, \text{mm}, W_3 = 0.3\, \text{mm}, S_1 = 0.25\, \text{mm}, S_2 = 0.65\, \text{mm}, S_3 = 0.7\, \text{mm}, S_4 = 0.6\, \text{mm}, L_1 = 8.35\, \text{mm}, L_2 = 7.1\, \text{mm}, L_3 = 0.8\, \text{mm}, L_4 = 1.7\, \text{mm}, L_5 = 1.05\, \text{mm}, L_6 = 2.3\, \text{mm}, L_7 = 1.8\, \text{mm}, d = 0.5\, \text{mm}. \) In Figure 5, shown the measured results and simulation results. The two passbands are centered at 2.4 GHz and 5.2 GHz, meeting the requirement of WLAN systems. The 3 dB bandwidths are about \((2.18 - 2.62\, \text{GHz})\) 18.3\% and \((4.97 - 5.43\, \text{GHz})\) 8.8\%, respectively. The minimum insertion losses measured for the two passbands in the same sequence are 1.1, and 1 dB, while the return losses are 13.7, and 20.9 dB, respectively. In addition, two center frequencies can be adjusted to the value at 2.4/5.2 GHz and fractional bandwidth of the first passband varies from 13.5\% (in section 3.1) to 18.3\%. However, the stop band only about 6 – 8 GHz. In here the slight frequency discrepancy may probably be caused by unexpected fabrication tolerance and measurement error.

3.3. Dual-band BPF for WLAN (2.4/5.2GHz) applications with good selectivity

In section 3.2, a dual-band for WLAN (2.4/5.2GHz) application has proposed. However, on both sides of the two passbands only has two transmission zeros, thus the selectivity of
Figure 6. Layout proposed filter using short-circuited stub loaded half-wavelength open loop resonator with good selectivity.

filter needs to be improved. In Figure 6, shows the proposed two-order dual-band (BPF) using an open loop resonator connected with a short-stub loaded in the midpoint. Two short circuited stubs are bent so that they can be connected by a common via ground hole, thus magnetic coupling also can be realized between two resonators at the first band frequency \( f_1 \), and its coupling coefficient \( k_1 \) is determined mainly by the length \( L_5 \) and the characteristic impedance of the common short circuited stub. The coupling strength between two open loops at the first band frequency is very weak and only has a little impact on \( k_1 \). Obviously, two coupling coefficients \( k_1 \) and \( k_2 \) can be regulated independently without affecting each other [10] and [12]. It should be noted that the design parameters must be chosen properly so that the central frequencies of two passbands can be placed at 2.4 GHz and 5.2 GHz rightly. Therefore, in the design procedure, the coupling spacing \( S_1 \) should be firstly chosen according to the desired coupling coefficient \( k_1 \) and \( k_2 \). Figure 7 shows the simulated response of the dual-band filter with respect to different values of source-load coupling space \( S_1 \). It is noted that all these transmission zeros will move close to their passbands with the decrease of \( S_1 \). Therefore, the selectivity is improved at the cost of stopband rejection level. A dual-band BPF with resonance frequencies at 2.4/5.2 GHz is presented. Furthermore, new coupling scheme is introduced in the filter configuration to produce four transmission zeros at the adjacent of two passbands. Thus, the bandwidths of this filter are controllable and the filter has good selectivity. In Figure 8, the resulting 3-dB fractional bandwidths (FBWs) for the dual passbands centered at 2.4 and 5.2 GHz are about 2.33–2.61 GHz (280 MHz), 5.1–5.29 GHz (190 MHz), respectively. The design parameters obtained 10.0 EM simulator are given as follows: \( W = 0.82 \, mm, W_1 = 0.6 \, mm, W_2 = 0.25 \, mm, S_1 = 0.45 \, mm, S_2 = 0.25 \, mm, S_3 = 0.2 \, mm, L_1 = 5.38 \, mm, L_2 = 2.3 \, mm, L_3 = 1.5 \, mm, L_4 = 0.75 \, mm, L_5 = 0.8 \, mm, d = 0.5 \, mm \). The fabricated dual-band BPF is compact with size of \( 0.09 \lambda_g \times 0.24 \lambda_g \) (4.2 mm \( \times \) 11.3 mm), where \( \lambda_g \) is the guided wavelength of the microstrip line at the center frequency of first band. Thus, it is very compact.
Figure 7. Simulated frequency response of the filter under different values of the $S_1$

Figure 8. Measured and simulated results of the dual-band BPF.
The simulation and measurement are accomplished by using Ansoft HFSS v.10.0 EM simulator and 8753ES network analyzer, respectively. Figure 8 depicts the measured and simulated results. Two passbands for WLAN applications are designed. The first passband is centered at 2.4 GHz, with the 3 dB bandwidth of 11.6%. The minimum insertion losses is 0.3 dB, while the return losses is 18 dB.

The second passband is located at 5.2 GHz. The 3 dB bandwidth is 3.6%. The minimum insertion loss is 0.3 dB. The return loss within the passband is greater than 18.8 dB.

Four transmission zeros are created at 1.55, 2.87, 4.89 and 6.56 GHz. They are near the two passbands edges and significantly enhance the rates of roll-off. Figure 9 shows the photograph of the fabricated dual-band BPF.

4. Conclusion

In this work, three compact dual-band bandpass filters are proposed and demonstrated, they are implemented with short-stub loaded half-wavelength resonator. Two passbands for WLAN and WiMAX can be realized between the transmission zeros adjusting by short stub loaded half-wavelength resonator and new couple-line structures. Good agreement is observed between the experiments and theoretical analysis, indicating the validity of the proposed design strategies.

References


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