A Compact Planar Dual-Band Bandpass Filter with Asymmetric Bandwidths

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Abstract

Two different lengths of open-loop rectangle-ring resonators were parallelly positioned at two sides of input/output microstrip lines with the same coupling gap and length. The longer open-loop rectangle-ring was designed to resonate at 1.23 GHz and 2.31 GHz. The shorter open-loop rectangle-ring was designed to resonant at 2.52 GHz to improve the bandwidth of 2.4 GHz band. The proposed filter revealed great filtering properties and was suitable to used for the applications of GPS (L2-band, 1.227 GHz), and WLAN (IEEE802.11b/g, 2.4~2.4835 GHz).

1. Introduction

Microstrip planar filters has found wide applications in microwave circuits because of low profile, compact size, easily fabricated, low cost, and easy integration [1]. Many modified methods are investigated to improve the filtering characteristics. Moreover, multi-band bandpass filters with compact size, wide-bandwidth, and high out-of-band rejection are essential requirements for the rapid advancement of modern wireless communication system [2]. In the past, the single-stage multi-band bandpass filters had the shortcomings of and narrow bandwidths and inadequate out-of-band rejections, which can not be used for commercial applications. In order to solve these problems, filters need to cascade more resonators to get wider bandwidth and larger out-of-band rejection. The insertion losses and the pattern sizes of designed filters are greatly increased and that will limit their utility. Besides, the requirements of many modern wireless communication protocols are quit different, especially in bandwidth. The cascaded method usually increases the bandwidths of all operating frequencies, and it is difficult to control the bandwidths of each passband [3]. Some bandwidths of passbands are too large for its narrow band protocol, for example, Global Position System (GPS) and Radio Frequency Identification (RFID), and that will bring lots of unwanted noise signal [4].

In this study, a compact microstrip planar dual-band bandpass filter was designed and fabricated on the Al_2O_3 substrate. The higher dielectric constant Al_2O_3 substrate (ε_r=9.4) is used for pattern minimization. At first, the proposed filter is based on a longer open-loop rectangle-ring resonator with the dual resonant frequencies of 1.23GHz and 2.31GHz. Another shorter open-loop rectangle-ring resonator is designed to resonate at 2.52GHz and is parallelly positioned at another side of input/output microstrip lines to improve the bandwidth of 2.4GHz band. The different resonant current distributions in two resonators are investigated to demonstrate the proposed theorem. The proposed filter has two operating frequencies with deep transmission zeros and it can be used for the GPS (L2-band, 1.227 GHz) and WLAN (IEEE802.11b/g/n, 2.4~ 2.4835GHz) dual-band.

2. Design of Planar Dual-Band Bandpass Filter

The dual-band bandpass filter was based on the structure of $\lambda_g/2$ open-loop rectangle-ring resonator (Resonator L) shown in Fig. 1. At first, we used 1.23 GHz as the fundamental resonant frequency and 2.31 GHz for the first spurious resonant frequency, because they could be used for GPS-L2-band and WLAN, respectively. According to the dielectric constant of Al_2O_3 ceramic (ε_r=9.4), the ideal fundamental resonant length could be determined by [5]:

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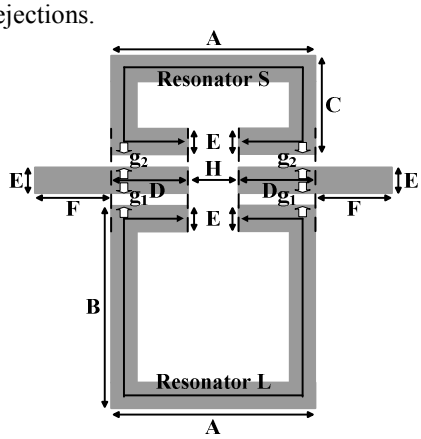
$$V = \frac{C}{\sqrt{\varepsilon_r}} = f\lambda_g \tag{1}$$

The calculated value of 1.23GHz for $\lambda_g/2$ open-loop microstrip line resonator path was about 39.78mm (abbreviated as Resonator L). The optimal coupling parameters between input/output microstrip lines and resonators were 0.1mm gap and 4mm length. The simulated property of Resonator L is showed in Fig. 2. However, even the bandwidth of 1.23GHz-band was adequate for GPS-L2-band protocol, the bandwidth of 2.31GHz was inadequate for the WLAN application. In order to improve the bandwidth of 2.4GHz band independently, another resonator resonated closely to 2.4GHz was needed. A shorter open-loop rectangle-ring resonator (Resonator S) was parallelly positioned at another side of the input/output microstrip lines with the same coupling conditions to the original longer resonator. The simulated property of Resonator S is also showed in Fig. 2. According to the reference, if phase difference for signals in two paths was equal to 180°, the energy cancellation was happened [6-7]. By suitable tuning, the signal could be coupled to different resonators and produced several resonant paths between the input/output transmission lines to enhance the bandwidths and produce the deep transmission zeros in the out-of-band rejections.



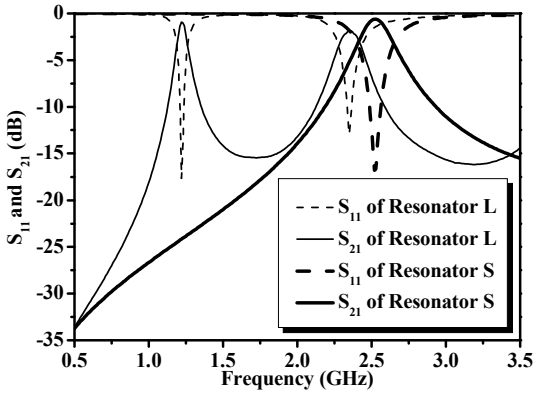
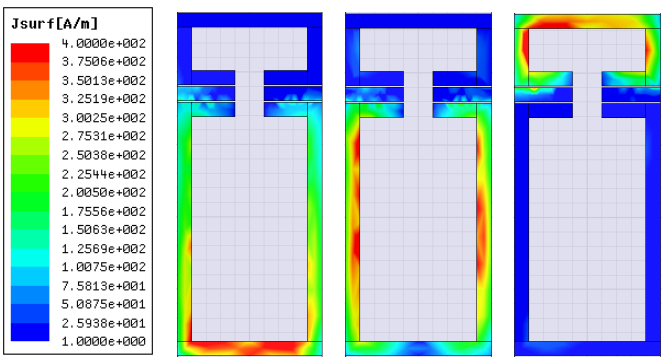


Fig. 1 The proposed dual-band bandpass filter. **Fig. 2** The simulated frequency responses of two different resonators.

Even some Rogers Duroid printed boards had the dielectric constant of 10.2 and low dielectric loss, it was indeed too expensive for commercial applications. The Al_2O_3 ceramic substrates had a comparably higher dielectric constant of 9.4, a smaller dielectric loss less than 0.001, and low price. For those reasons, the filter was built on 1 mm thickness Al_2O_3 ceramic substrates for pattern minimization. The low environmental pollution of printing method was used to fabricate the proposed dual-band bandpass filter. The printing method did not need using the $FeCl_3$ solution to etch the Cu plate from the surface of Duroid or FR4 substrates [1]. The Ansoft HFSS was used to adjust and optimize the pattern parameters. The printing mask was done according to the designed patterns and used to print the Ag/Pd paste on the Al_2O_3 ceramic substrates. After printing, the pattern was sintered in an oven at 700°C for 15min. Finally, two SMA connectors were welded as the input/output of the fabricated filters, and the microwave properties were measured by the network analyzer (Agilent-N5230A).

3. Simulated and Measured Results



(a) (b) (c)

Fig. 3 The simulated resonant current distributions of two resonators at (a) 1.23 GHz, (b) 2.31 GHz, and (c) at 2.52 GHz.

Fig. 3 shows the simulated current distributions of proposed bandpass filter. For the first resonant mode, as Fig. 3(a) shows, two zero points were present in the front/end edges and the maximum current density was present in the middle of resonator. It suggests that the length of Resonator L was equal to the half-guided-wavelength route of fundamental response at 1.227GHz. Fig. 3(b) shows that three zero points and two maximum current density zones were distributed in Resonator L. It means that the length of Resonator L was equal to the full-guided-wavelength route of the first spurious response at 2.31GHz. The current distribution of Resonator S in Fig. 3(c) presented the typical $\lambda_g/2$ fundamental response at 2.52GHz. Fig.4 shows the voltage phase diagrams of two resonators at the two resonant frequencies. The bandwidth of 1.227GHz band was produced by $\lambda_g/2$ fundamental response of Resonator L. The bandwidth of 2.4GHz band was produced by combining $1\lambda_g$ first spurious response of Resonator L and $\lambda_g/2$ fundamental response of Resonator S. The structure with 180° phase difference had designed to achieve deep transmission zeros and wider bandwidth at 2.4GHz.

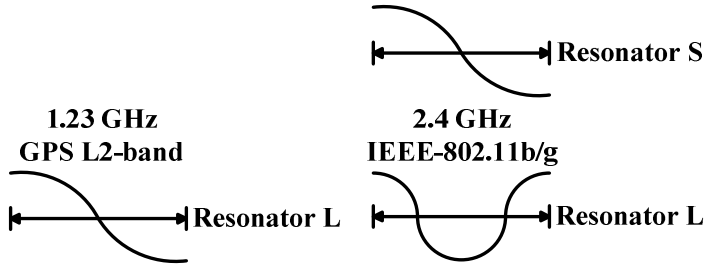
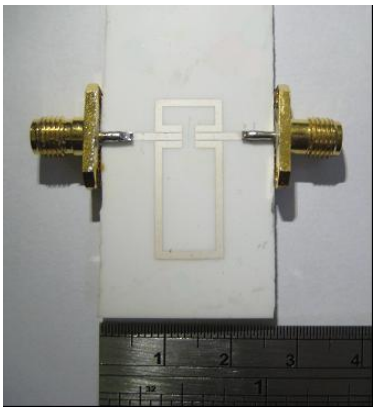


Fig. 4 The voltage phase diagrams of two resonators.

The 2.4GHz-band bandwidth of designed filter could be tuned by adjusting the length of Resonator S. Table 1 shows the resonant properties of proposed filter against different C value. As the C value increased from 5mm to 6mm, the responses of Resonator L had no apparent change, but the resonant frequency of Resonator S would shift to lower value. The bandwidth of 1.23GHz-band was almost unchanged and that of 2.4GHz-band decreased. Although the bandwidth in C= 5mm was wider than that in C= 5.5mm, the unwanted ripple would become seriously in the passband. If C was equal to 6mm, the resonant frequencies of two resonators were two closed, the cancellation in energy was not happened and no passband could be found around 2.4GHz. As a result, C=5.5mm was the optimal value in this study.

Table 1 Resonant properties of proposed filter against different C value.

C value	5	5.5	6
f_0 of Resonator L	1.23GHz	1.23GHz	1.23GHz
f_{s1} of Resonator L	2.31GHz	2.31GHz	2.31GHz
f_0 of Resonator S	2.61GHz	2.52GHz	2.39 GHz
BW at 1.23 GHz	60 MHz	60 MHz	60 MHz
BW at 2.4 GHz	280 MHz	240 MHz	--



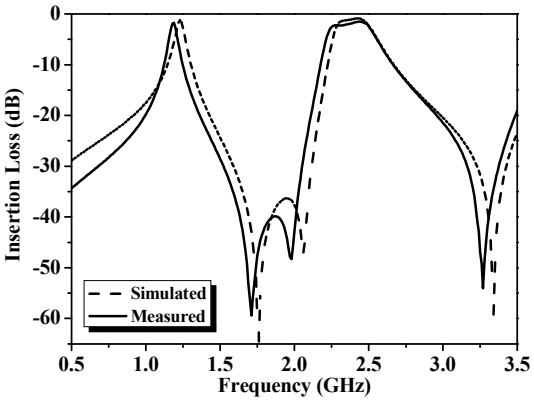


Fig. 5 The photographic of fabricated filter. **Fig. 6** The simulated and measured results of the proposed filter.

The optimal parameters of the proposed filter are A = 10mm, B=15.5mm, C=5mm, D=4mm, E=1mm, F=5mm, H=2mm, and g1=g2=0.1mm. Fig.5 is the photographic of the fabricated filter and Fig. 6 shows the simulated and measured results of proposed filter. The simulated results are S_{21}= 1.13dB, BW= 60MHz (4.88%) at 1.23 GHz and S_{21}=0.91 dB, BW= 260MHz (10.74%) at 2.42 GHz. The measured results are S_{21}=1.77 dB, BW= 60MHz (4.96%) at 1.21 GHz and S_{21}=1.96dB, BW= 310 MHz (12.55%) at 2.45GHz. The simulated transmission zeros are 64.55dB at

1.76GHz, 46.97dB at 2.05GHz, and 59.22dB at 3.34GHz. The measured transmission zeros are 59.39dB at 1.71GHz, 48.31dB at 1.98GHz, and 54.08dB at 3.32GHz, respectively. The fabricated filter was able to use in GPS-L2-band and WLAN systems.

4. Conclusion

A compact microstrip planar dual-band bandpass filter was achieved by two different lengths of open-loop rectangle-ring resonators. Two resonators had revealed even mode and odd mode with 180° phase difference around 2.4GHz and presented wide operating bandwidth and large out-of-band rejection. The measured results of the printed filter were S_{21}= 1.77dB, BW= 60MHz (4.96%) at 1.21GHz and S_{21}= 1.96dB, BW= 310MHz (12.55%) at 2.45GHz. The measured transmission zeros were 59.39dB at 1.71GHz, 48.31dB at 1.98GHz, and 54.08dB at 3.32GHz. The proposed filter revealed great filtering properties of GPS narrow band and WLAN wide band and deep transmission zeros.

5. Acknowledgments

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