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Novel High-selectivity Dual-band Substrate Integrated Waveguide Filter with Multi-transmission Zeros

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Abstract—A novel dual-band filter based on substrate integrated waveguide (SIW) is presented in this paper. The proposed filter is composed of two filters with different center frequencies and bandwidths, where they share the input and output ports with source-load coupling using rectangular SIW cavity structure. Multi-transmission zeros have been obtained through electrical coupling between the source and load, which improves the frequency-selective characteristics of the filter greatly. Finally, a Ku-band substrate integrated waveguide dual-band filter with bandwidths of 220 MHz and 120 MHz was finally designed, fabricated, and measured. The measurement results are found to be in good agreement with the simulation results.

1. INTRODUCTION

With the rapid development of dual- and multi-band wireless communication, dual-band microwave filter plays an important role in wireless communication systems. So far, much work has been done to develop a variety of dual-band filters. There are many methods to design dual-band filters. In [1, 2], the dual-band filter was constructed by cascading a broadband filter and a bandstop filter. Unfortunately, this type of dual-band filter may suffer from high insertion loss and large size. Dual-band filters can be realized by introduction of transmission zeros among a passband filter [3, 4] or by employing dual-resonance structures/step impedance resonator (SIR) [5–7], frequency transformation [8] etc..

Substrate integrated waveguide (SIW) has attracted considerable research interest in the field of dual-mode filters [11], half mode filter [12] or dual-band filter [13]. Compared with the traditional microstrip or coaxial cavities [9, 10], SIW takes the advantage of both waveguide and microstrip structures, such as high-Q factor, high power capacity, small size, low loss, and the possibility of integration. In Ref. [14], three K-band multi-passband filters with low loss using SIW structure, were easily realized by the synthesis of frequency transformation, but this method is only limited to design symmetrical dual-band filters. In Refs. [15, 16], miniaturized multi-band SIW filters were proposed by loading complementary open-loop resonator (CSRR). Although these methods are provided to deal with the problem of circuit size, they also cause some disadvantages, such as, insufficient stopband attenuation in lower frequency, and few design freedoms.

In this article, a novel dual-band filter is proposed by combining two SIW filters with source-load coupling [17] to improve the selectivity and achieve a high out-of-band rejection, where their bandwidth and order can be flexibly controlled. The electrical coupling is realized by introducing serpentine slot structure between the source and load, thereby generating four transmission zeros in finite frequency, which further improve frequency-selectivity characteristics of the filter. In addition, owing to folding layout, the circuit size will be reduced greatly compared with the conventional filter. Finally, the proposed filter is optimally designed, fabricated, and measured. Good agreement between simulated and measured results is achieved, which verifies the proposed design method.

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2. FILTER DESIGN

2.1. Principle of the Dual-passband Filter

Figure 1(a) shows the design schematic of the dual-band filter, and its frequency response is shown in Figure 1(b), where filter 1 and filter 2 are passband filter with different center frequencies, but they share the input-output coupling port. The two filters can be individually designed, which indicates that their order and bandwidth can be different. So this method can be widely used to design symmetric and asymmetric dual-band filters.

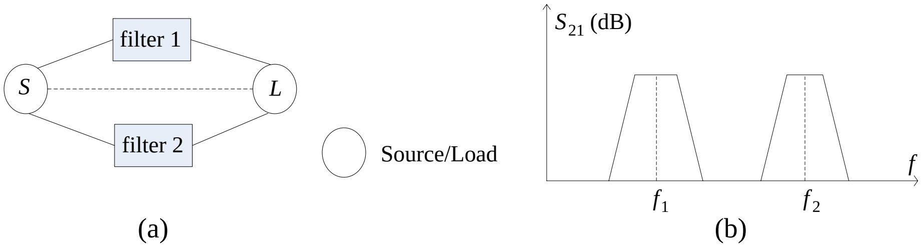


Figure 1. (a) The design schematic of the dual-band filter and (b) its frequency response.

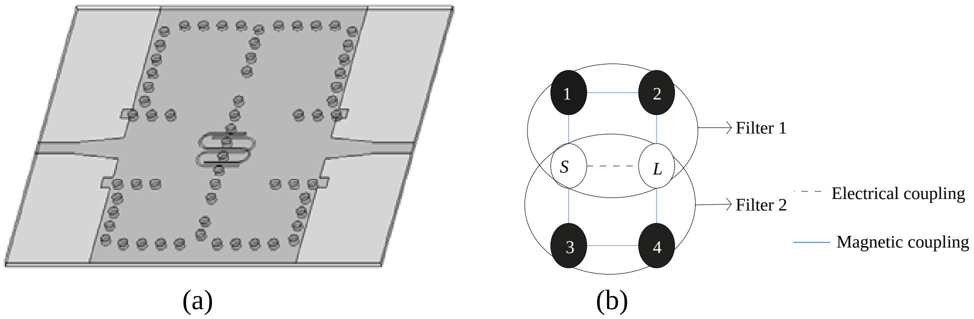


Figure 2. (a) The geometric configuration of the dual-band filter and (b) its schematic topology (Source: S, load: L).

For a detailed description of the design process based on the above design principle, a novel four-order Ku-band SIW dual-band filter is designed by using commercial full-wave electromagnetic simulator HFSS. Figure 2 shows the geometric configuration and schematic topology of the proposed SIW dual-band filter, where the black nodes 1, 2, 3, 4 represent the rectangular substrate integrated waveguide cavity resonator, respectively. The hollow nodes represent the source and load. Filter 1 and filter 2 sharing the input-output coupling structure can be designed independently. To obtain better stopband performance the electrical coupling is realized by etching complementary serpentine slot structure on the top and bottom metal surfaces of the shared coupling wall between the source and load. The other couplings between two SIW cavity resonators are obtained by the shared post-wall irises, which provide magnetic coupling. According to the cross-coupling theory [18], the source and load can be regarded as resonators, so the topologies of filter 1 and filter 2 can produce two transmission zeros in its high and low stopbands, respectively. Therefore, a novel SIW dual-band filter with four transmission zeros can be obtained by combining two filters, where two transmission zeros locate between two passbands, the other two appear in the upper and lower stopbands, respectively.

2.2. Coupling Matrix

A dual-band (12.98/15.44 GHz) BPF with 220 MHz (lower passband) and 120 MHz (upper passband) 3-dB bandwidths was designed. According to the synthesis method presented in [18], we can get the coupling matrix of each passband, as follows:

$$k_{\text{filter1}} = \begin{bmatrix} 0 & 0.0207 & 0 & -0.0002 \\ 0.0207 & 0 & 0.0286 & 0 \\ 0 & 0.0286 & 0 & 0.0207 \\ -0.0002 & 0 & 0.0207 & 0 \end{bmatrix} \tag{1}$$

$$k_{\text{filter2}} = \begin{bmatrix} 0 & 0.0094 & 0 & -0.0005 \\ 0.0094 & 0 & 0.0124 & 0 \\ 0 & 0.0124 & 0 & 0.0094 \\ -0.0005 & 0 & 0.0094 & 0 \end{bmatrix} \tag{2}$$

2.3. Design and Simulation

As shown in Figure 2(a), the electrical-coupling strength between the source and load is controlled by the length of the serpentine slot structure, while all of the remaining magnetic-coupling strengths between two SIW cavity resonators are determined by the width of post-wall iris. The input/output ports located at the top layer are excited by 50-microstrip lines with tapered microstrip line to provide a smooth impedance matching. Furthermore, each passband can be controlled independently by the dimensions of the presented filter 1 and filter 2. The design procedure can be summarized as follows:

First, the initial size of TE_{101}-mode-based SIW cavity of two filters can be determined by setting the resonant frequencies to the center frequencies using the following formula [19]:

$$f_0 = \frac{c_0}{2\sqrt{\varepsilon_r}}\sqrt{\frac{1}{a_{\mathit{eff}}^2} + \frac{1}{b_{\mathit{eff}}^2}} \tag{3}$$

where

$$a_{\mathit{eff}} = a - \frac{d^2}{0.95p} \quad b_{\mathit{eff}} = b - \frac{d^2}{0.95p} \tag{4}$$

a and b are the width and length of the TE_{101} SIW cavity, respectively. d and p are the diameter of metallized via-holes and center-to-center pitch between two adjacent via-holes, c_0 is the light velocity in vacuum, and ε_r is the dielectric constant of the substrate.

To determine the coupling coefficients of the two filters, a pair of coupled TE_{101} SIW cavity resonators with two resonant peaks f_e and f_m, can be obtained by eigenmode simulation in HFSS. Then the coupling coefficient between two coupled resonators can be extracted by using the following relation [20]:

$$k = \pm\left(f_e^2 - f_m^2\right)/\left(f_e^2 + f_m^2\right) \tag{5}$$

According to the above formulas, all the initial sizes of resonant units and coupling structure of the filters can be obtained by using HFSS. Figure 3 shows the full simulation models of two filters built in HFSS, while Figure 4 shows the optimized simulation response curves.

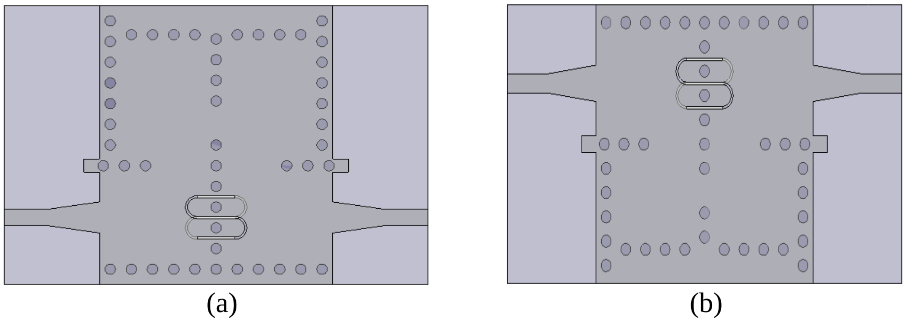
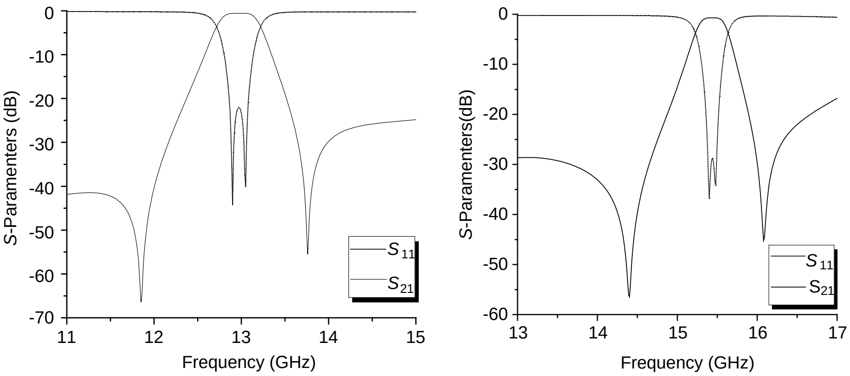

(a) (b)

Figure 3. Full-wave simulation models of (a) filter 1, (b) filter 2.



(a) (b)

Figure 4. The simulated response curves of (a) filter 1, (b) filter 2.

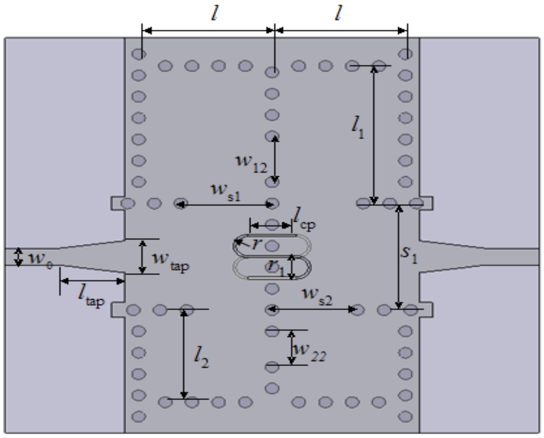


Figure 5. Plane geometry structure of the dual-band filter.

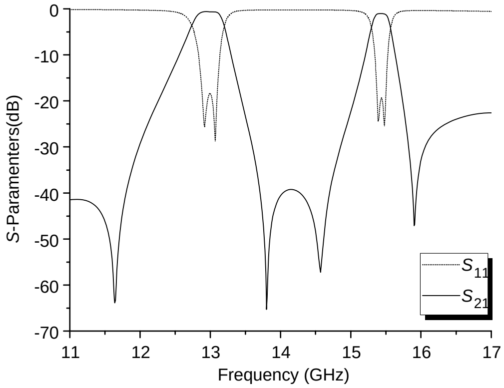
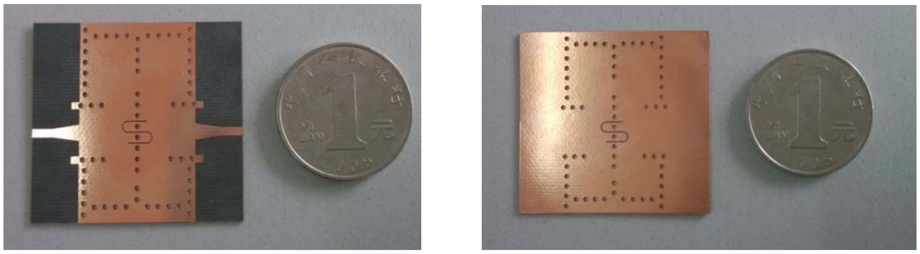


Figure 6. Simulated response for the combined filter.

Finally, the entire full simulation model of the proposed dual-band filter is carried out by combining filter 1 with filter 2. Figure 5 shows the plane geometry layout of the simulation model. By adjusting the length of the serpentine slots (l_{cp}) between source and load, the positions of the filter's transmission zeros varies. The larger l_{cp} is, the stronger the source-load coupling becomes, and transmission zeros are closer to passbands, leading to better stop-band suppressions, but the in-band attenuation will be worse. Considering the out-of-band suppression and in-band attenuation, the final length of l_{cp} is 3.6 mm. Figure 6 shows the simulated frequency responses of the dual-band filter. It is seen that the SIW dual-band filter operating at 12.87–13.09 GHz, and 15.38–15.50 GHz, whose in-band insertion losses are less than 1 dB, and the in-band return losses are below -19 dB. Meanwhile, the filter can produce four transmission zeros at about 11.65, 13.80 14.57, and 15.90 GHz, respectively. The attenuations at 11.65 and 13.80 GHz exceed 60 dB. Moreover, we can obtain good stopband characteristic with greater than 40 dB stopband from 13.71 to 14.71 GHz. The final optimized dimensions of the proposed filter are chosen as follows: $l = 10\,\text{mm}$, $l_1 = 12.88\,\text{mm}$, $l_2 = 8.64\,\text{mm}$, $l_{cp} = 3.6\,\text{mm}$, $S_1 = 10\,\text{mm}$, $W_0 = 1.56\,\text{mm}$, $W_{tap} = 5\,\text{mm}$, $l_{tap} = 3\,\text{mm}$, $W_{s1} = 6.82\,\text{mm}$, $W_{s2} = 6.39\,\text{mm}$, $W_{12} = 4.28\,\text{mm}$, $W_{22} = 3.36\,\text{mm}$, $r_1 = 2r = 1.2\,\text{mm}$.

3. FABRICATION AND MEASUREMENTS

Based on above design results, the proposed dual-passband substrate integrated waveguide filter is implemented on the substrate Rogers RT/Duroid 5880 with dielectric constant ε_r of 2.2, thickness



(a) (b)

Figure 7. Photograph of the fabricated dual-band filter, (a) top view, (b) bottom view.

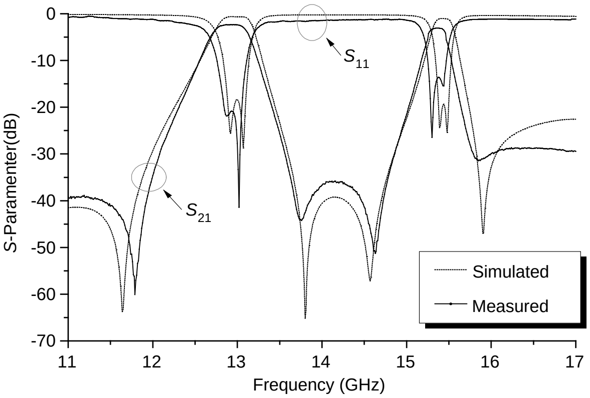


Figure 8. Comparison between the simulated and measured results.

of 0.508 mm, and loss tangent of 0.0009. Figure 7 shows a photograph of the fabricated dual-band BPF with the chip size of 40 mm × 37 mm × 0.508 mm. The fabricated dual-band BPF is measured by Agilent 8722ES network analyzer and Anritsu Wiltron 3680-K test fixture. The simulated and measured frequency responses are shown in Figure 8. It can be seen that the measured in-band return losses are below -15 dB with minimum insertion loss of 2.2 dB in the first pass-band and 2.6 dB in the second pass-band. As shown in Figure 8, there is a bit deviation of the measured result of the insertion loss level compared to that of the simulation. This may be due to measurement or fabrication error. Good selectivity and out-of-band rejection are achieved due to four transmission zeros lacated at 11.78, 13.76, 14.61, and 15.87 GHz, respectively. The measured results are in good agreement with the simulated ones, except that there is a slight frequency shift of approximately 0.3%.

4. CONCLUSION

In this article, a novel dual-band filter at 12.98/15.44 GHz is designed and fabricated using rectangular substrate integrated waveguide resonators. The performance of the presented dual-band filter has been demonstrated by using the simulated and measured results of the fabricated filter. Four transmission zeros by introducing electrical coupling between the source and load are realized, resulting in high selectivity. The simulated and measured results show that the proposed dual-band filter has the advantage of dual-passband high-frequency selectivity and good stop-band suppression characteristic. Excellent agreement between the measurement and simulation is obtained.

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