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A highly independent and controllable dual-band bandpass filter based on source-load coupling with stub-block isolation structure

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Abstract :

The main problem of dual-band bandpass filter (BPF) structures is to control each passband performance individually, separately, and independently. This Letter is proposed a dual-band BPF based on a source-load coupling structure with stub-block isolation to overcome the problem. The lower band resonator structure is placed on the top side, while the upper band resonator is placed at the bottom side, with the source-load (SL) coupling structure in the middle. An additional stub-block isolation structure is added to the center of the SL coupling structure. As a result, we have successfully designed an independent dual-band bandpass filter with highly controllable working frequency/ frequency center (f_c), bandwidth (BW), reflection coefficient (S_{11}), and isolation (ISO) between the passbands. The proposed dual-band BPF was fabricated on an RT/Duroid 5880 substrate. Furthermore, this dual-band BPF achieved an insertion loss/fractional bandwidth of 0.48 dB/7.71% and 0.35 dB/12.37% at 1.82 GHz and 2.58 GHz, respectively. The good agreement between the simulated and measured results validates the proposed method.

Keywords: Controllable dual-band BPF, source-load coupling, stub-block isolation

1. Introduction

A highly flexible RF device must be supported by a high-performance bandpass filter (BPF) that can be controlled. This requirement has motivated many researchers to produce BPFs with controllable performance [1]. Several methods have been proposed to control the frequency passband, such as the stepped impedance ring resonator (SIRR) with shorted stubs [2], defected and irregular stepped-impedance resonators (DI-SIRs) [3], multilayer resonator [4], loop resonator [5], cross resonator [6], substrate-integrated waveguide (SIW) cavities [7], and half-mode substrate integrated waveguide (HMSIW) [8]. Furthermore, to increase the isolation, some researchers have proposed a circular resonator [9] and a ring resonator. Moreover, a quasi-elliptical waveguide resonator was proposed by [10] to control the frequency and bandwidth. However, none of the proposed methods has a controllable performance frequency, bandwidth, reflection coefficient, and isolation simultaneously.

In this Letter, a dual-band BPF based on a source-load coupling structure is proposed, as shown in Fig. 1(a). It is clearly distinct from the microstrip structure used in [1–11]. The topology of the coupling structure is given in Fig. 1(b). Furthermore, M_{MN} denotes the coupling matrix values between two resonators for (*M=S, 1, 2, L* and *N=S, 1, 2, L*), can be derived as follows;

$$M_{MN} = \begin{pmatrix} 0 & 0.307 & 0 & -0.389 & 0 & 1.000 \\ 0.307 & 0.819 & 0.080 & 0 & 0 & 0 \\ 0 & 0.080 & -0.579 & 0 & 0 & 0.008 \\ -0.389 & 0 & 0 & -25.95 & -66.122 & 0 \\ 0 & 0 & 0 & -66.122 & -169.205 & -0.116 \\ 1.000 & 0 & 0.008 & 0 & -0.116 & 0 \end{pmatrix}$$

The coefficients of the coupling matrix are taken from the optimization process. By using this structure, the frequency, bandwidth, reflection coefficient, and isolation in each passband can be adjusted individually with convenience and robustness. The proposed method is validated by the good agreement between the simulated and measured results.

2. Dual-band BPF based on source-load coupling with stub-block isolation

The proposed dual-band BPF is constructed by using four important segments, i.e., a source-load coupling structure, a lower band resonator, an upper band resonator, and an additional isolation structure. The source-load (P_{IN} and P_{OUT}) coupling structure is positioned in the middle, where (W_1, L_1) represent the width and length, respectively. Furthermore, the lower band resonators are placed at the top, constructed by a coupled-resonator (R_{A1} and R_{A2}) with the back-to-back position, where (W_2, L_{2A}, L_{2B}) represent the widths and lengths of the lower band resonator. Moreover, the upper band resonators are arranged at the bottom. They are composed of a coupled-resonator (R_{B1} and R_{B2}) with a back-to-back position, where (W_3, L_{3A}, L_{3B}) represent the widths and lengths of the upper band resonator. The additional isolation is added at the center, with (W_4, L_4) representing the width and length isolation structures, respectively. Furthermore, (S_1, S_2, S_3, S_4) represent the gaps between the source/load and lower band, intercoupled lower band, intercoupled upper band, and source/load and upper band, respectively.

Fig. 1(c) shows the dual-band BPF response. If the lower band or upper band structures are applied separately, the transmission coefficient will respond separately. However, if they are combined, the interference between two passbands will increase. To reduce interference, the isolation structure should be added at the center. Furthermore, the odd-mode structure of lower band, even-mode structure of lower band, odd-mode structure of upper band, and even-mode structure of upper band are shown in Fig 2(a), 2(b), 2(c), and 2(d), respectively. The value input impedance of odd-mode at lower band $Z_{IN\text{-}LB\text{-}odd}$ can be derived:

$$Z_{IN(3)} = -jZ_3 \cot\theta_3 \quad (1)$$

$$Z_{IN(2)} = Z_2 \frac{Z_{IN(3)} + jZ_2 \tan\theta_2}{Z_2 + jZ_{IN(3)} \tan\theta_2} \quad (2)$$

$$Z_{IN-LB-odd} = Z_1 \frac{Z_{IN(2)} + jZ_1 \tan\theta_1}{Z_1 + jZ_{IN(2)} \tan\theta_1} \quad (3)$$

Moreover, equation (3) can also be expressed as:

$$Z_{IN-LB-odd} = Z_1 \frac{Z_2(-jZ_1 \cot\theta_1 + jZ_2 \tan\theta_2) + jZ_3 \tan\theta_3\,(Z_2 + Z_1 \cot\theta_1 \tan\theta_2)}{Z_3Z_2 + Z_3Z_1 \cot\theta_1 \tan\theta_2 + Z_2Z_1 \cot\theta_1 \tan\theta_1 - {Z_2}^2 \tan\theta_2 \tan\theta_3} \quad (4)$$

with

$$Z_2 = \frac{Z_{2,e} + Z_{2,o}}{2} \quad (5)$$

The resonant can be derived from admittance condition $Y_{IN-LB-odd} = 0$ or impedance condition $Z_{IN-LB-odd} = \infty$ [1], or it has a denominator equal with zero.

$$Z_3\left(\frac{Z_{2,e} + Z_{2,o}}{2}\right) + Z_3Z_1 \cot\theta_1 \tan\theta_2 + Z_1\left(\frac{Z_{2,e} + Z_{2,o}}{2}\right)\cot\theta_1 \tan\theta_1 - \left(\frac{Z_{2,e} + Z_{2,o}}{2}\right)^2 \tan\theta_2 \tan\theta_3 = 0 \quad (6)$$

Furthermore, the value input impedance of even-mode at lower band $Z_{IN\text{-}LB\text{-}even}$ can be derived:

$$Z_{IN-LB-even} = jZ_1 \tan\theta_1 \quad (7)$$

Moreover, the value input impedance of odd-mode at upper band $Z_{IN\text{-}UB\text{-}odd}$ can be derived:

$$Z_{IN(6)} = -jZ_6 \cot\theta_6 \quad (8)$$

$$Z_{IN(5)} = Z_5 \frac{Z_{IN(6)} + jZ_5 \tan\theta_5}{Z_5 + jZ_{IN(6)} \tan\theta_5} \quad (9)$$

$$Z_{IN-UB-odd} = Z_4 \frac{Z_{IN(5)} + jZ_4 \tan\theta_4}{Z_4 + jZ_{IN(5)} \tan\theta_4} \quad (10)$$

Equation (10) can also be derived as:

$$Z_{IN-UB-odd} = Z_4 \frac{Z_5(-jZ_4 \cot\theta_4 + jZ_5 \tan\theta_5) + jZ_6 \tan\theta_6\,(Z_5 + Z_4 \cot\theta_4 \tan\theta_5)}{Z_6Z_5 + Z_6Z_4 \cot\theta_4 \tan\theta_5 + Z_5Z_4 \cot\theta_4 \tan\theta_4 - {Z_5}^2 \tan\theta_5 \tan\theta_6} \quad (11)$$

with

$$Z_5 = \frac{Z_{5,e} + Z_{5,o}}{2} \quad (16)$$

The resonant can be derived from admittance condition $Y_{IN-UB-odd} = 0$ or impedance condition $Z_{IN-UB-odd} = \infty$ [1], or it has a denominator equal with zero.

$$Z_5\left(\frac{Z_{5,e}+Z_{5,o}}{2}\right)+Z_6Z_4\cot\theta_4\tan\theta_5+Z_4\left(\frac{Z_{5,e}+Z_{5,o}}{2}\right)\cot\theta_4\tan\theta_4$$

$$-\left(\frac{Z_{5,e}+Z_{5,o}}{2}\right)^2\tan\theta_5\tan\theta_6 = 0 \quad (17)$$

Furthermore, the value input impedance of even-mode at upper band $Z_{IN\text{-}LB\text{-}even}$ can be derived:

$$Z_{IN-UB-even} = jZ_4\tan\theta_1 \quad (18)$$

with the impedance (Z_N) and electric length (θ_N).

3. Result and discussion

Figs 3(a) and 3(b) show the relationship between the bandpass frequency /frequency center of the lower band response under various lengths L_{2B} and the bandpass frequency/frequency center of the upper band response under various lengths L_{3B}, respectively. The figures show that by increasing the dimension of L_{2B}, the bandpass frequency of the lower band will gradually shift to a lower frequency, while the upper band will remain stable. Moreover, the bandpass frequency of the upper band will be shifted by various lengths L_{3B}, while the lower band will remain stable.

Moreover, Figs 3(c) and 3(d) show the relationship the bandwidth characteristics of the lower band for various gaps S_2, and the bandwidth characteristics of the upper band for various gaps S_3, respectively. Moreover, the bandwidth of each passband can be controlled individually and separately by varying the gaps S_2 and S_3. It can be seen that by increasing the gap S_2, the bandwidth of the lower band will become narrower and can be adjusted separately. Furthermore, by decreasing gap S_3, the bandwidth of the upper band only will increase.

Figs. 4(a), 4(b), and 4(c) show the reflection coefficient characteristics of the lower band for various gaps S_2, the reflection coefficient characteristics of the upper band for various gaps S_3, and the isolation characteristics for various lengths L_4, respectively. The reflection coefficient value of the lower band can be controlled separately by varying gap S_2 without any impact on the upper band. Moreover, the reflection coefficient value of the upper band can be adjusted individually by varying gap S_3. Furthermore, the isolation characteristics can be changed by varying the length L_4 without affecting the frequency passband or bandwidth of the lower band and upper band. Moreover, Figs 5(a) and 5(b) show the current surface at lower-band of f_c = 1.82 GHz and upper band of f_c = 2.58 GHz, respectively. It can be seen that at the lower band, the surface current flows at upper part of BPF. Meanwhile, the surface current flows at lower part of BPF at the upper band.

The proposed dual-band BPF was fabricated on an RT/Duroid 5880 substrate with a permittivity of 2.2 and a thickness of 1.575 mm. A momentum simulation produced by the Advanced Design System (ADS) was used to optimize the structure. Furthermore, the R&S ZVA67 VNA was used to measure the BPF performance. The dimensions were as follows (all in millimetres): W_1 = 1.0, W_2 = 1.5, W_3 = 1.0, W_4 = 0.5, L_1 = 10, L_{2A} = 15, L_{2B} = 32, L_{3A} = 15, L_{3B} = 15, S_1 = 0.5, S_2 = 3.0, S_2 = 1.5, and S_4 = 0.5. The dual-band BPF insertion loss/fractional bandwidth was 0.48 dB/7.71% and 0.35 dB/12.37% at 1.82 GHz and 2.58 GHz, respectively. Figs. 5(c) and 5(d) show photographs of the fabricated dual-band BPF and comparisons of the simulated and measured results, respectively. Moreover, Table 1 shows comparison with some previous dual-band BPFs such as ref [11]–[19].

The proposed method is validated by the good agreement between the simulated and measured results. Furthermore, Table 1 gives the performance comparison of the dual-band BPF with some previous works, from which it can be deduced that the proposed BPF structure can enable adjustment of the frequency, bandwidth, reflection coefficient, and isolation of each passband individually with convenience and robustness.

4. Conclusions

We have successfully designed an independent dual-band bandpass filter with a highly controllable working frequency/ frequency center, bandwidth, reflection coefficient, and isolation between the passbands. This performance can be obtained by applying the source-load coupling with a stub-block isolation structure. The proposed dual-band BPF was fabricated on an RT/Duroid 5880 substrate. Furthermore, this dual-band BPF achieved an insertion loss/fractional bandwidth of 0.48 dB/7.71% and 0.35 dB/12.37% at 1.82 GHz and 2.58 GHz, respectively. The good agreement between the simulated and measured results validates the proposed method.

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