A Method for Improving Stopband Characteristics of a Dual-Band Filter

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Abstract

This paper presents a simple and effective method for improving stopband rejection characteristics of previously studied dual-band filters. Small electric couplings were applied to the symmetrically positioned shunt resonators, which divided each transmission zero into two transmission zeros without any effect on passbands. We were able to achieve improved stopband rejection characteristics by these additional transmission zeros. For the filter application, we designed a dual-band filter with improved stopband characteristics using microstrip quasi-lumped elements. The electric couplings that control the location of transmission zeros are controlled by the distance between symmetric open stubs of the filter. The filter was fabricated with a relative dielectric constant of 3.5 and a thickness of 0.76 mm. The fabricated filter has a small size (14.6×13.2×0.76 mm) and a low insertion loss when compared with conventional filters.

Key words: Dual-Band Filter, Electric-Coupling, Quasi-Lumped Elements, Transmission Zero.

I . Introduction

The development of modern communication systems requires that microwave components operate in either dual- or multi-bands. For this reason, dual-band filters have become key components in the RF front-end module. Several studies have focused on dual-band filters [1]~[9]. In [1], [2], dual-band filters were constructed by parallel connection of two filters operating at different frequency bands. In [3], a dual-band filter was achieved by cascading a wideband filter and a bandstop filter. However, these methods increase the overall circuit size and require extra impedance-matching circuits. In [4]~[6], dual-band filters were realized by a coupling matrix with multi-band responses. However, these solutions are performed through the cavity or waveguide structures. Microstrip cross-coupled structures were also proposed to realize the coupling matrix [7], but these filters showed high insertion losses, which resulted from the finite Q factor of resonators, and have a large size. For these reasons, we proposed an LC prototype for multi-band filters [8], [9]. This method has a simpler design process because it can be directly synthesized from the poles and zeros. In addition, the prototype filters can be realized by various methods such as planar or multi-layered structures. However, these filters have just one transmission zero in each stopband. The stopband characteristics have to be improved by adding additional transmission zeros.

Many methods have been reported for generating extra transmission zeros. In [10], a transmission zero was generated using input/output couplings. In [11], transmission zeros were generated using insert open stub structures in both input/output ports, but these methods increase the filter sizes.

In this paper, we proposed a simple and effective method for improving the stopband rejection characteristics of a previously studied dual-band filter. The suppression characteristics in the stopband can be used to generate transmission zeros at the stopband frequencies. Insertion of a small electric coupling onto the shunt resonators, which are symmetrically positioned in the lumped-element circuit, allows each transmission zero to be divided into two transmission zeros. The small electric coupling can divide the transmission zeros, with negligible effects on the passband performance. From this, we analytically show the effect of a small electric coupling on the location of transmission zeros.

We used the proposed method to design and fabricate a dual-band filter, which operates at the GSM (Global System for Mobile Telecommunication, 880~960 MHz) and ISM (Industrial, scientific and medical, 2,400~2,500 MHz) bands.

II . Equivalent Circuit Analysis

Fig. 1(a) shows the lumped-element circuit model of the dual-band filter that was previously studied in [8], [9]. This dual-band filter has two transmission zeros at the resonance frequencies of \( L_{d1}, C_{d1} \) and \( L_{d2}, C_{d2} \). Fig. 1(b) shows the proposed equivalent lumped-element cir-
circuit model with electrically coupled elements for improving stopband characteristics. We applied the small electric couplings \((C_{na} \text{ and } C_{no})\) to the symmetrically located resonators of the lumped-element circuit in Fig. 1(a). With the electric coupling, additional transmission zeros can be generated and stopband characteristics can be improved. We located the transmission zeros based on the network parameters \((Z, Y, S, \text{ or } ABCD)\) of the two-port microwave network shown in Fig. 1(b). \(S_{21}\), which represents the transmission coefficient from Port 1 to Port 2, is expressed by the admittance matrix [11]:

\[
S_{21} = -\frac{2Y_2Y_0}{(Y_{21} + Y_0)(Y_{22} + Y_0)} - Y_{12}Y_{21}
\]

where \(Y_0\) is characteristic admittance. The transmission zero generation condition is \(Y_{21} = 0\).

For the theoretical analysis of equivalent circuit, we calculated the \(ABCD\) matrix. The total network shown in Fig. 1(b) can be divided into an outer and inner network. The outer network is composed of \(L_{s1}, C_{s1}, \text{ and } C_{na}\) and the inner network is composed of \(L_{s2}, C_{s2}, L_{s3}, C_{s3}, \text{ and } C_{nc}. \) We can calculate an \(ABCD\) matrix for each network. Each \(ABCD\) matrix can also be transformed into an admittance matrix. From the total admittance matrix, we can find the location of the transmission zeros.

We first consider the outer network of Fig. 1(b). We analyzed the circuit by transforming the \(\pi\)-form (Fig. 2(a)) into the T-form (Fig. 2(b)), using the transformation formulas in (2)~(4) [12]:

\[
L_{p1} = L_{r1}
\]

\[
C_{p1} = \frac{C_{r1}^2 + 2C_{mo}C_{r1}}{C_{r1}}
\]

\[
C_{go} = \frac{C_{r1}^2 + 2C_{mo}C_{r1}}{C_{mo}}
\]

From Fig. 2(b), we can calculate the \(ABCD\) matrix of outer network:

\[
\begin{pmatrix}
    A_{out} & B_{out} \\
    C_{out} & D_{out}
\end{pmatrix} = \begin{bmatrix}
    1 + \frac{Z_{p1}}{Z_{go}} & 2Z_{p1} + \frac{Z_{p1}^2}{Z_{go}} \\
    \frac{1}{Z_{go}} & 1 + \frac{Z_{p1}}{Z_{go}}
\end{bmatrix}
\]

where:

\[
Z_{p1} = j\omega L_{p1} + 1/(j\omega C_{p1})
\]

\[
Z_{go} = 1/(j\omega C_{go})
\]

We can then calculate \(Z, Y, \text{ or } S\) parameters from the \(ABCD\) matrix [13]. According to (1), the transmission zeros are related to \(Y_{21}\), so we can obtain the \(Y_{21}\) of the outer network, as follows:

\[
Y_{21} = -1/B_{out} = \frac{1}{j\omega C_{p1}}
\]

\[
\omega^2 L_{p1} C_{p1} C_{go} - 2\omega^2 L_{p1} C_{p1} (C_{p1} + C_{go}) + 2C_{p1} + C_{go}
\]

Using the same method, we can also calculate the network parameters of inner network. We performed the circuit transform \(\pi\)-form into T-form, as shown in Fig. 3. The transformation formulas are as follows:

\[
L_{p2} = L_{r2}
\]
The inner network of Fig. 1.

\[ C_{p} = \frac{C_{s}^{2} + 2C_{m}C_{s}}{C_{s}} \]  
\[ C_{s} = \frac{C_{s}^{2} + 2C_{m}C_{s}}{C_{m}} \]  

We can then calculate the \( ABCD_{h} \) matrix:

\[
\begin{bmatrix}
A_{h} & B_{h} \\
C_{h} & D_{h}
\end{bmatrix} = \begin{bmatrix} 1 + Z_{s2}Y_{p2} & Z_{s2} \\
2Y_{p2} + Z_{s2}Y_{p2} & 1 + Z_{s2}Y_{p2} \end{bmatrix}
\]

(10)

where

\[ Y_{p2} = 1/(j\omega L_{p2} + 1/(j\omega C_{p2})) \]
\[ Z_{s2} = j\omega L_{s2} \]

The \( ABCD_{h} \) matrix is as follows:

\[
\begin{bmatrix}
A_{h} & B_{h} \\
C_{h} & D_{h}
\end{bmatrix} = \begin{bmatrix} 1 & 0 \\
j\omega C_{p} & 1 \end{bmatrix}
\]

(11)

The upper and lower networks are series-connected; therefore, the impedance parameter can be used to calculate these two networks:

\[ Z_{u} = \frac{1}{Y_{p2}(2 + Z_{s2}Y_{p2})} \left[ \begin{array}{cc} 1 + Z_{s2}Y_{p2} & 1 \\ 1 & 1 + Z_{s2}Y_{p2} \end{array} \right] \]

(12)

\[ Z_{l} = \frac{1}{j\omega C_{p}} \begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix} \]

(13)

where \( Z_{u} \) is upper impedance matrix and \( Z_{l} \) is lower impedance matrix. The total impedance matrix (\( Z_{u} + Z_{l} \)) of two networks is the sum of \( Z_{u} \) and \( Z_{l} \).

\[ Z_{\text{tot}} = Z_{u} + Z_{l} \]

(14)

From \( Z_{u} \), we can obtain the \( ABCD_{h} \) parameters. We can then calculate the \( ABCD_{h} \) parameters, as follows:

\[
\begin{bmatrix}
A_{h} & B_{h} \\
C_{h} & D_{h}
\end{bmatrix} = \begin{bmatrix} 1 & j\omega L_{s1} \\
0 & 1 \end{bmatrix} \begin{bmatrix} A_{h} & B_{h} \\
C_{h} & D_{h} \end{bmatrix} \begin{bmatrix} 1 & j\omega L_{s1} \\
0 & 1 \end{bmatrix}
\]

(15)

To find the transmission zeros, we have to obtain \( Y_{21} \) for the inner network from the \( ABCD_{h} \) parameter, as follows:

\[
Y_{\text{in},21} = -1/B_{m}
\]

\[
Y_{\text{in},21} = \frac{j\omega C_{p} + Y_{p2}(2 + Z_{s2}Y_{p2})}{-2\omega^{2}L_{s1}C_{p}(1 + Z_{s2}Y_{p2}) + j\omega Y_{p2}(2 + Z_{s2}Y_{p2})(2L_{s1} + C_{p}) - \omega^{2}L_{s1}^{2}C_{p}^{2}}
\]

(16)

The total admittance parameter \( Y_{\text{tot},21} \) of the network is then expressed by the sum of \( Y_{\text{in},21} \) and \( Y_{\text{in},21} \), giving us the following expression:

\[
Y_{\text{tot},21} = -1/B_{\text{out}} - 1/B_{m}
\]

(17)

From \( Y_{\text{tot},21} \), we can find the location of the transmission zeros. The transmission zeros are generated at the frequency, where the imaginary part of \( Y_{\text{tot},21} \) is equal to zero.
Fig. 5. The location of transmission zeros by various $C_{mi}$.

Fig. 4 shows the variation of the susceptances by the various $C_{mo}$ when $C_{mi}$ is fixed at 0 pF. The transmission zeros are generated when $\text{Im}(Y_{\text{out}_{21}})=-\text{Im}(Y_{\text{in}_{21}})$. When $C_{mo}$ is 0 pF, the two susceptibility graphs intersect in two places. Therefore, two transmission zeros appear at $\omega_1$ and $\omega_2$. As $C_{mo}$ is varied from 0.01 pF to 0.02 pF, each transmission zero is divided into two transmission zeros located on both sides of $\omega_1$ and $\omega_2$, which gives us four transmission zeros in the stopbands. When $C_{mo}$ increases, the distance between the transmission zeros on both the lower and upper stopbands increases.

Fig. 5 shows the variation of the susceptances by the various $C_{mi}$ with a fixed $C_{mo}$=0 pF. When $C_{mi}$ is 0 pF, the intersection points of two susceptances is two, so we can obtain two transmission zeros. As $C_{mi}$ is varied from 0.01 pF to 0.02 pF, the upper transmission zeros are divided into two transmission zeros, but the lower transmission zeros do not change, which gives us three transmission zeros in the stopbands. The distance between the transmission zeros on the upper stopband and $\omega_2$ in Fig. 6 show the comparison of circuit responses between the original and proposed filters.

Fig. 7. Microstrip layout of the dual-band filter with electric coupling. $w_{Ls}=0.2$, $w_{Lp1}=0.15$, $w_{Lp2}=0.2$, $l_{L1}=0.9$, $l_{L2}=12.5$, $l_{Lp1}=8$, $l_{Lp2}=3$, $l_{cp1}=4.6$, $l_{cp2}=3.7$. All dimensions are in millimeters.}

creases when $C_{mi}$ increases.

Fig. 4 and Fig. 5 show that a small electric coupling
divides one transmission zero into two transmission zeros. As we saw before, the outer electric coupling also has an effect on the lower and upper stopbands, but the inner electric coupling has an effect only on the upper stopband. Therefore, we can set the transmission zeros to the desired number in the desired frequency band simply by appropriate adjustments of $C_{oo}$ and $C_{nm}$ with no effect on the passband characteristics.

III. Dual-band Filter Design with Improved Stopband Characteristics

Using the results achieved so far, we designed a dual-band filter. The filter specifications are as follows:

- Passband : 880~960 MHz (GSM), 2,400~2,500 MHz (ISM);
- Stopband : 1,710~1,785, 1,805~1,880 MHz (DCS);
- Passband minimum return loss : 20 dB;
- Stopband minimum insertion loss : 30 dB.

Fig. 6 shows a comparison of the frequency responses of the original and the proposed networks. We use small electric couplings of 0.01 pF to $C_{oo}$ and $C_{nm}$. We can see that the frequency bands with the insertion loss under −30 dB are extended in both stopbands. The lower stopband is extended about 30%, from 198 MHz to 257 MHz, and the upper stopband is extended about 10%, from 2.123 GHz to 2.347 GHz. However, the passband characteristics are not significantly affected by the electric couplings because $C_{oo}$ and $C_{nm}$ are very small. The effects on $Y_{odd}$ and $Y_{even}$, which were studied in [8], [9], are also small. Therefore, we can achieve improved stopband rejection characteristics with negligible effects on the passband performance.

Fig. 7 shows the proposed dual-band filter with improved stopband characteristics. The filter has a very simple structure, which is designed using microstrip quasi-lumped elements [8], [9]. The electric couplings are realized by the distance control between the symmetric open stubs of the filter structure. The filter was fabricated from a 0.76 mm thick substrate with a relative di-electric constant of 3.5. The total filter size, excluding feed lines, is 14.6×13.2×0.76 mm, which is a very compact size compared with the conventional filter described in [7], [8]. Table 1 shows a comparison of previously reported dual-band filters. Fig. 8 shows the comparison of the measured results with the theoretical and simulated results. Within the first passband (880~980 MHz), the measured insertion loss is < 0.2 dB, whereas the return loss is > 18.87 dB. Within the second passband (2,400~2,500 MHz), the measured insertion loss is < 0.58 dB, whereas the return loss is > 24.8 dB. The measured results are in good agreement with both the theoretical and simulated values.

IV. Conclusion

This paper shows a simple and effective method for improving stopband rejection characteristics in a conven-
tional dual-band filter. By adding a small electric coupling to the symmetrically positioned resonators of lumped-element circuit model, each transmission zero is separated into two transmission zeros. We performed a theoretical analysis of the effect of electric coupling by dividing a circuit into inner and outer circuits. We were able to determine the effect of each electric coupling on the stopbands. For the practical application, we designed a dual-band filter. The filter has a very simple structure and a very small size compared with the conventional filter reported in [7], [8]. The fabricated filter shows a lower insertion loss and a smaller size than conventional filters, which makes the proposed filter suitable for use in many mobile or wireless systems.

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