Progress In Electromagnetics Research Letters, Vol. 37, 73–82, 2013

COMPACT MIXED-CROSS COUPLED BANDPASS FILTER WITH ENHANCED FREQUENCY SELECTIVITY

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Abstract—In this paper, a compact three-order mixed-cross coupled bandpass filter (BPF) with enhanced frequency selectivity is proposed. Multiple transmission zeros (TZs) can be obtained near the passband for high frequency selectivity by introducing mixed-cross coupling between the nonadjacent resonators. The frequency-dependent mixed-cross coupling matrix of the proposed filter is presented to explain the occurrence of the TZs caused by mixed-cross coupling. A new BPF centered at 2.7 GHz with 11.5% fractional bandwidth has been designed and fabricated to verify the validity of the proposed method. The measurement result shows four finite TZs in the stopband, located at 1.74 GHz with 52.16 dB rejection, 2.53 GHz with 24.67 dB rejection, 3.83 GHz with 47.52 dB rejection, and 7.75 GHz with 54.83 dB rejection, respectively. The circuit only occupies $6.2 \times 7.6\,\mathrm{mm}^2$.

1. INTRODUCTION

Compact microstrip bandpass filters with high selectivity and compact size are increasingly demanded in wireless and mobile communication systems to suppress harmonics and spurious signals. Transmission zeros are usually employed to enhance the rejection level of BPFs. The TZs can be realized by various methods such as non-adjacent coupling, signal-interference, mixed electric and magnetic coupling.

Received 11 December 2012, Accepted 7 January 2013, Scheduled 13 January 2013
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The non-adjacent coupling techniques including cross-coupling [1–5] and source-load coupling [6–10] generate transmission zeros by providing a multipath effect. The signal-interference technique is used to design BPFs with high skirt selectivity by forcing in-band signal enhancements and out-of-band destructive signal interferences to produce transmission zeros [11, 12]. The mixed electric and magnetic coupling is another method to generate transmission zeros [13–15], more transmission zeros can be generated in low order filter.

In this paper, to further improve the frequency selectivity of filter, mixed-cross coupling is proposed. Multiple finite TZs can be employed near the passband by introducing mixed-cross coupling between the nonadjacent resonators. The frequency-dependent mixed-cross coupling matrix of the proposed filter is presented and synthesized. To verify the validity of the proposed technique, a three-order filter based on the dual-metal-plane configuration with mixed cross coupling is designed and fabricated. Compared with the conventional cross coupled filter, the proposed filter has a higher selectivity and a more compact size.

2. SYNTHESIS OF THREE-ORDER MIXED-CROSS COUPLED FILTER

Figure 1 shows the equivalent circuit of proposed filter, where L, C, and G denote the inductance, capacitance, and conductance. Based on the Kirchhoff's circuit law, (1) can be obtained, where $L_{i,j}(C_{i,j})$ $(i, j = 1, 2, 3)$ represents the mutual inductance (capacitance) between adjacent resonators.

$$\begin{bmatrix} G_1+j(\omega C_1-\frac{1}{\omega L_1}) & -j\omega C_{1,2} & -j(\omega C_{1,3}-\frac{1}{\omega L_{1,3}}) \\ -j\omega C_{1,2} & j(\omega C_2-\frac{1}{\omega L_2}) & -j\omega C_{2,3} \\ -j(\omega C_{1,3}-\frac{1}{\omega L_{1,3}}) & -j\omega C_{2,3} & G_3+j(\omega C_3-\frac{1}{\omega L_3}) \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \\ v_3 \end{bmatrix} = \begin{bmatrix} i_s \\ 0 \\ 0 \end{bmatrix} \quad (1)$$

or

$$[Y] \cdot [v] = [i] \quad (2)$$

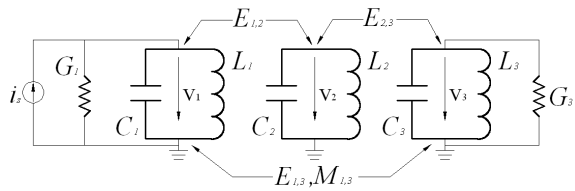

Figure 1. Equivalent circuit of three-order mixed-cross coupled filter.

in which $[Y]$ is an 3×3 admittance matrix

$$[Y] = \omega_0 C \cdot FBW \cdot \left[\overline{Y}\right] \tag{3}$$

where $\omega_0 = 1\big/\sqrt{L_i C_i}$ $(i = 1, 2, 3)$ is the midband angular frequency of filter, $FBW = \Delta\omega/\omega_0$ is the fractional bandwidth, and $\left[\overline{Y}\right]$ is the normalized admittance matrix, which in the case of synchronously tuned filter is given by

$$\left[\overline{Y}\right] = j\cdot\begin{bmatrix} \frac{-j\cdot G_1}{\omega_0 C\cdot FBW}+\omega' & -\frac{\omega}{\omega_0}E_{1,2}\cdot\frac{1}{FBW} & \left(\frac{\omega_0}{\omega}M_{1,3}-\frac{\omega}{\omega_0}E_{1,3}\right)\cdot\frac{1}{FBW} \\ -\frac{\omega}{\omega_0}E_{1,2}\cdot\frac{1}{FBW} & \omega' & -\frac{\omega}{\omega_0}E_{2,3}\cdot\frac{1}{FBW} \\ -\left(\frac{\omega_0}{\omega}M_{1,3}-\frac{\omega}{\omega_0}E_{1,3}\right)\cdot\frac{1}{FBW} & -\frac{\omega}{\omega_0}E_{2,3}\cdot\frac{1}{FBW} & \frac{-j\cdot G_3}{\omega_0 C\cdot FBW}+\omega' \end{bmatrix} \tag{4}$$

where $\omega' = \frac{1}{FBW}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)$ is the lowpass frequency variable. $M_{i,\ j} = \frac{L}{L_{i,\ j}}$, $E_{i,\ j} = \frac{C_{i,\ j}}{C}$ denote the magnetic coupling and electric coupling coefficient, respectively. According to the lowpass frequency variable,

$$\begin{aligned}\frac{\omega_0}{\omega}M_{1,3} - \frac{\omega}{\omega_0}E_{1,3} \ = \ & -\frac{E_{1,3}}{2}\left(\omega'\cdot FBW + \sqrt{\omega'^2\cdot FBW^2+4}\right) \\ & + \frac{2M_{1,3}}{\omega'\cdot FBW + \sqrt{\omega'^2\cdot FBW^2+4}}\end{aligned} \tag{5}$$

For a small FBW, the right-hand side of (5) can be expanded by using Taylor expansion at $\omega' = 0$. By omitting the high-order terms, (5) can be written by

$$\begin{aligned}\frac{\omega_0}{\omega}M_{1,3} - \frac{\omega}{\omega_0}E_{1,3} &= M_{1,3} - E_{1,3} - \frac{FBW}{2}\left(M_{1,3}+E_{1,3}\right)\omega' \\ &= -a\cdot\omega' + b\end{aligned} \tag{6}$$

where

$$a = \frac{FBW}{2}\left(M_{1,3}+E_{1,3}\right), \quad b = M_{1,3} - E_{1,3} \tag{7}$$

or

$$M_{1,3} = \frac{a}{FBW} + \frac{b}{2}, \quad E_{1,3} = \frac{a}{FBW} - \frac{b}{2} \tag{8}$$

Then the frequency-dependent mixed coupled matrix can be derived

$$[K] = \begin{bmatrix} 0 & -E_{1,2} & -a\cdot\omega'+b \\ -E_{1,2} & 0 & -E_{2,3} \\ -a\cdot\omega'+b & -E_{2,3} & 0 \end{bmatrix} \tag{9}$$

The S-parameters are determined by [16]

$$S_{11} = 1 + 2j\ [A^{-1}]_{1,1}, \quad S_{21} = -2j[A^{-1}]_{3,1} \tag{10}$$

where $[A]$ is a function of coupling matrix [16].

In (9), the numerator of $\left[A^{-1}\right]_{3,1}$ is

$$num(\left[A^{-1}\right]_{3,1}) = E_{1,2} \cdot E_{2,3} \text{ - } \omega' \left(-a \cdot \omega' + b\right) \cdot FBW \tag{11}$$

Therefore, the transmission zeros of S_{21} are

$$\omega' = \frac{b}{2a} \pm \sqrt{\left(\frac{b}{2a}\right)^2 - \frac{E_{1,2} \cdot E_{2,3}}{a \cdot FBW}} \tag{12}$$

which are introduced by mixed-cross coupling. Compared with the conventional cross coupled filter, more transmission zeros can be realized.

3. THREE-ORDER MIXED-CROSS COUPLED BPF WITH MULTIPLE TRANSMISSION ZEROS BASED ON DUAL METAL PLANE

Figure 2(a) shows the layout of the proposed BPF. The proposed filter consists of three patch-via-spiral resonators (PVSRs) fabricated in the dual-metal-plane configuration [17]. With the microstrip patch on the top plane serving as a capacitor C and linking to the quasi-lumped spiral inductor L on the bottom plane through a connecting via, the proposed dual-plane resonator structure located on the opposite sides of the single substrate may form a miniaturized one in the printed-circuit board fabrication, as shown in Figure 2(b). The resonance frequency of the single PVSR can be decided by the C and L. Figure 2(c) shows the coupling and routing scheme of the proposed filter. A mixed-cross coupling is introduced between the nonadjacent resonators. The microstrip patches of the coupled-resonator pair on the top plane provide the electric coupling, while the spiral inductors on the bottom plane offer the mutual magnetic coupling.

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{13}$$

The proposed trisection filter with a center frequency of 2.66 GHz and a fractional bandwidth of 11% is designed and fabricated on the Rogers RO4350 substrate. For given the transmission zeros located in 1.7 GHz and 2.5 GHz, respectively. Based on (9), using the synthesis method in [18, 19], the frequency-dependent coupling matrix of the

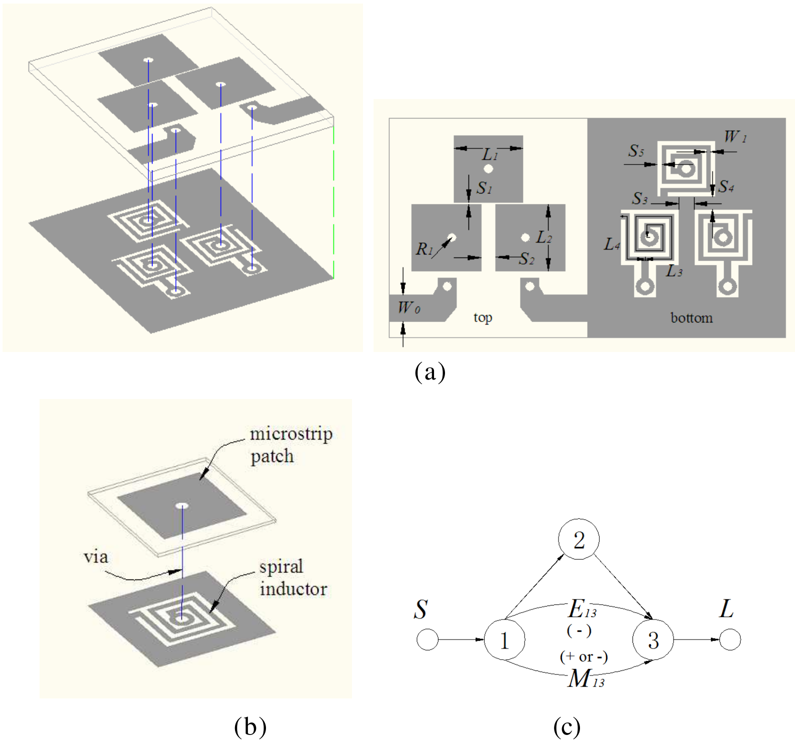

Figure 2. Three-order mixed-cross coupled filter. (a) Layout of proposed BPF. (b) Single PVSR. (c) Corresponding coupling and routing scheme.

proposed filter can be derived as (14). Since the magnetic coupling between the resonators 1, 2 and 2, 3 is much weaker than the electric coupling, the coupling efficient $K_{1,2}$ and $K_{2,3}$ are frequency independent. Therefore, the frequency dependent mixed coupling is only introduced in $K_{1,3}$.

$$K = \begin{bmatrix} -0.0110 & -0.1059 & -0.0091\omega' - 0.0933 \\ -0.1059 & 0.0233 & -0.1059 \\ -0.0091\omega' - 0.0933 & -0.1059 & -0.0110 \end{bmatrix} \quad (14)$$

From (8) and (14), the electric coupling $E_{1,3}$ and magnetic coupling $M_{1,3}$ can be deduced as $M_{1,3} = 0.0361$, $E_{1,3} = 0.1294$. The relation between geometric parameters and coupling coefficients are extracted with the aid of electromagnetic simulation, the extracted coupling results are given in Figure 3. $K_{1,2}$ and $K_{2,3}$ are extracted by conventional method [20]. As for the mixed coupling, the electric

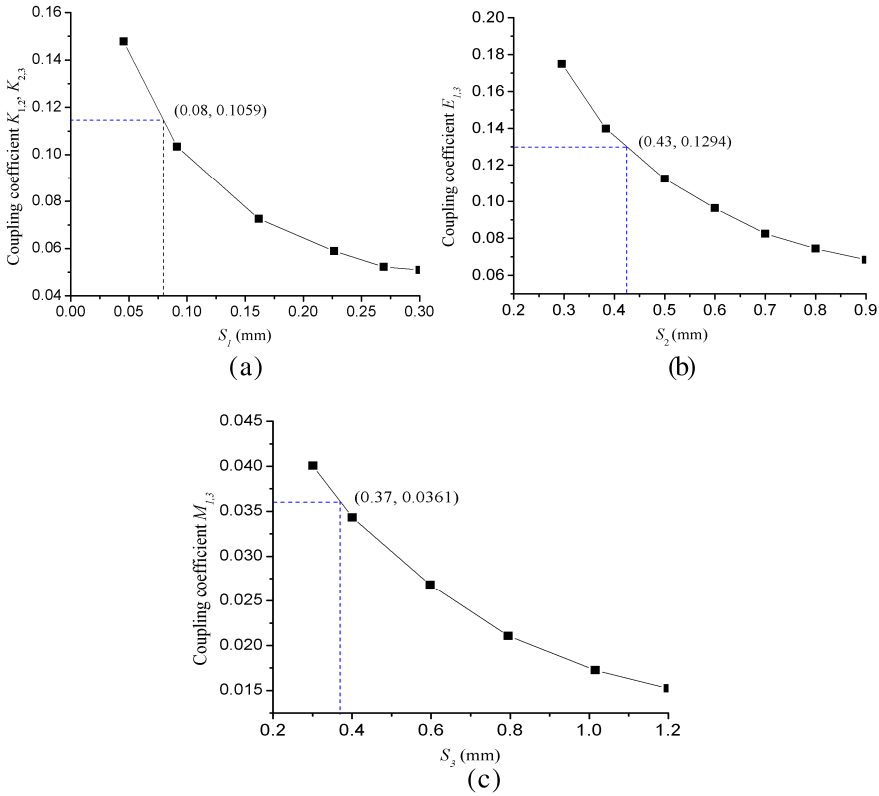

Figure 3. The extracting of coupling coefficients. (a) $K_{1,2}$, $K_{2,3}$. (b) $E_{1,3}$. (c) $M_{1,3}$.

coupling strength $E_{1,3}$ and magnetic coupling strength $M_{1,3}$ need to be extracted separately. While extract the electric coupling $E_{1,3}$, the magnetic coupling $M_{1,3}$ can be eliminated by keeping the two spiral inductors far away from each other, as shown in Figure 3(b). While extract the magnetic coupling $M_{1,3}$, the electric coupling$E_{1,3}$ can be eliminated by keeping the two microstrip patches far away from each other, as shown in Figure 3(c).

The proposed BPF was optimized by EM-simulation using Ansoft HFSS 10 and fabricated on the Rogers RO4350 substrate with a relative dielectric constant of 3.66 and a thickness of 0.508 mm. The geometrical dimensions are decided as: $L_1 = 2.7$ mm, $L_2 = 2.8$ mm, $L_3 = 8.9$ mm, $L_4 = 2.7$ mm, $S_1 = 0.10$ mm, $S_2 = 0.45$ mm, $S_3 = 0.5$ mm, $S_4 = 0.6$ mm, $W_0 = 1.1$ mm, $W_1 = 0.15$ mm, $R_1 = 0.2$ mm. Figure 4 shows the EM-Simulated results and impact of S_2on the

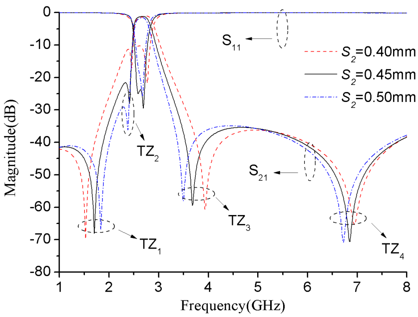

Figure 4. The simulation results and the impact of S_2 to the filter responses.

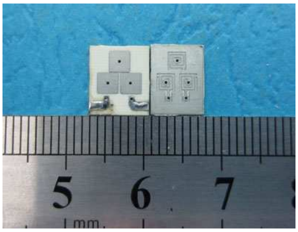

Figure 5. Photography of the proposed filter.

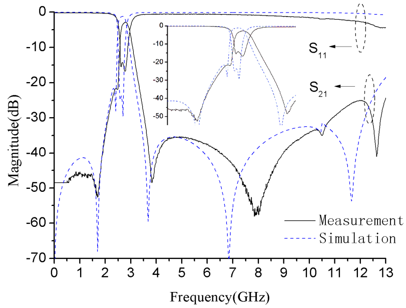

Figure 6. Comparison of simulated and measured frequency responses.

frequency responses. From the simulated results, four transmission zeros were achieved in the stopband, the TZ_1 and TZ_2 were realized by the proposed mixed-cross coupling. TZ_3 was caused by the tap-feeding scheme [21]. $TZ_4$is the harmonic of TZ_1. Keeping the parameters except of S_2 invariable, the TZ_1, TZ_3 and TZ_4 move near the passband, while TZ_2 moves left with the increase of the gap S_2.

4. FILTER FABRICATION AND MEASURED RESULTS

The BPF was fabricated and measured. Figure 5 shows the photograph of the proposed BPF. The proposed filter is measured using an Agilent E8363B network analyzer. Simulated and measured results of the proposed filter are compared in Figure 6 with good agreement. The measured results show that the passband is centered at 2.70 GHz with 11.5% FBW. In addition, the frequency selectivity was enhanced via introducing four finite TZs near the passband, located at 1.74 GHz with 52.16 dB rejection, 2.53 GHz with 24.67 dB rejection, 3.83 GHz with 47.52 dB rejection, 7.75 GHz with 54.83 dB rejection, respectively. And the spurious frequencies of the upper stopband are suppressed below 20 dB from 3.32 to 13.91 GHz. The total area of the proposed BPF is $6.2\times7.6\,\mathrm{mm}^2$ which corresponds to a size of $0.09\lambda_g\times0.11\lambda_g$, where λ_g is the guided wavelength at the center frequency of the passband.

5. CONCLUSIONS

A compact three-order mixed-cross coupled BPF with improved frequency selectivity using PVSRs is proposed. Multiple transmission zeros (TZs) can be obtained by introducing mixed-cross coupling between the nonadjacent resonators. The frequency-dependent mixed-cross coupling matrix of the proposed filter is presented and synthesized. It has been verified by simulation and measurement. The measurement result shows four finite TZs in the stopband, located at located at 1.74 GHz with 52.16 dB rejection, 2.53 GHz with 24.67 dB rejection, 3.83 GHz with 47.52 dB rejection, and 7.75 GHz with 54.83 dB rejection, respectively. The spurious frequencies of the upper stopband are suppressed below 20 dB from 3.32 to 13.91 GHz. The new BPF exhibits favorable selectivity and occupies a size of only $0.09\lambda_g\times0.11\lambda_g$.

ACKNOWLEDGMENT

This work was supported by the National Natural Science Foundation of China (51172034) and the Fundamental Research Funds for the Central University of UESTC (ZYGX2010J120).

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