Design of Quasi-Elliptic Function Filters With a Dual-Passband Response

Jen-Tsai Kuo, *Senior Member, IEEE,* and Hung-Sen Cheng

***Abstract*—Compact miniaturized hairpin resonators are used to design cross-coupled filters with a dual-passband response of elliptic function type. The hairpin resonators are treated as a stepped impedance resonator (SIR) in analysis. In a filter, two different hairpin resonators are used to establish proper couplings required for the two design frequencies. Two filters are fabricated and measured. The results show a good agreement with the simulation.**

***Index Terms*—Cross-coupled, dual-passband, elliptic function response, microstrip filter.**

I. Introduction

RECENT development in wireless communication systems has created a need of RF circuits with a dual-passband operation [1]–[5]. The intuitive idea for implementing dual-frequency operation is to build two circuits for respective bands in a direct-parallel connection [1], [2]. In this way, switch diodes can be used for channel selection. The interaction of the two circuits together with the switch diodes, however, may cause serious problems in increasing insertion loss of the circuit. Thus, it will be beneficial not only in the insertion loss but also in circuit size, number of components, and power consumption that a single circuit can be designed to simultaneously cover two bands [3]–[5]. The design techniques for dual-frequency circuits, however, are still challenging to circuit designers.

In this letter, a planar filter with a dual-passband elliptic function response is presented. Compact miniaturized hairpin resonators [6] are used as the building blocks. In establishing appropriate couplings among the resonators at two design frequencies, the resonators are designed to have different geometric dimensions. With detailed dimensions, two filters with passbands at 2.4/5.2 and 2.45/5.7 GHz are fabricated and measured. The measurement results are compared with those predicted by the IE3D [7] full-wave simulator. In the following, Section II gives the design procedure, Section III compares the measured with predicted responses, and Section IV draws the conclusion.

II. Design Procedure

Design procedure for a planar cross-coupled filter with an elliptic function response has been well documented in [8]. As shown in Fig. 1, the circuit has four hairpin resonators in a 2×2 configuration.

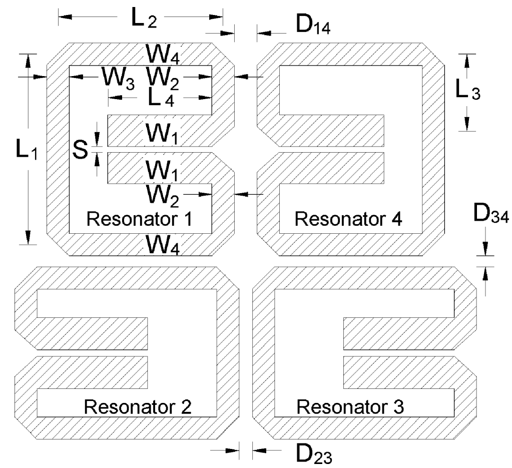

Fig. 1. Circuit layout of a four-pole cross-coupled filter.

In implementing a filter with a dual-passband response, the first step is to make each resonator have its leading two resonances coincident with the two design frequencies by adjusting its geometric parameters. A compact miniaturized hairpin resonator in Fig. 1 can be analyzed as a stepped impedance resonator (SIR) [6] consisting of a cascade of three transmission line sections, namely a microstrip in the middle and, at each end, an open microstrip representing the central coupled lines of the resonator in odd-mode or even-mode. Simple transcendental equations for calculating the resonant frequencies can be obtained.

The second step is to properly place the resonators so that four appropriate couplings in the structure, i.e., $K_{12} = K_{34}$, K_{23}, and K_{14} can be established. They are given by [9]

$$K_{12} = K_{34} = \frac{\Delta}{\sqrt{g_1 g_2}} \tag{1}$$

$$K_{23} = \frac{\Delta \cdot J_2}{g_2} \tag{2}$$

$$K_{14} = \frac{\Delta \cdot J_1}{g_1} \tag{3}$$

where Δ is the fractional bandwidth, g_1 and g_2 are the element values, and J_1 and J_2 are the admittance inverters in equivalent circuit of the low-pass prototype filter. All these coefficients are proportional to the fractional bandwidths. For simplicity, let the two passbands have identical fractional bandwidths. Recall that in design of cross-coupled filters with a single passband, the distances D_{14} and D_{23} have to be known before $D_{12} = D_{34}$ is determined. Thus, designing a 2.4/5.2-GHz filter, we begin with

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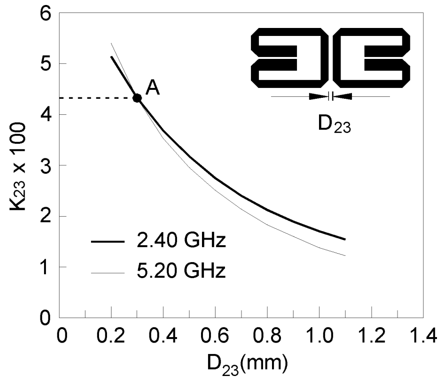


Fig. 2. Coupling coefficients between resonators 2 and 3 as a function of D_{23} at 2.4 and 5.2 GHz.

calculating K_{14} and K_{23} as a function of D_{14} and D_{23}, respectively, by using hairpin resonators with a square peripheral on a substrate with $\varepsilon_r = 2.2$ and thickness 0.508 mm. The calculation of the coupling between two resonators by an EM simulator is referred to [8], and typical results are shown as in Fig. 2. It is found that the two K_{23} curves for the two bands intersect at a point, e.g., point A in Fig. 2, of relatively high couplings. At the same time, the intersection of the two K_{14} curves has relatively low couplings. According to (2) and (3), the ratio K_{23}/K_{14} at the intersection points should be given by $J_2 g_1/J_1 g_2$. One possible way to decrease the K_{23}/K_{14} ratio is to decrease K_{23} by making L_1 shorter for resonators 2 and 3 and increase K_{14} by making L_3 longer for resonators 1 and 4. The change of either L_1 or L_3, of course, will lead to a change of all geometric parameters of a resonator, since the first and second resonant frequencies of each resonator should be kept unchanged.

There is still $K_{12} = K_{34}$ to be determined. Fortunately, the coupling K_{12} at 5.2 GHz is found slightly smaller than that at 2.4 GHz over a wide range of D_{12}, so that the fractional bandwidth for the second band is designed to be slightly less than that for the first one.

The last issue in circuit design is to locate the tap points at the input/output resonators. The tap position directly relates the circuit bandwidth and the singly loaded Q or $Q_{\rm si}$ of the compact miniaturized hairpin resonator. Again, a single tap point has to meet the specifications for both passbands. The value of $Q_{\rm si}$ can be calculated as [10]

$$Q_{\rm si} = \frac{R_L \omega_o}{2} \left.\frac{\partial B}{\partial \omega}\right|_{\omega_o} \tag{4}$$

where B is the total susceptance seen at the tap point looking into the resonator, R_L is the load impedance at the tap point seen by the resonator, and ω_o is the design frequency. In (4), B is a function of both the tap position and frequency. For simplicity, both passbands are assumed to have identical R_L. Then we can plot $\omega_o(\partial B/\partial \omega)$ against tap position at the two design frequencies, and locate the intersection of the two curves. The intersection point actually specifies R_L at the same time. The value of R_L is of course not necessary 50 Ω. Finally, we can design a dual-band transformer [4] to match R_L to 50 Ω.

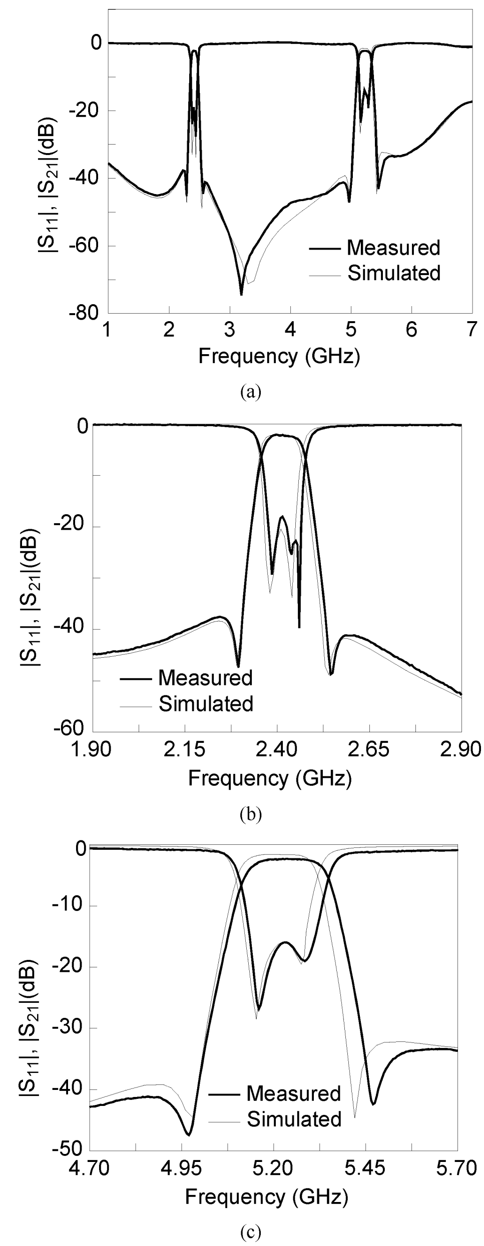
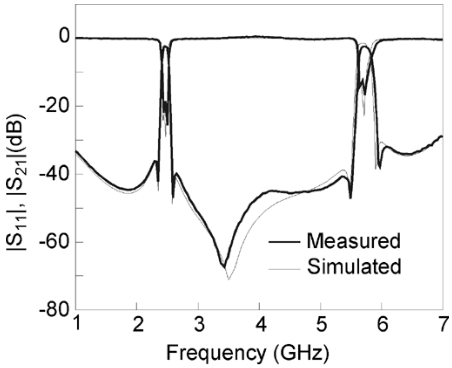


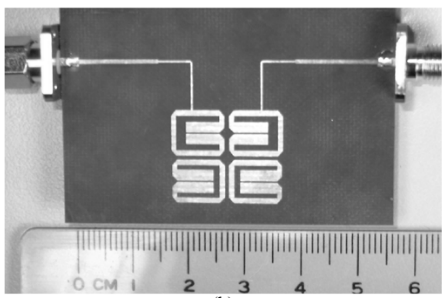
Fig. 3. Simulated and measured responses of filter 1. The design frequencies are 2.4 and 5.2 GHz. (a) Responses in a broad band. (b) Responses around the first passband. (c) Responses around the second passband. Substrate: $\varepsilon_r = 2.2$, thickness $= 0.508$ mm. $D_{12} = 0.55$, $D_{23} = 0.65$, and $D_{14} = 1.12$. Geometric parameters of resonators 1 and 4 are (see Fig. 1) $W_1 = 1.56, W_2 = W_3 = W_4 = 1.1, S = 0.26, L_1 = 9.4, L_2 = 8.14, L_3 = 4.07, L_4 = 5.15$, and those for resonators 2 and 3 are $W_1 = 1.6, W_2 = W_3 = W_4 = 1.1, S = 0.32, L_1 = 7.4, L_2 = 9.95, L_3 = 3.04$, and $L_4 = 5.45$. All are in mm.

III. Fabricated Filters and Measured Results

Fig. 3 plots the responses of a cross-coupled filter designed at 2.4 and 5.2 GHz with fractional bandwidths 5% and 4.7%, respectively. The R_L value at both passbands is 160 Ω, and the dual-band transformer has a 103 Ω-line and a 64 Ω-line of 14.9 mm in length. The plots in Fig. 3(a) cover a frequency band from 1 to 7 GHz, and those in Fig. 3(b) and (c) cover from 1.9 to 2.9 and from 4.7 to 5.7 GHz, respectively. The measured $|S_{21}|$ responses have 2.1 and 2.2 dB at the center passbands. The measured results of the fabricated circuit have a good agreement with the simulation responses.



(a)



(b)

Fig. 4. Simulated and measured responses of filter two. The center frequencies are 2.45 and 5.7 GHz. (a) $|S_{11}|$ and $|S_{21}|$ responses. (b) Photo of the circuit. Substrate: $\varepsilon_r = 2.2$, thickness $= 0.508$ mm. $D_{12} = 0.62, D_{23} = 0.67$, and $D_{14} = 0.85$. Geometric parameters of resonators 1 and 4 are $W_1 = 1.77$, $W_2 = 0.81, W_3 = 0.83, W_4 = 0.96, S = 0.21, L_1 = 6.78, L_2 = 8.34$, $L_3 = 2.88, L_4 = 5.80$, and those for resonators 2 and 3 are $W_1 = 1.89$, $W_2 = 0.81, W_3 = 0.92, W_4 = 0.96, S = 0.2, L_1 = 6.4, L_2 = 8.22$, $L_3 = 2.69$, and $L_4 = 6.33$. All are in mm.

Fig. 4(a) plots the responses of the second filter. The design frequencies are 2.45 and 5.7 GHz, and their respective fractional bandwidths are 5% and 4.4%. Due to the limit of the letter, only the results from 1 to 7 GHz are presented. The R_L at both passbands is 200 Ω, and the dual-band transformer has 109-Ω lines and 80-Ω lines of 13.8 mm in length. The measured $|S_{21}|$ are 2.3 and 2.6 dB at the design frequencies. Fig. 4(b) is the photo of the circuit. It is noted that we chamfer all right-angled bends. Our experiments show that the passband insertion losses are improved by 0.5 dB for both circuits.

IV. Conclusion

Planar cross-coupled filters with a dual-passband quasi- elliptic function response are presented. In the circuit, two resonators with different geometric dimensions are designed to have identical fundamental and the first higher-order resonant frequencies, and to establish appropriate couplings in the structure. Two circuits are designed and fabricated to have a dual-passband response 2.4/5.2 and 2.45/5.7 GHz. The measured results show a good agreement with the simulation.

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