Very Compact Stacked *LC* Resonator-Based Bandpass Filters With a Novel Approach to Tune the Transmission Zeros

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***Abstract*—This letter presents very compact stacked *LC* resonator structures and their coupling and excitation techniques for designing the bandpass filters operating in the 3.5 GHz WiMAX band. Using a coupled stacked *LC*-resonator structure not only greatly reduces the overall size but also can create multiple transmission zeros to enhance roll-off rate or desired stopband rejection for the designed bandpass filters. Furthermore, this letter develops a simple technique of etching a rectangular *C*-shaped slot in the bottom ground plane to flexibly adjust the transmission-zero frequencies without changing the size and shape of the filter.**

***Index Terms*—Bandpass filter (BPF), stacked LC resonator, transmission zeros.**

I. Introduction

MINIATURIZED structure, high selectivity, and large stopband rejection are critical for wireless communication bandpass filters. Most conventional microstrip filters to date are based primarily on transmission-line structures [1]. These filters require too much area because each resonator needs quarter- or half-wavelength to get resonance. Some other studies propose novel structures to reduce filter size, including stepped impedance resonator (SIR) [2], defected ground structure (DGS) resonator [3], patch-via-spiral resonator [4], and net-type resonator [5]. However, the resulting designs are still not size-competitive with low-temperature cofired ceramic (LTCC) bandpass filters. In addition to a miniature size, the transmission zeros are widely used to improve the selectivity or the rejection at the desired stopband frequencies in bandpass filter design. The cascaded quadruplet bandpass filters [2], [5] with a cross-coupling path always have two transmission zeros near both passband edges, but they are usually limited to a fourth-order design. Lower-order bandpass filters can use the source-load coupling technique [6] to effectively create two transmission zeros on both side of the passband, but they generally take more area to implement. In contrast, the tapped-line feed design [7] can also realize two transmission zeros for bandpass filters without adding too much space. In addition, it can tune the transmission-zero frequencies by varying the tapped position [7]. But a disadvantage of this way is to change the matching condition unless a quarter-wavelength transformer is used as [7].

This letter presents a stacked *LC* resonator-based bandpass filter embedded in a two-layer substrate that can achieve both miniaturization and transmission-zero creation for 3.5 GHz WiMAX applications. Tapped-line feed and cross-coupling path have been used to create the transmission zeros for second- and fourth-order design. A novel approach is proposed to adjust the transmission-zero frequencies caused by the tapped-line feed. This approach involves etching a rectangular C-shaped slot in the bottom ground plane without adding any space. Controlling the length of the slot can effectively tune the transmission-zero frequencies in the stopband.

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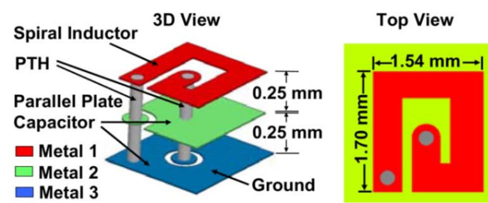
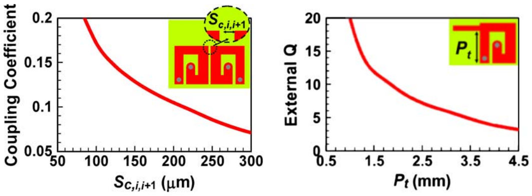


Fig. 1. Stacked *LC* resonator embedded in a two-layer RT/Duroid 6010 substrate.

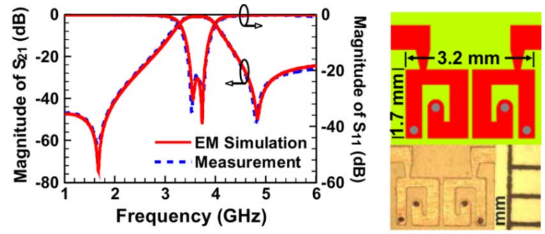
II. Design Methodology

The presented filter structure comprises coupled resonators embedded in a 0.5 mm thick two-layer RT/Duroid 6010 substrate with dielectric constant of 10.2 and loss tangent of 0.0028. Fig. 1 shows a single resonator with a dimension of 1.70 mm $\times$ 1.54 mm $\times$ 0.5 mm. It consists of a spiral inductor on the metal 1 layer and a parallel-plate capacitor using the metal 2 and 3 layers. The inductor and capacitor are stacked as a shunt parallel *LC* resonator by connecting each other to the metal 3 ground plane with two plated through holes (PTHs).



(a) (b)

Fig. 2. (a) Coupling coefficient versus the spacing between two adjacent stacked *LC* resonators. (b) External quality factor versus the tapped position of the input/output port for a single stacked *LC* resonator.



(a) (b)

Fig. 3. Results and structure of the designed second-order bandpass filter. (a) Comparison of the magnitude of S_{21} *and* S_{11} between electromagnetic simulation and measurement. (b) Top view layout and photograph.

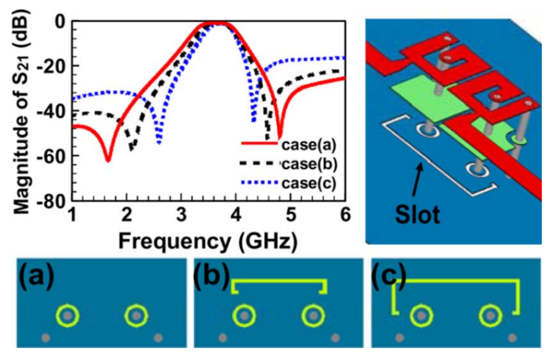


Fig. 4. Comparison of the two transmission-zero frequencies from the measured magnitude of S_{21} for the designed second-order bandpass filter with three different ground structures: Case (a) Without a slot; Case (b) With a 3 mm long slot; Case (c) With a 5.6 mm long slot.

The two most important parameters for designing a coupled resonator bandpass filter are the coupling coefficient $k_{i,i+1}$ between two adjacent resonators with indices i and $i+1$ and the external quality factor Q_e. The coupling coefficient was evaluated as a function of the spacing $(S_{c,i,i+1})$ between two adjacent stacked *LC* resonators, as shown in Fig. 2(a). From knowledge of two dominant resonant frequencies f_1 and f_2, the coupling coefficient can be calculated as [8]

$$k_{i,i+1} = \pm \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2}. \tag{1}$$

The external quality factor of a single stacked *LC* resonator can be derived in terms of the normalized input admittance (y_{in}) and the group delay (τ_Γ) with respect to the reflection coefficient at the resonant angular frequency ω_0 [8]. That is

$$Q_e = \frac{\omega_0 \tau_\Gamma(\omega_0)}{4}\left(1 - y_{in}^2(\omega_0)\right) \tag{2}$$

where y_{in} and τ_Γ can be obtained from the reflection coefficient and the derivative of its phase with respect to angular frequency, respectively. As a result, Fig. 2(b) shows the value of Q_e as a function of the tapped position (P_t) of the input/output port. Note that the results shown in Fig. 2(a) and (b) are based on full-wave electromagnetic simulation using Ansoft HFSS.

To determine the coupling spacing $(S_{c,i,i+1})$ and tapped position (P_t) for an nth-order bandpass filter design, one can use the relations of the $k_{i,i+1}$ and Q_e to the filter prototype elements. These relation are given as [8]

$$k_{i,i+1} = \frac{\Delta}{\sqrt{g_i g_{i+1}}} \tag{3}$$

$$Q_e = \frac{g_1}{\Delta} = \frac{g_n g_{n+1}}{\Delta} \tag{4}$$

where Δ is the fractional bandwidth and g_i is the ith prototype element value. For an nth order Chebyshev design, the filter passband insertion loss can be predicted by the formula given in [9] and shown as

$$IL_{\omega_0}(dB) \approx \frac{4.343}{\Delta Q_u}\sum_{i=1}^{n} g_i \tag{5}$$

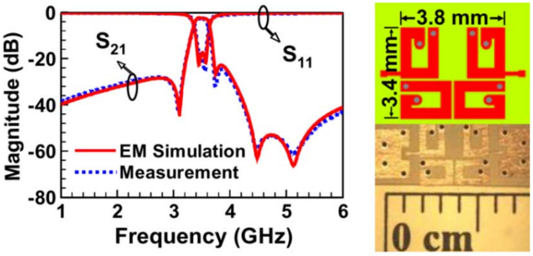
where Q_u is the unloaded quality factor of a stacked *LC* resonator at the resonant frequency.

III. Second-Order Bandpass Filter Design

A second-order 0.5 dB equal-ripple Chebyshev bandpass filter is designed at the center frequency of 3.5 GHz with a fractional bandwidth of 17% and the Q_u of a stacked *LC* resonator is 63. The prototype element values used are $g_1 = 1.403$, $g_2 = 0.707$, and $g_3 = 1.984$. The coupling coefficient $(k_{1,2})$ and the quality factor (Q_e) estimated by calculation are 0.17 and 8.25, respectively. It can be found from Fig. 2 that the $k_{1,2}$ and Q_e values correspond to a coupling spacing of $S_{c,1,2} = 100\ \mu\mathrm{m}$ and a tapped feed position of $P_t = 2.2\ \mathrm{mm}$. The insertion loss at the center frequency that can be evaluated by (5) is 0.86 dB.

Fig. 3(a) compares the magnitude of S_{21} and S_{11} between electromagnetic simulation and measurement for this second-order bandpass filter design, showing excellent agreement over the frequency range from 1 to 6 GHz. The passband insertion loss is approximately 0.9 dB, which is very close to the predicted value of 0.86 dB, and the passband return loss is larger than 22 dB. The tapped-line feeds shown in Fig. 3(b) can create two transmission zeros at 1.67 and 4.82 GHz on both side of the passband. These two transmission-zero frequencies come from the resonances of the two series resonant paths starting at the tapped position. One path starts from the tapped position, passes through a three-quarter turn spiral segment, and then connects to a parallel-parallel capacitor. The other path chooses to travel through a one-quarter turn spiral segment in another direction and then a PTH to the ground plane. The former path is treated as a series *LC* resonator, while the latter path is equivalent to a half-wave short circuit. The occupied area of the designed second-order bandpass filter excluding the feedlines is 3.2 mm $\times$ 1.7 mm.

This letter also presents a simple approach for tuning the transmission-zero frequencies. The proposed approach etches a rectangular C-shaped slot in the metal 3 ground plane to perturb



(a) (b)

Fig. 5. Results and structure of the designed fourth-order bandpass filter. (a) Comparison of the magnitude of S_{21} *and* S_{11} between electromagnetic simulation and measurement. (b) Top view layout and photograph.

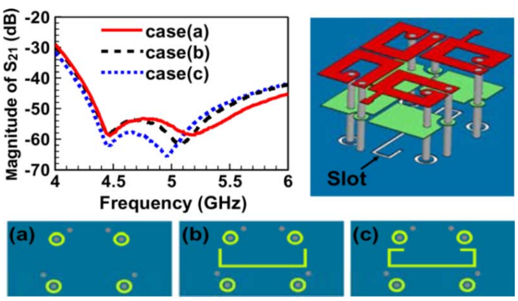


Fig. 6. Comparison of the higher stopband rejection from the measured magnitude of S_{21} for the designed fourth-order bandpass filter with three different ground structures: Case (a) Without a slot; Case (b) With a 4.2 mm long slot; Case (c) With a 5.2 mm long slot.

the series resonant paths for causing the transmission zeros. As shown in Fig. 4(b) and (c), different slot lengths on the ground plane result in different transmission-zero frequencies. Case (a) is the original ground plane without a slot. Case (b) and Case (c) shows the ground plane with a 3 mm and 5.6 mm long slot, respectively. It can be seen from Fig. 4 that the longer slot pushes the transmission-zero frequencies closer to the passband, causing a higher rolloff rate for passband-to- stopband transition. But it is accompanied by the disadvantage of lowering the minimum rejection level in the stopbands.

IV. Fourth-Order Bandpass Filter Design

Multiple stacked *LC* resonators can also be cascaded to form a higher order bandpass filter with enhanced selectivity and stopband rejection. In this letter, a fourth-order Chebyshev bandpass filter with 0.01 dB ripple level is designed with a center frequency of 3.5 GHz and a fractional bandwidth of 10%. The prototype element values used are $g_1 = 0.713$, $g_2 = 1.20$, $g_3 = 1.321$, $g_4 = 0.648$, and $g_5 = 1.101$. To reduce the insertion loss of the fourth-order bandpass filter, the Q_u of a stacked *LC* resonator is increased to 76 by widening the metal width of the inductor in the resonator under the same resonant frequency. The estimated coupling coefficients and external quality factor are $k_{1,2} = k_{3,4} = 0.108$, $k_{2,3} = 0.079$, and $Q_e = 7.13$. The insertion loss calculated at the center frequency is 2.22 dB. The required resonator coupling spacing and tapped feed position are found as $S_{c,1,2} = S_{c,3,4} = 195\ \mu\text{m}$, $S_{c,2,3} = 270\ \mu\text{m}$, and $P_t = 3.1$ mm.

As shown in Fig. 5(a), the electromagnetic simulated results are in good agreement with the measured results. In the passband, the measured insertion loss is approximately 2.4 dB, which agrees with the predicted value of 2.22 dB, and the return loss is better than 14.6 dB. Note that the rolloffs on both sides of the passband are very steep because of the two transmission zeros located at 3.09 and 3.74 GHz. These two transmission zeros are created by the cross-coupling path. Fig. 5(b) shows the top-view layout and photograph of the designed fourth-order bandpass filter. The tapped-line feeds used at the input and output ports can create two additional transmission zeros at 4.47 and 5.12 GHz, which achieves a stopband rejection of more than 45 dB from 4.3 to 6 GHz. The occupied area of the designed fourth-order bandpass filter excluding the feedlines is 3.8 mm $\times$ 3.4 mm.

The proposed rectangular C-shaped slots are also implemented on the metal 3 ground plane of the designed fourth-order bandpass filter, as shown in Fig. 6. Case (a), (b) and (c) represents no slot, a 4.2 mm long slot, and a 5.2 mm long slot, respectively. It can be seen from Fig. 6 that the longer slot can move the higher two transmission-zero frequencies toward the passband. However, this moving effect is not as obvious as in the second-order design. This is because the passband as well as the lower two transmission zeros near the passband are hardly changed by the presence of the slots, which forms a bigger obstacle to the movement of the higher two transmission zeros.

V. Conclusion

This letter has presented novel bandpass filter designs by mutually coupling the stacked *LC* resonators. The advantages Include very small size and ease of implementation using printed circuit technology. The tapped-line feeds used at the input and output ports create two transmission zeros for the proposed bandpass filters. Etching a rectangular C-shaped slot in the bottom ground plane provides an effective way to tune the frequencies of these two transmission zeros. As a result, the presented filter structure can achieve high performance and design flexibility under a stringent size constraint.

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