A Double Optimized Transmission Zeros Based On π-CRLH Dual-Band Bandpass Filter

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A Double Optimized Transmission Zeros Based on π-CRLH Dual-Band Bandpass Filter

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Abstract—In this paper, design and measurements of a highly selective π-CRLH dual-band bandpass filter, with transmission zeros optimized to serve Wi-Max applications, is presented. The dual-bands are designed at 5.2 and 5.7 GHz with a sharp rejection level between them and transmission zeros before and after the passbands. The filter is designed using coupled gap zeroth order composite right/left-handed (CRLH) resonators, which results in significant filter size reduction. Furthermore, two different coupled π-CRLH filters are discussed through the work development of this paper. The filter design concepts are verified and confirmed using electromagnetic simulations and experimental measurements. Presented results reveal that the proposed filter exhibits a rejection level greater than -20 dB, while maintaining 2 dB insertion loss and better than -25 dB for the transmission zeros with compact size ($12 \times 16\,\mathrm{mm}^2$) which is 70% smaller than similar conventional filters.

1. INTRODUCTION

Metamaterial (MTM) is an artificial material, and through altering its electromagnetic properties, it exhibits unusual propagation phenomena, which have made it very attractive for scientists in the last decade [1]. Left-handed metamaterials (LHMs) were presented as one type with simultaneous effective negative permittivity and permeability. This property has presented novel functionalities and size minimization of microwave components and antennas in a way that cannot be achieved using traditional materials [2]. The planar version of the LHM is called a composite right/left-handed transmission line (CRLH-TL) in which a series capacitor and shunt inductor are loading, periodically, a planar transmission line such as coplanar waveguide and microstrip. In [3], the composite right/left-handed transmission line (CRLH-TL) was first proposed. As a result, CRLH metamaterials have brought a revolution in the development of microwave circuits and aroused great interest in engineering and scientific communities.

There is a need for development in microwave filters properties with small size, simple design, low cost, and low insertion loss. Also, numerous conducted researches have focused on the design of multi-band filters since they are key components in most portable devices. However, these filters have a trade-off between circuit size and performance [4]. Conventional filters can operate only at fundamental frequencies and its integral harmonics, resulting in a massive circuit size. Hence, metamaterial CRLH based filters have presented different solutions for a goal where the filter functionality can be controlled arbitrarily using the loading elements only [5, 6]. By implementing and using a CRLH configuration, the ratio of the two central frequencies can be made arbitrary, and size minimization is achieved [7]. One possible performance with the size reduction is met using a zeroth-order resonator (ZOR) mode since it does not require a condition half-wavelength resonance [8]. Accordingly, several researches have been

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conducted using MTM-ZOR techniques for developing dual-bands such as highly selective filters, and in the design of bandpass filters (BPFs) in different frequency bands [9, 10]. Similarly, coupled mechanism bandpass filters can lead to a small size and sharp CRLH filters [11–13]. Different methods of compact and sharp multi-band CRLH filters are also proposed [14–23].

In this paper, a detailed design procedure of a compact dual-band bandpass π-CRLH filter with transmission zeros using coupled zeroth order gap CRLH resonators is introduced. Second and fourth order designs are fabricated on a Rogers's 6010 substrate with dielectric constant = 11.2, thickness = 1.27 mm, tanδ = 0.0023, and copper trace = 0.35 um. In Section 2, the ZOR-CRLH cell is introduced. In Sections 3 and 4, the second and fourth order π-CRLH filters are presented, respectively. The simulation has been done using commercial EM full wave simulators (ANSYS-HFSS).

2. π-CRLH DESIGN AND RESULTS

In Fig. 1(a), the equivalent circuit of a π-CRLH resonator is shown. In this circuit, the elements, C_L and L_L, refer to left-handed loading elements, where C_R and L_R refer to right-handed elements, which can be either loading or parasitic elements of the hosting planar transmission line. The π-CRLH cell has the advantage of simple matching and higher possible multi-band functionality than the T-CRLH cell configuration. The analysis of periodic π-CRLH cell can be done using either transmission line theory or periodic structure. The realization of the microstrip π-CRLH as a zeroth order gap resonator is shown in Fig. 1(b), where C_L is realized as an 8-finger interdigital capacitor, and L_L is realized using a simple via hole. The realization of the right-handed elements is done by making use of the parasitic elements of the microstrip transmission line. The cell is an air coupled to a $50\,\Omega$ Z_0 transmission line through a short feeding line.

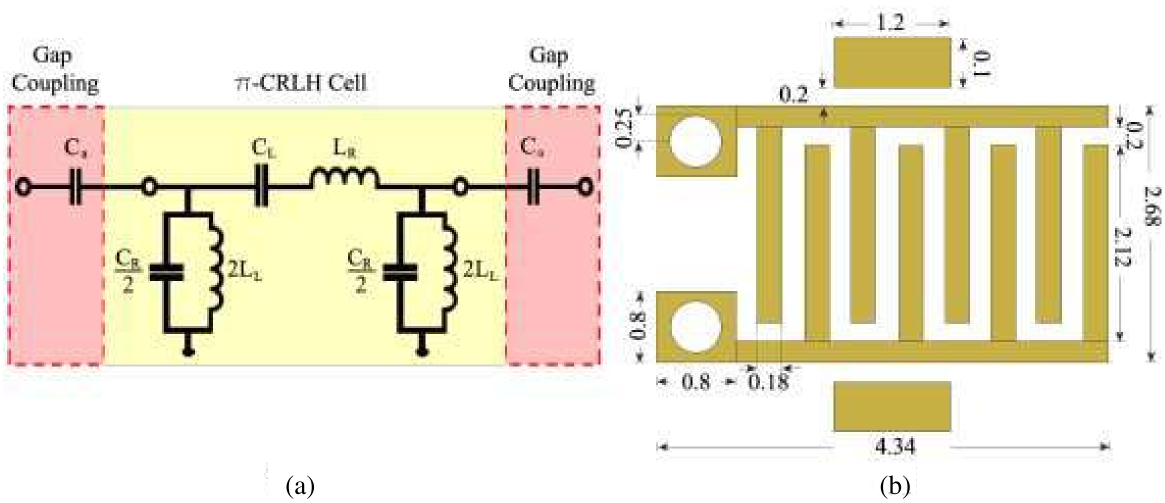


Figure 1. (a) Equivalent circuit of coupled gap π-CRLH resonator, (b) the 2D geometry of the proposed coupled gap π-CRLH resonator.

The π-CRLH cell was designed to demonstrate constant balanced $50\,\Omega$ characteristic impedance, in order to match the resonator to a $50\,\Omega$ feeding. Based on the condition, the matching design equation is fulfilled as in Eq. (1).

$$\sqrt{L_R/C_R} = \sqrt{L_L/C_L} \tag{1}$$

Along with this matching condition, the resonant condition can be adjusted as a zero phase as in Eq. (2).

$$\varphi_{CLRH}|_{5.2\,GHz} = -\beta l = \left(\frac{1}{\omega\sqrt{C_L L_L}} - \omega\sqrt{C_R L_R}\right) = 0 \tag{2}$$

In our design, the parasitic elements L_R and C_R were prespecified corresponding to a $50\,\Omega$ transmission line section of length $L = 2.5$ mm. This step was intended to control the size of the designed cell to achieve compactness. Solving Eqs. (1) and (2), C_L and L_L were calculated. The next step was synthesizing the designed elements on a Rogers substrate so that the π-CRLH cell dimensions were specified as in Fig. 1(b). The synthesis equations of the interdigital capacitor and via hole have been done using the formulas in [24] as in Eqs. (3) and (3d) where N_f is the number of the fingers, W_f the finger width = 0.2 mm, h_{sub} the substrate thickness = 1.27 mm, ε_r the substrate relative permittivity = 11.2, and R the via radius = 0.2 mm.

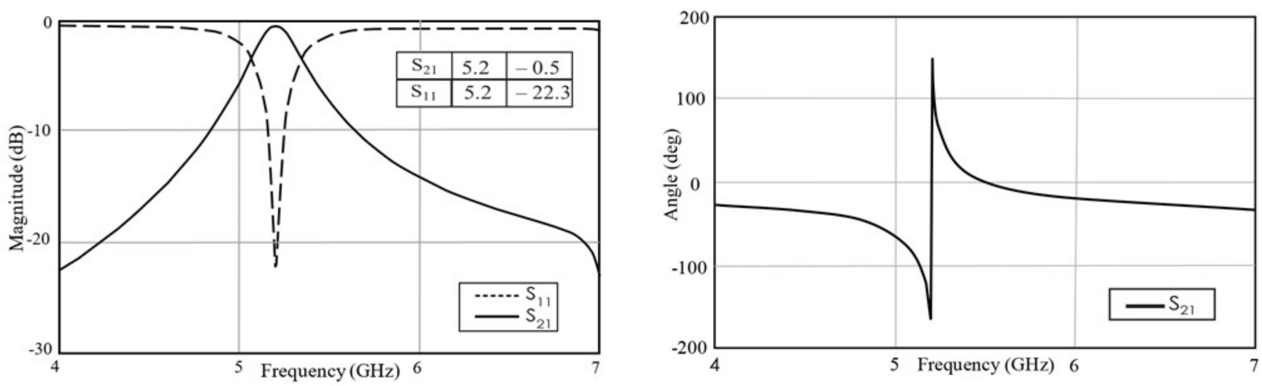
$$C_L(pF) = (\varepsilon_r + 1)\, L_f\, [(N_f - 3)A_0 + A_1] \tag{3a}$$

$$A_0 = 4.40 \tanh\left[0.55(h_{sub}/W_f)^{0.45}\right] \times 10^{-6} \tag{3b}$$

$$A_1 = 9.92 \tanh\left[0.52(h_{sub}/W_f)^{0.5}\right] \times 10^{-6} \tag{3c}$$

$$L_L(pH) = 0.2\left[h_{sub} - \ln\left(\frac{h_{sub} + \sqrt{r^2 + h_{sub}^2}}{R}\right) + 1.5\left(R - \sqrt{R^2 + h_{sub}^2}\right)\right] \tag{3d}$$

To confirm the zeroth order mode of the designed coupled gap π-CRLH resonator, the designed structure in Fig. 1(b) was simulated using 3D EM simulator, and the results are plotted in Fig. 2. In Fig. 2(a), it is obvious that the resonator demonstrates a sharp resonance at 5.2 GHz with S_{21} −0.5 dB and S_{11} lower than −20 dB of the unit cell. On the other hand, by plotting the transmission coefficient phase ($\angle S_{21}$) in Fig. 2(b), it is quite obvious that the resonator exhibits a very close value to zero phase at 5.2 GHz.



(a) (b)

Figure 2. The3D EM simulated (a) magnitude and (b) phase of the π-CRLH resonator.

Compared to conventional half wavelength transmission resonator, it can be confirmed that the designed coupled gap π-CRLH resonator length (L = 2.68 mm) is only $0.13\lambda_g$ at 5.2 GHz which represents a 75% size reduction.

3. A SECOND ORDER FILTER RESULTS AND DISCUSSION

The design of a second order filter is based on edge coupling two previously designed resonators. The filter topology and fabricated prototype are shown in Fig. 3(a) and Fig. 3(b), respectively. Thanks to the external gap coupling separation between the two resonators, an external coupling is added, and hence a second band is obtained. This second band has been optimized to be at 5.8 GHz. The relative feedings of the feeding lines (W_1 and W_2) are adjusted to phase difference between the two resonators enough to create a stopband between the two bands. The separation (S) between the two resonators is adjusted to control the second band.

The simulated and measured transmission coefficients (S_{21}) and reflection coefficients (S_{11}) of the second order filter are shown in Fig. 4. The results reveal that the filter has two passbands. The filter passbands are centered at 5.2 and 5.77 GHz with S_{21} = −2.4 dB and −0.35 dB, respectively, and return loss close to 20 dB. The 3 dB bandwidth for the designed filter is from 5.14 GHz to 5.27 GHz (2.5%)

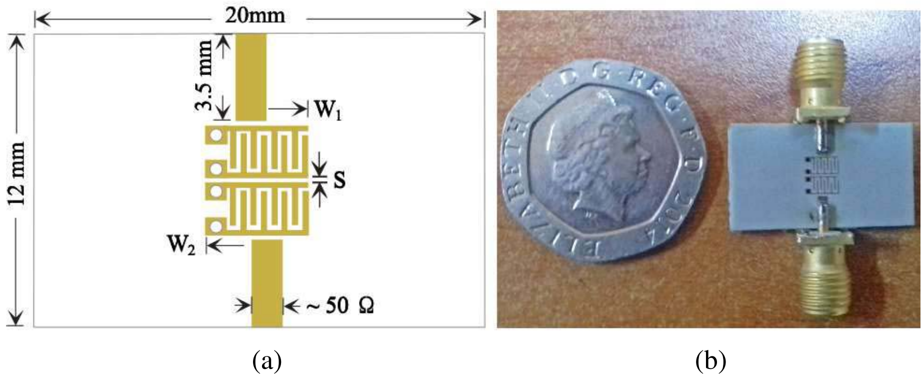


Figure 3. (a) The second order π-CRLH filter layout. (b) The fabricated filter prototype.

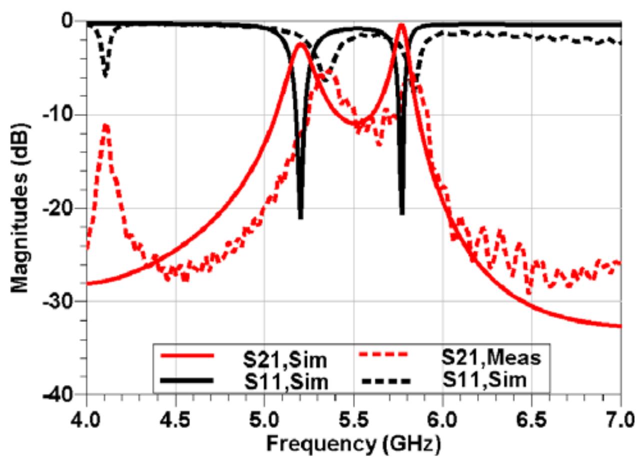


Figure 4. The simulated and measured scattering parameters of the second order π-CRLH filter.

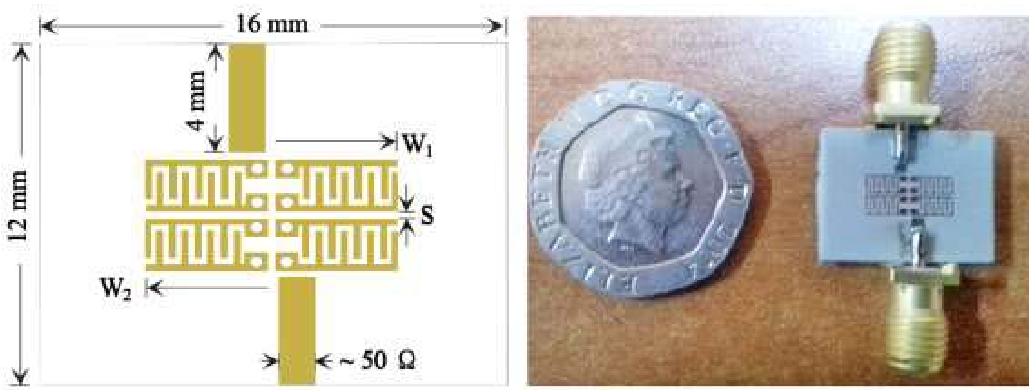
for the first band and from 5.72 GHz to 5.81 GHz (1.55%) for the second band. The filter total size is $12 \times 20\,\mathrm{mm}^2$ which demonstrates the compactness of the designed filter.

4. FOURTH ORDER π-CRLH FILTER

In the previous section, it is obvious that a dualband filter has been designed, but no transmission zeros were achieved. In this section, we present how the previously designed filter was modified to have transmission zeros. The idea of our modification is based on increasing the coupling order of the filter so that two transmission zeros have been designed before the first band and after the second band.

The final fourth order filter layout and prototype are shown in Figs. 5(a) and 5(b) respectively, with a size of $12\times16\,\mathrm{mm}^2$. This cross-coupling configuration is proposed to add extra coupling that introduces a double transmission zero before and after the passbands with a sharp rejection in-between, a result of the cross-coupling owing to the multiple paths without deteriorating the performance. Comparing the second and fourth order filters, it is authenticated that the fourth order has the advantage of having transmission zeros over the second order as a result of the initiated multiple paths. Path 1 is an electric coupling between the two cells capacitors (C_L). Path 2 is a magnetic coupling between the used magnetic vias.

The simulated and measured filter scattering parameters of the fourth order filter are shown in Fig. 6. It can be seen that the filter has dual passbands at 5.2 and 5.67 GHz with both $S_{21} = -2.8\,\mathrm{dB}$ and S_{11} close to 20 dB. The 3 dB bandwidth for the designed filter is from 5.16 GHz to 5.24 GHz (1.5%) for the first band and from 5.64 GHz to 5.69 GHz (1%) for the second band. In addition, the figure



(a) (b)

Figure 5. (a) The fourth order π-CRLH filter layout. (b) The fabricated filter prototype.

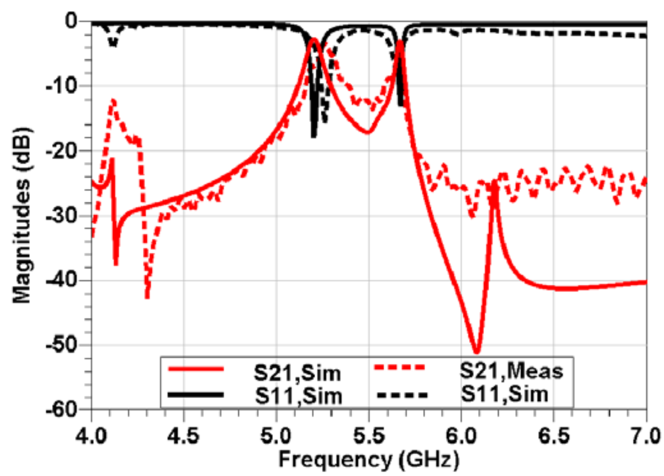


Figure 6. The simulated and measured scattering parameters of the fourth order π-CRLH filter.

Table 1. Performance comparison of our proposed dual-band bandpass filter with recent published filters.

References	Dual Centre frequencies (GHz)	Insertion loss (dB)	Return loss (dB)	3 dB Fractional bandwidth (%)	Electrical size ($\lambda_0 \times \lambda_0$) at f_{01}
[5]	0.9/2.4	1.0/0.85	18/22	12/12	0.175×0.065
[7]	1.5/2.3	2.5/3	22/25	5.76/4.98	54.3×30.3
[15]	0.81/2.42 0.6/2.3	1.4/2.2 1.6/2.5	12/20 > 12/11	20/19.5 16.5/4	0.14×0.03 0.14×0.016
[21]	2.4/5.2	3.6/3.1	15/23	5.8/6.4	0.15×0.16
[22]	3.5/5.7	0.98/0.81	22/25	3.3/6.4	0.21×0.18
[23]	2.4/3.57	0.87/1.9	20/15	7.5/6.16	0.5×0.2
This work, 2nd order 4th order	5.2/5.8 5.2/5.7	2.3/1.5 2.6/2.6	22/21 18/15	2.5/1.55 1.5/1	0.21×0.35 0.21×0.28

reveals that the filter has three transmission zeros at 4.1 GHz with −38 dB rejection, 5.5 GHz with −18 dB rejection, and 6.1 GHz with −51 dB rejection, respectively. Accordingly, it can be claimed that these transmission zeros improve the rejection level in the stopbands and the selectivity in the passbands. The measured filter results match the simulation with dual passbands at 5.2 and 5.7 GHz. In summary, the achieved transmission zeros have resulted in an improvement in skirt selectivity. Finally, it is worthwhile to comment that a good agreement among simulated, theoretical and measured results is achieved.

To validate the effectiveness of the proposed filter, performance of the proposed filter is compared to recent published works in Table 1. It is evident that the proposed filter has good competitive properties.

5. CONCLUSION

In this paper, a design of a CRLH dual-band BPF with an optimized double transmission zero is achieved. The design is done using a zeroth-order resonator unit cell to serve the Wi-Max band applications. The proposed structure introduces an optimized transmission zero before and after each passband center frequency, with a sharp rejection between the two passbands, as a result of the cross-coupling done for enhancing the selectivity. The CRLH filter size is $12 \times 16\,\mathrm{mm}^2$, which is about 70% smaller than conventional circuits such as the parallel coupled type, designed on the basis of the half-wavelength resonance.

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