Narrowband filters with tightly coupled resonators

W.M. Fathelbab  M.J. Almalkawi

Department of Electrical and Computer Engineering, South Dakota School of Mines and Technology, Rapid City, SD 57701, USA
E-mail: wfathelbab@ieee.org

Abstract: The design of narrowband combline filters implemented in double-layered microstrip is presented. The filters realise finite-frequency transmission zeros located within close proximity with respect to their passband edges. This is achieved by tightly coupling the resonators of the filters. Experimental evaluation of several fabricated filters demonstrates the principle.

1 Introduction

Microwave bandpass filters are essential components for a host of system applications such as test equipments, telecommunications and radars. The advanced functionality of modern systems has imposed stringent requirements on filters such as small size, minimal insertion loss and high selectivity. For example cellular radio applications have transmit and receive bands that are closely assigned because of the congested frequency spectrum. Consequently, there is always demand for narrowband filters with asymmetrical transfer characteristics as is the case with diplexers or multiplexers [1]. The design of asymmetrical filters [1, 2] requires the placement of finite-frequency transmission zeros either in the lower or the upper stopband and continues to be a subject of research interest.

The combline filter [1–6] has applications in radio and microwave frequency applications because of its compactness and excellent frequency performance. It is a known fact that the resonators of a combline filter are well spaced for a small fractional bandwidth and are closely spaced if the bandwidth is large. Contrary to this common belief, this communication will demonstrate that it is feasible to realise narrowband combline filters with closely spaced resonators. Furthermore, the transmission coefficients of such filters have finite-frequency zeros that are located close to their passband edges.

This work is organised as follows. Sections 2 illustrates the concept of requiring tight inter-resonator coupling for the realisation of combline filters with finite-frequency zeros. Section 3 presents the difficulties encountered in realising such filters in single-layered planar technology. Section 4 proposes a fabrication technique that overcomes such difficulties.

2 Coupled resonator filters with mixed coupling

Illustrated in Fig. 1a is a second-order bandpass circuit prototype capable of realising a finite-frequency transmission zero. The zero is generated because of the resonance of the coupling branch that may occur at a frequency below or above the passband of the circuit. The circuit of Fig. 1a can be easily realised in combline (see Fig. 3a) but first its properties will be scrutinised. Initially, four circuits were synthesised each realising either a lower or an upper stopband zero. The simulated responses of those circuits are shown in Figs. 1b and c. Subsequently, the capacitive inter-resonator coupling was eliminated and the transmission coefficient of each circuit evaluated. This was done after weakly coupling into the first resonator and out of the last resonator using decoupling capacitors as shown in Fig. 2a. The transmission coefficient of each of those circuits then becomes what is well known as the coupling bandwidth [1, 2]. Simulation of each case led to the plots that are illustrated Figs. 2b and c. It is clearly observed from Figs. 2b and c that as the finite-frequency zero moves closer to the edge of the passband the coupling bandwidth dramatically increases. This implies that for a selected fractional passband bandwidth (15% in all the presented cases) it is feasible to increase the coupling between...
CIRCUIT’S ELEMENT VALUES REALISING A LOWER STOPBAND ZERO – CASE 1

\[ C_1 = 7.114 \text{ pF}; L_1 = 2.883 \text{ nH} \]
\[ C_{12} = 13.582 \text{ pF}; L_{12} = 2.831 \text{ nH}; K = 50 \Omega \]

CIRCUIT’S ELEMENT VALUES REALISING A LOWER STOPBAND ZERO – CASE 2

\[ C_1 = 4.996 \text{ pF}; L_1 = 4.3471 \text{ nH} \]
\[ C_{12} = 40.985 \text{ pF}; L_{12} = 0.79 \text{ nH}; K = 50 \Omega \]

CIRCUIT’S ELEMENT VALUES REALISING AN UPPER STOPBAND ZERO – CASE 3

\[ C_1 = 8.829 \text{ pF}; L_1 = 3.581 \text{ nH} \]
\[ C_{12} = 9.01 \text{ pF}; L_{12} = 1.8733 \text{ nH}; K = 50 \Omega \]

CIRCUIT’S ELEMENT VALUES REALISING AN UPPER STOPBAND ZERO – CASE 4

\[ C_1 = 5.856 \text{ pF}; L_1 = 5.101 \text{ nH} \]
\[ C_{12} = 32.294 \text{ pF}; L_{12} = 0.62 \text{ nH}; K = 50 \Omega \]

Figure 1 A second-order circuit realising a finite-frequency zero in a 50 Ω system

a The prototype
b and c Typical frequency characteristics pertinent to this circuit
All circuits realise a response whose fractional bandwidth is 15% centred at 1 GHz

3 Single-layered planar implementations

Shown in Fig. 3a is the physical layout of a combline filter corresponding to the circuit prototype of Fig. 1a. Standard milling dictates that the minimum spacing, \( s \), between a pair of tracks that can be reliably manufactured is approximately 0.203 mm (8 mil). With this minimum spacing it was desired to build a second-order combline filter with mixed coupling to find out where its zero would be located with respect to its passband. This exercise was done for a passband centred at 1 GHz achieving a 15% fractional bandwidth. Two filters were designed on a Rogers 4003C (http://www.rogerscorporation.com) substrate one realising a lower stopband zero and another an upper stopband zero. The substrate has a dielectric constant of 3.38, loss tangent of 0.0027 and is 0.812 mm (32 mil) thick. The structures were simulated utilising CST Microwave Studio (CST studio suite, CST computer simulation technology, version 2006) resulting in the structure’s element values and responses as
shown in Figs. 3b and c. Each structure was then built leading to the measured frequency characteristics demonstrated in Figs. 3d and e. It is seen that the response of Fig. 3d is virtually symmetrical implying that its transmission zero is located at a much higher frequency above the passband. On the other hand, a lower passband transmission zero was realised leading to the asymmetrical response illustrated in Fig. 3e. In summary, a narrower spacing between the resonators (much less than 0.203 mm) would be required to position either of the zeros closer to the passband edges. The next section proposes a solution that alleviates this realisation difficulty.

4 Double-layered planar implementations

The new implementation approach is illustrated in Fig. 4a that depicts a main substrate and a uniform ground plane with a relative dielectric constant of $\varepsilon_r$, and a thickness of $h_1$, and a thin dielectric layer with a relative dielectric constant of $\varepsilon_{r2}$ and a thickness of $h_2$. This second-order structure is a lumped-distributed realisation of the lumped element circuit of Fig. 1a and basically is an alternative realisation to that of Fig. 3a. Here the first resonator of
Figure 3 The conventional lumped-distributed microstrip realisation of the circuit of Fig. 1a, simulations and measurements

a Physical layout
b One realising a zero in the upper stopband
c Another realising a zero in the lower stopband
d The measured frequency performance upon implementation of the two microstrip circuits one realising a zero in the upper stopband
e Another realising a zero in the lower stopband

Structure's element values realising an upper stopband zero

\[ C_1 = 3.4 \text{ pF}; C_2 = 2.45 \text{ pF}; C_{12} = 1 \text{ pF}; \]
\[ w_1 = 1.93 \text{ mm (76 mil)} = w_2; \]
\[ s = 0.203 \text{ mm (8 mil)}; \]
\[ l = 13.97 \text{ mm (550 mil)} \]

Structure's element values realising an lower stopband zero

\[ C_1 = 2.7 \text{ pF}; C_2 = 4.2 \text{ pF}; C_{12} = 5 \text{ pF}; \]
\[ w_1 = 2.794 \text{ mm (110 mil)} = w_2; \]
\[ s = 0.203 \text{ mm (8 mil)}; \]
\[ l = 13.97 \text{ mm (550 mil)} \]
Figure 4 The proposed lumped-distributed double-layered microstrip realisation of the circuit of Fig. 1a

a Physical layout

b and c The simulated frequency performance of two double-layered microstrip circuits
d and e The measured frequency performance of two double-layered microstrip circuits

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<tr>
<th>Structure's Element Values Realising an Upper Stopband Zero</th>
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<tbody>
<tr>
<td>( C_1 = 3.2 , \text{pF} ); ( C_2 = 2.5 , \text{pF} ); ( C_{12} = 6.3 , \text{pF} );</td>
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<tr>
<td>( w_1 = 1.219 , \text{mm (48 mil)} ); ( h_1 = 0.812 , \text{mm (32 mil)} );</td>
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<td>( l = 13.97 , \text{mm (550 mil)} ); ( l_1 = 10.67 , \text{mm (420 mil)} )</td>
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<th>Structure's Element Values Realising an Lower Stopband Zero</th>
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<td>( C_1 = 4.5 , \text{pF} ); ( C_2 = 1 , \text{pF} ); ( C_{12} = 27 , \text{pF} );</td>
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<tr>
<td>( w_1 = 3.556 , \text{mm (140 mil)} ); ( h_1 = 0.812 , \text{mm (32 mil)} );</td>
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<tr>
<td>( l = 13.97 , \text{mm (550 mil)} ); ( l_1 = 11.05 , \text{mm (435 mil)} )</td>
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Figure 5 A fourth-order circuit prototype realising a pair of finite-frequency zeros located in the upper stopband

a Prototype

b Its lumped-distributed double-layered microstrip realisation

c The simulated frequency performance of the double-layered microstrip circuit

d The measured frequency performance upon implementation of the double-layered microstrip circuit
the filter is milled on the main substrate whereas the second is milled on the thinner substrate. The second resonator is then carefully positioned over the first as shown in Fig. 4a. This way much tighter coupling between the resonators is achieved in comparison to the single-layered implementation of Fig. 3a. Thus by appropriately adjusting the values of $C_1$, $C_2$ and $C_{12}$, a finite-frequency transmission zero located very near to the passband edge is expected to be realised.

The configuration of Fig. 4a was simulated in CST Microwave Studio and two structures implemented each realising a zero either in the lower or upper stopband. Both the main substrate and the thin dielectric overlay were Rogers 4003C substrates of thicknesses 0.812 mm (32 mil) and 0.203 mm (8 mil), respectively. The simulated frequency responses of the structures together with their element values are shown in Figs. 4b and c. It is seen that the finite-frequency transmission zeros in the simulated responses of Figs. 4b and c are now located much closer to the edges of the passband in comparison to the simulated responses of Figs. 3b and c. This is of course due to the realisation of the required strong coupling between the pair of resonators comprising the filter. The response of Fig. 4c actually shows a minimum attenuation level of $-7$ dB in its lower stopband, which can be further increased upon increasing the order of the filter. A filter with a fast transition from passband to stopband should be capable of improving the performance of an RF frontend because of the suppression of nearby interferences. Subsequent implementation of the filters resulted in the measured characteristics illustrated in Figs. 4d and d. It is evident from Fig. 4d that the upper stopband zero is now located closer to the passband in comparison to the response of Fig. 3d that did not have it visible over the same frequency span. Also from Fig. 4c it is clear that the lower stopband zero is much closer to the lower passband edge when compared with the response of Fig. 3e. It is therefore possible to conclude that the proximity of the transmission zero to either of the passband edges is directly proportional to the thickness of the dielectric overlay, $b_2$.

In order to further demonstrate the presented principal, the fourth-order circuit prototype shown in Fig. 5a was utilised to design a filter with two upper stopband transmission zeros. An electrical layout of the synthesised prototype was simulated using CST Microwave Studio, which is shown in Fig. 5b together with its element values. As seen the two pairs of resonators are realised in a similar fashion as previously done in Fig. 4a. The simulation of the electrical circuit is shown in Fig. 5c whereas its measured response is depicted in Fig. 5d. It is seen that the measured and simulated results well match each other.

Finally, one of the upper stopband zeros of Fig. 5d was reconfigured to the lower stopband by manually altering some of the capacitor values of the implemented fourth-order filter. The result is shown in Fig. 6. Once again it is clearly evident that because of the realisation of the tight inter-resonator couplings, the zeros of the reconfigured filter were placed very near to the passband edges.

5 Conclusions

The design of narrowband combline filters with tightly coupled resonators was presented in this paper. The filters were fabricated in double-layered microstrip leading to the realisation of finite-frequency transmission zeros that are located within close proximity to their passband edges. Experimental measurements of fabricated filters have demonstrated highly selective and compact filters with minimal passband insertion loss levels.

It is therefore possible to conclude that the presented filters will prove advantageous in modern multi-layered microwave circuits such as microwave monolithic integrated circuits or low-temperature co-fired ceramic technologies.

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7 References


