MICROWAVE FILTERS WITH MULTIPLE CROSS-COUPLINGS AND MAXIMUM NUMBER OF CONTROLLED ATTENUATION POLES

G. Lojewski*, N. Militaru*, M.G. Banciu**

*University POLITEHNICA of Bucharest
E-mail: george.lojewski@munde.pub.ro

**National Institute of Materials Physics
E-mail: gbanciu@infim.ro

Abstract–*In this paper a novel configuration of microwave planar filters, with multiple cross-couplings and with a number of poles of attenuation NZ equal to the order N of the filter, is investigated. The position of the poles on the frequency axis can be controlled, allowing the design of band-pass filters with improved selectivity with respect to the adjacent channels. The new configurations were designed and verified by em-field simulation. The responses of the designed filters are in good agreement with the specification, confirming the possibilities of designing microwave band-pass filters of a relatively low order with moderate losses and with improved performances.*

Keywords: *band-pass filters, attenuation poles, cross-couplings, coupling matrix*

1. STRUCTURE OF THE INVESTIGATED FILTERS

It is well known that poles of attenuation in the transfer characteristic of a band-pass filter can be obtained only in the presence of one or more cross-couplings between the elements of the filter. The number of attenuation poles NZ cannot be greater than the order N of the filter, i.e. the number of resonators in the filter. The maximum number of poles, NZ = N, can be obtained only if the topology of the filter allows not only cross-couplings between resonators and/or between resonators and the lines, but also a direct coupling between the two access lines (Fig. 1).

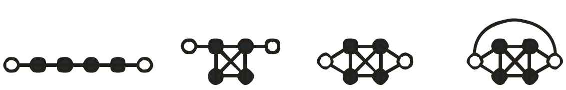

Fig. 1. Different types of band-pass filter topologies (here, case of N = 4 resonators): a) the in-line topology; b) with cross-couplings; c) with cross-couplings and with multiple couplings to the access lines; d) with cross-couplings, multiple couplings to access lines and with a direct coupling between these lines.

The generalized coupling matrix **M**, describing a topology like that shown in Fig. 1, has N + 2 rows and columns. It contains, in a normalized form, all the coupling coefficients between different resonators, the couplings between resonators and access lines represented by the corresponding loaded *Q*'s of the resonators, the possible direct coupling between input and output represented by the characteristic admittance of the corresponding inverter. This matrix **M** contains also the offsets of the individual resonators with respect to the central frequency of the filter, in a normalized form. All these key parameters of the filter can be derived from **M**, through a straightforward de-normalization procedure [1]. The normalized matrix **M** corresponding to some given specifications can be obtained through a basic synthesis procedure, presented in [1], [2].

The object of this paper is to investigate the practical possibilities of implementing topologies like that of Fig. 1, in a simple, cost-effective planar technology like microstrip.

The analysis was focused on the fourth-order filters, because the quadruplet with cross-couplings between resonators was intensively studied in the last years [3]. In order to obtain four attenuation poles an extra coupling has to be realized directly between the input and output lines, beside the cross-coupling possibilities offered by such a quadruplet.

2. DESIGN OF A FOURTH-ORDER FILTER, WITH FOUR PRESCRIBED POLES OF ATTENUATION

To illustrate the design procedure, a band-pass filter was designed, with the next main specifications:

- central frequency 3 GHz;
- bandwidth 75 MHz (2.5%);
- 50 Ohms terminal impedances;

- Chebyshev response with a 0.46 dB ripple in the passband, corresponding to a return loss of 10 dB;
- four attenuation poles, symmetrically located at normalized frequencies ±1.5 and ±3 (f_1 = 2.8969 GHz, f_2 = 2.948 GHz, f_3 = 3.052 GHz and f_4 = 3.104 GHz).

Starting with these specifications, the corresponding normalized matrix **M** was found by using a home-made program based on the methods presented in [1] and [2]:

$$\mathbf{M}=\begin{bmatrix} 0 & -0.7749 & 0 & 0 & 0 & 0.0129 \\ -0.7749 & 0 & 0.6729 & 0 & 0.2101 & 0 \\ 0 & 0.6729 & 0 & 0.7090 & 0 & 0 \\ 0 & 0 & 0.7090 & 0 & -0.6727 & 0 \\ 0 & 0.2101 & 0 & -0.6727 & 0 & 0.7749 \\ 0.0129 & 0 & 0 & 0 & 0.7749 & 0 \end{bmatrix}$$

The non-zero couplings needed for the implementation of this filter are shown in Fig. 2.

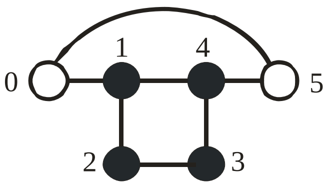

Fig. 2. Couplings needed for the filter corresponding to the matrix **M** above.

After de-normalization and disregarding for the moment the signs, the characteristic admittances (in Ohms) of the inverters which correspond to these couplings are:

Table 1. The characteristic admittances of the inverters

J_{in-out}	$J_{in-1} = J_{4-out}$	$J_{1-2} = J_{3-4}$	J_{2-3}	J_{1-4}
0.0129	1.9426	4.2272	4.4548	1.3203

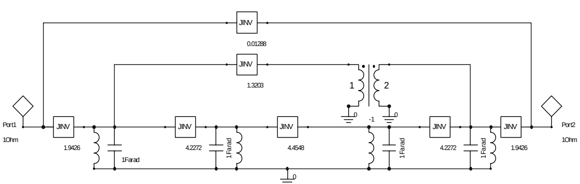

Fig. 3. A normalized model of the band-pass filter, with lumped resonators and ideal admittance inverters.

These values were verified by using a simple model of the filter, with ideal lumped resonators and ideal admittance inverters. This model is presented in Fig. 3 and its simulated response obtained with a circuit simulation software [4], is shown in Fig. 4. It is easy to notice the perfect match of this response with the filter specifications.

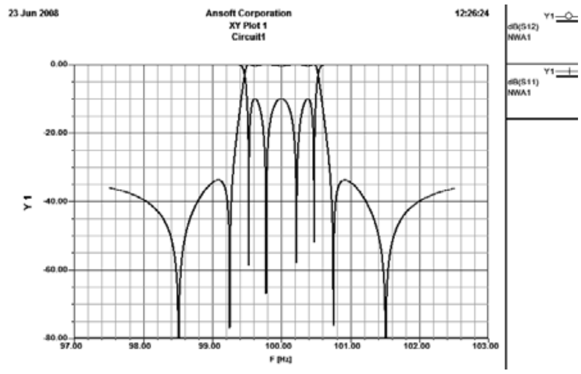

Fig. 4. Simulated response of the normalized filter shown in Fig. 3.

In Fig. 5, the theoretical response of the new structure with four attenuation poles is compared with an in-line fourth-order filter without any attenuation poles, and with a classical quadruplet with two attenuation poles. It shows clearly that the new structure allows a better attenuation of the adjacent channels.

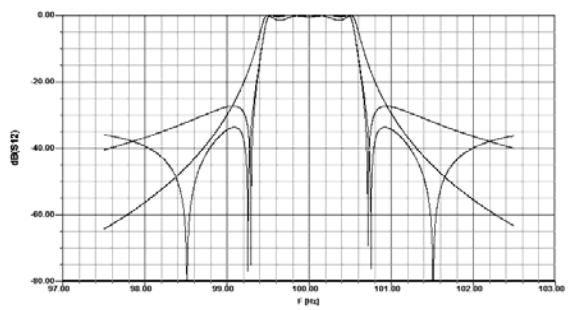

Fig. 5. Comparison between different fourth-order filters with similar in-band characteristics: a filter in-line, a classical quadruplet with two attenuation poles, and the new filter with four attenuation poles.

When trying to design this new type of filter in a practical manner, the main problem is to find a convenient way to realize the necessary direct coupling, through an admittance inverter, between the input and output lines. The inverter function can be approximated in practice by a λ/4 or a 3λ/4 piece of line (depending on the sign needed for this coupling), but these line inverters are working properly only in narrow frequency bandwidths. Other types of inverting circuits offer larger bandwidths. The solution chosen in this design is an inverter containing two capacities and a transmission line (Fig. 6).

The layout shown in Fig. 6 was designed accordingly to the filter specifications mentioned above, on a Rogers 3003 substrate with a thickness of 20 mils and with a relative permittivity ε_r of 3. This substrate was chosen due to its low losses, $\tan\delta = 0.0013$, at 3 GHz.

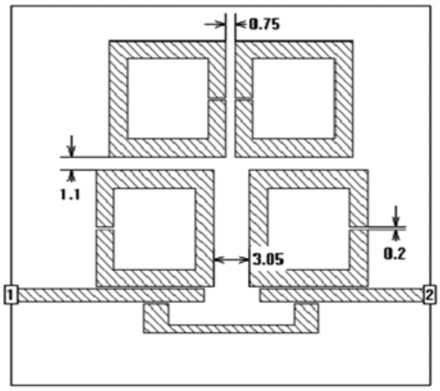

Fig. 6. The microstrip structure of the designed filter with four open-square resonators, a cross-coupling between resonators 1 and 4 and with an extra coupling between the input and output lines.

3. RESULTS

The designed structure was tested by using an electromagnetic field simulation software [5]. The results of the simulation are shown in Fig. 7. Generally speaking, they are in agreement with the filter specifications, but some important differences can be noticed.

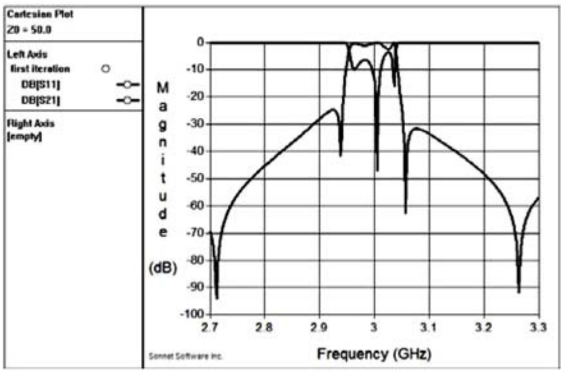

Fig. 7. Simulated response of the lossless microstrip structure shown in Fig. 6.

Some of these differences can be explained by the limited resolution of the grid in the layout design, resolution chosen based on the expected technological tolerances. However, the major differences can be explained only by an imperfect control of the desired couplings. In this sense, important errors can be generated by the fact that resonant frequencies of the end resonators are considerably influenced by their couplings with the lines, while the weak couplings, like the cross-coupling 1-4, are sensibly modified by the presence of other elements of the filter, in the final layout.

The design errors of this kind can be at least partially removed through a special optimization procedure [6], [7]. The optimized layout of the filter is shown in Fig. 8, where the corrections are represented by some slight changes in the geometries of the resonators and couplings.

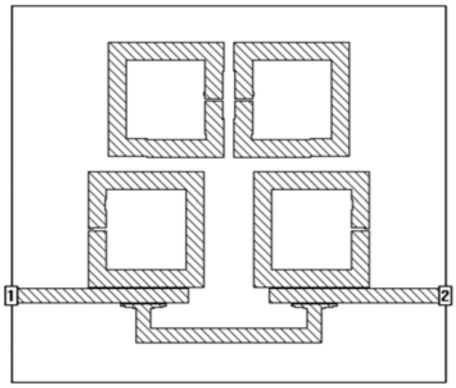

Fig. 8. Final layout of the designed filter (dimensions in millimeters).

The response of this corrected filter is presented in Fig. 9. It can be noticed that the optimization procedure was remarkably efficient: the simulated response is now very similar to the ideal, corresponding very well to the filter's specification.

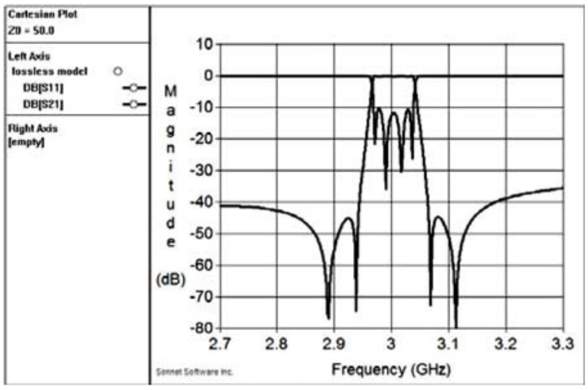

Fig. 9. Corrected response (after optimization).

The simulated results lead to the conclusion that such a design procedure which combines the electromagnetic field simulation with an optimization procedure is appropriate for designing filters with a maximum possible number of attenuation poles.

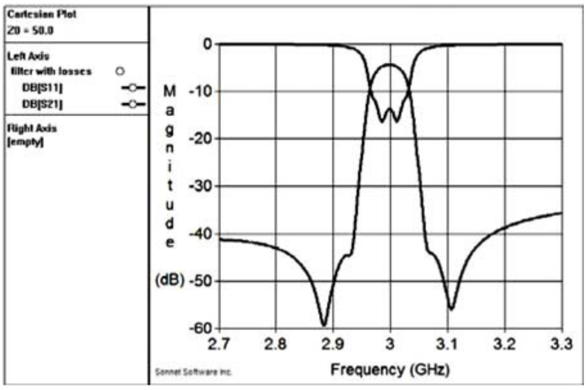

Fig. 10. The response of the filter, simulated in the presence of the dielectric and metal losses.

The results shown in Fig. 10 were obtained including in the simulation the known losses of the dielectric substrate and of the metallic sheets. It can be noticed that the insertion loss of the filter is moderate, despite its narrow bandwidth, as a consequence of its relatively low order.

4. CONCLUSIONS

The paper demonstrates that such filters, with a maximum number of attenuation poles, can provide a better selectivity with respect to the adjacent channels than the classical ones. This advantage can be achieved with only a slight change in the layout, while the area occupied by this new layout does not increase significantly.

In comparison to a classical filter with a similar selectivity, which must have a higher order, this novel solution offers a better miniaturization and can assure a lower insertion loss.

A drawback of this new filter can be its moderate far out-of-band attenuation. To compensate for this, the new filter can be cascaded with other simple filters designed with considerably larger bandwidths.

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