

MICROSTRIP FILTERS WITH MULTIPLE CROSS-COUPPLINGS AND MAXIMUM NUMBER OF ATTENUATION POLES

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Abstract. *In this paper a novel configuration of microwave planar filters, with multiple cross-couplings and with a number of attenuation poles equal to the order of the filter, is investigated. The location 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. A new configuration was designed, verified by em-field simulation, fabricated and tested. 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.*

INTRODUCTION

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).

The generalized coupling matrix \mathbf{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.

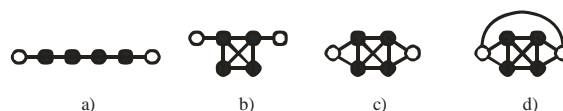


Fig. 1. Different types of band-pass filter topologies (here, case of $N = 4$ resonators):
a) the in-line topology; b) with cross-couplings; c) with cross-couplings and with multiple couplings to the access lines; d) with cross-couplings, multiple couplings to access lines and with a direct coupling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

This matrix \mathbf{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 \mathbf{M} through a straightforward de-normalization procedure [1]. The normalized matrix \mathbf{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.

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

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.

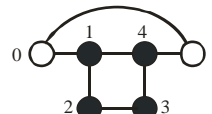


Fig. 2. Couplings needed for the filter corresponding to the matrix **M** above

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

TABLE 1 Values of the inverters' constants

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

These values can be 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 circuit simulation software [4], is shown in Fig. 4.

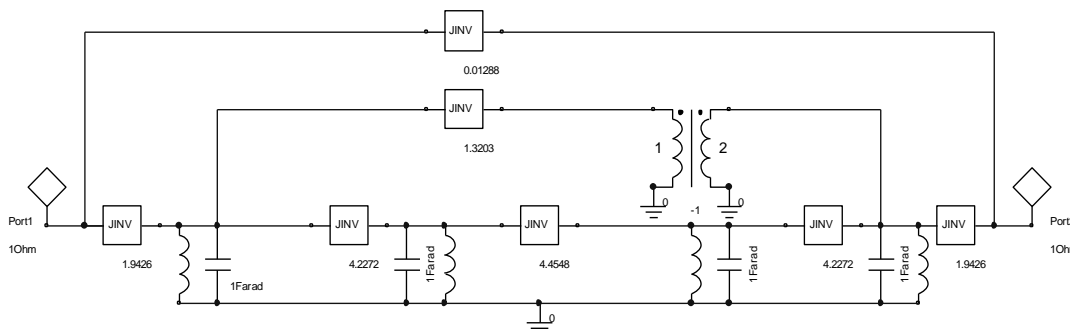


Fig. 3. A normalized model of the band-pass filter, with lumped resonators and ideal admittance inverters

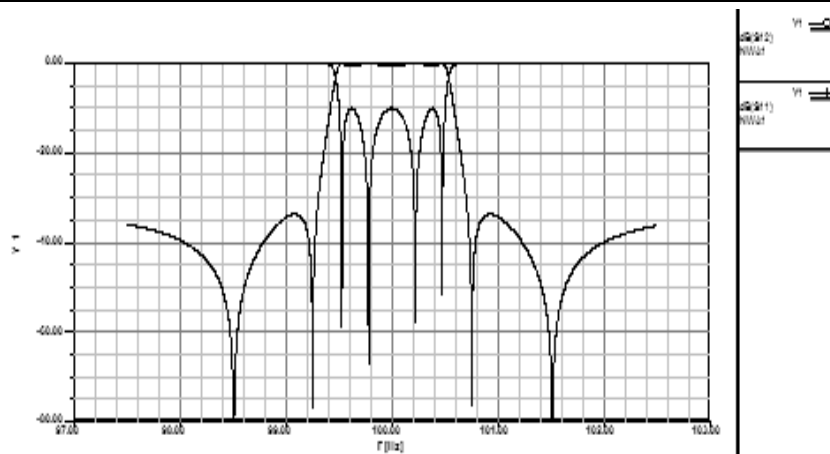


Fig. 4. Simulated response of the normalized filter shown in Fig. 3
 Fig. 5.

It is easy to notice the perfect match of this response with the filter specifications.

In Fig. 5, the theoretical response of the new structure with four attenuation poles is compared to 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 provides a better attenuation of the adjacent channels.

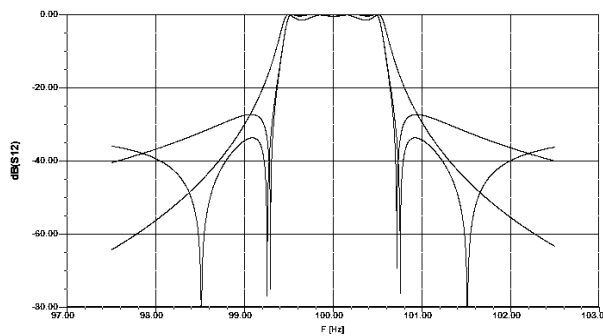


Fig. 6. Comparison between different fourth-order filters with similar in-band characteristics: a filter in-line, a classical quadruplet with two attenuation poles, and the new filter with four attenuation poles

When trying to design this new type of filter in a practical manner, the main problem is to find a way to realize the necessary direct coupling, through an admittance inverter, between the input and output lines.

The inverter function can be realized with different types of inverting circuits. The solution chosen in this design is an inverter containing two capacities and a transmission line, as shown in Fig. 6.

This layout 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 $\epsilon_r = 3$. This substrate was chosen due to its low losses, $\tan \delta = 0.0013$ at 3 GHz.

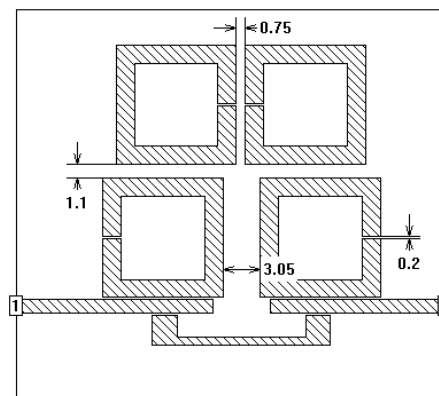


Fig. 7. The microstrip structure of the designed filter with four open-square resonators, a cross-coupling between resonators 1 and 4 and with an extra coupling between the input and output lines (dimensions in millimeters)

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

these differences can be explained by the limited resolution of the grid in the layout design, resolution chosen considering the expected technological tolerances.

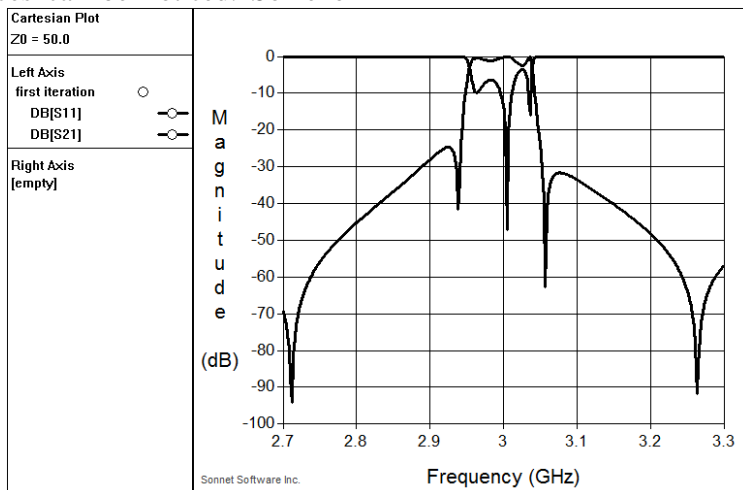


Fig. 8. Simulated response of the lossless microstrip structure shown in Fig. 6

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 access lines, while some 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]. 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. 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.

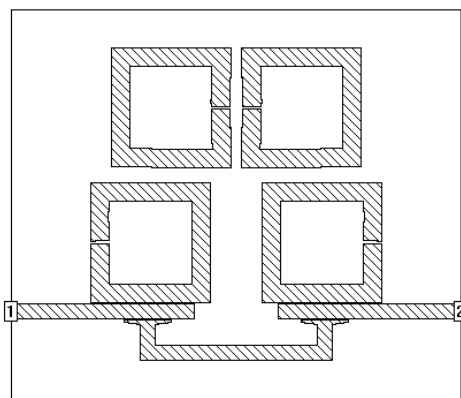


Fig. 9. Final (corrected) layout of the filter

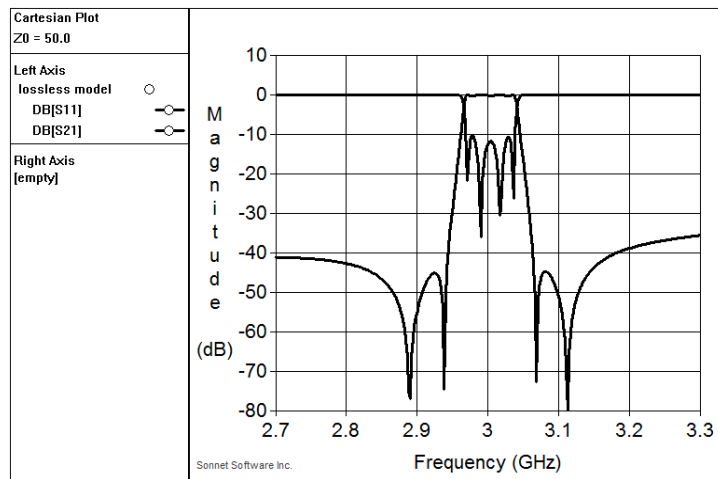


Fig. 10. Corrected response

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.

2. EXPERIMENTAL RESULTS

Based on EM-field simulation, the filter from Fig.8 was fabricated and tested. The photograph of the experimental model is shown in Fig. 10.

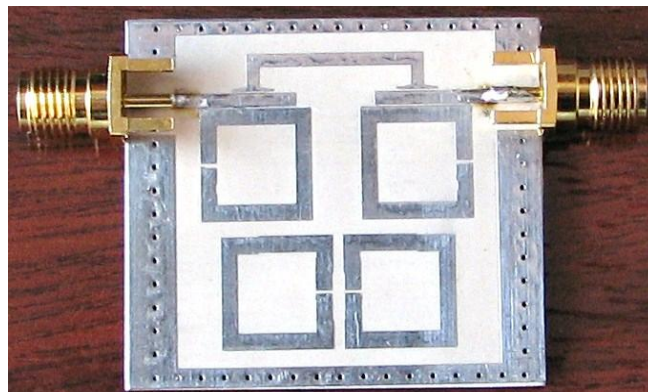


Fig. 11. Prototype of the designed bandpass filter

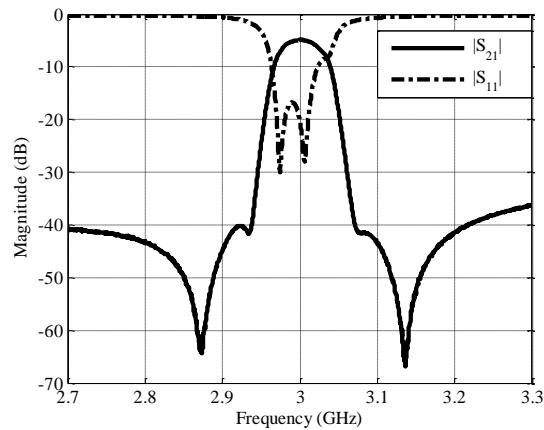


Fig. 12. Measured response of the prototype

The prototype exhibits a centre frequency of 3.001GHz and an in-band insertion loss of 4.8dB, as depicted in Fig. 11.

The measured 3dB-bandwidth of this filter is of 62MHz. The four attenuation poles are located at 2.87GHz, 2.93GHz, 3.07GHz, and 3.13GHz, respectively.

3. 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.

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.

4. REFERENCES

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