Microstrip Filtering Structure with Optimized Group-Delay Response for Wireless Communications

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Abstract: - In this paper the design of a filtering structure with improved group-delay characteristic in microstrip technology is presented. This improvement is achieved by using pairs of equalization poles. In addition, an improved selectivity with respect to the adjacent channels is achieved by introducing two attenuation poles in the stop-band, one on each side of the pass-band. As a design example, a sixth-order quasi-elliptic filter with an in-band Chebyshev response, having two prescribed attenuation poles and two optimized group-delay equalization poles is designed, simulated, fabricated and tested. The measured characteristics of the experimental model are found to be in good agreement with the results anticipated through electromagnetic-field simulation. In a similar way, filters acting as equalization circuits can also be obtained using a slightly different optimization procedure, in order to improve the overall group-delay of a specified communication system.

Key-Words: - Group-Delay, Attenuation Poles, Microstrip, Band-Pass Filter, Coupling Matrix, Planar Resonators

1 Introduction

Modern communication systems often require filtering structures with high electrical performances, small dimensions, and low cost.

The electrical performances of band-pass filters are related to the necessity of having low insertion loss, good selectivity with respect to the adjacent channels attenuation, and a flat group-delay response. These requirements can be satisfied by a proper choice of the filter's order and by a convenient use of the available poles of its transfer characteristic.

Usual approach for such requirements is to consider band-pass filters with quasi-elliptic Chebyshev responses. For such filters, the number of available poles is equal to the filter's order. Some of these poles (real zero-transmission frequencies) can be used to improve the adjacent channels attenuation. Part of the remaining poles (complex-conjugate zero-transmission pairs) can be used to improve the group-delay response of the filter. The proper choice of these equalization pairs can be done using an appropriate optimization procedure.

As a design example, a sixth-order quasi-elliptic filter with an in-band Chebyshev response, having two prescribed attenuation poles and two optimized group-delay equalization poles is designed, simulated, fabricated and tested. The measured characteristics of the experimental model are found to be in good agreement with the results anticipated through electromagnetic-field simulation.

2 Problem Formulation

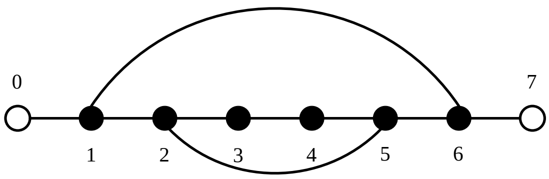
The optimization procedure applied in this paper, based on the Nelder-Mead algorithm [1], was implemented as a Matlab® application.

Based on the position of all poles, a standard synthesis procedure [2] can be applied to get the generalized coupling matrix in its canonical form. Then, using similitude transformations [2], the coupling matrix for the considered topology of the filter can be found.

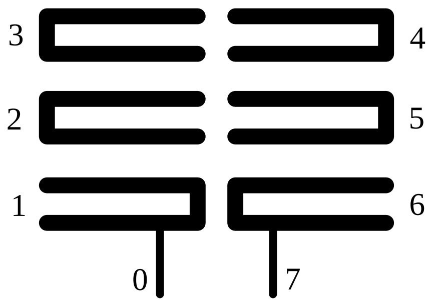
As a design example, a sixth-order quasi-elliptic microstrip band-pass filter with Chebyshev response in the pass-band, having two prescribed attenuation poles and two optimized group-delay equalization poles was designed, simulated, fabricated and measured. Two of the six poles of the proposed sixth-order Chebyshev filtering structure were used for obtaining a good selectivity with respect to the adjacent channels, while other two complex-conjugated poles were used for getting a flattened group-delay of the filter within 80% of its pass-band.

3 Design and Simulation

The design methods presented in the literature [2], [3] offer the possibility to synthesize normalized coupling matrices **M** for quasi-elliptic filters of different orders, starting from an imposed constant ripple in the pass-band and from specified positions of all transmission zeros. The individual resonance frequencies of the resonators in the designed filter and the necessary couplings between its elements can then be found from **M** through some simple de-normalization procedures.



a) Folded filter's topology



b) Filter's implementation with cross-coupled hairpin resonators

Fig.1. Sixth-order band-pass filter

Based on the topology from Fig.1, a sixth-order microstrip band-pass filter was designed, simulated, fabricated and tested. The designed filter has the next specification: center frequency f_0 = 3.5 GHz, bandwidth B = 140 MHz (fractional bandwidth w = 4 %, with cut-off frequencies f_1 = 3.430 GHz and f_2 = 3.570 GHz), a Chebyshev response with a constant ripple R = 0.0436 dB (corresponding to an in-band reflection loss L_R = 20 dB), and 50 Ω loads. The filter was designed to present two attenuation poles at the normalized values $f_{z1} = -2$ (3.36 GHz) and $f_{z2} = +2$ (3.64 GHz) in the two adjacent channels.

A flattened group-delay response of the filter in its pass-band can be obtained by using pairs of complex-conjugated transmission zeros. In this sense, two of the remaining available four poles of the designed filter were used to achieve an improved group-delay characteristic of its response. Unlike de imposed – therefore known – positions of the two attenuation poles used for getting an improved selectivity, the positions of the complex-conjugated transmission zeros, used for getting an optimized group-delay, are not known and are difficult to be obtained analytically. In this paper the optimum positions of these transmission zeros were found using a numerical optimization procedure based on the Nelder-Mead algorithm.

The Nelder-Mead optimization algorithm differs from other deterministic methods as it does not need one-dimensional optimization algorithms, as part of the computing strategy. It requires only function evaluations (no derivatives) and, therefore, it is a zero-order method. It is very effective in terms of computational effort measured in number of function evaluations [1].

This algorithm is used in our design to achieve the best approximation (in the mean-square sense) of the ideal constant group-delay characteristic. Practically, the goal of the optimization was to get a group-delay response as flat as possible within 80% of the filter's pass-band. This goal was achieved using a Matlab® application based on the Nelder-Mead algorithm. The procedure led to the following optimized positions of the complex-conjugated transmission zeros: $f_{z3} = -0.0000334 + 0.8809176i$ and $f_{z4} = -0.0000334 - 0.8809176i$. With these values, the group-delay peak-to-peak variations within the specified frequency band (80% from pass-band) decreased from 28.5 % to 5.8 %.

With another Matlab® application based on the methods presented in [2] and in [3], a synthesis procedure starting from the above specification leads to the following normalized coupling matrix M, in its canonical form:

$$\mathbf{M} = \begin{bmatrix} 0 & 0.3249 & -0.4646 & 0.4289 & -0.4289 & 0.4646 & -0.3249 & 0 \\ 0.3249 & 1.2138 & 0 & 0 & 0 & 0 & 0 & 0.3249 \\ -0.4646 & 0 & 0.9416 & 0 & 0 & 0 & 0 & 0.4646 \\ 0.4289 & 0 & 0 & 0.3094 & 0 & 0 & 0 & 0.4289 \\ -0.4289 & 0 & 0 & 0 & -0.3094 & 0 & 0 & 0.4289 \\ 0.4646 & 0 & 0 & 0 & 0 & -0.9416 & 0 & 0.4646 \\ -0.3249 & 0 & 0 & 0 & 0 & 0 & -1.2138 & 0.3249 \\ 0 & 0.3249 & 0.4646 & 0.4289 & 0.4289 & 0.4646 & 0.3249 & 0 \end{bmatrix}$$

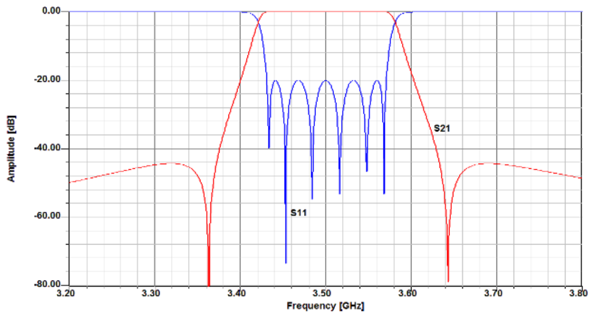
Finally, a general coupling matrix $\mathbf{M}'$ corresponding to the topology illustrated in Fig.1 was obtained, after several similitude transformations applied on **M**:

$$\mathbf{M}' = \begin{bmatrix} 0 & 1.0054 & 0 & 0 & 0 & 0 & 0 & 0 \\ 1.0054 & 0 & 0.8485 & 0 & 0 & 0 & -0.0361 & 0 \\ 0 & 0.8485 & 0 & 0.6104 & 0 & 0.1106 & 0 & 0 \\ 0 & 0 & 0.6104 & 0 & 0.5070 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0.5070 & 0 & 0.6104 & 0 & 0 \\ 0 & 0 & 0.1106 & 0 & 0.6104 & 0 & 0.8485 & 0 \\ 0 & -0.0361 & 0 & 0 & 0 & 0.8485 & 0 & 1.0054 \\ 0 & 0 & 0 & 0 & 0 & 0 & 1.0054 & 0 \end{bmatrix}$$

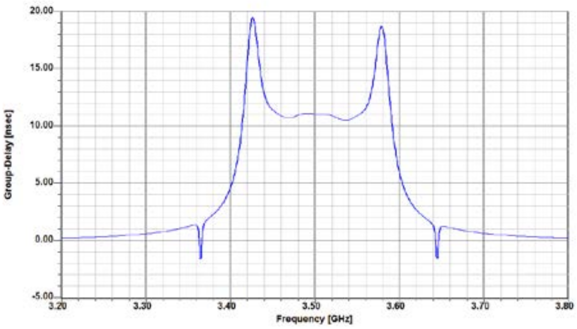
The results of this design procedure was tested and validated by considering a lumped-elements circuit model, composed of six ideal L–C resonators

ISBN: 978-1-61804-131-9

and ideal inverters as coupling elements. As can be seen in the Fig.2, the responses of this model meet near perfectly, as expected, the design requirements, validating the synthesis results. These simulated responses will be used later as terms of comparison for the microwave design.



a) Transmission (S_{21}) and reflection (S_{11}) amplitudes



b) Group-delay response

Fig.2. Simulated performances of the band-pass filter's lumped-elements model

The microstrip band-pass filter was designed on a Rogers3003® 0.508 mm-thick dielectric laminate substrate, having a dielectric constant ε_r of 3, a loss tangent tanδ of 0.0013, and a 0.017 mm thickness double copper metallization.

The microwave design of the filter starts by choosing the shapes of microstrip resonators to accommodate the coupling topology from Fig.1. All six resonators are synchronously-tuned on 3.5 GHz, the center frequency of the band-pass filter. The accurate dimensions of these resonators were found through electromagnetic-field simulation [5].

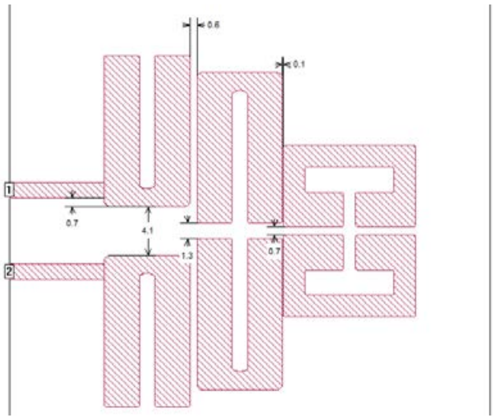
The necessary couplings between resonators, k_{ij}, and loaded quality factors Q of the resonators 1 and 6, as specified by $\mathbf{M}'$, are found through de-normalization:

$$k_{ij} = wM'_{ij},\ i, j \in \{1, 2 \ldots 6\},\ i \neq j,$$
$$Q_1 = \frac{1}{w\left(M'_{01}\right)^2},\ Q_6 = \frac{1}{w\left(M'_{76}\right)^2}, \qquad (1)$$

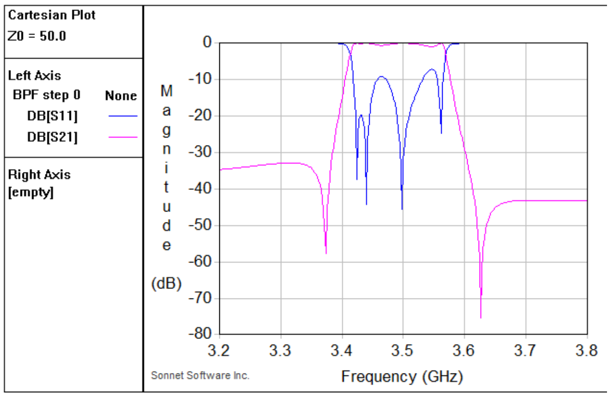
leading to the following numerical values:

$k_{12} = k_{56} = 0.03394$;
$k_{23} = k_{45} = 0.02442$;
$k_{34} = 0.02028$;
$k_{25} = 0.00442$;
$k_{16} = -0.00144$;
$Q_1 = Q_6 = 24.73$.

As can be noticed, the coupling coefficient k_{16} has a negative sign, which means that this coupling must be of opposite type than couplings 2 – 5 and 3 – 4 (see Fig.1.b). Consequently, the coupling 2 – 3 and 4 – 5 should be of the type–II mixed coupling [6]. In practice, such couplings should be avoided because this type of mixed coupling is very weak and does not depend monotonically on the distance between the two coupled resonators. For these reasons, in this design the shape of the resonators 3 and 4 were modified to get a mixed coupling slightly different from the typical type–II mixed coupling (see Fig.3.a).



a) Microstrip layout (dimensions in mm)



b) Simulated responses

Fig.3. The designed filter

The design started by considering a layout with a configuration similar to that shown in Fig.1.b. The accurate dimensions of the two types of resonators were found by electromagnetic-field simulation [6].

A similar procedure was applied to find the relative position of the two feed-lines with respect to the input and the output resonators, in order to obtain the required external quality factors Q, as well as the relative position of each pair of coupled resonators, in order to obtain the required coupling coefficients k.

The layout of this filter and its lossless-simulated response, are shown in Fig.3. It can be noticed that its performances are not as good as expected.

To improve the design, it is necessary to identify the errors in the individual resonance frequencies and in the couplings in the layout of Fig.3, in order to correct them. Indeed, up to this point, each element and each pair of elements from the filter's structure was studied and designed independently. In fact, when put together, all these parameters are slightly influenced by the presence of neighboring elements. When the whole structure of the filter is assembled, its parameters do not correspond any more exactly to the design intentions, so that some adjustments are needed.

The errors caused by all these influences can be corrected by applying a CAD-based optimization procedure that combines the accuracy of electromagnetic-field simulation with the speed and convenience of linear circuit optimization [7], [8]. In this sense a number of extra-ports are attached to the layout resonators. This transforms the layout into a 6 + 2 = 8 port device.

The values of the external elements found through this optimization procedure represent, in fact, the actual errors in the resonance frequencies and in the coupling values in the layout.

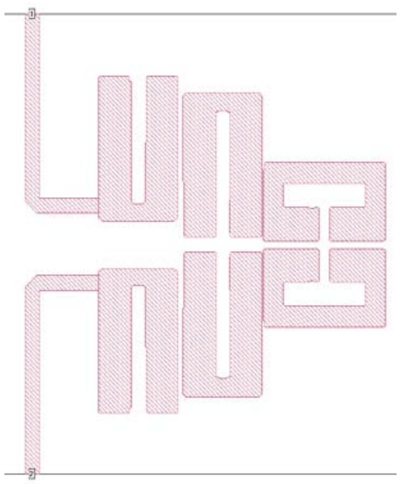
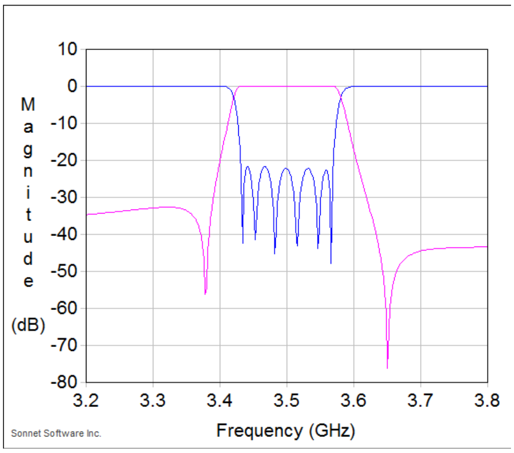


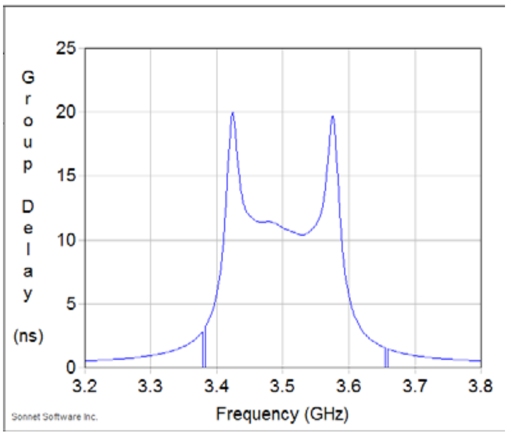
Fig.4. Layout of the optimized filter

Once these errors identified and quantified, they can be easily corrected resuming the analysis on individual resonances and couplings with the *em*-field simulator [5]. Consequently, the errors are eliminated by introducing small corrections in the filter’s geometry. If necessary, this hybrid optimization procedure can be resumed with a new iteration.

The optimized layout of the filter is shown in Fig.4 and its characteristics are presented in Fig.5. The effect of the optimization procedure on the filter’s layout can be noticed here as small corrections on the shapes of resonators.



a) Transmission (red) and reflection (blue) characteristics



b) Group-delay characteristic

Fig.5. Simulated performances of the optimized lossless band-pass filter from Fig.4

It can be noticed that the simulated characteristics of the optimized filter are in good agreement with the design specification.

4 Experimental Results

The filter from Fig.4 was fabricated in a classical microstrip technology. The experimental model is shown in Fig.6.

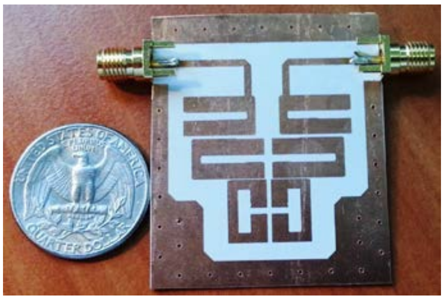
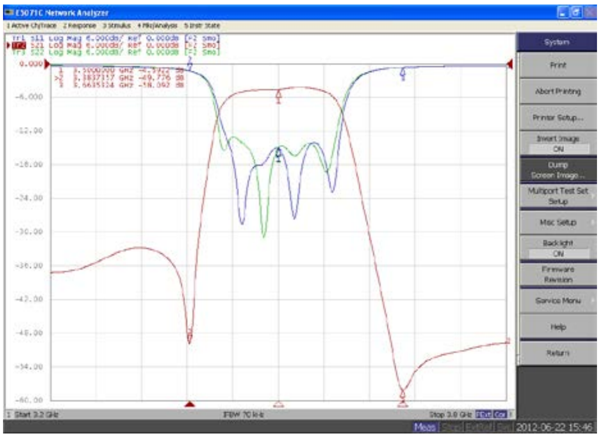
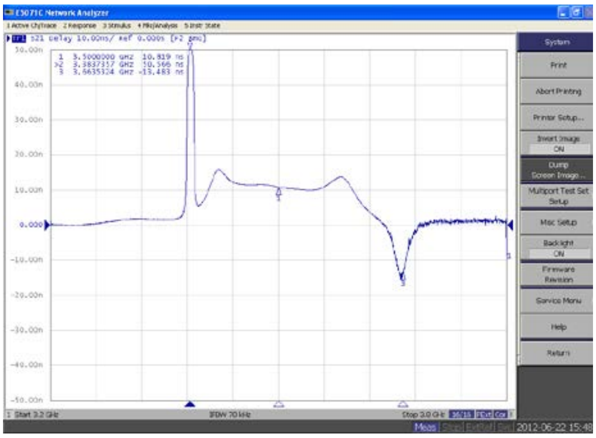


Fig.6. Experimental model of the designed filter

The experimental model was tested with an Agilent E5071C VNA. Its measured characteristics are shown in Fig.7.



a) Measured characteristics (S_{11}, S_{21}, S_{22})



b) Measured group-delay response

Fig.7. Performances of the experimental model

The experimental model exhibits a center frequency of around 3.5 GHz, an in-band insertion loss of about 4.5 dB, and a return loss better than 15 dB. The measured 3 dB-bandwidth is of about 140 MHz, as expected.

The group-delay response from Fig.8.b shows a good flatness within 80 % of the filter's pass-band, as required by the specification.

5 Conclusion

This paper demonstrates the possibilities of designing low-cost planar filtering structures with improved performances in terms of adjacent-channel attenuation and group-delay control. In the design example, the goal was to get a flat group-delay characteristic within a specific frequency range in the center of the pass-band. A similar procedure can be applied to obtain filters with different shapes of their group-delay responses, designed for channel group-delay equalization.

In this design an optimization procedure was applied to control the magnitude characteristics S_{21} and S_{11}. Future work will be focused on an optimization procedure concerning both magnitudes and group-delay characteristics.

The performances of the experimental model are relatively close to the design values, despite of the available low-cost technology, not very accurate.

The proposed filter structure is well-suited for miniaturization.

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