Accurate Computer Aided Design of Interdigital Filters Applying a Coupling Identification Method

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Abstract — **Interdigital filters are found to be difficult to adjust because of improper couplings that occur at high frequency. To solve this problem, a new optimization method for the design of high frequency interdigital filters is presented in this paper. This optimization method uses an accurate computer aided design method, based on the identification of each resonator coupling including parasitic ones. In order to show the feasibility of such a method an example of a 3.1GHz band-pass filter around 18.95GHz is studied.**

I. INTRODUCTION

Since sixties, interdigital and combline filters have attracted interests of many researchers [1]-[5]. These types of filters present advantages of low cost, compact size and adaptability to narrow and wide band applications. Nowadays, the microwave industry still shows a lot of consideration for these filters, particularly in microstrip technology that can be easily integrated. However, interdigital filters become difficult to adjust because of improper couplings that occur at high microwave frequencies. So, analytical methods [1]-[4], that do not consider parasitic couplings become limited to design efficiently such filters at high frequencies. In regards to this problem, the new tendency is to integrate electromagnetic simulators in optimization loop, applying space mapping [6], neural networks or other relevant numerical techniques. For example, the space mapping technique uses a fast analytic model, often called "coarse model" which describes the basic behavior of the filter ignoring subtle aspects like parasitic couplings. The coarse model can be properly used to determine the starting point in optimization. Furthermore it could be applied to drive the electromagnetic simulator to converge towards the solution. After each electromagnetic computation, the coarse model optimization parameters are adjusted to rectify the previous error.

The optimization method presented in this paper, is directly derived from a procedure which has already been successfully used to design waveguide filters [7]. This method combines an electromagnetic analysis, a rational approximation and a general lumped elements synthesis. After each electromagnetic analysis, all the couplings between resonators, including parasitic couplings are estimated from the spectral response. It becomes easy in such a case to relate the coupling values to the structure dimensions. The optimal structure dimensions may be finally determined applying an appropriate coupling synthesis.

In this paper, the design technique is applied to a five pole microstrip interdigital filter with a 3.1 GHz bandwidth around 18.95 GHz. The structure is presented in Fig. 1.

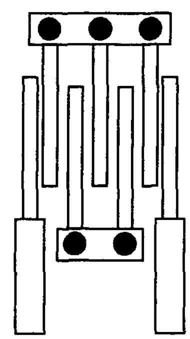

Fig. 1. Sketch of a microstrip interdigital filter

In the first part, the initial dimension synthesis is described. Then the optimization procedure, considering the parasitic couplings is presented.

II. INITIAL DIMENSION SYNTHESIS

This study begins with the G.L.Matthaei's straightforward procedure [1]. It enables to obtain the dimensions of a stripline interdigital filter with a narrow or moderate bandwidth. In our case, the procedure is applied to the design of a microstrip interdigital filter. Stripline and microstrip structures are presented in Fig. 2.

The G.L.Matthaei's synthesis starts with the transformation of the desired band-pass filter into a low-pass prototype, which is modified with appropriate inverters to result in a totally capacitive filter. These capacities described in Fig. 2, are calculated with respect to the filtering pattern. Finally, the geometrical stripline dimensions are determined using tables.

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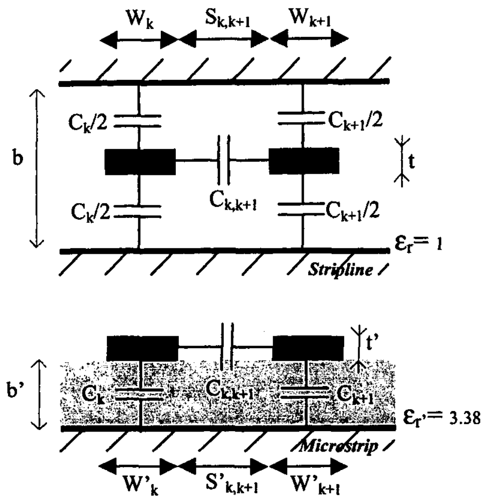

Fig. 2. Sketch of stripline and microstrip structure

In our example, the five pole filter has a 3.1GHz bandwidth around 18.95GHz with a 0.1dB ripple. The filter is synthesized considering a 7% larger bandwidth as suggested in the procedure. The synthesized capacities and stripline dimensions are presented in Table 1. We remind that each resonator is a quarter wavelength long at the midband frequency.

k	$C_{k,k+1}/\varepsilon$	$S_{k,k+1}/\varepsilon$
0 and 5	2.034	41 μm
1 and 4	0.501	278 μm
2 and 3	0.382	347 μm

k	C_k/ε	W_k
0 and 6	5.5	890 μm
1 and 5	2.636	402 μm
2 and 4	3.773	524 μm
3	3.751	502 μm

($\varepsilon_r'=3.38$, $b=817\mu m$, $t=17\mu m$)

Table 1. Capacities and stripline dimensions

Moreover, the G.L.Matthaei's procedure provides the ideal couplings between resonators, that define the following objective coupling matrix [8], [9]:

$$R_{IN}=R_{OUT}=0.934$$

$$[M_{OBJ}]=\begin{bmatrix} 0 & 0.797 & 0 & 0 & 0 \\ 0.797 & 0 & 0.607 & 0 & 0 \\ 0 & 0.607 & 0 & 0.607 & 0 \\ 0 & 0 & 0.607 & 0 & 0.797 \\ 0 & 0 & 0 & 0.797 & 0 \end{bmatrix}$$

Knowing, the ideal capacities, the microstrip geometrical dimensions have now to be found. An electromagnetic software based on the finite element method is applied in two dimensions to match the geometrical microstrip dimensions with the ideal capacities. The microstrip dimensions are presented in Table2. The resonators are still a quarter wavelength long.

k	$C_{k,k+1}/\varepsilon$	$S_{k,k+1}/\varepsilon$
0 and 5	2.034	83 μm
1 and 4	0.501	295 μm
2 and 3	0.382	357 μm

k	C_k/ε	W'_k
0 and 6	5.5	328 μm
1 and 5	2.636	109 μm
2 and 4	3.773	161 μm
3	3.751	151 μm

($\varepsilon_r'=3.38$, $b'=817\mu m$, $t'=17\mu m$)

Table 2. Capacities and microstrip dimensions

The G.L. Matthaei's synthesis neglects non-adjacent couplings, nevertheless it enables to find a good design starting point.

The Agilent's Momentum software, based on the method of moments, is used to compute the filter response considering the initial dimensions given in Table 2. The initial response of the filter is presented in Fig. 3.

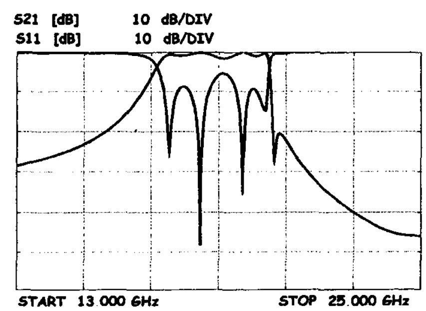

Fig. 3. Initial response

Obviously, the previous result is not optimal and a transmission zero appears due to parasitic couplings.

To improve the filter frequency response, an optimization method based on the identification of each resonator couplings is proposed.

III. Parasitic Coupling Identification

At this high working frequency, strong parasitic couplings make the optimization difficult. The identification procedure allows us to estimate all the couplings, including parasitic ones.

In the first step, the computed [S] parameters are approximated applying a rational approximation method. Then, the [S] parameters are expressed as follow:

$$[S] = \frac{1}{Q}\begin{bmatrix} P11 & R12 \\ R21 & P22 \end{bmatrix}$$

where $P11, R12, R21, P22$ and Q are frequency dependent polynomials.

Afterwards, a lumped element synthesis method [8]-[9] based on an appropriate equivalent circuit is applied to determine the coupling matrix. Many coupling matrices are suitable, but the preserved solution has to satisfy the structure geometry. Indeed, the coupling matrix topology is related to the non zero couplings.

In our case, the desired matrix has to satisfy the following coupling diagram:

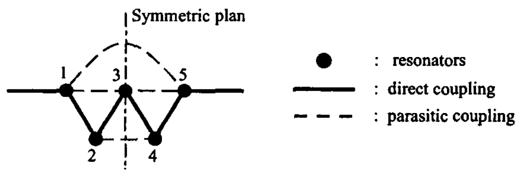

Fig. 4. Coupling diagram of the microstrip filter

The symmetry of the structure forces the matrix to be symmetrical about the second diagonal. Consequently the coupling matrix has the following topology:

$$[M] = \begin{bmatrix} m_{11} & m_{12} & m_{13} & 0 & m_{15} \\ m_{12} & m_{22} & m_{23} & m_{24} & 0 \\ m_{13} & m_{23} & m_{33} & m_{23} & m_{13} \\ 0 & m_{24} & m_{23} & m_{22} & m_{12} \\ m_{15} & 0 & m_{13} & m_{12} & m_{11} \end{bmatrix}$$

Giving these different conditions, we look for such a solution applying an appropriate sequence of rotations [8], [9]. The extracted coupling matrix that corresponds to the initial electromagnetic analysis is then:

$$R_{IN} = R_{OUT} = 0.560$$

$$[M] = \begin{bmatrix} -0.002 & 0.618 & -\underline{0.411} & 0 & -\underline{0.122} \\ 0.618 & 0.208 & 0.356 & -\underline{0.317} & 0 \\ -\underline{0.411} & 0.356 & 0.693 & 0.358 & -\underline{0.412} \\ 0 & -\underline{0.317} & 0.358 & 0.208 & 0.619 \\ -\underline{0.122} & 0 & -\underline{0.412} & 0.619 & 0 \end{bmatrix}$$

$\underline{X.XX}$: *parasitic coupling values*

This coupling matrix is completely different from the expected one. Particularly, the level of parasitic couplings is high. Such couplings cannot be neglected. Moreover, none of the geometrical dimensions enables to eliminate them directly. Our experience shown that parasitic couplings are also very sensitive to the geometrical dimensions. As a consequence, a complementary procedure is necessary to design the filter, taking into account parasitic couplings.

IV. Optimization Procedure

An optimization procedure is proposed to define an ideal response that takes parasitic couplings into account.

<u>First step</u>: *Equivalent model definition*

A sensitivity analysis is performed on the structure dimensions around the initial values. After each electromagnetic analysis, the coupling matrix is extracted. Consequently, all the coupling coefficients, including parasitic ones may be expressed as functions of the geometrical dimensions. Only a few time of electromagnetic analyses are necessary to determine the relations between the coupling coefficients and the geometrical dimensions. A lumped element equivalent model that integrates these relations is formed. This equivalent model can describe the variations of each coupling as a function of the filter dimensions.

<u>Second step</u>: *Geometrical dimension correction*

Afterwards, the equivalent model response is optimized using a circuit software. As a consequence, this step provides a new optimal set of geometrical dimensions. The optimized response of the equivalent model is presented in Fig. 5.

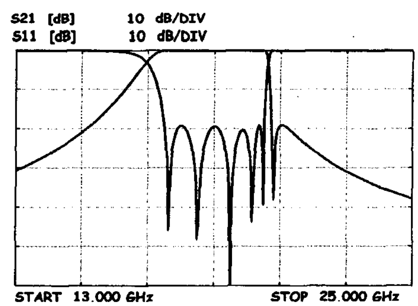

Fig. 5. Optimized equivalent model response

The new ideal coupling matrix is then:

$$R_{IN} = R_{OUT} = 0.997$$

$$[M] = \begin{bmatrix} -0.061 & 0.720 & -0.450 & 0 & -0.130 \\ 0.720 & 0.102 & 0.382 & -0.402 & 0 \\ -0.450 & 0.382 & 0.760 & 0.382 & -0.450 \\ 0 & -0.402 & 0.382 & 0.099 & 0.720 \\ -0.130 & 0 & -0.450 & 0.720 & -0.061 \end{bmatrix}$$

The set of optimal dimensions is given in Table 3.

k	W_{final}
0 and 6	328 μm
1 and 5	109 μm
2 and 4	161 μm
3	151 μm

k	$S_{final\ k,k+1}/\varepsilon$
0 and 5	41 μm
1 and 4	268 μm
2 and 3	325 μm

Table 3: Final microstrip dimensions

One can note that the line widths remain unchanged. However, the resonator lengths have been modified with respect to the quarter wavelengths:

$$L_k = \frac{\lambda}{4} + \delta L \quad \text{with}$$

k	δL
0 and 6	0 μm
1 and 5	-26 μm
2 and 4	7 μm
3	14 μm

Third step: *Electromagnetic verification*

The electromagnetic analysis is performed considering the previous geometrical dimensions.

The electromagnetic response that is presented in Fig. 6 is in good agreement with the equivalent model one.

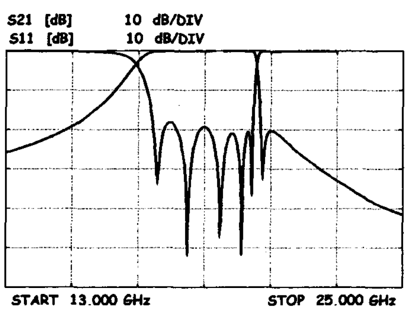

Fig. 6. Electromagnetic verification

The coupling matrix is finally extracted from the electromagnetic response:

$$R_{IN} = R_{OUT} = 0.994$$

$$[M] = \begin{bmatrix} -0.055 & 0.719 & -0.447 & 0 & -0.130 \\ 0.719 & 0.099 & 0.378 & -0.407 & 0 \\ -0.447 & 0.378 & 0.760 & 0.379 & -0.448 \\ 0 & -0.407 & 0.379 & 0.099 & 0.719 \\ -0.130 & 0 & -0.448 & 0.719 & -0.054 \end{bmatrix}$$

One can note that this matrix is close to the predicted one. Consequently, this result validates the equivalent model and the coupling identification method.

V. Conclusion

We have outlined an optimization procedure for high frequency interdigital filters that uses an accurate computer aided design method. This procedure is based on a sharp coupling identification that takes into account the parasitic couplings. The method has been applied to design a 5 pole interdigital microstrip filter with a 3.1 GHz bandwidth around 18.95GHz. The experimental realization is still in progress and the measurements will be presented in the symposium.

The present method may be generalized to design a wide range of microwave filters, especially when improper couplings affect the behavior.

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