# MODELLING STRUCTURE PARASITICS IN COMB-LINE FILTERS

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## ABSTRACT

Combline filters frequently show wider bandwidth then initially synthesized and analyzed. The main cause, as presented in the paper, is distortion of the TEM electromagnetic fields at the resonators open ends. These distortions depend on the three dimensional structure of this region, requiring heavy computation for 3-D field analysis.

A modification to the classic equivalent network of combline filters , which represents this effect is described and formulas to calculate its elements are given, which enables good approximation of the filter's actual performance by using inexpensive network analysis tools.

## PART 1: BACKGROUND

Combline filters are today widely used in commercial communication systems, either in metal or dielectric monoblock technology. The purpose of this work, which enhances previous works by the authors [9,10], is to represent 3-D field distortions in combline filters as electrical equivalent networks added to the classic equivalent networks of these filters. Thus, by simple network analysis tools a better design accuracy is achieved, requiring minimal or no design cycles to overcome bandwidth inaccuracy familiar to combline filter designers. G.L. Matthaei reported in [1] a 16% increase in measured bandwidth, compared to the design (i.e. 0.116 instead of 0.1). By applying the model suggested bellow, a similar bandwidth increase was calculated.

## PART 2: INTRODUCTION

Combline and interdigital filters, are synthesized and analyzed assuming that the filter structure is a finite length section out of an infinite length multi coupled TEM line array with appropriate end conditions (i.e. ground shorts and open ends reactive loading) . Therefor, the mathematical approach of classic synthesis and analysis of these filters, is done assuming that the Maxwellian capacitance matrix is uniform through the entire length of the resonators. The classic equivalent networks of these filters were derived based on the validity of this assumption [1-3,7].

Practically there are distortions in the electromagnetic field lines pattern at the open ends of the resonators. In this region the electrical field pattern is not perpendicular to the resonators [fig. 1], creating a TM propagation mode. As for the short circuit ground plane, it is proven in ref. 10&11, by using mirror image techniques around that ground plane, that the analyzed TEM propagation in the resonators is not affected by that ground plane.

The approach presented herein approximates the 3-D field problem as a deviation from the original TEM values. Therefor the values of the Maxwellian capacitance matrix vary along the coupled lines towards the open end region of the resonators.

Parameters of filters realized using these structures, such as coupling coefficients and slope parameters, will be varied as well, affecting the bandwidth and the ripple of these filters. Combline filters are highly sensitive to this effect, resulting mainly in bandwidth expansion, sometimes up to twice the synthesized (or analyzed) bandwidth. The classic equivalent networks do not include representation of this phenomenon, thus analysis of these networks will not show this effect.

In this paper, as detailed in part 2, the authors suggest an electrical equivalent network representing the open end field distortion effects in combline structures and formulas to predict the network elements values. The suggested network can be combined easily with the classic network, requiring no change in the original network values. Therefor, the use of the technique presented herein, results in an economical computational tool, which fairly approximates the actual response of the analyzed filter. An experimental combline filter designed based on ref. 1 & 2, was built in three versions , having identical cross sections and length of resonators, but different only in the structure



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of the open end ground plane [fig. 2,3 and 4]. The three versions had three different bandwidths [table 1].

## PART 3: THE MODIFIED EQUIVALENT MODEL

The field pattern distortion at the open end region of the combline structure shown in fig. 1 has two effects on the coupled resonators parameters. The first effect is the increased value of the open end self capacitance . This effect however is negligible, since in practice the total capacitance is tuned to a desired value . The second effect, is the reduction of the mutual capacitance, resulting in the reduction of the electric coupling.

In combline structure, the total coupling coefficient between two resonators is the difference between the "magnetic coupling coefficient " (Cm) and the "electric coupling coefficient" (Ce) [ref. 3-6,10,11].

1) C =Cm -Ce

A comprehensive discussion is given in ref. 10.

R.M. Kurzrok [5] suggested to use metal posts between resonators open ends in order to increase the total coupling by reducing Ce, a method widely used since. Similar reduction of Ce at the open ends region occurs when the electric fields is deflected towards the upper metal ground plane, which usually carries the tuning elements of the filter. This effect can be represented as an additional negative capacitance in parallel with the original Maxwellian mutual capacitance at the resonator open ends [fig. 5].

R. Pregla [ref. 8], used capacitors (positive, of course ) between combline resonator open ends in order to reduce the total coupling and to add elliptic properties to the filter. Such capacitors, either lumped or distributed, can be easily added to the equivalent network of a combline filter, in parallel to the series shorted stub. In order to prevent detuning of the modified equivalent network from it's original center frequency, the added capacitive network is expanded to the PI section shown in fig. 6.

## PART4: CALCULATING ADDED NETWORK VALUES

The parameter which significantly effect the 3-D field patterns at the open ends region are the distance H to the upper metal ground plane and the spacing S between resonators as in fig. 1. The following approximation is based on the assumption that most of the coupling mechanism is concentrated along the center longitudinal cross-section of the combline structure [figure 7] . Therefor a 2-D approximation is performed, giving satisfying accuracy.The odd mode excitation of adjacent resonators, coupled to each other is shown in figure 1. We identify two regions along the resonator where:

(a) The electric field lines are terminated at the adjacent resonators (i.e. perpendicular to the virtual electric wall created by the odd mode excitation), with length La.

(b) The electric field lines from the resonators are terminated at the upper metal ground plane, and at the open end of the resonator with length Lb [figure 1].

The total effect , due to the electric field deflection in region (b) , is presented as C - the value of the negative capacitance in the PI network which is added to the equivalent circuit [figure 5]. This approach also enable to calculate the parameter values of the distributed equivalent network (ZC, Lb) , which is presented in figure 6. Zc for each pair of coupled resonators equals the mutual impedance of that pair:

2) Zc(i,j) = Zij = (Zeven+Zodd)/(Zeven-Zodd)

Lb represents the length of the resonator section from which electric field is deflected. In the distributed equivalent model Lb will be the length of the capacitive stubs of the added PI network [Fig. 6]. For the general case [figure 1], Lb increases with the increase of S and with the decrease of H. For H>S/(2*PI) the end effects become sensitive to the other parameters. For that case H will be considered to equal S/(2*PI). The following general formula complies well with practical measurements:

3) Lb = k1*S - k2*H

k1 and k2 are practical coefficients that may vary with various media used to produce the filters. (e.g., round rods, rectangular bars, printed circuit, dielectric monoblocks etc.)

For air filled filters with equal round rods of 70 ohms impedance level, the practical values of k1 and k2 were equal to ½.3, thus for that case:

4) Lb = (S - H)/2.3

The values of Lb for the mentioned 8 section experimental filters are presented in table 2. Measured data and modified networks analysis data, presented intable 1, show excellent correlation.

## SUMMARY AND CONCLUSIONS:

The above work presents a simple equivalent PI network to aid accurate analysis and synthesis of combline filters using common analysis tools. It is important to mention that the resonator physical spacing S is determined from the normalized value of S/B, where B is the ground plane spacing of the coupled line TEM array. Thus for the same electrical design, different values for B will result in different values of S and different amount of bandwidth expansion.

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| | | Measured bandwidths | | |
|---|---|---|---|---|
| Bandwidth points | Original design analyzed bandwidth | Version 1 H= 0.0mm | Version 2 H= 2.3mm | Version 3 H= 5.0mm |
| 10 db attenuation | 74 MHz | 108 MHz | 102 MHz | 97 MHz |
| 20 db attenuation | 85 MHz | 123 MHz | 117 MHz | 107 MHz |
| 20 db return loss | 63 MHz | 95 MHz | 86 MHz | 82 MHz |

Table 1: Analyzed bandwidths of classical equivalent network Vs. Measured bandwidths of 8 section experimental filter 3 versions (f0=2.35GHz)

| | H = 0mm | H = 2.3mm | H = 5.0mm |
|---|---|---|---|
| S12 = S78 = 12.0 mm ( Zc = 1811 Ohm ) | 5.22mm | 4.22mm | 3.04mm |
| S23 = S67 = 13.9 mm ( Zc = 2731 Ohm ) | 6.04mm | 5.04mm | 3.87mm |
| S34 = S56 = 14.3 mm ( Zc = 2981 Ohm ) | 6.22mm | 5.22mm | 4.04mm |
| S45 = 14.4 mm ( Zc = 3040 Ohm ) | 6.26mm | 5.26mm | 4.09mm |

General structure data: Wall spacing B = 15mm;
Resonators equal diameter d = 5.9mm;
Resonator length = 19.7mm (55 deg. electrical length);

Table 2: Lb Vs. H and S for the 8 section filters

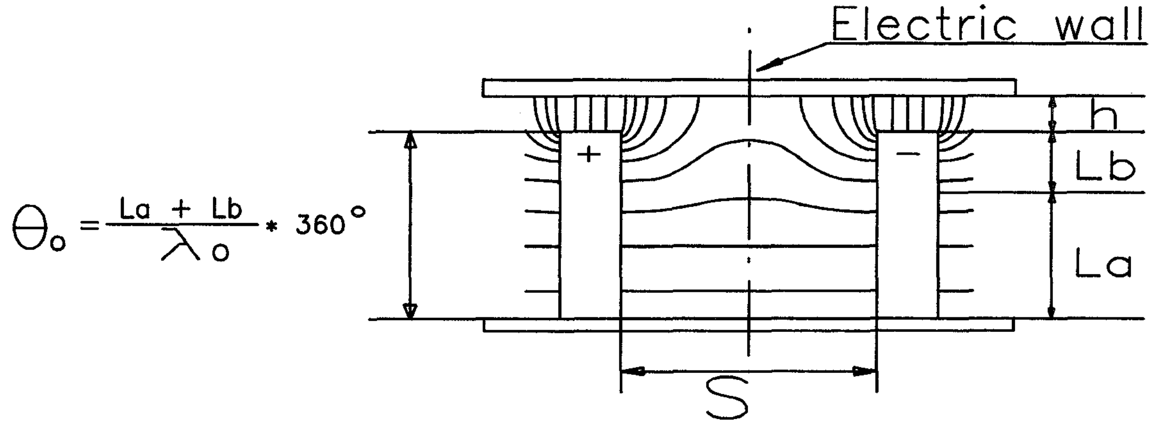


Figure 1 : Electric field pattern for odd mode excitation

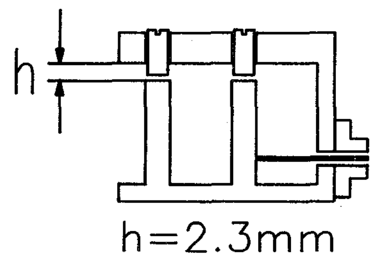


Figure 2

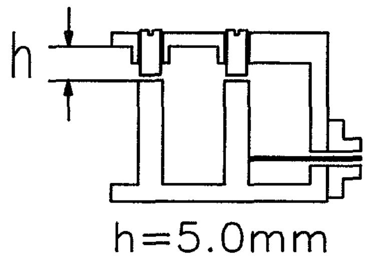


Figure 3

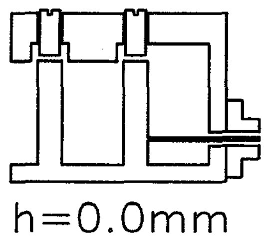


Figure 4

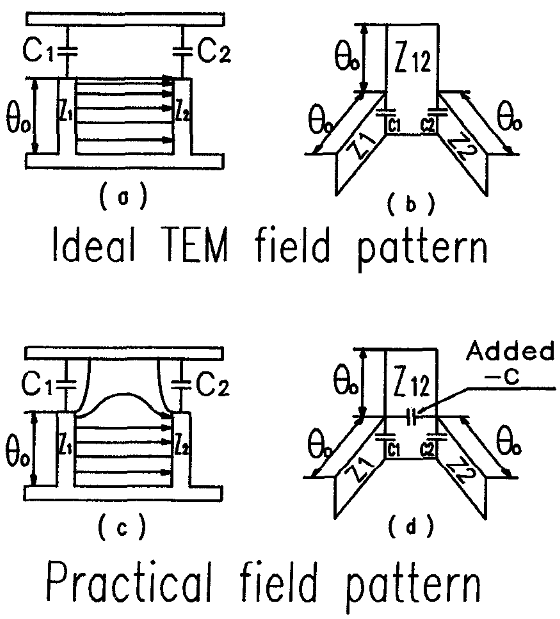


Figure 5

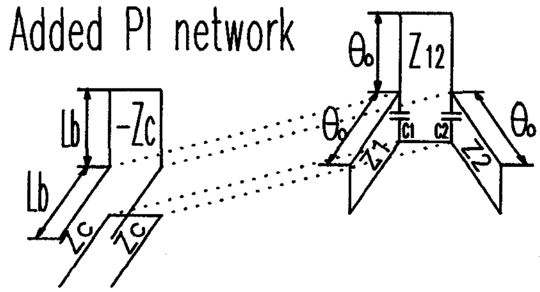


Figure 6

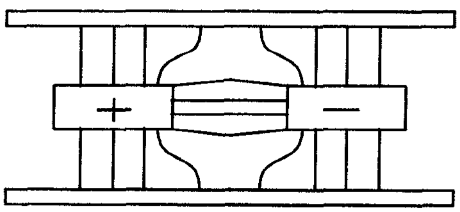


Figure 7