A novel compact coplanar filter

Thomas Paillot, Pierre Blondy, Dominique Cros, Pierre-Yves Guillon

IRCOM, Faculté des sciences, 123 avenue Albert Thomas 87060 Limoges cedex - France

Abstract — **A 30 GHz coplanar two pole filter is presented and consists of two quarter wavelength resonators. It has an original excitation structure based on magnetic coupling. To validate this approach a 2 pole, 10 percent relative bandwidth filter has been fabricated on fused silica. Low loss and good agreement between simulation and measurements have been achieved.**

I. Introduction

With the increasing demand in communications, high frequency systems are more and more complex. The integration of passive and active circuits is a difficult task as density of interconnection is increasing. To face this problem, CPW technology looks very interesting because circuit integration is made easier, compared with microstrip technology. Thanks to their uniplanar structure, coplanar waveguides are quite insensitive to the substrate thickness (unlike their microstrip counterpart), have low dispersive effects, and shunt and series connections are easily done (because there is no need for via hole). Finally, integration of active devices using flip-chip technology is easy.

In spite of all these advantages, relatively little has been done on coplanar filters compared to microstrip, especially at high frequencies [2]-[8]. This paper describes a 30 GHz two pole coplanar bandpass filter with 10% relative bandwidth.

Its design is as simple as possible with only two shunted quarter wavelength resonators feed by smaller coplanar lines [9] but other resonators might be added in the ground plane to achieve higher order filters. An original excitation scheme has been used to couple the energy at the input-output.

First we will see how does this two pole filter works and how has it been designed. Then, the simulation and measurements results will be presented. It will be shown that there is a good agreement between them validating the input-output coupling structure.

II. Filter design and coupling coefficients

A photograph of the filter is shown in fig. 1. The filter is made of two quarter wavelength resonators put in front of each other and capacitively coupled [10]-[11]. They are excited by a section of short circuited coplanar line. The first section is a 50 ohms feeding line, of length l_3 (fig. 1). It is followed by a larger section whose width w_2 and length l_2 are optimized to obtain the right external quality factor Qext. The coupling between the resonator and the input line is made by mutual inductance since their magnetic fields are closely coupled in this configuration (fig.2).

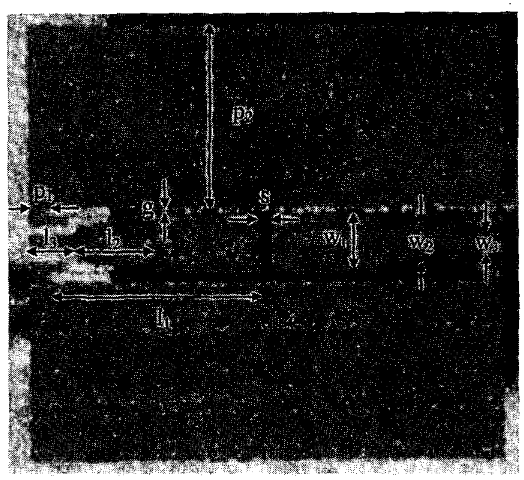

Figure. 1. Photograph of the coplanar filter

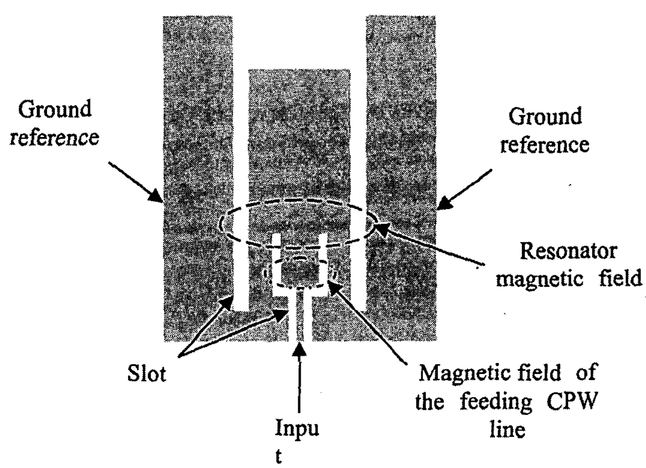

Figure. 2. Coupling between the feeding lines and the resonator

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The ratio between the widths of the input line and the resonator changes Qext as shown in fig. 3. Another way is to change the length of this section: the combination of these two dimensions allows to achieve a wide range of Qext values thus this coupling structure might be used in wide or narrow band applications. For the present example, the different dimensions are given on table. I.

TABLE I

VALUES OF THE DIFFERENT DIMENSIONS OF THE CIRCUIT

l_1	1330µm
l_2	490µm
l_3	300µm
w_1	400µm
w_2	280µm
w_3	150µm
s	60µm
p_1	145µm
p_2	1350µm
g (slot gaps)	30µm

This bandpass filter has been designed using the filter synthesis technique as detailed by Matthei, Young and Jones [1]. The chosen prototype (f0=29GHz, Δf=3GHz, Ripple=0.1dB) gave the following element values:

g0=1 g1=0.843 g2=0.622 g3=1.3554

Then, we can evaluate the Qext value and the inter-resonator coupling coefficient K12 with the following equations [1]:

$$Q_{ext} = f_0 \cdot g_0 \cdot g_1 / \Delta f = 16.86 \qquad (1)$$

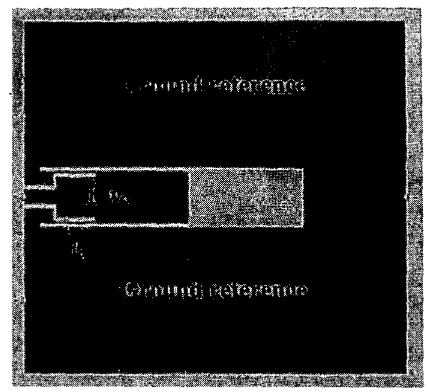
$$K_{12} = \Delta f / (f_0 \cdot \sqrt{g_1 \cdot g_2}) = 0.069 \qquad (2)$$

To approximately find which inter-resonator gap length corresponds to K12 or which coplanar line width corresponds to Qext, two circuits are studied using Agilent Momentum in the slot mode. The circuit designs are shown on figure. 3 and 4.

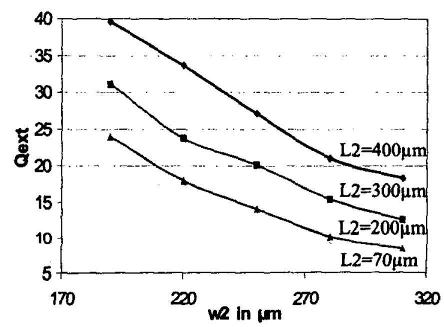
A. Qext (fig. 3.)

To determine the size of the feeding line that leads to the right Qext, the distributed circuit shown fig. 3(a). is simulated. By changing the distances w2 or l2, the S11 changes and by applying equation (3), the evolution of Qext as a function of the coplanar line width and length can be found. Then we find the right couple (w_2, l_2) which gives the required Qext (figure 3. (b)).

$$Q_{ext} = (f_{\angle f_0 - 90^\circ} - f_{\angle f_0 + 90^\circ}) / f_o \qquad (3)$$



(a)

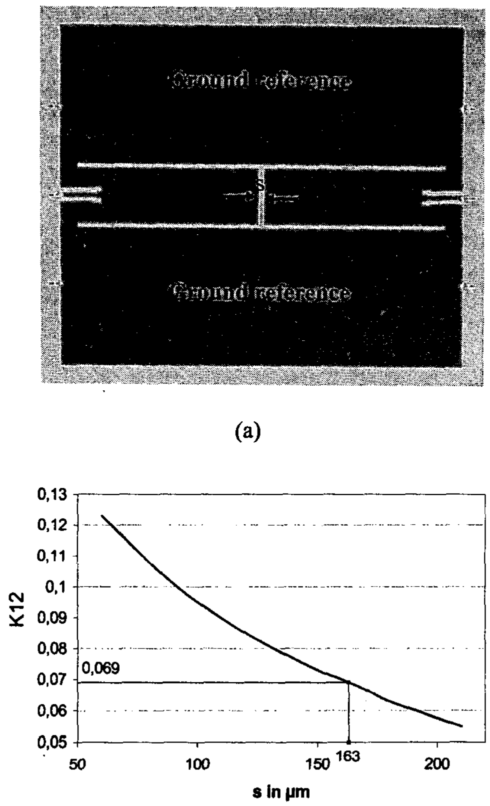


(b)

Fig. 3. CPW open-end resonator (a) and (b) Qext as a function of w and L.

B. K12 (fig. 4.)

To determine the gap length between the two resonators which gives the correct K12, the circuit shown fig. 4(a). is simulated. This circuit is almost the same as the final filter but is de coupled (the feeding lines are very small) in order to have a practically unloaded response. By changing the gap width between the two resonators and applying equation (4), the K12 function (fig. 4(b)) of the gap width s can be drawn so as to find the correct value of s.



(a)

(b)

Fig. 4. Unloaded two poles resonator (a) and (b) the K12 evolution as a function of s

$$K_{12} = (f_{om}^{2} - f_{oe}^{2}) / (f_{om}^{2} + f_{oe}^{2}) \qquad (4)$$

where f_{om} is the odd mode frequency and f_{oe} is the even mode.

Once the dimensions of the filter have been found, the last thing to do is to assemble the entire filter with the predetermined dimensions. Next, the filter dimensions are finely adjusted to get the optimal response. This is why the w2, l2 and s are not exactly the same as the values given before.

The final simulation and measurements are shown in fig. 5.

III. Simulation and Measurements

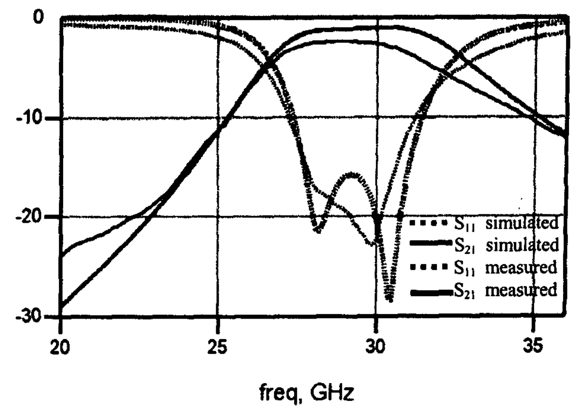


Fig. 5. Transmission and return loss characteristics in dB

The circuit has been realized on a 1mm thick fused silica substrate (ε_r=3.8) which provide low substrate losses and 1.5µm thick gold metallization (ρ=3.9 10e7 S.m). Measurements were taken using an HP 8510C and a cascade probe station. A SOLT calibration was used and the presented measurements were not deembeded. Measured and simulated results are shown on fig. 5. The measured central frequency is 28.9GHz with a 9% fractional bandwidth. It can be seen that they are in good agreement, with measured insertion losses around 2dB and return losses less than 17dB.

IV. Conclusion

This study focuses on a new design for a CPW 30GHz bandpass filter. An original input-output coupling technique has been presented. It has been shown that this coupling scheme allows a very flexible design of microwave CPW filters. We believe it will be useful for the design of various CPW filters. Furthermore, the response could be improved with a higher order filter that could be realized by adding other quarter wavelength resonators in the ground plane.

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