Parallel Coupled Microstrip Filters With Floating Ground-Plane Conductor for Spurious-Band Suppression

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***Abstract*—Floating strip conductors printed on the backside of the substrate are shown to be useful to suppress the spurious transmission band appearing at twice the central frequency of coupled-line microstrip filters. It is shown that proper adjustment of the dimensions of the floating conductors yields equal even and odd electrical lengths. An attractive feature of this design is its flexibility because the equalization of the modal electrical lengths can be achieved with various geometries. Additionally, the floating conductor provides an extra coupling mechanism that relaxes tolerances of strip width and spacing in those cases where tightly coupled sections are required. A fast quasi-TEM analysis is used to find the structure yielding equal mode phase velocities. Fine tuning to equalize the modal electrical lengths for each coupled stage is based on the use of a commercial electromagnetic simulator. Experimental verification is finally provided.**

***Index Terms*—Floating conductor, parallel-coupled microstrip filters, spurious response suppression.**

I. Introduction

PARALLEL coupled line microstrip filters (PCMFs), first proposed in [1], are among the most commonly used filters in microwave and millimeter-wave integrated systems. Some of the advantages of this kind of filters are their simple design procedure, planar character, and the possibility of achieving a wide range of filter fractional bandwidth [2]. However, the microstrip implementation of this kind of filters presents two relatively serious inconveniences. The first one is the degradation of the rejection level above the passband owing to the existence of a spurious passband at $2f_0$ (f_0 being the central frequency of the filter). This undesirable band is caused by the different phase velocities (and, consequently, different electrical lengths) of the even and odd modes supported by the inhomogeneous coupled-line sections. The second problem comes from the weak lateral coupling between the lines in the conventional structure [see Fig. 1(a)]. Strong coupling required by some filter specifications leads to very small values of strip width w and strip spacing s, which may not be accurately fabricated.

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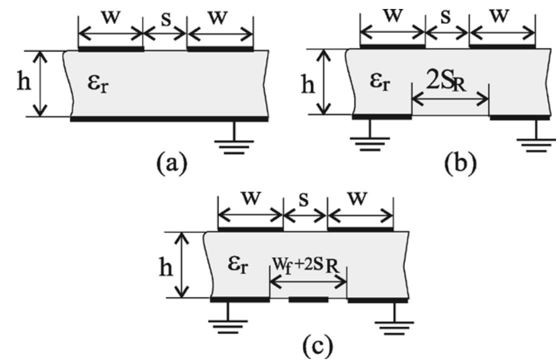


Fig. 1. Cross sections of basic building blocks of microstrip coupled lines for filter design. (a) Conventional. (b) Modified with ground-plane aperture. (c) Modified with floating conductor at the ground aperture.

In recent years, a number of methods have been proposed to suppress the spurious band in PCMFs [3]–[13]. If we center our attention in those papers in which the basic guiding structure is modified with respect to the conventional microstrip, but still preserving translational symmetry, the more recent contributions are those by Kuo and Jiang [9]–[12] and by Velázquez *et al.* [13]. In [9] and [10], the mode phase-velocity matching is achieved via substrate suspension. A similar procedure is employed in [11], but using a uniform dielectric overlay to suppress the spurious band. The technique of increasing the image impedance of the filter sections is combined in [12] with over-coupling at the end stages of the PCMF to improve the spurious-band reduction. In [13], a slotted ground plane in the basic guiding structure [see Fig. 1(b)] is used to pursue a double objective: spurious-band suppression and an increased level of coupling. These two goals can be also achieved with the basic guide structure proposed in this paper, in which a floating conductor strip is inserted in the ground plane slot [see Fig. 1(c)]. The structure with the floating conductor has two advantages with respect to that with a simple slot in the ground plane: the first one is its high flexibility because several (and not only one) geometries for each coupled-line stage match the even and odd impedances required by the filter specifications and the second one is that the presence of a floating conductor allows for mode phase-velocity matching in those cases involving low-permittivity substrates for which the slotted ground-plane solution does not provide a satisfactory performance (equalization of even- and odd-mode electrical lengths is not possible).

The use of a floating ground-plane conductor has been reported in the design of directional couplers with a high coupling level in microstrip [14], as well as in coplanar waveguide

[15], [16]. In this study, the starting hypothesis for a quasi-TEM analysis of the structure shown in Fig. 1(c) is that the total charge on the floating conductor is zero. Thus, the three-conductor transmission-line problem is reduced to an equivalent two-conductor transmission-line problem, one with only two propagation modes with even and odd symmetry. The rigorous three-mode analysis developed in [14] proves that the even/odd-mode theory yields correct results in many practical configurations such as those considered in this paper.

As a first step in the design, for each coupled stage we choose the floating conductor width w_f, and by means of a fast algorithm based on the quasi-TEM code for hybrid microstrip–coplanar-waveguide structures reported in [15], we obtain the corresponding values of w, s, and s_R to match the even and odd phase velocities and simultaneously obtain the modal-impedance values imposed by the filter specifications. However, because of the different influence of the open ends on the even- and odd-mode lengths [13], equalizing the even and odd phase velocities for the coupled lines does not ensure equal even and odd electrical lengths of each filter section. Owing to this fact, we attain this goal making use of an electromagnetic (EM) simulator (Ensemble) for fine tuning of the design. Basically this tuning consists of a perturbation of the length of the slot in which the floating conductor is inserted. Several selected examples are presented to illustrate the design procedure. Experimental confirmation of the theory is also provided.

II. Filter Design

A. *One-Pole Butterworth Filter of* $\Delta = 20\%$*, Substrate With* $\epsilon_r = 10$ *and* $h = 0.635$ *mm*

Let us consider, in order to illustrate the use of the block in Fig. 1(c), the simple case of a Butterworth bandpass filter centered at $f_0 = 2$ GHz having a bandwidth of $\Delta = 20\%$ and order $N = 1$. The substrate has relative dielectric permittivity $\epsilon_r = 10$ and thickness $h = 0.635$ mm. Following [1], for the two identical coupled sections, the filter specifications impose the modal-impedance values $Z_{\text{even}} = 77.67\ \Omega$ and $Z_{\text{odd}} = 38.04\ \Omega$. In this case, as shown in Fig. 2 (design curves obtained with the quasi-TEM approach), for each floating conductor width, we have a geometry that can be selected to fabricate the filter. Note that the value $w_f = 0$ corresponds to the simple slotted ground plane of Fig. 1(b). The flexibility in the geometry of the coupled sections comes from the fact that we have to fulfill *three conditions* (the equalization of mode phase velocities and the matching of the modal impedances with those required by the filter design) while we dispose of *four parameters* to vary: 1) w_f; 2) s_R; 3) , w; and 4) s. As mentioned before, these values can be calculated by means of an optimization procedure based on the quasi-TEM code developed in [15] or obtained from design curves such as those in Fig. 2, which have been generated with the same quasi-TEM code. The optimization strategy has been accelerated taking into account that, for a given value of floating conductor width w_f, the mode impedances are mainly controlled by strip width w and strip separation s, while the mode phase-velocity mismatch mainly depends on the slot width s_R. This fact can be visualized in Fig. 2, where the variations of w, as well as of s are

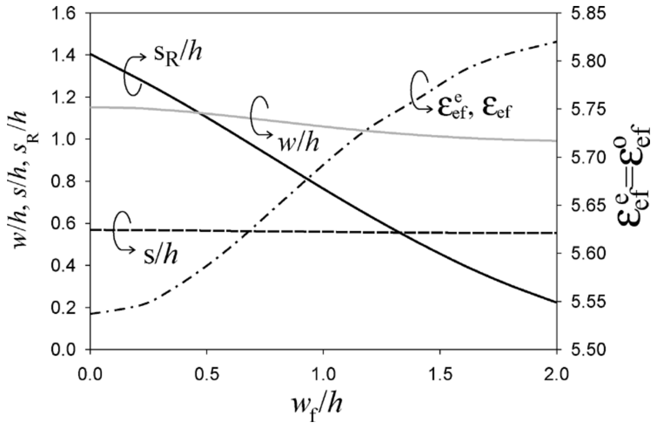
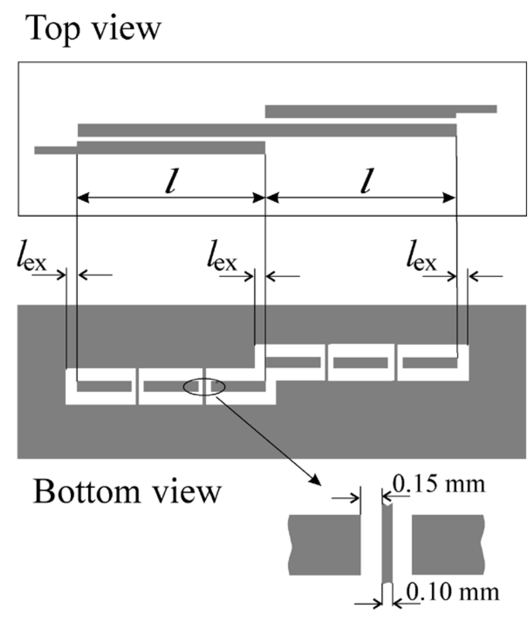


Fig. 2. Normalized values of strip width w/h, strip separation s/h, and slot width s_R/h versus floating conductor width w_f/h that simultaneously satisfy the required modal-impedance values for the coupled-line sections of filter A and provide modal phase-velocity matching. For each geometry, the effective dielectric constant is also shown.



	w_f	w	s	s_R	l	l_{ex}
Modified	0.5	0.689	0.356	0.577	14.50	1.3
Conventional		0.413	0.173		14.69	

Fig. 3. Layout and dimensions (in millimeters) of the designed filter A.

very smooth because all the geometries represented in this figure keep the same modal-impedance values. Thus, we have elaborated a specific-purpose gradient optimization method in which only one of the structural variables is changed in each iteration depending on the associated error function. Also in Fig. 2, we show the effective dielectric constant (equal for both even and odd modes) dependence on floating conductor width. Notice that the effective dielectric constant increases with the floating conductor width, thus resulting in a smaller circuit overall size.

In Fig. 3, we show the final values of the geometry together with a top and bottom view of the layout of the designed filter A. We have arbitrarily chosen the floating conductor width to be $w_f = 0.5$ mm. The other dimensions and parameters can

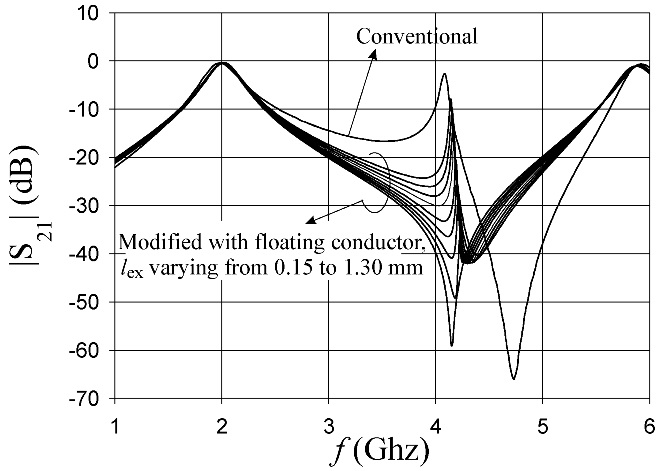


Fig. 4. Insertion loss of the designed filter A for several increasing values of the parameter l_{ex} until complete spurious-band suppression. The values of the other geometric parameters are shown in the table of Fig. 3. The response of the conventional filter is also shown for comparison.

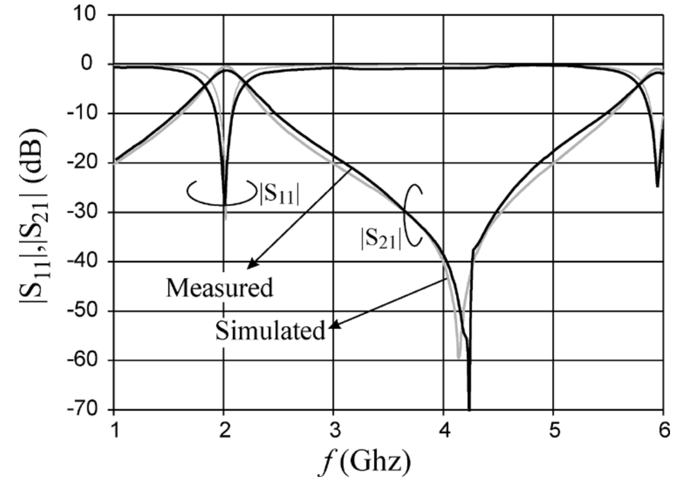


Fig. 5. Simulated and measured results for the final version of the filter design A.

be obtained from the curves in Fig. 2. We have also included the dimensions corresponding to a conventional filter of the same specifications [see Fig. 1(a)]. As expected, the values of w and s for that filter are smaller than for the modified version. Owing to the different edge effects for the odd and even modes of propagation, we have used the EM simulator Ensemble to find the exact value of the section length l that matches the desired central frequency f_0 and the extra length l_{ex} (see Fig. 3) that completely removes the spurious band. The optimization process can be visualized in Fig. 4, in which the insertion-loss response of the filter is shown for different values of l_{ex}. We have also included in Fig. 4 the simulated response of the conventional filter corresponding to the geometrical data shown in the table of Fig. 3. As can be seen from Fig. 4, without l_{ex}, the spurious band is reduced with respect to the conventional configuration of Fig. 1(a), but not eliminated. Complete elimination is achieved by increasing the value of l_{ex}. Therefore, l_{ex} can be viewed as a local perturbation useful to compensate for the different excess lengths for the even and odd propagation modes. The metal bridges joining the ground-plane sides shown in the bottom view of the filter play an essential role for a good filter performance because they cancel out undesired ground-plane slot modes [13]. The exact position of the metal bridges is not very important and we have chosen this parameter so as to divide the floating conductor into three identical sections. Moreover, metal bridges cut off the longitudinal currents on the floating conductor, thus making the hypothesis of null charge on it (used in the determination of the even- and odd-mode parameters) more acceptable. Imposed by the technology limits of our process (photoetching of a metallized plastic substrate), the width of all the bridges has been chosen equal to 0.1 mm and the longitudinal distance between the floating conductors and ground plane has been assigned to be 0.15 mm. Full-wave simulation of the final filter response is compared with measurements in Fig. 5. We have measured the scattering parameters using the HP8510B automatic network analyzer (ANA). As expected, spurious-band suppression is clearly achieved.

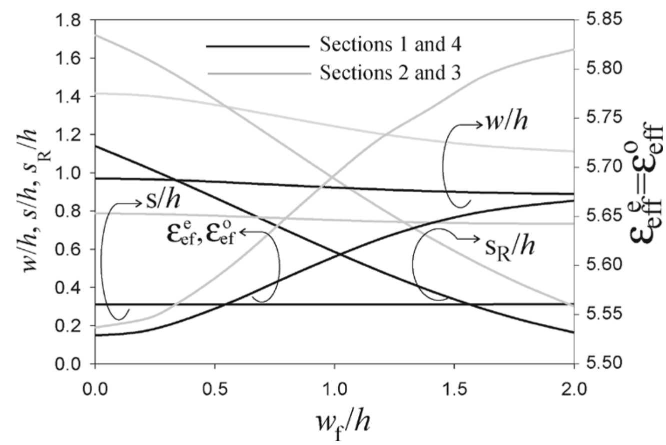


Fig. 6. Normalized values of strip width w/h, strip separation s/h, and slot width s_R/h versus floating conductor width w_f/h that simultaneously satisfy the required modal-impedance values for the coupled-line sections of filter B and provide modal phase-velocity equalization. The effective dielectric constant for each geometry is also shown.

B. Three-Pole Chebyshev Filter of $\Delta = 20\%$, Ripple 0.08 dB, Substrate With $\epsilon_r = 10$ and $h = 0.635$ mm

To extend the described technique to the design of higher order filters, in this second example, we have designed a $\Delta = 20\%$ bandwidth Chebyshev bandpass filter with $f_0 = 3$ GHz, order $N = 3$, and ripple 0.08 dB using the same substrate as in example A. Classical design requires $Z_{\text{even}} = 94.38\ \Omega$, $Z_{\text{odd}} = 37.72\ \Omega$ (sections 1 and 4) and $Z_{\text{even}} = 69.31\ \Omega$, $Z_{\text{odd}} = 39.55\ \Omega$ (sections 2 and 3). In our preliminary design, the section geometries that simultaneously fulfill impedance requirements and equal modal phase velocities can be obtained from the design curves in Fig. 6 or computed with the optimization algorithm. This provides the approximate version of the filter. To improve the filter response of this preliminary version, we have used Ensemble to simulate the response of the one-pole filter that could be constructed based on each section separately to find the length l of the corresponding section for a correct tuning of the central frequency. To obtain the final dimensions, an optimization process similar to that shown in Fig. 4 leads to the optimal extra-length l_{ex} that completely removes

TABLE I
DIMENSIONS (IN MILLIMETERS) OF CONVENTIONAL AND MODIFIED COUPLED-LINE BANDPASS FILTER B

Type of design	Sections 1, 4
Conventional	w =0.294, s =0.101, l =9.88 ϵ_{ef}^{e}=6.74, ϵ_{ef}^{o}=5.56
Modified	w_f =0.91, w =0.469, s =0.197, s_R =0.233, l =9.85, l_{ex} =0.4, $\epsilon_{ef}^{e}=\epsilon_{ef}^{o}$=5.59
Type of design	**Sections 2, 3**
Conventional	w =0.484, s =0.261, l =9.74 ϵ_{ef}^{e}=7.09, ϵ_{ef}^{o}=5.71
Modified	w_f =0.65, w =0.786, s =0.585, s_R=0.602, l =9.20, l_{ex} =0.4, $\epsilon_{ef}^{e}=\epsilon_{ef}^{o}$=5.62

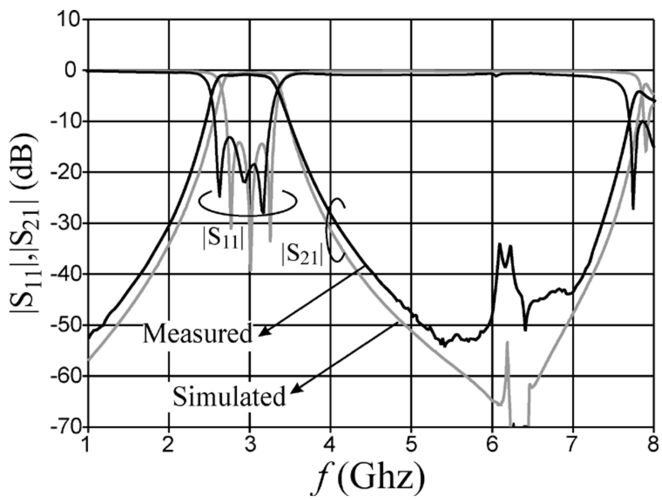


Fig. 7. Simulated and measured results for the final version of filter design B.

the spurious band. These dimensions are shown in Table I. In this table, we have also included the corresponding dimensions of the conventional filter in order to demonstrate again that the small values for strip separation are relaxed in the modified version of the filter. In Fig. 7, we compare Ensemble results for the final design and the measurements of the fabricated filter. Although a spurious band appears in the measured response, its level is below -30 dB. Discrepancies between simulated and measured results probably come from a lack of accuracy in the fabrication process, but it is clear that the spurious band is meaningfully reduced with respect to the one expected for a conventional design.

C. One-Pole Chebyshev Filter of $\Delta = 10\%$ and Ripple 0.5 dB, Substrate With $\epsilon_r = 2.43$ and $h = 0.49$ mm

With this last example, we want to illustrate one of the advantages of the proposed structure over the slotted ground-plane structure previously proposed by Velázquez *et al.* [13]: the possibility of perfect spurious-band suppression in low-permittivity substrate structures for which the slotted ground-plane geometry does not always provide perfect phase-velocity matching. In this case, our goal is the design of a $\Delta = 10\%$ bandwidth Chebyshev bandpass filter with $f_0 = 3$ GHz, order $N = 1$, and ripple 0.5 dB when using a substrate with relative dielectric permittivity $\epsilon_r = 2.43$ and thickness $h = 0.49$ mm. From the filter specifications, the required modal impedances for the two sections of the filter are $Z_{\rm even} = 85.95\ \Omega$ and

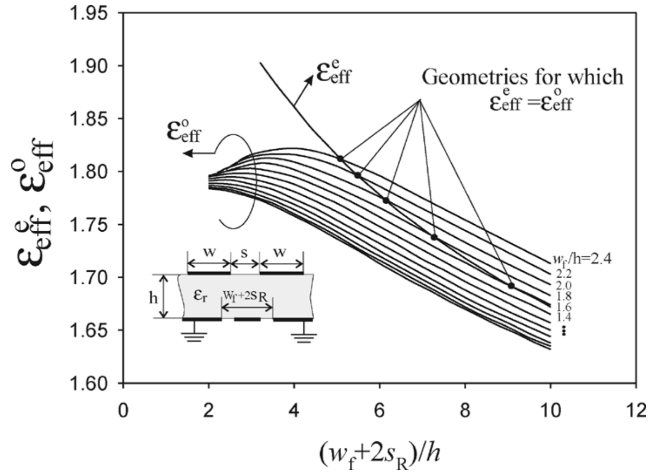


Fig. 8. Modal effective permittivities versus $w_f + 2s_R$ with w_f as parameter for the geometries of filter C satisfying the required modal-impedance values.

TABLE II
DIMENSIONS (IN MILLIMETERS) OF CONVENTIONAL AND MODIFIED COUPLED-LINE BANDPASS FILTER C

Type of design	Sections 1, 2
Conventional	w =0.881, s =0.0369, l =17.73 ϵ_{ef}^{e} = 2.08, ϵ_{ef}^{o}=1.78
Modified	w_f =1, w =1.764, s =0.161, s_R =0.965, l =16.0,l_{ex} =0.50,$\epsilon_{ef}^{e}=\epsilon_{ef}^{o}$=1.78

$Z_{\rm odd} = 37.53\ \Omega$. Looking for the geometries that fulfilled these modal-impedance requirements and simultaneously achieve modal phase-velocity equalization, we have depicted in Fig. 8 the dependence of the modal dielectric permittivities $\epsilon_{\rm eff}^{e}$ and $\epsilon_{\rm eff}^{o}$ with respect to the distance between both sides of the ground plane $w_f + 2s_R$ using w_f as a parameter. Please note that data for the curves in Fig. 8 corresponds to geometries with the specified modal impedances so w and s varies (slightly) accordingly to satisfy this condition. We have plotted a single curve for the even effective dielectric constant because this parameter keeps its value practically unaltered when w_f varies if the distance $w_f + 2s_R$ remains constant. This fact can be easily explained because, in the even mode, when $w_f = 0$ (i.e., in the slotted ground-plane case), the electric-field lines are mainly perpendicular to the slot and, thus, the presence of the floating conductor does not appreciably change the electric-field distribution. However, in the odd mode, the function of the floating conductor is to concentrate the electric field in the substrate region, increasing, in this way, the odd-mode effective permittivity. While in the first two examples involving a substrate of permittivity $\epsilon_r = 10$ we have found for any value of w_f (including $w_f = 0$) a geometry in which the modal phase velocities were equal (thus, the curves for $\epsilon_{\rm eff}^{e}$ and $\epsilon_{\rm eff}^{o}$ intersect for any w_f), in the case of filter C, in which a low-permittivity substrate is used, the width of the floating conductor must be larger than a minimum value (approximately $w_f/h > 1.6$) to achieve the modal phase-velocity equalization. Therefore, the spurious band cannot be eliminated in this filter by using a simple slot in the ground plane. The use of the floating conductor is imperative. The dimensions of the fabricated filter C are shown in Table II. We have arbitrarily chosen $w_f = 1$ mm

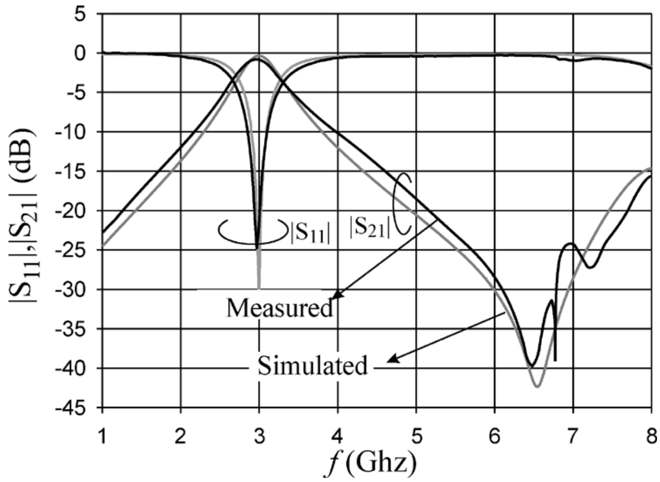


Fig. 9. Simulated and measured results for the final version of filter design C.

(thus $w_f/h = 2.04 > 1.6$). Dimensions of the conventional filter can also be found in this table. For this structure, the strip separation of the modified version of the PCMF is approximately four times that of the conventional filter (which is as small as $s = 37\ \mu$m). Final measured and simulated results show a good agreement, as can be seen from Fig. 9.

III. Conclusions

We have presented a new method to remove the spurious band at twice the central frequency of conventional PCMFs. The method is based on the insertion of floating conductors in the slot practiced in the ground plane along each coupled section. By means of a quasi-TEM approach, a preliminary version of the filter design can be efficiently obtained. The transversal dimensions of each section have been calculated to match the even- and odd-mode phase velocities. In this first step of the design process, no open-end effects are considered. To achieve equal even- and odd-mode electrical lengths, we have used a commercial EM simulator (Ensemble) to find out the optimal extra length that the slot must be lengthened in order to completely suppress the spurious band. The modified version of PCMFs using floating conductor also permits to fabricate tightly coupled sections with relatively large values of separation between strips, thus relaxing requirements on the fabrication technology.

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