New Microstrip "Wiggly-Line" Filters With Spurious Passband Suppression

Txema Lopetegi, *Student Member, IEEE*, Miguel A. G. Laso, *Student Member, IEEE*, Jorge Hernández, Miguel Bacaicoa, David Benito, María J. Garde, Mario Sorolla, *Senior Member, IEEE*, and Marco Guglielmi, *Senior Member, IEEE*

***Abstract*—In this paper, we present a new parallel-coupled-line microstrip bandpass filter with suppressed spurious passband. Using a continuous perturbation of the width of the coupled lines following a sinusoidal law, the wave impedance is modulated so that the harmonic passband of the filter is rejected while the desired passband response is maintained virtually unaltered. This strip-width perturbation does not require the filter parameters to be recalculated and, this way, the classical design methodology for coupled-line microstrip filters can still be used. At the same time, the fabrication of the resulting filter layout does not involve more difficulties than those for typical coupled-line microstrip filters. To test this novel technique, order-3 Butterworth bandpass filters have been designed at 2.5 GHz with a 10% fractional bandwidth and different values of the perturbation amplitude. It is shown that for a 47.5% sinusoidal variation of the nominal strip width, a harmonic rejection of more than 40 dB is achieved in measurement while the passband at 2.5 GHz is almost unaltered.**

***Index Terms*—Bandpass filter, coupled-line filter, harmonic suppression, microstrip.**

I. Introduction

PARALLEL coupled transmission-line filters in microstrip and stripline technologies are very common for implementations of bandpass and bandstop filters with required bandwidths up to a 20% of the central frequency. The design equations for the coupled line parameters (space-gap between lines and line widths and lengths) can be found in classical microwave books. This way, following a well-defined systematic procedure, the required microstrip-filter parameters can be easily derived for both Butterworth and Chebyshev prototypes [1]. Although this type of filter is indeed very popular and simple to implement, it does suffer from a fundamental limitation, namely, the presence of spurious passbands at the harmonics of the design frequency. To reject these harmonics, it is usually necessary to cascade additional filters that can reject the spurious passbands. This solution, however, increases the filter layout area and introduces additional insertion losses.

In this context, more compact microstrip bandpass filters that achieve spurious passband suppression have been proposed by using a uniplanar compact photonic-bandgap (UC-PBG) structure as a machined ground plane in a parallel-coupled-line microstrip filter [2]. The ground-plane structure in [2] introduces a periodic disturbance that rejects the spurious passbands of the microstrip filter and, at the same time, acts as a slow-wave structure that reduces the total physical size of the parallel-coupled-line microstrip filter itself, achieving a 20% shortening in the reported results. However, this slow-wave effect is achieved strongly in the even mode of the coupled lines, but very weakly in the odd mode, substantially increasing the difference between the phase constant of the even and odd modes [3]. As it is well known, the difference between these phase constants produces spurious effects in the frequency response of the bandpass filter [4], giving rise in the discussed case to low attenuation in the rejected band placed immediately to the right-hand side of the desired passband (around 20 dB in the results reported in [3]). Moreover, the design of the parallel coupled-line filter must be completely recalculated involving the need for new graphs to relate the physical and electrical parameters of the coupled lines once the chosen UC-PBG structure is introduced in the ground plane [3].

Unlike these microstrip bandpass filters with a UC-PBG ground plane [2] or the rejected-band photonic bandgap (PBG) microstrip devices with sinusoidal patterns etched in the ground plane proposed in [5] and [6], the standard filters described in the classical literature exhibited a periodic frequency response with harmonic bands. However, one common disadvantage for all these structures having a periodic pattern etched in the ground plane is that the resulting device is not simple to use in practical applications. This is because the whole structure must be suspended far from other ground conductors for the periodic ground plane to be effective.

To overcome this problem, and to broaden at the same time the applicability of the concepts introduced in [5] and [6] on ground-plane sinusoidal etching, in this paper, we propose to introduce a periodic pattern modulating the strip widths of a parallel-coupled-line microstrip bandpass filter while leaving the ground plane unperturbed. It will be shown that employing a continuous modulation pattern that follows a sinusoidal law, the replica of the response at the harmonic of the design frequency can be eliminated, thereby strongly improving the filter perfor-

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T. Lopetegi, M. A. G. Laso, D. Benito, M. J. Garde, and M. Sorolla are with the Electrical and Electronic Engineering Department, Public University of Navarre, E-31006 Pamplona, Spain.

J. Hernández was with the Electrical and Electronic Engineering Department, Public University of Navarre, E-31006 Pamplona, Spain. He is now with Telefonica Moviles, Madrid, Spain 28001.

M. Bacaicoa was with the Electrical and Electronic Engineering Department, Public University of Navarre, E-31006 Pamplona, Spain. He is now with CEIN, Noáim, Navarra, Spain 31110.

M. Guglielmi is with the RF Systems Division, European Space Research and Technology Centre–European Space Agency, 2200 AG Noordwijk, The Netherlands.

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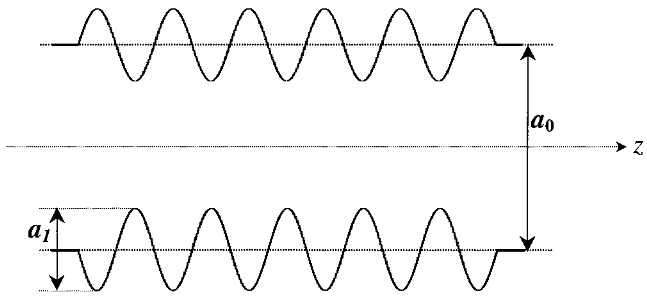


Fig. 1. Generalized waveguide with cross section $a(z)$ perturbed following a sinusoidal law along the propagation axis z.

mance. The design procedure is very simple and does not need a recalculation of the coupled-line dimensions (space between lines, line length, and line mean width) to introduce the wiggle. In fact, the classical design methodology for coupled-line microstrip filters is employed. This way, very effective rejection of the harmonic passband is obtained, while the design passband is almost unaltered.

II. Coupling of Modes and Bragg Effects in Waveguides

A continuous and periodic perturbation of the cross section of a waveguide results in a corresponding modulation of the wave impedance of the electromagnetic waves in the structure and it yields to Bragg reflection in some frequency bands. This phenomenon has been extensively studied in optics where it is well known that, for a weak periodic perturbation of a waveguide, the device frequency response in reflection is given, in a first estimation, by the Fourier transform of the coupling coefficient (closely related to the perturbation geometry) resulting from this perturbation [7]. This fact has been recently demonstrated for sinusoidal patterns etched in the ground plane of microstrip lines with a constant upper conductor strip width [5], [6].

This problem of mode coupling can be generalized for any waveguide mode, including forward- and backward-mode conversions. In a simplified way, it can be formulated for two modes in a generalized waveguide (Fig. 1) having the cross section $a(z)$ perturbed following a sinusoidal law, as in (1)

$$a(z) = a_0 + a_1 \cdot \cos(\Delta\beta \cdot z + \phi) \tag{1}$$

where $a(z)$ is the waveguide cross section perturbed along z, z is the propagation axis, a_0 is the mean value of the cross section, a_1 is the maximum value of the perturbation, and $\Delta\beta$ is the difference between the unperturbed phase constants of the two interacting modes, and it is given here in (2) as follows:

$$\Delta\beta = \beta_1 - \beta_2 = \frac{2\pi}{\lambda_B} \tag{2}$$

where λ_B is the beat wavelength of the two modes; that is the period of the perturbation along the z-axis that makes the two modes satisfy the resonance or coherence relationship. The initial phase ϕ in (1) can be adjusted for convenience.

The consequence of this perturbation is that, for the frequencies at which the coherence relationship is held, a continuous transfer of energy between the two modes is guaranteed and the structure acts as a mode converter [8]. Fourier integral transformation theory yields an additional general conclusion: the minimal length of the periodic structure for mode conversion must be at least of the order of the beat wavelength λ_β of the two modes of interest. The coherence relationship in (2) guarantees that mode conversion to other unwanted modes that could be coupled by the waveguide perturbations suffers from a destructive interference.

In the case of two interacting modes (one forward and one backward), the last one is considered in (2) including a minus sign in its phase constant. This means that, for Bragg reflection in the same, but counter-propagating mode, the phase constant mismatch $\Delta\beta$ must be two times the phase constant of the forward mode, as in (3)

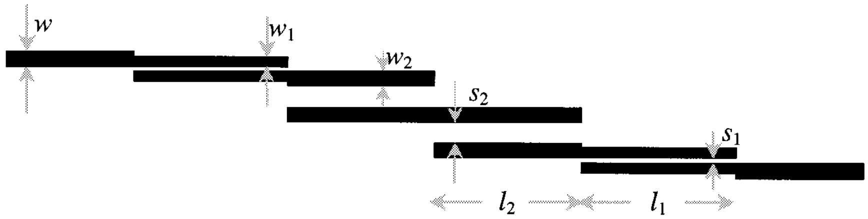
$$\Delta\beta = \beta_1 - (-\beta_1) = 2 \cdot \beta_1. \tag{3}$$

A rigorous description of the previous concepts in the frame of the coupled-wave theory (the cross-sectional method) for general nonuniform waveguides can be found in [8] for the case of bound modes. The generalized case of open waveguides including the continuous spectrum of modes is presented in [9]. A full employment of the coupled-wave theory would allow the designer to choose the periodicity, amplitude, and initial phase of the perturbation necessary to have the desired frequency response.

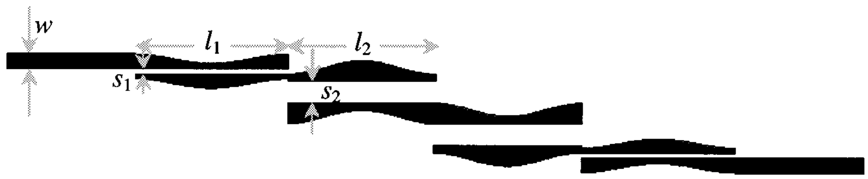
III. "Wiggly-Line" Filter Design

Parallel coupled "wiggly-line" microstrip bandpass filters apply the same concept on which sinusoidal etching of the ground plane [5], [6] for the rejection of frequency bands relies. A pattern formed by one raised sine (or several ones summed up) either implemented as a periodically removed ground plane or as a periodic variation of the upper conductor strip width could yield to improved filters due to the suppression of harmonics. The advantage of a uniform ground plane, in contrast with [2], is that it is not sensible to the interaction with other surrounding grounds.

The coupled-wave theory can be formulated for the case of parallel coupled microstrip lines with the same properties discussed in the previous section [10], [11]. To design the "wiggly-line" filter, we will take advantage of this theory to calculate the period of the perturbation necessary to reject the harmonic passband. Further studies are currently being carried out by the authors to apply the coupled-wave theory to obtain the quantitative frequency response of the structure introduced and, in this way, to characterize the amount of harmonic reduction/main passband distortion, as a function of the amplitude of the perturbation. This would help the designer in deciding the amount of perturbation to use. In Fig. 2, we show a conventional parallel coupled-line filter [see Fig. 2(a)] and a novel parallel coupled "wiggly-line" filter [see Fig. 2(b)] for the same passband requirements. Following the above considerations, the new filter should exhibit (if the perturbation is properly designed) the same passband at the design frequency as the unperturbed one, with an improved out-of-band behavior (harmonic suppressed).



(a)



(b)

Fig. 2. (a) Classical parallel-coupled-line bandpass filter showing the typical design parameters: conductor strip widths w_i, lengths of coupled-line sections l_i, and separation between the coupled lines of the section s_i. (b) "Wiggly-line" filter resulting after applying the strip width perturbation to (a).

TABLE I
PHYSICAL PARAMETERS FOR THE ORDER-3 BUTTERWORTH BANDPASS FILTER CENTERED AT 2.5 GHz WITH A 10% FRACTIONAL BANDWITH i, THE SECTION NUMBER, w_i THE STRIP WIDTH OF THE iTH SECTION, s_i, THE SEPARATION BETWEEN THE COUPLED LINES OF THE iTH SECTION, AND l_i THE LENGTH OF THE iTH SECTION. THE INPUT AND OUTPUT PORT STRIP WIDTHS ARE $w = 1.5$ mm $(Z_0 = 50\ \Omega)$

i	w_i, mm	s_i, mm	l_i, mm	$\varepsilon_{eff,e}$	$\varepsilon_{eff,o}$
1	0.75	0.42	11.98	7.244	5.354
2	1.09	1.70	11.54	7.448	6.086

To test the performance of these novel techniques, we have designed order-3 Butterworth bandpass filters centered at $f_{\rm d} = 2.5$ GHz with 10% fractional bandwidth (i.e., 250 MHz). The substrate employed has relative dielectric constant $\varepsilon_r = 10.2$ and thickness $h = 1.27$ mm. The first stage in the design process of the "wiggly-line" filter is to calculate a conventional parallel coupled microstrip filter to meet the specifications required. The design equations to obtain the layout parameters are very well known and can be found in classical microwave books [1]. The layout dimensions for our case are given in Table I, referring to the variables shown in Fig. 2 and to the effective dielectric constant of the even $(\varepsilon_{\rm eff,e})$ and odd $(\varepsilon_{\rm eff,o})$ modes. Using them, the phase constant is calculated as in (4)

$$\beta_{\rm e,o} = \frac{2 \cdot \pi \cdot f}{c} \cdot \sqrt{\varepsilon_{\rm eff,e,o}} = \frac{2 \cdot \pi}{\lambda_{\rm g,e,o}} \tag{4}$$

where $\lambda_{\rm g,e}$ and $\lambda_{\rm g,o}$ are the guided wavelengths of the even and odd modes and c is the light velocity in vacuum. Due to the nonideality of having different phase constants for the even and odd modes in the microstrip coupled lines, their mean value is taken to calculate the physical length of every section of coupled lines in order to have the prescribed electrical length of 90° $(\lambda_{\rm g}/4)$ at the design frequency.

The second stage in the design process is to calculate the period of the sinusoidal perturbation to adjust it to reject the harmonic passband at $2 \cdot f_{\rm d}$, where $f_{\rm d}$ is the design frequency. This case corresponds to backward coupling of the kind of Bragg reflection in the same, but counter-propagating mode, thus, (3) is applicable and the coherence relationship reduces to (5)

$$\Delta\beta = 2 \cdot \beta_1 = \frac{2\pi}{\lambda_B}. \tag{5}$$

Due to the nonideality of having different phase constants for the even and odd modes, their mean value is employed again to perform the calculations. Taking advantage of the moderate dispersion existent for the fundamental modes in the microstrip lines and neglecting it, the phase constant of the forward mode to be rejected at $2 \cdot f_{\rm d}$ is simply calculated as two times the phase constant at the design frequency $f_{\rm d}$ as follows:

$$\beta_1 = \beta(2 \cdot f_{\rm d}) \cong 2 \cdot \beta(f_{\rm d}) = 2 \cdot \frac{2 \cdot \pi}{\lambda_{\rm gd}} \tag{6}$$

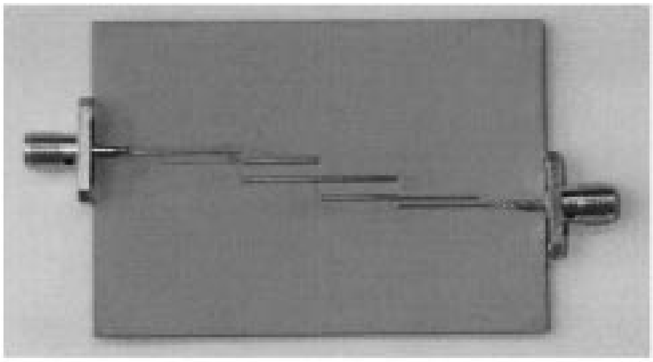
where $\lambda_{\rm gd}$ is the guided wavelength at the design frequency. Introducing this result in the coherence relationship of (5), we obtain

$$\Delta\beta = 2 \cdot \beta_1 = 2 \cdot \frac{4 \cdot \pi}{\lambda_{\rm gd}} = \frac{2 \cdot \pi}{\lambda_B}. \tag{7}$$

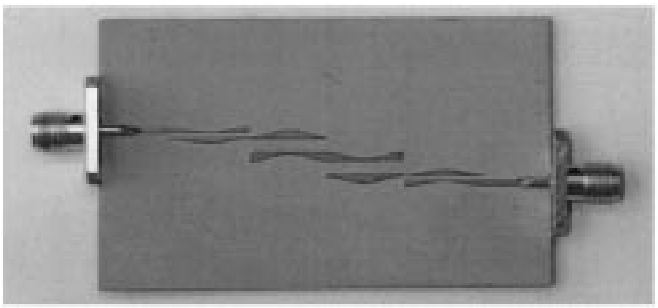
This means that the beat wavelength λ_B (the period of the perturbation necessary to satisfy the coherence relationship and this way to reject the mode) is equal to one-quarter of the guided wavelength at the design frequency of the microstrip filter, as indicated in (8)

$$\lambda_B = \frac{\lambda_{\rm gd}}{4}. \tag{8}$$

It must be noted that every coupled-line section of the filter will have its own mean phase constant value and, therefore, its own guided wavelength at the design frequency $\lambda_{\rm gd}$, but all of them will have the same electrical length of 90° at the design frequency, corresponding to $(\lambda_{\rm gd}/4)$. This means that, in every coupled line section, exactly a complete period of perturbation λ_B can be accommodated.

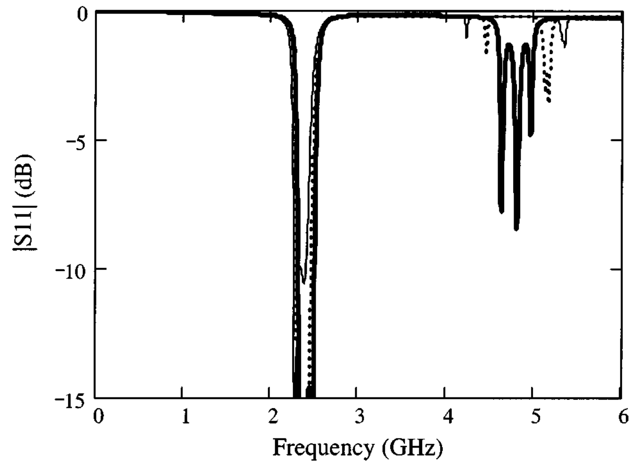


(a)



(b)

Fig. 3. (a) Classical coupled-line order-3 Butterworth bandpass microstrip filter centered at 2.5 GHz with a 10% fractional bandwidth. (b) "Wiggly-line" filter resulting after applying the strip width perturbation with $M = 47.5\%$ to (a).



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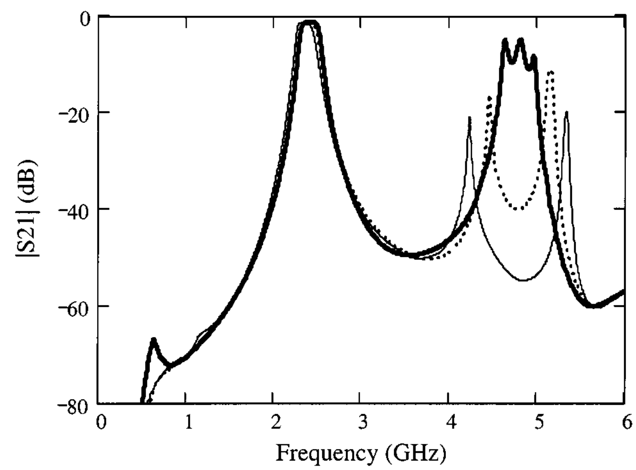


(b)

Fig. 4. (a) Simulated $|S_{11}|$- and (b) $|S_{21}|$-parameters for a classical coupled-line order-3 Butterworth bandpass microstrip filter centered at 2.5 GHz with a 10% fractional bandwidth (thick solid line), for a "wiggly-line" filter with $M = 37.5\%$ (dashed line), and for a "wiggly-line" filter with $M = 47.5\%$ (thin solid line).

The third and last stage in the design process consists of the introduction of the perturbation in the conventional filter previously designed. This perturbation will be introduced in an asymmetrical way, modulating the outer edge of the coupled lines, but keeping the inner edge (coupling region) unaltered (constant distance between the coupled lines and straight shape), as can be seen in Fig. 2. This way, the conductor strip-width variation $w_i(z)$ in the ith coupled line section can be expressed as in (9), where z varies from zero to the beat wavelength at this coupled-line section $\lambda_{B,i}$ (equal to the physical length of the section), and the initial phase is fixed alternatively to 0° and 180° as follows:

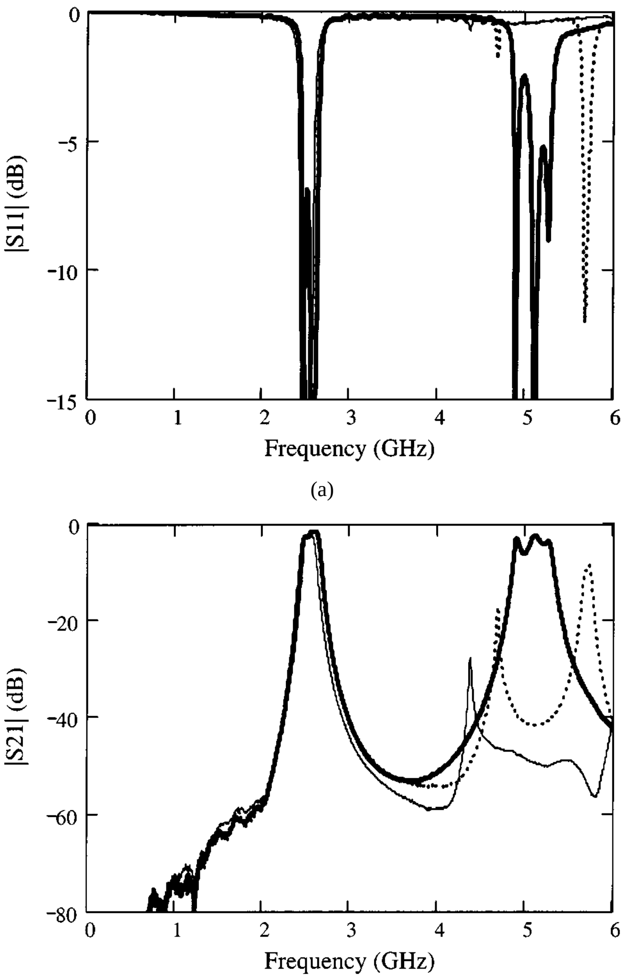
$$w_i(z) = w_i \cdot \left(1 + \frac{1}{2} \cdot \frac{M\,(\%)}{100} \cdot \cos\left(\frac{2 \cdot \pi \cdot z}{\lambda_{B,i}} + \phi\right)\right) \quad (9)$$

where $w_i(z)$ is the variable width of the conductor strips of the ith coupled line section, ϕ is their initial phases (0° and 180°), w_i is the constant widths calculated for the conventional filter, $\lambda_{B,i}$ is the beat wavelength (the period of the sinusoidal perturbation for the ith coupled-line section), and M is the strip-width modulation parameter expressed in percentage (see Fig. 2). The optimal initial phase ϕ of the perturbation in every coupled-line section has been chosen to maximize the rejection level achieved in the undesired passbands.

IV. Simulations and Measurements

A systematic simulation of multiple prototypes, designed as stated in the previous section, with different perturbation amplitudes (different M) has been performed to test the tradeoff between the spurious passband rejection level and the preservation of the shape and matching of the main passband. All the simulations have been done using an Agilent Momentum planar circuit simulator, considering a 1.27-mm-thick substrate with relative dielectric constant $\varepsilon_r = 10.2$. Several prototypes have been fabricated on a Rogers RO3010 substrate, using a numerical milling machine (Fig. 3). The measurements have been realized with a HP 8753-D vector network analyzer (up to 6 GHz).

In Figs. 4 and 5, the simulated and experimental results are presented for the conventional parallel coupled-line microstrip filter and two parallel coupled-"wiggly-line" prototypes with perturbations $M = 37.5\%$ and 47.5%. A very good agreement between simulations and experimental data has been obtained. The little differences found can be due to the tolerances in the fabrication process and to the presence of the connector junctions in the constructed prototypes. It can be observed that, the bigger the amplitude of the perturbation in the coupled lines, the higher the rejection level of the spurious passband. At the same time, the return losses and shape of the main passband are



(a)

(b)

Fig. 5. (a) Measured $|S_{11}|$- and (b) $|S_{21}|$-parameters for a classical coupled-line order-3 Butterworth bandpass microstrip filter centered at 2.5 GHz with a 10% fractional bandwidth (thick solid line), for a "wiggly-line" filter with $M = 37.5\%$ (dashed line), and for a "wiggly-line" filter with $M = 47.5\%$ (thin solid line).

maintained very similar in both the unperturbed and perturbed cases, in this way offering a very satisfactory performance. As an example, we can see that with a perturbation $M = 47.5\%$, a harmonic rejection of over 40 dB is achieved in measurement, while the main passband is kept almost unaltered.

V. Conclusions

In this paper, we have proposed the strip-width modulation of otherwise traditionally designed parallel coupled-line microstrip filters in order to suppress their spurious passband. The strip-width modulation for each coupled-line section follows a sinusoidal law whose spatial periodicity must be matched to the wavelength of the harmonic of the design frequency to be rejected. This way, enhanced out-of-band behavior can be obtained while maintaining the main passband virtually unaltered. This technique has important advantages since the wiggle of the lines does not force the prototype parameters to be recalculated. Moreover, the novel prototypes do not demand special fabrication processes or special installation requirements (e.g., suspended structures) different from those needed for classical coupled-line microstrip filters.

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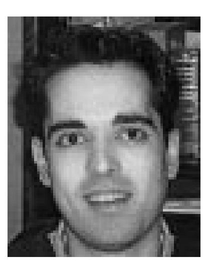
Txema Lopetegi (S'99) was born in Pamplona, Spain, in 1973. He received the Telecommunication Engineering degree from the Public University of Navarre, Navarre, Pamplona, Spain, in 1997, and is currently working toward the Ph.D. degree at the Public University of Navarre.

Since 1997, he has been with the Electrical and Electronic Engineering Department, Public University of Navarre, where he was an Assistant Professor from 1997 to 1999, Associate Researcher from 1999 to 2000, and again as an Assistant Professor from 2000 to the present. His research interests include passive devices and electromagnetic crystals in planar microwave and millimeter-wave technologies with a special concern for microstrip technology.

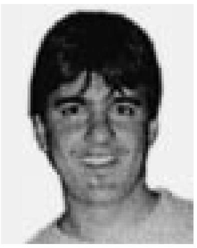


Miguel A. G. Laso (S'99) was born in Pamplona, Navarre, Spain, in 1973. He received the M.Sc. degree from the Public University of Navarre, Pamplona, Spain, in 1997, and is currently working toward the Ph.D. degree at the Public University of Navarre.

Since 1997, he has been an Associate Researcher in the Electrical and Electronic Engineering Department, Public University of Navarre. His main research interests include fiber Bragg gratings and fiber Bragg grating-based devices, as well as electromagnetic crystals in planar microwave technologies.

Jorge Hernández was born in Madrid, Spain, in 1976. He received the M.Sc. degree from the Public University of Navarre, Pamplona, Spain, in 2000.

He is currently with Telefonica Moviles, Madrid, Spain. His main research interests include passive microwave circuits and electromagnetic crystals.

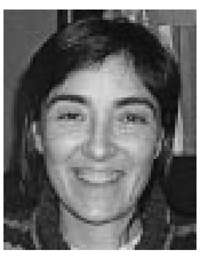


Miguel Bacaicoa was born in Pamplona, Spain, in 1976. He received the M.Sc. degree from the Public University of Navarre, Pamplona, Spain, in 2000.

He is currently with CEIN, Noáim, Spain. His main research interests include passive microwave circuits and electromagnetic crystals.

David Benito was born in Huesca, Spain, in 1965. He received the B.Sc., M.Sc., and Ph.D. degrees from the Polytechnic University, Madrid, Spain, in 1987, 1992, and 1999 respectively.

He is currently an Associate Professor in the Electrical and Electronic Engineering Department, Public University of Navarre, Pamplona, Spain. His main research interests include electrooptic modulators, fiber Bragg gratings and their applications to optical communications, and electromagnetic crystals.



María J. Garde was born in Pamplona, Spain, in 1964. She received the M.Sc. and Ph.D. degrees from the Complutense University of Madrid, Madrid, Spain, in 1987 and 1992, respectively.

She is currently a Professor in the Electrical and Electronic Engineering Department, Public University of Navarre, Pamplona, Spain. Her main research interests include acoustooptical modulators, fiber Bragg gratings, and electromagnetic crystals.

Mario Sorolla (S'82–M'83–SM'01) was born in Vinaròs, Spain, in 1958. He received the Telecommunication Engineer degree from the Politechnical University of Catalonia, Catalonia, Spain, in 1984, and the Ph.D. degree from the Politechnical University of Madrid, Madrid, Spain, in 1991.

From 1986 to 1990, he designed very high power millimeter waveguides for plasma heating in the Euratom-Ciemat Spanish Nuclear Fusion Experiment. From 1987 to 1988, he was an Invited Scientist at the Institute of Plasma Research, Stuttgart University, Stuttgart, Germany. He was involved with microwave integrated circuits and monolithic-microwave integrated circuits for satellite communications for Tagra and Mier Comunicaciones. From 1984 to 1986, he was a Professor at the Politechnical University of Catalonia, Vilanova i la Geltrú, Spain. From 1991 to 1993, he was a Professor at the Ramon Llull University, Barcelona, Spain. Since 1993 , he has been a Professor at the Public University of Navarre, Pamplona, Spain. His research interests include high-power millimeter waveguide components and antennas, coupled-wave theory, and applications of electromagnetic crystals to microwave circuits and antennas.

Marco Guglielmi (M'78–SM'97) was born in Rome, Italy, on December 17, 1954. He received the Ingegneria Elettronica Laurea degree from the University of Rome "La Sapienza," Rome, Italy, in 1979, attended the Scuola di Specializzazione in Elettromagnetismo Applicato, Rome, Italy, in 1980, received the M.S. degree in electrical engineering from the University of Bridgeport, Bridgeport, CT, in 1982, and the Ph.D. degree in electrophysics from the Polytechnic University, Brooklyn, NY, in 1986.

From 1984 to 1986, he was an Academic Associate at the Polytechnic University, and from 1986 to 1988, he was an Assistant Professor. From 1988 to 1989, he was an Assistant Professor at the New Jersey Institute of Technology, Newark, New Jersey. In 1989, he joined the RF System Division, European Space Research and Technology Centre, Noordwijk, The Netherlands, where he is currently involved in the development of passive microwave components for space applications. His professional interests include solid-state devices and circuits, periodic structures, phased arrays and millimeter-wave leaky-wave antennas, network representations of waveguide discontinuities, and microwave filtering structures.

Dr. Guglielmi was the recipient of a 1981 Fulbright Scholarship and a Halsey International Scholarship Program (HISP) Scholarship.