A Simple Method to Design Wideband Electronically Tunable Combline Filters

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Abstract— **A new systematic approach for designing wideband tunable combline filters is presented. New results on tunable combline filter theory are proposed and explicit design formulas, to obtain the filter design parameters from specifications, are included. These design parameters are: center frequency, resonator electrical length,** *instantaneous* **bandwidth and tuning capacitance. The proposed design technique is used to construct an X-band wideband microstrip tunable filter from** 8.0 GHz **to** 12.0 GHz **with commercial GaAs FETs as tuning elements. Parasitic effects and simulation problems are also discussed.**

Keywords— **electronically tunable filters, combline filters, frequency control.**

I. Introduction

Combline filters are broadly used as bandpass filters in modern microwave and millimeter-wave subsystems due to their compactness, excellent stopband and selectivity performance, and ease of integration [1], [2]. If they are built on microstrip substrate, very low cost and small size can be achieved making them very attractive for mobile communication [3] and wideband radar systems [4]. On the other hand, the need of flexibility in commercial and military radiofrequency applications demands the use of high-performance electronically tunable filters with high tuning speed, high Q-factor and broad tuning range [5].

Microstrip combline and interdigital tunable filters have been described by many authors. Hunter and Rhodes [6] describe a method to design combline filters. Although they provide basic expressions and useful results, their approach is neither suitable for microstrip technology nor gives the criteria to choose the filter design parameters. In [7] and [8] a varactor tunable, high-Q microwave filter is presented where the tuning element losses are compensated by the use of a FET circuit. However, this approach is not feasible for wide tuning ranges. The advantages and drawbacks of using either varactor diodes or MESFET varactors to tune the pass band center frequency are discussed in [9] along with a review of end-coupled microstrip-line bandpass filters. More recently in [10], a varactor electronically tunable interdigital filter is shown. Parasitic effects of the varactor series resistance and the resonator electrical length on the overall resonator quality factor and filter insertion losses are also discussed in detail.

However, these papers do not provide systematic and consistent criteria to establish the filter design parameters straightforward from the initial specifications. We present a simple method to design tunable combline filters systematically from the specifications, so that broad tuning ranges can be achieved. The method is applied to microstrip combline filters with tapped-line inputs. Its fundamentals are described in section II and an illustrative design example is shown in section III. Section IV includes simulations, measurements and comments on the most important non-ideal effects encountered in the practical implementation of the filter. This is a five pole filter with a tuning range from 8.0 GHz to 12.0 GHz. The measured *instantaneous* bandwidth range varies from 630 MHz to 1350 MHz. Commercial GaAs Fets in *cold* configuration are used as tuning devices.

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II. Tunable combline filters design procedure

The design method of tunable filters suggested in this paper can be split into two steps. The first step consists of the suitable selection of the design parameters required to undertake the second step. These are: design center frequency (f_0), resonator electrical length (θ_0) at f_0, *instantaneous* bandwidth (ΔB_{L0}) and tuning device capacitance (C_{s0}) at f_0. The second step comprises the design of a microstrip tapped-line combline filter with a fixed center frequency making use of the method in [11]. This method assumes a cuasi-TEM approximation and negligible nonadjacent-line coupling, and uses the equivalent circuit by Cristal [12]. Although its inaccuracy has been proved [13], it provides good agreement in and near the filter passband and can be used for design purposes. The characteristic admittances of the coupled-line structure can then be obtained using the expressions proposed in [14]. The filter physical dimensions can be calculated using synthesis formulas.

The necessary criteria for determining the above mentioned parameters (f_0, θ_0, ΔB_{L0} and C_{S0}) from the specifications constitute the main contribution of this work and are explained in next subsections. The most common specifications that a tunable filter must meet are next listed:

- Tuning range or tuning bandwidth ($[f_1, f_2]$): the range of frequencies in which the filter should be tuned.
- *Instantaneous* bandwidth[1] range ($[\Delta B_{Lmin}, \Delta B_{Lmax}]$) the filter has within the tuning range. This is given for a particular attenuation level ($L\ dB$).
- Maximum reflection coefficient (L_r) in the filter passband.
- Filter selectivity requirements; specified for example at a percentage of the filter passband edge frequencies.

[1] *Instantaneous* means in a specific tuning frequency within the range of tunable frequencies.

- Capacitance range ($[C_{smin}, C_{smax}]$) of the tunable elements used.

A. Design center frequency f_0

Due to the fact that rejection requirements are harder to attain at lower tuning frequencies [2], the frequency f_0 is selected to be

$$f_0 = \sqrt{f_1 \cdot f_2}\ , \tag{1}$$

where f_1 and f_2 are the tuning bandwidth lower and upper limits. A smaller value of f_0 may be chosen on the assumption that requirements at the highest part of the tuning range will be met without problems. In any case, equation (1) will be used in the design procedure.

B. Electrical length θ_0

Hunter and Rhodes [6] propose a three step transformation of the filter equivalent circuit consisting on 1) the admittance inverter equivalent circuit of the filter (see Fig. 4 of [6]), 2) scaling the network admittances by $\frac{\tan(\theta)}{\tan(\theta_0)}$, and 3) transforming the previous network to the lowpass prototype with admittance inverters by the inverse of a bandpass transformation. Assuming a narrow *instantaneous* bandwidth, it follows that

$$\Delta B_L = k\ \frac{\theta \cdot \tan(\theta)}{\tan(\theta) + \theta \cdot \left(1 + \tan^2(\theta)\right)}\ . \tag{2}$$

Equation 2 shows the filter *instantaneous* bandwidth dependence on the resonator electrical length, θ, at the tuning frequency. k is a constant that depends on the resonator geometry. If k is removed, the logarithm of the normalized *instantaneous* bandwidth, ΔB_{Ln}, can be defined as follows:

$$\begin{aligned}\Gamma &= 10\log_{10}(\Delta B_{Ln}) = \\ &= 10\log_{10}\left(\frac{\theta \cdot \tan(\theta)}{\tan(\theta) + \theta \cdot \left(1 + \tan^2(\theta)\right)}\right)\ .\end{aligned} \tag{3}$$

Equation (3) is depicted in Fig. 1. Due to the linear relationship between electrical length and frequency, this equation also shows the *instantaneous* bandwidth dependence on the tuning frequency. Since a typical specification consists of maintaining the *instantaneous* bandwidth within a range, $[\Delta B_{L\,\mathrm{min}}, \Delta B_{L\,\mathrm{max}}]$, equation (3) is very useful to select θ_0. As the maximum of this equation is reached when $\theta_{\mathrm{max}} \approx 53°$, it can be assigned θ_{max} to $\Delta B_{L\,\mathrm{max}}$ so that the maximum tuning range is attained fulfilling the bandwidth requirement as illustrated in Fig. 1.

If

$$\Delta\Gamma = 10\log_{10}\frac{\Delta B_{L\,\mathrm{max}}}{\Delta B_{L\,\mathrm{min}}}\ , \text{and} \tag{4}$$

$$\theta_{\mathrm{min}_1}, \theta_{\mathrm{min}_2} = \Gamma^{-1}\left(\Gamma(\theta_{\mathrm{max}}) - \Delta\Gamma\right) \tag{5}$$

then the filter must satisfy

[2]Due to the tuning element low Q factor at those low tuning frequencies [10].

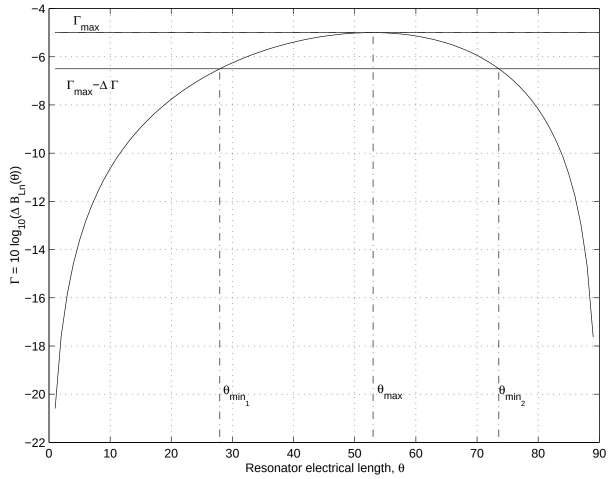

Fig. 1. Normalized instantaneous bandwidth $\Gamma = 10\log_{10}(\Delta B_{Ln})$. θ_{max} is the electrical length where Γ reaches its maximum value. Once $\Delta\Gamma$ is known, θ_{min_1} and θ_{min_2} can be obtained.

$$\theta_{\mathrm{min}_1} \le \theta_1 \le \theta_0 \le \theta_2 \le \theta_{\mathrm{min}_2}\ , \tag{6}$$

where θ_1 and θ_2 are, respectively, the resonator electrical length at f_1 and f_2. Constraint (6) can be written as follows

$$\theta_{\mathrm{min}_1}\frac{f_0}{f_1} \le \theta_0 \le \theta_{\mathrm{min}_2}\frac{f_0}{f_2}\ . \tag{7}$$

From (7) a range of possible θ_0 values can be obtained to satisfy the specified tuning range.

C. Instantaneous bandwidth ΔB_{L0}

The *instantaneous* bandwidth, ΔB_{L0}, (for $L\ dB$ attenuation) at f_0 is computed as it is done with θ_0. Using the relationship

$$\begin{aligned}k &= \frac{\min\limits_{\theta\in\{\theta_1,\theta_2\}}(\Delta B_L(\theta))}{\min\limits_{\theta\in\{\theta_1,\theta_2\}}(\Delta B_{Ln}(\theta))} = \\ &= \frac{\Delta B_L(\theta_0)}{\Delta B_{Ln}(\theta_0)} = \frac{\Delta B_L(\theta_{\mathrm{max}})}{\Delta B_{Ln}(\theta_{\mathrm{max}})}\ ,\end{aligned} \tag{8}$$

and the *instantaneous* bandwidth specification, an expression similar to the θ_0 constraint is obtained for ΔB_{L0}:

$$\begin{aligned}&\frac{\Delta B_{L\,\mathrm{min}}}{\min\limits_{\theta\in\{\theta_1,\theta_2\}}(\Delta B_{Ln}(\theta))}\ \Delta B_{Ln}(\theta_0) \le \Delta B_{L0} \le \\ &\le \frac{\Delta B_{L\,\mathrm{max}}}{\Delta B_{Ln}(\theta_{\mathrm{max}})}\ \Delta B_{Ln}(\theta_0)\ .\end{aligned} \tag{9}$$

Once ΔB_{L0} is known and the filter order is selected to meet the selectivity requirements, the Chebyshev approximation parameters can be computed through the conventional lowpass to bandpass transformation and the Chebyshev approximation [15].

D. Capacitance C_{s0}

The equation that relates a tuning frequency and its corresponding tuning capacitance is [6]

$$\frac{1}{Y_a} \cdot 2\pi f \cdot \tan(\theta) \cdot C_s(f) = 1 \ , \tag{10}$$

where θ is the resonator electrical length at the tuning frequency f, Y_a is the resonator characteristic admittance and $C_s(f)$ is the tuning capacitance that tunes the filter at f. Varactors and FETs in *cold* configuration are usually used as tuning elements.

The tuning device provides a variable capacitance within a range $[C_{s\,\min}, C_{s\,\max}]$. According to equation (10), the smallest capacitance tunes the highest tunable frequency, f_{Mtun}, and the largest capacitance the lowest one, f_{mtun}. Additionally, the maximum tunable frequency can be expressed as an implicit function of C_{s0} by means of

$$f_{Mtun}\ C_{s\,\min} \tan\left(\frac{\theta_0}{f_0} f_{Mtun}\right) = f_0\ C_{s0} \tan(\theta_0) \tag{11}$$

and the minimum tunable frequency by

$$f_{mtun}\ C_{s\,\max} \tan\left(\frac{\theta_0}{f_0} f_{mtun}\right) = f_0\ C_{s0} \tan(\theta_0) \ . \tag{12}$$

As can be noticed, the lower C_{s0} is, the lower f_{mtun} and f_{Mtun} are. Defining C_{s0_1} and C_{s0_2} as

$$f_{mtun}(C_{s0_1}) = f_1 \tag{13}$$

$$f_{Mtun}(C_{s0_2}) = f_2 \ , \tag{14}$$

the next constraint is obtained

$$C_{s0_2} \leq C_{s0} \leq C_{s0_1} \ . \tag{15}$$

Therefore the range of possible C_{s0} values is given by

$$\frac{f_2 \tan(\theta_0 \cdot f_2/f_0)}{f_0 \tan(\theta_0)}\ C_{s\,\min} \leq C_{s0} \leq \leq \frac{f_1 \tan(\theta_0 \cdot f_1/f_0)}{f_0 \tan(\theta_0)}\ C_{s\,\max} \ . \tag{16}$$

To illustrate the previous discussion, the reader may refer to Fig. 2, where the following parameters have been selected: tuning bandwidth $= [8.0, 12.0]\ GHz$, capacitance range $= [0.18, 0.6]\ pF$, $f_0 = 9.8\ GHz$ and $\theta_0 = 50°$. In Fig. 2 the variation with the capacitance C_{s0} of the lowest and highest tuning frequencies, f_{mtun} and f_{Mtun} respectively, is shown. The capacitances C_{s1} and C_{s2} are also depicted. In this case, the range of C_{s0} possible values results in $[0.337, 0.357]\ pF$.

III. Filter design example

In this section a specific design of an electronically tunable combline filter will be performed following the previously described technique. The proposed filter schematic is shown in figure 3.

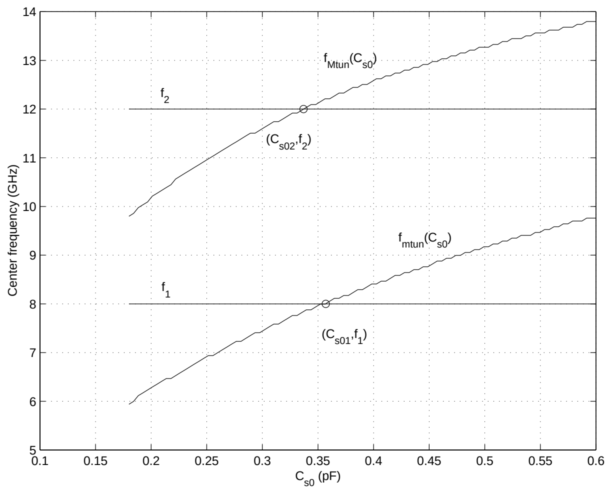

Fig. 2. Illustrative example of the C_{s0} selection. The dependence of f_{Mtun} and f_{mtun} on C_{s0} is shown. C_{s0} must be selected so that $f_{Mtun} \geq f_2$ and $f_{mtun} \leq f_1$ in order to achieve the desired tuning range.

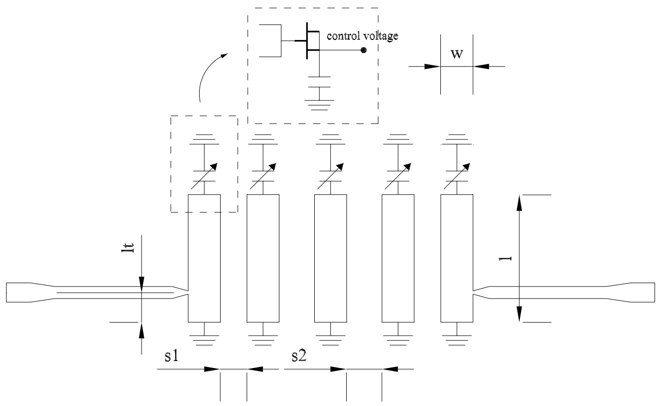

Fig. 3. Designed filter schematic.

A. Filter requirements and specifications

The initial design requirements and specifications are given in Table I, where f_{c1} and f_{c2} are the cut-off frequencies at $L = 1.5\ dB$. As tuning device the *NE71000* MESFET by *NEC* in *cold* configuration is employed. Its characterization is described in Appendix A.

B. Design parameters

The systematic design approach provided in Section II is next applied. First, from equation (1) the design center frequency f_0 is determined. Next, making use of (7) the resonators electrical length θ_0 is obtained. Once the previous parameters are determined the *instantaneous* bandwidth ΔB_{L0} is calculated using (9) and finally the parameter C_{s0} is obtained using equation (16). The design parameter values and their variation range obtained through the design technique are shown in Table II along with the Chebyshev approximation parameters.

TABLE I
DESIGN SPECIFICATIONS

f_1	8.0 GHz
f_2	12.0 GHz
ΔB_{Lmin}	600 MHz ($L = 1.5$ dB)
ΔB_{Lmax}	1000 MHz ($L = 1.5$ dB)
Rejection at $f_{c1} - 0.1f_{c1}$	15.0 dB
Rejection at $f_{c2} + 0.1f_{c2}$	15.0 dB
$[C_{smin}, C_{smax}]$	$[0.18, 0.6]$ pF
L_r	-10.0 dB

TABLE II
FILTER DESIGN PARAMETERS

f_0	9.79 GHz, $8.0 \le f_0 \le 12.0$ GHz
θ_0	$50°$, $28.29° \le \theta_0 \le 62.59°$
ΔB_{L0}	775 MHz, $645 \le \Delta B_{L0} \le 995$ MHz
C_{s0}	0.35 pF, $0.337 \le C_{s0} \le 0.356$
N	5
L_{ar}	0.1 dB
w	0.072

C. Physical Dimensions

The combline filter physical dimensions are calculated by means of [11], [14] and synthesis formulas. In our case, Linecalc^{TM} has been used to synthesize the physical dimensions (see Fig. 3) shown in Table III. The filter has been fabricated on a 254 μm alumina substrate with a dielectric constant of 9.9 and metal thickness of 10 μm.

TABLE III
FILTER PHYSICAL DIMENSIONS

widths (w)	0.400 mm
length (l)	1.590 mm
spacings (s1,s2)	$(0.331, 0.439)$ mm
length from ground to input line (lt)	0.370 mm

IV. SIMULATIONS AND MEASUREMENTS

First of all the filter response is simulated making use of the model given in [12] with ideal transmission lines. Capacitors with no parasitic effects are used as tuning elements. The shortcircuits at the end of the resonators are also assumed to be ideal. This response is shown in Fig. 4. As can be seen it meets the initial requirements given in Table I except the return losses specification at the extreme tuning frequencies, due to the narrow-band matching behaviour of the tapped-line inputs.

Secondly parasitic effects must be considered. These effects are: 1) the effects introduced by the low Q FETs used as tuning elements, 2) the resonators' shortcircuits inductive and resistive behaviour, and 3) the dispersive characteristics of microstrip transmission lines. The simulated response of the filter is obtained making use of Libra^{TM}.

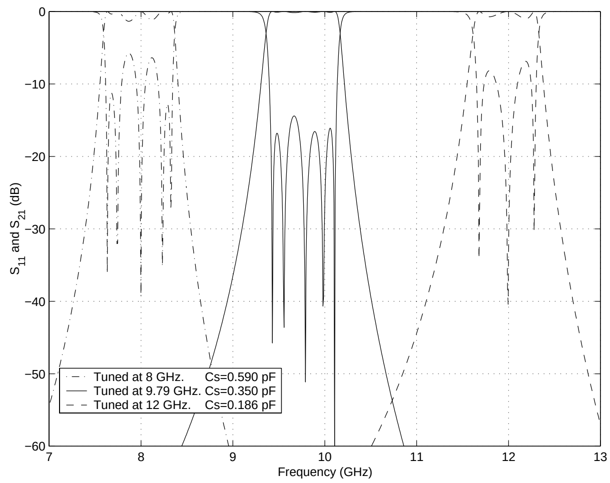

Fig. 4. Ideal filter simulated response.

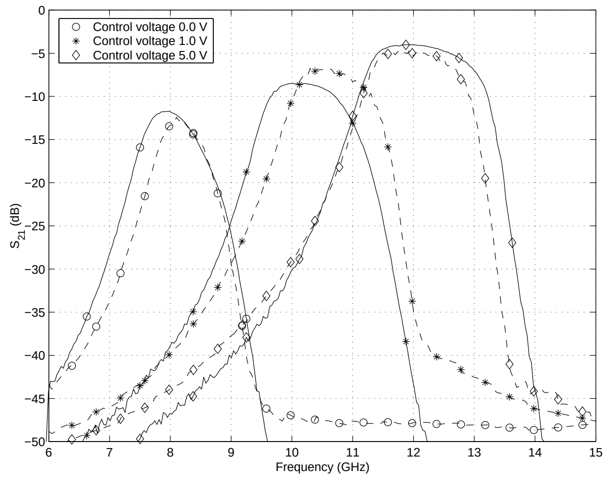

Fig. 5. Insertion losses for various control voltages: simulations (dashed line) and measurements.

This simulator does not support a coupling multiple resonator structure, so in order to perform the simulations it is necessary to use the technique described in [16]. Simulations after optimizing with Libra^{TM} and measurements of the final filter prototype are given in Figs. 5 and 6 for different control voltages . These simulations are performed using 1) real measurements of the FET (see Appendix A) and 2) an equivalent circuit for the non-ideal behaviour of the shortcircuits consisting on an RL series circuit.

Taking into account the high design center frequency (9.79 GHz), the good agreement between simulations and measurements should be remarked. The tuning frequency for a given control voltage is predicted with an error smaller than 5%. The discrepancy between the insertion losses measurements and simulations is always smaller than 2.5 dB. The typical behaviour of insertion losses reduction with tuning frequency [10] can also be noticed. The discrepancies between measured and simulated *instantaneous* bandwidth are due to the differences between the FETs' characteristics -they come from different batches- and to

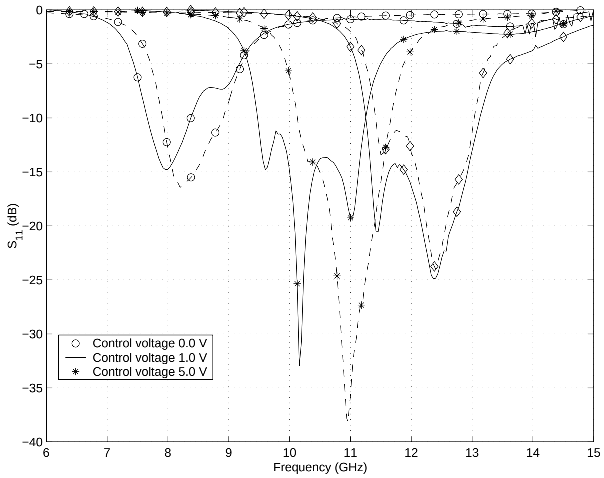

Fig. 6. Return losses for various control voltages: simulations (dashed line) and measurements.

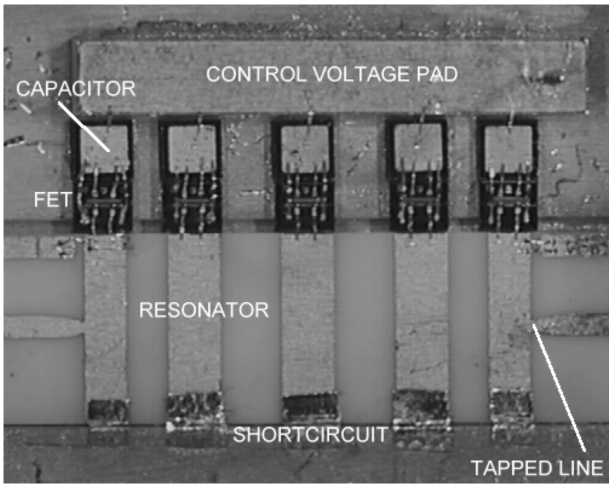

Fig. 7. Filter protoype.

the simulator models (see [16] for a comparison of different CAD tools). Table IV resumes the performance of the filter response shown in Figs. 5 and 6.

TABLE IV
Filter Measured Performance

f_1	8.0 GHz
f_2	12.0 GHz
ΔB_{Lmin}	610 MHz ($L = 1.5$ dB)
ΔB_{Lmax}	1380 MHz ($L = 1.5$ dB)
Rejection at $f_{c1} - 0.1f_{c1}$	> 15.0 dB
Rejection at $f_{c2} + 0.1f_{c2}$	> 15.0 dB
$[C_{smin}, C_{smax}]$	$[0.18, 0.6]$ pF
L_r	< -10.0 dB

Figure 7 shows a photograph of the filter prototype.

In order to fully characterize the designed filter other important parameters are also measured. The tuning time is measured to be smaller than 25.0 *ns* at 25° *C*. The measured noise figure, 1 *dB* compression point (P1dB) and third-order intermodulation intercept point (IP3) are given in Table V for different control voltages.

TABLE V
Noise Figure, Input $P1dB$ and Input $IP3$

Control Voltage (V)	NF (dB)	Input P1dB (dBm)	Input IP3 (dBm)
0.5	13.0	8.0	18.5
5	4.5	21.0	29.0

The IP3 is measured making use of two tones 200 MHz apart. The third-order mixing products are within the filter 1.5 *dB instantaneous* bandwidth. The small values of measured P1dB and IP3 at low control voltages are due to the fact that the FETs gate diode starts conducting with smaller signal levels.

V. Conclusions

A novel and easy to apply systematic procedure to design tunable combline filters has been presented. The proposed method sets the range of values of the key parameters necessary to design this kind of tunable filters (f_0, θ_0, ΔB_{L0}, C_{s0}) and allows the achievement of broad tuning ranges. The method has proved to be extremely useful in the design of an X-band FET electronically tunable microstrip combline filter. Additionally a discussion on the most relevant non-ideal effects has been presented.

Appendix

I. Tuning elements. Characterization and measurements

One of the key elements in designing a wideband tunable combline filter is the selection of the tuning element. In the case of this work a GaAs FET in *cold* configuration (see Fig 3) is chosen (*NE71000* by *NEC*) although other elements were also proved to be suitable for this kind of designs. The characterization of these tuning devices is essential if the range of available capacitance values needs to be accurately determined. From the S_{11} parameter measurement of the *cold*-FET it is noticed that a simple RLC series circuit is accurate enough to model its behaviour. Table VI gives the RLC series equivalent circuit values for the FET used as a tuning element for different control voltages, where R_{eq}, C_{eq} and L_{eq} come basically from the gate resistance, the gate-drain and gate-source parallel capacitance and the bonding wire inductance.

TABLE VI
FET Equivalent Circuit

Control Voltage (V)	R_{eq} (Ω)	C_{eq} (pF)	L_{eq} (nH)
0	2.8	0.40	0.23
1	2.16	0.21	0.23
5	1.78	0.15	0.23

In order to account for the bonding wire inductance effect on the reactance that loads the resonators, it is necessary to define an effective capacitance that is used to determine $[C_{smin}, C_{smax}]$ as given in equations 17 and 18.

$$\left.\frac{1}{2\pi f_1 C_{eq}} - 2\pi f_1 L_{eq}\right|_{V_{control}=0V} = \frac{1}{2\pi f_1 C_{smax}} \quad (17)$$

$$\left.\frac{1}{2\pi f_2 C_{eq}} - 2\pi f_2 L_{eq}\right|_{V_{control}=5V} = \frac{1}{2\pi f_2 C_{smin}} \quad (18)$$

Acknowledgments

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