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Hybrid tunable microwave devices based on Schottky-diode varactors

Emmanuel Pistono², Anne-Laure Perrier², Raffi Bourtoutian¹, Darine Kaddour¹, Akil Jrad³, Jean-Marc Duchamp², Lionel Duvillaret², Didier Vincent⁴, Anne Vilcot¹, Philippe Ferrari¹

Abstract - In this paper, tunable RF/microwave devices are demonstrated. First, a phase shifter is carried out, based on a modified classical topology. This modified topology is smaller than already published ones for a 20 dB return loss and exhibits lower insertion losses (2.4 dB for a 360° tunable phase shift). Second, two original topologies of tunable impedance transformers are carried out. 16.5 Ω to 280 Ω loads can be matched to 50 Ω with very compact transformers (30° electrical length) and small insertion losses (less than 1 dB). Then, tunable filters based on periodic structures showing interesting performances, especially in terms of compactness, selectivity and tuning, are demonstrated. A lowpass and a bandpass filter have been realized and measured. For the lowpass filter, a 16 dB return loss is obtained over a ± 17 % tunable bandwidth for a 130° filter electrical length. The attenuation slope is more than 300 dB/decade and the attenuation band is nine times the mean frequency with more than 25 dB of attenuation. For the bandpass filter, the -3 dB bandwidth can be tuned between ~50 MHz and ~78 MHz for a ± 18%-center-frequency tuning around 0.7 GHz, an insertion loss smaller than 5 dB and a return loss higher than 13 dB, at the center frequency. For a ~50 MHz fixed bandwidth, the center frequency can be tuned between 0.51 GHz and 0.81 GHz, leading to a relative ± 24%-center-frequency tuning. Finally, the total physical length of the prototype filter is about 0.27λ_c at 0.7 GHz.

Index Terms – Tunable devices, Semiconductor varactor, Hybrid, Filter, Phase shifter, Impedance transformer.

I. Introduction

Tunable devices constitute a great challenge in the field of modern RF/microwave telecommunications. In the near future, more and more applications will use multiple operating frequencies.

In this context, some researches are carried out at IMEP and LAHC to show the feasibility of various tunable devices that can be embedded in RF/microwave telecommunication systems like phase shifters, tuners, filters, and power dividers.

Several technologies can be used to achieve such tunable devices: ferroelectric materials, ferromagnetic materials, MEMS, piezoelectric devices, optical control, semiconductors.

In this paper, we will describe simulation and measurement results obtained at IMEP and LAHC on tunable devices like phase shifters, tuners (impedance transformers), and filters, in a hybrid technology, with the use of varactors (reverse-biased Schottky diodes). Varactor diodes have been used in multiple non linear

[1, 2] or linear [3, 4] applications, either in monolithic (MMICs) or hybrid forms.

All devices exhibited in this paper were designed in coplanar waveguide (CPW) technology because it facilitates the soldering of shunt varactors. To avoid odd-mode propagation, these shunt varactors are soldered on each side of the CPW central conductor to maintain symmetry. Reverse-biased silicon Schottky diodes are used as varactors to obtain tunable devices. Simulations and optimizations of these devices were carried out using ANSOFT Designer [5].

Measurements show the feasibility and potentialities of such devices. The next step will consist in implementing these devices in Silicon MMICs to demonstrate their feasibility for front-end systems.

II. Phase shifter

A) Background and basic principle

A new conception method is described to achieve tunable phase shifters, in a low cost hybrid technology. A transmission line is periodically loaded by varactors as

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shown in Figure 1. This principle has been described in [6].

Two innovations are carried out to realize low loss and well-matched phase shifters:

- addition of a series inductance with the varactors;
- addition of near- and far-end tapered sections.

A reverse-biased Schottky diode acts as a voltage variable capacitor. This allows the propagation constant tuning, and then the phase shifts.

Adding an inductance in series with the diodes leads to an improvement of C_{\max}/C_{\min} (maximum to minimum capacitance value) ratio, leading to shorter devices because fewer diodes are necessary to obtain a given tunable phase shift. This leads to devices with lower insertion losses. The equivalent capacitance of the varactors of capacitance C in series with an inductance L can be written as: $C_{eq} = C/[1-(\omega/\omega_0)^2]$, with $\omega_0 = 1/\sqrt{LC}$. Depending on the value of ω in comparison with ω_0 , C_{eq} can be much greater than C . It is easy to show that it is the same for C_{\max}/C_{\min} .

The return loss also constitutes a problem with this phase-shifter topology. The tuning of the loading capacitors leads to a variation of the phase-shifter characteristic impedance. To address this problem, tapered sections have been added at near- and far-end of the phase shifter. The bias of the near- and far-end diodes is different from the others, leading to the possibility of matching the device for every phase shift value. This contributes to the compactness of the phase shifter. Without tapered sections, a large C_{\max}/C_{\min} leads to a large characteristic impedance mismatch, and a compromise is necessary. If C_{\max}/C_{\min} increases, then the mismatch increases, but if C_{\max}/C_{\min} decreases, then more sections have to be used for the same tunable phase shift, leading to the use of more varactors and longer devices, and increased insertion losses.

B) Results

A prototype of phase shifter working at 1 GHz, with nine sections, has been produced and measured to validate our approach. Silicon diodes from MACOM have been used (MA46H071). The substrate was the Rogers RO4003 ($\epsilon_r = 3.36$, $h = 813 \mu\text{m}$ and $\tan(\delta) = 1.5 \times 10^{-3}$, leading to $\epsilon_{r\text{eff}} \epsilon_r \sim 1.4$ @ 1 GHz). The characteristic impedance Z_0 equals 180Ω , leading to a central conductor width $W_0 = 1 \text{ mm}$, and a gap width $G_0 = 5 \text{ mm}$. Physical lengths of near- and far-end tapered sections, and central sections are $d_{\text{tap}} = 17 \text{ mm}$ ($\sim 0.067 \lambda_c$) and $d_{\text{center}} = 10 \text{ mm}$ ($\sim 0.039 \lambda_c$) at f_c leading to a phase-shifter length of about $0.41 \lambda_c$, where λ_c is the guided wavelength at the operating frequency f_c .

Figure 2 (a) shows the measured phase shift in degrees versus the frequency in GHz. The reference bias is

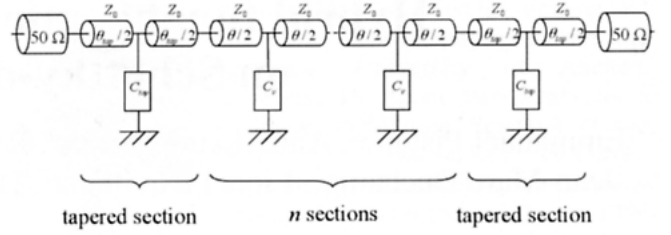


Fig. 1. Topology of phase shifter.

0.5 V for a 10 V bias voltage. Figure 2 (b) shows the insertion losses. They are 2.4 dB maximum for a 360° phase shift. Figure 2 (c) shows the return loss, it is better than 20 dB for a 15% bandwidth around 1 GHz.

These results show a great improvement compared to those presented in [6] (4 dB insertion losses and 10 dB return loss) and clearly validates the addition of an inductance in series with the varactors and the tapering

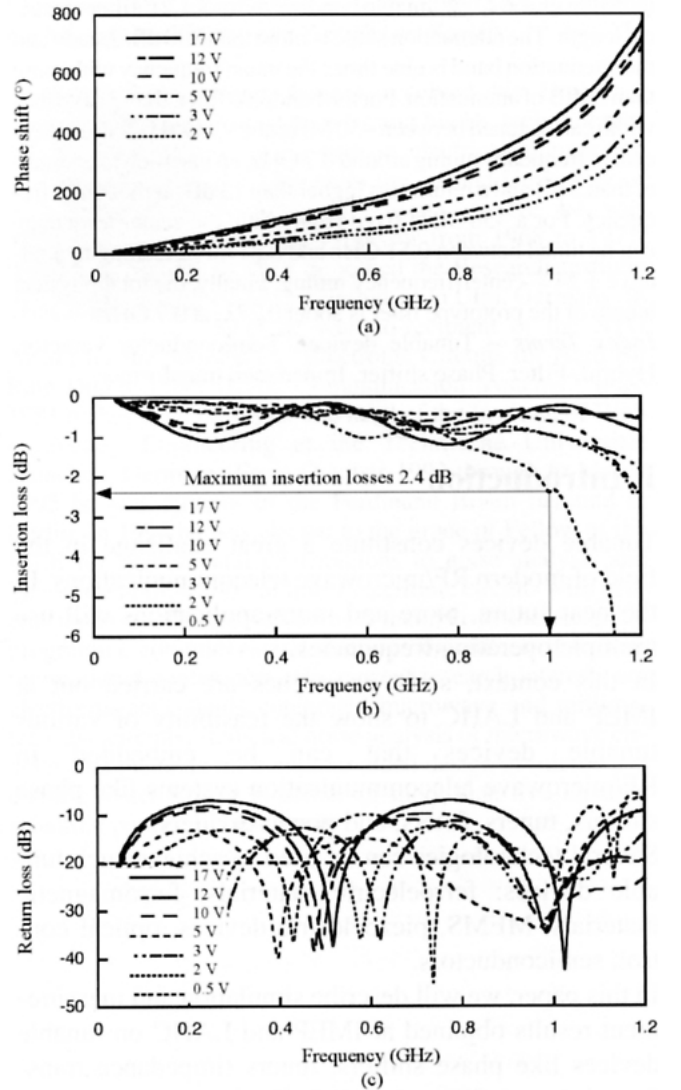


Fig. 2. Measurement results of the 1-GHz phase shifter. (a) Phase shift versus frequency. (b) Insertion loss. (c) Return loss.

of the near- and far-end sections to improve insertion losses and return loss. Moreover, the tapered sections allow a wide working frequency range, typically from 800 MHz to 1200 MHz, with the same characteristics: 2.4 dB insertion losses, 20 dB return loss and 15% bandwidth.

III. Impedance transformers

The goal was to improve the impedance transformer demonstrated in [7]. In [7], a *CLC* (variable capacitance – inductance – capacitance) topology was proposed. The voltage variable capacitance was realized by the use of a voltage variable capacitor embedded between two quarter-wave transformers. This leads to a half-guided-wave long device. In this paper, two topologies of compact impedance transformers have been studied. The first one is comparable to the phase shifter described above. A transmission line of about 120° electrical length is loaded by two varactors. The second one is based on the *CLC* structure used in [7], but the variable inductance is realized by a voltage variable capacitor in series with a high impedance small length (about 30°) transmission line and a fixed inductor.

A) Transformer 1

1) *Background and Basic Principle*: The principle of the first impedance transformer is given in Figure 3. A transmission line of characteristic impedance Z_0 is loaded by two reverse-biased diodes acting as voltage variable capacitors.

Both pairs of diodes are differently biased (V_0 and V_1 voltages). The device has been optimized to obtain a 20 dB return loss for a wide range of loads around 50 Ω .

2) *Results*: A prototype transformer has been realized with Philips BAT62-03W diodes on FR4 substrate ($\epsilon_r = 4$, $h = 1.5$ mm and $\tan(\delta) = 10^{-2}$), with a total physical length equal to $\lambda_c/3$, where λ_c is the guided wavelength at the operating frequency, and a characteristic impedance $Z_0 = 118 \Omega$, leading to a central conductor width $W_0 = 1$ mm, and a gap width $G_0 = 2$ mm.

These diodes exhibit poor performance – a variable capacitance from 350 fF to 220 fF, for a 0-40 V reverse bias. The series resistance is 6.6 Ω , leading to high insertion losses. However this first prototype allows validating our approach. Figure 4 shows simulation results after optimization and Figure 5 shows corresponding measurement results. The load is 50 Ω .

Figure 4 shows that the load can be tuned in a 15% bandwidth around 5 GHz. The large insertion losses are due to the diodes series resistance. Figure 5 shows

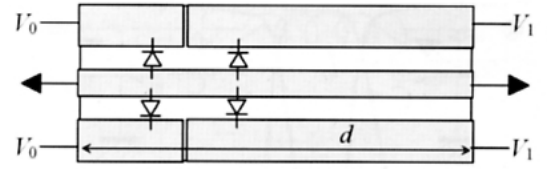


Fig. 3. First impedance transformer. Principle.

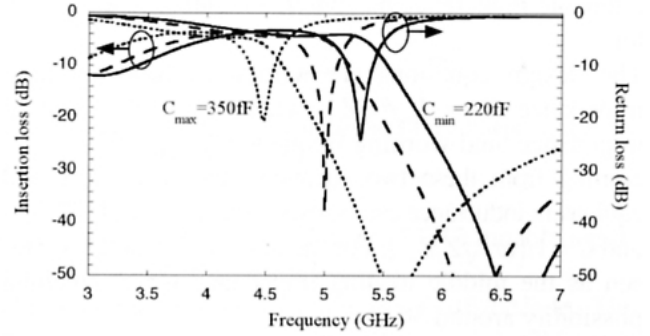


Fig. 4. First impedance transformer. Simulations.

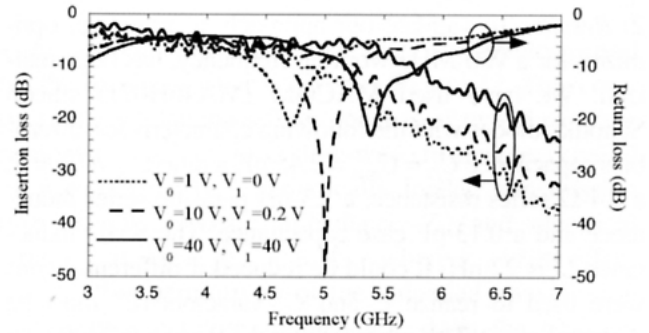


Fig. 5. First impedance transformer. Measurements.

that the agreement between simulations and measurements is quite good. The difference can result from an approximate parasitic model (case inductance and capacitance) given by the manufacturer.

B) Transformer 2

1) *Background and Basic Principle*: The second impedance transformer is based on the topology described in Figure 6.

C and C' are variable capacitances. The inductance L is realized in practice by a high characteristic impedance transmission line in series with a fixed inductance. The equivalent impedance of the LC' circuit can be written as: $\bar{Z} = [1 - (\omega/\omega_c)^2] / (jC'\omega)$ with $\omega_c = 1/\sqrt{LC'}$. When the $\omega > \omega_c$ condition is fulfilled, \bar{Z} corresponds to an equivalent inductance: $L_{eq} = [(\omega/\omega_c)^2 - 1] / (C'\omega^2)$. This inductance is a voltage variable inductance if C' is a voltage variable capacitance. Then it can be said that

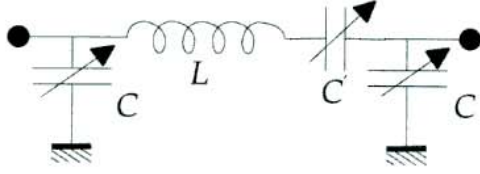


Fig. 6. Second impedance transformer. Electrical equivalent circuit.

a tunable inductance has been fabricated, like a varactor.

The design equations are written as follows: input impedance $Z_i = L_{eq}/(CZ_l)$, where Z_l is the loading impedance, and working frequency $\omega_f = 1/\sqrt{L_{eq}C}$.

Starting from these two equations, the capacitance and equivalent inductance can be expressed as $L_{eq} = \sqrt{Z_i Z_{in}}/\omega_f$ and $C = 1/(\omega_f \sqrt{Z_i Z_{in}})$. For the design, $Z_l = 50 \Omega$ is chosen as the middle loading impedance for a matching possibility around 50 Ω .

The device has been optimized in term of return loss, for the determination of the characteristic impedance and length of the transmission line used to realize the inductance L , and the diode choice.

2) *Results:* To validate our approach, a prototype, optimized for a 900 MHz working frequency, has been realized. We have used MACOM (MA46H071) silicon Schottky diodes with the following characteristics: a zero-bias capacitance $C_{j0} = C_{max} = 2.45$ pF, a ratio $C_{max}/C_{min} = 7$, a 0.4- Ω series resistance, a 1.3-nH parasitic series inductance and a 0.13-pF case capacitance. The SMD inductance L' is 22 nH. It could be reduced if different diodes were used to realize C and C' varactors (C' must be greater than C). The substrate was the Rogers RO4003 ($\epsilon = 3.36$, $h = 813 \mu\text{m}$ and $\tan(\delta) = 1.5 \times 10^{-3}$). The high characteristic impedance transmission line has the following properties: characteristic impedance $Z_0 = 102 \Omega$, length $d = \lambda/16$ central conductor width $W = 3.27$ mm and gap width $G = 2$ mm. The diodes are all identical but are differently biased, by voltages V_1 and V_2 .

The topology of the realized impedance transformer is given in Figure 7.

C_s is a decoupling capacitance of 1 nF, necessary for the separation of the two biases.

Figure 8 shows the simulation of the device $|S_{11}|$ (and $|S_{21}|$), after optimization, for a 50 Ω load.

Measurements for the same 50 Ω load are shown in Figure 9. The agreement between simulations and measurements is quite good. The shift in the working frequencies probably is due to mistaken specification of diode parasitics. Measurements show a very good return loss $|S_{11}| < -20$ dB over a wide bandwidth of $\pm 36\%$ around 900 MHz.

Figure 10 shows measurement results for the extreme loads of 16.5 Ω and 280 Ω . Outside this interval, the

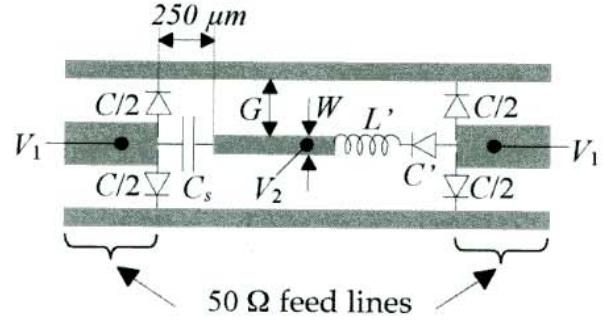


Fig. 7. Second impedance transformer. Topology.

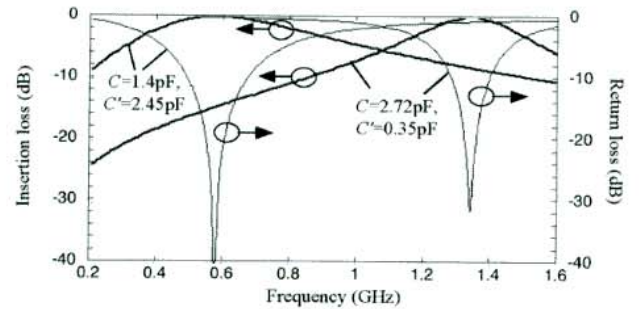


Fig. 8. Second impedance transformer. Simulations for a 50 Ω load.

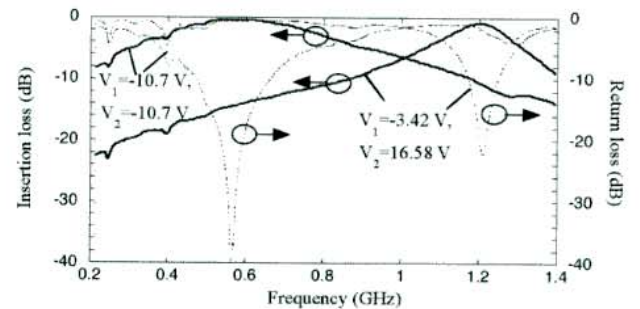


Fig. 9. Second impedance transformer. Measurements for a 50 Ω load.

load can no more be matched for $|S_{11}| < -20$ dB. For all measurements, insertion losses are always limited to 1 dB. These results are very promising for the achievement of a very compact and large bandwidth impedance transformer.

IV. Filters

Tunable filters constitute an interesting way to reduce size and cost of microwave multiband systems since they can perform requisites for all operating bands. The great number of publications demonstrating tunable microwave filters [8-12] confirms the importance of these solutions. In [11] and [12], tune-all bandpass

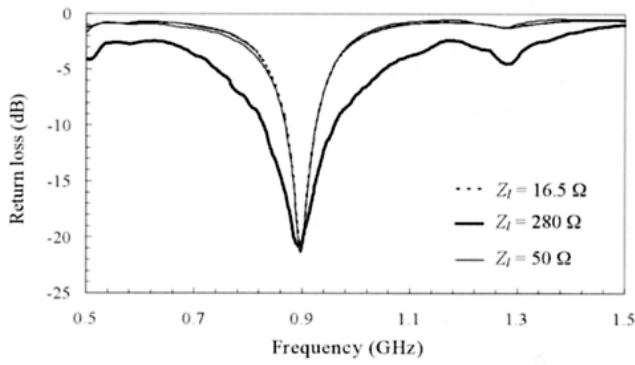


Fig. 10. Second impedance transformer. Measurements for 16.5 Ω and 280 Ω loads.

filters show wide tunings around 0.7 GHz. Nevertheless, undesirable peaks appear in the filter behavior, either in low frequencies [11], or in high frequencies [12].

In this section, compact CPW tunable lowpass and tune-all bandpass filters are exhibited, showing wide continuous tunings, without any spurious peak below five times the operating frequency.

These filters are carried out on Rogers RT Duroid 5880 substrate, with relative permittivity $\epsilon_r = 2.2$, dielectric loss, $\tan(\delta) = 1.5 \times 10^{-3}$ height $h = 1.5$ mm, and copper thickness $t = 35$ μm , leading to $\epsilon_{\text{reff}} \sim 1.5$ for the considered frequency range.

A) Lowpass filter

1) *Background and Basic Principle:* Based on periodic structures [13], these lowpass filters are very compact, selective (attenuation slopes reaching hundreds of dBs per decade are obtained), and does not exhibit any spurious peak below ten times the mean cut-off frequency. The basic principle for lowpass filters is described in [14], and the topology is that of the phase shifter shown in Figure 1. A transmission line is periodically loaded by reverse-biased Schottky diodes acting as voltage variable capacitors. Called tapered sections, the near- and far-end sections are different from the other ones to achieve a good return loss.

2) *Results:* A prototype filter has been carried out around the 0.66 GHz operating frequency. The transmission line characteristic impedance is 170 Ω , leading to a central conductor width $W_0 = 0.48$ mm, and a gap width $G_0 = 2$ mm. This lowpass filter constituted by four elementary sections has a 130° electrical length at the mean -1 dB cut-off frequency (0.66 GHz). MACOM (MA4ST1240) Schottky diodes have been used as varactors. Figure 11 shows results of the cut-off-frequency tuning range for a return loss better than 16 dB.

The agreement between simulations and measurements

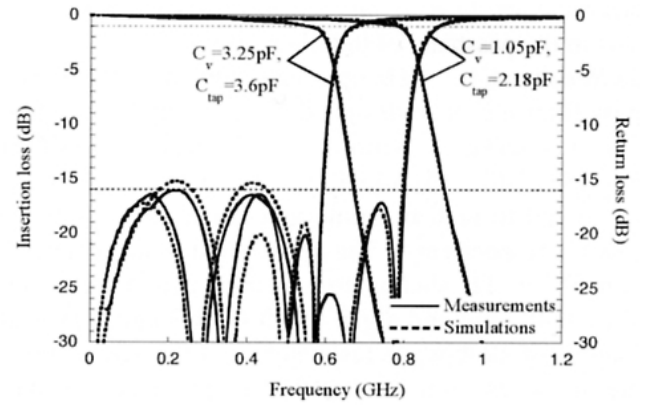


Fig. 11. Measurements and simulations of the lowpass filter for a return loss better than 16 dB.

is very good. Figure 11 exhibits a $\pm 17\%$ tunable bandwidth for a -16 dB return loss.

The attenuation slope is about -306 dB/decade. The attenuation is better than -25 dB up to 7.2 GHz, that is, up to ten times the mean cut-off frequency.

B) Bandpass filter

1) *Background and Basic Principle:* The topology used for the bandpass filter is given in Figure 12. This compact tune-all bandpass filter consists of two slow-wave resonators. Each of these slow-wave resonators is based on a capacitive (C_v) periodically loaded transmission line of characteristic impedance Z_0 , and electrical length, $\theta_0 = (2\pi \sqrt{\epsilon_{\text{reff}}} f d_0) / c$ where ϵ_{reff} is the effective relative permittivity of the unloaded transmission line at the operating frequency f , d_0 is the physical length of the unloaded resonator, and c is the velocity of light in vacuum. The slow-wave phenomenon leads to an electrical length of the resonator θ_l higher than θ_0 , leading to a compact device. When shunt capacitors C_v are replaced by varactors, the loaded electrical length θ_l of the slow-wave resonators can be tuned, leading to a center-frequency tuning. These resonators are loaded at near- and far-end by series varactors C_s to tune the couplings between resonators and hence to control the bandwidth of the filter. Finally, near- and far-end high impedance transmission lines called tapered sections are used. The characteristic impedance of these tapered sections is Z_0 , like the slow-wave resonators, but the

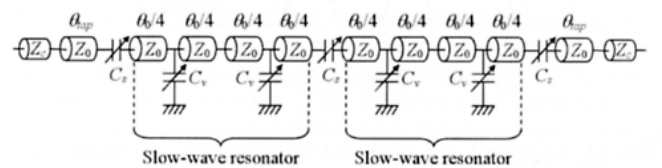


Fig. 12. Topology of the tune-all bandpass filter ($Z_c = 50$ Ω).

electrical length θ_{top} is optimized to improve the return loss in the passband of the filter.

2) *Results*: A 0.7-GHz optimized tune-all CPW bandpass filter has been designed. Commercially available reverse-biased Schottky diodes MACOM MA4ST1240 (MA46H071, respectively) were employed to realize shunt varactors C_v (series varactors C_s , respectively). The characteristic impedance Z_0 equals 170Ω , leading to a central conductor width $W_0 = 0.48$ mm, and a gap width $G_0 = 2$ mm. Physical lengths of slow-wave resonators and tapered sections are $d_0 = 24$ mm ($\sim 0.069 \lambda_c$) and $d_0 = 24$ mm ($\sim 0.063 \lambda_c$) at $f_c = 0.7$ GHz, leading to a filter length of about $0.27 \lambda_c$, where λ_c is the guided wavelength at the center operating frequency f_c .

Figure 13 (Figure 14, respectively) compares measured and simulated S parameters of the realized tune-all bandpass filter for a ~ 50 MHz (~ 78 MHz bandwidth, respectively). Shunt and series varactors bias $V(C_v)$ and $V(C_s)$ vary from 0 V to 2 V and 0 V to 5.9 V, respectively.

Figure 13 and Figure 14 show a good agreement between simulated and measured insertion losses. The simulations have been carried out considering:

- shunt varactors with a junction capacitance $C_j(V)$ varying from 3.2 pF to 8.8 pF, a series resistance $R_s = 1.6 \Omega$, a series inductance $L_s = 1.2$ nH, and a case capacitance $C_c = 0.11$ pF,
- series varactors with a junction capacitance $C_j(V)$ varying from 0.7 pF to 2.5 pF, $R_s = 0.7 \Omega$, $L_s = 0.45$ nH, and $C_c = 0.15$ pF.

Measured return loss is better than expected and is always better than 13 dB at f_c . The center-frequency tuning of this tune-all bandpass filter reaches about $\pm 24\%$ ($\pm 18\%$, respectively) for a ~ 50 MHz (~ 78

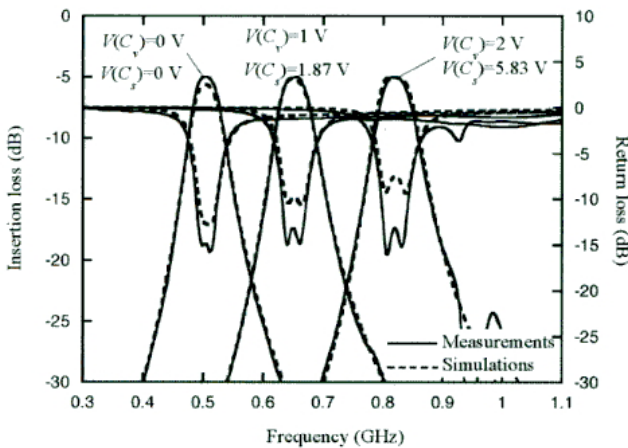


Fig. 13. Simulated and measured insertion loss and return loss for the tune-all bandpass filter: center-frequency tuning for a ~ 50 MHz bandwidth.

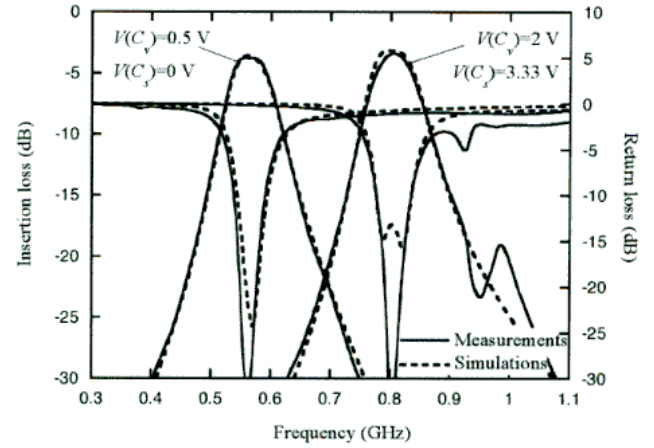


Fig. 14. Simulated and measured insertion loss and return loss for the tune-all bandpass filter: center-frequency tuning for a ~ 78 MHz bandwidth.

MHz, respectively) bandwidth and for an insertion loss smaller than 5 dB (3.8 dB, respectively) at f_c .

V. Conclusion

In this paper, some tunable RF device topologies have been demonstrated in a hybrid technology. Experimental results show the great interest of the studied topologies. For phase shifters, we obtain more compact devices than previously published with low insertion losses and high return loss. Very compact tunable impedance transformers have been produced, with low insertion losses (< 1 dB) and good return loss (better than 20 dB) for large tunable load impedances (from 16.5Ω to 280Ω). Tunable lowpass and bandpass filters have also been demonstrated, with wide continuous tunings, and a good return loss. These filters, based on periodic structures, lead also to spurious lobes far above the operating frequency (more than five times).

All the topologies exhibited in this paper can be implemented in a MMIC process to address the future challenge concerning the investigation of front-end systems with single signal processing path for a wide range of frequencies.

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Anne-Laure Perrier was born in France in 1980. She received the M. Sc. Degree in Optic, Optoelectronic and Microwaves from the INPG (National Polytechnical Institute of Grenoble), Grenoble, France, in 2003.

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Her research interest His research interest includes the theory, design, and production of tunable impedance transformers.



Raffi Bourtoutian was born in Lebanon, in 1981. He received the Electronics and Telecommunications Engineer degree from the Faculty of Engineering of the Lebanese University, Tripoli, Lebanon, in July 2004. He's pursuing studies to receive the M.S. degree in optics, optoelectronics and microwaves from the Institut National Polytechnique de Grenoble (INPG), France. In August 2003, he joined the

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