# A MMIC BROAD-BAND 90° POWER DIVIDER USING A NEW ALL-PASS ACTIVE FILTER

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**Abstract :**

**The design and performances of a new broad-band active MMIC 90° power splitter are described. The chip provides a gain of -1 dB with a 1° differential phase error, a 0.1 dB amplitude imbalance and input and output return losses better than 15 dB in the 0.5-3 GHz frequency range.**

## I- INTRODUCTION

In modern microwave transmission, digital modulation techniques such as phase shift keying (PSK) or quadrature amplitude modulation (QAM) are more and more used to achieve effective power and spectrum utilization. To get high-quality transmission, the modulator needs two microwave carriers in quadrature with both very low phase error and low amplitude imbalance. Today, 90° power dividers are realized by branch line couplers or equivalent lumped components, Lange couplers or low-pass/high-pass filters, featuring 3 dB transmission losses and limiting the band to about 40%.

In this context, we developed an original broad-band 0.5-3 GHz MMIC 90° power splitter based on a new compact architecture of an all-pass active filter. In addition to achieving low quadrature phase error and low amplitude imbalance, this structure also provides active matching, constant output amplitude and the possibility to tune the gain of each quadrature component within a 15 dB range.

## II- NEW ARCHITECTURE OF AN ALL-PASS ACTIVE FILTER

This new architecture (depicted in Fig. 1) consists in three stages: a first all-pass stage, a common source amplifier and a second all-pass stage. All-pass stages include a common source MESFET or HEMT with a feedback parallel R-C circuit between the drain and gate. Such a single all-pass stage has been proposed earlier by Rubin for a bipolar transistor [1].

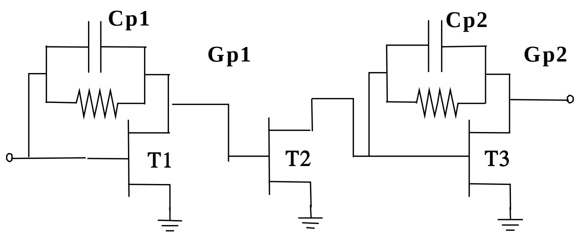



*Fig. 1: New all-pass active filter*

Using a simple transistor model, it can be shown that the transfer scattering parameter $S_{21}$ can be written as the product of two first order all-pass functions as follows:

$$S_{21} = K * \frac{1 - \frac{j\omega}{\omega_1}}{1 + \frac{j\omega}{\omega_1}} * \frac{1 - \frac{j\omega}{\omega_2}}{1 + \frac{j\omega}{\omega_2}} \quad \text{with}$$

$$\omega_1 = \frac{gm_1 - Gp_1}{Cp_1} \quad \text{and} \quad \omega_2 = \frac{gm_3 - Gp_2}{Cp_2}$$

provided that the following all-pass conditions are satisfied: $Gp_1 = gm_1 / 2$ and $Gp_2 = gm_3 / 2$ , where $gm_1$ and $gm_3$ are the transconductances of T1 and T3.

## III- BROAD-BAND 0.5-3 GHz 90° POWER DIVIDER

The 90° active power divider is designed using two all-pass active filters as described in section II, with their inputs joined (see Fig. 2). The middle common source amplifier stage is a gain control amplifier with a biasing circuit proposed by Rumelhard et al. [2]. Input and output HEMTs are scaled to provide active matching [3] and feedback resistors are chosen to satisfy all-pass conditions.

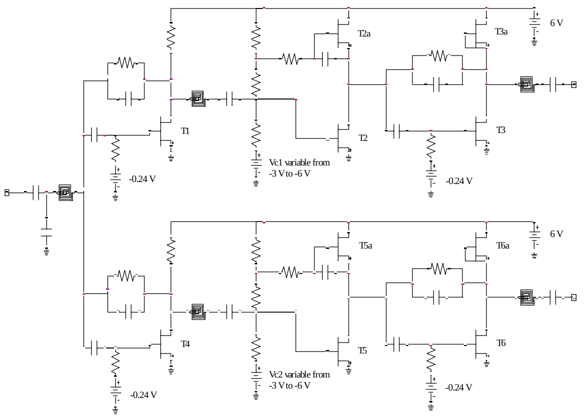



Fig. 2: Electric circuit of 90° power divider

The two all-pass filters are quasi symmetrical except for the values of the resonant frequencies $\omega_{1i}$ and $\omega_{2.j}$. The transfer scattering parameters $S_{21}$ and $S_{31}$ can be written as follows:

$$S_{21} = K * \frac{1 - \frac{j\omega}{\omega_{11}}}{1 + \frac{j\omega}{\omega_{11}}} * \frac{1 - \frac{j\omega}{\omega_{12}}}{1 + \frac{j\omega}{\omega_{12}}}$$

and

$$S_{31} = K * \frac{1 - \frac{j\omega}{\omega_{21}}}{1 + \frac{j\omega}{\omega_{21}}} * \frac{1 - \frac{j\omega}{\omega_{22}}}{1 + \frac{j\omega}{\omega_{22}}}$$

Feedback capacitors are adjusted to obtain a differential phase $\Delta\phi$ between the two paths equal to $90 \pm 1°$ on the possible maximum bandwidth. This is realized assuming that the resonant pulsations satisfy the following conditions:

$\omega_{11} = \omega_0 * \cot(x_1)$, $\omega_{12} = \omega_0 * \cot(x_2)$, $\omega_{21} = \omega_0 * \tan(x_1)$ and $\omega_{22} = \omega_0 * \tan(x_2)$, where $\omega_0$ is the central pulsation and $x_1$ and $x_2$ are two variables defined by $x_1 + x_2 = \frac{3\pi}{8}$.

Fig. 3 shows that, by varying $x_1$, it is possible to obtain phase quadrature between the two outputs with less than 2° error on a minimal frequency band of 40% and a maximal frequency band of 330%, i.e. more than a decade.

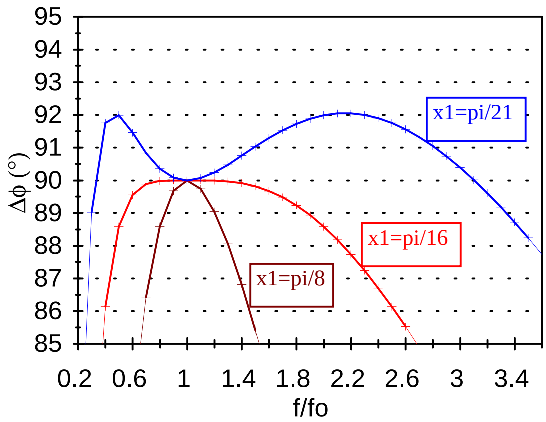



*Fig. 3: Phase difference between the two all-pass filters*

To design a MMIC 90° power divider, we chose $f_0 = 1.2$ GHz and $x_1 = \pi/19$ so as to obtain $\Delta\phi = 90 \pm 1°$ on the frequency band 0.5-3 GHz.

The simulations of electric circuit show that, on the 0.5-3 GHz band, the input and output return losses are better than 15 dB while the transmission scattering parameters $S_{21}$ and $S_{31}$ present a maximal gain of $-0.4 \pm 0.2$ dB with an amplitude imbalance less than 0.1 dB. Moreover, the gain control amplifier allows the gain of the two quadrature components to be attenuated independently on a range of 15 dB with less than 10° differential phase shift.

## IV- EXPERIMENTAL RESULTS

Our layout pattern preserves symmetry for all devices (Fig. 4). The chip size is 2 mm x 1.5 mm and has been manufactured using the PML ED02AH process with 0.2 µm PHEMT.

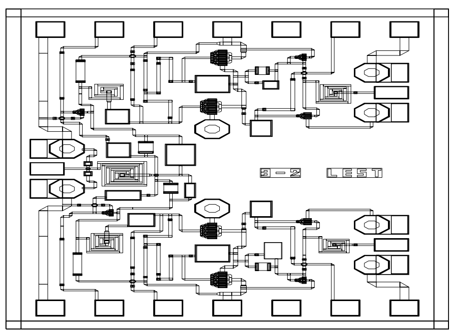
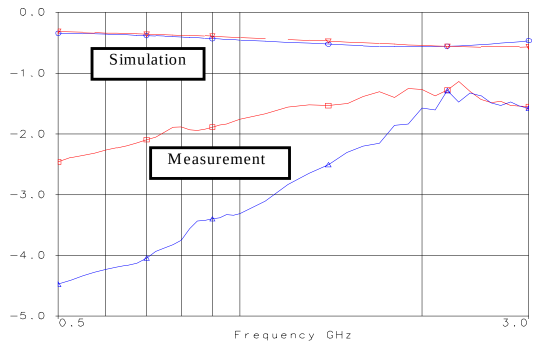
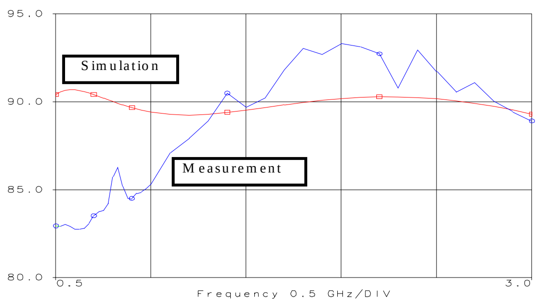


*Figure 4: MMIC of 90° power divider*

The simulation and measured results obtained from microwave probing are presented in Fig. 5-a and 5-b with the control voltages $Vc_1$ and $Vc_2$ set to get maximal gain.




*a)*




*b)*

*Fig. 5: Simulation and experimental results*
*a)* $|S_{21}|$, $|S_{31}|$ *in dB*
*b)* $\Delta\phi$ ( $|S_{21}|$ / $|S_{31}|$ ) *in degree*

Simulation results do not fit well with measurements, in particular for the magnitudes on the lower frequency range. These differences are due to: a) the shift of the threshold voltages Vt of the manufactured PHEMTs compared to the values used for modeling; b) the parasitic effects of DC probes which act like an LC parallel resonant circuit and to the lack of DC decoupling capacitor on chip.

## V- REVERSE ENGINEERING

To confirm these hypotheses, we led the reverse engineering of the 90° power divider, changing the threshold voltages Vt in PHEMTs models and the DC probes by the equivalent resonant circuit of Fig. 6.

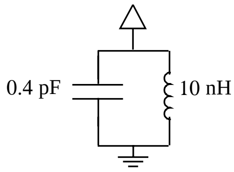
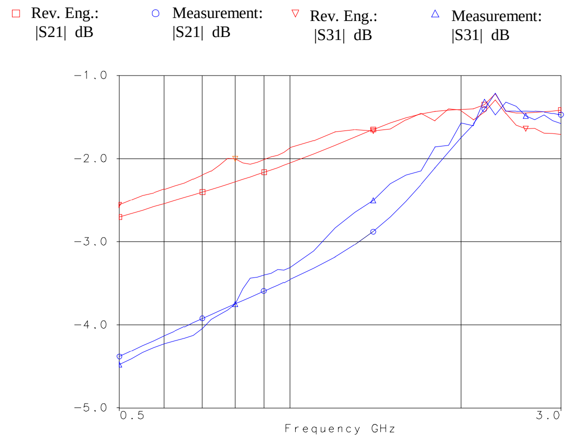
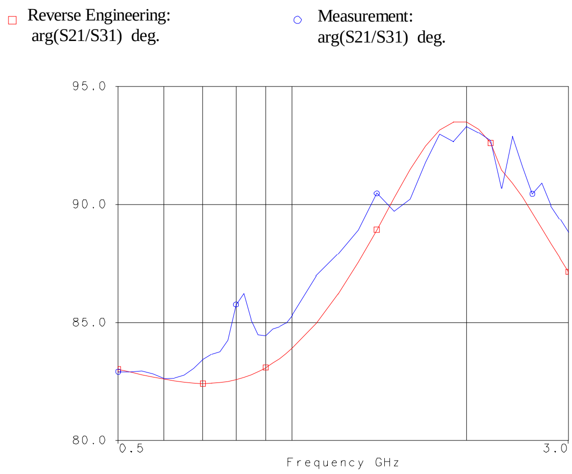



*Figure 6: DC probe equivalent circuit*

Fig. 7 displays the new simulation and experimental results for the same bias points. The two results being in good agreement validate this reverse engineering survey.



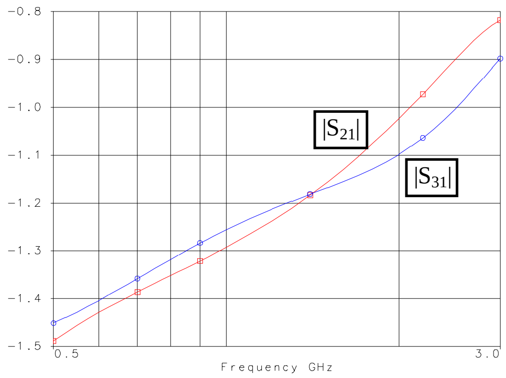

*a)*



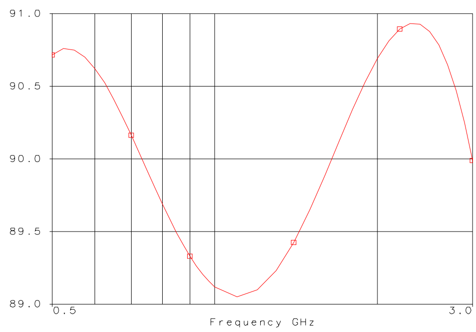

*b)*

*Figure 7: Reverse engineering and measured results*

Considering these results, we have then tuned the gate voltage of the HEMTs all-pass stages to meet the first predicted performances. New simulation results presented in Fig. 8 show that it is possible, by decoupling DC supplies with off chip capacitors, to obtain a maximal gain in each quadrature component of -1.2 ± 0.4 dB with less than 1° phase error and 0.1 dB amplitude imbalance.




*a)* $|S_{21}|$ *and* $|S_{31}|$ *in dB*




*b)* $\Delta\phi$ *in degree*

*Figure 8: New simulation results*

## VI- CONCLUSION

A new broad-band MMIC 90° power splitter using an original all-pass active filter has been proposed and demonstrated experimentally. The device presents good active matching and constant output magnitude. It exhibits also low quadrature phase error and low amplitude imbalance. Moreover, each quadrature component amplitude can be adjusted on a 15 dB range with less than 10° differential phase shift. In comparison with large-sized, lossy and narrowband passive structures, this new 90° active power divider is very attractive for the design of vector generators and microwave direct conversion I-Q modulators in MMIC technology.

## VII- REFERENCES

[1] H. Rubin and R.K. Even: « Single transistor all-pass networks », IEEE Transactions on Circuit Theory, Vol. 20, N° 1, pp 24-30, Jan. 1973.

[2] C. Rumelhard and B. Carnez : « A new biasing circuit for high integration density GaAs MMIC », Proc. EuMC, Sept. 1985.

[3] K. Niclas : « GaAs MESFET feedback amplifiers, design considerations and characteristics », Microwave Journal, Vol. 23, pp. 39-48, Mar. 1980.