A NEW TIME DOMAIN WAVEFORM MEASUREMENT SETUP TO INVESTIGATE INTERNAL NODE VOLTAGES IN MMICs.

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INTRODUCTION

Time domain waveforms engineering at internal nodes of MMICs is essential for the phenomenologic understanding of the nonlinear elements behavior inside of MMICs. This topic is extremely enriching in various fields as varied as the circuits reliability, the nonlinear parmetric stability and, in a general way, the optimal design and behavior survey of MMICs.

This approach, usually performed in simulation, is possible henceforth in instrumentation and characterization. The description of the measurement setup and the calibration procedure is the object of this paper.

The heart of the measurement setup have to be an instrument which provide the capabilities to measure RF multi-harmonics time domain waveforms. The 'Large Signal Network Analyser' (LSNA) is a perfect solution to perform this kind of measurements. It enables time domain waveform measurements of a device under test, provides a NIST validated calibration (specilly a phase calibration) procedure and is commercially available through 'Maury Microwave'.

This paper presents a way to modify the classical LSNA system and its calibration procedure in order to enable time domain waveforms at internal nodes of MMICs with high impedance probes (HIP). The HIP calibration and measurement procedures can be easily added to the original LSNA driver software. Indead, the HIPs are considered like a classical test-set through the software and do not modify the traditionnal use of the LSNA.

This work is applicated here to the characterization of a class-F amplifier at S band.

THE LARGE SIGNAL NETWORK ANALYSER (LSNA) [1]

The LSNA could be seen as a vectorial network analyser (VNA) enhanced to nonlinear devices characterization. The waves (incident / reflected waves or voltages and currents) are measured in an absolute way : the LSNA enables waves magnitudes and phases measurements (and not ratios only like with VNAs) and performs measurements at differents frequencies simultaneously (a CW fundamental and all harmonics).

The characterization tool [2]

The LSNA hardware architecture is very close from the VNA one. The main difference is the receiver part (downconverter process) : a sampling convertor take the place of the classical mixer. The sampling rate is around 20 MHz and leads to a signal frequency-compressed within a 10 MHz bandwidth. This low-frequency downconverted signal is sampled with ADC.

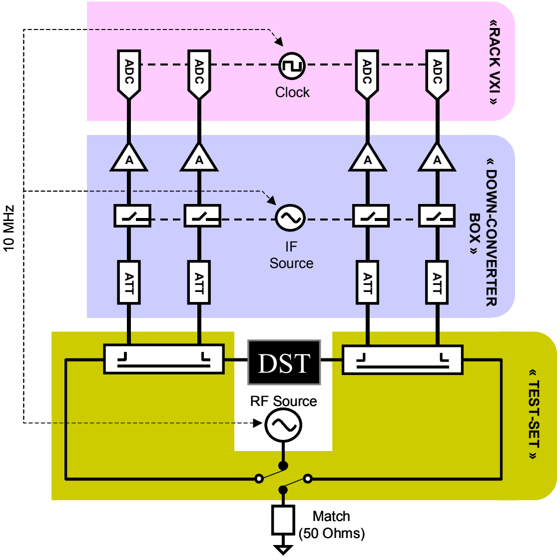


Fig. 1. LSNA : the simplified architecture

The calibration procedure [3]

The calibration error-model of the LSNA is defined as a 8-terms matrix. 7 of thoses 8 terms are known by a classical VNA relative calibration (SOLT or LRRM). The last (8th) term defines the absolute calibration in order to measure accuratly both amplitude and phase of a single travelling wave. This term is calculated from a calibration with an amplitude standard (powermeter) and a phase standard (the harmonic phase reference : HPR). The relationship between the non-corrected raw-data (r_i) and the real physical travelling waves at the reference plane (a_i and b_i) can be written as the 8-term following matrix :

$$\begin{pmatrix} a_1 \\ b_1 \\ a_2 \\ b_2 \end{pmatrix}_{DUT} = \left|\tilde{K}\right| . e^{j.\varphi\{\tilde{K}\}} \bullet \begin{bmatrix} 1 & \beta_I & 0 & 0 \\ \gamma_I & \delta_I & 0 & 0 \\ 0 & 0 & \alpha_{II} & \beta_{II} \\ 0 & 0 & \gamma_{II} & \delta_{II} \end{bmatrix} \bullet \begin{pmatrix} r1 \\ r2 \\ r3 \\ r4 \end{pmatrix} \qquad (1)$$

The LSNA calibration procedure is proceed in 3 steps :

- A classical VNA relative calibration to define 7 of the 8 errors terms ;
- An amplitude calibration by measuring a forward waveform with both the LSNA and a powermeter to define the magnitude of the 8th error term ;
- A phase calibration by connecting the reference phase generator at a port of the LSNA to deduce the phase correction of the 8th error term.

Then, the complete calibration matrix is computed as follow :

$$\begin{pmatrix} v1 \\ i1 \\ v2 \\ i2 \end{pmatrix}_{LSNA} = \begin{bmatrix} \alpha_1^{CAL} & \beta_1^{CAL} & 0 & 0 \\ \gamma_1^{CAL} & \delta_1^{CAL} & 0 & 0 \\ 0 & 0 & \alpha_2^{CAL} & \beta_2^{CAL} \\ 0 & 0 & \gamma_2^{CAL} & \delta_2^{CAL} \end{bmatrix} \bullet \begin{pmatrix} r1 \\ r2 \\ r3 \\ r4 \end{pmatrix}_{LSNA} \qquad (2)$$

The phase standard (HPR) is a step recovery diode (SRD). It generates a very stable pulse. This pulse contains a lot of frequencies. The phase relationship between these frequencies has been precharacterize either with a RF scope calibrated with the ‘nose to nose’ technique [3] (the Agilent’s method) or with a electro-optic sampling system [4] (the NIST’s method).

ABOUT THE USE OF HIGH IMPEDANCE PROBES (HIP)

Basic considerations

As it has been shown in the calibration matrix expression, the corrected quantities which are measured by the LSNA are the voltage v and current i. These terms are deduced from the voltage traveling waves (or voltage normalized pseudo-waves) a and b such as :

$$\begin{cases} a = V_+ \\ b = V_- \end{cases} \text{ and } \begin{cases} v = a + b \\ i = \dfrac{a-b}{Z_0} \end{cases} \text{ with } Z_0 = 50 \text{ ohms} \qquad (3)\ (4)$$

We are going to use the same formalism about the high impedance probe in order to deduce the pseudo-wave a and b from the corrected measured voltage v. With this approach, Z_0 is the caracteristic impedance of the probed line and thus could include an imaginary part at low frequencies [5].

According to the telegraphists equations, the incident (a or V_+) and reflected (b or V_-) pseudo-waves can be deduced from 2 simultaneous voltage measurements (v_1 and v_2) on a lonely probed propagation line as follow :

$$\begin{cases} V_+ = \dfrac{v_1.e^{+\gamma.L} - v_2}{2.\mathrm{Sinh}(\gamma.L)} \\ V_- = \dfrac{v_1.e^{-\gamma.L} - v_2}{-2.\mathrm{Sinh}(\gamma.L)} \end{cases} \qquad (5)$$

where γ is the propagation constant of the propagation line and L the distance between the two probing locations.

Thus, we need the propagation constant of the probed line (γ) to obtain the voltage travelling waves and both the propagation constant and the characteristic impedance (Z_0) to calculate the current travelling waves.

Previous works about HIPs

The first works about this topic have been realized by J.C.M. Hwang from the Lehigh university. He has used for that purpose a Microwave Transcient Analyser (MTA) [6] [7] [8]. The results have illutrated the great interest of this kind of measurement setup but the calibration procedure was not fully described. Indead, the high impedance probe have to be preliminary characterized. The MTA is not a fully calibrated mesurement instrument and introduces phase distortion which is neglected but not negligible [2].

In [9], U. Arz describes a classical and a new optimized method to get the S-parameters of a HIP with a VNA. Those S-parameters provide a practical mean for correcting raw measurements. In [10], P. Kabos illustrates and validates calibrated time-domain oscilloscope measurements using these previous HIP S-parameters. He comes to the following expression :

$$V(f) = V_{OSC}(f).[K(f)] \qquad (6)$$

$$\text{with } K(f) = \frac{1 - S_{22}(f).\Gamma(f)}{S_{21}} + S_{11}(f).\frac{1 - S_{22}(f).\Gamma(f)}{S_{21}} + S_{21}(f).\Gamma(f) \qquad (7)$$

V(f) is the corrected voltage at the tip of the HIP. K(f) define a voltage transfert function of the HIP. $\Gamma(f)$ is the reflexion coefficient of the scope and the S parameters are the HIP's ones.

At last, theses methods (with the MTA or the scope) are complex and require several measurements and connections / disconnections. The LSNA architecture enables more flexibility with the use of HIP. Then it becomes very easy to determine the value of K(f) in (6).

THE NEW SETUP AND ITS CALIBRATION PROCEDURES

Our measurement setup use the LSNA in a « receiver mode » like. Two HIPs are connected to 2 receiver channels of the LSNA and are similar to a lumped reflectometer. This is the port 1 of our new test-set located into the device under test. The port 2 use the classical test-set (reflectometer) to perform measurements at the output of the circuit.

To calibrate this new measurement setup, we just have to use the LSNA standard calibration code to generate a 8-terms calibration matrix and then to swap the 4-error terms relatives to the 'port 1' with the new ones relatives to the HIPs. A new 'Mathematica' module, developped for this new approach, performs this tranformation automatically : finally, the user drives the measurement setup like a classical LSNA.

The calibration procedure is performed in 3 steps.

The first step consists in calibrating the LSNA with a probe station in a classical way (LRRM process, powermeter and HPR). This procedure generates the complete 8-terms calibration matrix. Then, we disconnect the input channels 1 and 2 of the downconverter box from the test-set and connect the HIPs as illustrated figure 2. The port 2 is calibrated yet, and will be consider as our new voltage (or current) standard in order to calibrate high impedance probes.

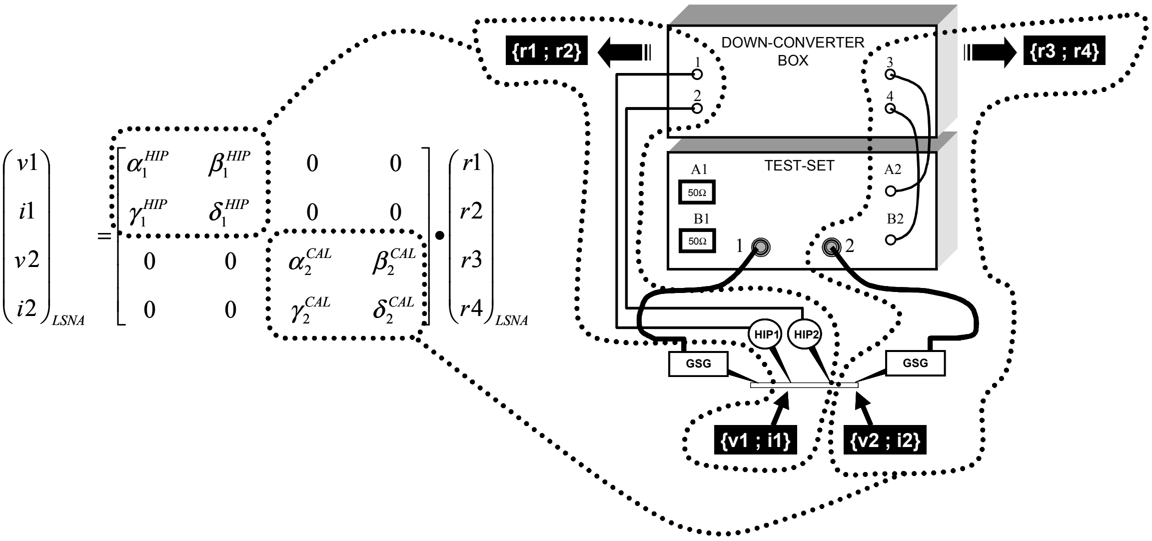
The second step consists in probing sequentially the thru line of the calibration kit with the HIPs at the port 2 reference plane. Then we can calculate the ratio of 2 measurements : v2 (the error corrected measurements provided by port 2) and r1 or r2 (the down-converted raw data corresponding to the probing HIPs). This ratio (calculated in the frequency domain for each frequency of a sweep-sin) looks like K(f) in (6) ; but here, the voltage transfert function describe not only the HIP but also a bias tee (placed between the HIP and the down-converter box to protect the sampling convertor from continuous current within disable DC measurements) and the downconverter hardware in the receiver :

$$\tilde{K}_1(f)=\frac{v2(f)}{r1(f)} \text{ and } \tilde{K}_2(f)=\frac{v2(f)}{r2(f)} \tag{8}$$

The third step is the generation of a new calibration sub-matrix associated to the port 1 of the test-set as follow :

$$\begin{pmatrix} v1 \\ i1 \end{pmatrix}=\begin{bmatrix} \alpha_1^{HIP} & \beta_1^{HIP} \\ \gamma_1^{HIP} & \delta_1^{HIP} \end{bmatrix}\bullet\begin{pmatrix} r1 \\ r2 \end{pmatrix} \tag{9}$$

We have to calculate 4 terms and include this sub-matrix in the whole system calibration matrix as illustrated in figure2. Figure 2 shows the final measurement setup. r1 and r2 are the uncorrected raw data coming from the high impedance probes. r3 and r4 are the raw data from port 2. v2 and i2 are the corrected measurements at the calibrated port 2 reference plane (tip of the GSG probe). v1 and i1 are deduced from r1 and r2 and represent the physical quantities at the tip of one high impedance probe.



$$\begin{pmatrix} v1 \\ i1 \\ v2 \\ i2 \end{pmatrix}_{LSNA}=\begin{bmatrix} \alpha_1^{HIP} & \beta_1^{HIP} & 0 & 0 \\ \gamma_1^{HIP} & \delta_1^{HIP} & 0 & 0 \\ 0 & 0 & \alpha_2^{CAL} & \beta_2^{CAL} \\ 0 & 0 & \gamma_2^{CAL} & \delta_2^{CAL} \end{bmatrix}\bullet\begin{pmatrix} r1 \\ r2 \\ r3 \\ r4 \end{pmatrix}_{LSNA}$$

Fig. 2. LSNA associated with HIPs : the measurement setup. The port 2 is used in a traditional way.
The port 1 is used s a “receiver mode”.

There are 3 differents methods to calculate the port 1 calibration sub-matrix. The 2 first methods use as an assumption the relation (6). A third method don’t need any assumption but require accurate values of the propagation constant and the caracteristic impedance of the probed line even to deduce a single voltage value.

Figure 3 illustrate K(f) (the HIP voltage transfert function versus the frequency) obtained at the second calibration step. This information is essential to build the calibration sub-matrix at the third step by using one of the 2 first methods.

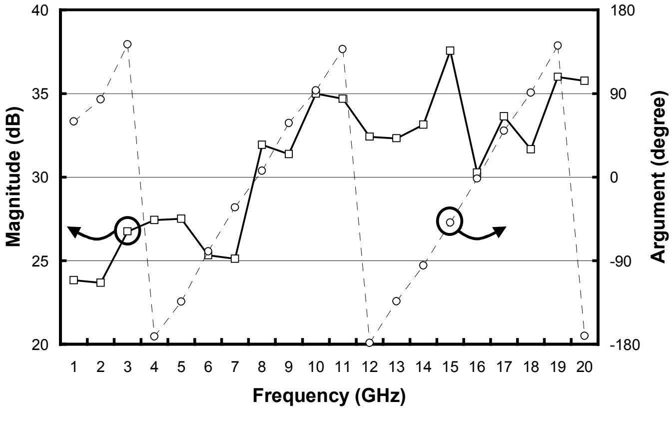


Fig. 3. Magnitude and phase of K(f) : the voltage transfert function between the tip of one high impedance probe and the associated raw data.

Method 1

This method is an easy way to enable 2 simultaneously corrected voltage measurements with HIPs. This method consists on the use of K_1 and K_2 (the pre-characterizated ratio of 2 HIPs) to define the matrix (9) :

$$\begin{pmatrix} v1 \\ i1 \end{pmatrix} = \begin{bmatrix} \tilde{K}_1 & 0 \\ 0 & \tilde{K}_2 \end{bmatrix} \bullet \begin{pmatrix} r1 \\ r2 \end{pmatrix} \tag{10}$$

With this new calibration sub-matrix, we can obtain the voltage at the tips of both HIP but not the currents. Thus, in the LSNA software, the i1 waveform is not the current but the voltage at the tip of the second HIP. This method has to be used when we require independents HIPs.

Method 2

Some applications require the measurement of the current. Then, a second possibility has been implemented in the software to fix the errors terms consequently. According to the telegraphists equations, the user have to know the distance between the 2 HIPs, the propagation constant of the probed line (γ) and its characteristic impedance (Z_C).

$$\begin{pmatrix} v1 \\ i1 \end{pmatrix} = \begin{bmatrix} \tilde{K}_1 & 0 \\ \left(\dfrac{\tilde{K}_1 . \mathrm{Coth}(\gamma . L)}{Z_C} \right) & \left(\dfrac{-\tilde{K}_2}{Z_C . \mathrm{Sinh}(\gamma . L)} \right) \end{bmatrix} \bullet \begin{pmatrix} r1 \\ r2 \end{pmatrix} \tag{11}$$

But the use of the K(f) ratio cancel any coupling effect between the 2 HIPs because the calculated β_1^{HIP} in (9) is null with this method. To take into account this coupling effect, we have to proceed to a new second step in the calibration procedure and define a sub-matrix instead of a single K(f) ratio.

Method 3 : Calibration of coupled HIPs

This method don't use any assumption about a K(f) voltage ratio. The calibration of the high impedance probes has to be done with 2 HIPs simultaneously on a transmission line. One of the HIPs has to be placed at the port 2 reference plane (standard of our HIP calibration). The aim of the calibration, with this method, is to deduce the complete 4-terms errors matrix (9) instead of the K(f) ratio (8).

To find out the 4 errors terms associated to the HIPs, we have to solve 2 equations with 4 unknown factors, so we need 2 different voltage/current standards. It's possible by removing the classical 50 ohms termination at the output of the test set (port 'RF IN 2') with an 'Open' or a 'Short'. The corresponding modified voltage and current standards are measured at the reference plane with the calibrated port 2. This procedure provides additive measurement necessary to determine the 4-terms error matrix relative to the HIPs. Here come the calibration sub-matrix :

$$\begin{pmatrix} v1 \\ i1 \end{pmatrix} = \begin{bmatrix} \left(\dfrac{r2_{C2}.v2_{C1} - r2_{C1}.v2_{C2}}{r1_{C1}.r2_{C2} - r1_{C2}.r2_{C1}} \right) & \left(\dfrac{r1_{C2}.v2_{C1} - r1_{C1}.v2_{C2}}{r1_{C2}.r2_{C1} - r1_{C1}.r2_{C2}} \right) \\ \left(\dfrac{r2_{C2}.i2_{C1} - r2_{C1}.i2_{C2}}{r1_{C1}.r2_{C2} - r1_{C2}.r2_{C1}} \right) & \left(\dfrac{r1_{C2}.i2_{C1} - r1_{C1}.i2_{C2}}{r1_{C2}.r2_{C1} - r1_{C1}.r2_{C2}} \right) \end{bmatrix} \bullet \begin{pmatrix} r1 \\ r2 \end{pmatrix} \tag{12}$$

where the subscripts C1 and C2 indicate the condition of measurements ('Open' or 'Short' at the output of the test-set) for both the raw data (r1 and r2) and the standards at the GSG tip reference plane (indeed, the corrected measurements v2 and i2).

During the measurements, the distance between HIPs, the propagation constant and the characteristic impedance must be equal to the ones used (but unknown) during calibration.

This restriction implies that we have to use the same substrate during calibration and measurements.

Method 3 : Measuring on a substrate different from the calibration one

To proceed a HIP calibration on a substrat and perform some measurements on another substrate, we need to know all the characteristic parameters during the calibration and the measurements which are the propagation constants, the characteristic impedances and the distances between HIP of both substrates. Then, according to the telegraphists equations and assuming that 2 HIPs behave like a lumped-reflectometer with 4 wave-coupling factors as illustated figure 4, we can create a new sub-matrix error from the one obtained during the calibration procedure and the characteristic parameters values by extracting theses 4 coupling factors values.

We can deduce a new calibration sub-matrix correspounding to a measurement substrate (subsripted 'MEA') from the complete calibration sub-matrix (9) identified with the calibration substrate (subsripted 'CAL') and the physical parameters of both substrats (the distance between the 2 probe locations, the propagation constant γ and the caracteristic impedance Z_C) :

$$\alpha_{1\,\mathrm{NEW}}^{\mathrm{HIP}} = \frac{\mathrm{Sinh}(\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})}{\mathrm{Sinh}(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}})}.\left[\left(\alpha_1^{\mathrm{HIP}}.\mathrm{Cosh}((\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})-(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}}))\right)-\left(Z_{\mathrm{C\,CAL}}.\gamma_1^{\mathrm{HIP}}.\mathrm{Sinh}((\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})-(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}}))\right)\right] \tag{13}$$

$$\beta_{1\,\mathrm{NEW}}^{\mathrm{HIP}} = \beta_1^{\mathrm{HIP}}.\frac{\mathrm{Sinh}(\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})}{\mathrm{Sinh}(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}})} \tag{14}$$

$$\gamma_{1\,\mathrm{NEW}}^{\mathrm{HIP}} = \frac{1}{Z_{\mathrm{C\,MEA}}}.\frac{\mathrm{Sinh}(\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})}{\mathrm{Sinh}(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}})}.\left[\left(Z_{\mathrm{C\,CAL}}.\gamma_1^{\mathrm{HIP}}.\mathrm{Cosh}((\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})-(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}}))\right)-\left(\alpha_1^{\mathrm{HIP}}.\mathrm{Sinh}((\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})-(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}}))\right)\right] \tag{15}$$

$$\delta_{1\,\mathrm{NEW}}^{\mathrm{HIP}} = \delta_1^{\mathrm{HIP}}.\frac{Z_{\mathrm{C\,CAL}}}{Z_{\mathrm{C\,MEA}}}.\frac{\mathrm{Sinh}(\gamma_{\mathrm{CAL}}.L_{\mathrm{CAL}})}{\mathrm{Sinh}(\gamma_{\mathrm{MEA}}.L_{\mathrm{MEA}})} \tag{16}$$

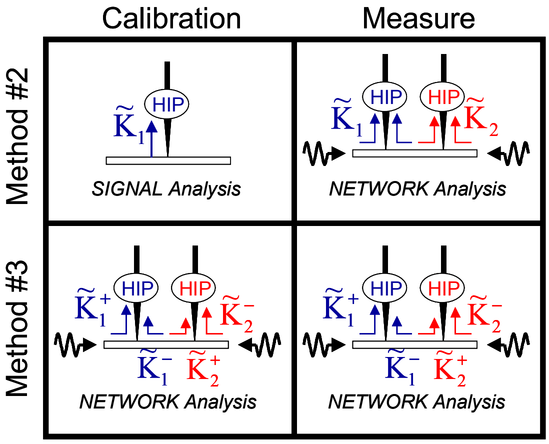


Fig. 4. Calibration and measurements with methods 2 and 3. Method 2 does not take into account the coupling effect between the 2 HIPs : the incident (+) and reflected (-) waves have the same transfert function.

CALIBRATION COMPARISON (GSG AND HIP)

The method number one has been used, first to measure the voltage along a line and secondly to test the design of a power amplifier.

Several voltage measurements have been made on a 50Ω line during a 'sweep-sin' excitation (different power levels and frequencies of the RF source). Figure 5 shows the voltage measured with a calibrated GSG probe (conventional use of the LSNA) and with a calibrated HIP on a propagation line which not use during the calibration. The comparison is quite good.

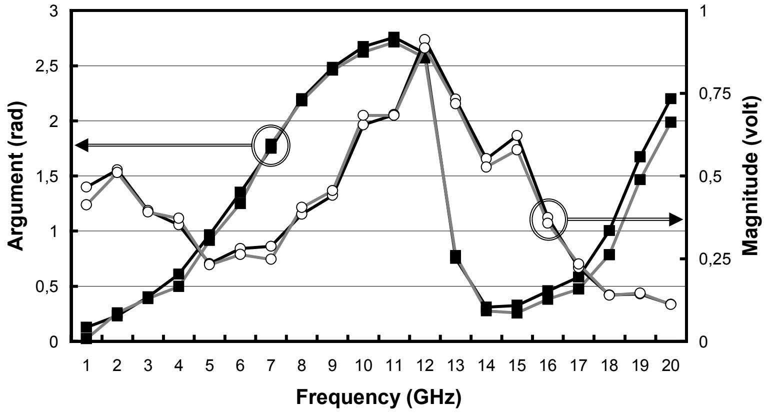


Fig. 5. Calibrated voltage measurements at the tip of a GSG probe and a HIP during a sweep-sin.

APPLICATION TO THE CHARACTERIZATION OF A CLASS-F POWER AMPLIFIER

A class-F power amplifier has been measured. Basically, an optimized class-F amplifier have to present a square voltage waveform at the output of the transistors but not necessarily at the output of the amplifier because of the filtering effects of the matching and the power combining networks. Internal waveform probing is the lonely experimental way to check the pertinent and successful design and realization of such topology.

The device under test

The device under test is a MMIC S-band one-stage power amplifier. The circuit is built in HBT technology. Design goal was to produce an optimized class F for a fundamental frequency equal to 2.15 GHz [11]. The bias are Vbe0 = 1 volts and Vce0 = 9 volts. Because of the class F, the expected efficiencies of the transistor are about 70%.

We probed the voltage at differents location on the MMIC for several input power. Figures 6 to 8 illustrate the measurements results. The time domain waveforms are displayed upon 930 ps (2 fundamental periods).

Measurement Results

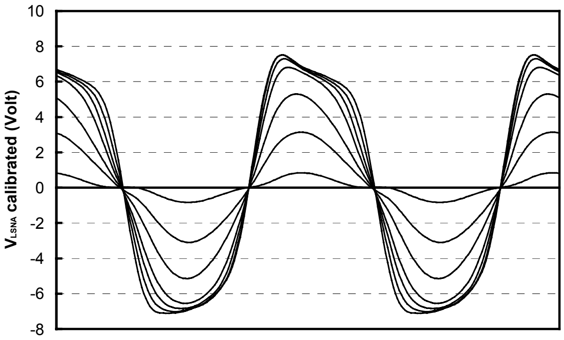


Fig. 6. Calibrated voltage measurements at the output of the MMIC.

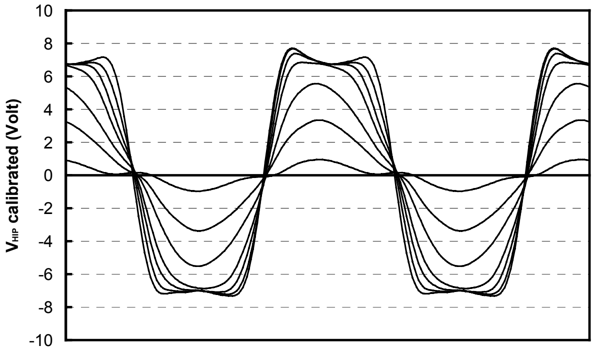


Fig. 8. Calibrated voltage measurements at the output of a HBT.

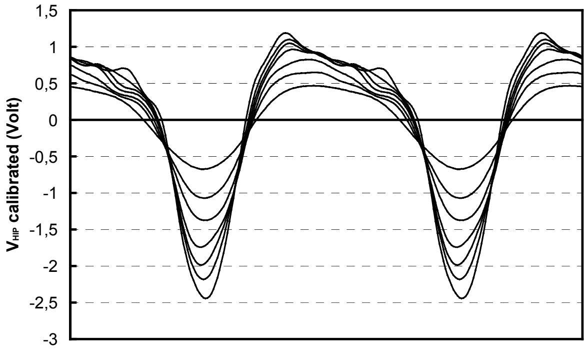


Fig. 7. Calibrated voltage measurements at the input of a HBT.

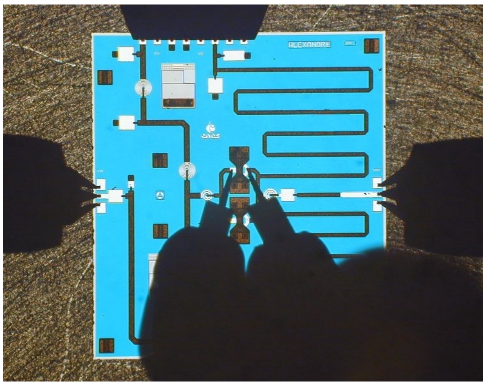


Fig. 9. Picture of the MMIC during measurements.

Figure 9 shows a picture of the experiment. We have used one dual positioner (FPD 100 : Fine-Pitch Dual Positioner) to handle 2 high impedance probes (Cascade FPS-20X : Fine Pitch Signal) [12]. The probes are not connected to the ground, neither during calibration nor measurements [9].

The output of the MMIC has been measured with both a GSG probe (LSNA's port 2) and a calibrated HIP. The time domain waveforms are strictely the sames. The measurements at the input of the transistors lead to typical waveforms. At high input power, we can see a quasi-square voltage waveform at the output of each transistor which validate the design topology and method for this class-F amplifier demonstrator.

CONCLUSION

High impedance probe calibrations and measurements procedures have been described and implemented in the LSNA software (a single 'Mathematica' module). This module has to be linked to the original 'Mathematica' kernel of the LSNA driver. This tool provides a friendly and easy way to investigate waveforms at internal nodes of MMICs.

An application has been presented concerning the validation of an high efficiency amplifier design. This measurement system is expected to be essential for the stability analysis of MMIC because it offers an unique experimental verification of even and odd ocillation modes [13].

As a conclusion, the use of the LSNA with HIPs strongly reinforces the links needed between the measurement and the simulation environments. It enables a better understanding of nonlinear effects and contributes to improved optimized designs of reliable MMICs.

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