A novel fast calibration method for NVNAs based linearity setup

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***Abstract*—This paper deals with a new calibration procedure dedicated to on-wafer passive load-pull NVNA setups that allows the search of optimal impedances (associated with e.g. maximum Power Added Efficiency (PAE max), optimal linearity (NPR max). The originality of this work stands in a new calibration procedure for the load-pull setup that permits a drastic reduction of time compared with a conventional calibration. This setup is driven by a tailored Unequally Spaced Multi Tones (USMT) signal that has been already proved to be an efficient tool for the measurement of the Noise Power Ratio of nonlinear devices. The search for the optimal load impedance to reach the PAE max is applied to an UMS Foundry on-wafer 4 x 50 µm GaN transistor in Ka-band.**

Keywords*— *load-pull measurements, Power Added Efficiency, USMT signal, NPR measurements

I. Introduction

Solid-state Power amplifier (SSPA) circuits remain the most critical element of radiofrequency (RF) telecommunications and radars systems. 50-ohms SSPA (either in connectorized module or MMIC die) linearity is traditionally assessed by quasi-static AM/AM and AM/PM conversions or even with complex metrics as EVM, ACPR when signal generator and analyzer are available. However, there is a great need for such linearity characterizations dedicated to unmatched power transistors, that allows to bridge linearity requirements from the transistor selection and sizing, to the final circuit validation.

A connectorized load-pull setup based on a NVNA has been developed [1] for linearity (Noise Power Ratio: NPR) measurements using an original signal, termed Unequally Spaced Multi Tones signal (USMT). It is based on a multi frequency signal that uses a specific set of frequency tones (number of tones ≥8) and whose none of intermodulation tones overlap.

This paper presents an extension of this previous set-up dedicated to connectorized power modules to on-wafer load-pull linearity measurements of power transistors. In this case, the calibration procedure becomes much more complicated because it is impossible to measure with connectors in the DUT reference planes and because of the multiple positions of the tuner. Therefore, a new calibration procedure is proposed in this paper.

This paper is organized as following: section II describes the measurement setup with the detailed calibration procedure. Section III shows the load-pull measurement results to find the maximum Power Added Efficiency.

II. Measurement set-up

The upgraded set-up is shown in Fig. 1.

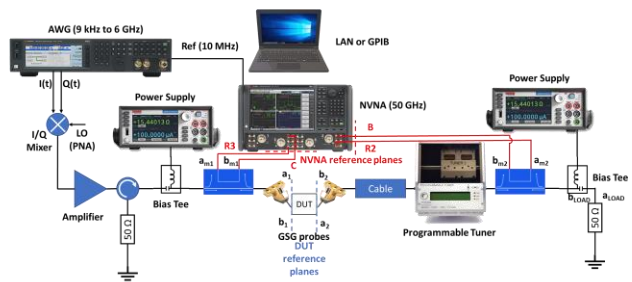

Fig. 1: Passive load-pull measurement bench for linearity measurement

The measurement set-up requires the use of couplers to allow the measurement of the incident and reflected waves (a_{m1}, b_{m1}, a_{m2}, b_{m2} in Fig. 1). Then it is necessary to calibrate the setup to determine relationships between the NVNA reference plane waves and the DUT reference plane waves (Fig. 1). A SOLR calibration procedure [3] allows to find the calibration matrix.

As shown in Fig. 2, the bias tee and the coupler are connected at the output of the tuner to reduce losses between the probes (DUT plane) and the tuner. The calibration matrix $\mathbf{M_1}$ (1) depends on the position of the tuner carriage and probes. Thus, the test bench must be calibrated for each tuner position. As the tuner has a huge number of positions, it is a very time-consuming task to calibrate all of them.

$$\begin{bmatrix} a_2 \\ b_2 \end{bmatrix} = [\mathrm{M}_1] \times \begin{bmatrix} a_{m2} \\ b_{m2} \end{bmatrix} \qquad (1)$$

The method we propose is to find a relationship between the tuner S-parameters and the calibration matrix. The first step is to perform S-parameter characterization of the tuner for a set of tuner positions, and is to be done only once.

Then, a standard calibration can be done, to determine the $\mathbf{M_1}$ matrix (1) which links the a_2, b_2 DUT reference plane waves and a_{m2}, b_{m2} NVNA reference plane waves. Note that the calibration matrix depends on the tuner position (x, y).

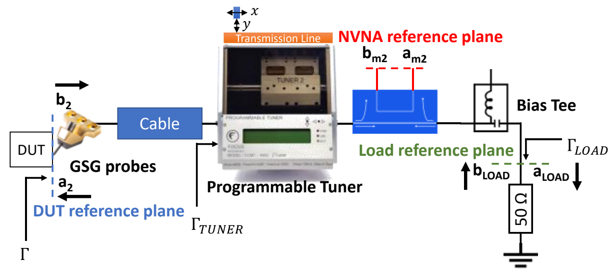

Fig. 2 : Setup output with reference planes

This relation (1) is rearranged to be expressed in terms of T-parameters. The matrix **T** links the T-parameters of the cable, the tuner, the bias-tee and the coupler (2).

$$\begin{bmatrix} b_2 \\ a_2 \end{bmatrix} = [T(x,y)] \times \begin{bmatrix} b_{m2} \\ a_{m2} \end{bmatrix} \quad (2)$$

with $[T(x,y)] = [T_{cable}] \times [T_{tuner}(x,y)] \times [T_{bias}] \times [T_{coupler}]$.

The matrix **T** depends on the tuner position, so a calibration matrix is to be extracted for each tuner position, which is very cumbersome. We propose an original solution by replacing these multiple measurements by only four measurements.

By vectorizing the 4 x 4 **T** matrix, we can link this matrix to the tuner matrix $\boldsymbol{T_{tuner}}$ as (3). A key observation is that this transformation allows to isolate the contribution of the tuner from the other components, namely the coupler, the bias tee and the cable, whose matrices $T_{coupler}$, T_{cable} and T_{bias} form the **C** matrix, that remains fixed. The problem can be rearranged as in (3). A detailed derivation is beyond the scope of this paper.

$$\overrightarrow{T^V} = [\mathbb{C}].\overrightarrow{T^V_{tuner}} \Rightarrow$$

$$\begin{bmatrix} T_{11}(x,y) \\ T_{12}(x,y) \\ T_{21}(x,y) \\ T_{22}(x,y) \end{bmatrix} = \begin{bmatrix} C_{11} & C_{12} & C_{13} & C_{14} \\ C_{21} & C_{22} & C_{23} & C_{24} \\ C_{31} & C_{32} & C_{33} & C_{34} \\ C_{41} & C_{42} & C_{43} & C_{44} \end{bmatrix} . \begin{bmatrix} T_{11}^{tuner}(x,y) \\ T_{12}^{tuner}(x,y) \\ T_{21}^{tuner}(x,y) \\ T_{22}^{tuner}(x,y) \end{bmatrix} \quad (3)$$

The **C** matrix is independent of the tuner position and only depends on the cable, bias-tee and coupler matrices which have fixed values. This **C** matrix can be obtained by only four measurements of **T** for 4 positions of the tuner. The Smith chart in Fig. 3 a) shows all the tuner positions measured once in blue while the tuner positions used to compute the **C** matrix are indicated by red points. From the knowledge of **C**, it is possible to calculate the calibration matrix **T** (3) for any position of the tuner without having to re-measure it.

Of course, this method includes the ability to determine the reflection coefficient at the DUT reference plane (**Γ** in the Fig. 2) as a function of tuner positions. The reflection coefficient **Γ** can be linked to the reflection coefficient $\boldsymbol{\Gamma_{LOAD}}$ (Fig. 2) (6), and it is possible to synthesize the desired value of **Γ**. However, the calibration matrix $\mathbf{M_1}$ (1) only links the waves at the DUT's output and NVNA reference planes. Since the waves are measured in the NVNA plane, it is then necessary to perform a SOL calibration to obtain the calibration matrix $\mathbf{M_2}$ linking the a_{m2}, b_{m2} NVNA reference plane waves and a_{load}, b_{load} load reference plane waves (4). The coefficient $\boldsymbol{\Gamma_{LOAD}}$ is frequency dependent since the connected load is not an ideal matched load.

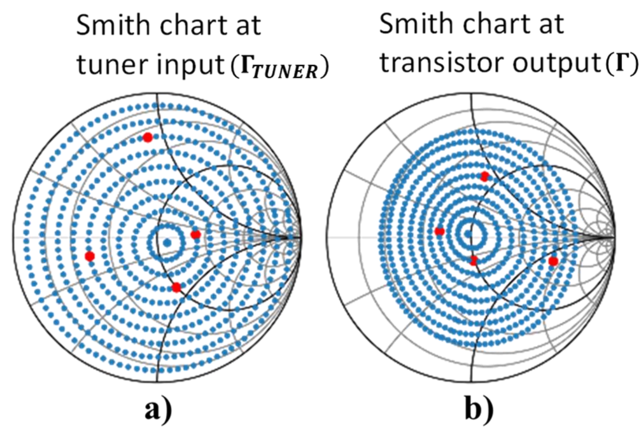

Fig. 3 : in blue: Measured tuner positions, in red: 4 tuner positions to compute C matrix

$$\begin{bmatrix} a_{load} \\ b_{load} \end{bmatrix} = [M_2] \times \begin{bmatrix} a_{m2} \\ b_{m2} \end{bmatrix} \quad (4)$$

The two calibration matrices $\mathbf{M_1}$ and $\boldsymbol{M_2}$ can be combined to produce a resulting matrix **M** that links the waves at the DUT and load reference planes (5), as shown in Fig. 2.

$$\begin{bmatrix} a_2 \\ b_2 \end{bmatrix} = [M] \times \begin{bmatrix} a_{load} \\ b_{load} \end{bmatrix} \quad (5)$$

with $[M] = [\mathrm{M}_1] \times [\mathrm{M}_2]^{-1}$

It is then possible to calculate the reflection coefficient **Γ** at the output of the device (6) A limitation of this method is that this reflection coefficient can only be estimated for the coordinates where the S-parameter characterization has been performed (blue points in Fig. 3 a)).

$$\boldsymbol{\Gamma} = \frac{1+M'_{12}\cdot\Gamma^{LOAD}}{M'_{21}+M'_{22}\cdot\Gamma^{LOAD}} \text{ where } \mathrm{M}'_{ij} = \frac{\mathrm{M}_{ij}}{\mathrm{M}_{11}} \quad \text{i,j} = 1,2 \qquad (6)$$

Fig. 3 shows the reflection coefficient $\boldsymbol{\Gamma}$ of the tuner obtained from an S-parameter characterization (a)) and the estimated reflection coefficient at the output of the DUT (b)). The four reflection coefficients selected for calibration are shown in red. At this frequency, it is shown that, unfortunately, in this version of the setup, the cable and connectors losses between the probe and the tuner reduce the range of magnitude of the synthesized gamma at the device output at about 0.6 than can be not sufficient for larger power devices.

III. Measurements and Results

A. CW operating mode

The following results have been measured with 4 x 50 µm HEMT GaN transistors from the UMS foundry (GH15) at 150 mA/mm (Class AB) and 20 V drain bias voltage. The objective is to determine the load impedance for the maximum Power Added Efficiency (PAE max) at 29 GHz.

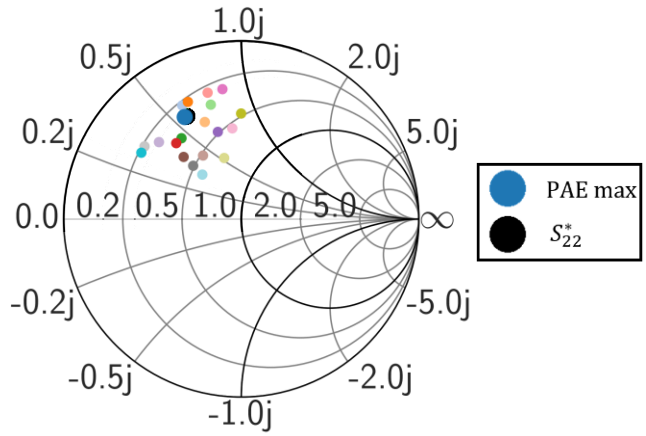

Fig. 4 : Load impedances presented to the transistor.

To avoid going through all the Smith chart we locate our search around the transistor S_{22}^* point (in black in Fig. 4) [4]. The chosen reflection coefficients are the nearest around S_{22}^* among the points measured in the tuner characterization.

The impedance giving the PAE max is the blue curve with a maximum value of 23% @ 29 GHz. For each impedance point, the gain, the DC drain current, output power, have been measured as shown in Fig. 5.

Due to the setup limitation (maximum reflection coefficient of 0.6 in the DUT reference plane), it is not possible to reach the PAE max of 48% measured by UMS for $|\Gamma| = 0.79$ at 30 GHz with a gain of 13 dB.

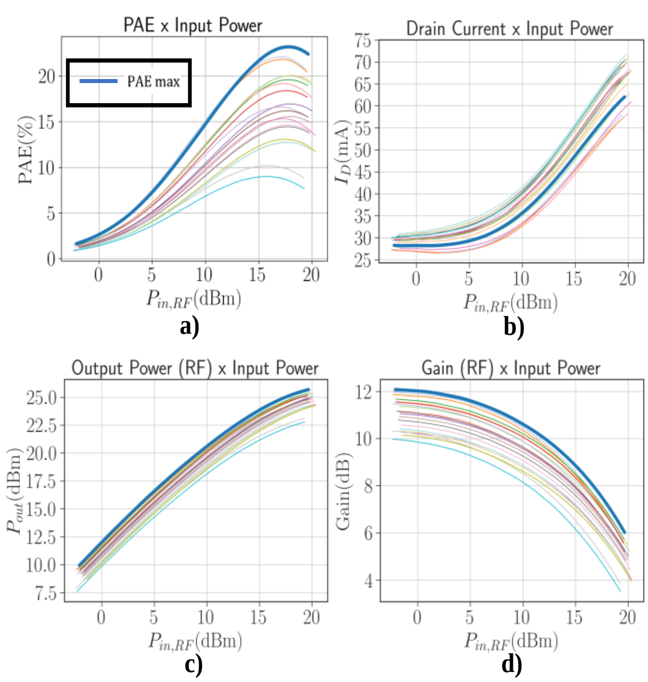

Fig. 5 : Measured power-added-efficiency (a), DC drain current (b), output power (RF) (c) and gain (d) versus transistor input power for the impedances presented in Fig. 4.

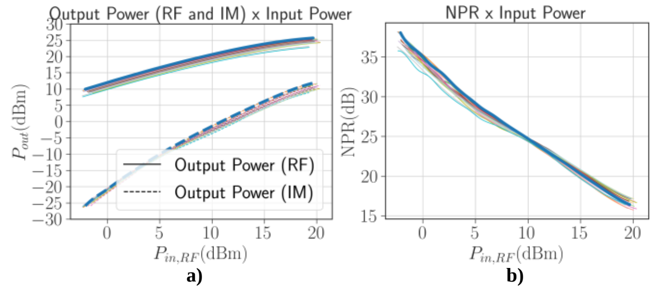

Fig. 6 : Output power, IM3 power (a) and NPR (b) versus input power for the impedances presented in Fig. 4.

B. *USMT signal operation mode*

To measure the Noise Power Ratio, the transistor is driven with an USMT signal explained in detail in [1]. The USMT signal baseband frequencies applied to the device are presented in Table I. These frequencies are translated in the upper band of the frequency $f_0 = 29$ GHz thanks to an external I/Q modulator (Fig. 1).

TABLE I. EIGHT BASEBAND FREQUENCIES

f_1 (MHz)	f_2 (MHz)	f_3 (MHz)	f_4 (MHz)
3	3.201	3.403	3.609
f_5 (MHz)	**f_6 (MHz)**	**f_7 (MHz)**	**f_8 (MHz)**
3.827	4.081	4.443	5.129

Due to the narrowband modulation, the magnitude of the reflection coefficient varies less than 8% in the band.

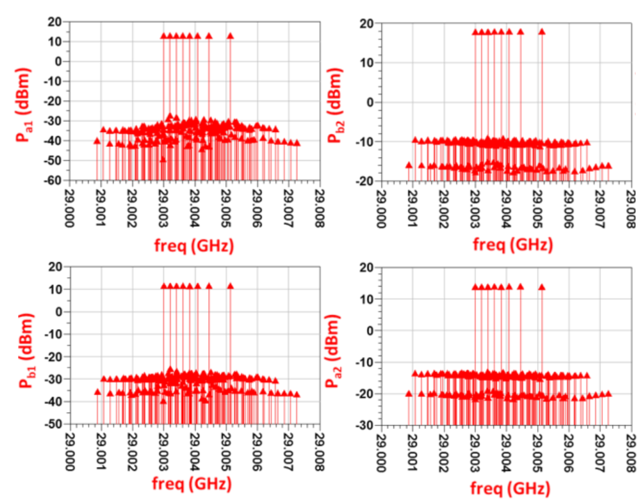

Fig. 7 : Spectrum of the USMT signal for an input power of 18.6 dBm for the impedance giving the PAE max.

Due to the non-overlapping properties of the USMT signal, average powers are simply calculated as the sum of each power of corresponding IM_3 frequencies [1]. To compute the NPR only frequencies between the 1st and 8th tones are considered. The input power (RF) is calculated by adding the input power of each of 8 carriers. Similarly, the RF output power is the sum of the power of each carrier at the device output. The component intrinsic NPR is computed by calculating the difference between the output NPR and the input NPR. As an example, the measured waves spectrum at the DUT reference planes can be visualized in Fig. 7 for an input power of 18.6 dBm.

For the impedance giving the PAE max, the Fig. 6a) shows the output power and IM3 power versus input power. It is verified that the slope of the intermodulation products is 3:1.

The Fig. 6b) is the NPR versus input power. The NPR for the PAE max point is plotted in blue.

Conclusion

This paper deals with an upgraded, on-wafer, NVNA set-up dedicated to linearity transistors measurement. The main difficulty is the calibration procedure which is more complex than connectorized configuration. Moreover, an original procedure allows a significant gain in measurement time by avoiding the systematic measurement of all the impedance points of the tuner for each measurement. This setup has been used to find the PAE max load impedance for an UMS Foundry transistor. For each measured point, by using a dedicated signal, it's possible to quickly measure the NPR. The extension of this bench is to search the load impedance of maximum NPR and to develop active load pull to reach reflection coefficients with higher modules than those obtained in Fig. 3 b).

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