Simultaneous AM-AM/AM-PM Distortion Measurements of Microwave Transistors Using Active Load-Pull and Six-Port Techniques

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***Abstract*—A programmable active load-pull measurement system using two six-port reflectometers and three passive two-port standards has been developed to obtain load-pull contours of the transistor's input-output phase shift variations over a wide dynamic range of the input power. The output power, gain, power-added efficiency, and phase shift are measured simultaneously at the transistor's input and output reference planes. The phase distortion versus input power, $\phi \sim P_{\text{in}}$, and the AM-PM conversion coefficient at various power levels, $k \sim P_{\text{in}}$, are obtained for different load impedances by post-measurement calculations. A NE8001 MESFET is tested at f = 1.7 GHz for the class A operation. The experimental results are also given.**

I. Introduction

Low phase distortion MESFET power amplifiers and limiters are key components in systems where the information is related to the phase of the microwave signal. For an accurate and quick design of linear solid state power amplifiers (SSPA's) and power limiters in HMIC and MMIC technologies, transistor characterization in terms of phase distortion measurements is highly recommended. The AM-PM distortion behavior can be described by the relative phase shift versus input power ($\phi \sim P_{\text{in}}$) and the AM-PM conversion coefficient k in degrees per dB. Various methods for measuring the AM-PM conversion coefficients at intermediate frequencies have been published [1], [2]. However, for microwave/millimeter wave amplifiers, it is less complex and tedious to obtain the characteristic $\phi \sim P_{\text{in}}$ by single-carrier power sweep measurements. Then, if needed, k can be found by derivation of ϕ with respect to P_{in} ($d\phi/dP_{\text{in}}$). Actually, the measured $\phi \sim P_{\text{in}}$ curve as a phase transfer function is more general to characterize the AM-PM distortion performance of microwave transistors and various power amplifiers. In addition, $\phi \sim P_{\text{in}}$ characterization is especially useful for microwave limiters, where the variation range of the phase shift over a given power range beyond saturation is the main concern [3].

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This paper presents an approach and the corresponding experimental setup to simultaneously investigate the AM-PM characteristics along with AM-AM characteristics as the functions of the power level and the load impedance over the Smith chart. For a designed performance in terms of output power, efficiency and AM-PM distortion, this comprehensive characterization can provide data for improving the design of limiting amplifier or SSPA magnifying signals having time variable envelopes. The proposed approach is based on the principles of a dual six-port network analyzer [4] and the active loading techniques [5]. In comparison with the setups employing heterodyne network analyzers, the advantages of this measurement setup are: 1) the impedance and power measurements are performed at the actual power level of the device—no extra attenuators are needed for power device characterization; 2) the AM-PM characterizations are carried out at the input and output reference planes of the device under test, loaded by arbitrary impedances; and 3) the cost of this system is much lower than the cost of the setups using two automatic network analyzers for AM-PM distortion load-pull measurements [6].

The effects of load impedance on the phase shift variations have been previously reported for a class AB power amplifier terminated by three different loads in [7]. To the best knowledge of the authors, the load-pull contours of the AM-PM conversion coefficients have been published only once in [6] for a fixed input power value. In this paper, the load-pull contours of the relative input-output phase shift at different power levels including the strong saturation region of the tested transistor are presented, using an active loading technique. Also, the phase distortion and AM-PM conversion coefficient versus input power are obtained for different load impedances. The effects of the loads on AM-PM conversion performance of the tested MESFET are discussed.

II. Measurement Approach and Setup

A. Principle

The measurement system is shown in Fig. 1. As is well known, two six-port junctions, associated with the amplitude and phase controllers, can be utilized for active load-pull techniques. Meanwhile, the measurement system can be viewed as a three-port network with reference planes 1, 2, and 3. In order to measure the input-output phase shift of the Transistor Under

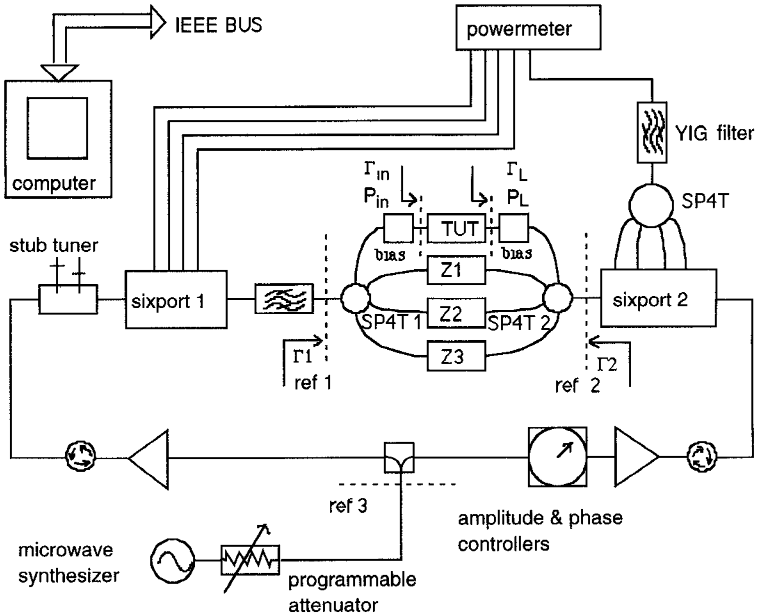


Fig. 1. Block diagram of the load-pull of phase distortion measurement system.

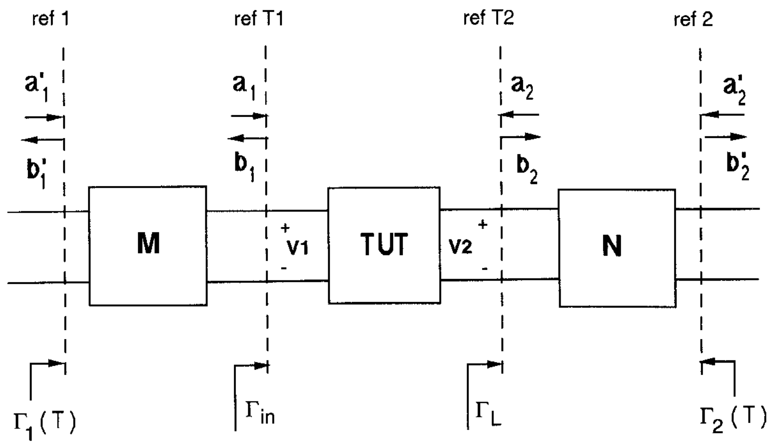


Fig. 2. Error-box model for de-embedding procedures and computations.

Test (TUT), three two-port passive standards Z1, Z2, and Z3 are introduced to obtain a calibration coefficient g, as shown below.

Due to the fact that the three-port network has fixed network scattering parameters during the four measurements of the TUT, Z1, Z2, and Z3 for a fixed setting of the amplitude and phase controllers, we can obtain the following equations expressed in matrix format:[1]

$$\begin{bmatrix} S_{12}(1) & \Delta(1) - \Gamma_1(1)S_{22}(1) & S_{12}(1)\Gamma_1(1) \\ S_{12}(2) & \Delta(2) - \Gamma_1(2)S_{22}(2) & S_{12}(2)\Gamma_1(2) \\ S_{12}(3) & \Delta(3) - \Gamma_1(3)S_{22}(3) & S_{12}(3)\Gamma_1(3) \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} = \begin{bmatrix} \Gamma_1(1) - S_{11}(1) \\ \Gamma_1(2) - S_{11}(2) \\ \Gamma_1(3) - S_{11}(3) \end{bmatrix} \tag{1}$$

where x_i is related to the scattering parameters of the three-port network for a given amplitude and phase setting; $\Delta(k) = S_{11}(k)S_{22}(k) - S_{12}(k)S_{21}(k)$ and $S_{ij}(k), k = 1,2,3$ are the known S parameters of three standards Z1, Z2, and Z3.

$\Gamma_1(k), k = 1,2,3$ are measured by six-port 1, when the test path is connected to Z1, Z2 or Z3, respectively.

The three two-port standards Z1, Z2, and Z3 are three coaxial transmission lines with different lengths. Note that we have purposely selected them to make the phase of $S_{12}(1), S_{12}(2)$, and $S_{12}(3)$ deviate from each other about $120°$, in order to ensure a good condition number of (1).

The following equation can also be derived [4] to calculate the coefficient g, i.e., the ratio of the incident waves at reference planes 2 and 1 (see Fig. 2):

$$g = \frac{a'_2}{a'_1} = \frac{x_1 + x_3\Gamma_1(T)}{1 + x_2\Gamma_2(T)} \tag{2}$$

where x_i is obtained by (1); $\Gamma_1(T)$ and $\Gamma_2(T)$ are measured by six-port 1 and 2, respectively, when the test path is connected to the TUT.

[1] If $\Delta - \Gamma_1 S_{22}$ is replaced by $\Gamma_2(S_{11} - \Gamma_1)$, (1) is the same as the one reported in [8]. We prefer (1) because only Γ_1 needs to be measured.

Then, the coefficient g has to be de-embedded to the input and output ports of the TUT (reference planes T_1 and T_2). As shown in Fig. 2, the TUT includes the MESFET NE8001. The network M (or N) consists of the single-pole four-throw switch SP4T1 (or SP4T2), the bias tee and the half part (input or output) of the test fixture. We obtain

$$\begin{aligned} b_1' &= S_{11}^m a_1' + S_{12}^m b_1 \\ a_1 &= S_{21}^m a_1' + S_{22}^m b_1 \end{aligned}$$

and

$$\begin{aligned} a_2 &= S_{11}^n b_2 + S_{12}^n a_2' \\ b_2' &= S_{21}^n b_2 + S_{22}^n a_2' \end{aligned}$$

where S_{ij}^m and S_{ij}^n are the S parameters of the networks M and N.

Also, we have $\Gamma_1(T) = b_1'/a_1'$ and $\Gamma_2(T) = b_2'/a_2'$. From the above equations, a_2/a_1 can be developed and related to coefficient g as follows:

$$\frac{a_2}{a_1} = g \cdot \frac{S_{12}^m}{S_{21}^n} \cdot \frac{\Gamma_2(T)S_{11}^n - \Delta^n}{\Gamma_1(T)S_{22}^m - \Delta^m} \tag{3}$$

where $\Delta^m = S_{11}^m S_{22}^m - S_{12}^m S_{21}^m$ and $\Delta^n = S_{11}^n S_{22}^n - S_{12}^n S_{21}^n$. Since the switches SP4T1 and SP4T2 are operated under linear conditions, the S parameters of networks M and N are constants when P_{in} sweeps, and they can be obtained using de-embedding method such as TRL [9].

According to the definition of the normalized waves a and b, the ratio of the voltages at the output and input ports of the TUT for any arbitrary load at a given P_{in} is determined as follows:

$$\frac{V_2}{V_1} = \frac{a_2 + b_2}{a_1 + b_1} = \frac{a_2}{a_1} \cdot \frac{1 + 1/\Gamma_L}{1 + \Gamma_{\text{in}}} \tag{4}$$

where a_2/a_1 is obtained by (3). Γ_{in} is the input reflection coefficient of the TUT and Γ_L is the load reflection coefficient presented to the TUT.

In brief, by means of switches, $\Gamma_1(1), \Gamma_1(2), \Gamma_1(3)$ and $\Gamma_1(T), \Gamma_2(T)$ can be measured when the Z1, Z2, Z3 and TUT are tested in turn for a fixed setting of the amplitude and phase controllers. Meanwhile, it's easy to obtain $\Gamma_{\text{in}}, P_{\text{in}}$ and Γ_L, P_L with de-embedding techniques when $\Gamma_1(T)$ and $\Gamma_2(T)$ are measured by six-port 1 and six-port 2 [5], [9]. Therefore, the phase shift of the TUT can be determined by (1)–(4). The harmonics are filtered, as shown in Fig. 1, to ensure that the signals detected by the powermeter are the fundamental components only.

The phase distortion, ϕ, is defined as an input-output phase shift, relative to the reference phase shift value at a given lower input power $P_{\text{in}}^{\text{ref}}$ (small-signal operation mode) for a given load impedance. Thus, the pertinent relative phase shift corresponds to the change of the angle of V_2/V_1 when P_{in} increases. The measurement of the AM-PM distortion behavior, $\phi \sim P_{\text{in}}$, followed by the calculation of the derivative $d\phi/dP_{\text{in}}$, provides the AM-PM conversion coefficient.

Similarly, the AM-AM distortion is described by the deviation of the gain from its value at a small-signal input power level. Due to the fact that P_L and P_{in} are already measured, the operating power gain $G_p \sim P_{\text{in}}$ can be deduced and the AM-AM conversion coefficient can be determined by the derivative dG_p/dP_{in}.

B. Measurement Procedure

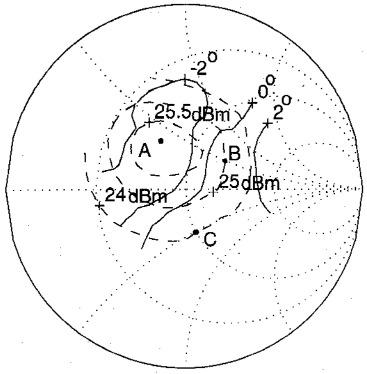
Traditionally, the load-pull has to be carried out for each given P_{in} with variable Γ_L. This results in a large number of measurements and adjustments. The present experimental procedure performs the active load-pull measurements by fixing the positions of the amplitude and phase controllers and sweeping the input power. In such condition, Γ_L changes, as P_{in} is swept. Then, the amplitude and phase controllers are adjusted at new positions and the measurements are performed with another sweep of P_{in}. Finally, the load-pull contours for a given P_{in} can be extracted from the recorded measurement database.

In addition, the characteristics $P_L \sim P_{\text{in}}$ and $\phi \sim P_{\text{in}}$ for a given Γ_L can be obtained as a post-measurement calculation result. The power-added efficiency PAE and G_p can also be deduced. It is known that in an active load-pull system, when P_{in} changes it is difficult to maintain the required Γ_L value by adjusting the amplitude and phase controllers. This problem can be circumvented by means of the above alternative procedure.

The error sources in the proposed approach are mainly due to the following four aspects. First, the measurements are made by increasing the input power and not by amplitude modulation as in the actual operation mode. Therefore, the dynamic effect of the AM modulation, which becomes more significant by increasing the modulation frequency, is ignored. Second, the self-heating effect can introduce a drift in the electrical operational conditions. To minimize this effect, the temperature of the fixture of the transistor investigated was maintained almost constant by circulating a flux of air during characterization. Third, the measurement accuracy can also be diminished if the power level exceeds the operational range where the switches are linear. It is preferable to use mechanical switches instead of solid-state switches when characterizing high-power devices. Fourth, the extraction of the pertinent data from the raw database and the post-measurement calculations using interpolation routines might introduce error. This effect can be reduced by increasing the amount of experimental data.

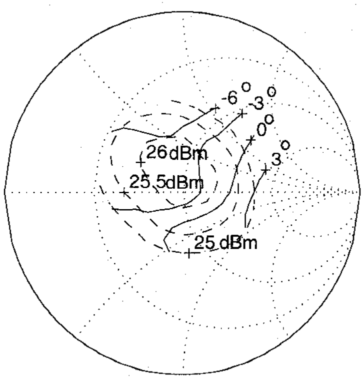
III. Experiment Results

A NE8001 MESFET was measured at $f = 1.7$ GHz under a class A bias condition. Fig. 3(a)–(c) shows the load-pull contours of the relative input-output phase shift, ϕ, superposed with the output power, P_L, corresponding to $P_{\text{in}} = 10$, 15, and 19 dBm. The reference input power level was -2 dBm. The contours were zoomed around the load impedance region of maximum output power and the saturation of the output power can be observed. It can also be seen that the phase distortion increases sharply when P_{in} goes to 19 dBm. In addition, the variations of the phase shift for different Γ_L in a given range of P_{in} are not the same. Examining these figures, we expect



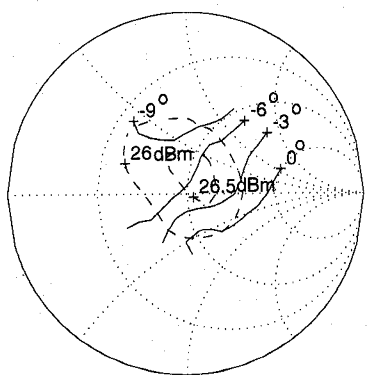
P(f0)/f0

(a)



P(f0)/f0

(b)



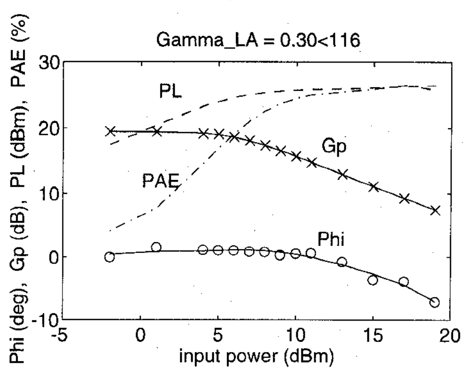
P(f0)/f0

(c)

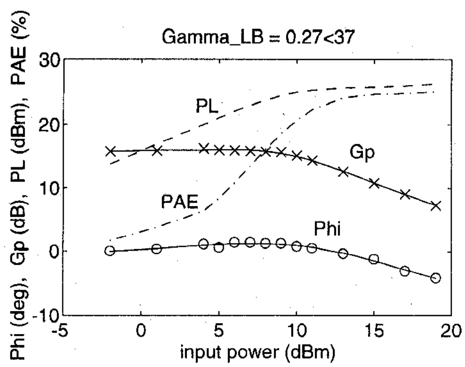
Fig. 3. Load-pull contours of relative input-output phase shift ϕ (in degrees) and output power P_L (in dBm) at different P_{in}: (a) $P_{\text{in}} = 10$ dBm, (b) $P_{\text{in}} = 15$ dBm, (c) $P_{\text{in}} = 19$ dBm.

that the load impedances in the right part of the Smith chart have better phase distortion performance than the left part.

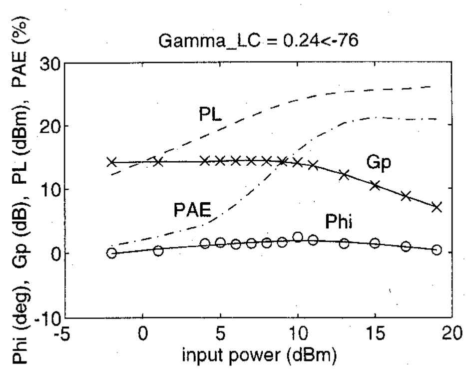
The relative phase shift versus input power, $\phi \sim P_{\text{in}}$ for each Γ_L, can be extracted by interpolation from the above measured database. For example, three loads Γ_{LA}, Γ_{LB} and Γ_{LC} were chosen, corresponding to the $P_{L\max} = 2\ 5.8$ dBm, $P_L = 25$ dBm, and $P_L = 24$ dBm for $P_{\text{in}} = 10$ dBm [see Fig. 3(a)]. For these different loads, Fig. 4(a)–(c) shows the characteristics of $P_L \sim P_{\text{in}}, Gp \sim P_{\text{in}}, \text{PAE} \sim P_{\text{in}}$ and ϕ (i.e., *Phi*) $\sim P_{\text{in}}$. By differentiation, the AM-PM conversion coefficient $k = d\phi/dP_{\text{in}}$ was obtained and shown in Fig. 4(d).



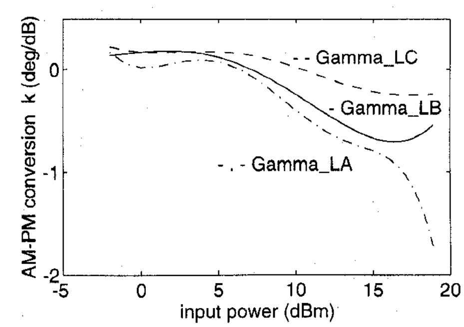
(a)



(b)



(c)



(d)

Fig. 4. (a) P_L (output power), (b) G_p (operating power gain), PAE (power-added efficiency) and (c) *Phi* (i.e., ϕ, relative input-output phase shift) versus P_{in}. (d) AM-PM conversion coefficient k versus P_{in}.

The experimental results demonstrate that the load impedance has an effect on the phase shift. It may be different at various input power levels. When P_{in} exceeds 10 dBm, the transistor presents different behaviors of the AM-PM conversion characteristic (about 1°–2°/dB deviation) as the load changes. It is found that Γ_{LA} has higher output power and efficiency, but its AM-PM distortion is worse than the others. The selection of Γ_{LC} offers an improvement in AM-PM distortion. However, the power and efficiency are low. Γ_{LB} is the best tradeoff for the power, power-added efficiency, and phase distortion among these loads.

IV. Conclusion

The proposed approach for measuring AM-PM distortion along with AM-AM distortion, using active load-pull and six-port techniques, is a new attempt. The phase shift variations of a MESFET can be measured at variable load impedances in the Smith chart and over a wide dynamic range of power, including the strong saturation region.

The power and phase load-pull contours and the P_L (or Gp, PAE) $\sim P_{\text{in}}, \phi \sim P_{\text{in}}$ characteristics can be obtained by automatic sweeping the input power for variable active loading, followed by post-measurement calculations.

The measurement system is especially useful for large-signal phase distortion characterizations of microwave transistors. It also introduces an experimental method to provide data for improving the design of low phase distortion power amplifiers and limiters.

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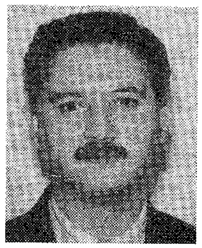

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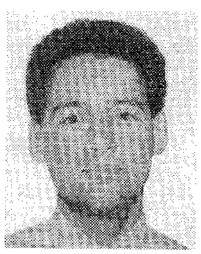

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