Automated Characterization of HF Power Transistors by Source–Pull and Multiharmonic Load–Pull Measurements Based on Six-Port Techniques

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***Abstract*—An original measurement system for nonlinear microwave power-transistor characterization using six-port reflectometers is presented. It allows independent active tuning of the output impedances at f_0 and $2f_0$ (multiharmonic load–pull) and variation of the source impedance at the input port at f_0 (source–pull). An appropriate search algorithm enables automatic optimization of the output impedances and leads to fast user-friendly operation of the system. Experimental results are shown for a commercial GaAs MESFET power transistor at $f_0 = 2$ GHz.**

***Index Terms*—Load–pull, measurement automation, multiharmonic, nonlinear, six-port, source–pull.**

I. Introduction

EXPERIMENTAL characterization of microwave power transistors is essential for optimization of output performance and verification of models for nonlinear simulation.

The dependency of the device performance on the output impedance is usually examined by applying a fixed input power, varying the output load, and measuring quantities like output power $P_{\rm out}$ and power-added efficiency (PAE). This approach is called "load–pull." Variation of the output load can be passively carried out by means of a tuner or actively carried out by injecting a power wave toward the transistor output and, thus, forcing the desired wave ratio. The load variation can take place either at the fundamental frequency f_0 alone or at the fundamental frequency and a number of its harmonics (multiharmonic load–pull). The importance of the harmonic load impedance (especially at the second harmonic $2f_0$; for some operational classes (e.g., class-F), at the third harmonic $3f_0$ too), in particular on PAE, is known [15].

Passive tuning structures can be carried out by variable mechanical elements or electronic circuits (*solid-state tuner*). Tuners are commercially available, can handle high powers, are easy to use, and are of comparatively low cost. Despite these advantages, they are of limited use because of inherent losses, which impose limitations in reflection-coefficient magnitude. The maximum magnitude decreases with frequency and with the number of elements and length of cables connected between the measurement plane and tuner, which is especially disadvantageous for on-wafer measurements and measurements of highly mismatched devices. The active load–pull principle does not suffer from limitations in load reflection-coefficient magnitude. By simulating a reflection coefficient with the help of a power wave injected toward the device output, unity magnitude can be achieved for any physical setup and at any frequency. Phase and magnitude of the injected wave are controlled by a variable attenuator and a variable phase shifter or by an IQ modulator, followed by a power amplifier. Two basic principles of load variation are commonly used: either the signal injected to the device output is synchronized with the one applied to the input ("synchronous sources," introduced by Takayama [18]) or the signal generated by the device itself is fed back to its output with variable phase and magnitude via an "active-load loop" [1]. While the first solution shows no risks of oscillation, the second solution provides a constant reflection coefficient independent from the device's output signal.

Deshours *et al.* [5] studied some active and passive load–pull systems by comparative measurements of a power transistor.

Just like the load impedance, the impedance of the source providing the input signal for the power transistor also plays a role on the transistor behavior. The source impedance presented at the device input can influence output characteristics like linearity or optimum load impedance. This is true for devices with a nonlinear input part (like bipolar transistors) and for mixing devices. The source impedance also affects the noise figure. Attempts have been made to realize a variable source impedance using the active-load loop technique [2] or a mechanical tuner [11], [14]. However, no solution has been proposed for simultaneous measurement of both the input reflection coefficient of the device-under-test (DUT) and the reflection coefficient of the synthesized source with variable impedance.

Unless precalibrated tuner systems equipped with power meters are used, a device for measuring power-wave ratios and absolute powers is needed. This can be an automatic network analyzer (ANA) of heterodyne type or a double six-port network analyzer (DSPNA). Appropriate calibration

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methods assure vector corrected measurements at the device's input and output ports. The use of an ANA for multiharmonic measurements requires a frequency converter either in form of an additional device [9] or built-in [4], [17] in order to enable the ANA to measure at harmonic frequencies. In case of a DSPNA which does not need a reference signal, frequency selection can be carried out by switched filters [12] or tunable filters.

Computer-controlled operation and automation of the measurement tasks are key elements of any measurement system, as the number of measurements necessary for a complete characterization of the DUT is very high (typically hundreds of settings and readings). In addition, intelligent search algorithms should be used for load optimization in order to improve characterization speed and avoid the measurement of useless points or impedances that lead to device destruction.

In this paper, we present an original large-signal measurement system for high-frequency (HF) power transistors. It features load variation at the fundamental frequency and the second harmonic (f_0 and $2f_0$) by two completely independent branches at the DUT's output, each of which is equipped with a six-port reflectometer. Furthermore, the input branch, equipped with a third six-port reflectometer, contains an active-load loop that allows source impedance variation (source–pull). By means of a new structure, this reflectometer can measure both the input reflection coefficient of the DUT and the source reflection coefficient.

The principle of six-port reflectometer operation and its calibration are explained in Section II. The load–pull capabilities of our system are described in Section III-A, the source–pull principle in Section III-B. In Section IV, we discuss the high-level automation software developed for this system. Finally, Section V reports experimental results.

II. THE SIX-PORT REFLECTOMETER AND ITS CALIBRATION

Six-port operation is based on scalar power measurements at four of the six ports of a passive interferometric junction [6]. The measured powers P_i (i: detection port 3–6) depend on linear combinations of the incident and emerging waves in the measurement plane.

The calibration of a six-port reflectometer has been subject of many papers and will not be discussed here in detail. It is usually divided into two parts: first, the so-called six- to four-port reduction or $P \rightarrow w$ transform is carried out to derive a complex wave ratio w from the four scalar power readings $P_i (i = 3, \cdots, 6)$ [7]. Second, a bilinear transform $w = (A\rho + B)/(C\rho + 1)$ corrects for the actual measurement plane. The complex A, B, C are the so-called errorbox. In our case, two reflectometers form a DSPNA, and the two errorboxes are found by the well-known thru-reflection line (TRL) method [8].

In our setup, six-port 1 and six-port 2 form a DSPNA at f_0, while six-port 1 and six-port 3 form a DSPNA at $2f_0$. For this reason, six-port 1 must be broad-band and cover both f_0 and $2f_0$.

For absolute power measurements, a constant k_P is needed as a factor of proportionality between the reference power

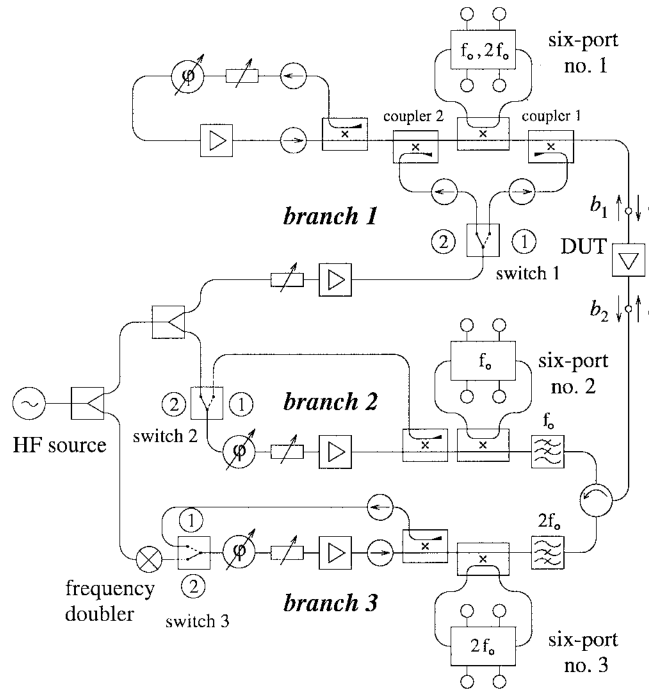


Fig. 1. Simplified structure of the measurement system.

detector readout P_3 and the actual incident power in the measurement plane [13].

For further use in the load–pull setup, it is important to mention that the calibration constants for the $P \rightarrow w$, the $w \leftrightarrow \rho$ transform, and absolute power measurements of each single reflectometer do not depend on the inner impedance of the driving signal source.

Our special calibration procedure that enables the system to carry out vector corrected measurements of wave ratios and absolute power values in an on-wafer or in-fixture environment has been presented in [3].

Fig. 1 shows the structure of the measurement system. Some filters and isolators, necessary at the outputs of the power amplifiers, are left out for better comprehension and clarity.

A. Principle of Load–Pull Measurements

Two branches (branches 2 and 3) are connected to the output of the DUT via a filter/circulator network providing frequency separation. The fundamental component of the output signal is treated in branch 2, harmonics in branch 3. Both of the output branches are equipped with an attenuator, phase shifter, and power amplifier, and are thus able to inject a power wave to the transistor output at f_0 (branch 2) and $2f_0$ (branch 3) and simulate variable reflection coefficients Γ_L (f_0) and Γ_L $(2f_0)$. Higher harmonics are terminated in 50 Ω by appropriate filters and isolators. Switches 2 and 3 offer the choice of the "active-load loop" mode (position "1") and the "synchronous sources" mode (position "2"). For the latter case, a frequency doubler acts as a $2f_0$ signal source for branch 3.

By means of the two six-port reflectometers in the two branches, this configuration allows completely independent setting and measuring the fundamental and second harmonic

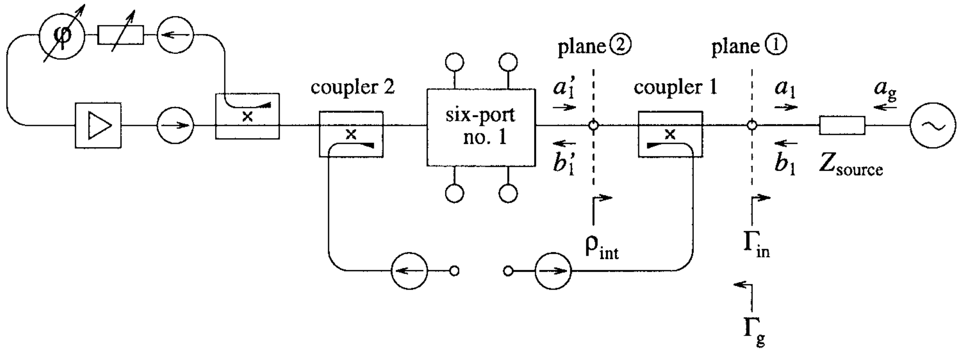


Fig. 2. Fictitious setup for explanation of source impedance measurement.

load and respective absolute powers. No variable filter element is necessary, so that the system does not suffer from repeatability problems (switched filters) or low measurement speed (tunable yttrium–iron–garnet (YIG) filters). As the relative bandwidth $f_{\max}/f_{\min}$ for each six-port is limited to 1.5 (due to the circulator/filter network with fixed cutoff frequencies), very low-cost reflectometers in microstrip technology can be used.

B. Principle of Source–Pull Measurements

In normal operation (i.e., a DUT is measured), switch 1 is in position 2 and the signal generated by the source is fed to the device's input via coupler 2 (see Fig. 1). The active-load loop of branch 1 is used to vary the reflection coefficient if one looks into port 1. This whole configuration represents a signal source with variable impedance [1]. As the six-port reflectometer calibration constants are valid for any source impedance, six-port 1 correctly measures the wave ratio b_1/a_1, which is the input reflection coefficient of the DUT, and the absolute incident power, for any position of the active-load loop

$$\rho_1 = \frac{b_1}{a_1} = \Gamma_{\text{in}}. \tag{1}$$

In order to explain the measure of the reflection coefficient of the synthesized source, let us study the following fictitious case (see Fig. 2).

Neither via coupler 1 nor coupler 2 is a signal fed to the measurement port. At the measurement port (plane 1), a signal source is connected instead of a DUT. In this case, the wave ratio b_1/a_1 is determined by the entire structure at the left of plane 1 and equals the reciprocal of the reflection coefficient Γ_{g} of this structure.

As the six-port constants are still valid (via the transforms $P \rightarrow w$ and $w \leftrightarrow \rho$, they describe the linear relationship between the waves a_1, b_1 in the reference plane and the waves measured by the power detectors), this wave ratio is well measured by the six-port 1

$$\rho_1 = \frac{b_1}{a_1} = \frac{1}{\Gamma_{\text{g}}}. \tag{2}$$

Notice that this wave ratio is independent from the source impedance Z_{source}.

Just like the wave ratio $\rho_1 = b_1/a_1$, the wave ratio $\rho_{\text{int}} = b_1'/a_1'$ in plane 2 (introduced for explanation purposes) is also linked to the intermediate variable w by a bilinear transform (see Section II). Its parameters need not be known.

Let us now consider that the structure of Fig. 2 is unchanged and, in particular, the parameters of the active-load loop remain the same. The signal is now fed into the structure via coupler 1 (switch 1 in position ①) instead of a signal source at the measurement port. In this case, the wave ratio ρ_{int} is unchanged too. Thus, the value of the intermediate variable w must be the same, too, and the result of the $w \leftrightarrow \rho$ transform giving ρ_1 is also the same. The six-port reflectometer "sees" the same ratio. That means that six-port reflectometer 1 is able to measure the reflection coefficient presented by the whole structure to the input of the DUT if the excitation signal is fed via coupler 1 instead of coupler 2.

Thus, depending on the position of switch 1, six-port 1 measures either the input reflection coefficient of the DUT or the reflection coefficient presented to its input by the synthesized source with variable impedance. For both cases, the calibration constants found by the classical calibration method are valid.

In total, the system is able to measure the following quantities:

- input reflection coefficient of the DUT at f_0
$$\Gamma_{\text{in}}(f_0) = b_1(f_0)/a_1(f_0)$$
- source reflection coefficient at f_0 $\Gamma_{\text{g}}(f_0)$
- load reflection coefficient at f_0 and $2f_0$
$$\Gamma_{\text{L}}(f_0) = a_2(f_0)/b_2(f_0)$$
$$\Gamma_{\text{L}}(2f_0) = a_2(2f_0)/b_2(2f_0)$$
- input power at f_0
$$P_{\text{in}} = 1/2|a_1|^2(1 - |\Gamma_{\text{in}}|^2)$$
- output power at f_0 and $2f_0$
$$P_{\text{out}}(f_0) = 1/2|b_2(f_0)|^2(1 - |\Gamma_{\text{L}}(f_0)|^2)$$
$$P_{\text{out}}(2f_0) = 1/2|b_2(2f_0)|^2(1 - |\Gamma_{\text{L}}(2f_0)|^2)$$
- power gain at f_0
$$G = P_{\text{out}}/P_{\text{in}}$$

- PAE at f_0 (if $P_{\rm dc}$, dissipated dc power, is known)

$$\text{PAE} = (P_{\rm out} - P_{\rm in})/P_{\rm dc}.$$

III. Automation Software

In view of the large amount of measurement points needed for a comprehensive device characterization, automation of the measurement process is imperative. Basic software routines concern attenuator, phase shifter and switch setting, detector readout, and input power stabilization. In addition to these low-level operations, algorithms for load optimization and load–pull contour tracking are very advantageous compared to statistic approaches carrying out a search over the whole Smith chart. They reduce the number of points to be measured and help limit the risk of device destruction by highly mismatched loads.

In the past, several approaches have been undertaken to automate the load search for optimum transistor performance [10], [16]. In the case of a tuner, as well as in the case of active-load simulation, two control signals usually determine the load reflection coefficient presented to the device's output. Unfortunately, the dependence of the simulated reflection coefficient on the control signals is not precisely known. In the case of our system, the load reflection coefficient is determined by the positions of the variable attenuator and phase shifter. The phase shifter exhibiting a variable attenuation when changing the phase value, the attenuator having a variable phase when changing attenuation, and the other elements being potentially slightly mismatched, the positions of the attenuator and the phase shifter allow only rough estimates of the simulated reflection coefficient; thus, a precise prediction is impossible.

For our system, varying the phase-shifter position for a given attenuation corresponds to load reflection coefficients that lie, for ideal hardware, on a circle around the Smith chart origin. The radii of these circles are determined by the attenuation (decreasing attenuation means increasing radius). For nonideal hardware, the shape of the resulting plots can differ significantly from circles, and the centers do not coincide with the Smith chart origin and may vary with the radius. An automatic algorithm is needed that is insensitive to those hardware imperfections.

Our algorithm (written in QuickBASIC and FORTRAN) proceeds as follows: starting with a high attenuation (small circle), it optimizes the phase position for this attenuation using a method of decreasing intervals. When the maximum is found, the attenuation is decreased, leading to a higher circle radius, and phase optimization is started again. Attenuation is decreased until the absolute maximum is found. This method assures the determination of the optimum load by following the gradient of the load–pull contours. It avoids critical loads for the transistor by starting near the Smith chart origin (matched load) and orientating from the beginning toward the optimum load. Fig. 3 illustrates this procedure.

If desired, a number of points on both sides of the gradient path found by the optimum search is measured and treated mathematically (MATLAB) in order to obtain the load–pull contours.

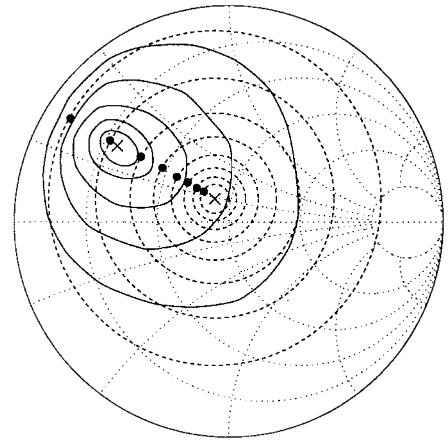


Fig. 3. Schematic representation of the load optimization method. Solid lines (—): constant output power or constant efficiency contours, dashed lines (- - -): reflection coefficients for a given attenuation and variable phase shifter position, •: optimum phase position for each attenuation value.

The optimization criterion for output load optimization can be maximum output power $P_{\rm out}$ max or maximum efficiency $\text{PAE}_{\rm max}$. We use this algorithm for synthesizing a certain user-specified reflection coefficient too. In this case, the optimization criterion is the distance between the current and desired reflection coefficient.

IV. Practical Results

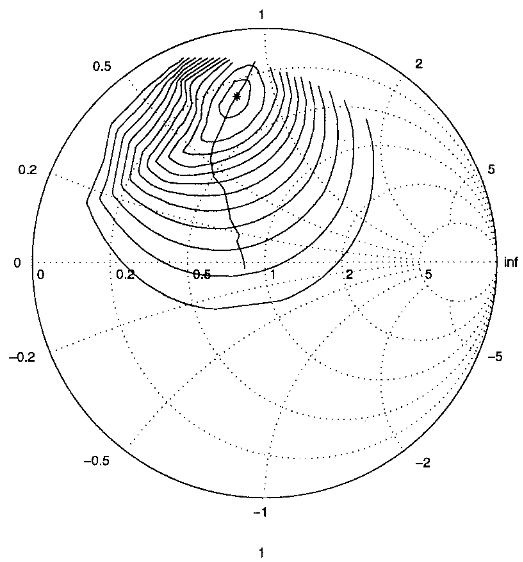
In order to illustrate the capabilities of our system, we present experimental results obtained for a commercial power MESFET (Fujitsu FLL101ME) at $f_0 = 2$ GHz, mounted into a test fixture (Inter-Continental Microwave) and biased in class AB ($V_{\rm DS} = 10$ V, $V_{\rm GS} = -1.53$ V, $I_{\rm DS0} = 35$ mA).

Fig. 4(a) and (b) shows the optimum load search for maximum $\text{PAE}_{\rm max}$ for the loads at the fundamental frequency and second harmonic.

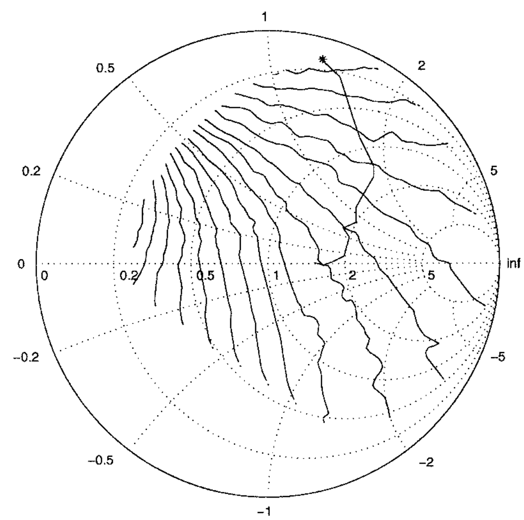
Fig. 5 shows the output power at f_0 $P_{\rm out}$ (f_0), at $2f_0$ $P_{\rm out}$ ($2f_0$) and the PAE for three cases: the values plotted in dotted lines ("$\cdots$") are measured for matched loads at f_0 and $2f_0$, dashed lines ("- - -") correspond to the optimum load at f_0 and a matched load at $2f_0$, and the solid lines ("—") are taken for the optimum loads at both f_0 and $2f_0$.

The previous measurements have been taken for a matched source ($\Gamma_{\rm g} = 0$). Table I summarizes the numerical results obtained for this case and shows, in addition, results obtained for a mismatched source (the source reflection coefficient could not be set to the conjugate of the input reflection coefficient, as this would lead to instability of the transistor; thus, a reflection coefficient with the same phase, but lower magnitude, was chosen). All values are measured at 10-dBm input power.

Optimization of the fundamental load has a large effect on transistor performance. It increases PAE from 22.6% with a matched load to 71.1% with the optimum load. The measurements show that harmonic load optimization has an important effect also, as it gives further improvement of PAE by 5% to 76.1%. At the same time, output power is increased from 23.1 to 26.9 dBm (optimum load only at f_0) and 27.3 dBm (optimum loads at f_0 and $2f_0$). It is interesting to notice that

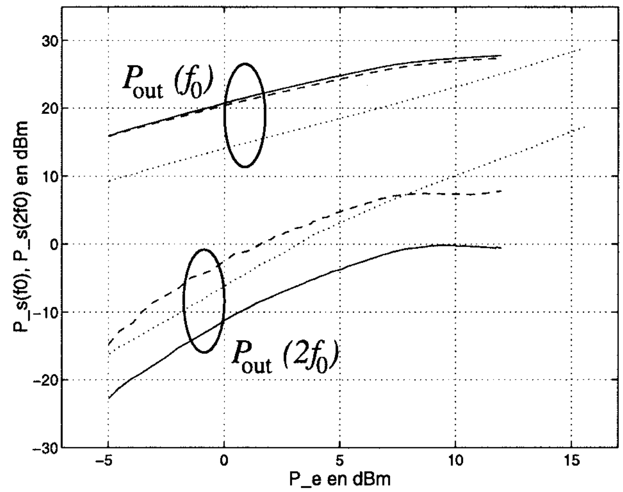


(a)

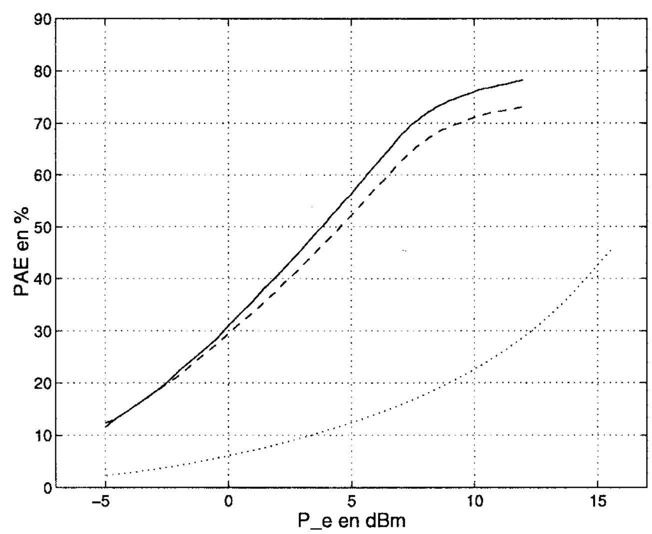


(b)

Fig. 4. Optimum load search at (a) f_0 and (b) $2f_0$. Both (a) and (b) show the search paths, constant efficiency contours (both in solid lines) and the optimum load (marked with $*$). The step between the contours for constant efficiency is 5% for f_0 and 0.5% for $2f_0$.



(a)



(b)

Fig. 5. (a) Output power at f_0 and $2f_0$ and (b) PAE as a function of the input power. $\Gamma_g = 0$.

TABLE I
OPTIMUM LOADS FOR PAE FOR DIFFERENT SOURCE REFLECTION COEFFICIENTS. INPUT POWER 10 dBm

source refl. coeff.	load reflection coeff. $\Gamma_L(f_0)$	load reflection coeff. $\Gamma_L(2f_0)$	P_{out} (dBm)	PAE (%)
0.00/–	0.00/–	0.00/–	23.1	22.6
	0.68/105.5°	0.00/–	26.9	71.1
	0.68/105.5°	0.93/51.7°	27.3	76.1
0.55/153.0°	0.67/104.9°	0.98/36.9°	27.3	75.8

the optimum load at $2f_0$ leads to an output power at $2f_0$ that is about 10 dB lower than for a matched load at $2f_0$. The shapes of the P_{out} (f_0) and PAE plots show that the transistor is deeper in compression than with a 50-Ω load.

Changing the source reflection coefficient has, for this transistor and these bias conditions, no effect on the optimum load at f_0, but it changes the optimum load at $2f_0$ ($|\Delta\Gamma| = 0.25$). The output performance remains virtually unchanged.

V. CONCLUSIONS

This paper deals with a new active large-signal transistor measurement system. Original ideas concerning the structure and automation have been presented. Two output branches allow independent load control at the fundamental frequency and second harmonic. Thanks to one six-port reflectometer in each output branch and the possibility to use microstrip technology, variable filter elements with the associated problems are unnecessary and the whole system is kept simple and low cost. At the transistor input, a new structure allows us to simulate a variable source impedance and to simultaneously measure the input reflection coefficient and the synthesized source impedance. The proposed automatic load-optimization algorithm makes the system user friendly, speeds up the characterization process, and helps prevent the transistor being operated under inappropriate load conditions. Experimental results for a commercial MESFET power transistor have illustrated the system performances.

Future work will focus on pulsed RF measurements and on an extension of the input structure to multiharmonic source–pull measurement capabilities.

REFERENCES

[1] G. P. Bava, U. Pisani, and V. Pozzolo, "Active load technique for load–pull characterization at microwave frequencies," *Electron. Lett.*, vol. 18, no. 4, pp. 178–180, Feb. 1982.

[2] ——, " 'Source–pull' technique at microwave frequencies," *Electron. Lett.*, vol. 20, p. 4, Feb. 1984.

[3] G. Berghoff, E. Bergeault, B. Huyart, and L. Jallet, "On-wafer calibration of a double six-port reflectometer including constants for absolute power measurements," *IEEE Trans. Instrum. Meas.*, vol. 46, pp. 1111–1114, Oct. 1997.

[4] F. Blache, J. N. Nebus, P. Bouysse, and J. P. Villotte, "A novel computerized multiharmonic active load–pull system for the optimization of high efficiency operating classes in power transistors," in *IEEE MTT-S Symp. Dig.*, Orlando, FL, May 14–19, 1995, pp. 1037–1040.

[5] F. Deshours, E. Bergeault, F. Blache, J. P. Villotte, and B. Villeforceix, "Experimental comparison of 'load–pull' measurement systems for nonlinear power transistor characterization," *IEEE Trans. Instrum. Meas.* to be published.

[6] G. F. Engen, "The six-port reflectometer: An alternative network analyzer," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, pp. 1075–1080, Dec. 1977.

[7] ——, "Calibrating the six-port reflectometer by means of sliding terminations," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 951–957, Dec. 1978.

[8] G. F. Engen and C. A. Hoer, "Thru-reflect-line: An improved technique for calibrating the dual six-port automatic network analyzer," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 987–993, Dec. 1979.

[9] A. Ferrero and U. Pisani, "A generalized harmonic load–pull system," in *3rd Asia–Pacific Microwave Conf. Proc.*, 1990, pp. 449–452.

[10] ——, "Novel hardware and software solutions for a complete linear and nonlinear microwave device characterization," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 299–305, Mar. 1994.

[11] F. M. Ghannouchi and R. G. Bosisio, "Source–pull/load–pull oscillator measurements at microwave/mm-wave frequencies," *IEEE Trans. Instrum. Meas.*, vol. 41, pp. 32–35, Feb. 1992.

[12] F. M. Ghannouchi, R. Larose, and R. Bosisio, "A new multiharmonic loading method for large-signal microwave and millimeter-wave transistor characterization," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 986–992, June 1991.

[13] T. E. Hodgetts and G. J. Griffin, "A unified treatment of the six-port reflectometer calibration using the minimum of standards," Royal Signals and Radar Establishment, Tech. Rep. 83003, 1983.

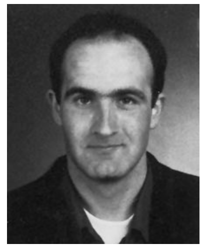
[14] D.-L. Lê and F. M. Ghannouchi, "Source–pull measurements using reverse six-port reflectometers with application to MESFET mixer design," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1589–1595, Sept. 1994.

[15] S. R. Mazumder, A. Azizi, and F. E. Gardiol, "Improvement of a class C transistor power amplifier by second-harmonic tuning," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 430–433, May 1979.

[16] S. M. Perlow, "New algorithms for the automated microwave tuner test system," *RCA Rev.*, vol. 46, pp. 341–355, Sept. 1985.

[17] B. Roth, D. Köther, M. Coady, and Th. Sporkmann, "A combined on-wafer measurement stand for linear and nonlinear microwave measurements," in *24th European Microwave Conf. Proc.* vol. 1, Sept. 1994, pp. 962–967.

[18] Y. Takayama, "A new load–pull characterization method for microwave power transistors," in *IEEE MTT-S Symp. Dig.*, June 1976, pp. 218–220.



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