Computer-Aided Error Correction of Large-Signal Load-Pull Measurements

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***Abstract*—A versatile system is described for the large-signal characterization of microwave power MESFET's. High accuracy is obtained through vector error-correction techniques. The system is calibrated using a procedure based on conventional automatic network analyzer calibration measurements and a series of simple insertion loss measurements. The measurement system provides accurate reflection coefficient and RF power data over a wide range of device loading conditions.**

I. Introduction

An IMPORTANT STEP in the design of microwave power amplifiers and oscillators is the accurate large-signal characterization of the active devices. The load-pull method [1]–[6] has gained wide acceptance as a useful large-signal characterization technique. The basic method in load-pull measurements is to terminate the device under test (DUT) with adjustable impedances. Parameters such as power gain, output power, and dc to RF power conversion efficiency are then measured as a function of input power level, device terminations, and dc bias conditions. The chief advantages of this approach are that the device is subjected to realistic operating conditions during the measurement [7], and that the measured data can be used readily in design calculations [8].

The accuracy of the load-pull method is limited primarily by the accuracy with which the terminations and RF power levels can be determined. The termination impedances (or reflection coefficients) are normally obtained using manual network analyzers. This can seriously limit the measurement accuracy due to the well-known directivity, mismatch, and cross-coupling errors associated with the network analyzer components. Errors in the measurement of RF input power and RF output power depend on the hardware configuration, but can be large. For example, if the output power meter is separated from the DUT by the output tuner [1], [3], unknown tuner losses will introduce uncertainty into the measurement. These losses can be as large as several decibels [9]. In systems where the RF power levels are determined using directional couplers at the input and/or output of the DUT [2], [5], [6], finite coupler directivity and connector mismatches can result in significant errors in the measured RF power. These errors are particularly troublesome if the VSWR in the directional coupler is high. With a 25-dB directivity, errors in the measured power can be as large as ± 1 dB.

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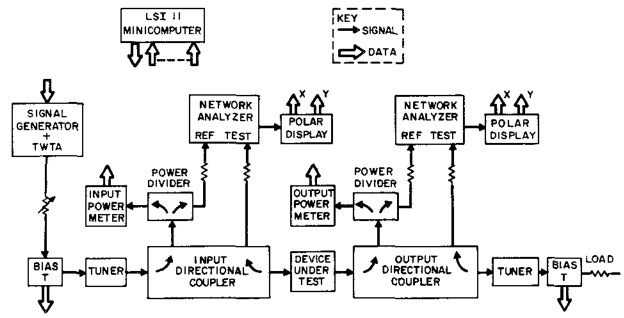

Fig. 1. Block schematic of the load-pull system.

This paper describes a computer-aided load-pull measurement system which achieves a significant improvement in measurement accuracy. Reflection coefficients and RF power levels are vector error-corrected at both ports of the DUT. The system is calibrated in a straightforward manner, using a series of conventional small-signal network analyzer calibration measurements and a series of simple insertion loss measurements. The system is controlled by a small computer, which gives a virtually instantaneous display of corrected RF power levels and reflection coefficients. Load-contour maps [1] are generated automatically by the computer, using an interpolation procedure.

II. Hardware

A block schematic of the computer-aided load-pull system is shown in Fig. 1. The DUT is fed from a high-power microwave source, and the source and load impedances are controlled by adjustable tuners at the input and output. (If the DUT is an oscillator, only the output part of the system is needed.) Uncorrected large-signal reflection coefficients at the input and output of the DUT are monitored using dual directional couplers and RF network analyzers. Attenuators ensure that the signal levels at the inputs of the harmonic frequency converters are within the safe operating range. If only one network analyzer is available, coaxial switches can be used to connect it back and forth between the input and output circuits.

At each port of the DUT, part of the signal for the network analyzer reference channel is coupled into a power meter using a resistive power divider. Since the reflection coefficients are known, only one power meter is required at

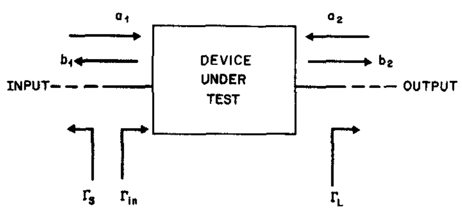

Fig. 2. Power waves and reflection coefficients at the DUT reference planes.

each port [10]. The power meters monitor the (uncorrected) power incident on the input port of the DUT and the (uncorrected) power generated at the output of the DUT. In an earlier computer-corrected load-pull system described by the authors [10], the output power meter was connected to the other coupled port of the output directional coupler. Under certain conditions this led to reduced accuracy, due to very small power meter readings when the load reflection coefficient was close to zero. The present system avoids this problem.

A small computer (LSI-11) is interfaced to the various items of hardware as shown. It performs the tasks of instrument control, data acquisition, error correction, and storage. For a fixed input signal frequency, the computer outputs a continuously updated display of error-corrected quantities, such as reflection coefficients, RF input and output power, and power-added efficiency. This enables the operator to manually adjust the dc bias, RF drive, and load conditions in order to optimize a particular performance objective.

The input and output tuners are both passive devices (Maury Microwave 2640D). However, a variety of other tuning schemes can be used, including two-signal methods [2] or active tuning [5]. Cusack *et al.* [1] used a computer-controlled servo-driven output tuner to achieve automatic load contour mapping. In the present system, the tuners are manually controlled. A computer routine generates load contours by interpolating measured data obtained at a number of points on the load reflection coefficient plane. About 50 data points are usually required. They should be spread in an approximately uniform manner across the region of interest. However, their exact placement on the load reflection coefficient plane is not critical. Sufficient data for a complete set of contours at one frequency can usually be obtained within ten minutes.

III. Error Correction

The objective of the load-pull system is to provide error-corrected values for the reflection coefficients and RF power levels at the reference planes of the DUT. Fig. 2 defines the traveling power waves at these planes and shows the reflection coefficients of interest. These are the load reflection coefficient $\Gamma_L = a_2/b_2$ and the large-signal input reflection coefficient $\Gamma_{in} = b_1/a_1$. Also shown in Fig. 2 is the source reflection coefficient $\Gamma_S = a_1/b_1$ which is measured with the RF source in Fig. 1 switched off, and a test signal injected into the input coupler from the right. The present system is intended primarily for load-pull characterizations and does not provide a direct output of Γ_S. However, the system can be readily modified for source-pull measurements. For normal load-pull measurements, the input power to the DUT is $P_{in} = |a_1|^2 - |b_1|^2 = |a_1|^2(1-|\Gamma_{in}|^2)$, and the output power is $P_{out} = |b_2|^2 - |a_2|^2 = |b_2|^2(1-|\Gamma_L|^2)$. The network analyzers and power meters in Fig. 1 give direct (but uncorrected) measurements of Γ_{in}, Γ_L, $|a_1|^2$, and $|b_2|^2$.

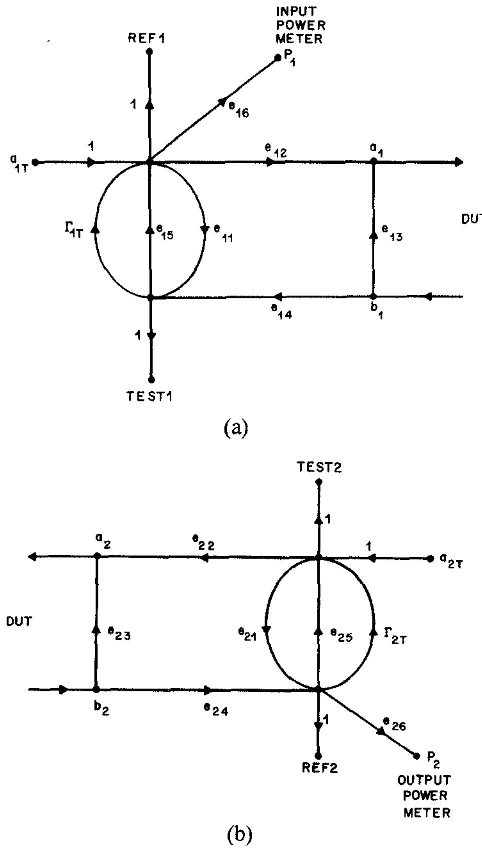

Fig. 3. Signal flowgraph error model of the directional coupler, power divider, power meter, and network analyzer. (a) Input port. (b) Output port.

A. Flowgraph Model

The error-correction techniques are based on a signal flowgraph model of the complete system. A detailed signal flowgraph model has been obtained for the directional coupler, power divider, power meter, and network analyzer at each port. The model includes all major sources of error, such as the finite coupler directivity, connector mismatches, and cross-coupling between the reference and test channels.

The detailed error model has been reduced using signal flowgraph reduction techniques [11]. The only assumption required in this reduction is that the reflection coefficients of the power meter sensor heads and the reference and test channel inputs of the harmonic frequency converters are constant and independent of RF power level. The reduced flowgraph models of the input and output networks are shown in Fig. 3(a) and (b), respectively. These models form the basis of a unified approach to the vector error correction of RF power measurements and reflection coefficient measurements. They are similar to error models commonly used in automatic network analyzer (ANA) systems [12], but include a number of additional terms.

In Fig. 3(a), a_{1T} is proportional to the input signal from the high-power microwave source, Γ_{1T} is the reflection coefficient presented to the input directional coupler by the input tuner, e_{16} represents coupling to the input power meter, and e_{15} represents directivity errors in the reference channel of the input network analyzer. All other terms have the usual significance [12]. For the output network in Fig. 3(b), e_{26} represents coupling to the power meter, and e_{25} represents directivity errors in the test channel of the network analyzer. The tuner is represented by the path Γ_{2T} and the node a_{2T}. This is a general form which describes passive, two-signal, and active tuners. A passive tuner has a nonzero Γ_{2T} term and zero a_{2T}, while two-signal and active tuners have nonzero a_{2T}. One step in the calibration procedure described below requires that a test signal is injected from the right of the output coupler. In this case, a_{2T} is nonzero.

The error-corrected reflection coefficients are easily obtained from Fig. 3 in terms of the uncorrected network analyzer reading at the input (TEST1/REF1) and at the output (TEST2/REF2)

$$\Gamma_s = e_{13} + \frac{e_{12}e_{14}\text{REF1/TEST1}}{1 - e_{11}\text{REF1/TEST1}} \tag{1}$$

$$\Gamma_{\text{in}} = \frac{\text{TEST1/REF1} - e_{11}}{e_{12}e_{14} - e_{11}e_{13} + e_{13}\text{TEST1/REF1}} \tag{2}$$

$$\Gamma_L = e_{23} + \frac{e_{22}e_{24}\text{TEST2/REF2}}{1 - e_{21}\text{TEST2/REF2}}. \tag{3}$$

The error-corrected input and output RF power levels are also obtained from Fig. 3

$$P_{\text{in}} = |P_1|^2 \cdot \left|\frac{e_{12}}{e_{16}}\right|^2 \cdot \frac{(1 - |\Gamma_{\text{in}}|^2)}{|1 - \Gamma_{\text{in}}e_{13}|^2} \tag{4}$$

$$P_{\text{out}} = \frac{|P_2|^2}{|e_{24}e_{26}|^2} \cdot \left|1 - e_{21}\frac{\text{TEST2}}{\text{REF2}}\right|^2 \cdot (1 - |\Gamma_L|^2) \tag{5}$$

where $|P_1|^2$ and $|P_2|^2$ are the readings of the input and output power meters, respectively. Note that (1)–(5) are independent of the tuner reflection coefficients Γ_{1T} and Γ_{2T}, and the directivity terms e_{15} and e_{25}. Thus, it is not necessary to obtain explicit values for these four terms.

B. Calibration-Input Port

1) The error terms e_{11}, e_{13}, and the product $e_{12}e_{14}$ are obtained by conventional ANA calibration techniques, in which the DUT is replaced by a series of calibration standards. A variety of different standards can be used, but in the present work a short circuit, an offset short circuit, and an open circuit [13] were employed. This provides all of the error terms in (1) and (2), and all but the $|e_{12}/e_{16}|^2$ error term in (4).

2) The $|e_{12}/e_{16}|^2$ term is obtained by connecting a matched power meter in place of the DUT. The ratio of this power meter reading to the input power meter reading is $|e_{12}/e_{16}|^2$.

C. Calibration-Output Port

1) The error terms e_{21}, e_{23}, and the product $e_{22}e_{24}$ are also obtained by conventional ANA techniques using a short, offset short, and an open circuit as calibration standards. The input signal for this stage of the calibration (a_{2T}) is injected from the right in Fig. 1 with the signal generator and TWTA connected in place of the load. This stage of the calibration gives all the error terms (3) and (5), except for $|e_{24}e_{26}|^2$.

2) The $|e_{24}e_{26}|^2$ term is obtained by measuring the magnitude of the insertion loss of the output coupler and power divider between the DUT output reference plane and the output power meter. If the test signal injected into the coupler at the DUT reference plane is supplied from a matched source, and if the output tuner is replaced by a matched load, then the power insertion loss I_p is given by

$$I_p = \frac{|e_{24}e_{26}|^2}{|1 - e_{21}e_{25}|^2}. \tag{6}$$

Since e_{21} and e_{25} are both small, the denominator is close to unity, and the insertion loss measurement gives $|e_{24}e_{26}|^2$ directly.

For maximum accuracy, the system is initially calibrated at coaxial (APC-7 or APC-3.5) reference planes close to the test fixture. The reference planes are then extended into the microstrip test fixture using accurate models of the coaxial-to-microstrip transition [14] cascaded with the models in Fig. 3.

D. Accuracy Check

The accuracy of the computer-corrected output power measurement was checked in the following way. The available signal power P_a at the output of the TWTA was measured using a matched power meter. The TWTA was then connected at the input port of the output directional coupler, in place of the DUT. The computer-corrected output power was obtained for a range of different output tuner settings across the Γ_L plane. These power measurements were compared with the calculated output power $P_a(1 - |\Gamma_L|^2)$, obtained using the measured P_a and the computer-corrected Γ_L (which was also checked for accuracy). The maximum difference between the computer-corrected and calculated output power was ± 0.15 dB.

IV. Examples

The load-pull system has been used to characterize a variety of power MESFET's at frequencies to 15 GHz. It is illustrated here with a number of examples of measured data for a packaged Dexcel DX3615A power MESFET.

A. Power Transfer Characteristics and Load Contour Maps

Fig. 4 shows an example of measured power transfer and power-added efficiency characteristics of the DX3615A at 8 GHz, with a drain–source voltage of 8 V and a drain current of 200 mA. The characteristics are shown for three different values of load reflection coefficient, ranging from the optimum load for maximum output power at an input

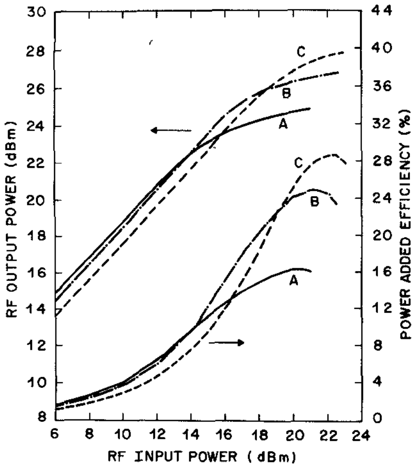

Fig. 4. Power transfer and power-added efficiency characteristics of the DX3615A at 8 GHz. In curves A, B, and C, the load reflection coefficient was adjusted for maximum output power at input power levels of 6 dBm, 17 dBm, and 23 dBm, respectively.

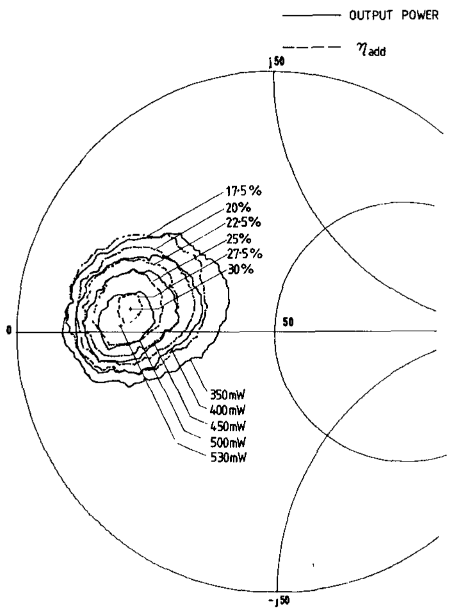

Fig. 5. Contours of constant output power (solid contours) and constant power-added efficiency (broken contours) on the Γ_L plane for DX3615A at 8 GHz.

power of 6 dBm, to the optimum load for maximum output power at an input power of 23 dBm. For each curve, the input tuner was also adjusted for maximum output power at the same reference input power level. This ensures a good large-signal match at the input. In Fig. 5 is shown a set of computer-generated load-pull contours of constant output power and constant power-added efficiency for the same device at the same frequency and with an input power of 125 mW (21 dBm). The input tuner was adjusted for maximum output power at this power level. The high resolution of the measurements enables the load for maximum output power to be clearly distinguished from the load for maximum power-added efficiency. The fine structure on the contours is due to noise in the interpolation scheme. Smoother contours can be obtained if additional data points are used.

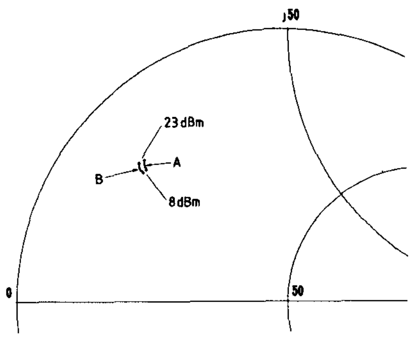

Fig. 6. Large-signal Γ_{in} of DX3615A at 8 GHz, for input power levels between 8–23 dBm. Curve A and curve B are for two different values of Γ_L (see text).

B. Input Reflection Coefficient

In the design of power MESFET amplifiers, it is often assumed that the input impedance of the device is linear [15] and is independent of the load impedance [16]. The validity of these assumptions for the DX3615A were investigated using the load-pull system. Fig. 6 shows the large-signal input reflection coefficient Γ_{in} of the DX3615A at 8 GHz as a function of input power level from 8 dBm to 23 dBm, and for two different values of the load reflection coefficient Γ_L. Curve A is for a Γ_L giving maximum output power at an input power of 6 dBm, and curve B is for a Γ_L giving maximum output power at an input power of 23 dBm. It is clear from these results that Γ_{in} is almost constant.

C. Harmonic Effects

An important consideration in the load-pull characterization of power MESFET's is the effect of harmonics generated in the DUT. Of particular interest are the effects of harmonic power detected by the output power meter and the influence of the load reflection coefficient at the second harmonic [3]. The spectrum of the output signal was observed for the DX3615A at frequencies above 5 GHz and for input power levels up to 24 dBm. It was found that the second harmonic component of the signal at the output power meter was always at least 20 dB below the carrier, and that associated errors in the measured output power were less than 0.05 dB. For devices where significant higher harmonics are present, a low-pass filter can be placed between the power divider and the output power meter to reduce the harmonic signal components detected by the power meter.

In order to determine the dependence of fundamental-frequency output power $P_{out}(\omega)$ on the load reflection coefficient at the second harmonic frequency $\Gamma_L(2\omega)$, a second tuner was connected in cascade with the output tuner. With this arrangement, it was a simple matter to adjust the fundamental frequency load $\Gamma_L(\omega)$ and the second harmonic load $\Gamma_L(2\omega)$ to different values. The load-pull system was fully calibrated both at the fundamental frequency (6 GHz) and at the second harmonic frequency (12 GHz) and the input signal was periodically switched between these two frequencies. The computer

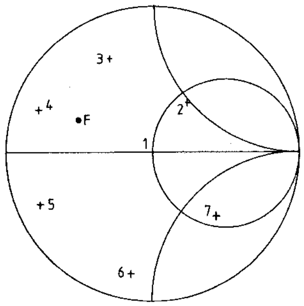

Fig. 7. Fundamental (F) and second harmonic (1–7) load reflection coefficients used for harmonic load-pull of the DX3615A. The chart is normalized to 50 Ω.

provided an almost continuous output of error-corrected values for $\Gamma_L(\omega)$, $\Gamma_L(2\omega)$, and $P_{out}(\omega)$. The load reflection coefficients used in this experiment are shown in Fig. 7. The point F represents $\Gamma_L(\omega)$, which was fixed at a value of $0.54\angle 157^\circ$. Seven different values of $\Gamma_L(2\omega)$, (labeled 1–7 in Fig. 7) were used. The output power $P_{out}(\omega)$ was measured for each $\Gamma_L(2\omega)$ and for two different values of RF input power (19 dBm and 21 dBm). It was found that the maximum variation of $P_{out}(\omega)$ with $\Gamma_L(2\omega)$ was only ± 0.04 dB for $P_{in} = 19$ dBm, and ± 0.06 dB for $P_{in} = 21$ dBm. Thus, the output power of DX3615A shows very little dependence on the second harmonic load.

V. Conclusions

A versatile system has been described for the accurate large-signal load-pull characterization of power transistors. Reflection coefficients and RF power levels are error-corrected using vector methods at both ports of the device under test. A significant feature of the system is that it treats the measurement of reflection coefficient and RF power in a unified manner. This leads to a simple calibration procedure which requires a minimum of calibration standards. The method is based on well-known automatic network analyzer techniques. It is, therefore, a straightforward matter to upgrade existing small-signal automatic network analyzer systems to include large-signal load-pull capabilities.

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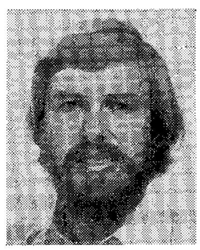

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