1. Experimental Characterization of Nonlinear Active Microwave Devices

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1. INTRODUCTION

The worldwide diffusion of wireless telecommunication systems opened new perspectives for microwave technologies. High-volume microwave monolithic integrated circuit (MMIC) production became a reality, revolutionizing the approach to microwave-circuit design. Nowadays, this is based on the intensive use of accurate simulation and computer-aided-design (CAD) tools. While scattering parameter measurements provide a reasonable accuracy for circuit modeling for linear operating conditions, large-signal applications still suffer the lack of accurate device models and a reliable experimental procedure.

A scheme related to the techniques and measurement systems for device and circuit microwave characterization is shown in Figure 1.

		LINEAR CHARACTERIZATION	NON-LINEAR CHARACTERIZATION	
		S-PARAMETERS	LOAD-PULL CHARACTERIZATION	NON-LINEAR MODELS
MEASUREMENT TECHNIQUES	HETERODYNE DETECTION	TWO-PORT NETWORK ANALYSIS MULTI-PORT NETWORK ANALYZERS	NETWORK ANALYZER-BASED INTERGRATED SYSTEM	MICROWAVE TRANSITION ANALYZER NON-LINEAR NETWORK ANALYZERS
	POWER DETECTION	SIX-PORT REFLECTOMETER	SIX-PORT-BASED LOAD-PULL SYSTEMS	————

Figure 1. Microwave measurement techniques and systems for device and circuit characterization.

When dealing with devices operating in large-signal (nonlinear) conditions, a strictly application-oriented technique was proposed by *Soares* [*1988*], where the device is tested in a configuration similar to the operating configuration. For instance, to design a single-stage power amplifier, the transistor is tested between two variable-impedance transformers, adjusted in order to measure the *optimum* performance in terms of gain, power output, etc. In other words, the aim is not to obtain a complete representation of the large-signal behavior of the DUT (device under test), but to find the design parameters that satisfy the final circuit specifications (i.e., the best loading conditions). These time-consuming and complex experimental procedures, known as *load-pull characterizations*, are carried out by using a special purpose test set, based on power meters or vector network analyzers. The latter take advantage of the vector information to measure many useful device parameters, and to perform sophisticated procedures for systematic error correction.

Other attempts were made to obtain a large-signal characterization, suitable for nonlinear model-based time-domain waveforms. These so-called nonlinear network analyzers represent a new approach, which combines a traditional load-pull system with a time-domain receiver, to offer a deeper investigation of a nonlinear model. These systems, working in the time-domain as very-large-bandwidth sampling oscilloscopes, acquire microwave voltage or current waveforms at the device terminals [*Hewlett-Packard Company*, *1991*]. The distorted waveforms, acquired by these so called *microwave transition analyzers*, allow establishing the complex relationships among the electrical quantities involved in the device model.

The alternative solution to frequency conversion is given by *six-port systems* [*Engen*, *1997*], which obtain microwave vector information from multiple microwave-power measurements. Nevertheless, power detection is inherently broadband, so that it is difficult to adapt six-port systems for harmonic-content measurements, which play a fundamental role when nonlinear measurements are involved.

2. MODERN MEASUREMENT SYSTEMS FOR MICROWAVE DEVICE AND CIRCUIT CHARACTERIZATION

Commercial test equipment is based mainly on heterodyne receiving. Automatic vector *network analyzers* have become the fundamental instrument for microwave device and circuit characterization. In their simplest versions, they allow swept-frequency S-parameter measurements of two-port circuits in small-signal conditions. They usually include: a microwave synthesized source; a test set to separate incident and reflected waves; a multi-channel microwave receiver to down-convert the separated signals; and a central unit for IF detection, A/D conversion, data processing, and presentation [*Hewlett-Packard Company*, *1991*].

The main factors limiting the accuracy of microwave vector measurements are the systematic errors due to the non-ideal nature of the test-set components. Imperfect coupler directivity and matching, as well as cable phase-shift and loss, modify the magnitude and phase relationship of the signals acquired by the microwave receiver with respect to the actual waves at the ports of the DUT. Consequently, *calibration procedures* have to be performed, in order to evaluate the systematic errors and to remove their effects from the measurements of the DUT.

Up to their introduction in the seventies, network analyzers were used as basic equipment for linear microwave characterization. More recently, their possible uses have been extended to large-signal applications.

Commercial network analyzers can be used to measure spectral content by locking the reference channel to the particular harmonic frequency to be measured [*Hughes, Ferrero, and Cognata, 1992*]. In principle, when compared to a spectrum analyzer, this technique should yield better accuracy, since the measurement results can be vector corrected [*Ferrero and Pisani, 1993*]. Nevertheless, it should be possible to reconstruct the distorted time-domain waveforms by using the inverse Fourier transform, providing that the relative phases of each spectral component were correctly measured. This involves practical problems, since the phase coherence is lost if the tuning frequency is changed to acquire the different tones. Several approaches were suggested to overcome this problem in the last few years [*Hewlett-Packard Company, 1991*; *Barataud, Arnaud et al., 1998*]. System calibration plays a fundamental role in microwave measurements. Owing to the complexity of the test set, calibration is responsible not only for systematic error correction, but – in general – deals with the extraction of the characteristics of the DUT from the raw instrument readings. Modern automatic load-pull systems [*Ferrero, Sanpietro et al., 1994*] are an example, where the network analyzer is used as a generic multi-channel receiver, and its readings are processed to obtain all the quantities of interest at the ports of the DUT (source and load reflection coefficients, input and output powers, harmonic content, and so on).

The conceptual steps involved in the calibration process are:

The error model. The relationship between the instrument readings and the quantities of interest of the DUT is defined by choosing a proper error model. Its parameters are usually called *error coefficients*, and they are characteristic of the test set .

The calibration procedure. This defines how the current values of the error coefficients are computed. This step usually involves measuring a proper sequence of reference circuits (i.e., *standard devices*), or performing a set of measurements with reference instruments connected to the test set.

The de-embedding is the final step. It consists of using the error model and the calibration results to extract – or *de-embed* – the characteristics of the DUT from the instrument's raw data.

A general scheme for a microwave measurement system, based on heterodyne receiving, is shown in Figure 2. The DUT is connected to the vector microwave receiver and to the other test-set components by a multi-port circuit: the *reflectometer junction*. This circuit usually consists of directional couplers and other passive components to feed the DUT, and to separate the incident and reflected signals.

The measurements are made in the frequency domain, and the quantities of interest are the DUT waves $a_i(f)$, $b_i(f)$. The target of the calibration process basically consists of computing the DUT's waves from the readings, $m_i(f)$. Under the assumption of a linear and invariant reflectometer junction, the error-model definition can be carried out for a generic *M*-channel microwave receiver. The result is that the DUT's waves are unique, *linear* combinations of the instrument's readings.

The calibration is carried out by connecting a set of reference circuits with known parameters – the *standards* – to the network analyzer. Usually, these circuits are simple, passive devices, so that their model can be theoretically predicted by physical measurements (e.g., geometrical dimensions, physical properties of the materials) [*Wong, 1992*], and electromagnetic simulations.

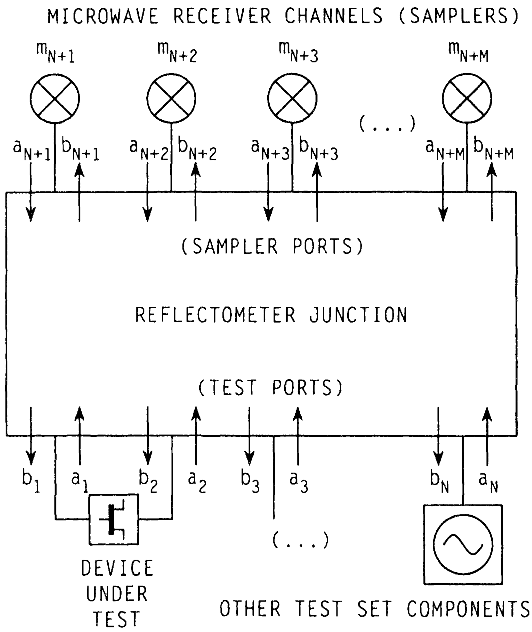


Figure 2. The general architecture of a microwave measurement system, based on heterodyne detection.

Different calibration techniques involve different standard connection sequences. For instance, the most traditional technique for two-port network-analyzer calibration, the SOLT algorithm [*Rytting*, *1982*], uses three one-port standards at each port: Short-Circuit, Open-Circuit, and Load (i.e., a matched termination), plus a Through connection (a direct connection between the measurement ports).

Not all the standard parameters have to be perfectly known: *self-calibration algorithms* use partially unknown standards, and compute their parameters as byproducts of the same calibration procedure. For instance, the most important self-calibration algorithm, called the TRL technique [*Bianco*, *Parodi*, *and Ridella*, *1976*; *Engen and Hoer*, *1979*], uses a perfectly known Through, plus other two standards: a transmission line with known characteristic impedance but unknown length; and a one-port reflective, unknown termination, connected, in turn, to both measurement ports.

3. SOURCE AND LOAD-PULL TECHNIQUES

For power amplifiers, transistor nonlinearities play a fundamental role in determining the optimum loading conditions. These are significantly different from the linear case, where the optimum Γ_L and Γ_S are directly evaluated from S parameters. Load-pull systems allow finding the proper load values experimentally, since the active device is driven by the microwave source at a single frequency, and its performance is measured while physically changing Γ_L (*load-pull*) or Γ_S (*source-pull*), as shown in Figure 3. The quantities of interest are typically the output power, the operating power gain and gain compression, the power-added efficiency, etc. Moreover, by driving the device with two-tone or CDMA-modulated signals, intermodulation or adjacent-channel power-ratio (ACPR) measurements can be performed to test the linearity of the amplifier under different loads.

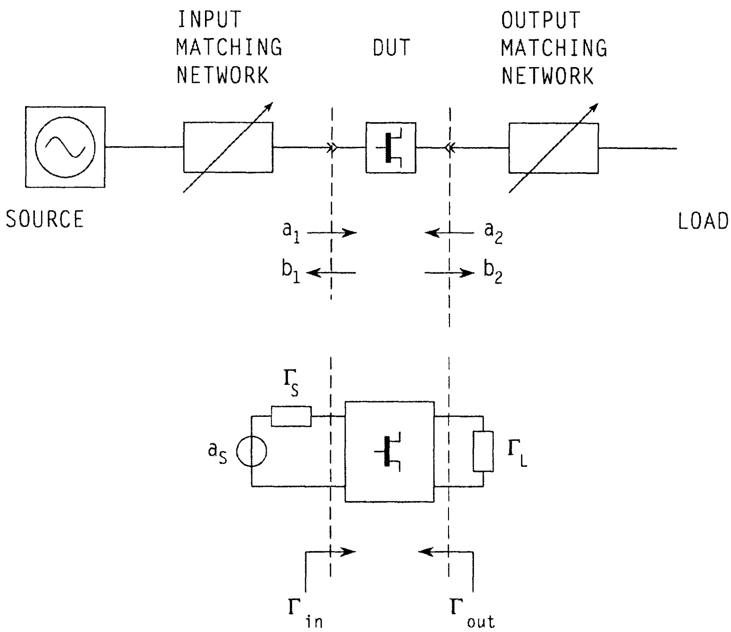


Figure 3. The definition of the quantities of interest.

The loading conditions at harmonic frequencies may also significantly affect the device performance, as already proven by both theory [*Snider*, *1967*] and experiments [*Ferrero and Pisani*, *1990*]. Harmonic source and load-pull systems allow changing Γ_S and Γ_L values at a discrete set of frequencies (typically, two or three).

These systems can also be used for nonlinear model validation and, in general, to find the optimum loading conditions of a microwave device operating nonlinearly, such as mixers [*Le and Ghannouchi*, *1998*] and oscillators [*Ghannouchi and Bosisio*, *1992*]. Moreover, source-pull systems have important applications in low-noise amplifier design, where the best noise figure and, in general, the transistor noise parameters are found by applying different source impedance values [*Adamian and Uhlir*, *1973*; *Caruso and Sannino*, *1979*; *Davidson*, *Leake*, *and Strid*, *1979*; *Le and Ghannouchi*, *1995*.]

3.1 LOAD REFLECTION COEFFICIENT SETTING AND MEASUREMENT

Designing a load-pull system involves dealing with three kinds of problems:

• setting a particular load value;

• measuring the imposed load reflection coefficients;

• measuring the required device performance.

Originally, load-pull systems [*Heiter, 1973*; *Cusack, Perlow, and Perlman, 1974*] were based on manual tuners and power meters. To get a variable load, passive networks with manual or automatic variable elements can be used (slug tuners, solid-state tuners, etc.) [*Sechi, Paglione, Perlman, and Brown, 1983*; *Adamian, 1989*; *Tsironis, 1992*; *McIntosh, Pollard, and Miles, 1999*].

At lower microwave frequencies (i.e., up to few gigahertz), the simple use of passive tuners is still the most effective and economical way to control the loading conditions of the DUT. The introduction of harmonic tuners [*ATN Microwaves, 1996*; *Focus Microwaves, 1998*], recently provided a compact and economic solution for harmonic-source and load-pull setup. On the other hand, the need for better measurement accuracy while characterizing highly mismatched transistors suggested the use of pre-matching networks [*Sevic, 1998*; *Tsironis, 1999*].

At higher frequencies, a test set based on passive tuners cannot offer highly reflective load conditions due to losses. The problem is especially evident for harmonic tuning, since the optimum harmonic termination is typically on the edge of the Smith chart. The alternative is an *active load*, which electronically synthesizes the required reflection coefficients by properly amplifying, phase-shifting, and combining microwave signals [*Takayama, 1976*; *Bava, Pisani, and Pozzolo, 1982*]. The active-load solution, which compensates for the losses, allows setting the reflectance at almost unity, and probably represents the most reliable scheme for microwave and millimeter-wave load-pull test sets. Concerning the problem of load measurement, systems which use power meters do not allow real-time measurement of the load. A preliminary network-analyzer characterization of the tuners impedance is required for all different tuner positions to be later acquired during the measurement. This operation can be really time-consuming, especially if harmonic tuning is required. Modern systems, based on vector network analyzers, perform a real-time vector-corrected load measurement while measuring the active device performance.

3.2 Active Load-Pull Systems

The first technique described in the literature for active load-pull is often referred to as the *two-signal technique.* It was originally due to *Takayama* [*1976*] [*Mazumder and Van der Puije, 1978*]. Figure 4a shows the test-set configuration: a power divider splits the source signal into two parts: the first drives the input port of the device, while the second is properly amplified, phase-shifted, and injected into the output port of the DUT as b_2. The load reflection coefficient is controlled by changing the attenuator and phase-shifter settings. This method is still considered for developing systems at millimeter wavelengths [*Bonte, Gaquiere et al., 1998*]. However, it suffers from a drawback, due to the difficulty of keeping the load value constant when the input power or the characteristics of the DUT change.

The second configuration is known as *the active loop technique*, and the basic scheme is sketched in Figure 4b. The DUT output's outgoing power wave is sampled by a directional coupler, properly controlled in amplitude and phase, and sent back to the DUT. The magnitude of Γ_L is ideally proportional to the loop gain, while the load phase depends on the loop phase shift. Initially, this technique presented severe problems in the form of oscillations due to the relative broad bandwidth of the loop components. The solution was reached by introducing a high selectivity bandpass filter, tuned to the measurement frequency, into the loop.

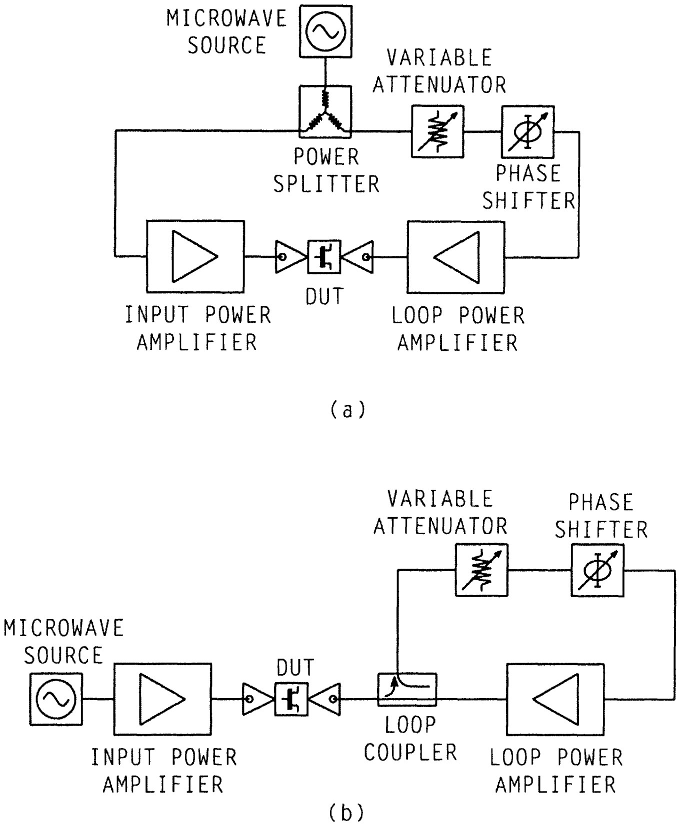
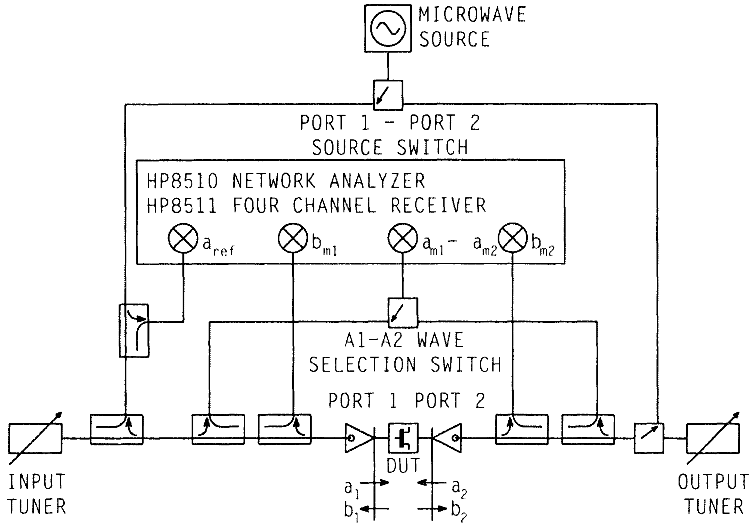


Figure 4. Active load-pull system configurations: (a) The two-signal technique; (b) The active loop technique.

Once the oscillation problem is solved, it is much simpler to control the load magnitude and phase with the active-loop solution. Moreover, since the loop is frequency selective, the same scheme can be easily extended for harmonic load-pull measurements [*Hughes*, *Ferrero*, *and Cognata*, *1992*], with a separate active loop for each harmonic. As an example, a 0.5-18 GHz harmonic system, based on this technique, is shown in Figure 5a. Signal a_{ref} provides a stable reference for phase locking, and the other waves are coherently sampled. This is a well-established technique, already experimented with in many load-pull systems [*Hughes*, *Ferrero*, *and Cognata*, *1992*; *Barataud, Arnaud*, *et al.*, *1998*], and it allows fast acquisition of all four DUT waves. For S-parameter measurements and load-pull at only the fundamental frequency, the main source is used for both driving the DUT and for providing a stable reference to the receiver. For harmonic measurements, two sources are used, as demonstrated in *Hughes*, *Ferrero*, *and Cognata* [*1992*], and shown in Figure 5b. The complete block diagram of the system, realized at the Politecnico di Torino, is shown in Figure 6 [*Ferrero*, *Pisani*, *and Madonna*, *1999*]. Two independent 0.5-4 GHz and 2-18 GHz active loops are enclosed, using a coaxial phase shifter, a broadband PIN diode attenuator, and a YIG filter as loop components.

The realized system allows performing S-parameter measurements without disconnecting the DUT by means of configuration switches, which bypass the input power amplifier and the active load. Reflectometers are mounted directly on the probe to improve the accuracy, while few other components are left outside the load unit for easy reconfiguration according to the user's needs. By changing the filter center frequency, each loop can be tuned on the fundamental or harmonic frequencies.

(a)



(b)

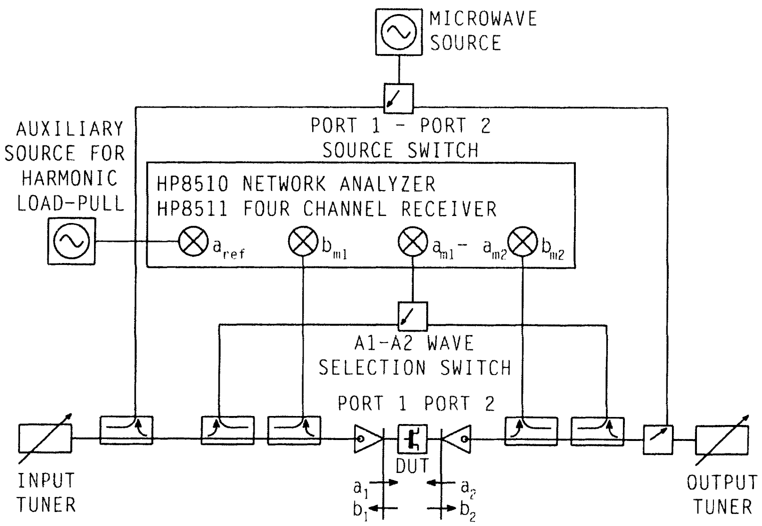


Figure 5. Network-analyzer-based systems: (a) Fundamental load-pull, using a single microwave source; (b) Harmonic load-pull, using an auxiliary source to tune the receiver.

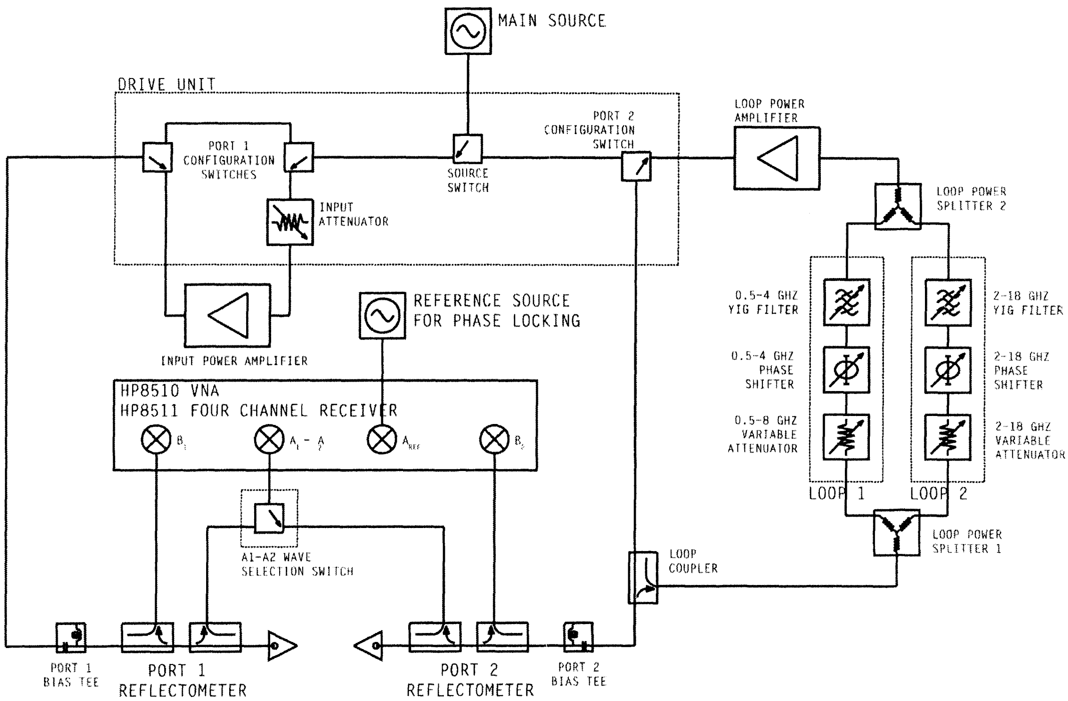


Figure 6. The 0.5-18 GHz measurement system, configured for harmonic load-pull.

The user can independently set:

- the frequency at which the DUT is driven (*source frequency*);
- the frequencies (one or more) at which the DUT is measured (*measurement frequencies*);
- the frequencies (one or two) at which YIG filters are tuned, i.e. the load reflection coefficients are controlled (*loop frequencies*).

In other words, it is possible to drive the DUT at one frequency, while controlling the loading conditions at two other frequencies and measuring the DUT at a third set of frequencies, all of which are not even harmonically-related with each other.

Vector error-correction theory for load-pull network-analyzer-based measurements is primarily described in three basic papers, by Tucker [*Tucker and Bradley, 1984*], *Hecht* [*1987*], and the authors [*Ferrero and Pisani, 1993*]. It is quite different from the traditional S-parameter calibration. The S parameters are defined as wave ratios. Consequently, the error coefficients are evaluated unless a normalizing constant – an additional parameter, which allows the setting of the absolute power levels at the DUT input and output ports – has to be found. That can be accomplished by connecting a power sensor at port 1, and measuring the coefficient:

$$|K_{11}|^2 = \frac{P_m}{P_{in}}, \tag{1}$$

where P_m is the power-sensor reading, and $\mathcal{P}_{in}$ is computed from the network-analyzer readings when the power sensor is connected.

A problem arises when on-wafer devices have to be measured with particular probes having port 1 or port 2 in the form of a coplanar line. Since power sensors are normally realized with the input in a coaxial line, the power measurement at port 1 is replaced by a power measurement at the coaxial tuner port (see Figure 7) when a direct (through) connection between ports 1 and 2 is made.

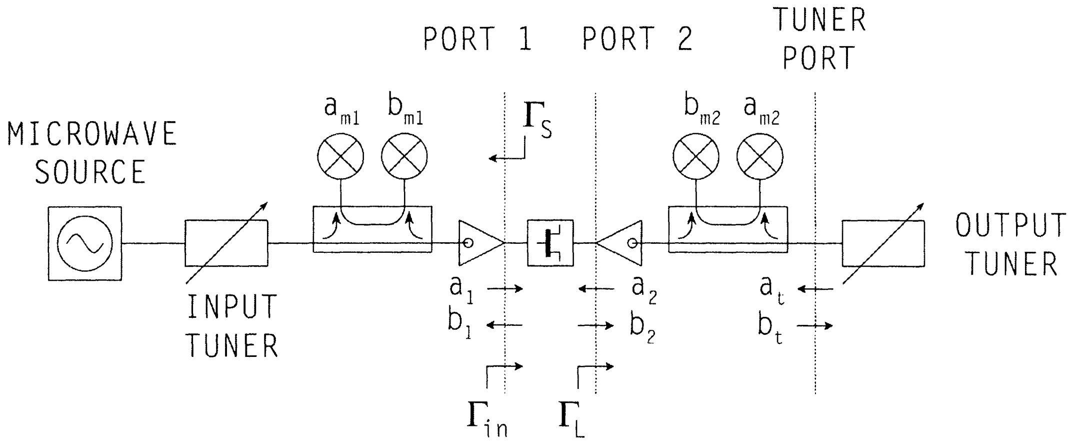


Figure 7. The reference planes for calibration of a generic source and load-pull system.

Typical measurements from a load-pull system are as follows:

- The *input power* P_{in} is found from

$$P_{in} \equiv |a_1|^2 - |b_1|^2 = |a_1|^2 \left(1 - |\Gamma_{in}|^2\right), \tag{2}$$

where $\Gamma_{in} = b_1/a_1$ is the device input reflection coefficient.

- The *available input power*, i.e., the maximum input power achieved by varying Γ_S when the condition $\Gamma_S = \Gamma_{in}^*$ is reached. The result is that

$$P_{av} \equiv P_{in}\Big|_{\Gamma_S=\Gamma_{in}^*} = |a_S|^2 \frac{1}{1-|\Gamma_S|^2}. \tag{3}$$

- The *output power*, i.e., the power delivered to the load Γ_L :

$$P_{out} \equiv |b_2|^2 - |a_2|^2 = |b_2|^2 \left(1 - |\Gamma_L|^2\right). \tag{4}$$

• Two different power gains are usually considered:

– the *operating power gain*, G_{op}, defined as the ratio of the output and the input power:

$$G_{op} \equiv \frac{P_{out}}{P_{in}} \tag{5}$$

– the *transducer gain*, G_t, i.e., the amplifier power gain when the input stage is matched:

$$G_t \equiv \frac{P_{out}}{P_{av}}. \tag{6}$$

• The *power-added efficiency* (PAE) is the net amount of RF power introduced by the active device, normalized by the supplied DC power:

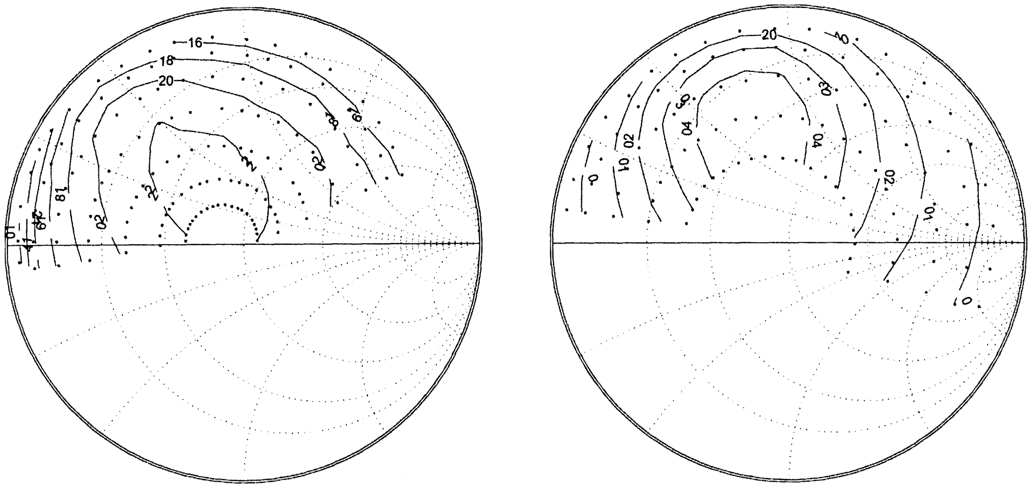
$$PAE \equiv \frac{P_{out} - P_{in}}{P_{dc}}. \tag{7}$$

All the previous quantities are functions of the reflection coefficients seen by the DUT from its input and output ports. Their dependence on the output load, Γ_L, at the fundamental frequency can be represented (for instance) by contour plots on the Smith chart.

Examples of load-pull contours are reported in Figures 8a and 8b. Figure 9 shows the power-swept measurement data acquired at a fixed Γ_L value for the DUT output power, gain, and efficiency.

3.3 Intermodulation Measurements

Intermodulation measurements are performed to test the linearity of power devices. A typical application is the power-amplifier design for radio links using digital QAM transmissions, where the performance of the telecommunication system is heavily affected by amplitude distortion. We suppose that the device is driven non-linearly by two tones, f_0 and f_1, with $f_0 < f_1$. Intermodulation products appear at frequencies $(mf_0 + nf_1)$, with $m, n = \pm 1, \pm 2, \ldots$. If $|f_1 - f_0| \ll f_0$, they fall within the amplifier's pass band, and they affect the signal amplitude of the channels adjacent to f_0 and f_1. Simple but fundamental contributions to modeling non-linearities in microwave transistors are due to *Heiter* [*1973*] and *Tucker and Rauscher* [*1977*],



(a) (b)

Figure 8. Large-signal load-pull contours: (a) Output power, in dBm, at 1 dB of gain compression; (b) Power-added efficiency, in percent, at 2 dB of gain compression.

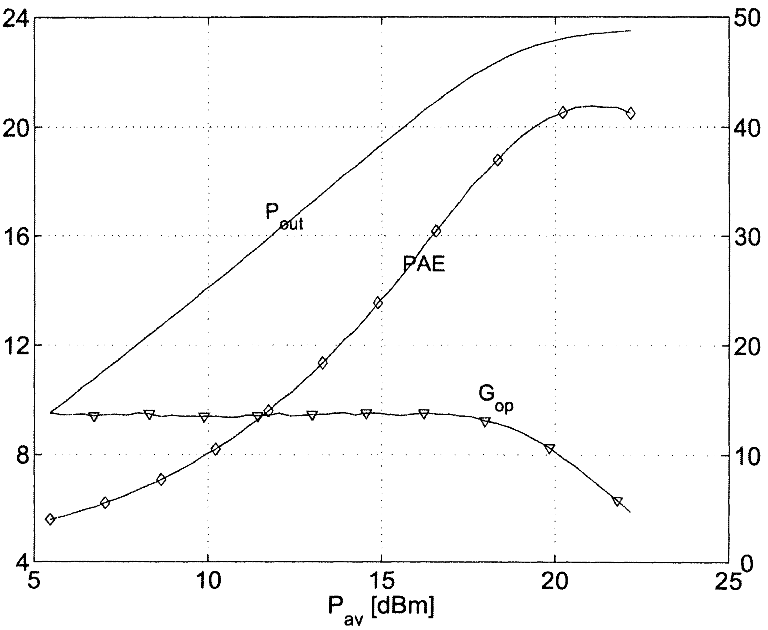


Figure 9. Power-swept measurement data for output power, operating gain, and PAE at $\Gamma_L = 0.32e^{j143°}$. The left vertical axis is P_{out} in dBm and G_{op} in dB; the right vertical axis is PAE in percent.

[*Tucker, 1979*]. The distorted-signal spectral components at frequencies $mf_0 - nf_1$ and $mf_1 - nf_0$ are called $(m+n)$ *th-order intermodulation products*. Their power levels are often written as

$$IM_{(n+m)L} \equiv P_{out}\left(mf_0 - nf_1\right) \tag{8}$$

and

$$IM_{(n+m)R} \equiv P_{out}\left(mf_1 - nf_0\right). \tag{9}$$

Figure 10 shows the typical amplitude spectrum of a two-tone, distorted signal. Since the distortion is weak, only *third-order intermodulation products* appear (i.e., at frequencies $2f_0 - f_1$ and $2f_1 - f_0$). Their power levels are IM_{3L} and IM_{3R}. The subscripts L and R stand for *left* and *right*, respectively, since the corresponding spectral components appear on the left and on the right sides of the two tones.

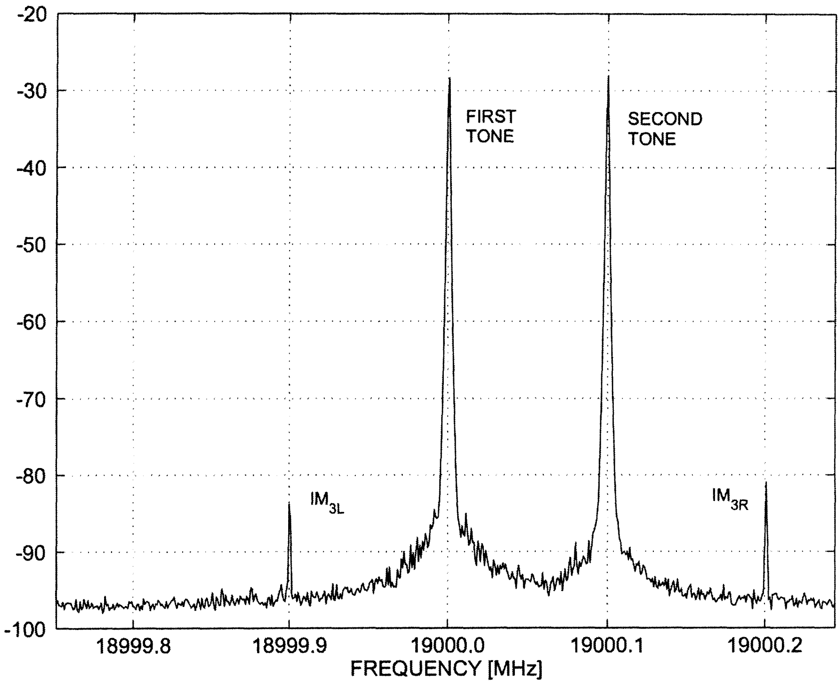


Figure 10. The amplitude spectrum of a two-tone, weakly distorted signal. Third-order intermodulation products appear next to the two carriers. The vertical axis is P_{out} in dBm.

Carrier-to-Intermodulation (C/I) ratios are defined as the power at the fundamental frequency divided by the intermodulation products:

$$C/I_{(n+m)L} \equiv \frac{P_{out}(f_0)}{P_{out}(mf_0 - nf_1)}, \tag{10}$$

$$C/I_{(n+m)R} \equiv \frac{P_{out}(f_0)}{P_{out}(mf_1 - nf_0)}. \tag{11}$$

Like all of the other device characteristics, intermodulation depends on the loading conditions. *Demmler et al.* [*1995*] described how to extend the load-pull system in [*Hughes, Ferrero, and Cognata, 1992*] for intermodulation measurements. Figure 11 shows the output power and the most significant intermodulation products as a function of the available input power, P_{av}.

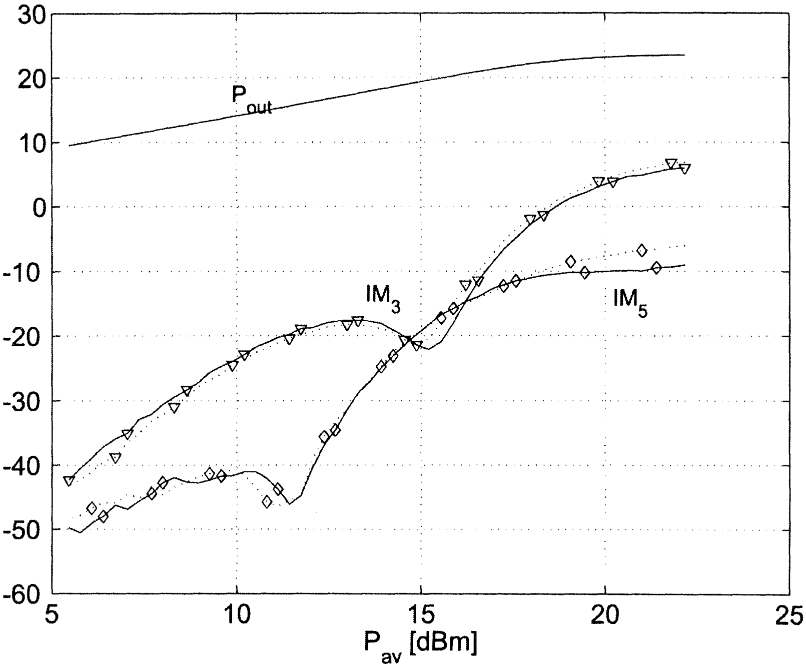


Figure 11. Output power and intermodulation: The continuous lines (—) represent the *left* third and fifth intermodulation products (IM_{3L} and IM_{5L}), while the dotted lines $(\cdots)$ represent the *right* products (IM_{3R} and IM_{5R}). The vertical axis is P_{out}, IM_3, and IM_5, all in dBm.

These last results were obtained by modifying the test set according to the scheme in Figure 12. Two tones were generated by two microwave synthesizers, each followed by an isolator. Their signal amplitudes were set separately to guarantee the same power level at the device's input port. The electronically-controlled input attenuator allowed automatic sweeps of the carrier power. At each loading condition and input power level, the following steps were performed. First, one-tone vector-corrected measurements were performed with the network analyzer by turning off the f_1 synthesizer and setting the f_0 synthesizer's output power to the nominal value P_0. Then, both synthesizers were turned on with nominal power $(P_0 - 3\,\text{dB})$ (so that the total input power was unchanged), and the C/I measurements were performed by the spectrum analyzer. The tone spacing, $\Delta f = f_1 - f_0$, had to be chosen to be compatible with the loop bandwidth.

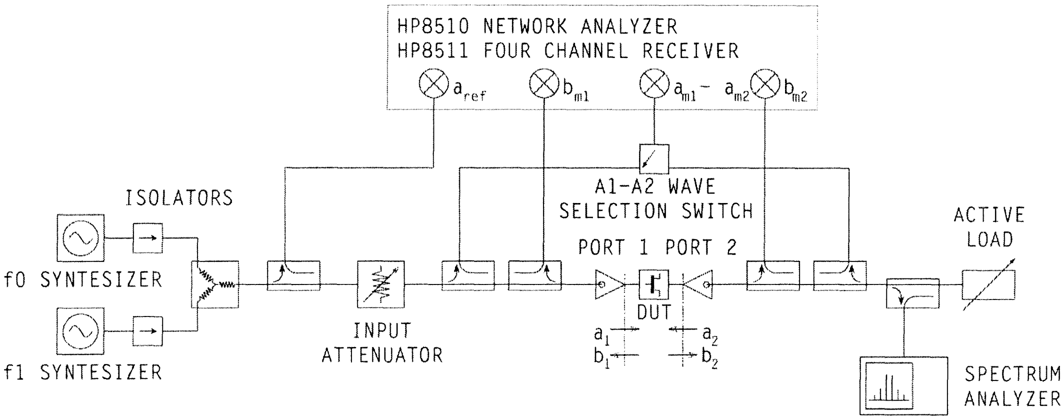


Figure 12. A load-pull system for intermodulation measurements.

4. SOURCE REFLECTION COEFFICIENT MEASUREMENT

After the calibration procedure described above, the system is able to measure the input reflection coefficient of the circuit connected to its test port. For example, the port 1 reflectometer in Figure 7 allows a calibrated measurement of the DUT input reflection coefficient, Γ_{in}, to be carried out using

$$\Gamma_{in} = \frac{b_1}{a_1}. \tag{12}$$

Equation (12) defines the relationship between the waves at the input reference plane, set by the DUT. On the other side, the microwave source sets

$$a_1 = a_S + \Gamma_S b_1, \tag{13}$$

where Γ_S is, by definition, the source reflection coefficient. From Equation (13), it results that

$$\Gamma_S = \frac{a_1}{b_1}\left(1 - \frac{a_S}{a_1}\right). \tag{14}$$

Therefore, a measurement of a_1, b_1 is not sufficient to compute the source reflection coefficient. In facts, Γ_S is equal to the ratio a_1/b_1 only if $a_S = 0$, i.e., if the internal generator is switched off. *Hughes and Tasker* [*1990*] proposed the solution shown in Figure 13a. First, the source switch is set to position 1, and the DUT input gamma is computed by Equation (12). Then, the source switch is turned to position 2, and a second acquisition of waves a_1, b_1 is performed. From Equation (14), the source reflection coefficient is simply the ratio $\Gamma_S = a_1/b_1$, since the source term is null.

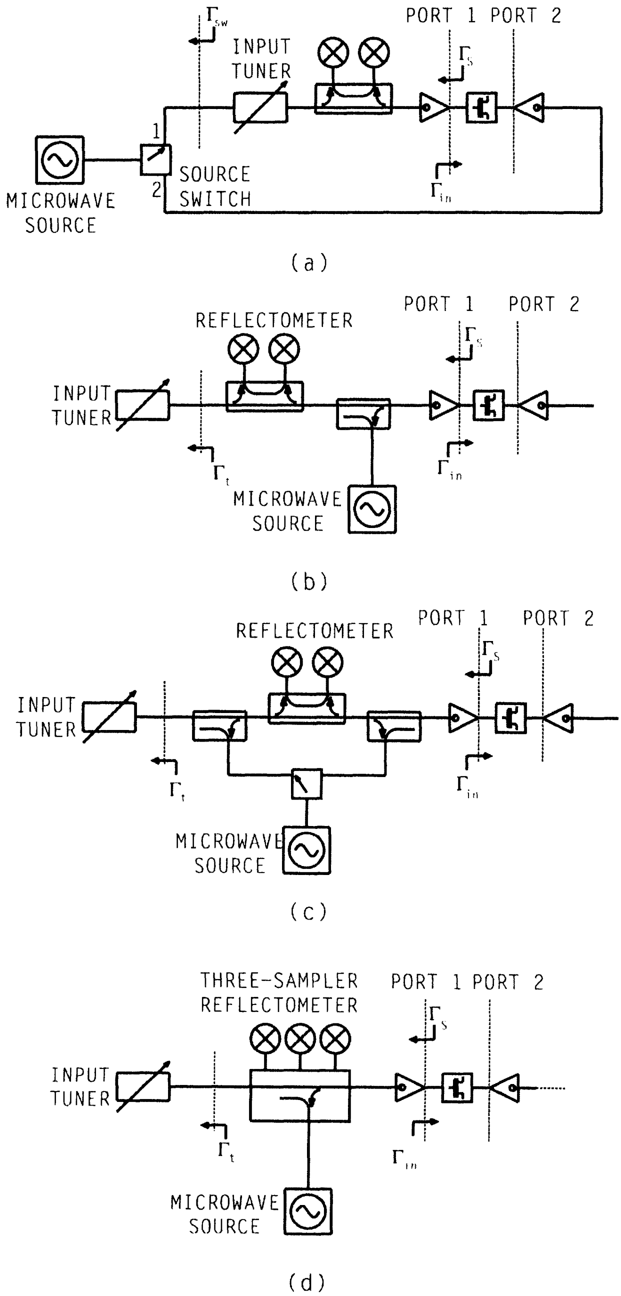


Figure 13. Existing solutions for source reflection coefficient measureme

A quite different approach is described in *Le and Ghannouchi* [*1994*], and sketched in Figure 13b. Here, the signal from the microwave source is combined with the wave reflected by the tuning element, and injected into the DUT. The reflectometer is used in an unconventional configuration (referred as *reverse*), and it directly monitors the tuner coefficient, Γ_t. After a proper calibration procedure, Γ_S is directly available, but – this time – it is the DUT reflection coefficient, Γ_{in}, that cannot be determined. The method shown in Figure 13c solves the latter problem in two steps [*Berghoff, Bergeault, et al., 1998*]. First, the microwave signal is injected before the reflectometer and the DUT input characteristic is computed. Then, it is switched immediately afterwards, and the source reflection coefficient is measured by the reflectometer in the reverse configuration. Again, the switch reflection coefficient is assumed constant while changing the switch position.

As common feature, all of the previous techniques measure the DUT and the source reflection coefficients in two different steps. That can be time consuming for fast and automatic characterization of active devices. A simple, yet rigorous, method has been recently presented in [*Ferrero and Madonna, 1999*] for determining the source reflection coefficient while characterizing active devices. Briefly, it consists of measuring the waves at the input reference plane under two different DUT bias conditions (a *double-bias* technique). The variations of the DUT input waves due to the bias change give enough information to compute the source reflection coefficient with an accuracy sufficient for most applications. An alternative technique, proposed in [*Madonna, Pirola, Ferrero, and Pisani, 1999*], is based on the concept of a *three-sampler reflectometer* (see Figure 13d), which allows the simultaneous determination of the source and DUT input gamma. This technique is indeed fast and accurate, but it is based on an unconventional error model, and it requires a special-purpose calibration procedure.

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