BROADBAND NOISE PARAMETER AND S-PARAMETER MEASUREMENT TECHNIQUE

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Abstract

An accurate noise parameter measurement technique is presented that uses an amplifier/attenuator noise source, a multi-octave tuner and a directional coupler for noise injection at all tuner states. The presented concept allows measurements with more than 30 different ENR steps and the ability to calibrate the ENR during system calibration. A system that demonstrates the concept was built for the frequency range from 0.5 - 6.0 GHz.

Introduction

Traditionally, noise parameter measurement systems use a diode in reverse breakthrough as a noise source [1]. Two different noise temperatures are generated by turning the diode on and off, resulting in excess noise ratios (ENR) between a few dB and up to 30 dB depending on the setup used. The lower noise temperature is normally at room temperature. Noise figure is then measured using the hot/cold noise source (Y-factor) technique [2]. This approach has three major drawbacks: (1) The ENR of the noise source has to be known over the measurement frequency range, (2) the source impedance of the noise source varies as the diode is turned on and off, and (3) there are only two different ENR's available for one specific noise source. The concept proposed in this paper overcomes all three problems by using a amplifier/attenuator noise source and a method to calibrate the ENR's [3].

Broadband noise parameter measurements require an equally broadbanded tuner in order to present at least four different source impedances (more to improve accuracy) to the device under test (DUT) [4,5]. The tuner has to be under electronic control with fast switching speed and good repeatability in order to make fast, computer controlled measurements [6]. We present a tuner concept, that fulfills all of these requirements.

Some systems make true hot/cold noise figure measurement only at a source impedance of 50 Ω. At all other source impedances, output noise power at this tuner state is measured and compared to the thermal noise power (cold source measurement technique [7]). Especially for DUT with low 50 Ω gain, such as low power transistors, this results rather inaccurate accurate measurements. By using a directional coupler, the noise figures can be measured with more than two source temperatures for all tuner states.

Measurement system

The measurement system [3] used to demonstrate the proposed concept (Fig. 1) combines three major parts: (1) A S-parameter measurement system using a HP 8753B

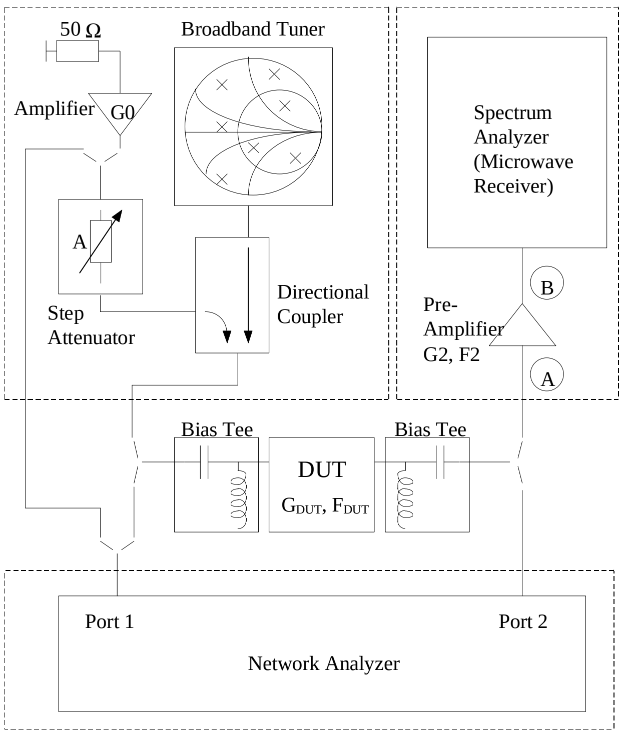

Figure 1: Measurement system setup

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Network Analyzer, (2) a noise source setup with a broadband amplifier as noise source, a 0 - 70 dB step attenuator (1 dB steps) for ENR generation, a broadband, variable PIN-diode tuner and a 16 dB directional coupler for noise injection, (3) an Anritsu 2602A Spectrum Analyzer with preamplifier as microwave receiver.

<u>Noise source:</u>

The noise power is generated by broadband amplification of the thermal noise of a 50 Ω resistor. The output of the amplifier contains white noise with a power of $P_{50} \bullet F_0 \bullet G_0$ where P_{50} is the 50 Ω thermal noise power, F_0 is the noise figure of the amplifier and G_0 is the gain of the amplifier. We may postulate that this amplifier is noiseless if we increase the figure for G_0 mathematically. After passing a step attenuator, a directional coupler and a bias tee, the available noise power presented to the DUT is

$$P_{DUT} = (1 + \frac{A-1}{G_0}) \cdot P_{50} \cdot G_0 \cdot \frac{1}{A} \qquad (1)$$

where A is the overall attenuation between noise source and DUT. By choosing an amplifier with a certain gain, we may obtain virtually every upper limit of ENR desired, while the lowest achievable noise temperature is at ambient for infinite attenuation. High ENR figures are important when measuring devices with no 50 Ω gain. The change of reflection coefficient for a change in ENR is less than 0.0006 at 1 GHz. ENR drift is monitored at low frequencies and a drift compensation may be used.

<u>Tuner:</u>

The source impedance of the noise power is dominated by the PIN diode tuner [5,8]. We built this tuner using 13 PIN diode short circuits logarithmically spaced on a 50 Ω microstrip transmission line (Fig.2). A 2 bit step attenuator at the input of the tuner shifts the reflections more towards the center of the Smith chart if desired. For each measurement frequency, a set of four to six shorts are used.

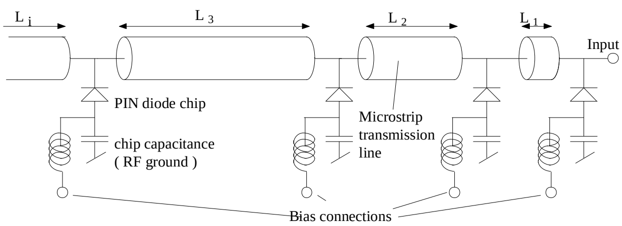

Figure 2: Tuner built by logarithmically spaced PIN diodes

The PIN diodes are mounted directly on a chip capacitor located in a substrate via hole. The bond wires are used to tune the parasitic capacitance of the PIN diodes when in off condition (Fig. 3, Fig. 4).

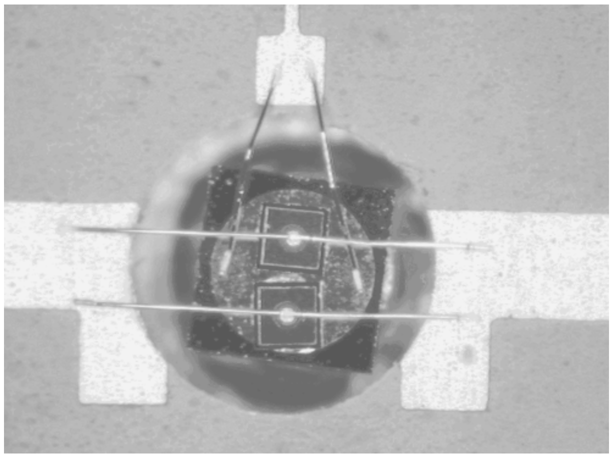

Figure 3: Photograph of a short circuit built by two PIN diodes mounted on a chip capacitor in a substrate via hole

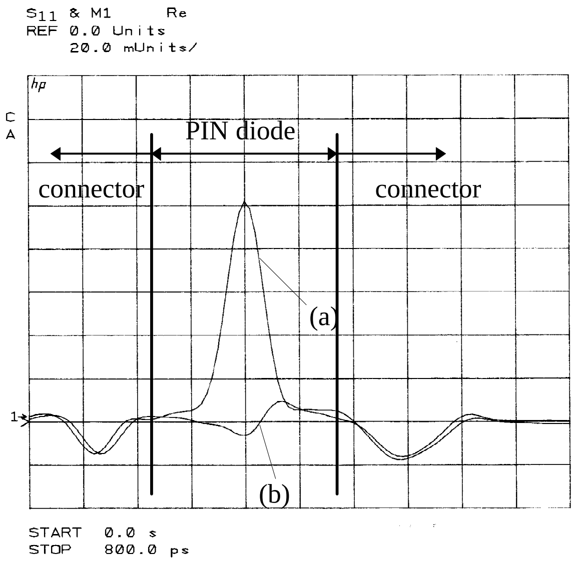

Figure 4: TDR response of one PIN diode short circuit in off condition; short built by (a) one diode (b) two diodes in parallel

This concept has several advantages compared to other tuner systems: (1) For all measurement frequencies, there are sets of shorts available with more or less equal electrical length. This way, the tuner states cover the Smith chart evenly independent of frequency (Fig. 5).

(2) This tuner concept is extremely broadband and is only limited by the available maximum length of the microstrip lines for low frequencies and by the minimal distance of the PIN diodes at high frequencies. (3) For high frequencies, the shorts close to the input of the tuner are used. The loss introduced by the transmission line and PIN diode combination are therefore small because of the small

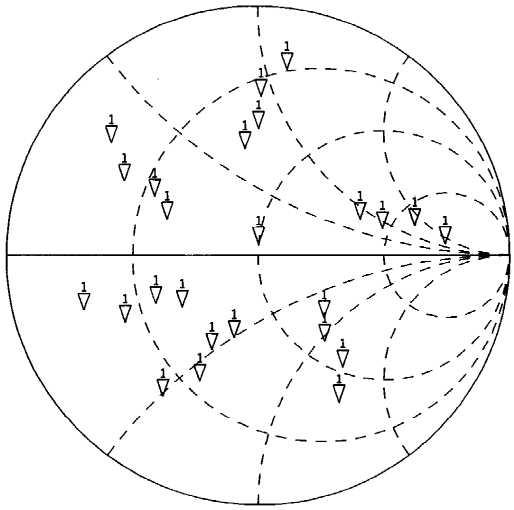

Figure 5: Typical set of noise source impedances at 1 GHz as seen by the DUT

number of shorts involved for these high frequencies. For lower frequencies, shorts with more distance to the input are used, but the loss at these frequencies is smaller. This results in a quite constant maximum reflection coefficient for all measurement frequencies. (4) Resonances due to multiple reflections are completely eliminated, provided that the unused PIN diodes present a small reflection (Fig. 4) coefficient and that at the far end of the tuner a 50 Ω termination absorbs all power that passed the short. (5) The loss of the tuner is minimized, because there is no switchable through connection used.

<u>Microwave receiver:</u>
As microwave receiver, a spectrum analyzer with a low noise broadband preamplifier was used. For the measurement of the noise power, we have two possibilities: (1) The Anritsu 2602A has a noise measurement option. The spectrum analyzer displays the total power in a adjustable frequency span. (2) The 1st IF output of the spectrum analyzer at 521.4 MHz can by converted down to 21.4 MHz and after filtering, a Boonton 4300 RF power meter displays the total noise power that passed this last filter. Both methods may be used, but the second one is more accurate because of the very exact power measurement with the RF power meter.

Calibration

Calibration of the system for a given measurement frequency range requires seven steps:

(1) The relative attenuation steps in the noise source step attenuator are measured with the 0 dB setting as reference attenuation.
(2) A S-parameter calibration is performed using a standard LRM method. At the end of this calibration procedure, the DUT is replaced by a through connection.
(3) All possibly used tuner reflection coefficients are measured from the port 2 side of the NWA.
(4) The reflection coefficient of the microwave receiver as presented to the DUT is measured from the port 1 side of the NWA.
(5) The loss between DUT and input of the preamplifier G_2 is determined by connecting a short, open and load respectively instead of the preamplifier G_2.
(6) The amplification of the preamplifier G_2 is measured in two steps. The NWA port 1 outputs a sine wave at the calibration frequency. The microwave receiver is connected once to point A in Figure 1 and the sine wave power is measured. Then, the preamplifier G_2 is connected between A and B and the sinewave power is measured again. The difference between the two measurements corresponds to the gain of the preamplifier G_2. The filter response of the microwave receiver can be calibrated at this time as well.
(7) For all attenuator and tuner steps, the noise power received by the spectrum analyzer is measured.

Now, the second stage noise figure and the available noise power that is presented to the DUT can be calculated for all tuner states. The second stage noise figure is determined when the noise source is maximally attenuated. It is the difference between the measured noise power (P_{50}•G_2•F_2). and the noise power we should measure for thermal noise power with no second stage noise contribution (P_{50}•G_2) plus the loss between DUT and preamplifier G_2.
A noise amplification factor G_0’ instead of G_0 is calculated using equation (1). G_0’ is a mathematical combination of the noise figure F_0, the gain of the noise amplifier G_0 and the overall loss A’ of the step attenuator plus the loss between step attenuator and DUT. The absolute value of loss A’ between noise amplifier and

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DUT has NOT to be known. Only the relative steps are required, because an offset in the loss A' is equalized by an offset in the calculated figure for G_0'. Because many different noise powers are measured, we can increase accuracy by averaging. This calculation of G_0' corresponds to a calibration of the ENR and now the complete system is calibrated by knowledge of the figures F_2, G_2, A and G_0' for all tuner states.
The accuracy of this system relays on the accuracy of S-parameter measurements and the linearity of power measurement and not on the accuracy of a diode noise source.

Results

Measurement is very similar to conventional systems [4,9,10,11]. The differences are: (1) We may use many different ENR with an input noise power level where the DUT has a good responsitivity: More noise power for low-gain / high-noise devices, less for high-gain / low-noise devices; (2) ENR averaging increases accuracy.

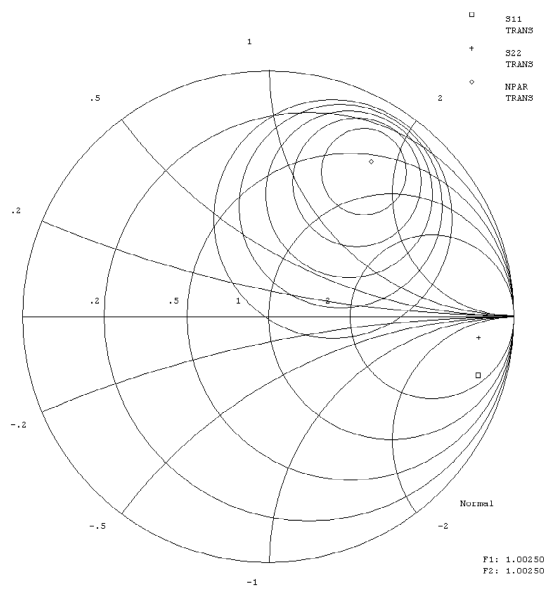

Figure 4: Noise parameters of a GaAs-FET measured at 1.025 GHz; noise circles correspond to 1,2,3,4,5 dB respectively. S11 and S22 of the FET are also shown

As a demonstration, the noise parameters of a GaAs enhancement FET was measured (Fig. 4) using a total of nine different tuner states. The FET has a gate length of 0.7 µm and a total gate width of 200 µm. The measured noise parameters at 1 GHz correspond well with design manual figures and were determined to be [4,10,12]:

Fmin	Rn	$\lvert Gamma_{opt} \rvert$	$\angle Gamma_{opt}$
0.678	5.61 Ω	0.756	56.6°

Conclusions

A broadband noise parameter measurement concept is presented. It uses an amplifier/attenuator combination with many different ENR's instead of a diode noise source with only two equivalent different noise temperatures, and it has the ability to calibrate the ENR.
Additionally, an ultra broadband tuner concept is shown that allows an increase in measurement frequency range especially towards lower frequencies.
Accuracy is based on S-parameter measurements, and not on knowledge about a fix noise source ENR.
To demonstrate the capabilities of the proposed concept, a computer controlled system for the frequency range from 0.5 GHz to 6.0 GHz was built and the noise and S-parameters of a low power GaAs FET were measured.

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