The Impact of Long-term Memory Effects on Diode Power Probes

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***Abstract*—This paper presents an analysis of long term-memory effects on power measurements with diode power probes. We show that a power probe calibrated with a single-tone sinusoidal excitation can provide erroneous values when used with modulated signals. This fact is ascribed to the low-frequency response imposed by the power probe baseband circuit. This hypothesis is first theoretically demonstrated by use of a Volterra series, and then validated by simulations and measurements using a diode power probe.**

***Index Terms*—Diode Power Probe, Long-term Memory Effects, Nonlinear Devices, Power Measurement.**

I. Introduction

DIODE power probes have been used for many years as high-speed power probes, and the results have been quite satisfactory when the power being measured is the power of a simple signal such as a sinusoid [1-3]. The nonlinearity of the diode in these probes rectifies an RF signal, providing a representation of the DC power in the signal. However, the nonlinear response of the power probe may depend on the bandwidth of the RF input signal, which can be an issue due to increased signal bandwidth of state-of-the-art wireless systems. This behavior can be ascribed to the dynamic interaction of the baseband impedance response and the low-frequency voltage and current excited when a nonlinear device under test is excited by any modulated excitation [4]. If not corrected, these dynamic effects may impact the reliability of the measurements. Commercially available microwave power sensors have been designed to work in a 50 Ω environment, eliminating impedance mismatches. However, simple diode power detector circuits are used in, for example, cell phone applications, to monitor the received power from the base station. For these circuits, the baseband embedding impedance can play a key role.

In this paper, we will study and analyze the impact of dynamic long-term memory effects on measurements made with diode power probes. This analysis will be done by comparing single-tone and two-tone excitations. The DC voltage corresponding to the detected power will be studied to explain the changes caused by dynamic effects.

Long-term memory effects in power amplifiers have been studied for many years. They are normally attributed to the low-frequency behavior of the amplifier, due mainly to the bias networks. Both input and output bias matching networks may cause these effects [5]. The frequency response of these networks imposes a change in the nonlinear behavior of the device, mainly introducing asymmetries in the upper and lower third-order intermodulation distortion products as a function of the bandwidth of interest.

In envelope detectors, this low-frequency interaction is intuitively expected. This is because, for this case, the objective is to down-convert the signal from RF to the baseband.

The impact of nonlinear distortion created by the change of operation to diode power probes is not so obvious. Because we are searching for the DC voltage created by the rectification of the diode, we may not expect that the effects of nonlinear distortion in the kilohertz or megahertz range would affect this DC value [1-3]. However, the nonlinear junction capacitance of the diode means that the superposition theorem is not valid anymore. As a result, the output of the diode is different for single or multiple tones. In addition, while the use of a single sinusoid does not create any baseband spurious signals, in a multitone arrangement, the baseband spurious signals have a bandwidth equivalent to the RF bandwidth.

These facts have motivated the authors to consider that baseband impedance terminations can affect the accuracy of the DC voltage measured by the probe. In the present work, we will study the impact of these terminations, first for a single sinusoid and then for a multitone signal. We also will develop methods to predict the effect of the baseband impedance on the measured DC values.

We will start first by presenting this problem with a mathematical approach based on Volterra series analysis. Next in Sections III and IV, an analysis of the impact of the baseband impedance will be carried out by use of both simulations and measurements when the system is excited by a single- and a two-tone signal. Finally, some conclusions will be drawn.

II. Mathematical Analysis

In order to understand the basic nonlinear mechanisms of diode power probes, consider the following circuit, in this case a very simple probe, used for demonstration purposes.

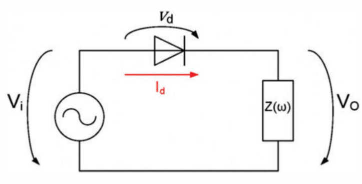


Fig. 1- Ideal,simplistic diode power probe.

A Volterra series-based mathematical approach will be applied to the circuit of Fig.1, considering a minimum number of terms to simplify the understanding of the sought-after behavior.

The diode model [6] is first approximated by a polynomial series expansion around a quiescent point truncated to order four. Thus the current through the diode is given by:

$$I_d = I_s + k_1 v_d + k_2 v_d^2 + k_3 v_d^3 + k_4 v_d^4 + \cdots \quad , \qquad (1)$$

$$\text{where } v_d = v_i - v_O \quad . \qquad (2)$$

The polynomial has been truncated to order four because we are using the minimum number of terms for a correct understanding of the nonlinear mechanism. Considering that we would like to obtain the output voltage of the probe versus the input RF signal, we can develop the Volterra series [4] as follows:

$$H_n(\omega_1, \omega_2, \ldots, \omega_n) = \frac{Y_{BB}(\omega_1+\omega_2+\cdots+\omega_n)}{X_{RF}(\omega_1+\omega_2+\cdots+\omega_n)} \quad , \qquad (3)$$

where $Y_{BB}(.)$ is the output voltage at baseband and $X_{RF}(.)$ is the input voltage at RF. This strategy was applied to the circuit of Fig. 1. The first two Volterra nonlinear operators are shown as examples in (4) and (5):

$$H_1(\omega_1) = Z(\omega_1)\frac{k_1}{1+k_1 Z(\omega_1)} \quad \text{and} \qquad (4)$$

$$H_2(\omega_1,\omega_2) = Z(\omega_1+\omega_2)\frac{k_2[1+H_1(\omega_1)+H_1(\omega_2)+H_1(\omega_1)H_1(\omega_2)]}{1+k_1 Z(\omega_1+\omega_2)}. \qquad (5)$$

If these formulas are further simplified, the change in DC voltage for a sinusoid excitation will be:

$$Y_{dc}(0) = H_2(\omega_1;-\omega_1)A_1^2 + H_4(\omega_1;-\omega_1;\omega_1;-\omega_1)A_1^4, \qquad (6)$$

where A_i is the amplitude of each tone. We see that the DC value is only dependent on ω_1-ω_1, or ω_1+ω_1, with the latter term equal to zero because of the low-pass behavior of the output matching network.

If we use a two-tone signal at the input, we have for the DC value

$$\begin{aligned} Y_{dc}(0) &= 2!\,H_2(\omega_1;-\omega_1)A_1^2 + 2!\,H_2(\omega_2;-\omega_2)A_2^2 \\ &+4!\,H_4(\omega_1;-\omega_1;\omega_1;-\omega_1)A_1^2A_2^2 \\ &+4!\,H_4(\omega_2;-\omega_2;\omega_2;-\omega_2)A_1^2A_2^2 \quad . \qquad (7) \\ &+4!\,H_4(\omega_1;-\omega_1;\omega_2;-\omega_2)A_1^2A_2^2 \end{aligned}$$

In this case, some terms will depend on ω_1-ω_1, as in the single-tone case, but also on the nonlinear mixing product falling at $\Delta\omega=\omega_2-\omega_1$. No matter what the relative phase offset between tones, the overall DC value will be equal due to the fact that the $Y_{dc}(0)$ is always a group of complex conjugated mixings. Thus, we see that the baseband impedance may affect the DC value. Assuming that $Z(\omega) \approx 0$ for RF frequencies (low-pass filter behavior), then H_2 and H_4 simplify to

$$H_2(\omega_1,\omega_2) = \frac{Z(\omega_1+\omega_2).k_2}{1+k_1 Z(\omega_1+\omega_2)} \qquad (8)$$

$$H_4(\omega_1,\omega_2,\omega_3,\omega_4) = Z(\omega_1+\omega_2+\omega_3+\omega_4) \times \frac{\left[\begin{array}{c} k_2\left(\frac{1}{3}H_2[\omega_1,\omega_2].H_2[\omega_3,\omega_4]+\frac{1}{3}H_2[\omega_1,\omega_4].H_2[\omega_2,\omega_3]+\cdots\right) \\ +k_3\left(-\frac{1}{2}H_2[\omega_1,\omega_2]-\frac{1}{2}H_2[\omega_1,\omega_4]-\frac{1}{2}H_2[\omega_2,\omega_4]-\frac{1}{2}H_2[\omega_3,\omega_4]+\cdots\right)+k_4 \end{array}\right]}{1+k_1 Z(\omega_1+\omega_2+\omega_3+\omega_4)}. \qquad (9)$$

To illustrate the concept behind these formulas, a load impedance with the behavior shown in Fig. 2 will be used. This impedance was designed to exhibit a resonance frequency around 750 kHz, to simulate a memory effect.

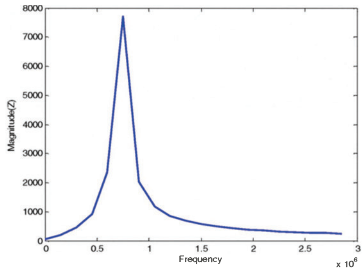


Fig. 2- Load impedance used as an illustrative example.

Using this impedance in (7) together with standard diode model values [6], $Y_{dc}(0)$ will present different values for different excitations. The DC response for a single-tone and two-tone excitation is shown in Fig. 3. In the one-tone case, we sweep the frequency of the tone over a 2.5 MHz frequency range near the carrier. In the two-tone case, we sweep the tone spacing from near 0 Hz to 2.5 MHz.

As shown in the figure, with a single tone, the DC output voltage is equal for all values of frequency. With two excitation tones, the resonance in the baseband impedance has a noticeable effect on the DC output.

This mathematical approach illustrates that the calibration of the overall system with a single sinusoid will not guarantee the calibration of the instrument for other type of excitations because they may be corrupted by long-term memory effects. For the case shown here, the maximum deviation appears at the resonant frequency.

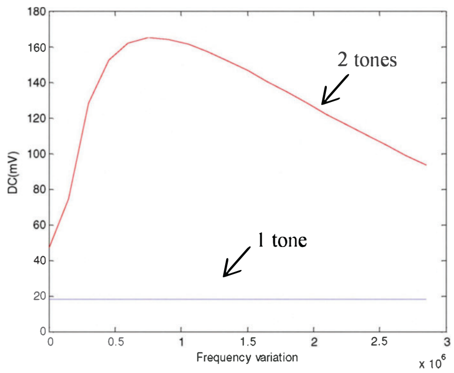


Fig. 3- Output voltage of the diode probe obtained through the Volterra model.

III. DC Probe Simulations

In order to validate the mathematical concept presented in the previous section, a diode power detector circuit designed for operation at 5.8 GHz, was simulated by use of a commercially available circuit simulator. The schematic of the simulated circuit is presented in Fig. 4.

The Vigo model of the diode was used [6]. An output-matching circuit was built in order to optimize the performance of the circuit, both at 5.8 GHz and at baseband, presenting a low-pass filter behavior, as illustrated in Fig. 5. A load impedance with a resonance at 2.25 MHz was inserted in order to clearly mimic the effect of the baseband impedance.

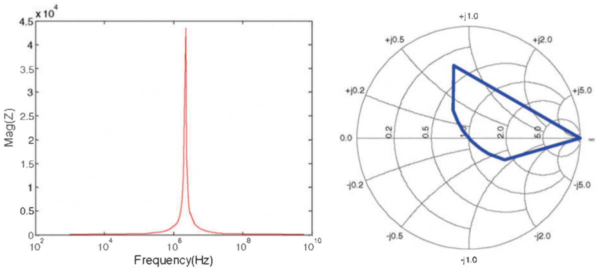
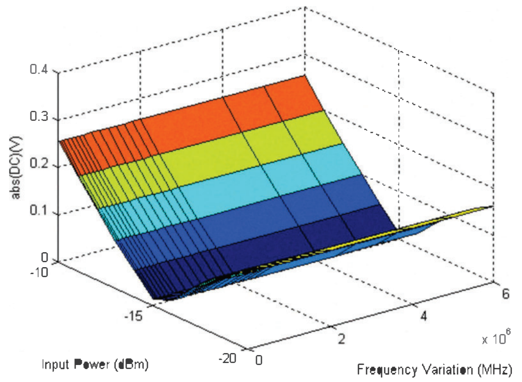


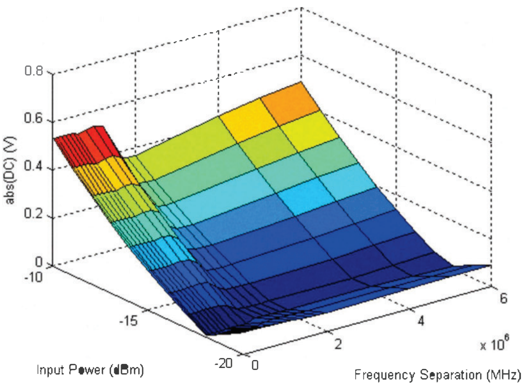
Fig. 5 – Diode's simulated output load impedance.

To simulate the behavior of the diode power detector, two simulations were conducted. In the first, a single tone was centered at 5.8 GHz and swept in frequency within a 6 MHz band around the carrier. The results are shown in Fig. 6(a). In the second simulation, a two-tone excitation was centered at 5.8 GHz. In this case, the tone separation was swept from 0 to 6 MHz. The power of each tone was half the power of the single-tone case, and varied from -20 dBm to -10 dBm. Results are shown in Fig. 6(b).

We expect that the voltage measured at the output of the diode probe for the two-tone excitation is almost twice that of the one measured with the single tone. In fact, this is the case for frequency separation near 0 Hz. But for other frequency spacings, especially close to the resonance frequency, the detector's output for the two-tone excitation changes significantly. The dip seen in both the single-tone and two-tone cases at a certain input power level (at -15 dBm for the single-tone and -18 dBm for the two-tone case) is due to the non-quadratic behavior of the diode for low values of input power. Normally, this is compensated for with a single-tone calibration. These simulated results validate our initial idea and mathematical analyses, which state that a calibration for the one-tone case will not guarantee the correct calibration for more complex forms of excitations.



(a)



(b)

Fig. 6 – Simulated results showing the DC voltage measured at the output of the diode power detector with (a) one- and (b) two-tone excitation.

IV. Measurements

The circuit of Fig. 4 was then fabricated and tested under the same conditions used in the simulation. The circuit is presented in Fig. 7. Our measurement set-up (Fig. 8) consisted of two RF sources to generate a two-tone signal, a spectrum analyzer, and a voltmeter.

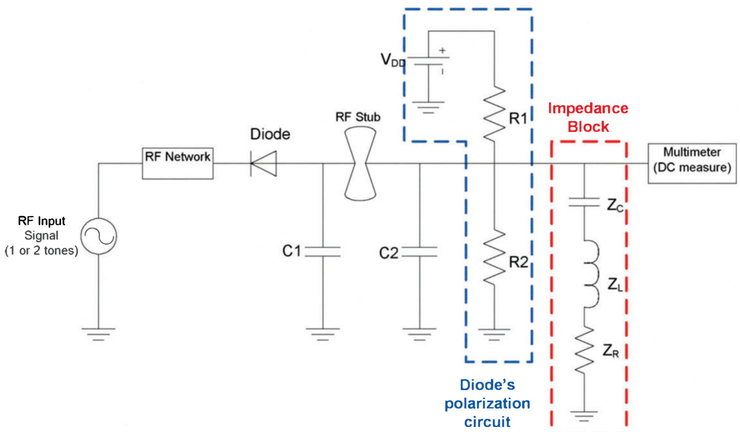


Fig. 4 – Schematic of the diode power detector circuit used in our simulations.

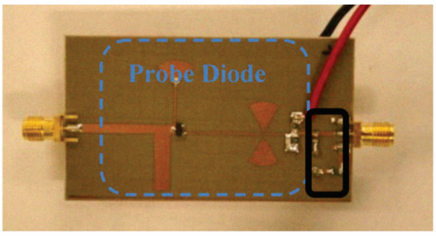


Fig. 7 – Power probe prototype used in laboratory tests

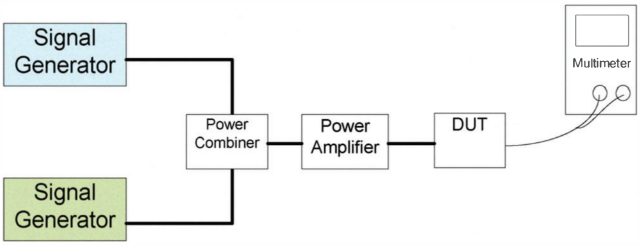


Fig. 8- Block diagram of the laboratory measurement set-up

We first measured the output impedance of the diode probe with a vector network analyzer. This measurement was made both with and without a multimeter attached. Because the measurements were made at low frequencies, the multimeter can actually play an important role in this characterization, as shown in Fig. 9.

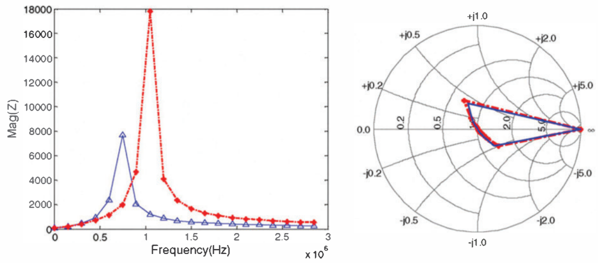
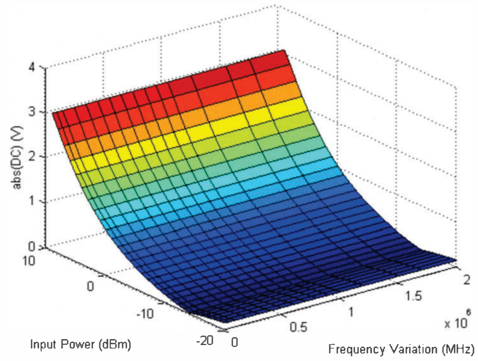


Fig. 9 – Diode power detector's output load impedance: impedance without multimeter (red dashed) and impedance with multimeter (blue solid) attached.

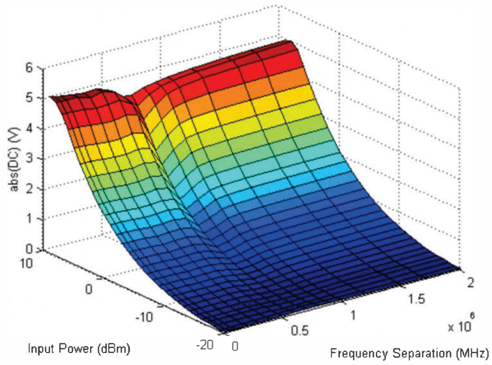
In Fig. 10 (a) and 10 (b), the measurement results for a single sinusoidal signal and two-tone excitation are presented, respectively. As can be seen in these figures, we observe a change in the DC value of almost 1V near the resonance frequency. The resonant frequency values are not equal to the simulated ones because of an inexact representation of the baseband impedance in the simulations. However, these results demonstrate our initial assumption that if the probe is calibrated with a single tone, it can give incorrect results when excited by a two-tone signal when the baseband impedance is a function of frequency, as is often the case for reactive matching networks.

V. CONCLUSIONS

This paper analyzed the impact of long-term memory effects on measurements made with DC power probes. The analysis, demonstrated mathematically, by simulations and by measurements, that the calibration of a power probe with a single-tone excitation does not guarantee the correct calibration for more complex forms of excitations. Further work is underway proposing new mechanisms for diode power probe calibration.



(a)



(b)

Fig. 10- Laboratory measurement results with one-tone (a) and two-tone (b) excitations.

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