Frequency Characterization of a 2.4 GHz CMOS LNA by Thermal Measurements

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Abstract — **This paper presents a new technique to obtain electrical characteristics of analog and RF circuits, based on measuring temperature at the silicon surface close to the circuit under test. Experimental results validate the feasibility of the technique. Simulated results show how this technique can be used to measure the bandwidth and central frequency of a 2.4 GHz low noise amplifier (LNA) designed in a 0.35 microns standard CMOS technology.**

Index Terms — **RF testing, RF characterization, thermal testing.**

I. Introduction

The scaling down of CMOS technologies has enabled the possibility of integrating a whole system on a single silicon chip (System on Chip, SoC), with the benefits of low cost, high reliability and low power consumption. A drawback of SoC integration is the significant loss of observability it entails, since fewer nodes are accessible from the outside. As a consequence, the complexity and cost of the test, monitoring and characterization of the individual parts of the system increase, specifically in RF front-ends where the operating frequency is in the range of gigahertzs [1].

The inclusion of built-in testers together with the circuit under test (CUT) [2] is one of the solutions proposed to increase the observability of analogue and RF systems. Requirements for this on-chip monitoring circuitry are low area overhead and almost inappreciable loading effects on the CUT.

This paper presents a novel procedure to characterize high frequency and RF analog circuits, consisting in measuring the temperature at the silicon surface, close to the circuit under test (CUT). As we present along this paper, the amplitude of some spectral components of the temperature near the CUT depends on the electrical characteristics of the circuit.

Measuring temperature to extract the electrical characteristics is an innovative and attractive solution as the CUT is not electrically loaded. The temperature measuring system gets the information thanks to the intrinsic thermal coupling provided by the silicon substrate. Thus, the electrical performances of the CUT are unaffected. Temperature can be sensed with embedded sensors, which would allow in-field test, as well as with external temperature sensors [3].

This paper is organized as follows: section 2 explains the principle of the procedure. Experimental measurements that validate its feasibility are in section 3, while section 4 shows an application of the technique based on simulated results: the bandwidth and central frequency of a 2.4 GHz LNA designed with a 0.35 μm CMOS technology are estimated. Finally, section 5 concludes the paper.

II. Principle of the technique

The proposed technique is based on three circuital properties:

i) The existing relation between the power dissipated by a device/circuit and its electrical characteristics.

ii) The frequency shift that can be observed in the spectral components of the power dissipated by a device/circuit when compared with the spectral components of the electrical signals that drive it.

iii) The thermal coupling provided by the silicon die, which allows the observation of the dissipated power by measuring temperature.

To illustrate the first two properties, let's consider a linear resistor R. When driven with a voltage sum of two tones of amplitude A and frequencies f_1 and f_2:

$$V_{in}(t) = A \cdot \left(cos(2\pi f_1 t) + cos(2\pi f_2 t)\right), \quad (1)$$

its dissipated power is:

$$P(t) = \frac{A^2}{2R}\Big(2 + 2cos\big(2\pi(f_2 - f_1)t\big) + +2cos\big(2\pi(f_2 + f_1)t\big) + cos\big(2\pi 2 f_2 t\big) + cos\big(2\pi 2 f_1 t\big)\Big). \quad (2)$$

As it can be seen, the electrical signals have only spectral content at frequencies f_1 and f_2, whereas the power dissipated has spectral components at f_2-f_1, f_1+f_2, $2f_1$ and $2f_2$. This effect can be called frequency mixing. In the case of having a resistor driven with (1) plus a DC bias, other spectral components of the power appear, for instance, at f_1 and f_2.

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On the other hand, the amplitudes of the spectral components of the dissipated power depend on the value of *R*, i.e., they carry information about electrical characteristics of the circuit considered. Then, by estimating the power dissipated we will obtain information about the circuit. The power will be estimated by measuring the temperature near the circuit, thanks to the thermal coupling of the silicon substrate.

Specifically, if we consider the power dissipated at f_2-f_1:

$$P_{f_2-f_1}(t) = \frac{A^2}{R} cos\left(2\pi\left(f_2 - f_1\right)t\right), \quad (3)$$

we observe the dependence of the power dissipation amplitude with *R*. Since in the example considered *R* is independent of the frequency, so it is the amplitude of P_{f2-f1}. But it must be remarked that by measuring the power dissipated at f_2-f_1 (which can be a low frequency if f_1 and f_2 are close to each other), the information that we obtain comes from high frequency, in this case f_1, f_2.

To show this, let us suppose that *R* depends on the frequency, R(f). In such a case, the amplitude of the power dissipated at f_2-f_1 becomes:

$$P_{f_2-f_1}(t) = \frac{A^2}{2}\left(\frac{1}{R(f_1)} + \frac{1}{R(f_2)}\right) cos\left(2\pi\left(f_2 - f_1\right)t\right), \quad (4)$$

Assuming that f_1 and f_2 are close and inside the working bandwidth of *R(f)* (i.e., f_2-f_1 is much smaller than the bandwidth of *R(f)*), then the power dissipated at f_2-f_1 is:

$$P_{f_2-f_1}(t) \approx \frac{A^2}{R(f_1)} cos\left(2\pi\left(f_2 - f_1\right)t\right), \quad (5)$$

From the last expression is obvious that the power dissipated at a low frequency as f_2-f_1 depends on a characteristic of the circuit at high frequency. Since the dissipated power will be estimated by thermal measurements, this down conversion is very useful, thus thermal coupling of the silicon substrate behaves as a low pas filter: some works have reported thermal measurements, corresponding to power dissipated, up to 100 kHz [3],[4].

If instead a so simple circuit as a resistor a more complex one is considered, for example an amplifier, is possible to find that the power dissipated at f_2-f_1 when two tones are applied depends also on the characteristics of the circuit, for example, the voltage gain of the amplifier.

From the analysis performed in this section, we propose a new technique for characterizing analog and RF circuits consisting in driving the circuit with the sum of two tones and, if the spectral component of the power dissipated at f_2-f_1 carries electrical information about the characteristics of the circuit at high frequencies, to derive these characteristics by thermal measurements at f_2-f_1.

III. Feasibility of the Technique

The goal of this section is to show the feasibility of the temperature measurements needed to track the spectral content of the power dissipated at f_2-f_1.

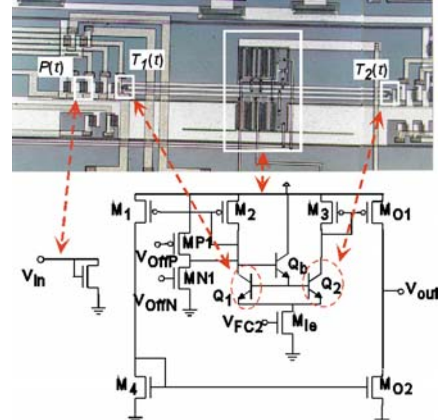

Fig. 1 Experimental set-up and sensor schematic.

Fig. 1 shows the layout of the IC used in the experimentation. It contains MOS devices in diode configuration (Fig. 2) which behave as dissipating devices and a differential temperature sensor embedded in the same silicon die. The output voltage of differential temperature sensors is proportional to the difference of temperature at two points of the silicon surface. In our case, the temperature at the sensing points is named $T_1(t)$ and $T_2(t)$ [3][5].

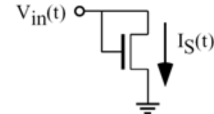

Fig. 2. Dissipative NMOS transistor in diode configuration.

The aspect ratio of the dissipating device is 10/1.2 and it is placed at 47 microns from the temperature monitoring point $T_1(t)$. Fig. 3 shows the I/V DC characteristic of the dissipating device. This plot can be fitted to a third order polynomial, resorting to the following expression:

$$\begin{array}{l} I_s(V_{in}) = m_0 + m_1 \cdot V_{in} + m_2 \cdot V_{in}^2 + m_3 \cdot V_{in}^3 \\ m_0 = -0.14 \cdot 10^{-3} \quad m_1 = -0.25 \cdot 10^{-3} \\ m_2 = 0.56 \cdot 10^{-3} \quad m_3 = -49.74 \cdot 10^{-6} \end{array}, \quad (6)$$

Fig. 4 (left) shows the experimental set-up: the dissipating device is activated with two tones of frequencies f_1 and f_2. The sensor's output is connected to a spectrum analyzer. Fig. 4 (right) shows the spectrum of the sensors output when f_1 = 1 kHz and f_2 = 1.130 kHz. In this

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figure the temperature frequency components at f_2-$f1$, $2f1$-f_2, $2f_2$-$f1$, f_1 and f_2 are clearly observed. If we apply two tones plus a DC voltage for polarizing the diode, the power dissipated at frequency (f_2-f_1) is:

$$P_{f_2-f_1} = \left(A^2 m_1 + 3A^2 m_2 V_{DC} + 6A^2 m_3 V_{DC}^2 + 3A^4 m_3\right) \quad (7)$$

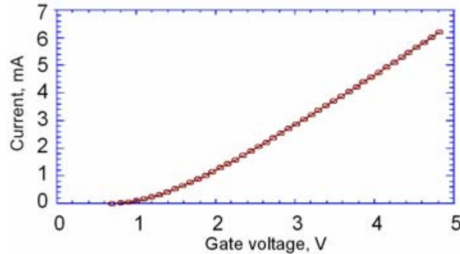

Fig. 3. I/V DC characteristic of the dissipative transistor.

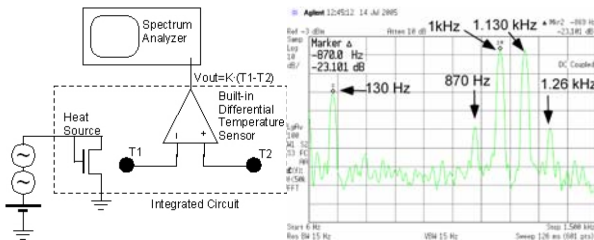

Fig. 4. Experimental set-up (left) and temperature sensor output spectrum (right).

Analyzing this expression we can see that the first term $A^2 m_1$ corresponds to the linear behavior of the diode (similar to the case of the resistor), and the other three terms correspond to the non-linearity of the diode.

To validate experimentally expression (7) we have replaced the spectrum analyzer of Fig. 4 by a Look-in Amplifier. Fig. 5 shows the amplitude of the spectral component at (f_2-f_1) of the sensor's output voltage versus amplitude A. We have fixed (f_2-f_1) at 1 kHz for three different values of f_1: 100 kHz, 1 MHz and 10 MHz, with V_{DC} fixed to 2V.

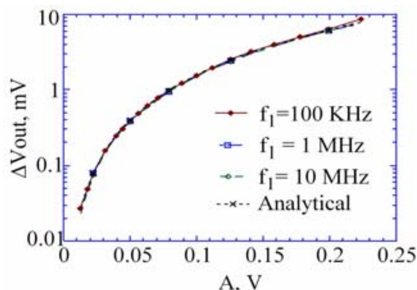

Fig. 5. Amplitude measurements from the Look-in Amplifier (V_{out} sensor), and equation (7) multiplied by the sensitivity of the sensor (all values are at f_2-f_1).

Additionally, the estimated amplitude of sensor's output is plotted in the same figure, which is obtained by multiplying analytical expression (7) by the sensitivity of the sensor (81.73 V/W [5]). In Fig. 7 the power amplitude at (f_2-f_1) estimated from analytical equation (7) is drawn.

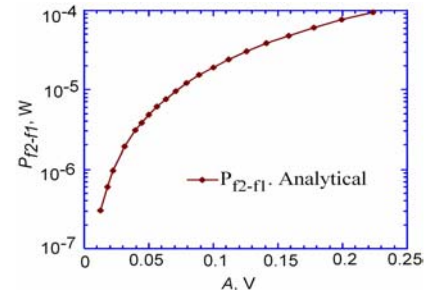

Fig. 6. Power amplitude estimation at f_2-f_1 from analytical equation (7).

From these figures we observe:

i) By comparing V_{out} for the three values of f_1, we see that the magnitude of the spectral component of the temperature at f_2-f_1 does not depend on the absolute value of f_1. This means that the m's coefficients in expression (13) do not depend on the frequency in the range considered.

ii) We also observe a good matching between the measured and the estimated output voltage V_{out}, that validates the analysis performed.

iii) We can also observe from Fig. 6 that power dissipation magnitudes lower than 1 µW can be detected with temperature sensors embedded with the CUT.

IV. Application of the Technique

In this section we consider a tuned amplifier, whose gain is frequency dependent. Since the amplitude of the power dissipated at f_2 -f_1 depends on the gain of the amplifier, if the value of f_1 and f_2 are swept (keeping constant the value f_2 -f_1), then it is possible to obtain the bandwidth and the central frequency of amplifiers by plotting the magnitude of the dissipated power as a function of f_1.

The circuit considered is a fully differential 2.4 GHz Low Noise Amplifier (see Fig. 7), designed in a 0.35 µm CMOS technology. Its electrical performances are: 24 dB voltage gain (16 in linear), 15 dB input return loss, 5 dB NF, -10 dBm 1 dB compression point and 10 mA DC current consumption. We have applied to its input two tones of equal amplitude (-15 dBm) and frequencies f_1 and f_2=f_1 + 1 kHz. By sweeping f_1 from 2 GHz to 3 GHz (and keeping always 1 kHz between the two tones) we have obtained by means of simulation both the voltage gain A_v

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inside this frequency range, as well as the amplitude of the power dissipated by the *MCn* transistor at 1 kHz, P_{1kHz}^{MCn} .

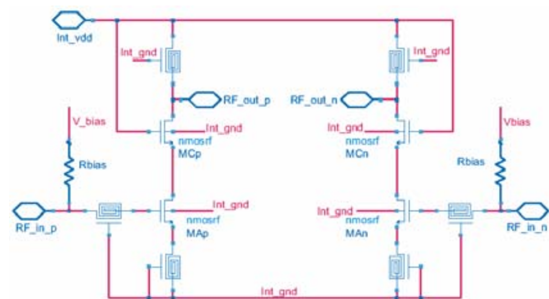

Fig. 7. LNA schematic.

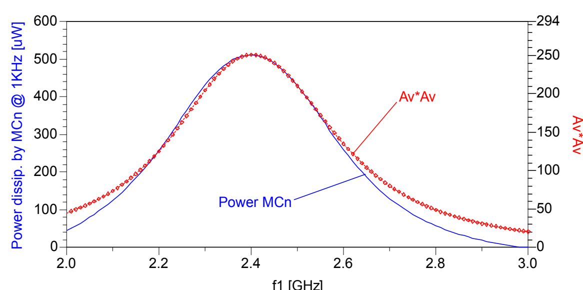

Fig. 8. LNA voltage gain squared (right axis) and power dissipated at 1 kHz by the MCn transistor, P_{1kHz}^{MCn} (left axis).

In Fig. 8 we show P_{1kHz}^{MCn} and A_v squared (note that the power dissipated at f_2-f_1 by an active device depends on the voltage gain squared). The two curves shown in the figure agree well in the band of interest. From the P_{1kHz}^{MCn} curve (solid line, left axis) it is possible to estimate that the LNA -3 dB bandwidth is *BW*=402 MHz and that its central working frequency is f_0=2.398 GHz. If we use directly the $A_v \times A_v$ curve (dashed line, right axis), the values obtained are BW=414 MHz and f_0 =2.400 GHz. The error between the original *BW* and f_0 characteristics and their values obtained from the low-frequency dissipated power is 3 % and 0.1 %, respectively. This demonstrates that some electrical characteristics of the LNA at 2.4 GHz can be effectively estimated by monitoring the power dissipated by one of its devices at 1 kHz (power that, as has been shown in the previous section, can be obtained by thermal measurements).

V. Conclusion

In this paper we have presented a novel strategy to observe electrical characteristics of analog/RF circuits. It is based on measuring the amplitude of some spectral components of the temperature when sensed at the silicon surface, close to the circuit under test.

This test and characterization strategy presents the following advantages:

i) The circuit under test is not electrically loaded, and then its electrical performances are unaffected.

ii) There is no need to have direct access to electrical nodes of the CUT to measure electrical signals.

iii) Temperature sensors can be embedded with the CUT, allowing in-field test and characterization.

iv) The information is read at f_2-f_1, independently of the absolute values of f_1 and f_2. Thus, temperature sensors and filters of the testing/characterizing system are designed for this low frequency, independently of the specific spectral band at which f_1 and f_2 are placed.

The experimental results presented have shown that it is possible to measure power amplitudes lower than 1 µW using differential temperature sensors embedded with the CUT. Simulated results have shown an application of the technique by estimating the BW and central frequency of a 2.4 GHz LNA designed in a 0.35 µm CMOS technology.

Acknowledgement

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