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DC Temperature Measurements to Characterize the Central Frequency and 3 dB Bandwidth in mmW Power Amplifiers

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***Abstract*— This letter shows how a temperature sensor and a simple DC voltage multimeter can be used as instruments to determine the central frequency and 3 dB bandwidth of a 60 GHz linear power amplifier (PA). Compared to previous works, the DC temperature monitoring now proposed requires a much simpler and convenient measurement set-up. In this example, the temperature sensor is embedded in the same silicon die as the PA. Being placed in empty layout spaces next to it, it is proposed as a built-in test circuit.**

***Index Terms*— CMOS millimeter wave integrated circuits, Design for Testability, temperature measurement, built-in test.**

I. Introduction

Conventional CMOS technology is used today to implement mmW communications circuits like those under the IEEE 802.11ad standard. To obtain tolerance to process-voltage-temperature variations and aging, circuit overdesign is usually applied, which ultimately implies increasing the power consumption to guarantee the targeted performance. Embedding sensors that are able to monitor the circuit operation together with active tuning strategies, is one solution proposed to push up circuit performance [1].

Thanks to the Joule Effect, the dynamic operation at high frequencies is translated to very low frequency dynamics in the thermal domain. This allows using low-frequency thermal sensors to monitor the high-frequency performance of circuits, even those working at the millimeter waves. Besides moving the measurement to low frequencies, this thermal approach has the advantage of not loading any node of the circuit under test, thus can be applied to existing circuits without redesign.

In a recent letter published in this journal [2], low frequency temperature measurements were proposed to determine the central frequency and the 3 dB bandwidth of a 60 GHz power amplifier (PA). In that work, the PA was driven with two tones whose frequencies were f_{IN} and f_{IN}+5 kHz. While f_{IN} was swept from 50 GHz to 67 GHz, the spectral component of the temperature at 5 kHz was monitored by measuring at the output of a built-in thermal sensor. The strategy of using two tones to drive the circuit under test in order to achieve a low frequency temperature increase is known as heterodyne temperature measurement [3].

One disadvantage of this heterodyne approach is the difficulty of generating two high-frequency tones with a small and precise spacing –six order of magnitude difference, in [2]– and on top of that sweep them over a large range. Although it is possible to use two signal generators with external synchronization [2], the complexity of the set-up is still remarkable. In addition, the measurement requires a lock-in amplifier, or a FFT representation of the measured waveform, to extract the 5 kHz component of the output.

To avoid this measurement complexity, an alternative approach known as homodyne temperature measurement uses a single input tone and DC temperature measurement [3]. This strategy has been used before [4-7] to monitor several other metrics of RF PAs, and in [8] as monitor of the structural integrity (structural test) of a RF LNA, but never used to characterize the frequency response in mmWave amplifiers. In this letter we demonstrate the feasibility of the homodyne technique to measure the central frequency and 3 dB BW of a mmW PA, using both a theoretical analysis and proof-of-concept experimentation. Compared to the heterodyne approach, a single signal generator and a simple DC multimeter are enough to perform the measurements. Also, the DC measurement enables on-chip digital calibration techniques by using embedded non-invasive monitoring.

II. Circuit Analysis

The simplified model of a linear (non-switched) PA can be described as a transconductor and a resonant RLC load connected to the output. This tuned load resonates at frequency ω_0 and is responsible for the frequency behavior of the amplifier.

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A PA is far from working in small signal, but a linearization of its input-output transfer characteristic in the frequency domain can be used to quantify the relationship between the input voltage amplitude and the amplitudes of the harmonics generated at the output node. By definition, the relationship between the output and input amplitudes at the input frequency ω_{in} is the describing function for the nonlinear gain of the transconductor [9]. Assuming G_m is the magnitude of the describing function of the PA and assuming that when the PA is driven by a single tone of amplitude A the harmonics generated are negligible –both because they are significantly lower than the principal and because they are filtered out by the load–, the output voltage of the amplifier can be expressed as:

$$V_{out}(t) = V_{outDC} - G_m \cdot \left| Z_L|_{\omega_{in}} \right| \cdot A \cdot \cos\left(\omega_{in} t + \theta_{Z_L}|_{\omega_{in}}\right) \tag{1}$$

where $\left| Z_L|_{\omega_{in}} \right|$ and $\theta_{Z_L}|_{\omega_{in}}$ are the magnitude and phase, respectively, of the load impedance at input frequency ω_{in}.

The instantaneous power dissipated by the transconductor can be obtained as:

$$\begin{aligned} Pot_M(t) &= V_{out}(t) \cdot I_D(t) = \\ &\left(V_{outDC} - G_m \cdot \left| Z_L|_{\omega_{in}} \right| \cdot A \cdot \cos\left(\omega_{in} t + \theta_{Z_L}|_{\omega_{in}}\right)\right) \cdot \left(I_{bias} + G_m \cdot A \cdot \cos(\omega_{in} t)\right) \end{aligned} \tag{2}$$

whose DC component can be written as:

$$Pot_M|_{DC} = V_{outDC} \cdot I_{bias} - \frac{G_m^2 \cdot A^2 \cdot \left| Z_L|_{\omega_{in}} \right| \cdot \cos\left(\theta_{Z_L}|_{\omega_{in}}\right)}{2} \tag{3}$$

At the resonance frequency ω_0 the load impedance module is R and its phase is zero, i.e.:

$$\left| Z_L|_{\omega_0} \right| \cdot \cos\left(\theta_{Z_L}|_{\omega_0}\right) = R \tag{4}$$

while at the -ω_{3dB} frequency it can be shown that the load impedance module is $R/\sqrt{2}$ and its phase is $\pi/4$, i.e.:

$$\left| Z_L|_{-3dB} \right| \cdot \cos\left(\theta_{Z_L}|_{-3dB}\right) = \frac{R}{\sqrt{2}} \cdot \cos\left(\pi/4\right) = \frac{R}{2} \tag{5}$$

The first term in expression (3) is exclusively produced by the DC biasing and independent of the input signal, whereas the second term is the power delivered to the output load in response to the input. Therefore, variations in the RF output power will appear as variations of the total DC power dissipated. By sensing the temperature variations on the silicon surface, the 3 dB bandwidth of the PA can be obtained through the following steps:

1. A temperature measure with no input signal (A=0) is necessary to allow the calibration of the first term in expression (3).
2. A sweep of the input frequency ω_{in} for a given amplitude A is then performed in order to obtain:
   a. The central frequency ω_0, that corresponds to the input frequency at which the observed temperature variation is maximum.
   b. The 3 dB bandwidth ω_{-3dB}, measured at the frequencies at which the temperature variation has half the value (equation (5)) of that obtained at ω_0, once the calibration of step 1 is done.

III. Experimental Validation

A. Circuit Description

As a proof-of-concept validation a 60 GHz PA operating as class-AB has been used as circuit under test, in order to observe its frequency response. As shown in Fig. 1, the PA consists of three cascaded push-pull stages and was implemented in a 65 nm CMOS process. A differential temperature sensor was implemented in the same IC and placed close to the last stage of the PA in order to monitor its power dissipation. Details of the sensor circuit and the PA + sensor layout can be seen in [2, 3, 7]. Fig. 2 shows a detail of the temperature transducer of the sensor Q_1 placed next to the transistors in the 3rd stage of the PA. In the figure, the position of a diode connected NMOS transistor is as well highlighted ($Heat_1$) which is used in this experiment to generate heat to calibrate the temperature sensor.

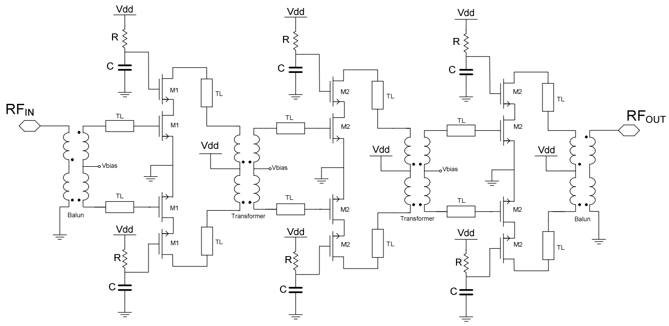

Fig. 1. Schematic of the 3-stage 60 GHz PA under test.

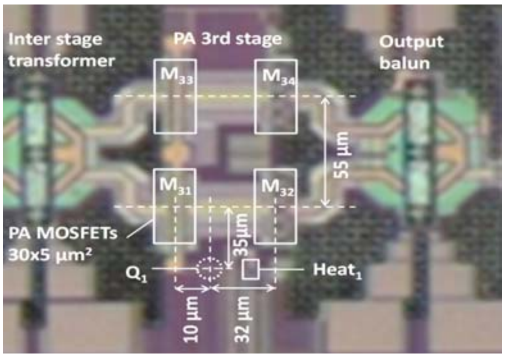

Fig. 2. Detail of the layout of the PA 3rd stage together with the thermal transducer (Q_1) and heat source for the sensor characterization.

Fig. 3 plots the measured characteristic response of the sensor DC output voltage vs. power dissipated by $Heat_1$ (PA powered off). When no power is dissipated by $Heat_1$ (i.e., sensor detects no temperature difference), the nominal output voltage is 780 mV, and then decreases as the power dissipated is increased with a maximum sensitivity of -55 mV/mW. Note that the sensor has offset-correction knobs, thus in practical PA measurements the curve will be shifted so as to place the sensor in the maximum sensitivity region [7].

B. *Homodyne Temperature Measurements*

The naked IC was accessed using multicontact probes for the power and DC bias, and Cascade GSG probes for accessing the input and output of the PA. The measurement instrumentation consisted simply of a mmW signal generator for the PA input, a multimeter connected to the temperature sensor output, and a spectrum analyzer to measure the PA output power simultaneously with the sensor output. The PA was initially powered to a supply voltage of 1.8 V, while the signal generator was disabled. The offset-correction knobs were then adjusted to place the sensor in the maximum sensitivity region, specifically at a 425 mV output. This value will be used as a reference, and thus corresponds to the DC power dissipation of the PA with no input signal (first term in expression (3)).

The PA is next excited by setting a -5 dBm tone at the generator, which corresponds to an effective input power of approximately -20.5 dBm (because of the losses of cables and probe). The fraction of the power dissipation due to the PA output signal will then manifest at the sensor output as variations respect to the 425 mV reference. As observed in equation (3), an increase in the output power will result in a decrease of the total dissipated power. Given the inverting characteristic of the sensor shown in Fig. 3, an increase in the sensor output voltage is in fact observed.

The frequency of the input signal is then swept from 57 GHz to 62 GHz in order to observe the central frequency and estimate the bandwidth of the PA. The thermal settling time for each measurement is below 1 ms [8], as only the temperature difference between two temperature transducers needs to be stable, not the absolute temperature value. Fig. 4 shows the measurements of the PA output power P_{out}, together with the sensor voltage increase, expressed as $10\log(\Delta V_{sensor})$. Note that a 10 factor is used because the sensor output voltage is directly proportional to dissipated power changes. A good matching between the curves is observed. Due to the frequency-dependent losses of the input set-up, the PA gain did not match exactly the P_{out} curve, thus gain is also represented in Fig. 4, in the same 1 dB/div scale.

According to the gain measurement, the central frequency of the PA response is 58.1 GHz, while the curve provided by the temperature sensor predicts 58 GHz, which in fact matches the P_{out} peak. The sensor predicts a 3 dB bandwidth of 1.65 GHz, while for the power gain curve the bandwidth is 1.6 GHz. According to measurements of the S21 parameter reported in [2], the central frequency is at 58.45 GHz and the bandwidth was 1.8 GHz. Note that a 500 MHz step was used in the measurements.

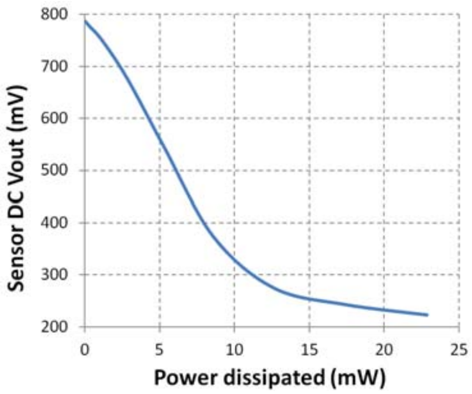

Fig. 3. Measured characteristic response of the sensor DC output voltage vs. power dissipated.

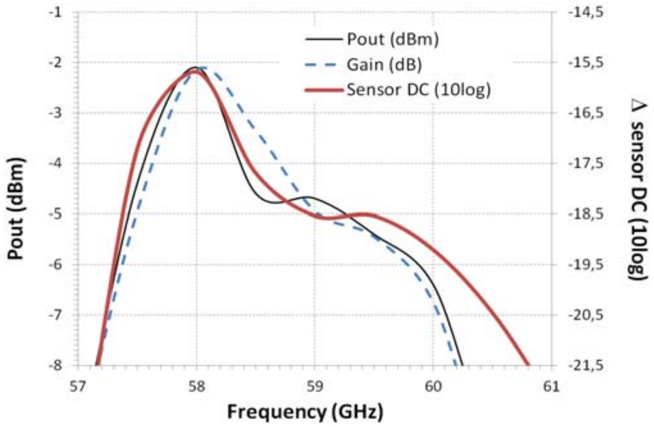

Fig. 4. Measured PA output power vs. frequency (left axis), together with the DC shift measured by the thermal sensor, in 10log scale (right axis). The measured PA gain is also represented, in the same 1 dB/div scale.

IV. Conclusions

Low-frequency temperature observation had already shown to provide a non-invasive methodology to monitor the frequency response in mmWave power amplifiers [2]. This letter shows an homodyne methodology in which the DC component of the on-chip temperature is monitored. Contrary to the previous methodology, it only requires a signal generator to feed the PA input, and a simple multimeter to monitor the DC output voltage of a temperature sensor. An experimental proof-of-concept demonstration has shown that, even with a challenging 60 GHz PA, the frequency response of the circuit (central frequency and bandwidth) can be obtained with the advantages of (1) low-frequency measurement with extremely simple set-up, (2) non-invasive measurement, which avoids any impact on the PA or any need to redesign it, and (3) negligible area overhead, making it suitable as a built-in tester.

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