Thermal Memory Effects Modeling and Compensation in RF Power Amplifiers and Predistortion Linearizers

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***Abstract*—Memory effects, which influence the performance of RF power amplifiers (PAs) and predistortion-based linearizers, become more significant and critical in designing these circuits as the modulation signal bandwidth and operation power increase. This paper reports on an attempt to investigate, model, and quantify the contributions of the electrical nonlinearity effects and the thermal memory effects to a PA's distortion generation, as well as how to compensate for these effects in designing baseband predistortion schemes. The first part of this paper reports on the development of an accurate dynamic expression of the instantaneous junction temperature as a function of the instantaneous dissipated power. This expression has been used in the construction of an electrothermal model for the PA. Parameters for the new proposed behavior model were determined from the PA measurements obtained under different excitation conditions (e.g., small-signal and pulsed RF tests). This study led us to conclude that the effects of the transistor self-heating phenomenon are more important under narrow-band signal (e.g., enhanced data for global evolution of global system for mobile communications) than for signals with wide modulation bandwidth (CDMA2000, Universal Mobile Telecommunications System). In the second part of this paper, the newly developed model has also been used to design a temperature-compensated predistortion function to compensate for these effects. The linearized PA output spectrum and error vector magnitude show a significant performance improvement in the temperature-compensated predistortion function over a memoryless predistortion. The results of these measurements that have been conducted on a 90-W peak lateral double-diffused metal–oxide–semiconductor PA are in agreement with those obtained from simulations using the developed PA and the predistorter models implemented in an ADS environment.**

***Index Terms*—Behavior model, distortion, electrothermal, measurements, memory effects, RF power amplifier (PA).**

I. INTRODUCTION

THE design of RF power amplifiers (PAs) for wireless base-station applications is becoming an increasingly complex task. These applications are characterized by a continuous growing bandwidth and envelope variation, and thereby impose stringent constraints on system designers to meet performance requirements in terms of power efficiency and linearity. The choice of transistor class of operation and load-matching circuits are important aspects of system design that can help overcome these challenges. Furthermore, several linearization methods, reported in the literature, have been considered as important approaches to increase power efficiency while keeping good linearity. However, the relatively limited performance of such RF PAs and linearizers [1], [2] is due to several factors that have not been taken into account during system design. The importance of the degradation factors may differ depending on RF PA technology (e.g., bipolar junction transistor (BJT), field-effect transistor (FET), ...), topology (e.g., switching, push–pull, balanced, single-ended amplifier) and the linearization method (e.g., predistortion, feedforward, ...) used. Considered a major source of degradation in the performance of PAs, memory effects have recently become an active area of investigation. A solid knowledge of the sources of memory effects and the development of accurate methods to measure and quantify their impact are critical steps in the design of high PAs and linearizers. In the literature [3] and [4], two categories of memory effects have been identified, i.e., electrical and electrothermal memory effects. The predominant factor that causes the electrical memory effect is the variation of terminal impedances (biasing and matching circuit's impedances) over the input signal bandwidth around the carrier frequency and its harmonics, as well as at baseband frequencies. Careful design of the matching and biasing circuit would minimize these effects, especially in the case of FET-based amplifiers. However, gain variation caused by transistor temperature-dependent electrical parameters leads to unavoidable electrothermal memory effects.

Memory effects have traditionally been defined as the measured PA distortion dependency on tone spacing under two-tone drive. In fact, it is assumed that, for a given value of two-tone spacing and wide-range amplitude, the PA behaves in a similar way to that of a signal with a modulation bandwidth equal to the two-tone spacing. It is seen as a practical way to get around the difficulties encountered while using other metrics such as adjacent channel power ratio (ACPR) and error vector magnitude (EVM) in case of band pass modulated signal.

A measurement system, which uses two network analyzers that are capable of providing information about the memory effects on the fundamental signals was proposed by Bosch and Gatti [4]. This system, however, does not account for their effects on intermodulation distortion (IMD) that are essential in the linearization case. Other systems [5] that are capable of measuring the relative phase of third-order intermodulation

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distortion (IMD3) have also been suggested. These systems are based on a reference nonlinearity generator (low-frequency MESFET). This should provide a constant phase IMD3 over a range of wide-band modulation frequencies. The main drawback of such systems is the dependency of their performance on the quality of the reference nonlinearity, which should be an ideal third-order distorter. Vuolevi *et al.* [3] proposed a measurement setup, suitable for a predistortion application, based on the envelope injection technique. This approach requires a tedious and lengthy calibration procedure, which calls for many specialized instruments such as three generators, two vector network analyzers, two spectrum analyzers, down-converters, etc.

As previously mentioned, memory effects are defined as distortion phase and amplitude changes over the modulation bandwidth. In the literature, one can distinguish two means for minimizing these effects at the design stage or by adding a compensation module. Some designers of RF PAs treat memory effects as a design stage optimization parameter. Consequently, previous efforts [6] have concentrated on the biasing circuit feed structure in order to minimize the drain biasing circuit impedance variation over the modulation bandwidth. As a result, better-balanced lower and upper intermodulation products of the output spectrum were observed. However, the scope of this study was limited to the study of memory effects that perturb the magnitudes of intermodulation products over the modulation bandwidth, while discarding their effect on the phases of the intermodulation products. In practice, this omission could strongly compromise the performance of any predistortion correction scheme based and designed using only magnitude information of the intermodulation products.

On the other hand, certain designers prefer measuring these effects in order to be able to synthesize a dedicated compensation module. The compensation module could be realized by adding a memory effects compensator to a memoryless predistortion polynomial function [7]. Vuolevi *et al.* [8] proposed an envelope injection method for memory effects pre-compensation. This method was used to compensate for terminal impedance variation versus the modulation frequency. Even though memory effects are related to the modulation frequency, they have an impact on the amplitude attributed to interaction with higher order nonlinearities. Simulation and measurement results reported in this study were limited to two-tone signals and third-order distortion [9]. Moreover, the envelope injection-based compensation method does not take into account the dynamic changes of the junction temperature, which cause electrothermal memory effects. Consequently, this compensation method should, in principle, be more appropriate to compensate for the electrical memory effects.

This paper reports on an attempt to thoroughly investigate, model, and quantify the contributions of the electrical nonlinearity effects and the electrothermal memory effects to a PA's distortion generation, as well as how to compensate for these effects in designing baseband predistortion linearizers. In particular, the development of an accurate expression of the junction temperature as a function of the instantaneous dissipated power and the input signal level for PAs is proposed. This expression is described in the first part of this paper and it will

TABLE I
THERMAL AND ELECTRICAL QUANTITIES EQUIVALENCE

	THERMAL QUANTITY		ELECTRICAL QUANTITY
P_{dissip}	Power heat flow (W)	I	Current flow (A)
T_j	Temperature (K)	V	Voltage (V)
R_{th}	Thermal resistance (K/W)	R	Electrical resistance (Ω)
C_{th}	Thermal capacitance (J/K)	C	Electrical capacitance (F)

subsequently be used in the construction of an electrothermal model suitable for RF PAs. Following this, a brief discussion is given to explain how the electrothermal behavior of the amplifier influences the generation of the IMD for different types of drive signals such as two-tones, wide-band code division multiple access (WCDMA), and enhanced data for global evolution of global system for mobile communications (EDGE GSM), etc.

Based on the developed PA electrothermal model and in order to compensate for the memory effects on the output spectrum, a temperature-compensated memoryless predistorter is developed and reported in the second part of this paper.

II. ELECTROTHERMAL MODEL

The electrothermal model of the PA requires knowledge of the junction temperature of the transistor as a first step. Vuolevi *et al.* [3] expressed the junction temperature of the transistor in the form described in (1) in order to explain its dynamic changes with the drive signal. Equation (1) utilizes the following as the thermal impedance at the envelope frequencies and as the dissipated power at these frequencies:

$$T_j = T_{\mathrm{amb}} + R_{\mathrm{th}} * P_{\mathrm{dissip}}(\mathrm{DC}) + Z_{\mathrm{th}}(w_1 - w_2) * P_{\mathrm{dissip}}(w_1 - w_2). \quad (1)$$

The instantaneous dissipated power determines the instantaneous rate of heat that is applied to the transistor. Furthermore, due to the finite mass of the component, thermal impedance includes a capacitive part in addition to the resistive one. Thermal resistance describes just the steady-state behavior, and thermal capacitance is essential for the description of the dynamic behavior. Thermal resistances and capacitances together lead to exponential rise and fall times characterized by a thermal *RC* time constant, similar to the electrical *RC* constant. The expression of the instantaneous junction temperature of the transistor was developed by making use of the existing duality [10] between heat transfer and electrical phenomena summarized in Table I.

Fig. 1 shows the thermal network models of the transistor including the silicon chip, package, and heat sink relating the junction and ambient temperatures to the dissipated heat amount. Since thermal constant $R_{\mathrm{th,heat\,sin}\,k} \times C_{\mathrm{th,heat\,sin}\,k}$ is too large when compared to $R_{\mathrm{th}} \times C_{\mathrm{th}}$, Fig. 1(a) was simplified to (b).

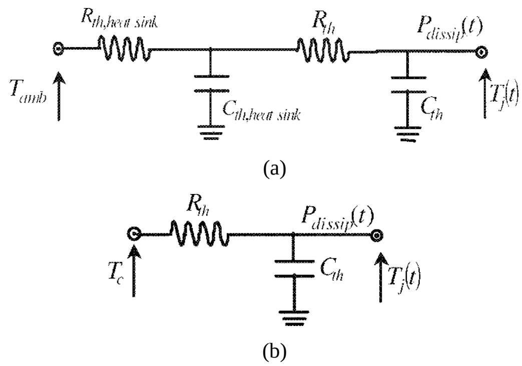

Fig. 1. Transistor thermal model.

Based on Fig. 1(b), the instantaneous temperature can be expressed as a solution to the following first-order nonhomogeneous differential equation:

$$\frac{\partial T_j(t)}{\partial t}+\frac{1}{R_{\rm th}C_{\rm th}}T_j(t)=\frac{1}{R_{\rm th}C_{\rm th}}\left(R_{\rm th}P_{\rm dissip}(t)+T_c\right) \quad (2)$$

where

$$P_{\rm dissip}(t)=V_{\rm DS,dc}\times I_{\rm DS,dc}(t)+P_{\rm RF,in}(t)-P_{\rm RF,out}(t) \quad (3)$$

or

$$P_{\rm dissip}(t)=\left(1-\eta(t)\right)\times P_{\rm RF,out}(t) \quad (4)$$

and $\eta(t)=$ instantaneous power efficiency.

Equation (2) has the form

$$\frac{\partial}{\partial t}T_j(t)+a(t)T_j(t)=b(t) \quad (5)$$

where

$$a(t)=\frac{1}{R_{\rm th}C_{\rm th}}$$

and

$$b(t)=\frac{1}{R_{\rm th}C_{\rm th}}\left(R_{\rm th}P_{\rm dissip}(t)+T_c\right).$$

The general solution of (5) has the following form:

$$T_j(t)=e^{\left(-\int a(t)dt\right)}\left(\int e^{\left(\int a(t)dt\right)}b(t)\,dt+K\right). \quad (6)$$

This equation is equivalent to

$$T_j(t)=e^{-t/\tau}\left(\int\frac{1}{\tau}e^{(t/\tau)}\left(R_{\rm th}P_{\rm dissip}(t)+T_c\right)dt+K\right) \quad (7)$$

where $R_{\rm th}C_{\rm th}=\tau$ is the thermal time constant.

The integral on the right-hand-side expression of (7) can be rewritten as follows:

$$T_j(T)=e^{1/\tau}\left\{\int\frac{\partial e^{(1/\tau)}\left(R_{\rm th}P_{\rm dissip}(T)+T_c\right)}{\partial t}dt -\int R_{\rm th}e^{(1/\tau)}\frac{\partial p_{\rm dissip}(t)}{\partial t}dt+K\right\}. \quad (8)$$

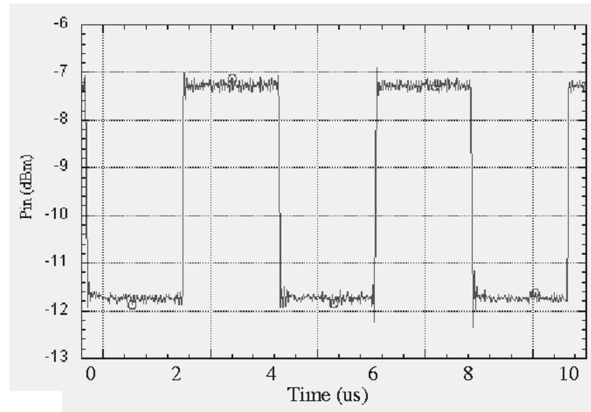

Fig. 2. Input pulsed signal envelope (period 4 μs, duty cycle $=0.5$).

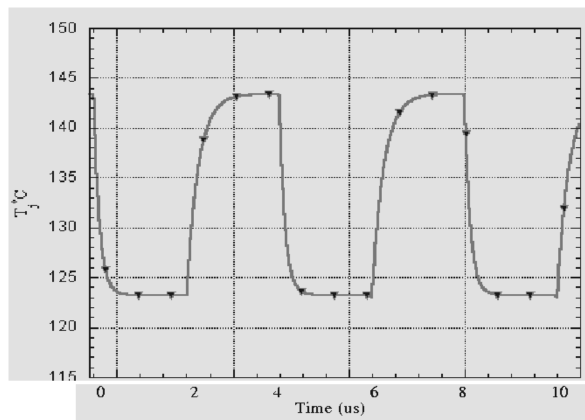

Fig. 3. Junction temperature variation versus time for pulsed signal (period 4 μs, duty cycle $=0.5$).

In the particular case of a step input signal excitation, the instantaneous power is constant; hence, the instantaneous dissipated power also remains constant. Therefore, one can write

$$P_{\rm dissip}(t)=\begin{cases}P & t_0\le t\le T\\ P_0 & t<t_0\end{cases};\Rightarrow\frac{\partial P_{\rm dissip}(T)}{\partial t}=0 \quad (9)$$

with $\tau<>T$.

In such a case, (8) becomes

$$T_j(t)=T_{j,s}+(T_{j,0}-T_{j,s})\times e^{-\Delta t/\tau} \quad (10)$$

where $T_{j,0}=T_c+R_{\rm th}P_0$ and $T_{j,s}=T_c+R_{\rm th}P$.

The time variation of the junction temperature according to (1) for the pulsed signal, shown in Fig. 2, is given in Fig. 3.

III. Model Identification and Validation

The expression of the instantaneous junction temperature defined in Section II is used to complete the free-self-heating behavior model of the PA [1], as shown in Fig. 4. In this study, an LDMOS 90-W peak PA is used. Several experiments were conducted on this amplifier in order to characterize its thermal behavior. Figs. 5 and 6 show the measured gain compression, phase, and amplitude versus junction temperature. The amplifier under test was operated in this stage of measurements in its small-signal region. Consequently, the measured gain compression results only from the junction temperature variation and not from the electrical nonlinearity of the amplifier. These curves are

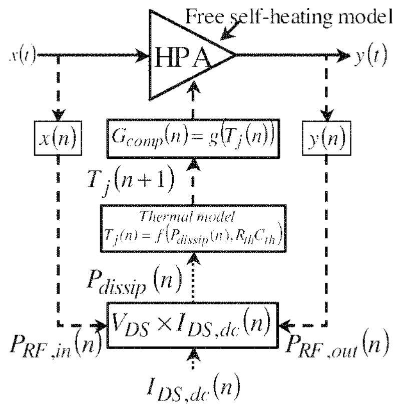

Fig. 4. Dynamic temperature calculation block diagram.

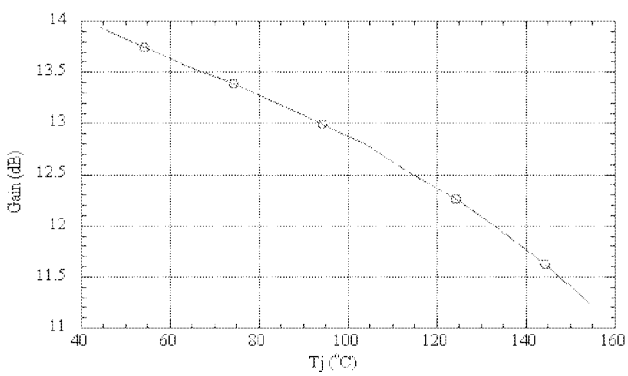

Fig. 5. Measured gain compression versus junction temperature.

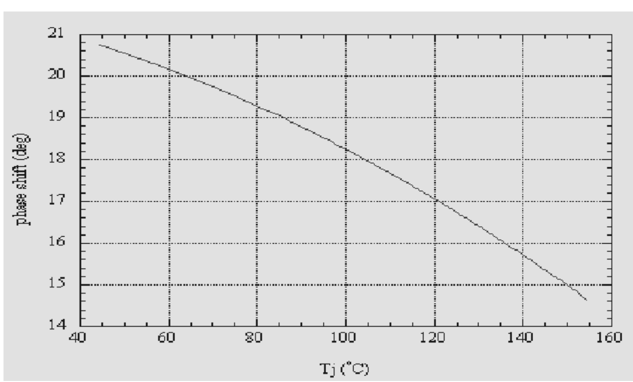

Fig. 6. Measured phase compression versus junction temperature.

used in the behavior model to determine the complex gain compression corresponding to each input signal level as a function of its corresponding junction temperature. Fig. 7 shows the pulsed measurement results of the drain current versus the input signal for different case temperatures. This curve is used in the amplifier model given in Fig. 4 for the calculation of instantaneous dissipated power, which, in turn, will serve to calculate the instantaneous junction temperature. Fig. 8 shows the simulated and measured output waveforms of the PA under a 50% duty cycle square input signal. This figure also clearly shows the exponential fall and rise of the signal as a response to the junction temperature exponential variation predicted by (10). Moreover, the agreement between the simulated and measured waveforms in Fig. 8 demonstrates the accuracy of the developed model to predict the electrothermal behavior of the tested amplifier.

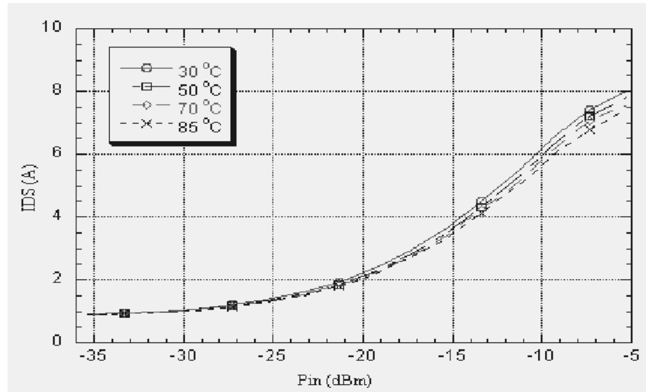

Fig. 7. Pulsed measurements of the drain current versus drive.

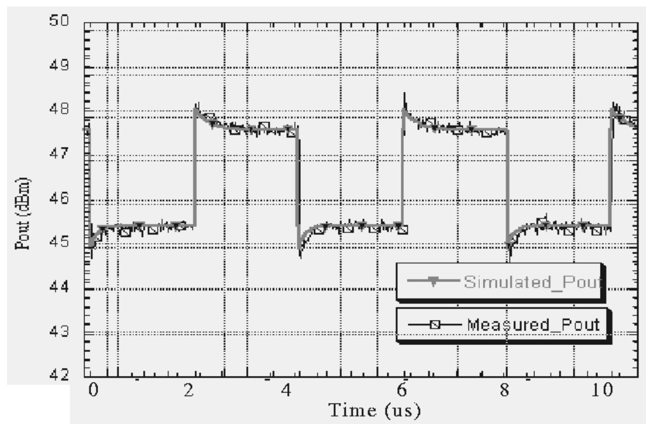

Fig. 8. Measured and simulated output envelope waveforms (period 4 μs, duty cycle $= 0.5$).

IV. Discussion

The temperature changes have repercussions on electrical parameters of the transistor that influence the amplifier's behavior. Instantaneous transistor junction temperature variation depends on instantaneous dissipated power and, consequently, on the input signal time variation. As a result, the dependency of the complex gain of high PAs (amplitude and phase shifting) on junction temperature can be considered as a source of nonlinearity. To demonstrate and quantify these effects, the actual PA was substituted with an ideal amplifier (perfect limiter) in the simulations. Fig. 9 shows the simulation results of the IMD3 products generated only by the thermal effects for different frequency spacing. One can clearly observe the decrease of the IMD3 product level as the frequency spacing increases until they reach a constant level where the thermal effects has no more significance on the IMD3. In addition, it is well established that the level of IMD3 generated by a pure electrical memoryless nonlinearity is independent from the two-tone frequency spacing. The combination of both effects explains the dependency, often seen and reported [1], [3], [6] of the IMD3 level on the two-tone frequency spacing interval.

This mechanism provides an explanation for the dependency of IMD levels (third and higher orders) on the variation in tone spacing. Thus, the IMD vectors emerging from thermal and electrical nonlinearity sources result in a varying IMD vector that depends on frequency tones spacing, as well as on the transistor average thermal condition. Fig. 10 supports the above

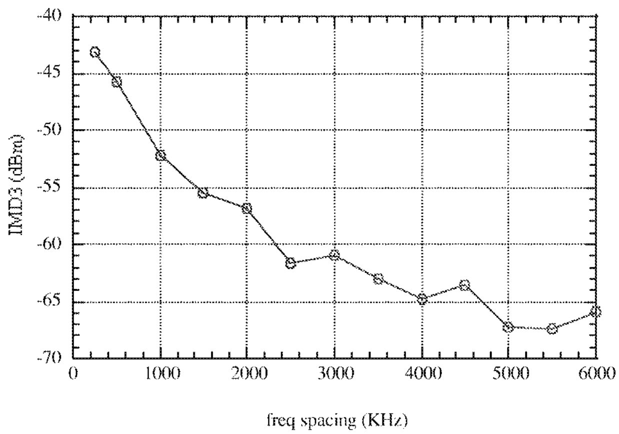

Fig. 9. IMD3 amplitude due to thermal effects versus frequency spacing.

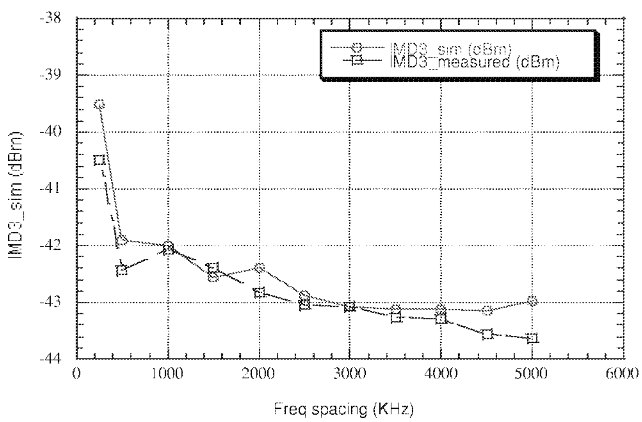

Fig. 10. Measured and simulated IMD3 amplitude versus frequency spacing.

statement; it shows the simulated and measured IMD3 level versus the frequency spacing of a two-tone signal at the output of the 90-W peak lateral double-diffused metal–oxide–semiconductor (LDMOS) PA resulting from the combination of the electrical and electrothermal nonlinearity sources.

The above results lead to the conclusion that the behavior of the PA under high varying amplitude and wide-band signals (high-speed signals) may be considered as free of self-heating effects. Indeed, instantaneous junction temperature will not experience large fluctuations and, therefore, the influence of electrothermal nonlinearity sources will be minimized. Furthermore, the instantaneous temperature will be close to that fixed by input average power and, consequently, to the average dissipated power. Accordingly, electrical memory effects may well be considered as the main source of memory effects in PAs driven by wide-band and high varying signals, such as multicarrier WCDMAs. However, in the case of narrow-band signals (e.g., global system for mobile communications (GSM) signals) with a channel separation as small as 25 kHz, the PA dynamic electrothermal behavior should be well taken into account to assure good performance.

V. Self-Heating Effects Compensator

The previous electrothermal model of the PA is used in Section V to develop a temperature-compensated predistortion function in order to compensate for self-heating effects. Fig. 11 shows the scheme of the amplifier model including the dynamic temperature-compensation circuit where the functions f and g represent the memoryless nonlinear complex gain (AM/AM and AM/PM) function and the electrothermal model function of the PA, respectively.

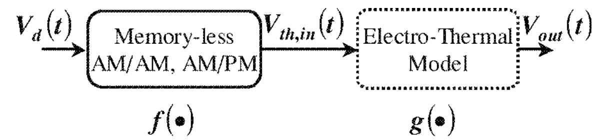

Fig. 11. PA model scheme.

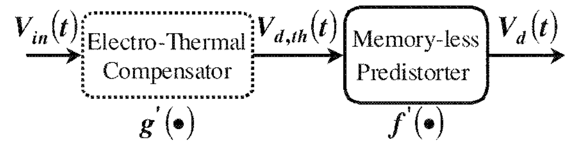

Fig. 12. Temperature-compensated predistorter scheme.

One can easily express the PA's electrothermal sub-model (see Fig. 11) output voltage as function of its input as follows:

$$V_{\text{out}} = g\left(V_{\text{th,in}}\right) * V_{\text{th,in}} \tag{11}$$

with $V_{\text{th,in}} = f(|V_d|)*V_d$, where V_d is the PA's input voltage.

Fig. 12 illustrates the scheme of the temperature-compensated predistortion function used to compensate for both nonlinear distortion and electrothermal memory effects of the PA modeled in Fig. 11. f' and g' represent the inverse function of f and g, respectively.

To obtain a linearized PA over a wide temperature range, the cascade of the amplifier and its temperature-compensated predistorters must satisfy the following constraints:

$$\begin{aligned} f(|V_d|)*V_d(t) &= f\left(\left|f'(|v_{d,\text{th}}|)*\right.\right.\\ f'(|V_{d,\text{th}}|)*V_{d,\text{th}} &= K \end{aligned} \tag{12}$$

and

$$\begin{aligned} g(V_{\text{th,in}})*V_{\text{th,in}} &= g(v_{d,\text{th}})*V_{d,\text{th}} \\ &= g(g'(v_{\text{in}})*V_{\text{in}})*g'(V_{\text{in}})*V_{\text{in}} \\ &= 1 \end{aligned} \tag{13}$$

where K represents the small-signal linear gain of the PA.

V_{in} and $V_{d,\text{th}}$ represent the input signals of the electrothermal compensator and memoryless predistorters, respectively.

In order to correct for the instantaneous compression due to the thermal effect in the PA, one has to predict the instantaneous junction temperature of the transistor using the *RC* equivalent thermal network, given in Fig. 1. The complex-gain expansion value of g', corresponding to the predicted junction temperature, is computed as the inverse of the complex gain compression value of g for the same junction temperature. Fig. 13 shows the details of the thermal effect compensation function g'.

In order to evaluate the contribution of the self-heating effect nonlinearity compensation on the output signal quality, we have carried out simulations using an EDGE-GSM signal. This signal was synthesized using the EDGE library of ADS. The synthesized EDGE signal uses the $3\pi/8$ eight phase–shift keying (8PSK) modulation technique with a symbol rate equal to 270.833 kHz and a Gaussian shaping filter frame. Fig. 14

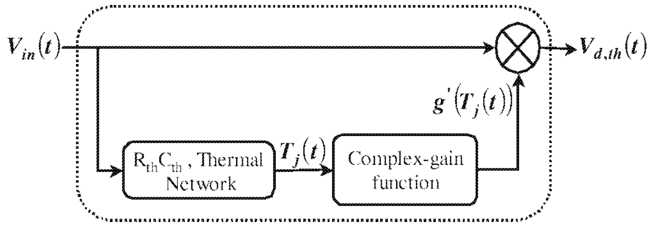

Fig. 13. Thermal effect compensation function.

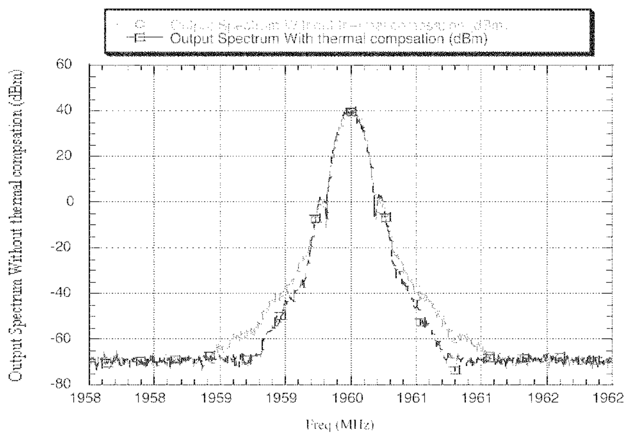

Fig. 14. Linearized output spectrum with and without thermal effect compensation under an EDGE-GSM signal.

includes the output spectra of the PA when linearized using a memoryless digital predistorter before and after adding the self-heating compensator. One can clearly conclude that the temperature-compensated predistorter allows better reduction of the out-of-band emission at the output of the linearized amplifier. Moreover, the proposed self-heating compensation function could be easily implemented using either a digital signal processor (DSP) or an field-programmable gate array (FPGA) module.

To further highlight the contribution of the thermal compensation circuit in high-speed applications, simulations using the EDGE standard were carried out. The EDGE standard uses both amplitude and phase modulation, which is translated into more stringent linearity requirements for the PA. EVM is a critical parameter used to measure the transmitter performance. In this context, the EVM was evaluated as the vector difference, in time, between the simulated and expected carrier magnitude and phase at the output of the amplifier. The computed EVM explicitly shows the contribution of the amplifier nonlinearity to the overall EVM of the transmitter. The evaluation of the EVM values was done for a peak input power equal to -10.8 dBm, which corresponds to a peak output power of 49.1 dBm, in the following cases.

- Case 1) PA without predistortion.
- Case 2) PA with a memoryless predistorter.
- Case 3) PA with temperature-compensated predistortion with the self-heating compensator.

Table II summarizes the rms EVM values (in percent) obtained for the three previous cases. It can be concluded that the self-heating compensation function reduces the EVM introduced by the PA and, hence, enhance the quality of the signal at the output of the transmitters.

TABLE II
EVM (IN PERCENT) OF THE PA WITHOUT PREDISTORTION, WITH MEMORYLESS PREDISTORTION, AND WITH TEMPERATURE-COMPENSATED PREDISTORTION

	EVM (%)
Without pre-distortion	5.95
Memory-less pre-distortion	2.5
Temperature compensated pre-distortion	0.2

VI. CONCLUSION

In this paper, a dynamic electrothermal behavior model for PAs has been proposed along with its parameter identification procedure. Implementation of the model within an ADS simulator and its validation have been carried out for a 90-W peak-power LDMOS amplifier. Satisfactory results have been obtained for pulsed signals. This study is an attempt to deeply investigate and model the electrical and electrothermal effects' respective contribution to the PAs' distortion generation. This has led to the decoupling and quantification of both effects and has allowed us to conclude that the effects of the transistor's self-heating phenomenon are more important under narrow-band signals (e.g., EDGE GSM) than under signals with wide modulation bandwidths [e.g., multicarrier third-generation (3G) CDMA2000, Universal Mobile Telecommunications System (UMTS)]. The latter types of signals are more sensitive and influenced by electrical-induced memory effects than are thermal memory effects.

The newly developed model has been used to design a temperature-compensated predistortion function to compensate for the self-heating effects. The linearized PA output spectrum and EVM clearly show the great improvement of the temperature-compensated predistortion function performance over a memoryless predistortion.

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