An Active Pulsed RF and Pulsed DC Load–Pull System for the Characterization of HBT Power Amplifiers Used in Coherent Radar and Communication Systems

Caroline Arnaud, Denis Barataud, Jean-Michel Nebus, Jean-Pierre Teyssier, Jean-Pierre Villotte, and Didier Floriot

***Abstract*—This paper presents a new automated and vector error-corrected active load–pull system allowing the characterization of microwave power transistors under coherent pulsed RF and pulsed dc operating conditions. In this paper, the use of this system is focused on the characterization of a 240-μm^2 GaInP/GaAs heterojunction bipolar transistor (HBT) (Thomson CSF-LCR, Orsay, France). On one hand, source and load–pull measurements of such a transistor are reported for different pulsewidths. On the other hand, nonlinear simulations based on an electrothermal model of an HBT have been performed and are compared with experiments. Power variations and RF carrier phase shift within the pulse versus input power and junction temperature of the transistor are shown.**

***Index Terms*—Electrothermal HBT model, pulsed load–pull measurements, RF carrier phase shift.**

I. INTRODUCTION

BOTH MODERN nonlinear computer-aided design (CAD) techniques (such as the envelope transient technique) and improved nonlinear transistor models, taking into account low-frequency dispersive effects (self-heating or trapping effects), require validation based on appropriate large-signal measurements [1], [2]. If well-suited large-signal measurements of elementary amplifying cells are performed early in the design process of a power amplifier, it gives a maximum chance to design an optimized power amplifier.

In such a context, a pulsed load–pull system reveals to be a very useful tool [3]. The fact that pulsed RF and pulsed dc capabilities are combined in a general active load–pull system, like the one presented in this paper, enforces the possibility of a deep insight into the influence of low-frequency dispersive effects on power performances of transistors. Such a system is an efficient tool providing valuable information like design rules concerning the optimum combination of amplifying cells with embedding circuits (biasing circuits—RF matching circuits) in order to minimize nonlinear memory effects.

Obviously, this can be the case to minimize the RF carrier phase shift within the output RF pulse of power amplifiers used for coherent radar applications. Investigations on power amplifiers dedicated to time-division multiple-access (TDMA) applications represent another potential aspect of our measurement setup.

Our pulsed dc and pulsed RF measurement system is based on the use of a Wiltron pulsed vector network analyzer (VNA) and a two-channel HP 8115A dc pulsed generator. An overall system synchronization is achieved and allows RF and dc pulsewidths down to 250 ns. A specific analysis window (down to 50 ns) can be swept within the RF pulse for vector measurements of the device-under-test (DUT). Variable load impedance at the fundamental frequency is monitored using the active loop principle [4]. Source impedance of the DUT can be controlled by means of a mechanical tuner.

GaInP/GaAs heterojunction bipolar transistors (HBTs) including a thermal shunt and a proprietary ledge [5] offer very good potentialities for power amplification due to their high current density. This paper focuses on the characterization of these devices for pulsed RF operation.

II. SYSTEM DESCRIPTION

The heart of the measurement system is a Wiltron pulsed VNA. This system, which uses a vector receiver principle, allows phase and magnitude measurements of power wave ratios at each frequency of the $\sin x/x$ signal spectrum obtained under pulsed conditions. A two-channel HP 8115A pulse generator is used to pulse the bias of transistors under test. The pulsed bias and pulsed RF subsystems are synchronized by a 10-MHz signal. Furthermore, four identical narrow pulse modulators (profile modulators) connected on measurement channels allow to move a measurement window, which can be swept within the RF pulse stimulus. The measurement principle is sketched in Fig. 1. A symbolic representation of signals at different ports is depicted Fig. 2.

These pulsed RF and pulsed bias subsystems are coupled to a load–pull setup, as shown in Fig. 3. Load impedance at the fundamental frequency is synthesized thanks to an active loop. Source impedances can be tuned by using a mechanical tuner. Linear 20-W power amplifiers are used to drive the input of the transistor and to synthesize load impedances. This system covers L- and S-bands and allows the characterization of up to 4-W unmatched transistors. A scope allows the measurement of pulsed dc currents and voltages.

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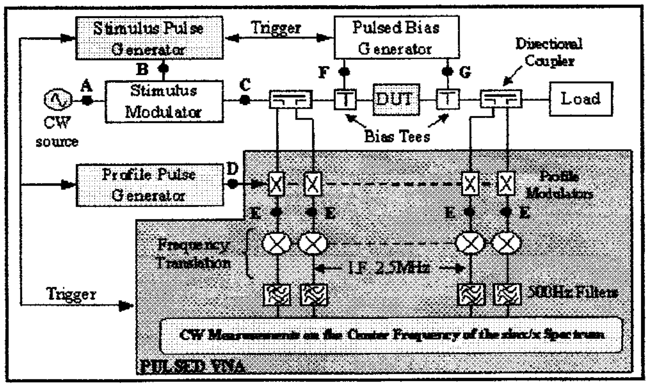

Fig. 1. Pulsed RF and pulsed bias environment.

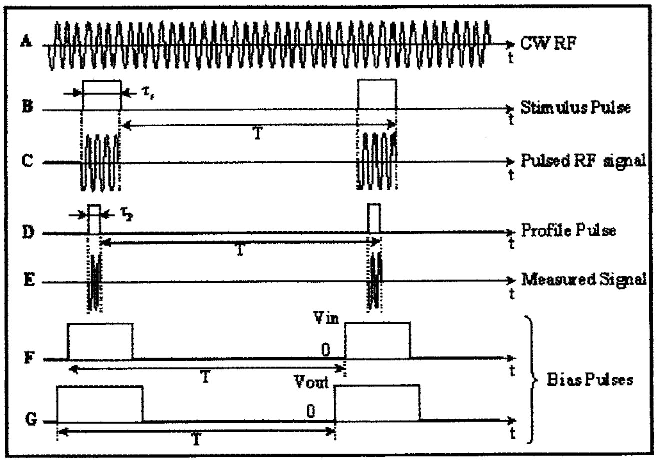

Fig. 2. Pulsed RF and pulsed bias signal representation.

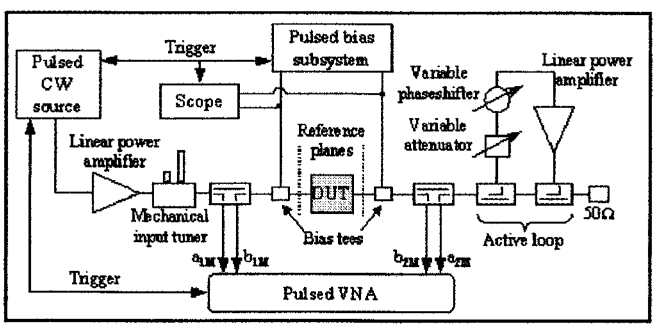

Fig. 3. Block diagram of the active pulsed load–pull system.

III. Measurement Principle and System Calibration

A. Measurement Principle

The characterization principle is based on the measurement of power wave ratios at the center frequency of the $\sin x/x$ spectrum obtained under pulsed conditions. For that purpose, the vector receiver operation mode of the network analyzer is used. Therefore, when a profile (acquisition) window width τ_p (in nanoseconds) and a repetition rate T are fixed, the absolute power of the spectrum center frequency of each power wave a_{1M}, a_{2M}, b_{1M}, and b_{2M} can be measured using the selective power meter capability of the VNA.

Let us take this power P_1 for the reference incoming wave a_1. The corresponding RF pulse power P_{1p} is simply calculated using the relationship

$$P_{1p} = P_1 \left(\frac{T}{\tau_p}\right)^2.$$

P_1 is measured and, consequently, P_{1p} is calculated for various delays of the profile window within the RF pulse. In a similar manner, power wave ratios at the center frequency (RF carrier frequency)

$$\left.\frac{\mathbf{b_i}}{\mathbf{a_i}}\right|_{\mathbf{M}} (\mathbf{f_o})$$

are measured for various positions of the profile window within the RF pulse, thus allowing the determination of the shape of the carrier power and phase shift within the RF pulse. In a sense, this looks like an experimental way of performing the envelope transient technique limited here at the fundamental frequency.

B. System Calibration

A conventional thru-reflection line (TRL) calibration routine is performed and applied to the correction of measured power wave ratios

$$\left.\frac{\mathbf{b_i}}{\mathbf{a_i}}\right|_{\mathbf{M}}$$

at the RF carrier frequency (center frequency). For this purpose, the DUT is removed and classical calibration standards are connected to the DUT reference planes. For the absolute-power calibration procedure, a power meter is connected to each reference plane. The power difference between the peak power measurements and the power at the center frequency of the spectrum of the incident power waves a_{1M} and a_{2M} measured with the VNA is recorded. This power difference is then used to correct raw data of the VNA in order to compute RF carrier powers at the DUT reference planes during the measurement procedure.

IV. Measurement Results

First, a class-C MACOM 8-W silicon bipolar matched amplifier was measured at 2.7 GHz. In this case, dc biases were not pulsed. The base–emitter voltage was set to zero (short-circuit) and the collector–emitter voltage was fixed to 36 V and kept constant.

The goal was to demonstrate the capability of our system to accurately determine RF power variations and RF carrier phase shift within the pulse, and the sensitivity of these parameters, when the load impedance is pulled away from 50 Ω.

The RF pulse was set to 200 μs and the repetition rate to 2 ms (duty cycle of 10%). A 20-μs acquisition window width was fixed and different measurements were performed when the position of this window is moved from the beginning to the end of the 200-μs RF pulse. Input/output power characteristics and input/output carrier phase shift of this amplifier are given in Fig. 4.

Another kind of measurement results concerns an $8 \times 30\ \mu\text{m}^2$ GaInP/GaAs HBT (Thomson foundry, Orsay, France) recorded

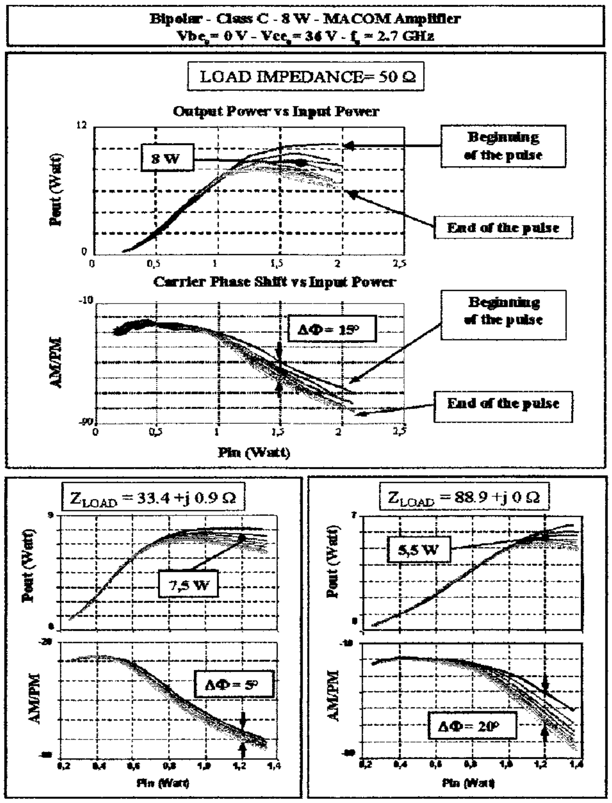

Fig. 4. Output power and RF carrier phase shift measurements of an 8-W amplifier.

at 2 GHz for different pulsewidths. In that case, both pulsed RF and pulsed dc conditions were achieved.

Fig. 5 shows measurement results obtained for the following conditions: carrier frequency: 2 GHz, pulse width: 100 μs, duty cycle: 10%, acquisition window width: 10 μs, optimum load impedance for maximum added power, source impedance: 50 Ω, bias conditions: pulsed base–emitter voltage from 0 to 1.6 V with a 8-Ω series resistor (self biasing class), constant collector–emitter voltage: 8 V.

A maximum 2° RF carrier phase shift is observed within the pulse. In fact, as HBT is matched for maximum power-added efficiency (PAE), the temperature variation within the pulse is limited. Therefore, the variations of nonlinear capacitances of the transistor versus temperature, which are responsible of RF carrier phase shift, are also limited.

Similar measurements were performed when the input of the transistor was matched using a mechanical tuner (Fig. 6). In this case, a significant RF carrier phase shift versus the input power was observed. This clearly indicates a nonlinear resonance phenomena at the input of the transistor due to a strongly nonlinear behavior of the base–emitter capacitance.

Other measurements were performed at 2 GHz in the following conditions: Pulse width: 5 μs, duty cycle: 20 μs, acquisition window width: 200 ns, pulsed base–emitter voltage from 0 to 1.6 V, constant dc collector–emitter voltage: 8 V. With a 5-μs stimulus pulse and a 200-ns acquisition window, we expected to finely see the self heating of the transistor (short thermal time constant). With a long 20-μs repetition rate, we let the transistor

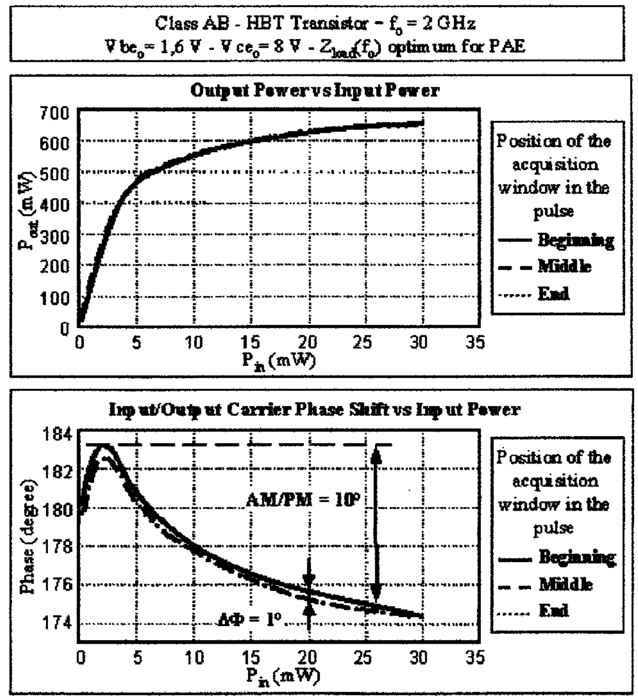

Fig. 5. 240-μm^2 GaInP/GaAs HBT measurements (source impedance of 50 Ω, load impedance $(79 + j28)$ Ω).

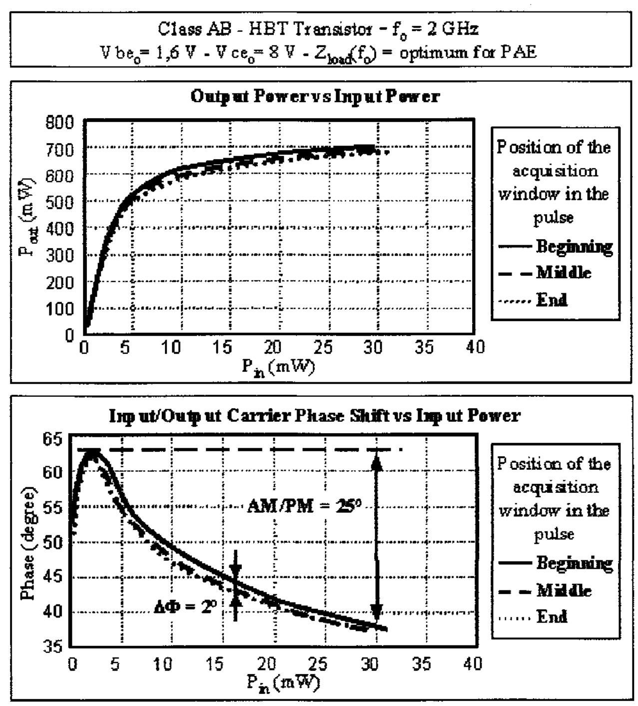

Fig. 6. 240-μm^2 GaInP/GaAs HBT measurements (RF input matched impedance, load impedance $(79 + j28)$ Ω).

come back to a cold thermal state from the end of a pulse to the beginning of the next pulse (longer thermal time constant). Therefore, during successive pulses, the transistor is in the same thermal state. This is important for the case of pulse-to-pulse coherent radar applications. Measurement results are given Fig. 7.

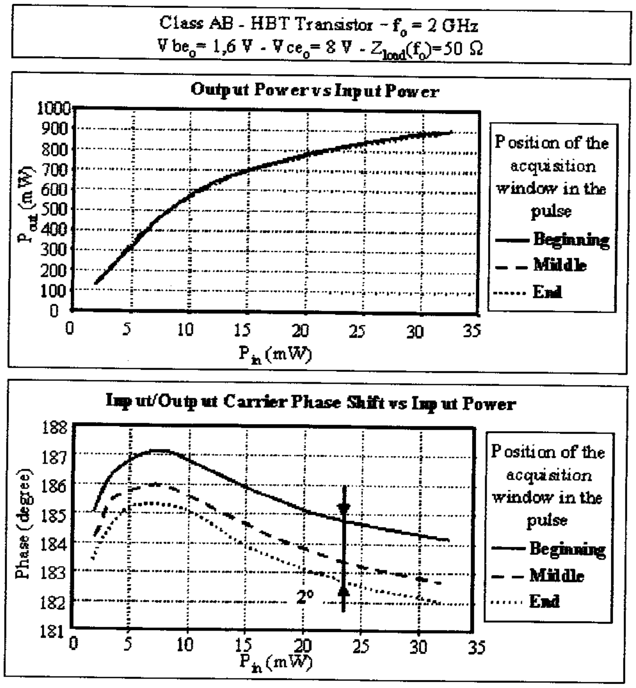

Fig. 7. 240-μm^2 GaInP/GaAs HBT measurements (5-μs pulsewidth, 10% duty cycle, 200-ns acquisition window).

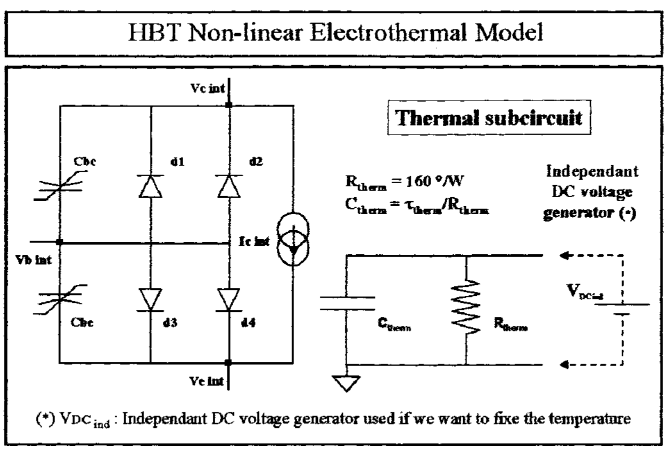

Fig. 8. Electrothermal HBT model.

No significant RF carrier phase shift is recorded (2° maximum). It could indicate that the major amount of self-heating effects stands in the first 200 ns of the pulse.

V. Simulations

To compare measurement results with simulations, a nonlinear electrothermal HBT model was used. This model was extracted from pulsed *I/V* and pulsed S-parameters measurements performed with an on-wafer thermal measurement system [2].

The nonlinear model topology is given in Fig. 8.

The thermal resistance was found to be equal to 160 °C/W. The thermal time constant is taken to 1 μs. Nonlinear continuous wave (CW) harmonic balance (HB) analysis was performed for different junction temperatures (from 25 °C to 150 °C). These temperatures can be imposed to the thermal subcircuit with an external independent voltage source connected to this subcircuit.

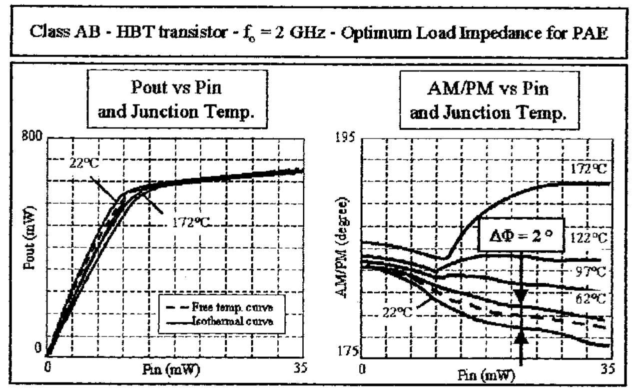

Fig. 9. HBT simulation results (50-Ω source impedance and load impedance of $(80 + j26.3)$ Ω).

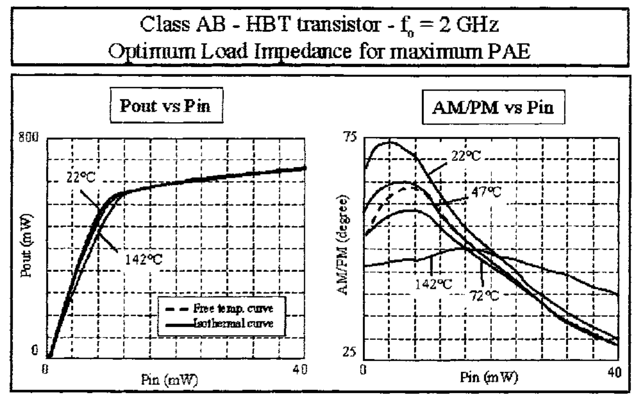

Fig. 10. Input matched HBT simulation results (RF input matched, load impedance of $(80 + j26.3)$ Ω).

This quasi-static approach allows to determine the range of minimum/maximum RF carrier phase shift versus temperature for any input RF power of interest. Simulations results performed at 2 GHz are shown in Fig. 9. In this case, the transistor is optimally loaded for maximum PAE, and the RF input is not matched (a 50-Ω generator is directly connected to the transistor).

The doted line shown on the AM/PM graph indicates the RF carrier phase shift versus the RF input level obtained in the case of self-heating effect of the transistor. Other lines indicate phase variations that would occur if the temperature was kept constant (isothermal curves). At any fixed input RF power, a vertical line allows to see the minimum/maximum RF carrier phase shift versus temperature, thus giving an idea of phase shift within the pulse (cold temperature at the beginning of the pulse and hot temperature at the end of the pulse).

The 2° RF carrier phase shift measurements shown in Section IV are predicted here by the simulations. Fig. 10 shows the same kind of simulations obtained when the RF input of the transistor is matched.

The important carrier phase shift versus input RF power is shown (nonlinear input RF resonance). RF carrier phase-shift range versus junction temperature at any input level can be also quantified.

VI. Conclusion

An efficient experimental tool allowing the characterization of power transistors under pulsed dc and pulsed RF conditions has been presented. A first main important application lies in the characterization and optimization of operating conditions of power amplifiers for coherent radar applications. A second interesting point concerns the validation of nonlinear models of transistors, taking into account low-frequency dispersive effects (thermal and trapping effects). Comparison between our pulsed measurements and simulations using an electrothermal model of an HBT [6] has been presented. In-depth investigations by simulations on dynamical carrier phase shift using the envelope transient technique is the next step under study. For that, a refinement of the thermal subcircuit of the HBT model is needed and should be validated by our narrow RF pulse measurement system.

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Didier Floriot, photograph and biography not available at time of publication.