INVESTIGATIONS OF LINEARITY CHARACTERISTICS FOR LARGE-EMITTER AREA GaAs HBT POWER STAGES

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ABSTRACT

In this paper the linearity properties of large-emitter-area GaAs heterojunction bipolar transistors are experimentally investigated. The approach is based on AM-AM and AM-PM conversion measurements, performed on-wafer with an active harmonic load-pull system. The measured data are then processed to evaluate the transistor linearity in terms of intermodulation distortion (IMD) and adjacent channel power ratio (ACPR). As an important result, the harmonic source impedance at the second harmonic plays a significant role in determining the maximum output power allowed for a given IMD or ACPR level.

INTRODUCTION

Power amplifier stages are critical elements in mobile telephones, since they have a significant influence on talktime, size and cost. GaAs heterojunction bipolar transistors (HBTs) represent a promising solution. They have good high-frequency performance in conjunction with high breakdown voltage, achievable already with relaxed 3 μm lithography. In addition, HBTs allows high linearity at low voltage operation, high power density (leading to small chip size) and single voltage supply operation. Last but not least, GaAs technology offers some advantages with respect to Si, in terms of low parasitics and high-Q passive elements for matching.

For technology development as well as for transistor model assessment and validation (i.e., for efficient MMIC design), accurate large-signal RF measurements are essential. The aim of the investigations reported here was to study experimentally how the linearity characteristics of our GaAs HBTs are influenced by the bias settings and the RF load impedances at the input and output ports.

Class B and AB performances were separately investigated. In class B mode, the DC collector current moves from a small quiescent value to the normal operating point while the RF input power level is increased, since the RF signal is rectified by the base-emitter diode. As a consequence, the bias is automatically adjusted according to the RF signal level (commonly referred to as *self-biasing*) and the power-added efficiency (PAE) is improved (with respect to constant-bias class A operations) over a wide dynamic range. As a drawback, the performance of the amplifier in terms of signal distortion is somewhat compromised, even for small RF power levels. For this reason, class B operations are interesting for applications which involve only constant envelope modulation (for example, GSM), so that the transistor can be pushed even into hard saturation (where the highest efficiency is obtained). On the other hand, power stages for 2.5G and 3G mobile communication systems have more demanding linearity specifications, and a trade-off between linearity and efficiency must be found. Therefore, the power amplifier must operate below the saturation region, and class AB mode is preferred to ensure constant gain versus input power level.

EXPERIMENTAL APPROACH

The investigations were carried out on-wafer using our active harmonic load-pull system, which is similar to the one described in Ferrero et al., (1). Vector-corrected RF measurements are performed by a four-sampler microwave receiver (network analyzer), which allows the acquisition of the four waves at the DUT input and output ports. As a result, the input and output powers are accurately determined, as well as the load reflection coefficient, the DUT input gamma and, last but not least, the AM-AM and AM-PM conversion characteristics of the device. The programmable RF load is based on the active loop architecture (Bava et al. (2), Mueller and Gyselinckx (3)). Two separate loops are available, so that the system can be easily reconfigured to perform source- or load-pull, controlling the RF impedance at up to two different frequencies. A traditional passive tuner can be used as well for load-pulling, taking advantage of the network analyzer measurement capabilities.

Linearity characteristics of an amplifier are normally specified in terms of intermodulation distortion (IMD) or adjacent channel power ratio (ACPR). In both cases, the device is driven by a modulated (spread spectrum) RF signal, and the unwanted spectral content caused by the non linearity is compared to the desired signal power. In a real amplifier circuit, it can be assumed that the load impedance at the output stage is constant over the whole signal bandwidth (typically some MHz), since the matching network consists only of a few lumped passive elements. Unfortunately, the same assumption cannot be made for the active load-pull system, since long cables and high-Q elements are included in the active loop. As a result, the load reflection coefficient typically shows an unacceptable phase shift (more than ten degrees in few MHz) that can seriously degrade the accuracy of the measurement. A possible solution consists in reducing the bandwidth of the input RF signal, e.g. by reducing the spacing of the two tones used in the intermodulation measurements. Unfortunately, this is often not possible, because of the thermal frequency response of the amplifier (McIntosh and Snowden, (4)). As an alternative, a passive broadband tuner could be used to set the load impedance. This solution has, however, two severe limitations. First, the magnitude of the reflection coefficient is limited by losses of the measurement setup between the tuner and the on-wafer probe tips. This is a problem when measuring power stages operating at low supply voltages (typical for mobile communication applications), since the optimum load impedance is in the order of a few ohms. Second, the independent control of the load impedance at harmonic frequencies is difficult, or even impossible.

The solution proposed here consists in performing only simple single-tone measurements to determine the AM-AM and AM-PM conversion characteristics, and computing the linearity performances (in terms of IMD or ACPR) from these. This method is based on an approach well-known in literature (see, for example, Kaye et al. (5), Sevic and Staudinger (6)). It considers the complex envelope of the modulated microwave signals and describes the modulated waves at the input at output port as

$$a_1(t)=\left|\tilde{a}_1(t)\right|\cos(2\pi f_0 t) \qquad b_1(t)=\left|\tilde{b}_1(t)\right|\cos\left(2\pi f_0 t+\angle\tilde{b}_1(t)\right) \qquad (1),(2)$$

$$a_2(t)=\left|\tilde{a}_2(t)\right|\cos(2\pi f_0 t+\angle\tilde{a}_2(t)) \qquad b_2(t)=\left|\tilde{b}_2(t)\right|\cos\left(2\pi f_0 t+\angle\tilde{b}_2(t)\right) \qquad (3),(4)$$

where f_0 is the RF signal frequency. The quantities $\tilde{a}_\bullet(t)$, $\tilde{b}_\bullet(t)$ are complex-valued, slowly time-varying functions – with respect to $\cos(2\pi f_0 t)$ – and they represent the envelope carrying the digital information.

In the case of single tone measurements, those functions are constant and represent the magnitude and phase of the RF signals. The relationship between $\tilde{b}_1$, $\tilde{b}_2$ and $\tilde{a}_1$ describes the non-linear behavior of the device:

$$\tilde{b}_1=\Gamma_{in}\left(\left|\tilde{a}_1\right|\right)\cdot\tilde{a}_1 \qquad \tilde{b}_2=S_{21L}\left(\left|\tilde{a}_1\right|\right)\cdot\tilde{a}_1 \qquad (5),(6)$$

The relationship $\tilde{a}_2=\Gamma_L\cdot\tilde{b}_2$ is imposed by the output load. Since the device is non-linear, the two quantities S_{21L} (large-signal gain) and Γ_{in} (input reflection coefficient) depend on $\left|\tilde{a}_1\right|$ (i.e. on the input power level) and they represent the AM-AM and the AM-PM conversion characteristics of the transistor. They are automatically measured by the network-analyzer-based system, as a result of a simple power sweep, and they depend on the bias settings and the loading conditions (source and load, fundamental and harmonic).

Now we will consider the case of a modulated RF input signal. If we assume that the transistor has no memory effect within the bandwidth of the modulated signal (Staudinger, (7)), then the complex envelope of the RF waves can be directly computed from $\left|\tilde{a}_1(t)\right|$, S_{21L} and Γ_{in} by means of (5) and (6):

$$\tilde{b}_1(t)\approx\Gamma_{in}\left(\left|\tilde{a}_1(t)\right|\right)\cdot\tilde{a}_1(t) \qquad \tilde{b}_2(t)\approx S_{21L}\left(\left|\tilde{a}_1(t)\right|\right)\cdot\tilde{a}_1(t) \qquad (7),(8)$$

Once $\tilde{a}_1(t)$, $\tilde{b}_1(t)$, $\tilde{a}_2(t)$, $\tilde{b}_2(t)$ have been computed in the time-domain, their Fourier transforms (FFTs) provide the spectrum of the modulated RF signals, so that all the quantities of interest can be computed, such as input and output powers, third-order intermodulation distortion (IM_3), ACPR, etc. *As a result, the linearity characteristic of the device can be evaluated through the standard quantities (IM_3, ACPR) after simple, fast, single-tone load-pull measurements.*

The experimental verification of this approach was carried out by investigating small power transistors. In this case, a broadband passive tuner was sufficient to set reasonable loading conditions. Therefore, the linearity characteristics evaluated through AM-AM and AM-PM conversion could be compared to IMD and ACPR measured with a spectrum analyzer. We observed a good agreement, that is suitable for most of the applications. Afterwards, the same approach was used to investigate the large-emitter-area HBTs.

MEASUREMENT RESULTS AND DISCUSSION

As an example, in the following we will investigate the linearity of a 2880 μm^2 emitter-area HBT power cell, designed to operate at 3 V with a maximum collector current of about 600 mA for normal operation (current density: $J_{C,max}$ = 0.2 mA/μm^2). The fundamental frequency f_0 was set to 1 GHz. In this conditions, the optimum load impedance for maximum output power is mainly real, and its resistance is in the order of 5 Ω.

Understanding the self-biasing effect and the quantities by which it is influenced is the key point to study the non-linear behavior of the transistor in class B and AB mode, since S_{21L} depends strongly on the DC collector current. Figure 1 shows a logarithmic plot of the measured collector current as a function of the output power. It can be noted that the quiescent current chosen for class B operation is about 30 mA (1/20 $J_{C,max}$), while for class AB mode is about 120 mA (1/5 $J_{C,max}$). Figure 1 shows also the measured gain $|S_{21L}|$. In class B mode, it starts from a relative low value (due to the low quiescent current), rises to a maximum with increasing RF power, and finally drops due to output compression. The lower the load impedance, the higher is the self-biasing and the gain expansion. A similar effect can be seen for class AB operations, but the gain expansion is reduced, since the collector current starts from a higher quiescent value. Note that a lower load impedance results in a higher output power at which the gain drop appears. The AM-PM characteristics ($\angle S_{21L}$ vs. P_{out}, not shown for brevity) behaves similarly.

Figure 2 shows the third-order intermodulation distortion IM_3, computed by the method described in the previous section, as a function of the output power for different values of the load impedance at fundamental frequency. For class B operation in the back-off region, the base-emitter diode distortion is the main non-linear effect. A lower load impedance leads to a higher IM_3 level, since gain expansion is maximized and the output signal envelope is strongly distorted. On the other hand, when output compression happens, a new non linear contribution arises. First, it causes a partial cancellation of the intermodulated component (see the IM_3 minimum around +25 dBm in Figure 2), then it becomes dominant, with an abrupt increase of the IM_3. Class AB behavior is similar, with a reduced IM_3 level in the back-off region, due to the higher quiescent current. As an example, for an hypothetical application with a maximum IM_3 level of -40dBc, the optimum load would be $\Gamma_L(f_0) = -0.8$, with a maximum output power of about +24 dBm. If the IM_3 specification were relaxed down to -30 dBc, then a load of $\Gamma_L(f_0) = -0.9$ would allow an output power up to +27 dBm.

One of the advantages of the described method lies in the possibility of investigating the power stage linearity as a function of the input and the output port harmonic terminations. For this, the AM-AM and AM-PM conversion characteristics are required for each harmonic impedance setting of interest. As an example, the IM_3 curves of our 2880 μm^2 emitter-area transistor were computed for $\Gamma_L(f_0) = -0.8$ and for different values of the 2^{nd} harmonic source impedance $\Gamma_S(2f_0)$. The expectations were that the harmonic tuning at the input could influence the linearity of the device, since a relevant portion of the waveform distortion is due to the base-emitter diode exponential characteristic.

The linearity curves for class AB harmonic source-tuned operation are shown in Figure 3. The interesting $\Gamma_S(2f_0)$ values were found on the unitary circle close to the short circuit point. In this area of the Smith chart, we observed extreme behaviors. From one side ($\Gamma_S(2f_0) \approx 1\angle +180°$) self-biasing is minimized, but output compression happens for relatively low power levels. On the opposite side ($\Gamma_S(2f_0) \approx 1\angle +170°$), self-biasing is further enhanced, and the output power allowed for maximum IM_3 level of –40dBc is extended from +24dBm (obtained with $\Gamma_S(2f_0) = 0$) up to +26dBm. A similar behavior has been found for ACPR performance and for class B operations (not reported for brevity).

CONCLUSIONS

The linearity performance of HBT power stages operating in class B and AB was studied by computing the intermodulation distortion and ACPR curves from the AM-AM/AM-PM conversion characteristics, which were easily determined by single-tone load-pull measurement data. As an interesting result, the input termination at the second harmonic significantly affected the maximum output power for a given distortion level. Therefore, in the power MMIC design the value of the second harmonic source impedance should be carefully tuned to improve the linearity performance of the amplifier.

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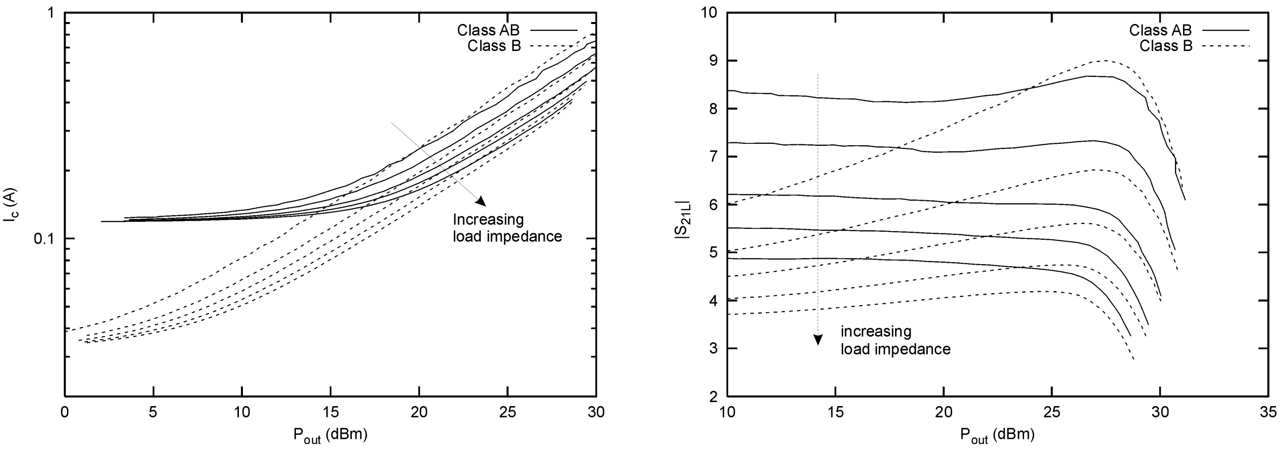

Fig.1. DC collector current and AM-AM conversion characteristics as a function of RF output power for different load impedance values ($\Gamma_L(f_0) = -0.7, -0.75 ... -0.9$).

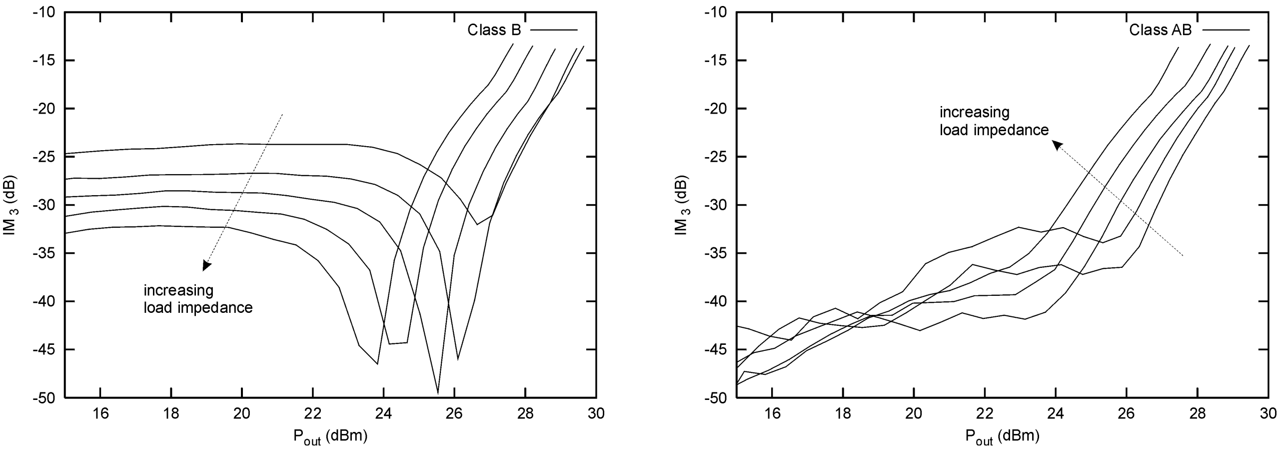

Fig.2. Class B and AB third-order intermodulation versus RF output power for different load impedance values ($\Gamma_L(f_0) = -0.7, -0.75 ... -0.9$).

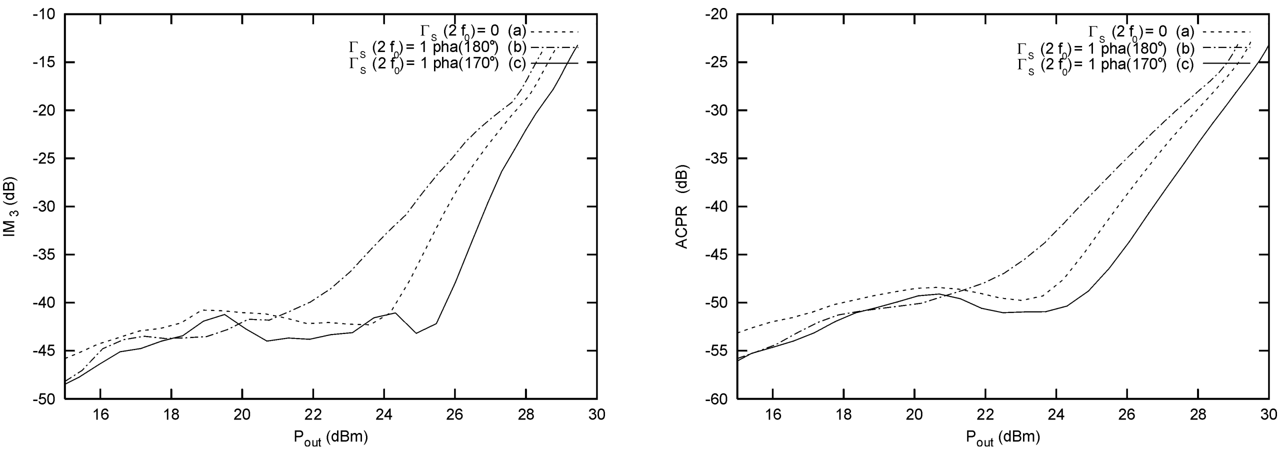

Fig.3. Class AB IM_3 and ACPR curves versus RF output power for $\Gamma_L(f_0) = -0.8$ and different second harmonic source terminations: (a) 50Ω ($\Gamma_S(2f_0) = 0$), (b) short circuit ($\Gamma_S(2f_0) = 1\angle 180°$), (c) $\Gamma_S(2f_0) = 1\angle 170°$.