Self-Linearizing Technique for L-Band HBT Power Amplifier: Effect of Source Impedance on Phase Distortion

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***Abstract*— L-band power amplifiers operating with high efficiency and high linearity at a single and low supply voltage are in strong demand for mobile communication systems. This paper presents a new self-linearizing technique for power heterojunction bipolar transistors (HBT's). Utilizing the nonlinear input conductance of the device itself and setting the source impedance to the self-linearizing impedance, the phase distortion and the adjacent channel leakage power (ACP) for π/4-shift QPSK modulated signal of our InGaP/GaAs power HBT's have been greatly improved. As a result, the HBT exhibited the ACP at 50 kHz offset frequency of −49.2 dBc with a power-added efficiency (PAE) of 56% at an output power ($P_{\rm out}$) of 31 dBm under a supply voltage of 3.5 V.**

I. Introduction

HETEROJUNCTION bipolar transistors (HBT's) are one of the most promising devices for future mobile communication handsets due to their superior power characteristics even at low-bias voltage with a only single supply, and their higher power handling capability with a smaller chip size [1]. In addition, the advantages of InGaP/GaAs HBT's over AlGaAs/GaAs HBT's for low-voltage-operation amplifiers are their smaller knee voltage, due to the smaller collector-emitter offset voltage of less than 0.1 V, and the higher reliability even at a high current density [2]. The large current handling capability enables high output power with a low supply voltage at a low cost [3].

Moreover, recent digital communication systems require power amplifiers with much higher linearity. The linearity of HBT's, however, has not been studied in detail so far. In particular, the effects of load and source impedance on phase distortion and adjacent channel leakage power (ACP) for a π/4-shift QPSK-modulated signal are not fully understood.

We discuss the nonlinear effect of device parameters on phase distortion and present a new self-linearizing matching technique utilizing a nonlinear input conductance of the device itself.

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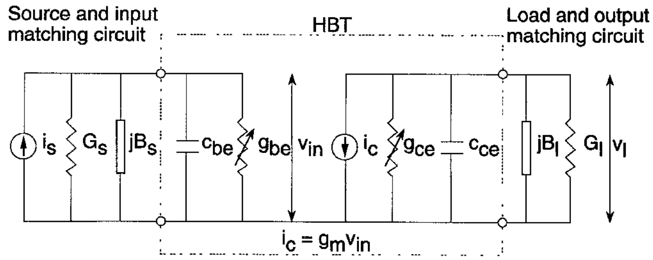


Fig. 1. Equivalent circuit of a common-emitter HBT.

II. Phase Distortion Analysis

ACP originates from the nonlinearity of a device. Power gain compression and phase distortion result in ACP increase as the output power increases. The nonlinear effect of device parameters on phase distortion is discussed here using the simplified equivalent circuit of a common-emitter HBT shown in Fig. 1. In our analysis, we assume that 1) a feedback capacitance, c_{bc}, and a feedback conductance, g_{bc}, are ignored due to their small values and simplifying the analysis, 2) the nonlinearity of the output capacitance, c_{ce}, can be ignored, 3) the input and output conductance, g_{be} and g_{ce}, increases with the output power, and 4) the nonlinearity of g_{be} is greater than that of the input capacitance c_{be} when the base-emitter is forward-biased [4].

The relationship between a signal source current, i_s, and the output voltage, v_l, is shown below

$$v_l = \frac{g_m i_s \exp j(\pi - \theta_l - \theta_s)}{|g_{ce} + G_l + j(\omega c_{ce} + B_l)||g_{be} + G_s + j(\omega c_{be} + B_s)|} \tag{1}$$

$$\tan \theta_l = \frac{\omega c_{ce} + B_l}{g_{ce} + G_l} \tag{2}$$

$$\tan \theta_s = \frac{\omega c_{be} + B_s}{g_{be} + G_s} \tag{3}$$

where G_s is the conductance of the input matching circuit, B_s the susceptance of the input matching circuit, G_l the matching load conductance, and B_l the matching load susceptance.

The gain compression can be explained by (1). As the input power increases, g_{ce} increases and g_m decreases because of the current clipping at saturation region in the common-emitter I-V curves. This results in compression of the output voltage swing v_l.

TABLE I
DC AND SMALL-SIGNAL CHARACTERISTICS OF THE UNIT HBT

Emitter size	2 μm x 20 μm
$R_{ballast}$	2.5 Ω
$V_{ce(offset)}$	80 mV
BV_{ceo}	14 V
h_{fe}	73
f_T	40 GHz
f_{max}	110 GHz
MTTF	1×10^6 hours @200 °C, Jc=6×10^4 A/cm^2

The output phase deviation, $\Delta\theta_l$, and the input phase deviation, $\Delta\theta_s$, can be calculated from the differentials of the θ_l and θ_s given in (2) and (3), respectively

$$\Delta\theta_l = -\frac{\omega c_{ce} + B_l}{(g_{ce} + G_l)^2 + (\omega c_{ce} + B_l)^2}\Delta g_{ce} \tag{4}$$

$$\Delta\theta_s = -\frac{\omega c_{be} + B_s}{(g_{be} + G_s)^2 + (\omega c_{be} + B_s)^2}\Delta g_{be} + \frac{g_{be} + G_s}{(g_{be} + G_s)^2 + (\omega c_{be} + B_s)^2}\omega\Delta c_{be}. \tag{5}$$

The total phase distortion, $\Delta\phi$, is given by

$$\Delta\phi = -\Delta\theta_l - \Delta\theta_s \tag{6}$$

$$= \frac{\omega c_{ce} + B_l}{(g_{ce} + G_l)^2 + (\omega c_{ce} + B_l)^2}\Delta g_{ce} + \frac{\omega c_{be} + B_s}{(g_{be} + G_s)^2 + (\omega c_{be} + B_s)^2}\Delta g_{be} - \frac{g_{be} + G_s}{(g_{be} + G_s)^2 + (\omega c_{be} + B_s)^2}\omega\Delta c_{be}. \tag{7}$$

When an input-matching circuit is conjugate-matched for maximum gain, $\omega c_{be} + B_s$ is zero. When an output matching circuit is tuned for maximum output power or efficiency, $\omega c_{ce} + B_l$ is not zero because the load impedance deviates from the conjugate-matched impedance. From (7) it can be seen that the phase distortion, $\Delta\phi$, depends mainly on Δg_{ce} and Δc_{be} in the above matching condition. In this paper, we propose that the phase distortion, $\Delta\phi$, can be canceled with the nonlinearity of input conductance, Δg_{be}, in (7). In HBT's, nonlinearity of the input conductance, g_{be}, is great when the input base-emitter circuit is forward-biased. By optimizing an input-matching circuit impedance, that is, the susceptance, B_s, the sign and value of the second term in (7) can be adjusted to cancel the other terms. In other words, phase distortion can be self-compensated by the nonlinear input conductance of the device itself and the susceptive component of the input-matching circuit impedance. We can reduce the phase distortion $\Delta\phi$ and achieve a superior ACP performance with this self-linearizing technique.

III. FABRICATION AND BASIC CHARACTERISTICS

The InGaP/GaAs HBT structure grown by MOCVD consists of an n^+-InGaAs cap layer, an n-InGaP emitter layer, a C-doped p^+-GaAs base layer, and n/i-structure collector layers [3].

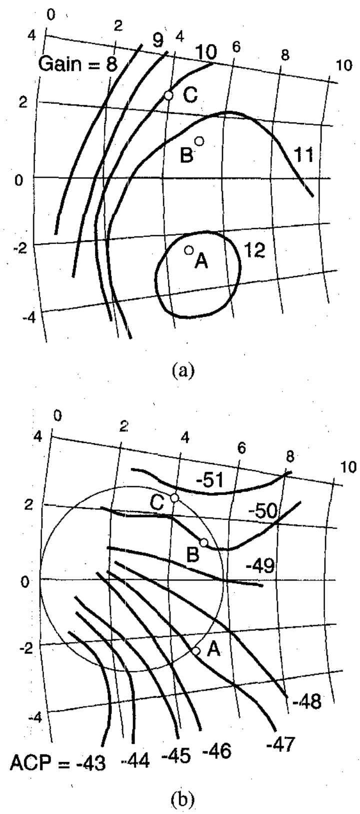


Fig. 2. Source-pull measurement results of InGaP/GaAs HBT. Figure (a) shows the contour lines for power gain and (b) shows the contour lines for ACP at 50 kHz offset frequency for a PDC 1.5-GHz $\pi/4$-shift QPSK signal with a P_{out} of 31 dBm, a V_{ce} of 3.5 V under a class A-B operation. Point A is the gain-matched impedance, points B and C are the points with changed susceptance values (show Table II). Constant conductance circle is also plotted in (b).

Table I summarizes the DC and small-signal characteristics of an HBT with an emitter area of 2 μm $\times$ 20 μm. We used a ballasting resistance of 2.5 Ω for thermal stability [3]. The collector emitter offset voltage, $V_{ce(\text{offset})}$, is 80 mV, which is 0.1–0.3 V lower than that of the AlGaAs/GaAs HBT we fabricated. BV_{ceo} is 14 V. f_T and $f_{\max}$ of the InGaP/GaAs HBT are 40 GHz and 110 GHz respectively, at a collector current I_c of 16 mA and a collector bias V_{ce} of 2.5 V. The MTTF of our HBT's is 1×10^6 h with a current density of 6×10^4 A/cm^2 at a junction temperature T_j of 200°C [2].

The fabricated power HBT's have 64 fingers of 2 μm $\times$ 20 μm emitter. The total emitter area is 2560 μm^2, and the structure of the HBT is a so-called plated heat sink (PHS) with via holes.

IV. EXPERIMENTAL RESULTS AND DISCUSSION

Fig. 2 shows the source-pull measurement results. Fig. 2(a) shows the contour lines for several values of power gain, and Fig. 2(b) shows the contour lines of ACP at 50 kHz offset frequency for the Japanese personal digital cellular (PDC) 1.5-GHz $\pi/4$-shift QPSK signal with a P_{out} of 31 dBm, a V_{ce} of 3.5 V under a class A-B operation, and a quiescent collector current $I_{c(\text{dc})}$ of 400 mA. The load impedance is set to a minimum ACP with a PAE of over 50% when the source impedance is gain-matched. In Fig. 2 point A is the

TABLE II
CONDUCTANCE AND SUSCEPTANCE OF THE SOURCE IMPEDANCE POINTS A, B, AND C

Impedance Point	Conductance (S)	Susceptance (S)
A	0.182	0.092
B	0.199	-0.041
C	0.192	-0.116

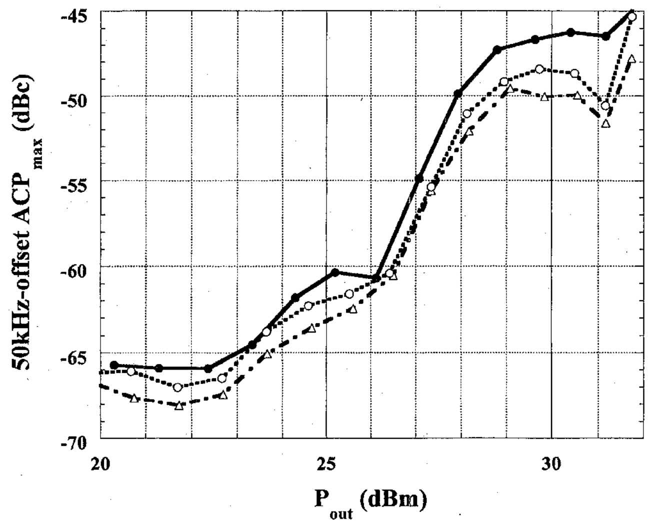


Fig. 3. 50 kHz-offset ACP versus output power for source impedance A (solid line with closed circles), B (dashed line with open circles), and C (dashed line with open triangles). V_{ce} is 3.5 V and $I_{c(\mathrm{dc})}$ is 400 mA under a class A-B operation.

gain-matched impedance, points B and C are the impedance changed by subtracting susceptance values, B_s, from point A while keeping the conductance at the same value (see Table II). In Fig. 2(b) a constant conductance circle is also plotted. From point A to point C the ACP improves while the gain decreases. The constant ACP contour lines are perpendicular to the constant conductance circle near point A.

Fig. 3 shows the ACP at 50 kHz-offset frequency as a function of output power for source impedance A (solid line with closed circles), B (dashed line with open circles), and C (dashed line with open triangles). V_{ce} is 3.5 V and $I_{c(\mathrm{dc})}$ is 400 mA under a class A-B operation. For A to C when the output power is over 26 dBm, the 50 kHz offset ACP value improves as susceptance decreases below the gain-matched impedance.

Fig. 4 shows phase distortion as a function of output power at source impedance A (solid line with closed circles), B (dashed line with open circles) and C(dashed line with open triangles). Fig. 5 shows gain compression as a function of output power at source impedance A (solid line with closed circles), B (dashed line with open circles) and C (dashed line with open triangles). For A to C when the output power is over 25 dBm, phase distortion is greatly reduced as susceptance decreases below the gain-matched impedance. However, there is only a very slight corresponded degradation in gain compression.

At the gain-matched impedance A, the phase distortion, $\Delta\phi$, increases with output power. According to the previous discussion, it can be said that the sum of the first and third

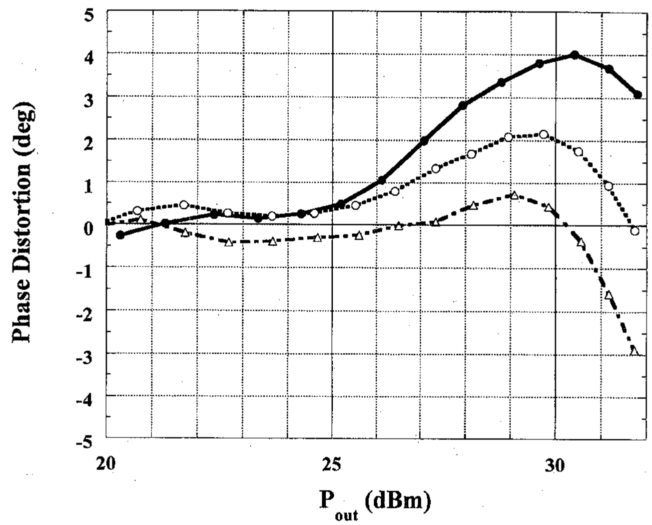


Fig. 4. Phase distortion versus output power for source impedance A (solid line with closed circles), B (dashed line with open circles), and C (dashed line with open triangles).

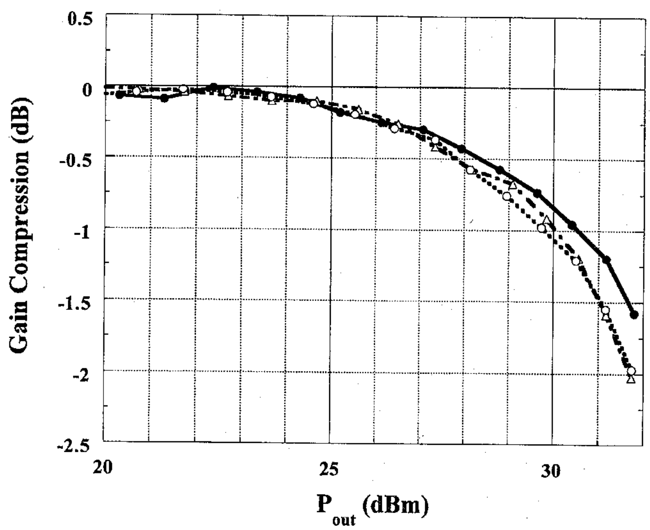
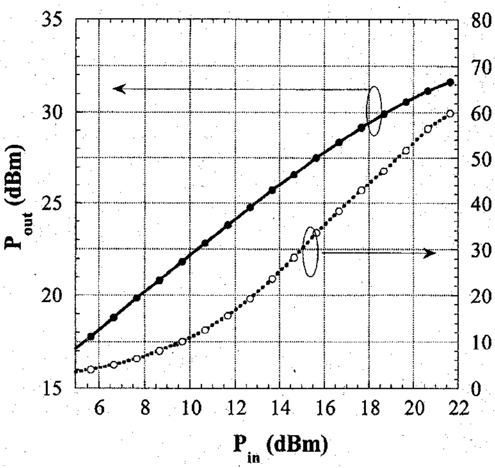


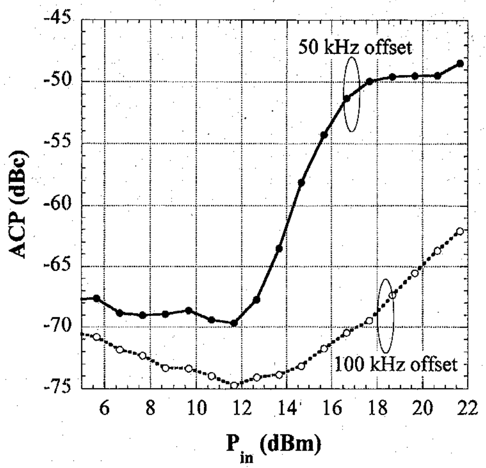
Fig. 5. Gain compression versus output power for source impedance A (solid line with closed circles), B (dashed line with open circles), and C (dashed line with open triangles).

terms in (7) has a positive value because the second term in (7) is almost zero due to conjugate matching of source impedance. To reduce the phase distortion, $\Delta\phi$, the value of the second term in (7) must be negative, so $\omega c_{be} + B_s$ must be negative, because Δg_{be} is positive. This means that the susceptance value must be decreased below the gain-matched impedance to reduce the phase distortion. Figs. 2–4 clearly demonstrate that the phase distortion and 50-kHz-offset ACP of the HBT are greatly improved for impedance A–C where the susceptance value is subtracted from the gain-matched impedance A. These results agree very well with our phase distortion analysis.

Hayashi *et al.* discussed phase distortion of GaAs MESFET's [5]. In FET's, phase distortion is dominated by the nonlinearity of output conductance, Δg_{ds}. However, the nonlinearity of input conductance, Δg_{gs}, is small compared to the nonlinearity of the input capacitance, Δc_{gs}, and it is



(a)



(b)

Fig. 6. (a) Output power and PAE versus input power and (b) 50 kHz-offset and 100 kHz-offset ACP versus input power at the self-linearizing source impedance for an efficiency-matched load impedance.

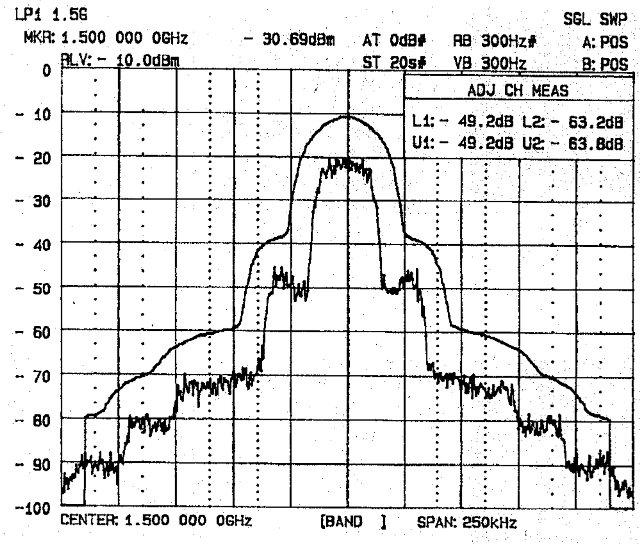


Fig. 7. Output power spectrum at an output power of 31 dBm for a 1.5 GHz modulated input.

difficult to compensate the output phase distortion with source impedance. They used a common-gate FET with a negative phase deviation in the input circuit to compensate for the positive output phase deviation in common-source FET. On the contrary, in HBT's phase distortion can be improved by setting the source impedance to the self-linearizing impedance.

Fig. 6(a) shows the output power and PAE. In Fig. 6(b) the ACP at offset frequencies of 50 and 100 kHz is shown as a function of input power at the self-linearizing source impedance for an efficiency-matched load impedance. Fig. 7 shows the output power spectrum at an output power of 31 dBm for a 1.5 GHz modulated input. As a result, the HBT exhibited the ACP of -49.2 dBc at 50 kHz offset frequency and -63.2 dBc at 100 kHz offset frequency with a PAE of 56% and a P_{out} of 31.0 dBm at a bias of 3.5 V.

V. Conclusion

We have discussed the nonlinear effect of device parameters on phase distortion in InGaP/GaAs HBT's, measured the effect of the source impedance on phase distortion, and presented a new self-linearizing technique utilizing a nonlinear input conductance of the device itself and adjusting only susceptance value of the input matching circuit impedance. With this technique, our HBT exhibited the 50 kHz offset ACP of -49.2 dBc with a P_{out} of 31.0 dBm and a power added efficiency of 56% at a bias of 3.5 V.

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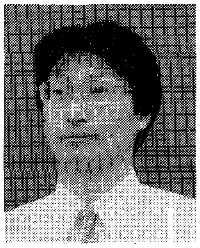

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