Intermodulation in Heterojunction Bipolar Transistors

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Abstract:

This paper examines the modeling of small-signal intermodulation distortion (IM) in heterojunction bipolar transistors (HBTs). We show that IM current generated in the exponential junction is partially cancelled by IM current generated in the junction capacitance, and that this phenomenon is largely responsible for the unusually good IM performance of these devices. Finally we propose a nonlinear HBT model suitable for IM calculations, show how to measure its parameters, and verify its accuracy experimentally.

I. Introduction

One of the most delightful properties of the heterojunction bipolar transistor (HBT) is its unusually high linearity at relatively low levels of dc bias power. For example, Nelson et. al. have reported a third-order intermodulation intercept point (IP_3) of 33 dBm in a small-signal amplifier using two HBTs and 150 mW dc power [1]. Other researchers have reported similar results [2]. This high linearity is most remarkable in view of the exponential dependence of the HBT's emitter current, I_e, on base-to-emitter voltage, V_{be}, an extremely strong nonlinearity. Furthermore, the junction capacitance, consisting primarily a diffusion capacitance, is nearly as strongly nonlinear (in terms of its charge/voltage characteristic) as the junction I/V.

The reason for this unusually high linearity has never been explained adequately, not even in papers on intermodulation distortion (IM) in homojunction BJTs [3]-[5]. One of the most common conjectures is that the output resistance of these devices is very high, and thus does not generate IM current. However, the output resistance is rarely a dominant contributor to IM in other solid-state devices; in MESFETs and HEMTs, for example, it is clearly nonlinear but still only minimally significant. The more strongly nonlinear $I_e(V_{be})$ is the logical candidate for the device's dominant nonlinearity.

We have resolved this quandary through a Volterra-series analysis of an HBT equivalent circuit. We have drawn the counterintuitive conclusion that the largest output distortion current components generated by the resistive junction and those generated by the junction capacitance have a 180-degree phase difference, and, in theory, cancel almost exactly. Thus, paradoxically, it is the strong nonlinearity of both elements that causes the IM levels to be low; if only one of these elements were nonlinear, the device's intercept points would be much lower.

Relatively few nonlinear models of HBTs have been reported, and none of these have been intended specifically for IM analysis [2, 6]. The requirements of a device model for accurate IM calculations are generally more severe than for single-tone, large-signal amplifier analysis [7]. Here we propose such a model, and prove its validity experimentally.

II. Intermodulation in the HBT

Consider the equivalent circuit of an HBT (Figure 1). The resistance R_b represents the sum of the base resistance and the source resistance; the load resistance is R_L. Initially we make the approximation $R_{ee} \cong 0$ (in our devices R_{ee} is less than 2Ω) and treat all capacitances except the base-to-emitter junction capacitance as negligible. The nonlinearities are modeled by a diode (the base-to emitter junction) and a nonlinear capacitance. Using a Volterra-series analysis, we can show that the second-harmonic output current is given by

$$I_{o,2} = \frac{\alpha V_{be,1}^2}{2R_{je}c_1}(R_{je}c_1 g_2 - c_2) - \frac{\alpha V_{be,1}^2 (1-\alpha) g_2}{4R_{je}c_1 j\omega_1} \qquad (1)$$

where c_n and g_n are the Taylor-series coefficients of the junction's resistive and reactive nonlinearities, and $V_{be,1}$ is the excitation-frequency junction voltage. The second term is virtually always negligible. Thus, if g_2 and c_2 are both positive, the second-harmonic components of distortion current cancel.

The c_n and g_n coefficients are

$$1/R_{je} = g_1 = \delta I_e,\; g_2 = \delta^2 I_e/2,\; c_1 = \tau\delta I_e,\; c_2 = \tau\delta^2 I_e/2$$

Substituting these into (1), we find that the term $\frac{\alpha V_{be,1}^2}{2R_{je}c_1}(R_{je}c_1 g_2 - c_2)$ cancels exactly, leaving only the negligible second term.

A similar situation occurs in third-order IM, although it is somewhat more complicated because of the large number of frequency components involved. We observe the same IM cancellation in both the third harmonic and two-tone intermodulation products. This cancellation occurs above a corner frequency given by $\omega_1 > \frac{1}{2R_b c_1}$.

Perfect cancellation does not occur for a number of reasons. One is that the relation for the junction charge, $Q=\tau I_e$, is not accurate at microwave frequencies, and the junction capacitance includes a substantial depletion component. It is interesting--and perhaps tantalizing--to note that if perfect cancellation did in fact occur, the mid-frequency second- and third-order intercept points of these devices would be on the order of 70 dB. Thus, there may be considerable room for improving the linearity of these devices by closely matching their capacitive and reactive nonlinearities.

III. Modeling the HBT

To model the HBT we use the equivalent circuit of Figure 1. This circuit includes three nonlinearities: the resistive junction, modeled as an ideal junction diode, the capacitive junction, and the nonlinear current gain $\alpha(I_e)$. We also include a number of parasitics that were ignored in the previous section; most important of these is the emitter resistance, R_{ee}.

The parameters of the model are found from a combination of dc and S-parameter measurements. We have measured the parameters of the junction *I/V* characteristic in two ways. The first is to plot $\log(I_e)$ as a function of V_{be} at low current levels (so-called *Gummel plots*). The parameters of the junction can be found in a straightforward manner [8]. It is not possible to find R_{ee} from this plot, however, because of heating in the junction. Instead, the sum of R_{je} and R_{ee} can be found easily by converting measured S parameters to Z parameters; then R_{ee} is found from $Z_{12} = R_{je} + R_{ee}$. Alternatively, one can plot Z_{12} as a function of $1/I_e$; the extrapolated y-intercept is R_{ee}.

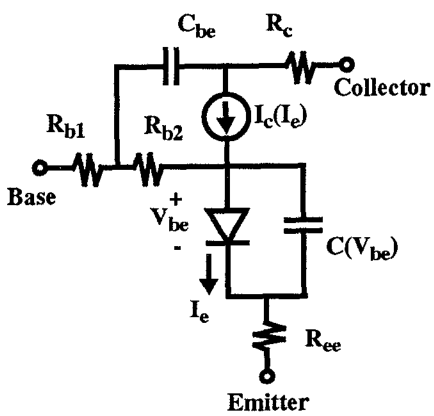

Fig. 1. HBT equivalent circuit

The collector current I_c is a nonlinear function of emitter current, and there is a time delay τ between the two currents. Because of numerical difficulties in differentiating $I_c(I_e)$, the nonlinearity in α is best found by differentiating $\beta(I_e)$(measured at low RF frequencies). We express I_c as a Taylor series, and must find τ and the α_n coefficients:

$$I_c = \alpha_1 I_e(t-\tau) + \alpha_2 I_e^2(t-\tau) + \alpha_3 I_e^3(t-\tau) \qquad (2)$$

These are found by measuring $\beta=H_{21}$ at a number of bias currents and deriving the α_n coefficients from the derivatives of β. We have found that the current gain is a very significant nonlinearity affecting IM in HBTs; neglecting it results in several dB error in the second- and third-order intercept points.

The nonlinear base-to-emitter capacitance is the parameter most difficult to model. Because the standard expression for the charge in this element, $Q=\tau I_e$, is not adequate at microwave frequencies, we must determine the c_n coefficients by measurement. The first coefficient, c_1, is the linear capacitance; it is found by fitting the model to measured S parameters. Once $I_c(I_e)$ and $I_e(V_{be})$ are established, c_2 is the only parameter affecting second-order intermodulation; thus it can be adjusted so that calculated and measured IM levels are identical at some convenient point in the mid-frequency region (the measurements are made with 50-ohm source and load impedances). Similarly,

Table 1: HBT Model Parameters

TRW 2x10 micron quad-emitter;

V_{ce}=3.0, I_c=16 mA

Parameter	**Value**
R_{ee}	1.7
R_{je}	1.75
R_{b1}	3.7
R_{b2}	3.7
R_c	3.0
C_{bc}	$58.7 \cdot 10^{-15}$
η	1.065
I_0	$1.0 \cdot 10^{-23}$
c_1	$1.85 \cdot 10^{-12}$
c_2	$1.2 \cdot 10^{-11}$
c_3	$4.0 \cdot 10^{-11}$
α_1	0.9764
α_2	0.22605
α_3	-6.259
τ	$3.66 \cdot 10^{-12}$

when c_2 is known, the only remaining parameter affecting third-order IM is c_3; this parameter is adjusted to give the correct third-order IM levels.

This process is adequate for characterizing the Q/V coefficient at a single dc bias point. However, for large-signal modeling (an eventual goal of this work) it will be necessary to find more general ways to characterize this nonlinearity.

Table 1 lists the equivalent-circuit parameters of a TRW 2x10 micron, quad-emitter HBT. One may note from Table 1 that the c_2 and c_3 coefficients are considerably smaller than one would obtain by assuming that they represent ideal diffusion capacitances.

Figure 2 compares the measured and the modeled second-order (second-harmonic) intercept points, and Figure 3 shows the

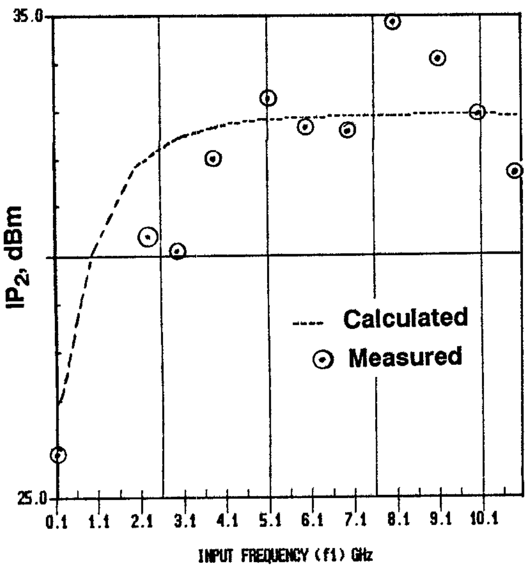

Fig. 2. Second-harmonic intercept point. $Z_s = Z_L = 50+j0$

third-order. To verify the modeling of both the resistive and reactive junction nonlinearities, the figures include intercept-point measurements at frequencies above and below the corner frequency $\omega=1/2R_bc_1$. We estimate the accuracy of the measured intercept points to be no better than ±1 dB. The accuracy is limited by the HBT's pronounced sensitivity to its terminating impedances, and the imperfect VSWR (approximately 2.0:1) of the probes used in the on-wafer measurement. We are currently examining the possibility of using network-analyzer calibration techniques to characterize the output and input circuits; this should improve the accuracy considerably.

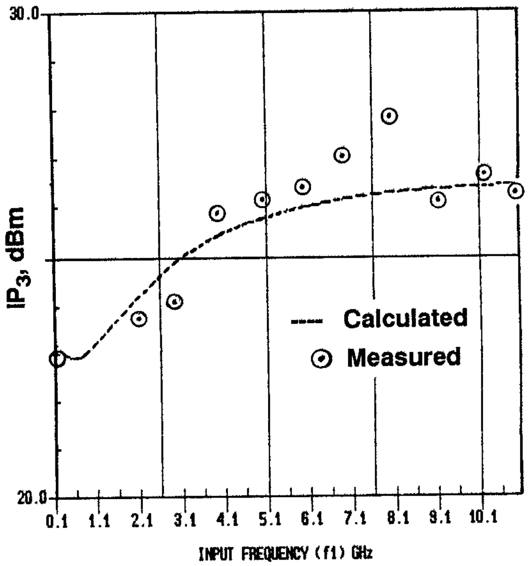

Fig. 3. Third-order intercept point $(2f_2 - f_1)$

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