A NEW SPICE-TYPE HETEROJUNCTION BIPOLAR TRANSISTOR MODEL FOR DC, MICROWAVE SMALL-SIGNAL AND LARGE-SIGNAL CIRCUIT SIMULATION

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ABSTRACT

Accurate modelling of the microwave large-signal characteristics of Heterojunction Bipolar Transistors (HBTs) is extremely useful for microwave power applications of this device. This paper presents a new type of HBT large-signal model which is valid for DC, small-signal and large-signal AC modes of operation. The model may be used over a wide range of operating conditions and includes allowance for self-heating effects which are very important for HBTs. Through the use of several novel features, the model is differentiated from traditional Ebers-Moll or Gummel-Poon BJT representations. The new model is accompanied by a very simple parameter extraction process requiring only a series of conventional DC measurement and multi bias point small-signal S parameter measurements. The model is validated by independent power sweep measurements on HBTs from two different manufacturers.

INTRODUCTION

Heterojunction bipolar transistors (HBTs), based either on the material system AlGaAs/GaAs or GaInP/GaAs, are widely used in many analogue and digital applications [1]. For modern CAD-based circuit design, there is therefore a need for a unified HBT circuit model which should ideally be valid for DC, microwave small-signal and large-signal operation.

Many designers have directly used conventional BJT models such as the Ebers-Moll or Gummel-Poon for circuit designs incorporating the HBT [2-6]. However, the models which result are usually only useful in restrictive areas of application because HBTs exhibit several significant differences compared to Silicon BJTs, which cannot be adequately represented by traditional BJT models. Many existing HBT-specific models are only valid for DC or small-signal AC, although some of them can be extended into non-linear operation by using Volterra series for simulation at medium input signal level and at a fixed bias point. Grossman and Choma [5] have presented an Ebers-Moll type large-signal model which includes many high order effects to achieve accuracy, but this model is rather complex for use in commercial software environments, and the associated parameters are difficult to extract.

Until the time of writing, no unified HBT model which can cover DC, small-signal AC and microwave large-signal applications has been reported. This paper presents such a model, which is quite different from the conventional Ebers-Moll or Gummel-Poon models, and is best described as a semi-experimental-based model, valid for the entire application region from DC to large-signal microwave conditions. The main novel features of this model are: (i) a new basic topology is proposed, involving a division of the base region into several sub-nodes, (ii) two temperature-dependent diodes are used to simulate self-heating effects; (iii) two different ideality factors are used for the above two diodes and also two non-linear resistors to simulate non-uniform gain, (iv) a new method is introduced to reconstruct the base non-linear charge in order to simulate base storage charge, (v) the new model also yields a very simple, direct and reliable parameter extraction process.

THE PROPOSED HBT MODEL

Conventional BJT models such as Ebers-Moll and Gummel-Poon are based on the assumption that surface recombination effects can be ignored. This kind of model uses constant current gain β_F to relate the base and collector currents of the BJT. This description is true for most modern BJTs which normally present constant current gains for nearly the whole forward active region except the regions with very low or very high base and collector currents. Since the base-emitter heterojunction can introduce many recombination centres on the interface of the junction, modern HBTs rarely present constant current gains in forward regions [4]. The Gummel-plot measurement of an HBT normally produces two non-parallel straight lines (Ib and Ic) in the forward active region, corresponding to nonuniform current gain. Since the most important requirement for a large-signal model is its ability to give correct predictions of the current gain in the forward active region, the nonuniform gain characteristic of the HBT leads us to abandon a modelling approach based on conventional BJT models. To create a new model for the HBT which fits the experimental data, we first assume that the relationship between the voltage across the BE heterojunction and the base current (Ib) should follow the diode equation, and that high current effects such as the Early effect and the Kirk effect can be ignored[5]. Another very important factor in modelling is the self-heating effect which is caused by the high power density of the HBT and the low thermal conductivity of the GaAs

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substrate. The output characteristics of an HBT often presents a negative slope which becomes more severe when dissipated power increases. It is very difficult to analyse the temperature distribution across the HBT and we assume in the following that this distribution is uniform. Based on these assumptions, the new HBT model has been developed and is shown in Figure 1.

The static part of the proposed model includes two temperature-dependent diodes and two nonlinear resistances in order to track accurately the DC characteristics of the HBT. To simulate the non-uniform current gain, the base region is actually split into several nodes, whilst two diodes with different characteristics simulate Ib and Ic independently. Of course this represents a considerable departure from conventional BJT modelling, but it does provide a new way to simulate non-uniform gain effects for the HBT. The equations for these two diodes are as follows:

For Dbe_ib:

$$I = I_{S,b}(T_j)\cdot\left(e^{V/(n_b(T_j)\cdot V_T)}-1\right) \qquad (1)$$

For Dbe_ic:

$$I = I_{S,c}(T_j)\cdot\left(e^{V/(n_c(T_j)\cdot V_T)}-1\right) \qquad (2)$$

where V and I are the voltage and current applied to the diode ports, and Tj is proportional to the temperature of the junction. Temperature-dependence in the diode characteristics is introduced by means of the following expressions for the saturation current and the identity factor [5]:

$$I_s(T_j) = I_s(T_0)\cdot e^{\left(\frac{1}{T_0}-\frac{1}{T_j}\right)\cdot E_g} \qquad (3)$$

$$n(T_j) = n(T_0)\cdot\left(1+A_n\cdot\Delta T+B_n\cdot\Delta T^2\right) \qquad (4)$$

where T_0 is room temperature and Tj is the *computation junction temperature*. The latter term is used here as distinct from the real junction temperature, which at some particular points in the device might differ from Tj. The computation junction temperature is a tool to connect the dissipated power inside the HBT with the electrical behaviour of the device. It is proportional to the average temperature of the junction area, although that temperature distribution across the junction might have a much more complex pattern. The computation junction temperature is calculated from the dissipated power inside the HBT by the following equations:

$$P_D = I_B\cdot V_{BE} + I_C\cdot V_{CE} \qquad (5)$$

$$f(P_D, \Delta T, G, t) = 0 \qquad (6)$$

where P_D, ΔT, t are the dissipation power, junction temperature increment and time. G is a parameter set which is related to the HBT's geometry and physical parameters (such as the thermal conductivity and the location of heat-sink), which can be used to calculate the temperature of the junction. A complete description of G would involve thermodynamic analysis and its electrical analogue equivalent might be a multi-section RC network or RC transmission line. For the simplest case, i.e. a first order approximation, equation (6) can be represented by:

$$P_D = \frac{\Delta T}{R_{th}} + C_{th}\cdot\frac{d(\Delta T)}{dt} \qquad (7)$$

where $\Delta T = T_j - T_0$, Rth is the *computation thermal resistance* and Cth is the *computation thermal capacitance*. Again, these two terms are used to avoid confusion with the normal definition of thermal resistance and capacitance. The equivalent circuit to describe equations (5) and (7) is also shown in Figure 1.

The two nonlinear resistances Rbi_ib and Rbi_ic in Figure 1 are used to adjust the high current region current gain and in the work reported here are found to be well approximated by:

$$R_b = f_1(I_b) = K\cdot I_b\cdot e^{H\cdot I_b} + D \qquad (8)$$

$$R_c = f_2(I_c) = K\cdot I_c\cdot e^{H\cdot I_c} + D \qquad (9)$$

where K, H and D are constants which can be extracted from the DC measured data. In the new model the base current and collector current are two separate components which are controlled by the base-emitter voltage via two nonlinear resistances. The traditional proportional relationship between Ib and Ic for a normal BJT via a constant current gain is replaced by a more complex implicit relationship. To support this, it has been observed that most AlGaAs/GaAs HBTs exhibit a region in the Gummel plot where log(Ic) and log(Ib) are straight lines. This implies that one can use an exponential function or, more specifically, the Shockley diode equation to describe these two current components. On the other hand, these two lines are not parallel as one would expect from a normal BJT. This fact renders the concept of a constant current gain β_F meaningless, which is why some previous research has attempted to use a current-dependent β_F to describe the HBT

$$\beta_F \propto I_C^N$$

where N is a constant normally less than 1, and dependent on the process [7]. Based on equations (1) and (2) given earlier, the N in [7] is found from the following expression

$$N = \frac{n_{be,b}}{n_{be,c}} - 1$$

Since the BC junction is a homojunction, the reverse characteristics can be described by the same method as a normal homojunction BJT with a reverse current gain β_R. It is not necessary to simulate the BC junction as a temperature-dependent diode since it does not affect the forward active region which is the most important operation region. For most HBTs β_R is very low and less than 0.01. This is caused by the fact that the reverse injection from base to emitter is suppressed by the BE heterojunction.

The model also includes several parasitic resistances, Rbx, Re and Rc, where Rbx and Re are mainly the contact resistance for the base and emitter and Rc is the combination of the bulk resistance and contact resistance for the collector region.

The dynamic part of the model is simulated by a nonlinear charge component Qb which represents the charge storage in the neutral base region. When the device is forward-biased, the capacitances related with this charge are the diffusion capacitance for the BE junction and the depletion capacitance for the BC junction. Since some base parameters such as base width are a function of Vbc, the charge is also modulated by Vbc. We can use a simple approximate function which is obtained from the measured small-signal data (multi-bias S-parameters) to simulate the charge components in the forward bias region:

$$Q_b = f(V_{be}, V_{bc}) \quad (16)$$

where the function can be constructed from measured data. For the simplest case, the charge can be assumed to be partitionable and represented as Cbe and Cbc. The new model is a quasi-static model and, if desired, a carrier-base-transit time delay can be included into the charge term.

It is worth noting that this new model is valid not only for the HBT, but also for modern BJTs, as they can exhibit similar departures from normal behaviour such as non-uniform current gain and self-heating effects.

The whole parameter set can be extracted by a simple step by step process involving DC and broadband small-signal S parameters measurements.

VERIFICATION OF THE NEW HBT MODEL

The full parameter extraction process has been carried out for several AlGaAs/GaAs HBTs obtained from Siemens and Rockwell. Before the small signal measurement was carried out, a dummy structure was measured to enable de-embedding of the probe pads [8].

Model verification is separated into three parts: DC, broadband small-signal S parameters and large-signal power sweep measurement for two different HBT from Siemens and Rockwell. The parameters of the model are fixed for all three situations so the results are consistent. The simulation was carried out on HP MDS and the model can also be added to SPICE without any difficulties.

For the DC situation, we use the parameters extracted by our method to perform the simulation and then compare them with the measurement results (Figure 2). The results show very good agreement. For the same parameters and model, we use HP MDS to generate the small-signal S parameters, and these are compared with the measured results in Fig.3, 4 and again show good agreement at all bias points. The final and most important step to verify the model is using the power sweep measurement. Since all the model parameters were extracted from DC and broad band S parameter measurements, the 50Ω amplifier power sweep is an independent verification method to validate the model. The measurement and simulation results give excellent agreement as shown in Figure 5 for Siemens and Figure 6 for Rockwell HBTs. It must be emphasised that no post adjustment of the large-signal model was made to improve the fit in the data comparisons shown.

Hence, the final HBT model is valid for DC, small-signal and large-signal conditions as evidenced by these verification tests. It is a unified model, i.e. it uses one parameters set for all operating conditions, which can work well for nearly all situations. The final test of the model is illustrated in Figure 7 & 8, which shows that the bias point shift with varying input power level has been predicted quite accurately for both Siemens and Rockwell HBTs.

CONCLUSION

A new HBT large-signal model has been developed which includes self-heating effects. The new model is based on experimental observation and physical analysis which leads to a very direct parameter extraction process. The introduction of a new topology to split the base into several regions allows the variation of the base and collector currents to be tracked very accurately. A thermal sub-circuit which includes a simplified *compositional thermal resistance and computational thermal capacitance* to simulate self-heating effects. This model is a practical model and can be added to the commercial softwares such as SPICE, HP MDS and Harmonica without undue difficulty.

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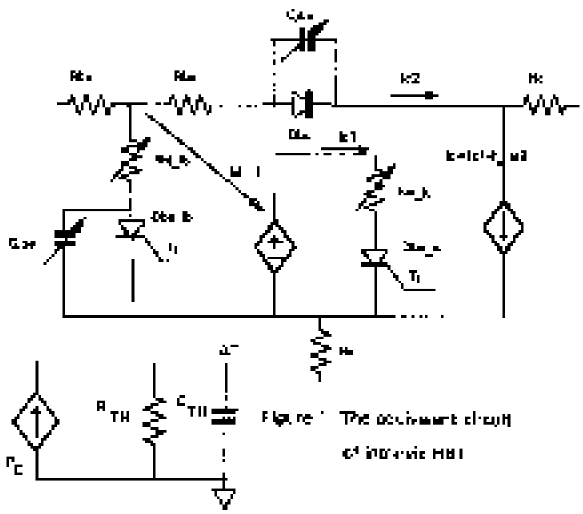

Figure 1. The equivalent circuit of intrinsic HBT

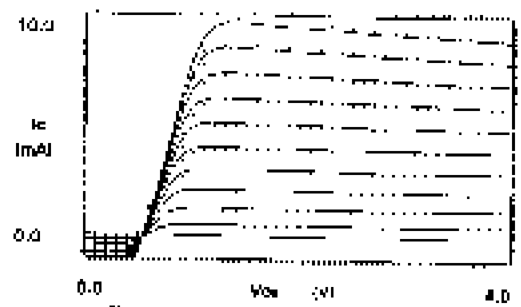

Figure 2. Measured (circles) and simulated (solid line) DC output characteristics of HBT

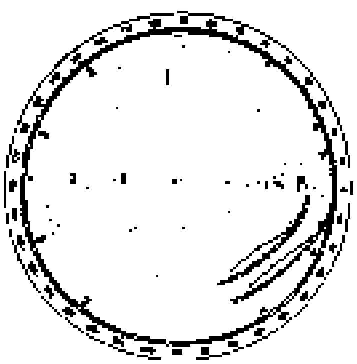

Figure 3. Comparison of measured S11, S22 (circles) and simulated S11, S22 (solid lines) for Siemens HBT

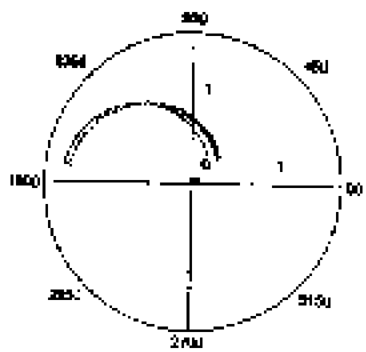

Figure 4. Comparison of measured S21, S12 (circles) and simulated S21, S12 (solid lines) for Siemens HBT

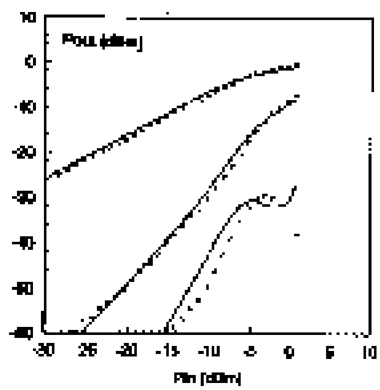

Figure 5. Comparison of measured (circles) and simulated (solid lines) first three harmonic components for Siemens HBT

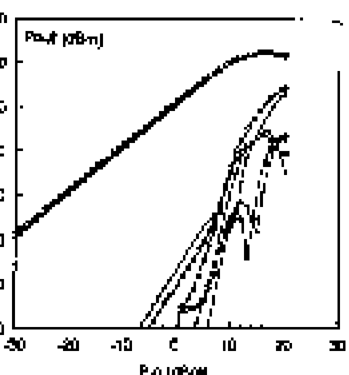

Figure 6. Comparison of measured (circles) and simulated (diamonds) first four harmonic components for Rockwell HBT

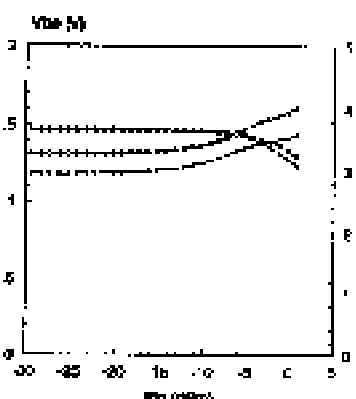

Figure 7. Comparison of measured (circles) and simulated (diamonds) Vbe and Ic vs. Pin for Siemens HBT

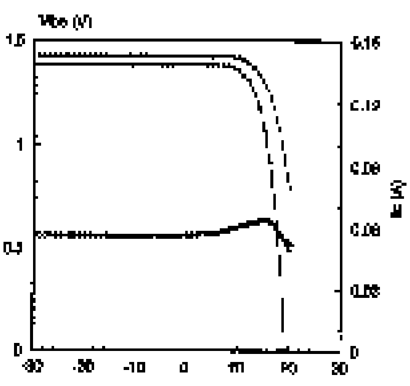

Figure 8. Comparison of measured (circles) and simulated (diamonds) Vbe and Ic vs. Pin for Rockwell HBT