A PHYSICS-BASED HETEROJUNCTION BIPOLAR TRANSISTOR MODEL FOR INTEGRATED CIRCUIT SIMULATION

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ABSTRACT

A new physics-based dc model for a heterojunction bipolar transistor (HBT) has been developed. This model offers features not found in previous analytical or physics-based HBT models, such as consideration of a cylindrical emitter-base geometry and direct implementation into SPICE (Simulation Program with Integrated Circuit Emphasis). The model parameters are determined from a knowledge of the device material, geometry, and fabrication process. The HBT model is developed by using semiconductor physics to calculate modified parameters for the existing SPICE bipolar junction transistor (BJT) model. Model generated data was obtained via a SPICE dc simulation for devices with different geometries and doping profiles. The resulting current-voltage characteristics were compared to the corresponding measured curves and excellent agreement was obtained. The model predicted the device's performance over its entire dc range of operation to within $\pm 5\%$. Although a few of the model parameters were determined empirically, the physical nature of the model provides insight into new device designs by directly relating the material, geometry, and process specifications to the model parameters.

I. INTRODUCTION

The design and fabrication of heterojunction bipolar transistors (HBTs) has received increased attention in recent years. The interest is due primarily to the significant performance improvement that can be obtained from HBTs compared to traditional bipolar

junction transistors (BJTs). The most technologically mature HBTs are fabricated from $\text{Al}_x\text{Ga}_{1-x}\text{As}/\text{GaAs}$, although other III-V compounds have been used. Higher speed devices with greater efficiency are the most notable advantages offered by HBTs. The HBT is an attractive device for high-performance military and commercial integrated circuits (ICs).

Simulation is crucial to the cost-effective design and fabrication of integrated circuits. Several authors have developed physics-based [1]-[4] and empirically curve fit [5]-[7] HBT models. Minimizing, or even eliminating curve fit parameters provides more insight into device and IC design. Moreover, by modifying the existing SPICE BJT model parameters, circuits can be designed with HBTs without procuring additional computer-aided design (CAD) tools or software updates. A methodology to determine physics-based HBT model parameters for implementation in the widely-accepted IC simulator, SPICE, is reported in this paper.

II. ASSUMPTIONS

This paper considers $\text{Al}_x\text{Ga}_{1-x}\text{As}/\text{GaAs}$ HBTs and the corresponding material parameters and expressions that are unique to $\text{Al}_x\text{Ga}_{1-x}\text{As}/\text{GaAs}$ semiconductors. Most HBTs are fabricated from these materials; however, the proposed methodology is applicable to other materials if the material constants are known. The approach further assumes the following: 1) The model can be represented by the dc SPICE BJT equivalent circuit topology. 2) Carrier transport across the base-emitter heterojunction is character-

ized by drift-diffusion and not by thermionic emission. 3) Standard, non-degenerate Boltzmann statistics apply. (Despite the fact that the GaAs base of a typical HBT may be degenerately doped, this assumption is made as a starting point. If the Boltzmann approximation is suspected to hinder model accuracy, then this assumption can be reconsidered). 4) There is uniform doping in the emitter, base, collector, and subcollector regions (i.e., no built-in drift fields). 5) Carrier mobility in AlGaAs can be sufficiently approximated by using the empirical mobility expressions for GaAs. 6) The base-emitter junction and contacts have a cylindrical geometry (i.e., emitter dots as compared to the typical emitter stripes). The proposed model does not include the effects of temperature, and assumes device temperature is constant at 300 K. This assumption is generally valid for low collector current density. At high collector current densities, a departure of the model data from the measured data due to device self-heating is expected, and will be readily identifiable. In a junction transistor, all significant effects are one-dimensional; thus, the model will be one-dimensional.

The model's simplicity is demonstrated through direct implementation in SPICE. The equivalent circuit model parameters are calculated using semiconductor physics and a knowledge of: a) material properties and related expressions, such as carrier mobility, lifetime, intrinsic carrier concentration, bandgap, and permittivity, b) device geometry, such as junction area and configuration of contacts, and c) process parameters, such as doping profile, Al mole fraction, and layer thicknesses.

Mathcad 3.1 [8] software was used to solve the semiconductor physics equations determining the SPICE model parameters, which were directly included in the SPICE BJT model statement. The version of SPICE used to simulate the developed model is Meta-Software's HSPICE version H92 [9]. Axum 3.0 [10] data analysis package was used to compare the measured and modeled data.

III. DC MODEL DEVELOPMENT

The classic Ebers-Moll model [11] for a junction transistor can be regarded as two back-to-back diodes. The SPICE BJT model expands upon the basic Ebers-Moll model by adding resistors, capacitors and diodes. Recombination currents, which can be significant in HBTs, are not accounted for in the basic Ebers-Moll model. These recombination currents are components of the base current, I_B , and are primarily the surface, space charge region (SCR), and

neutral base recombination currents. A composite recombination current, which is dependent on the junction voltage, can be modeled with a non-ideal diode (represented analytically by a non-unity emission coefficient). In order to model the behavior of the transistor in all modes of operation accurately, a recombination current diode is needed for each junction. The complete dc equivalent circuit is shown in Fig. 1 [12], where $V_{B'E'}$ and $V_{B'C'}$ are the intrinsic voltages seen by the base-emitter and base-collector junctions, C_{je} and C_{de} are the base-emitter depletion and diffusion capacitances, and C_{jc} and C_{dc} are the base-collector depletion and diffusion capacitances. The intrinsic voltages, $V_{B'E'}$ and $V_{B'C'}$, must be differentiated from the applied voltages V_{BE} and V_{BC} due to the voltage drop across the base (R_B), collector (R_C), and emitter (R_E) series resistances. The diodes I_1 and I_2 represent the currents through the base-collector and base-emitter junctions. I_3 and I_4 are the net base-collector and base-emitter recombination currents, and I_{CT} is the total current through the transistor from collector to emitter. The currents I_1 , I_2 , I_3 , I_4 , and I_{CT} are given by the following relations [13]:

$$I_1 = \frac{IS}{BR} \left[\exp \left(\frac{V_{B'C'}}{NR \cdot V_T} \right) - 1 \right] \quad (1)$$

$$I_2 = \frac{IS}{BF} \left[\exp \left(\frac{V_{B'E'}}{NF \cdot V_T} \right) - 1 \right] \quad (2)$$

$$I_3 = ISC \left[\exp \left(\frac{V_{B'C'}}{NC \cdot V_T} \right) - 1 \right] \quad (3)$$

$$I_4 = ISE \left[\exp \left(\frac{V_{B'E'}}{NE \cdot V_T} \right) - 1 \right] \quad (4)$$

$$I_{CT} = IS \left[\exp \left(\frac{V_{B'E'}}{NF \cdot V_T} \right) - \exp \left(\frac{V_{B'C'}}{NR \cdot V_T} \right) \right] \quad (5)$$

where IS is the transport saturation current, ISC and ISE are the base-collector and base-emitter recombination saturation currents, BF and BR are the ideal maximum forward and reverse common-emitter current gains, NF and NR are the forward and reverse current emission coefficients, NC and NE are the base-collector and base-emitter leakage emission coefficients, and V_T is the thermal voltage.

Semiconductor device physics can be used to determine values for each parameter excluding the emission coefficients. Because of their complex dependence on the device material, geometry, and fabrication process, accurate values for the emission coefficients are better determined through some form of parameter extraction or curve fitting.

The HBTs modeled in this effort have emitter dots rather than stripes which are found in more common transistor designs. Figures 2 and 3 show the layout of one of the HBTs. In particular, the layout shows five three-micrometer diameter emitter dots on one base finger (3 μ 5d1f). Thus the expressions for calculating the base series resistance, R_B , had to be reconsidered for cylindrical geometry [13].

The doping profile in Table 1 shows the layer structure for one of the three devices modeled. The other two devices were fabricated to Table 1 with the following exceptions: the 3 μ 1d1f device has a collector doping concentration of $2 \times 10^{16} \text{ cm}^{-3}$; the 2 μ 6d2f device has a collector doping concentration of $1 \times 10^{17} \text{ cm}^{-3}$ and a collector thickness of $3.5 \times 10^{-5} \text{ cm}$. The intrinsic AlGaAs buffer layer between the GaAs substrate and the GaAs subcollector serves primarily to minimize substrate leakage. The p-type base dopant is carbon.

The devices were fabricated via metalorganic chemical-vapor deposition (MOCVD). Thus the base-emitter junction was unintentionally graded over approximately 150 Å. Assuming that this graded region is fully within the emitter, the conduction band spike characteristic of abrupt AlGaAs/GaAs heterojunctions is almost completely eliminated. The resulting metallurgical base-emitter junction can be treated as a homojunction where carrier flow is assumed to be by drift-diffusion process. Because the Al_{0.35}Ga_{0.65}As emitter still exists beyond the grading, the effective bandgap difference, ΔE , between the emitter and base materials remains unchanged.

Another parameter that is dependent upon both the material and fabrication process is the specific contact resistance, ρ_c . For the devices modeled, there is a different ρ_c value for contacts to n-type and p-type semiconductors: $\rho_{cn} = 1 \times 10^{-6} \Omega \cdot \text{cm}^2$ and $\rho_{cp} = 5 \times 10^{-6} \Omega \cdot \text{cm}^2$. Au was used to contact the emitter semiconductor, Au/Ge/Ni was used to contact the collector, and Ti/Pt/Au was used to contact the base semiconductor.

IV. RESULTS AND DISCUSSION

After determining the device's physical characteristics solely from its material, geometry and fabrication process, SPICE model parameters were calculated to accurately predict the device's electrical performance. The dc steady-state model is fully characterized by the topology of Fig. 1 and by Eqs. (6) and (7), which are the HSPICE expressions for the collector

and base terminal currents, respectively [9]:

$$I_C = I_{CT} - (I_1 + I_3) \quad (6)$$

$$I_B = I_1 + I_2 + I_3 + I_4 \quad (7)$$

Consider the common-emitter current-voltage (I-V) characteristics. The series resistances, though not shown in Eqs. (6) and (7), directly influence the junction voltages, $V_{B'E'}$ and $V_{B'C'}$. Both R_E and R_C affect the slope of the I-V curves in saturation. While in saturation, an increase in R_C means a higher V_{CE} needs to be applied to maintain the same I_C . The transistor will remain in saturation at higher V_{CE} (a decreased slope of I_C vs. V_{CE} through saturation region) since more voltage is dropped across a larger resistor, decreasing the voltage seen by the internal collector node, $V_{C'}$. The active mode is reached when the base-collector junction becomes reverse biased ($V_{C'} > V_{B'}$).

A similar scenario occurs when R_E is increased. Given a constant I_B for each curve, to maintain the same I_C , the potential at $V_{E'}$ must increase due to the increased voltage drop across R_E . Consequently, the input voltage V_{BE} must increase to maintain the same $V_{B'E'}$. Kirchhoff's voltage law then demands that the output voltage V_{CE} also increase resulting in a lower I_C slope through saturation region. The slope is more sensitive to changes in R_E and decreases linearly with increasing R_C or R_E .

R_B has no effect on the I-V curves. Since I_B is a constant for each curve, any change in R_B results in a corresponding change in V_{BE} necessary to maintain the proper $V_{B'E'}$ and $V_{B'C'}$ relationship.

a) 3 μ 1d1f and 3 μ 5d1f Device Results

The comparison between modeled and measured common-emitter I-V characteristics for the 3 μ 1d1f device is shown in Figs. 4 and 5. The set of HSPICE parameters used to generate the model data of Figs. 4 and 5 is shown in Table 2. In the devices investigated in this research the base-collector junction is a homojunction, and thus NR was assumed to be close to unity. Although HSPICE was not used to optimize any model parameters, BR and NF were manually fit to best match the measured data. First, BR was chosen small enough to provide very good agreement in the reverse active region. Next, NF was chosen to best match the data in the saturation region and at the offset voltage.

The 3 μ 1d1f device has very good agreement in the forward and reverse active regions. The slight discrepancies in the saturation region may be related to

the bias dependence of R_C . Increasing R_C will decrease the slope of each I-V curve. However, HSPICE does not consider that the actual value of R_C changes with $V_{B'C'}$. To do this, HSPICE would require information on the doping and thickness of the collector layer. Instead, HSPICE regards R_C as a constant. Thus, the slopes of all the curves change by the same amount when R_C is changed (i.e., independent of bias). Carrier recombination is another phenomenon that effects the slopes of the curves in saturation, but only at low bias. If a sufficient number of data points were recorded, the measured curves for the few lowest I_B values would show a distinctly lower slope than the corresponding modeled curves. This is due to recombination hindering a faster rise in I_C at the lower bias levels. Such bias dependence is not modeled by the constant empirical BF ; therefore, the modeled curves have steeper, bias insensitive slopes. Another bias dependence not accounted for with the constant empirical BF is that of the offset voltage. The offset voltage of the modeled curves increases with increasing bias. The proper relationship has offset voltage inversely proportional to bias [5].

The same modeling procedure was performed on the 3 μ 5d1f device. The resulting I-V characteristics are shown in Figs. 6 and 7, and the model parameters are listed in Table 3. Because the emitter dot diameter is the same for both the one dot and five dot devices, R_E and R_B for the 3 μ 5d1f device are exactly one fifth of the respective 3 μ 1d1f values. This is because each of the dots are in parallel. R_C has a more complex geometry dependence, but is approximately one fifth of the 3 μ 1d1f value for the same reason. The 3 μ 5d1f saturation current is larger than the value calculated for the 3 μ 1d1f device primarily due to the larger junction areas.

The effects of device self-heating are readily seen in the highest four curves at $V_{CE} \geq 3$ V. In fact, the top curve ($I_B = 500\mu A$) is effected almost immediately after entering the forward active region. Like the 3 μ 1d1f device, the 3 μ 5d1f device model slightly overestimates the active region collector current for the lowest four curves ($50\mu A \leq I_B \leq 200\mu A$). This is due to the model having a constant BF that was chosen to match the near constant empirical current gain of the highest six curves ($250\mu A \leq I_B \leq 500\mu A$).

Overall, the physics-based models for the 3 μ 1d1f and 3 μ 5d1f devices provide very good correlation with the measured data. The 3 μ 1d1f model overestimates I_C by less than 4% in the reverse active region, (0-0.1 V). Excellent agreement is obtained in the forward active region (1-2 V) with a difference of less

than 5%. The overestimation at the two highest V_{CE} values (3 and 4 V) is due to transistor self-heating which reduces the measured I_C values.

The 3 μ 5d1f dc model performs just as well as the 3 μ 1d1f model. In the reverse active region (0-0.1 V), the model overestimates I_C an average of less than 5%. Excellent correlation is obtained in the forward active region (1-2 V) with an average difference of about 5%. As with the 3 μ 1d1f device, the 7% average difference at 3 and 4 V can be attributed to transistor self-heating.

b) 2 μ 6d2f Device Results

Accurate physics-based modelling of a non-constant current gain transistor is a difficult task primarily because the I-V curves are highly sensitive to NE and NC . The default values of ISE and ISC in SPICE are zero. This means that unless values of ISE and ISC in the range of 10^{-14} to 10^{-20} A are put in the model card, no recombination current will be modeled. Since the forward common-emitter current gain of the 2 μ 6d2f device was not constant, unlike for the 3 μ 1d1f and 3 μ 5d1f devices, the recombination currents had to be considered [13].

To develop a robust dc model for the 2 μ 6d2f device, the optimization function of HSPICE was used to curve-fit the measured data and simultaneously determine values for NF , NE , NC , ISE , and ISC . Measured data points that deviated noticeably from the constant I_C line due to device self-heating were not included in the optimization so as not to confuse HSPICE or corrupt the temperature independence of the model. The I-V characteristics for the 2 μ 6d2f device model that considers the bias dependence of recombination using NE , NC , ISE , and ISC are shown in Figs. 8 and 9. The corresponding model parameters are listed in Table 4. Notice that IS is several orders of magnitude smaller than either ISE or ISC . This is expected since the recombination saturation currents are significant in most HBTs.

Notice in Fig. 9 that the modeled curves show a decreasing offset voltage with increasing I_B . Since the recombination currents have less impact with increasing bias, the base-emitter junction is able to turn-on more quickly, hence lowering offset voltages. The modeled offset voltage has not only the proper bias dependence, but the modeled and measured offset voltages are nearly identical for each curve. This distinct improvement over the 3 μ 1d1f and 3 μ 5d1f models is directly attributed to modeling the bias dependence of recombination with NE , NC , ISE , and ISC .

The agreement in the forward active region (1-2 V) is excellent due to the curve fitting of measured data to extract the three emission coefficients and recombination saturation currents: NF , NE , NC , ISE , and ISC . The slight disagreement of the model through saturation ($0.2 \leq V_{CE} \leq 0.7V$) is indicative of the difficulty in precisely matching the variable saturation slope due to the bias dependence of R_C . Excellent agreement is obtained as the model leaves the saturation region and begins to enter the active region (0.8 and 0.9 V) with an average difference in collector current of less than 5%.

Unfortunately, there is no good way to calculate the HSPICE BJT model emission coefficients (NF , NE , NC) physically. These three values which are so crucial to the ultimate fit of the model can only be determined via parameter extraction. The dc models derived in this paper are physics-based, with the exception of BF , NF , NE , NC , ISE , and ISC . The series resistances (R_B , R_C , and R_E) and transport saturation current (IS) are determined only from a knowledge of the device material, geometry, and fabrication process. Aside from the agreement in saturation (which can be imputed to the bias dependence of R_C and recombination), the $3\mu 1d1f$ and $3\mu 5d1f$ models fall within the $\pm 5\%$ difference criterion for model success. This level of performance is at least as good as models found in current literature [1,2,4,5]. The $2\mu 6d2f$ model has good agreement in the reverse active region, good agreement in the saturation region, and excellent agreement in the forward active region (neglecting differences due to self-heating). Additionally, the models have proven to support a cylindrical emitter-base geometry, and are readily implemented in SPICE.

V. CONCLUSION

The objective of the research presented in this paper was to derive a physics-based model for an HBT that would accurately predict the device's dc electrical behavior. A model that considers a cylindrical emitter-base geometry was directly implemented into SPICE. Using semiconductor physics, the device model parameters were determined from a knowledge of the device material, geometry, and fabrication process.

A Mathcad file was written, which calculates all of the necessary SPICE BJT model parameters required to model an HBT accurately. The dc model was successful at producing data that was within $\pm 5\%$ of the measured data. Various integrated circuit components can now be designed using the developed physics-based SPICE HBT model. Although a few

of the model parameters were determined empirically, the physical nature of the model provides insight into new device designs, by directly relating the material, geometry, and process specifications to the model parameters.

ACKNOWLEDGEMENT

This research was sponsored by the Solid-State Electronics Directorate, Wright Laboratory, Wright-Patterson Air Force Base, Ohio 45433. The authors would like to thank B. Bayraktaroglu, C. Huang, and L. Liou for their fabrication and the dc measurements of the devices used in this research.

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Fig. 1. The complete dc equivalent circuit.

Fig. 2. Mask layout of a 3μ5d1f AlGaAs HBT.

Fig. 3. Side view of Figure 2 cut along AB.

Fig. 4. I-V characteristics for the 3μ1d1f device with I_B swept in 20 μA increments.

Fig. 5. Reverse I-V characteristics for the 3μ1d1f device with I_B swept in 20 μA increments.

Fig. 6. I-V characteristics for the $3\mu 5d1f$ device with I_B swept in $50\ \mu A$ increments.

Fig. 8. I-V characteristics for the $2\mu 6d2f$ device with I_B swept in $100\ \mu A$ increments.

Fig. 7. Reverse I-V characteristics for the $3\mu 5d1f$ device with I_B swept in $50\ \mu A$ increments.

Fig. 9. Reverse I-V characteristics for the $2\mu 6d2f$ device with I_B swept in $100\ \mu A$ increments.

Table 1. Layer Doping Profile for 3 μ 5d1f Device.

Layer	x	Doping (cm ⁻³)	Thickness (cm)
In _{0.5} Ga _{0.5} As cap	0.5	2×10^{19}	3×10^{-5}
In _{0.5} Ga _{0.5} As grading	0.5 $\rightarrow$ 0	1×10^{19}	2.5×10^{-6}
GaAs emitter	0	5×10^{18}	1×10^{-5}
Al _{0.3} Ga _{0.7} As grading	0 $\rightarrow$ 0.35	5×10^{17}	5×10^{-6}
Al _{0.3} Ga _{0.7} As	0.35	5×10^{17}	5×10^{-6}
GaAs Base	0	5×10^{19}	7×10^{-6}
GaAs Collector	0	8×10^{15}	1×10^{-4}
GaAs Subcollector	0	3×10^{18}	1×10^{-4}
Al _{0.3} Ga _{0.7} As Buffer	0.3	—	3×10^{-5}
GaAs substrate	0	—	5×10^{-2}

Table 3. 3 μ 5d1f SPICE Model Parameters.

Parameter	Units	Value Used in HSPICE BJT Model Card
BF	—	83.429
BR	—	0.0569
NF	—	1.1393
NR	—	1.0000
ISE	A	default 0
ISC	A	default 0
IS	A	1.1169×10^{-25}
RE	Ω	6.7341
RB	Ω	8.6569
RC	Ω	16.785

Table 2. 3 μ 1d1f SPICE Model Parameters.

Parameter	Units	Value Used in HSPICE BJT Model Card
BF	—	83.541
BR	—	0.1512
NF	—	1.1121
NR	—	1.0000
ISE	A	default 0
ISC	A	default 0
IS	A	2.2345×10^{-28}
RE	Ω	33.6707
RB	Ω	43.2846
RC	Ω	68.713

Table 4. 2 μ 6d2f SPICE Model Parameters.

Parameter	Units	Value Used in HSPICE BJT Model Card
BF	—	83.926
BR	—	0.6338
NF	—	1.1049
NR	—	1.0000
NE	—	1.7000
NC	—	1.7595
ISE	A	2.432×10^{-19}
ISC	A	4.393×10^{-18}
IS	A	1.111×10^{-25}
RE	Ω	6.7896
RB	Ω	4.8379
RC	Ω	2.20