AN EMPIRICAL HBT LARGE SIGNAL MODEL FOR CAD

I. Angelov[1], K. Choumei[2] and A. Inoue[2]

[1]Microwave Electronics Laboratory, Chalmers University of Technology, S-41296 Göteborg, Sweden,
[2]Mitsubishi Electric Corporation, 4-1 Mizuhara Itami Hyogo, Japan.

Abstract — **Extensive measurements were performed at different temperatures on GaAs HBT's in order to obtain a simple Large Signal Model usable in CAD. The model was experimentally evaluated by DC, S and Power Spectrum measurements and good correspondence was obtained between measurements and experiment.**

I. INTRODUCTION

Heterojunction Bipolar Transistors (HBT) have become very promising devices for different applications at microwaves and millimeter wave frequencies. An important condition for any successful design work is availability of an accurate large signal model (LSM). In recent years more than 100 publications in respectable journals were devoted on the creation of a compact model for use in CAD tools and procedures for extraction of parameters for these models [1-26]. Despite the tremendous work already done on the subject we still do not have unified, accurate model standard in industry together with the automatic extraction procedure. The reason for this is that the device physics is very complicated, the range of currents is broad, and power density at which the device operates is very high. All this together with poor thermal conductivity of the III-V materials makes the problem of the creation of universal LSM for HBT more difficult. Many of the existing HBT models are based on solid physical background, but again because of the difficulties of the problem they end up with many empirical coefficients, that are difficult to extract.

In some cases the HBT models can be " excessively" accurate and include effects that may not be so important and model simplification is appropriate in such cases. When the model is complicated, an additional difficulty to the extraction problems is that such a model could show problems with the convergence.

II. DEVICE MODELING

The HBT used in this study are AlGaAs/GaAs HBTs with a 4 x 20 um2 emitter [27].

Systematic DC, S- multi-bias measurements were performed at different temperatures (-25-+125C) in order to obtain a general picture of the device behavior. ICCAP (Agilent) was used in the measurement and extraction of basic parameters.

Figures 1-3 shows some results for I_b, I_{ce}, and Forward Gummel (FG) β obtained from these devices that are typical for many HBT. As can be seen, when keeping the I_b (or Ie) constant the voltage shift of the BE junction is almost linear and can be used to monitor the device temperature. Increasing the temperature will decrease the base voltage required to sustain the same current and the temperature coefficient T_cV_{bj} is~0.001. A problem with this type of transistors is that β cannot be considered constant, Fig. 3. The models considering β constant like Gummel-Poon model produce significant error and separate equations for I_b and I_{ce} should be used in the model. There is a change of the V_{be} at which we have maximum of β and this should be considered in the LSM.

The transistor can be described with a conventional equivalent circuit shown in Fig.5. Nonlinear are current sources I_{ce} and C_{be} and C_{bc}, the remaining elements can be considered linear and there is a significant amount of papers describing the extraction of the small signal equivalent circuit [14-26]. For devices exhibiting substrate effects an extra port should be added.

In order to use the fact that when keeping I_b constant the junction voltages varies linearly with the temperature, we can modify diode current definition. It is a common practice to describe the complicated I_b dependence by several diodes (respective exponential functions) in order to improve the accuracy and describe the different physical phenomena occurring in the device [1-12]. According to the device physics we use an exponential function to describe the semiconductor junction, but the reference point is changed to V_{bj} above the knee voltage at currents where we normally operate the device. In order to have a possibility to fit a variety of factors and effects influencing the junction current, we can describe the I_b as a power series:

$$I_b = I_{be} + I_{bc}$$

$$I_{be} = I_{jbe}(\exp(P_{be}) - \exp(P_{be0})), (1)$$

$$I_{bc} = I_{jbc}(\exp(P_{bc}) - \exp(P_{bc0})), (2)$$

$$P_{be} = P_{be1}(V_{be} - V_{je}) + P_{be2}(V_{be} - Vje)^2 \dots,$$

$$P_{bc} = P_{bc1}(V_{bc} - V_{jc}) + P_{bc2}(V_{bc} - V_{jc})^2 \dots \quad (3)$$

$$P_{be0} = -P_{be1}.V_{je} + P_{be2}.V_{je}^2 \dots,$$

$$P_{bc0} = -P_{bc1}.(V_{jc}) + P_{bc2}.(-V_{jc}^2) \dots (4)$$

This definition is equivalent to the conventional diode equation, but describes the currents with highest accuracy in the useful operating range, because extraction point is changed. When $V_{be}= V_{je}$ the base current I_{be} is equal to I_{jbe} and at $V_{be}=0$, $I_{be}=0$.Usually 1-3 terms of the power series are enough to obtain accuracy 2%. If higher accuracy is required more terms can be added to improve the fit at very small currents.

Using Eq.1-4 we can reduce the number of the diodes (exponents) and keep the accuracy. This in turn will improve the convergence of the large signal model. It is a common practice to limit the exponential growth of the current using a modified exponent (soft exponent). In this case this is not required any more, because we can limit the exponential growth of the current by selecting the coefficients of the arguments of Eq.3 in a proper way.

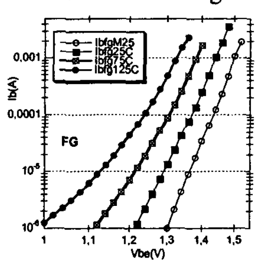


Fig.1. Ib FG

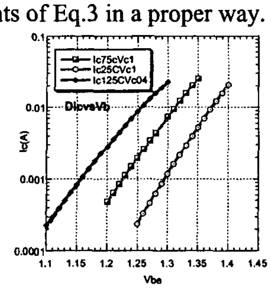


Fig.2.Ic vs. Vbe

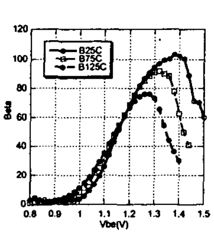


Fig. 3. FG β. vs. T

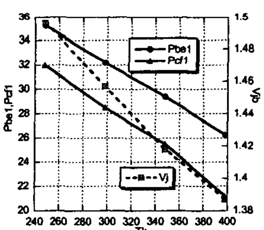


Fig. 4. Vjb, Pbe1,Pcf1 vs. T

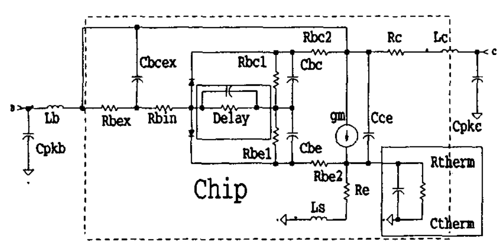


Fig. 5. Equivalent circuit

In Fig. 4. and table 1 are shown extracted parameters of the base part using one term in the power series of Eq. 1-3. The changes of the parameters are linear with respect the temperature and this can be use to model self-heating in a simple way using linear dependencies.

We can keep the basic form of the Gummel-Poon model [10] but define collector current source as a power series:

$$I_c = I_{pkc}(exp(P_{cf}) - exp(P_{cr}))(1 - \frac{V_{be}}{V_{ar}} - \frac{V_{bc}}{V_{af}})(5)$$

P_{cf} is a power series centered at V_{bep} and with a variable V_{be}, i.e.:

$$P_{cf} = P_{cf1}(V_{be} - V_{bep}) + P_{cf2}(V_{be} - V_{bep})^2 \dots (6A)$$

$$P_{cr} = P_{cr1}(V_{bc} - V_{bcp}) + P_{cr2}(V_{bc} - V_{bcp})^2 \dots (6B)$$

$$V_{bcp} = V_{bep} + \Delta V_{bep} * (1 + Tanh(P_{cf1} * V_{ce})$$

When $V_{be}=V_{bep}$ the collector current is equal to I_{pkc} and the transconductance is equal to $g_m=I_{pkc}.P_{c1}$.

Typically with 3 terms of the power series the error is less than 2 %. There is freedom for the selection of V_{bep} and I_{pkc}, but probably the best is to select V_{bep} equal to the voltage at which β is maximum and usually even a simple fitting program will converge and find the optimum values.

Normally the accuracy at negative V_{ce} voltages is not so critical and corresponding parameters from Eq.6A.and 6.B can be considered equal.

The collector current I_{ce} can be expressed also as:

$$I_c = I_{pkc}(exp(P_{cf}).\tanh(\alpha.V_{ce})(1 + \lambda.V_{ce})(5a)$$

$$\alpha = \alpha_r + \alpha_s.(\exp(P_{cf1}.V_{ce}) - 1)$$

This definition will behave better from numerical point of view, because *tanh* is limited function and at $V_{ce}=0$, $I_{ce}=0$.

The exponent's argument P_{cf} varies with collector voltage for number of reasons - change of the carrier velocity, temperature... This can be reflected using following approach:

$$P_{c1m} = P_{c1}(1 + B_1 / Cosh^2(B_2.V_{ce})(7)$$

The coefficient B_1 ($B_1 \approx 0.1$-0.3) is extracted at low collector voltages and $B_2 \approx 10$-30 is a fitting coefficient, which will determine how quickly the gain is changing when V_{ce} is increased.

HBT Capacitances.

Devices can operate at high currents and voltages and their nonlinear capacitances should include both depletion and diffusion parts. The total C_{be} capacitance shows rapid increase and then decrease at high bias. That is similar to what can be found in homo-junction transistors [7,14], but in HBT this increase will be larger than the increase we observe in the homo-junction.

Diffusion capacitances are described as [1]:

$$C_{bedif} = C_{bep} + C_{be0}.(1 + \tanh[C_{be10} + C_{be11}.V_{be}]), (8)$$

$$C_{bcdif} = C_{bcp} + C_{bc0}.(1 + \tanh[C_{bc20} + C_{bc21}.V_{bc}]), (9)$$

Integrating with the terminal voltages we can obtain the Q_{be} and Q_{bc} when using the charge definition:

$$Q_{bedif} = C_{be0}\frac{\log[\cosh[C_{be10} + C_{be11}.V_{be}]]}{C_{be11}} + (C_{bep} + C_{be0})V_{be}$$

$$Q_{bcdif} = C_{bc0}\frac{\log[\cosh[C_{bc20} + C_{bc21}.V_{bc}]]}{C_{bc11}} + (C_{bcp} + C_{bc0})V_{bc}$$

(10)

The ordinary equation for the depletion part can create convergence problems when $V_{bc}=V_{bci}$ and to avoid this the equation for the depletion capacitance equation is modified to:

$$C_{dep} = C_{depp} + C_{dep0}(x^2 + m)^{-n-1}(m - (2n-1)x^2)(11)$$

where $x=V_{bc}-V_{bci}$, ***n***- grading coefficient, ***m***- parameter determines the maximum- minimum capacitance ratio ($0<$ ***m*** <0.5). Eq.11 gives results identical to the textbooks depletion capacitance definition, but it is well defined at V_{bci}. The function in Eq. 11 is positive and without poles in the interval $-\infty < V_{bc} < \infty$. For the simulators using charge definition, a closed form of the charge can be derived from Eq.11 and combined with Eq.10.

Self-heating modeling

Since the temperature coefficients of different model parameters are small (in the order of 10^{-3}) [8], the changes of the model parameters vs. temperature can be considered linear in a limited temperature range (Eq.12). If required more terms can be added. The values of I_c and V_c are available in harmonic balance simulator and are used to calculate dynamically junction temperature T_j:

$$P_{tot} = I_c.V_c + I_b.V_b;\ T_j = P_{tot}.R_{therm} + T_{amb}\ (12)$$

$$I_{jbeT} = I_{jbe}(1 + T_{cIjbe}(T_j - T_r)), I_{pkcT} = I_{pkc}(1 + T_{cIpc}.(T_j - T_r))$$

$$PP_{beT} = PP_{be}(1 + T_{cPPbe}(T_j - T_r))$$

$$V_{jeT} = V_{jeT}(1 + T_cV_{je}.(T_j - T_r))$$

$$V_{bepT} = V_{bep}(1 + T_cV_{bep}.(T_j - T_r))$$

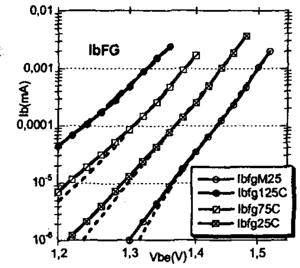
where P_{tot} is the dissipated power, R_{therm} is the thermal resistance. For higher accuracy R_{th} can be made temperature dependent.

$$R_{thT} = R_{th}(1 + T_{cRth}(T_j - T_r))$$

Evaluation of the model

The extraction of parameters starts with the extraction of the DC parameters. When using eq.5.a there are Many fitting programs (Mathematica, MathCAD, Kaleidagraph) can be used to extract basic parameters of the model. Fig.6, 7 show some results from measured and modeled I_b, I_{ce} (measured-points and modeled lines).

There is a large amount of papers on extraction of a small signal equivalent circuit [14-26] of HBT's. The model was implemented as a SDD in MDS (Agilent) with a self-heating and delay part and was experimentally evaluated using DC, S-parameter and Power Spectrum Measurements.



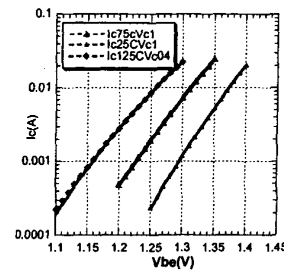
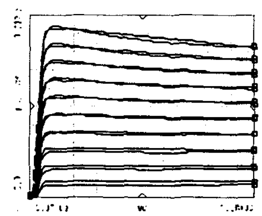


Fig.6. Meas. and modeled Ib Fig.7. Meas. and modeled Ice



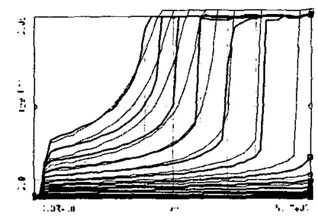
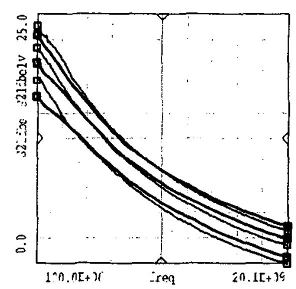


Fig. 8. Ice: Ib=20->200µA step 20µA. Fig.9. Ice,Vbe1.28->1.4



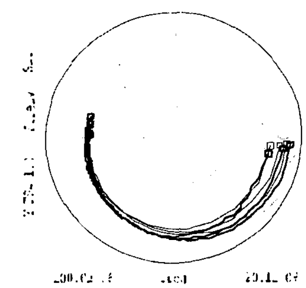


Fig. 10. S21 Vbe: 1.36- 1.38V Fig. 11. S11, Vbe: 1.36- 1.38V.

In Fig.8, 9 are shown measured and modeled I_{ce}. In the example shown in Fig. 8 the base current I_b is a parameter and in Fig. 9 V_{be} is a parameter. The fit is good even with such a simple definition of the model and using only 2 terms in the power series for I_{ce}. Despite the simplicity the model is able to describe the thermal runaway, Fig. 9.

Fig. 10,11 shows some results of S-parameter measurements and simulations. The model accurately describes the small signal behavior.

The large signal properties of the model were evaluated using a power spectrum (PS) method [28]. Fig. 12 shows results of the measurements and simulations and the fit is good.

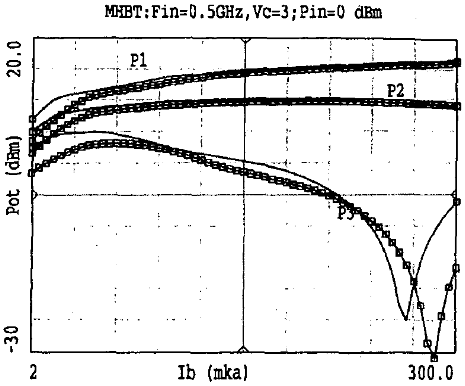


Fig. 12. Measured and modeled PS, Vce=3V,Ib->300 μA.

III. CONCLUSIONS

A simple bias dependent HBT model applicable to CAD was proposed and implemented. The model was experimentally evaluated with a DC; S-parameter and PS measurements and good correspondence was obtained between the measurements and the model.

ACKNOWLEDGEMENT

The authors wish to acknowledge Agilent and Ansoft Software for the donation of high-frequency simulation software, S. Maas for valuable discussions and A. Olaussen for the help.

REFERENCES

1. P. C. Grossman and J. Choma, Jr. "Large Signal Modelling of HBT Including Self—heating and Transit time Effects," IEEE Trans. MTT, Vol. 40, N.3, March 1992, pp.449-464.
2. Colin et al., "VBIC95, The Vertical Bipolar Inter Company Model," IEEE J. SSC, Vol. 31, N. 10, Oct. 1996, pp.1475-1483
3. R. Anholt, "Electrical and Thermal Characterization of MESFET's HEMTs, and HBTs," Artech House, Boston, 1995.
4. Ke Lu, P. Perry, T. Brazil "A new Large Signal AlGaAs/GaAs HBT Model including Self-heating Effects, With Corresponding Parameter Extraction Procedure," IEEE Trans. MTT, Vol.43, N.7, July 1995, pp.1433-1445.
5. Lovell, H Camnitz et al., "An accurate large Signal High Frequency Model for GaAs HBT, " GaAs IC Symposium, 1996, pp. 303-306.
6. UCSD Model, UCSD Electrical and Comp. Eng. Dep., 1995
7. S. Tiwari et al., "Analyses of the Operation of GaAlAs/GaAs HBT's," IEEE Trans. ED, Vol.36, N.10, 1989, pp.2105.
8. C. Wei. et al., "Large Signal Modelling of Self-Heating, Collector Transit Time and RF breakdown effects in Power HBTs," IEEE Trans. MTT, Vol. 44, N.12, Dec.1996, pp.2641.
9. Rudolph, R. Doerner, P. Heymann, "New GaInP/GaAs-HBT Large-Signal Model for Power Applications," EUMC1998, pp.231.
10. K. Gummel, H. C. Poon, "An Integral Charge Control Model of Bipolar Transistors, BSTJ Vol. 49,May, 1970,pp.827-852
11. G. M. Kull at. Al. IEEE Trans. ED, Vol. ED- 32, N.6, June 1985, pp.1103-1113.
12. J. Fossum, " Modelling Issues for Advanced Bipolar Device/ Circuit Simulations," Bip. Trans. Meetings, 1989, pp. 234-241
13. J. Scott, " Nonlinear III-V HBT Compact Models: Do we have what we need? " 2001 IEEE MTTS Digest, pp. 663-666.
14. S. Laux, " Technique for Small-Signal Analysis of Semiconductor Devices," IEEE Trans. ED, Vol.ED-32, N.10, Oct 1985, pp.2028-2037.
15. K. Joardar," A new Technique for Measuring Junction Capacitance in Bipolar Transistors," IEEE
16. H. C. Poon, H. K. Gummel, " Modelling of Emitter Capacitance," Proceedings of the IEEE, Dec. 1969, pp.2181.
17. R. Anholt, J. Gerber, R. Tayrany, et al.," HBT Model Extractor for Spice and Harmonic Balance Simulations," 1994 IEEE MTT-S pp. 1257.
18. S. Maas, D. Tait, " Parameter-Extraction Method for Heterojunction Bipolar Transistors," IEEE Microwave and Guided Wave Letters, Vol.2, N. 12, Dec. 1992, pp.502-504.
19. D. Pehlke, D. Pavlidis, "Evaluation of the factors Determining HBT High Frequency Performance by Direct Analysis of S-parameter Data," IEEE Trans, on MTT, Vol. 40, N. 12, Dec. 1992, pp. 2367-2373.
20. D. Dawson, A. Gupta, " CW measurement of HBT Thermal Resistance," IEEE Trans, ED, Vol. 39, N. 10, Oct. 1992, pp.2235.
21. W. J. Klosterman et al., " Improved Extraction of Base and Emitter Resistance from Small Signal High Frequency Admittance measurements," IEEE BCTM 1999 pp. 93-96.
22. HBT Static Model IC-Cap Users' Meeting 1996
23. D. Costa et al.," Direct Extraction of the AlGaAs/GaAs Heterojunction Bipolar Transistor Small-Signal Equivalent Circuit, " IEEE Trans, ED, Vol. 38, Sept1991, pp. 2018-2024.
24. P. Baureis, D. Seitzer," Parameter Extraction for HBT's Temperature Dependent large Signal Equivalent Circuit Model, " 1993 IEEE GaAs IC Symp, pp.263-267.
25. Y. Suh, D. Heo, A. Raghavan, E. Gebara et al.," Direct Extraction and Modelling Method for Temperature Dependent Large Signal CAD Model of Si-BJT," 2001 MTTS-Digest, pp.971-974.
26. M. Schröter et al., "Physics-Based Minority Charge and Transit Time Modelling for Bipolar Transistors," IEEE Trans. ED Vol.46, N2, pp.75.
27. T.Shimura et al., "High efficiency AlGaAs/GaAs power HBTs at a low supply voltage for digital cellular phones," IEICE Trans. Electronics, vol. E80-C, no.6, pp. 740-745, June 1997.
28. L. Bengtsson, C. Hednas, I. Angelov, "Verification of an HBT Gummel Poon model by Power Spectrum," EUMC 1996, pp.911.

Table1

I_{pkc} mA	P_{cf1}	P_{cf2}	V_{bep} V	D_{vpkc} V	α_r	α_s	B_1	B_2
5.3	28.2	-30	1.31	0.002	18	0.1	0.1	20

I_{jb} mA	P_{be1}	P_{be2}	P_{be3}	V_{jbe}	P_{bc1}	V_{jbc}
1.31	29.8	0	0	1.45	29.8	1.45