An Empirical Table Based HBT Large Signal Model

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Abstract — **An Empirical Table Based HBT Large Signal Model is proposed and experimentally evaluated. The important features of the model are that the model is a combination between an Empirical HBT model and a Table Based Model. The main model parameters are determined directly from measurements. The model was evaluated by DC, S-parameters, Power Spectrum and LSNA measurements. Good correspondence was obtained between measurements and simulations.**

Index Terms- Nonlinear models for active devices, nonlinear circuit design, microwave circuits.

I. INTRODUCTION

Heterojunction Bipolar Transistors (HBTs) have become very promising devices for different applications at microwave and millimeter wave frequencies. An important condition for any successful design work is the availability of an accurate large signal model (LSM). Generally, there are 3 types of models- Physics-based, Empirical and Table & Behavior models (TBM) [1-25]. Only models based on a solid physical background will describe HBTs accurately and will be easy to extract and understand. Often, because of the difficulties of the problem, they end up with many empirical coefficients, that are difficult to extract.

When the model is complicated, an additional difficulty to the extraction problems is that such a model could show problems with the convergence in circuit simulators.

II Large Signal Model Development

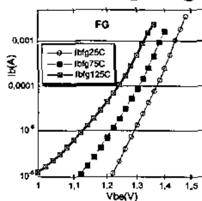

Fig.1. I_{be} FG

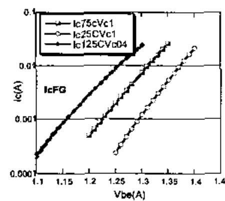

Fig.2.I_{ce} vs. V_{be}

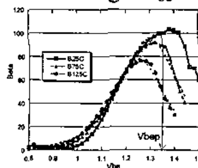

Fig. 3. FG β. vs. V_{be}&T

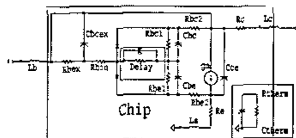

Fig. 4. Transistor EC

HBT Devices (GaAs and InGaP) from 3 foundries were studied in order to collect more data and to find a common strategy for optimum model structure and extraction procedure. In this paper some results of InGaP devices from Mitsubishi are shown. Multi-bias S-parameter and DC measurements were used to extract the model from devices under study. ICCAP (Agilent) was used in the measurement and extraction of basic parasitic parameters like resistances and capacitances and bias dependencies of C_{be}, C_{bc}.

Figures 1-3 show some typical results for measured Forward Gummel (FG) I_{be}, I_{ce}, and β obtained for 4x20 um finger devices from the DC measurements. When keeping the I_{be} (or I_c) constant, the voltage shift of the BE junction is almost linear and can be used to monitor the device temperature. This is a general feature for all bipolar devices. Increasing the temperature will decrease the V_{jbe} and the temperature coefficient is small, T_cV_{jbe} ~0.002.

The logarithmic plots for I_{ce} and I_{be} are close to a straight line, as should be from the device physics, and this means that the main function describing the device current should be exponential. The important inflection point for β maximum is used as a center plane for the model extraction.

The transistor can be described with a conventional equivalent circuit as shown in Fig.4. Nonlinear elements are the current source I_{ce} with transconductance g_m, and capacitances 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 [9-16].

We can simplify the self-heating modeling if we use the fact that the junction voltages vary linearly with the temperature, however this requires modification of the diode current definition. It is a common practice to describe the complicated I_{be} dependence by several diodes (respective exponential functions) in order to improve the accuracy and describe the different physical phenomena occurring in the device [1-8]. According to device physics we use an exponential function to describe the semiconductor junction, but the reference point is changed to the currents and respective voltages at which we normally operate the device, close to currents (voltages) at which β is maximum in Fig.3. In addition to the change of the reference point, the argument of the equation for the junction current can be described as a power series [23] or data set as it is in this work. This gives us a possibility to fit a variety of cases, many factors and effects.

A. Base current

The expressions for the base current are given by:

$$I_{be} = I_{bej}(\exp(P_{be}) - \exp(P_{be0})); I_{bej} = I_{pkc} / \beta_{\max}; (1)$$

$$P_{be} = (19.347 / N_{b1}).\tanh(2 \cdot [(V_{be} - V_{je})]);; (2)$$

$$P_{be0} = (19.347 / N_{b1}).\tanh(2 \cdot [-V_{je}]); ; (3)$$

$$P_{be1} = q_e / K_b T_{ambK}.N_{b1} = 1 / V_t.N_{e1} = 38.695 / N_{b1}); (4)$$

where q_e is the electron charge, K_b is the Boltzmann constant and N_{bl} is ideality factor. The coefficient $38.695 \cong 1/V_t$ at room temperature. When $V_{be} = V_{je}$ the base current I_{be} is equal to I_{jbe}; at $V_{be}=0$, $I_{be}=0$.

B. *Collector current*

The expressions for the collector current are:

$$I_c = I_{cf}.\tanh(\alpha.V_{ce})(1+\lambda.V_{ce}) \approx I_{cf}.\tanh(\alpha.V_{ce})e^{\lambda.V_{ce}}$$

$$I_{cf} = I_{pkc}(\exp(P_{cf}) - \exp(P_{cf0}))/\cosh(B_{be}(V_{be}-V_{bep}); (5)$$

$$P_{cf} = (19.347/N_{c1}).\tanh(2\cdot[(V_{be}-V_{bep})]); (6)$$

$$P_{cf0} = (19.347/N_{c1})\tanh(2\cdot[-V_{bep}]); (7)$$

$$a = \alpha_r + \alpha_s(1+\tanh(\frac{38.695}{N_{c1}}.V_{ce})); \ (8)$$

The parameters α_r, α_s reflect the dependence of the collector current at small collector voltages, parameter B_{be} in the term $1/\cosh(B_{be}(V_{be}-V_{bepm}))$ predefines the bell shape of β dependence vs. V_{be} and λ the output conductance dependence at high V_{ce}.

For devices with complicated doping profile, it is difficult to fit the IV characteristics and therefore, to speed up the extraction, the approach in [23] can be further extended. This can be done as the model is arranged as a mixed Empirical-Table Based Model (ETB). In this case the power series is replaced with a data set calculated from measured data. The basics of the Table-Based approach were given in works [20-22]. Generally, it is possible to combine the Table-Based approach and Empirical approach and extract the best from both [24].

Devices on the wafer usually have similar characteristics, but parameters depend on the device size. In addition the HBT, BJT characteristics are strongly dependent on the mounting conditions and users should be able to change some basic parameters like R_{therm}. There is a spread of parameters and it is good to have some flexibility to tune these basic parameters without re-measuring and recreating the complete model. This can be done with a proper arrangement of the ETB model giving access to important parameters like β, I_{pkc}, N_{c1}, N_{be1}, C_{be0}, R_{therm}, combining the Empirical and the Table-Based models with a possibility to tune these parameters and easily scale the model.

The envelope for the Table Based model is the Empirical model, implying that the problem with spline function selection and convergence is solved and correct exponential functions with proper derivatives are used. The model will be limited and valid out of the measured range, because the data set is limited from the measured data and linear extrapolation out of the measured data range will be adequate. In this case the model is transformed to simple, compact form:

$$I_{be} = I_{bej}(\exp(N_{b1}.(D_{sIb}-D_{vpk})); (9)$$

$$I_{bej} = I_{pkc}/\beta_{\max}; (10)$$

$$I_c = I_{cf}.\tanh(D_{Salpha})(1+(D_{s\lambda}).V_{cb}); (11)$$

$$I_{cf} = I_{pkc}(\exp((D_{sIc}-D_{vpk})/\cosh(D_{sBbe}); (12)$$

The convergence will be simplified, because the argument is calculated from measured data and a priory result in a limited function. All parameters are extracted, but the user has an access to use N_{b1}, N_{c1}, β_{max}, and D_{vpk} to fit the spread of these parameters. These several parameters are usually enough to handle practical cases.

The BJT HBT parameters are strongly temperature dependent and generally the data sets should be dependent on the voltages V_{be}, V_{ce} and temperature. This will work well, but for the extraction we will need measurements done at different temperatures. But even in the case we have the temperature controlled measurement system, processing 3 dimensional data I_{ce}= Ds (V_{be}, V_{ce}, T) is more complex than 2 dimensional I_{ce}= Ds (V_{be}, V_{ce}). That is why, it is beneficial if we can make transformation from 3 to 2 dimensional data set. This can be done using the junction property; i.e., the linear change of junction voltage vs. the temperature arranged as:

$D_{sIc}(T)=D_{sIc}-T_cV_{pk}*DT_j$, (13)

$D_{sIb}(T)=D_{sIb}-T_cV_{pk}*DT_j$, (14)

This is important part of the transformation. The V_{bep} and I_{pkc}, β are taken directly from the measurements. The best choice is to select V_{bep} equal to the voltage at which β is maximum, but usually even a simple fitting program will converge and extract the optimum values of I_{pkc}, V_{bep}, β_{max}, because of the presence of an inflection point.

The model structure is such, that data sets are usually dependent on one voltage V_{be} or V_{ce}. This simplifies the extraction, but to keep the structure of the data sets the same, the DS are arranged as dependent on V_{be}, V_{ce}. This approach leads to compact, bounded current model with correct derivatives without discontinuities which will converge using linear interpolation for the dataset.

C. *HBT capacitances*

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 [9,10,25], but in HBT this increase will be larger than the increase we observe in homo-junctions. It is a matter of tradition to treat the device capacitance as consisting of 2 parts, depletion and diffusion, connected in parallel described as[1-10,15-18, 23,25]:

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

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

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

$$C_{bc} \approx C_{dep} * (1+\frac{1}{\exp(V_{be}-V_{\min})} +$$

$$C_{dbe2}(\exp(\frac{38.695}{N_e}\tanh(V_{be}-V_{\min})); (15)$$

where $x=V_{bc}/V_{bci}$, n is the grading coefficient, m is a parameter that determines the maximum- minimum capacitance ratio. C_{depp} can be associated with parasitic elements.

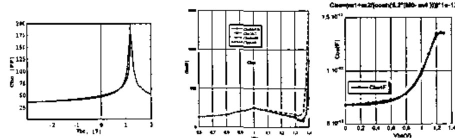

Fig.5 C_{bc} vs. V_{bc}; Fig.6 C_{bc} vs. V_{be} C_{be} vs. V_{be}

Figs. 5 and 6 show some results of capacitance extraction and a comparison between the measurements and simulations. As expected, the standard depletion capacitance model has a singularity at V_{bci}. The C_{bc} dependence is quite complicated- at some voltage will show minimum and then a sharp increase at high currents(V_{be}). The spline function for the capacitance models should be a function that has a natural peak. A good candidates for eq.15 is 1/cosh(Ds(Ax)):

$C_{be}=C_{bep}+C_{be0}/\cosh(DsCbe)$, (16)

$C_{bc}=C_{bcp}+C_{bc0}/\cosh(DsCbc)$; (17)

C_{bc} increases at voltages $V_{be}>V_{min}$. This approach corresponds to the device physics, it is stable in the extraction, because a limited function is used to provide the envelope for the data set. The fit of C_{be} is good using 1/cosh as envelope and models the decrease of C_{be} at voltages larger then V_{bci}. For faster execution function 1/cosh(Ax) can be replaced with: 1/(1+0.5Ax^2.

D. Self-heating modeling

Since the temperature coefficients of different model parameters are small (on the order of 10^{-3}) [3,19], the changes of the model parameters vs. temperature are considered linear in a limited temperature range ±100C. The dissipated power P_{tot} used to calculate dynamically the junction temperature T_j:

$$P_{tot} = I_{ce} * V_{ce}; V_{therm} = R_{therm} * I_{therm}; T_{chanK} = T_{ambK} + V_{therm}$$

$$V_{bepT} = V_{bepm} * (1 + T_{cVjbe} * (T_{chanK} - T_{refK}))...$$

(18)

where P_{tot} is the dissipated power, R_{th} is the thermal resistance. Thermal parameters are normalized to the reference temperature T_{ref} at which we extract parameters.

E. Delay modeling

The importance of the proper delay modeling was very well described in the classical papers of Lovell, Scott, Schrotter, Curtice, and Rudolf [5,8,15-18]. The simplified model of the delay can be replaced with the data set using controlling voltages V_{be},V_{ce}:

$$T_{ff}=T_f*(1+0.5.X_{tf}.(1+\tanh((38.68/N_{be1})(V_{be}-V_{bep}))))*(1-\tanh(V_{ce}/V_{TF})); \quad T_{ff}=T_f.Ds(V_{be},V_{ce}); \qquad (19)$$

where T_f is independent part of the forward delay, X_{tf} is the bias dependent part of the forward delay, V_{tf} is a fitting coefficient as they are defined in other commercial software packages and the user have access to TF. In ADS the bias dependent delay was implemented using the delay function of the SDD.

III. Extraction of the Model

The extraction of parameters starts with the extraction of the DC parameters. The basis of the proposed model is that the main parameters and DS are taken directly from the measurements. The I_{bej}, V_{je} and I_{pkc}, V_{bep}, β_{max} are taken from currents for maximum β typically at V_{ce} ~0.6-0.8V. The argument of the current functions is: $P_b=\ln(I_b/I_j)$; $P_c=\ln(I_c/I_{pc})$. The ideality factors N_{be1},N_{c1} are calculated from the derivative of $P_{c1}=38.695/N_{c1}$ at I_{be}, I_{ce} at V_{bep}.

The data set for B_{be} is found from β vs. V_{be} dependence, dataset for α at $V_{ce}<V_{kn}$. The output conductance parameter λ will fit the I_{ce} at voltages above the knee V_{kn} and this give the possibility to extract dataset for α, and λ separately. In ordinary cases it is not necessary to use a data set $\lambda=D_{sl}$ but when there is a breakdown, using dataset simplifies the model. There are many programs (Mathematica, MathCAD, Kaleidagraph, XLfit3,…) that can be used to extract basic parameters of the model.

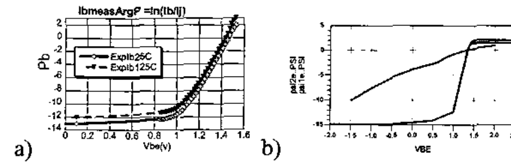

Fig. 7a Extracted argument P_b for I_{be} b Data sets DsIc, DsIb, Dsalpha.

There is a large amount of papers on extraction of the parasitic in the small signal equivalent circuit [10-18] of HBT's. Some of the parasitic like resistances, capacitances, used in this study are available directly from measured data from ICCAP. The rest of the parasitics are found with optimization using ADS in multiple bias S-parameter optimizations.

IV. Evaluation of the Model

The model was implemented as an SDD in ADS and was experimentally evaluated using DC, S-parameter and Power Spectrum Measurements (PSM). Figs. 8 and 9 show measured and modeled I_{ce}. In the example shown in Fig. 8 the base current is a parameter. Fig. 9 shows an important test; how the model describes the thermal runaway when device is biased with a voltage source. The good fit shows that simplification of the data set organization from I_{ce}=Ds (V_{be}, V_{ce}, T) to I_{ce}=Ds (V_{be}, V_{ce}) is adequate even with a simple linear model of the voltage shift of V_j due to the temperature change. If for some reasons a higher accuracy is required, the linear dependence can be replaced with a DS.

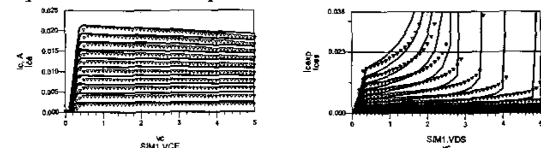

Fig. 8. I_{ce}: I_{be}=20-to200μA step 20μA Fig.9.I_{ce}, V_{be}>1.2 to 1.4V

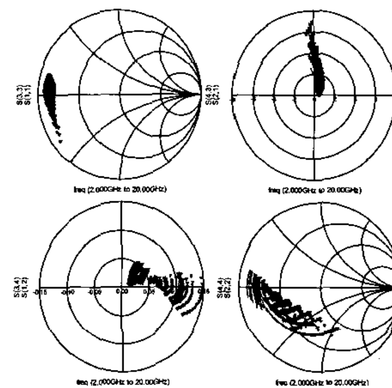

Fig. 10. S-parameters vs. Vce, Vbe parameter

Fig. 10 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 PS method and LSNA. Fig.11-13 shows some results of the measurements and simulations and the fit is good. The models for currents and capacitances facilitate the convergence in large signal applications, because of the use of well-defined functions and symmetry.

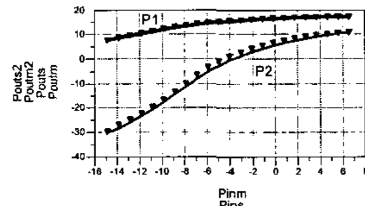

Fig.11. LSNA measured and modelled PS at 1GHz

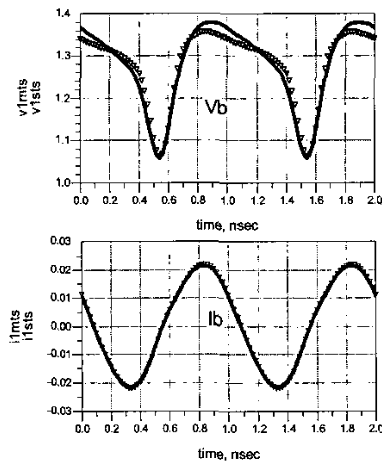

Fig 12 LSNA measured and modelled Vb and Ib at 1GHz

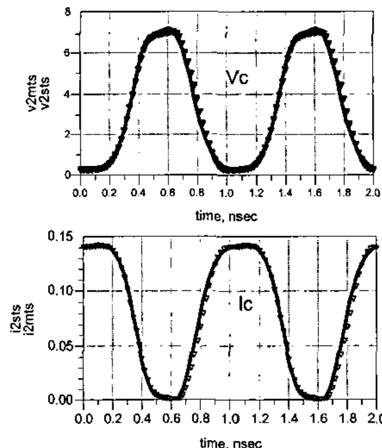

Fig 13 LSNA measured and modelled Vc and Ice at 1Ghz.

CONCLUSIONS

An Empirical Table Based HBT model was proposed and implemented. The model was experimentally evaluated with a DC, S-parameter, LSNA and PS measurements. Good correspondence was obtained between the measurements and the model.

ACKNOWLEDGEMENT

The authors wish to acknowledge A. Inoue, T. Hirayama, H. Zirath and E. Kollberg, the European NoE Target and SSF for the help and support of this work, as well as Agilent for the donation of high-frequency simulation software.

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