Modelling the Vertical PNP - Transistor using ICCAP and VBIC

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Abstract: This paper deals with modelling the vertical pnp-transistor (VPNP) using the VBIC95 model. The vertical pnp-transistor is an important element in modern BiCMOS technologies. It reaches higher transit frequencies and higher current driver capabilities as the usual lateral or substrate pnp-transistor. On the other hand, based on it's complicated structure, several parasitic effects may occur, which are investigated in this article. Furthermore, measurement and simulation results in comparison of the VBIC and the Siemens SQ3 model are presented.

[1] This paper is intended for the HP ICCAP User Meeting, June 1999 in Marseille,
Rev: UM150.DOC, 7.6.99, 8.6.99

1 Introduction

The VBIC model [1] became more and more public during the last years and several papers were published, referring to model the npn transistor using VBIC [3][4][5][10]. On the other hand, so far no results announced regarding to the pnp version of the model. Based on it's equivalent circuit, the VBIC model seems to be especially suitable to model the VPNP transistor. That is why, this paper is concerned with modelling the VPNP using VBIC in comparison to the Siemens SQ3 model. Section 2 of this paper outlines some details of the VPNP transistor and the appropriate measurement circuits. In Section 3, important features for the VBIC and the SQ3 model are explained. Finally, in Section 4 DC- and AC- measurement and simulation results are presented.

2 VPNP Transistor

2.1 Cross Section

In contrast to the well known integrated npn-transistor, the vertical bipolar transistor (VPNP) consists of five technological layers: p-emitter, n-base, p-collector, n-pocket and p-substrate (Fig. 1). The VPNP reaches higher transit frequencies and current driver capabilities as a lateral pnp- or substrate pnp-transistor. In comparison to the substrate pnp-transistor, the VPNP collector is not grounded, resulting in a more flexible usage in the circuit design.

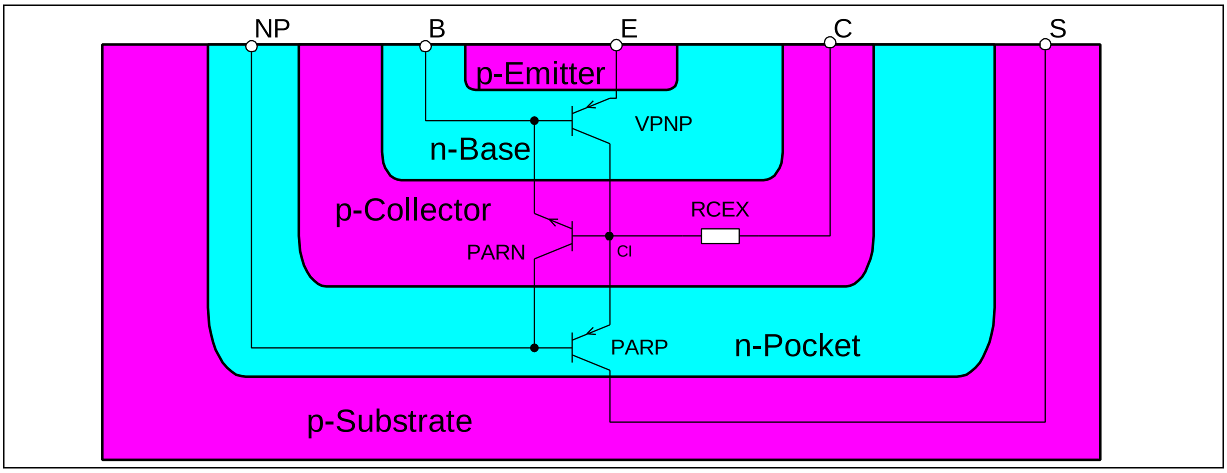


Fig. 1: VPNP cross section

Considering the VPNP structure it is obviously, that there are two parasitic transistors additional to the main VPNP transistor: the parasitic npn-transistor PARN (consisting of n-base, p-collector and n-pocket) and the parasitic pnp-transistor PARP (consisting of p-collector, n-pocket and p-substrate). Both these transistors, together with high collector resistance values, may cause several parasitic effects, explained in the next section in detail. The occurrence of these effects depends on technological parameters (e.g. collector sheet resistance), on geometrical parameters (transistor design) and the operation point conditions.

2.2 Parasitic Effects

2.2.1 Latch Up

If Latch Up occurs, we can observe a steep increase of both the n-pocket- and the substrate currents in the Gummel plot (Fig. 3). Here the inner R_C voltage drop biases the CN-junction in forward direction. The parasitic npn-transistor (PARN) enters the active inverse mode, it's emitter current gains the VPNP base current. Both the collector current and the R_C voltage drop increase. This results in a positive feedback.
PARN creates the n-pocket current, whereas the parasitic pnp-transistor (PARP) creates the substrate current, working in the active forward mode. Both currents are limited only by external sources. That is why, the operating points of an integrated circuit may be affected, if Latch Up occurs.
A high collector resistance value is a prerequisite for Latch Up. It is caused by the BC-depletion layer extension into the collector. Consequently Latch Up appears only at high collector voltages (Fig. 4).

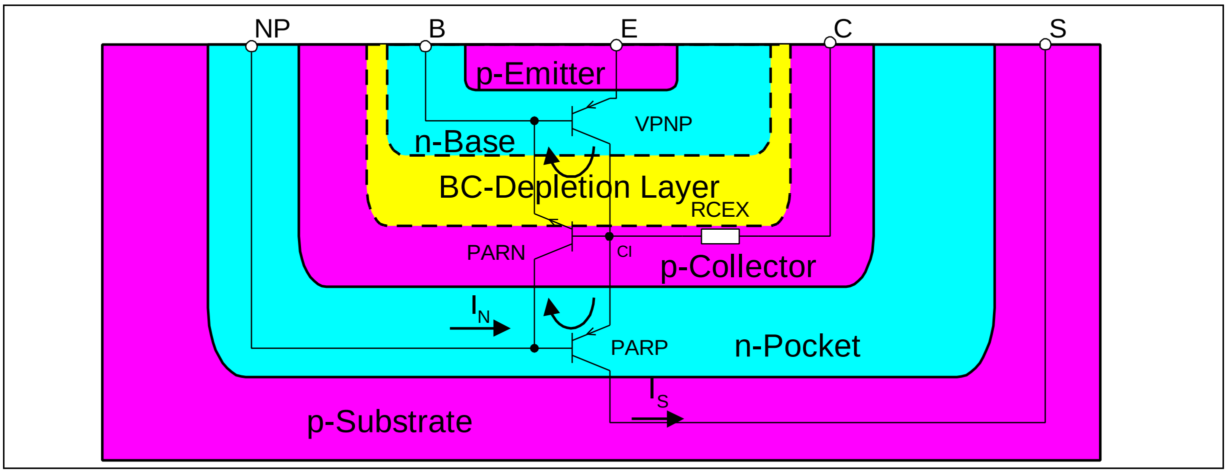


Fig. 2: Latch UP, RC is pinched by the BC-depletion layer

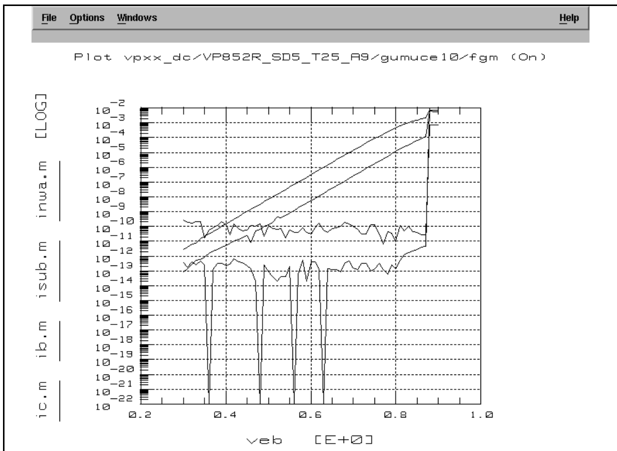


Fig. 3: Gummel Plot I_C , I_B , I_N , I_{SUB} = f (V_B), V_N = V_C + 4 V, V_C = - 10 V,
Latch Up is indicated by the steep increase of both I_N and I_{Sub} at V_{EB} = 0.85 V

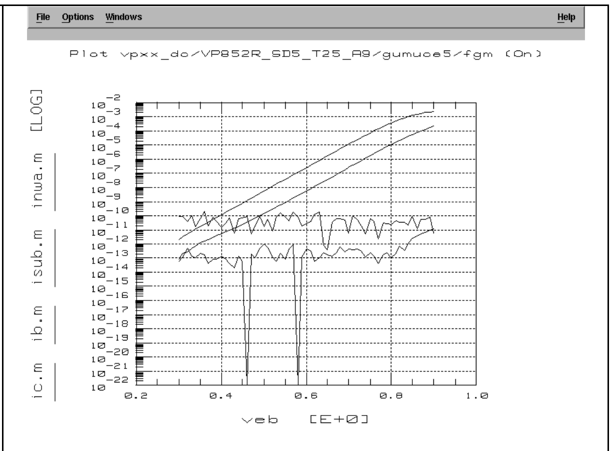


Fig. 4: Gummel Plot I_C , I_B , I_N , I_{SUB} = f (V_B), V_N = V_C + 4 V, V_C = - 5 V,
No Latch Up occurs at lower V_C

2.2.2 Collector Current Compliance

Another parasitic effect is the collector current compliance. This effect may occur, if the CN-depletion layer extension into the collector is high enough to pinch off the collector resistance. The current path through the collector is pinched off similarly to a JFET. Consequently the collector current is limited. Fig. 5 illustrates the situation in the cross section.
Fig. 6 shows a typical output characteristic [15]. This effect may be avoided using a constant offset voltage at the CN-depletion layer (Fig. 7).

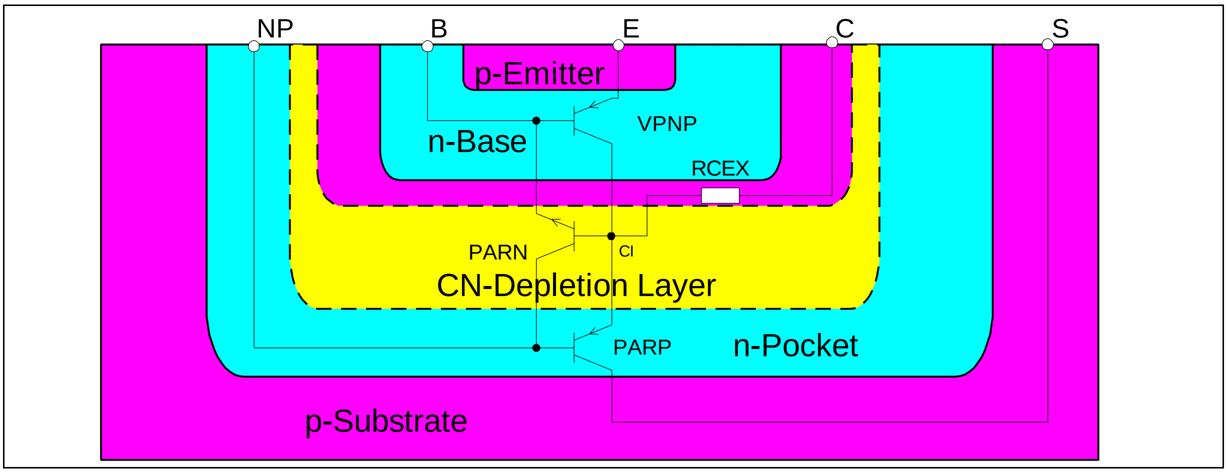


Fig. 5: Collector resistance pinch off, RC is pinched by CN-depletion layer

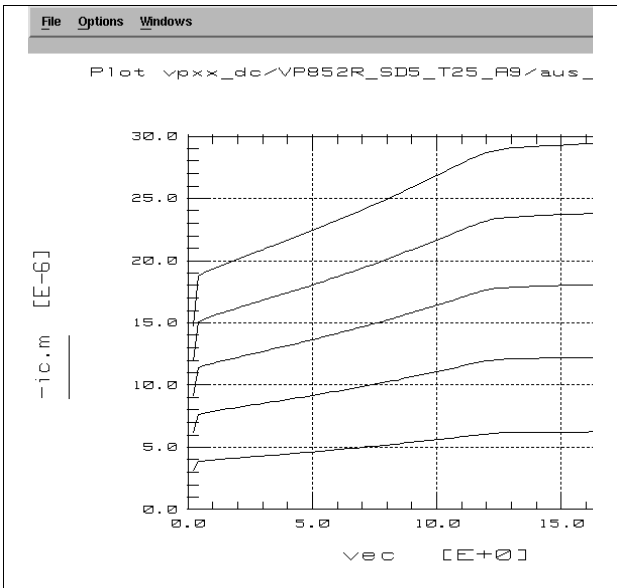


**Fig. 6: Output characteristic $I_C = f (V_C)$, I_B = parameter ;
collector current compliance is caused by RC pinch off**

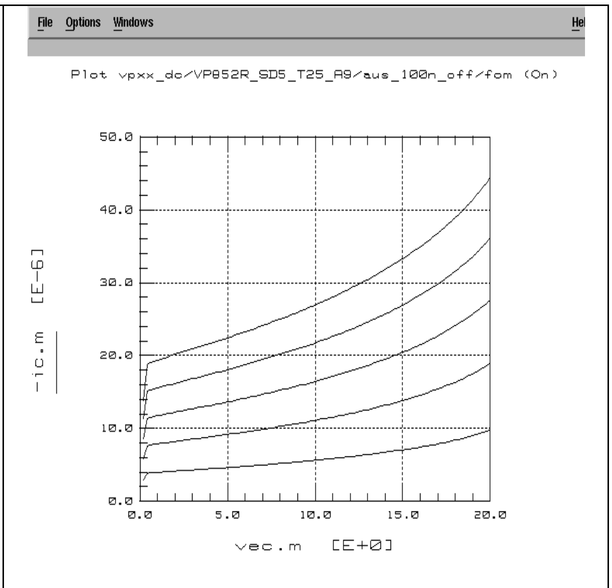


**Fig. 7: Output characteristic $I_C = f (V_C)$, I_B = parameter;
using an offset voltage V_{off} = + 4 V no collector current compliance appears**

2.2.3 Quasisaturation

Quasisaturation does not occur only for VPNP transistors, however, a VPNP is particularly affected by this effect.
Generally, quasisaturation occurs, if the outer base-collector junction is reverse biased ($V_{BCvpnp} > 0$, as in the active forward mode) , while the inner base-collector junction is already forward biased ($V_{B'C'vpnp} < 0$, as in saturation). Quasisaturation is caused by a current and voltage dependent collector resistance. At high collector currents the internal voltage drop V_{RC} is high enough to drive the transistor into quasisaturation. Considering an forward output characteristic, QS appears as an increase of the output conductance with decreasing collector voltage (Fig. 9).

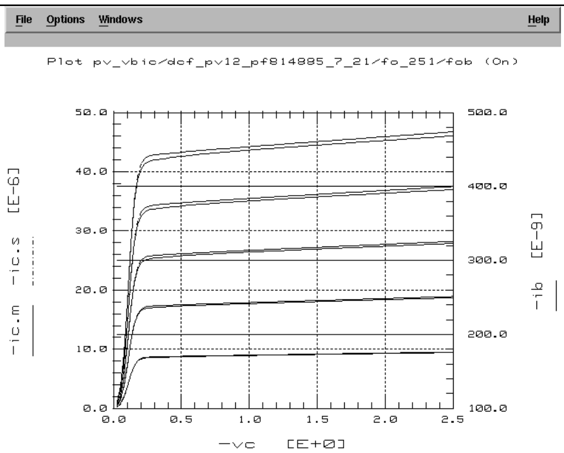


Fig. 8: Output characteristic I_C = f (V_C), I_B = Par. ; no Quasisaturation at low collector currents

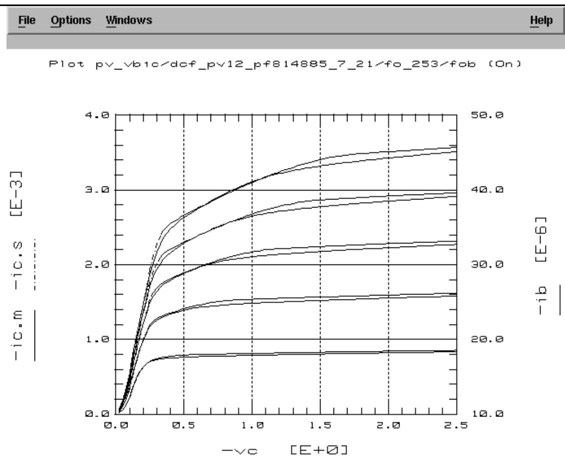


Fig. 9: Output characteristic I_C = f (V_C), I_B = Par. Quasisaturation occurs at high collector currents

2.3 Measurement Circuits

According to the VPNP equivalent circuit, a number of measurement circuits is necessary to characterize the main VPNP and the two parasitics PARN and PARP. In Fig. 10 and Fig. 11 the general DC- and AC- measurement circuits are shown for an ICCAP environment, where SMU's [2] are available. The appropriate SMU settings for the particular circuit are defined in Table 1[3].

[2] SMU = Source Monitor Unit (e.g. in HP4142)
[3] This table does not contain definite values, because they depend on the used technology.

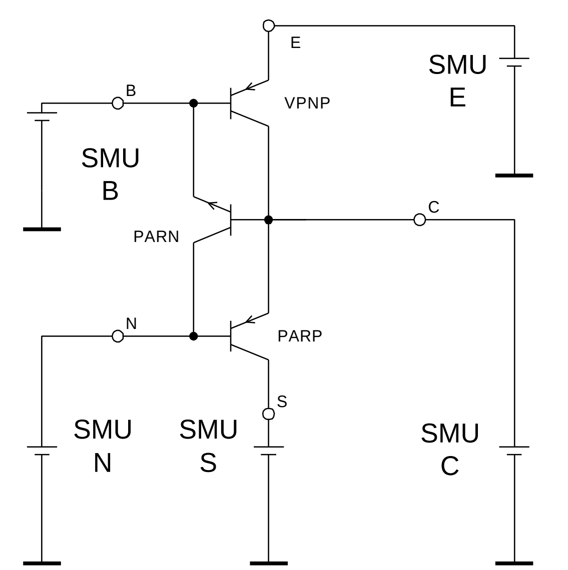


Fig. 10: General VPNP DC- measurement circuit

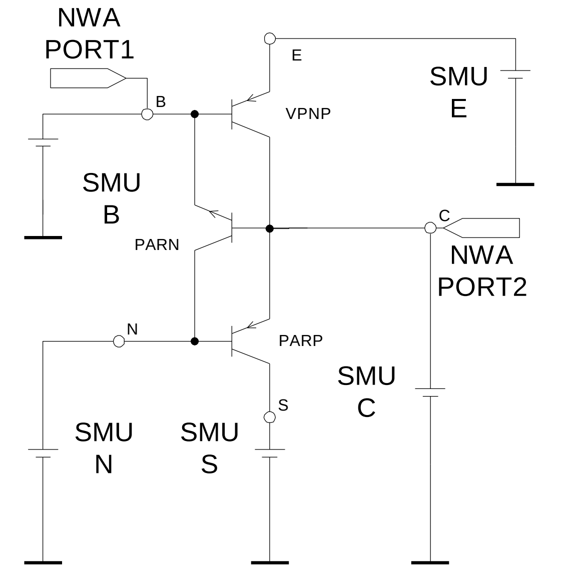


Fig. 11: General VPNP AC- measurement circuit

Table 1: SMU settings for VPNP measurement circuits

Circuit	SMU1 E	SMU2 B	SMU3 C	SMU4 N	SMU5 S	VPNP	PAR N	PAR P
fg_vpnp	0	- V_B = sweep	- V_C	0	-1	fwd	off	off
fg_vpnp (V_N = offset)	0	- V_B = sweep	- V_C	V_C + V_{offset}	-1	fwd	off	off
rg_vpnp	-V_E	- V_B = sweep	0	0	-1	rev	fwd	off
fg_parp	0	+ 1	+ V_C = sweep	0	-1	off	rev	fwd
rg_parp	0	0	0	0	+ V_S =sweep	off	off	fwd
re_vpnp	0	- I_B = sweep	I_C = 0	I_N = 0	I_S = 0	Sat	off	off
rc_vpnp	0	- I_B = sweep	I_C = B_{FB} * I_B	I_N = 0	I_S = 0	Sat	off	off
fo_vpnp	0	- I_B = sweep1	- V_C = sweep2	0	-1	fwd	off	off
fo_vpnp (V_N = offset)	0	- I_B = sweep1	- V_C = sweep2	V_C + V_{offset}	-1	fwd	off	off
ro_vpnp	- V_E = sweep2	- I_B = sweep1	0	0	0	rev	off	off
fo_parn	- 1	0	+ I_C = sweep1	+ V_N = sweep2	- 1	off	fwd	off
ro_parn	0	+ V_B = sweep2	+ I_C = sweep1	0	- 1	fwd	off	off
acmf_vpnp f_M = sweep1	0	- V_B = sweep2	- V_C = sweep3	0	-1	fwd	off	off
acsf_vpnp f_M = const.	0	- V_B = sweep1	- V_C = sweep2	0	-1	fwd	off	off

3 VPNP Modelling

3.1 VBIC - Model

The VBIC95 was developed by a committee of US semiconductor companies to overcome the deficiencies of the SGP[4] model, which has remained unchanged in the last 20 years. The main features of the VBIC95 model are [1] [3] [5]:

- Base current definitions independent of the transfer currents, no beta parameters are used
- First order distributed-base model
- Early effect model is based on the junction depletion charge
- Modified Kull model for quasisaturation
- Parasitic substrate transistor
- Single piece depletion capacitance model
- Improved temperature scaling
- Weak avalanche for BC junction
- Parasitic overlap capacitances
- Self heating model

A complete description of the model equations is behind the scope of this paper (see [2][6]), we will focus on the equations, describing the Early effect, the quasisaturation and the transit time.

3.1.1 Depletion Charge Model

The VBIC95 model is focused on charges, not capacitances. Normalised charges are used both for DC - and AC – equations. Given the voltage dependent capacitance of a reverse biased junction as

$$C(V) = \frac{CJO}{\left[1 - \frac{V}{P}\right]^{M}} \qquad (1)$$

where are P and M the built-in potential and the grading coefficient, respectively, we have after integration the following relation for the voltage dependent charge:

$$Q(V) = \int_0^V C(V)dV = CJO \cdot \frac{P}{(1-M)}\left[1 - \left(1 - \frac{V}{P}\right)^{(1-M)}\right] \qquad (2)$$

In VBIC95 the second factor of eqn (2) is realised as a function, called qj. This function represents a normalised charge[5]. Using this normalised charge, the charge itself is given by multiplication with the zero bias capacitance, given as model parameter:

$$Q(V) = CJO * qj\left(V, P, M\right) \qquad (3)$$

Based on this principles, the normalised depletion charges are calculated in the model as fundamental values[6]. Note, that the normalised charge depends on the voltage across the junction and on both the model parameters P , M as well, whereas the parameter CJO does not affect qj.

[4] SGP = Spice - Gummel Poon Model
[5] Note, that the dimension of this normalised charge is Volt.
[6] The VBIC95 model includes two depletion charge models: the regional model (SGP model) and a new single-piece model, which limits the capacitance to a constant value for bias values greater than the built in potential. The appropriate model is selected using AJ as a flag: if AJ <= 0, the regional model is used, otherwise the single-piece model.

3.1.2 Normalised Base Charge

The normalised base charge is defined as

$$q_b = \frac{1}{2}\left[q_1 + \sqrt{q_1{}^2 + 4q_2}\right] \tag{4}$$

where

$$q_1 = 1 + \frac{q_{dbe}}{VER} + \frac{q_{dbc}}{VEF} \tag{5}$$

$$q_2 = \frac{I_{tf}}{IKF} + \frac{I_{tr}}{IKR} \tag{6}$$

The Early voltages VER, VEF and the knee currents IKF, IKR are the model parameters. Note, that for q_1 the normalised charges q_{dbe} and q_{dbc} (unit V) are used, instead of branch voltages V_{be} and V_{bc} , as in the SGP model. As explained before, the normalised charges q_{deb} and q_{dbc} are dependent on the internal voltages V_{bei} and V_{bci} again.

3.1.3 Transport Currents

The forward and reverse ideal transport currents are defined as[7]

$$I_{tf} = I_S \cdot \exp\left(\frac{V_{bei}}{NF \cdot V_t} - 1\right) \tag{7}$$

$$I_{tr} = I_S \cdot \exp\left(\frac{V_{bci}}{NR \cdot V_t} - 1\right) \tag{8}$$

where are IS the transport saturation current, NF and NR the emission coefficients and V_t the temperature voltage. Based on these equations we have the forward and reverse nonideal transfer currents I_{tzf} and I_{tzr} as[8]

$$I_{tzf} = \frac{I_{tf}}{q_b} \tag{9}$$

and

$$I_{tzr} = \frac{I_{tr}}{q_b} \tag{10}$$

where q_b is the normalised base charge. In this way a voltage dependent output conductance is realised in VBIC.

[7] The index “i” denotes a voltage between the internal nodes.

[8] The separate definition of I_{tzf} and I_{tzr} is necessary, because the zero phase current I_{tzf} is used in the additional excess phase network of the VBIC95 model as the control current. Using this network, the current I_{txf} with an additional excess phase is calculated. If the excess phase parameter TD is specified, the current I_{txf} is used in the model instead of I_{tzf}.

3.1.4 Collector Resistance and Quasisaturation

The collector resistance consists of two parts: the constant external resistance R_{CX} and the epi layer resistance. The epi layer resistance depends on both the collector current and the epi layer voltage drop. According to the KULL model [7] , the current through the epi layer I_{ohm} is calculated as:

$$I_{ohm} = \frac{V_{rci} + V_t \cdot \left[K_{bci} - K_{bcx} - \log\left[\frac{1+K_{bci}}{1+K_{bcx}} \right] \right]}{RCI} \tag{11}$$

with

$$K_{bci} = \sqrt{1 + GAMM \cdot \exp\frac{V_{bci}}{V_t}} \tag{12}$$

$$K_{bcx} = \sqrt{1 + GAMM \cdot \exp\frac{V_{bcx}}{V_t}} \tag{13}$$

where RCI (the epi resistance under equilibrium condition) and GAMM (the epi charge coefficient) are the model parameters. Eqn (11) takes into account the epi layer resistance reduction, if the transistor enters quasisaturation in the range of ohmic behaviour.
Examining the effect of the quasisaturation parameters GAMM and RCI, it is useful to consider eqn (11) more in detail. Rearranging this relationship, it reduces to a simple equation of the form I = V / R:

$$I_{ohm} = \frac{V_{rci} + V_{cor}}{RCI} \tag{14}$$

where

$$V_{rci} = V_{bci} - V_{bcx} \tag{15}$$

and

$$V_{cor} = V_t \left[K_{bci} - K_{bcx} - \log\left[\frac{1+K_{bci}}{1+K_{bcx}} \right] \right] \tag{16}$$

V_{COR} here is a correction voltage, depending on both the epi layer doping parameter GAMM and the epi layer voltage drop V_{rci} . This correction voltage is added to V_{rci} . In the active normal case ($V_{bcx} < 0$, $V_{bci} < 0$) V_{COR} is zero, but in quasisaturation V_{COR} increases. For a given RCI value this results in an increasing value for the current I_{ohm} . In this way the effective collector resistance is reduced.

Additional to the epi layer resistance reduction, the effect of carrier velocity saturation (nonohmic behaviour) may occur in quasisaturation. Eqn (17) takes into account this effect by modifying the current I_{ohm} :

$$I_{rci} = \frac{I_{ohm}}{\sqrt{1 + \left[\frac{RCI \cdot I_{ohm}}{VO \cdot \left[1 + \frac{0.5 \cdot \sqrt{V_{rci}^2 + 0.01}}{VO \cdot HRCF} \right]} \right]^2}} \tag{17}$$

where VO (saturation velocity voltage) and HRCF (high current RC factor) are the new model parameters. Eqn (17) shows, that the VBIC95 model modifies the value of RCI using the factor sqrt(1+ RCI * I_{ohm} / VO) instead of the factor (1+ (V_{rci} / VO)^2) used by the KULL model.

$$I_{rci} = \frac{I_{ohm}}{\sqrt{1+\left[\frac{RCI \cdot I_{ohm}}{VO}\right]^2}} \tag{18}$$

Considering the simplified eqn (18), we can see: if the voltage drop RCI * I_{ohm} is lower than VO, I_{rci} and I_{ohm} are nearly identical. If the voltage drop reaches VO, the denominator increases and I_{rci} decreases ($I_{rci} < I_{ohm}$).
Additional, eqn (17) shows that the value of VO is changed depending on V_{rci} and the model parameter HRCF. Thus the effect of I_{rci} reduction increases with V_{rci} .
To avoid numerical problems |V_{rci} | is calculated in eqn (17) as ($V_{rci}{}^2$ + 0.01) ^{½} .

3.1.5 Transit Time

The forward transit time is defined as:

$$t_{ff} = TF \cdot \left(1+QTF \cdot q_1\right) \cdot \left(1+XTF \cdot \left[\frac{I_{tf}}{I_{tf}+ITF}\right]^2 \cdot \exp\left(\frac{V_{bci}}{1.44 \cdot VTF}\right)\right) \tag{19}$$

This definition is identical to the SGP model, except the additional term (1+ QTF * q1). Note, that

- q1 includes a dependence on the internal branch voltages V_{bei} and V_{bci} using the normalised charges q_{dbe} and q_{dbc} and
- the ideal current I_{tf} is used in eqn.(19).

3.2 SQ3 - Model

The Siemens SQ3 model is an improved Spice Gummel Poon model (SGP). The main advantages over the SGP model are:

- base current definitions independent of the transfer currents, no beta parameters are used
- improved temperature scaling
- improved transit time equation

A detailed description of the SQ3 model is again behind the scope of this paper (see [13][14]). We will consider here only the SQ3 base charge and the transit time equation.

3.2.1 Normalised Base Charge

The used normalised base charge definition is identical to the SGP:

$$Q_B = \frac{Q_1}{2} * \left[1 + \sqrt{1 + 4Q_2}\right] \tag{20}$$

where

$$Q_1 = \frac{1}{1 - \frac{V_{B'C'}}{VAF} - \frac{V_{B'E'}}{VAR}} \tag{21}$$

and

$$Q_2 = \frac{I_F}{IKF} + \frac{I_R}{IKR} \tag{22}$$

This base charge definition results in an voltage independent output conductance.

3.2.2 Transit Time

The SQ3 transit time formulation is different to the VBIC model:

$$T_{FF} = \frac{TF}{\left[1 - \frac{V_{bc}}{VTA}\right]^{MTA}} * \frac{1 + \left[\frac{I_C}{ITK * \left[1 - \frac{V_{cb}}{VTK}\right]^{MTK}}\right]^{KTK}}{1 + \left[\frac{I_C}{ITG * \left[1 - \frac{V_{bc}}{VTG}\right]^{MTG}}\right]^{KTK}} \tag{23}$$

Simplifying this equation as

$$T_{FF} = A * \frac{B}{C} \tag{24}$$

we can consider the terms A, B and C more closely (Fig. 12). Term A describes the voltage dependence of T_{FF} , using the well known space charge capacitance formulation (model parameters VTA, MTA).

Term B is used to describe the T_{FF} – increase with increasing collector current. Again, the space charge capacitance formulation is used to introduce an additional voltage dependence of the knee current parameter ITK, using the parameters VTK, MTK.

At very high collector currents the increase of T_{FF} attenuates and T_{FF} approaches a limiting value. This behaviour may be modelled using term C and the model parameters ITG, VTG, MTG.

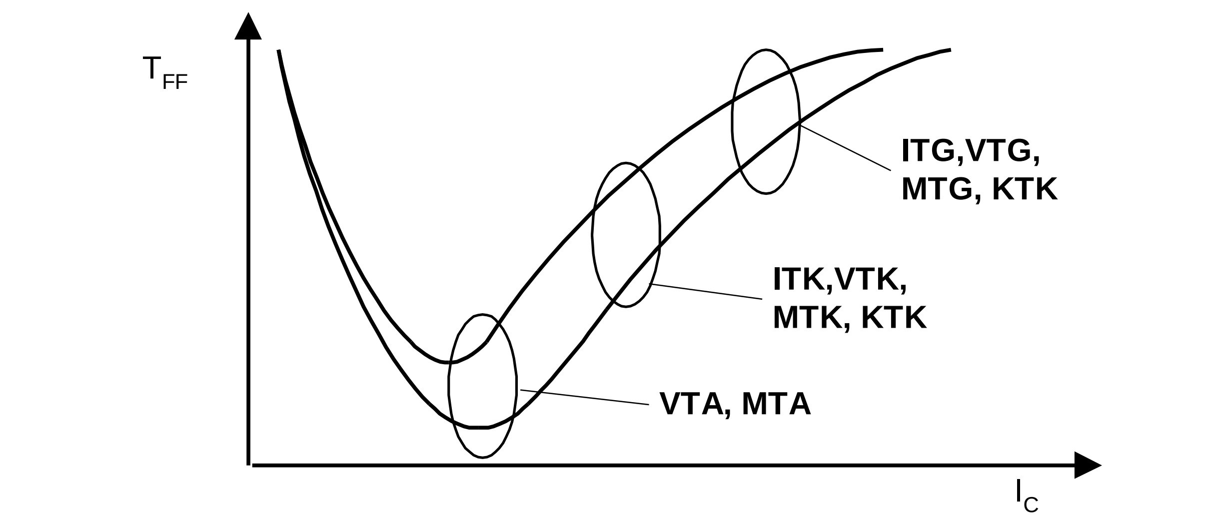


Fig. 12: Effect of SQ3 - parameters on T_{FF}

Equation (23) allows a good fit for various types of T_{FF} – curves, but it may be complicated determine the model parameters, because the terms A, B and C affect each other.
The SQ3 model does not take into account the Quasisaturation and, moreover, it does not contain parasitic devices. That is why a subcircuit model is always necessary to model integrated transistors using SQ3.

3.3 Subcircuit Models

The VBIC model contains one parasitic transistor, sufficient for npn–modelling. For VPNP modelling however, a subcircuit model is necessary to take into account the parasitic substrate transistor PARP (Fig. 13).
The SQ3 subcircuit model consists of three transistors, each one modelled by an SQ3 model, and an external collector resistance (Fig. 14).

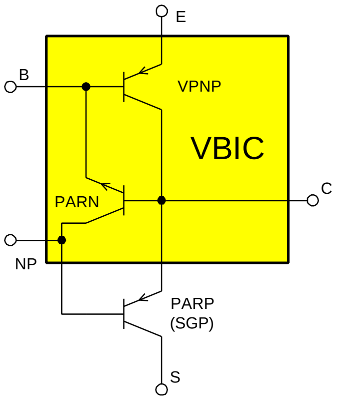


Fig. 13: VBIC subcircuit model for VPNP

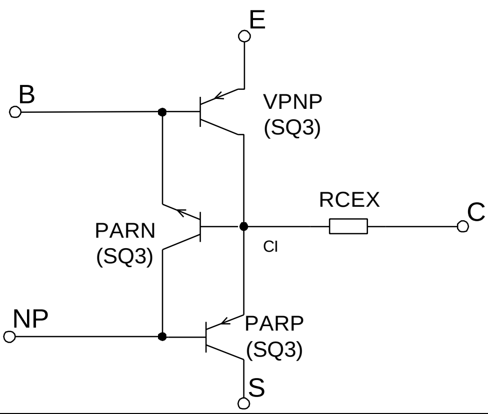


Fig. 14: SQ3 subcircuit model for VPNP

4 VPNP Measurement and Simulation Results

In this section some measurement and simulation results are presented. For the sake of an practical assessment we will consider the VBIC simulation results in comparison to the SQ3 – subcircuit model, used so far for the VPNP transistor at Infineon Technologies [15] [16]. The simulations are made using ICCAP 5.2 and Spectre 4.4.3 for the VBIC model respectively Saber 4.1.1. for the SQ3 model.

Gummel Plot (Fig. 15)

Fig. 15 shows forward Gummel characteristics for various collector voltages. At V_C = -3V the transistor is in the active forward mode, at V_C = -1V the device enters saturation at V_B = -0.9V, indicated by the increasing n-Pocket current and at V_C = -0.2V the device is in saturation during the whole V_B – sweep. The simulated n-Pocket current I_N is for both the models good in agreement with the measurement. The substrate current does not appear under active forward and saturation operating conditions, as explained in section 2.2. In the case of Latch Up both the models are not able to model substrate current exactly, because the RC parameter value, sufficient for active forward and saturation operating conditions, is to low.

Output Characteristic (Fig. 16, Fig. 17)

The output characteristics were measured at various base current ($-I_B$ = 1 ... 250 µA) and collector voltage sweeps ($-V_{Cmax}$ = 2.5 and $-V_{Cmax}$ =15 V).
As can be seen from Fig. 16, in the low current range simulations and measurements in agreement for both the models. In the mid current range ($-I_B$ = 10 ... 50 µA), VBIC delivers betters results, because the SQ3 does take into account the quasisaturation. This statement is true in the high current range ($-I_B$ = 50 ... 250 µA) as well, however only in the low voltage range up to $-V_C$ = 5V.
Considering Fig. 17 we have to note, that the output conductance in the high current range for $-V_C$ > 5V is modelled badly by VBIC. Contrary to an npn transistor, there is no way to model the output conductance both in the low and high current range exactly using VBIC, despite it's improved Early voltage formulation (eqn(5)). The strong increase of the output conductance vs. collector current seems to be typical for VPNP transistors. Note, that this is not caused by self heating or avalanche effect. That is why, there is no possibility to model the effect by the appropriate self heating or avalanche parameters of the VBIC model.
Surprisingly, on the other hand, the SQ3 model delivers better results in the high voltage range ($-V_C$ = 5 ... 15V) as the VBIC, despite it uses the simple SGP Early voltage formulation (21).

Transit Frequency Characteristic (Fig. 18, Fig. 19)

The Transit frequency characteristics where measured at various collector voltages: $-V_C$ = 0.2, 0.5, 1, 3, 5, 10V. Modelling these characteristics, we are faced with two problems:

- a strong dependence of f_T vs. V_C and
- a steep decrease of f_T vs. I_C at higher currents

As expected, the SQ3 model is able to simulate the transit frequency from $-V_C$ = 1V up to $-V_C$ = 10V sufficiently in agreement with the experimental data (Fig. 18). In the saturation range ($-V_C$ = 0.2, 0.5 V), however, considerable deviations appear (Fig. 19).

The VBIC model is, contrary to the SQ3 model, not able to simulate the f_T dependence of f_T vs. V_C . The simulated f_{Tmax}-value at $-V_C$ = 10V is nearly 20% lower than the measured value and, moreover, the f_T decrease at high currents is not modelled. Although for the VBIC model the SGP f_T – equation is modified by QTF (eqn(19)), it is not possible to get a better fit of f_T vs. V_C using the parameter QTF.
In the saturation range, similarly to the SQ3 model, considerable deviations appear

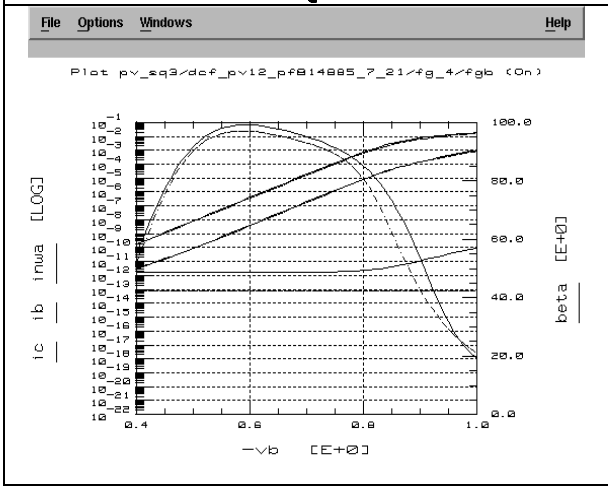
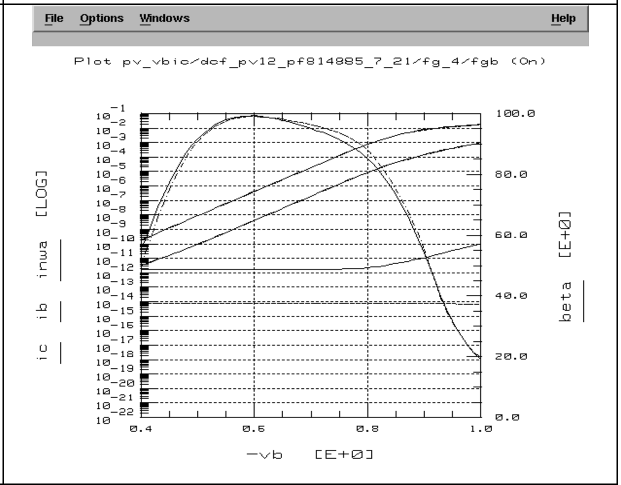
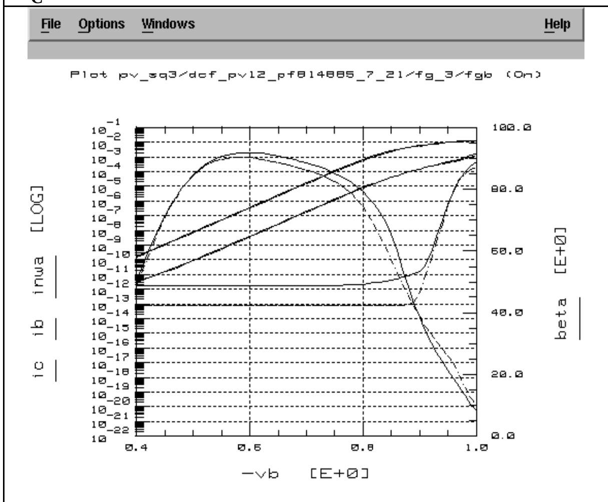
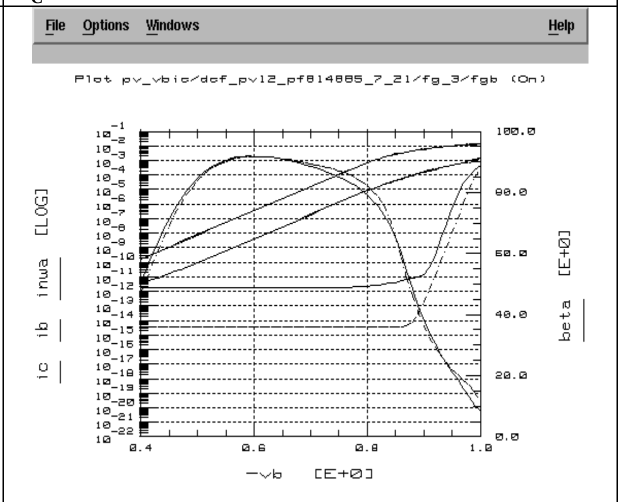
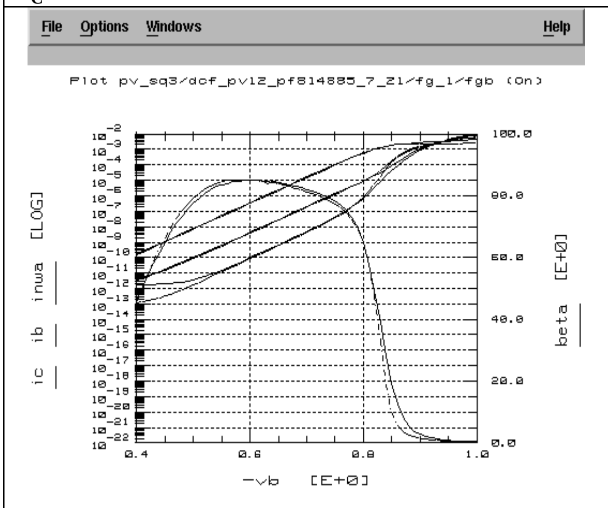
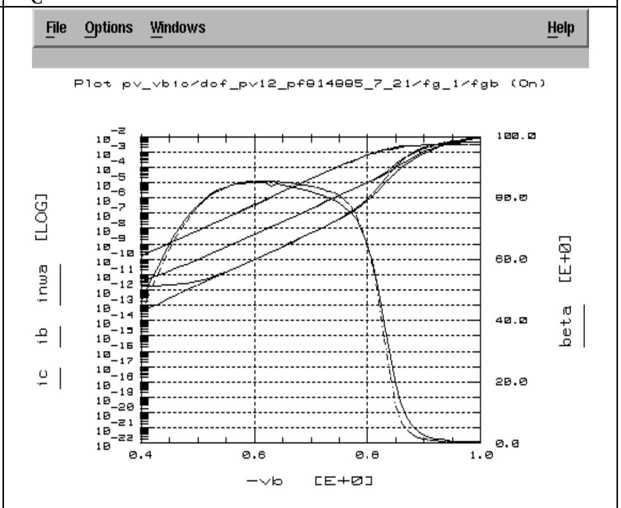
SQ3	VBIC
	
V_C = - 3 V	V_C = - 3 V
	
V_C = - 1 V	V_C = - 1 V
	
V_C = - 0.2 V	V_C = - 0.2 V

Fig. 15: Gummel Plot I_C , I_B , I_N , I_{Sub} , B_N = f (V_B), V_C = Parameter

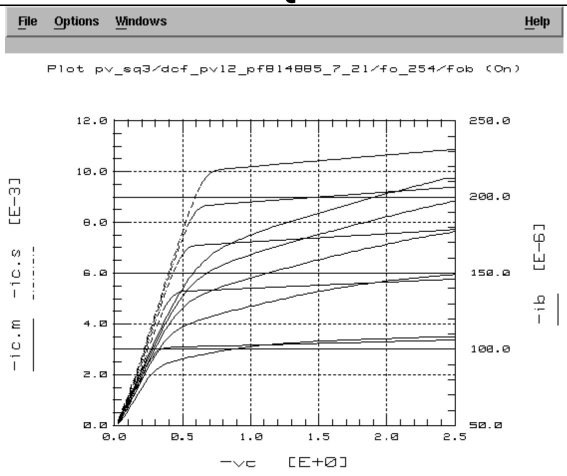
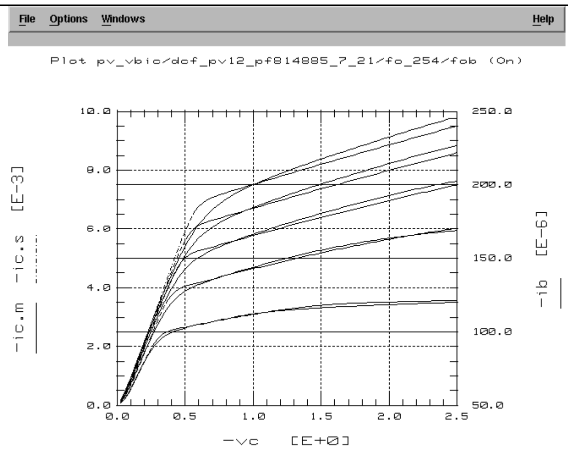
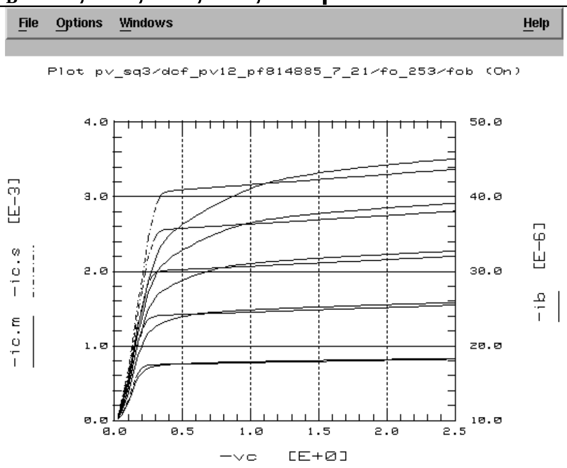
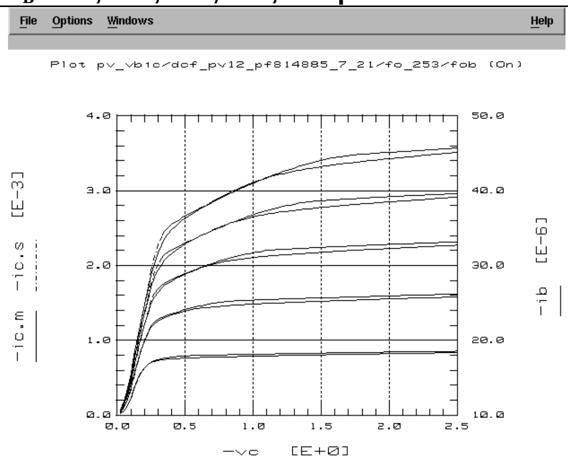
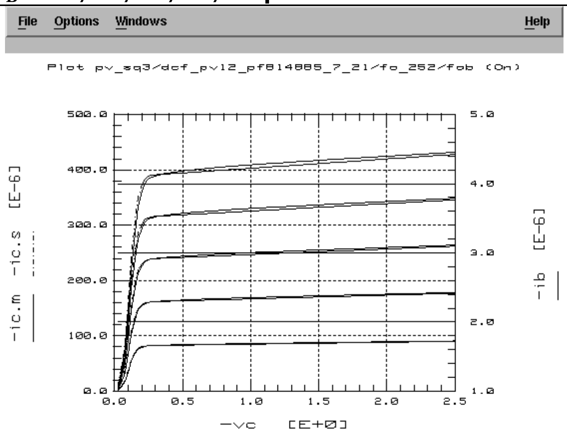
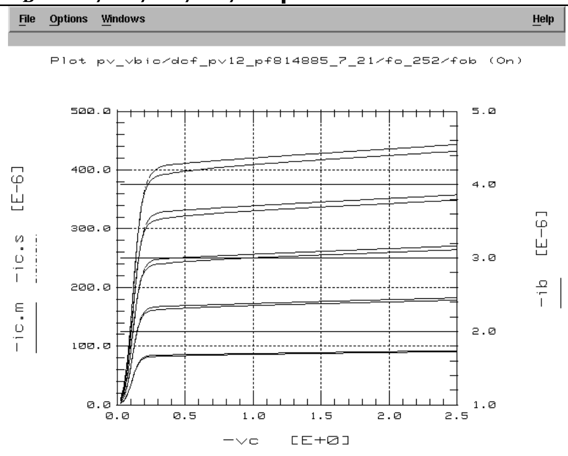
SQ3	VBIC
	
- I_B = 50, 100, 150, 200, 250 µA	**- I_B = 50, 100, 150, 200, 250 µA**
	
- I_B = 10, 20, 30, 40, 50 µA	**- I_B = 10, 20, 30, 40, 50 µA**
	
- I_B = 1, 2, 3, 4, 5 µA	**- I_B = 1, 2, 3, 4, 5 µA**

Fig. 16: Output characteristic I_C = f (V_C), I_B = Parameter, V_{Cmax} = -2.5V

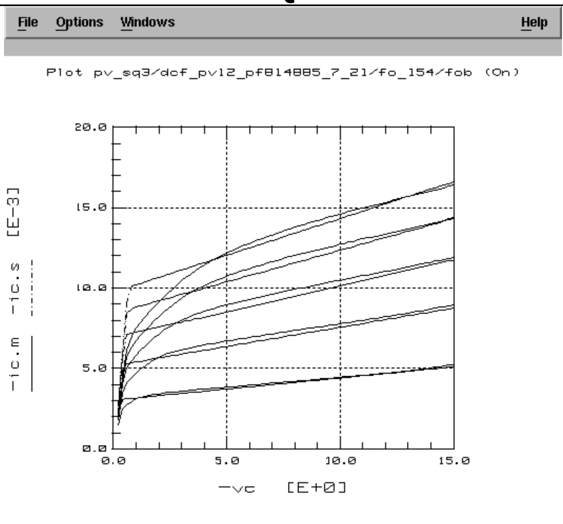
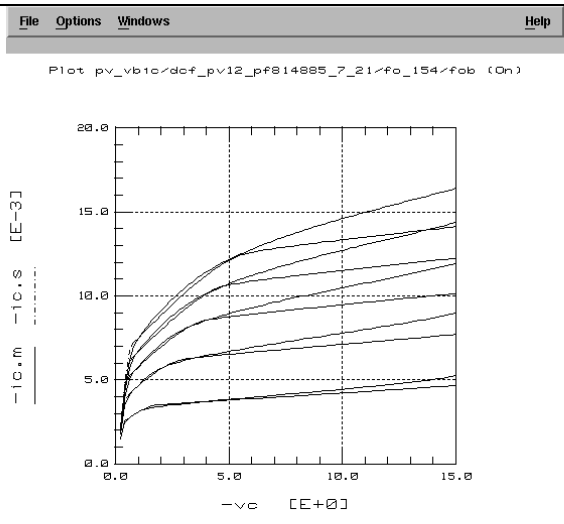
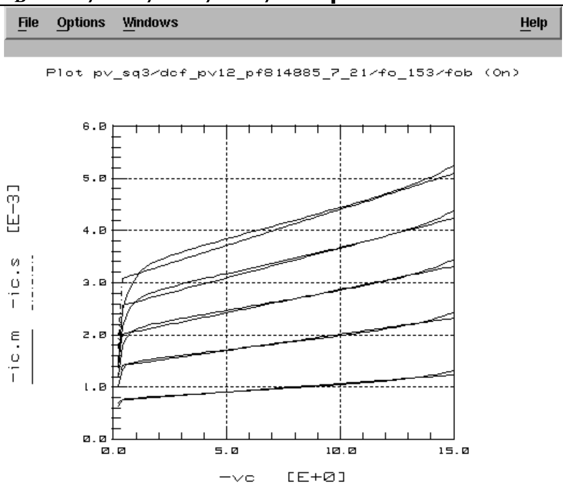
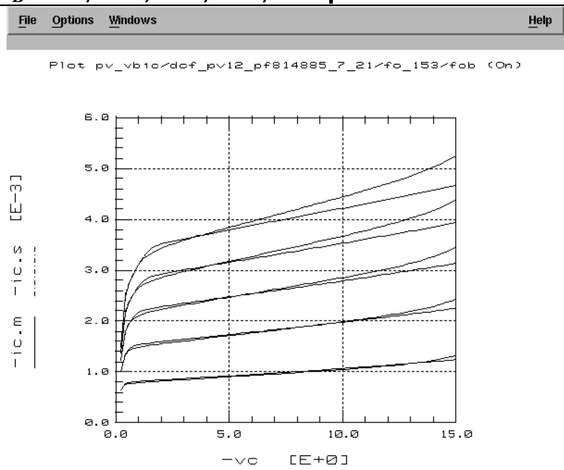
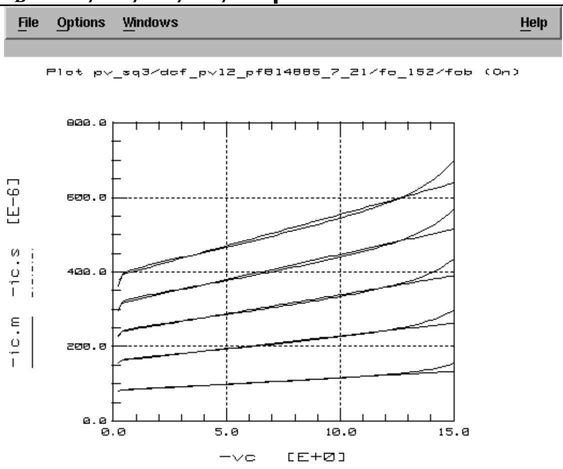
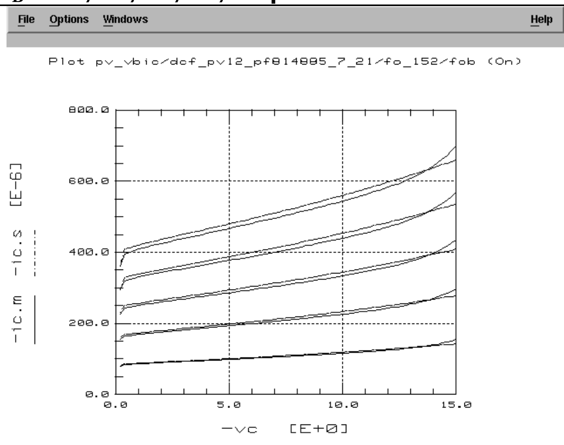
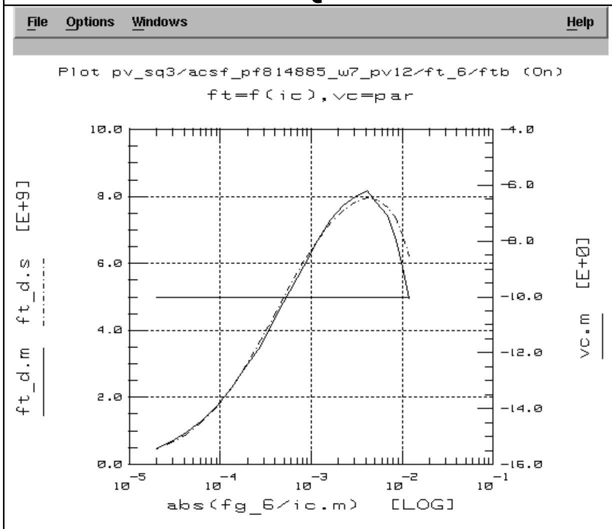
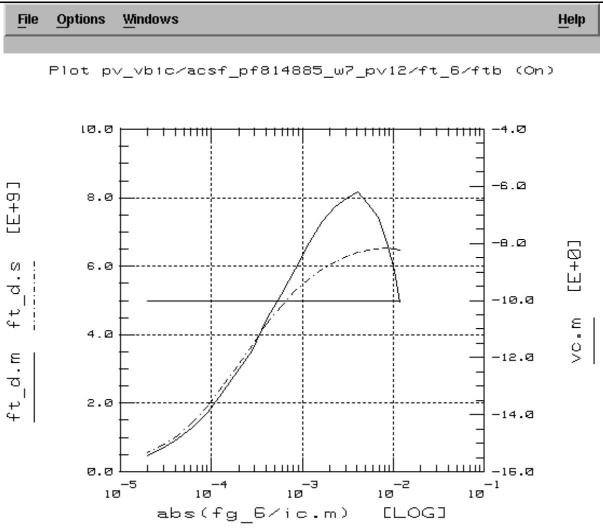
SQ3	VBIC
	
- I_B = 50, 100, 150, 200, 250 µA	**- I_B = 50, 100, 150, 200, 250 µA**
	
- I_B = 10, 20, 30, 40, 50 µA	**- I_B = 10, 20, 30, 40, 50 µA**
	
- I_B = 1, 2, 3, 4, 5 µA	**- I_B = 1, 2, 3, 4, 5 µA**

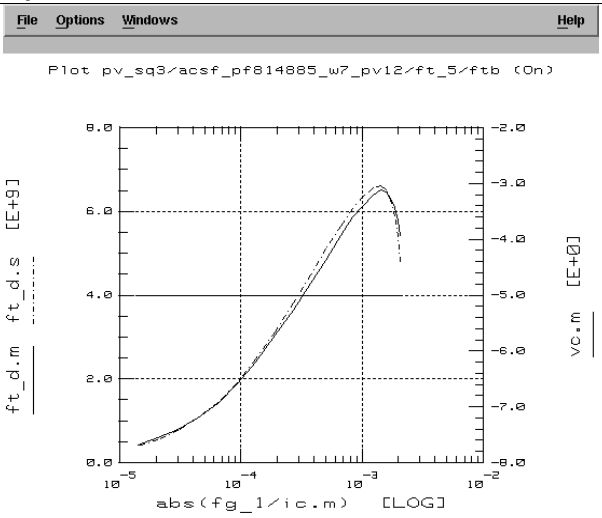
Fig. 17: Output characteristic I_C = f (V_C), I_B = Parameter

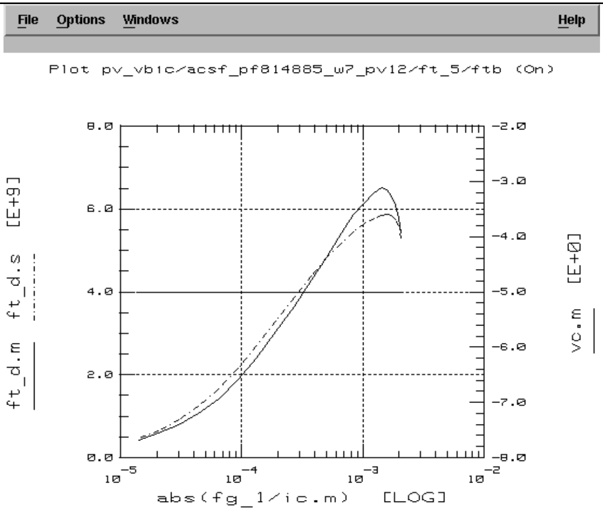
SQ3	VBIC



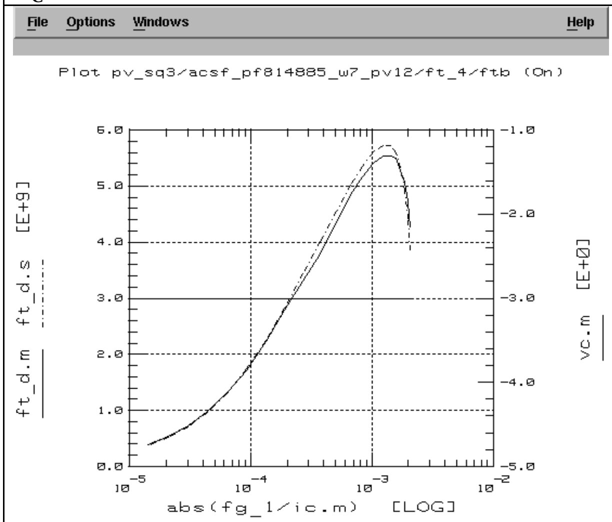


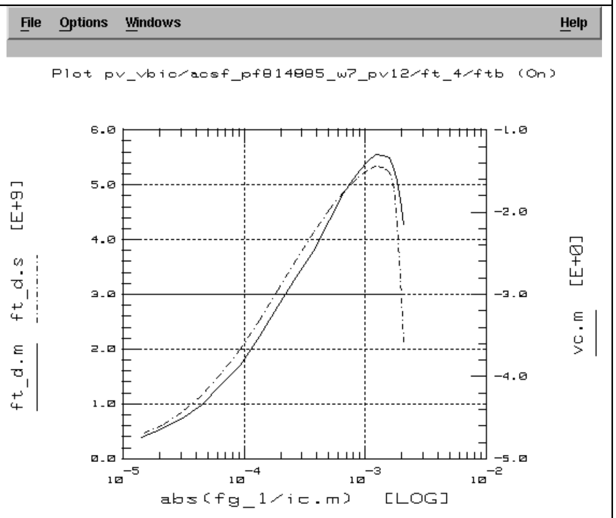
V_C = - 10 V





V_C = - 5 V





V_C = - 3 V

Fig. 18: Transit frequency characteristic f_T = f (I_C), -V_C = 3, 5, 10

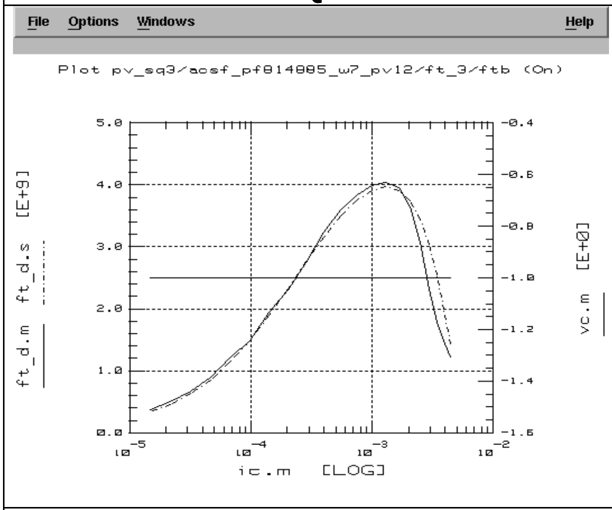
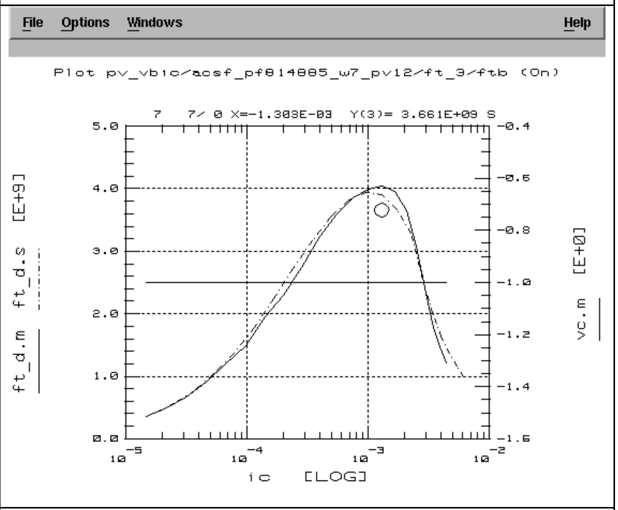
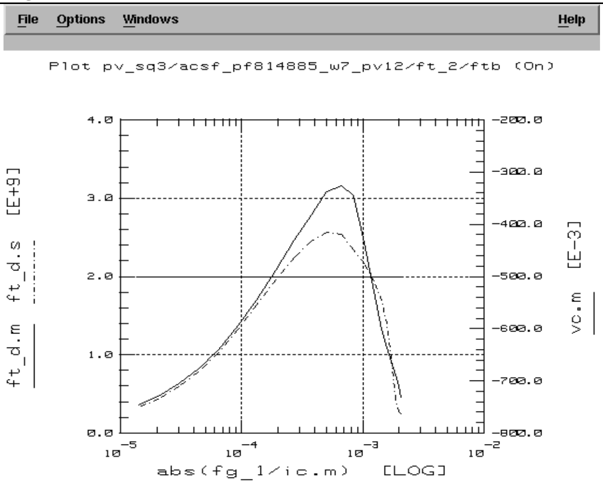
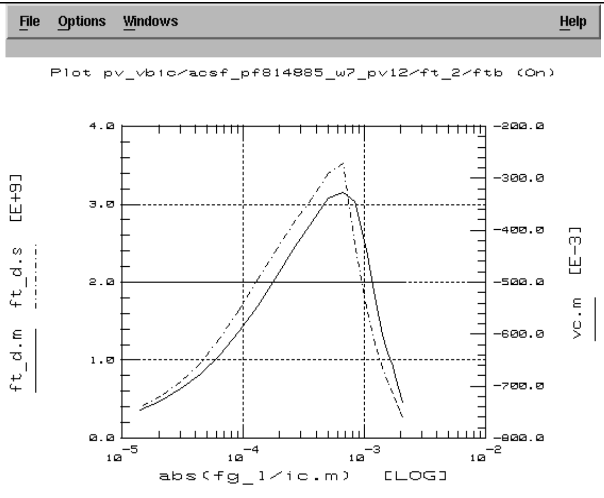
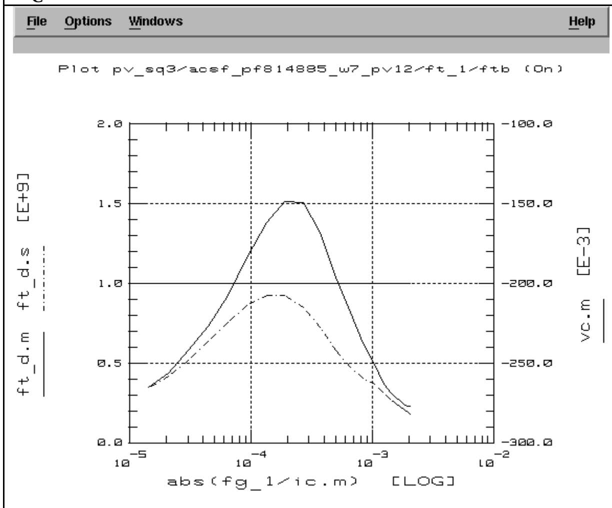
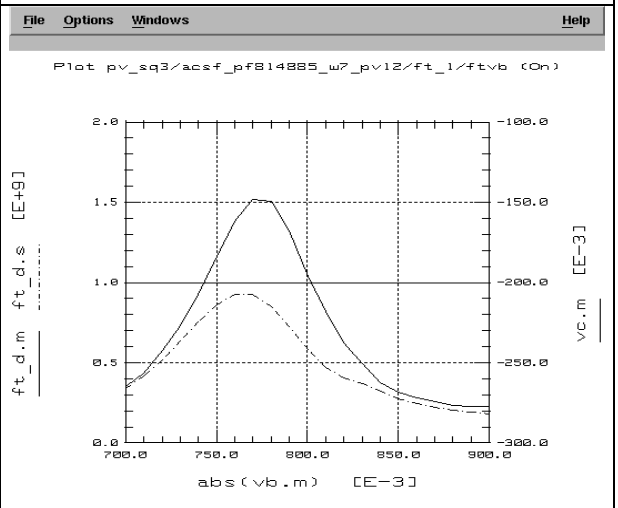
SQ3	VBIC
	
V_C = - 1 V	
	
V_C = - 0.5 V	
	
V_C = - 0.2 V	

Fig. 19: Transit frequency characteristic $f_T = f(I_C)$, $-V_C$ = 0.2, 0.5, 1V

5 Summary

In this paper the suitability of the VBIC model for VPNP modelling was investigated. The VPNP structure and typical parasitic effects as Latch Up, collector current compliance and quasisaturation were explained. DC- and AC- measurement circuits suitable for an ICCAP environment where shown. Some main features of the VBIC95 and the SQ3 model where explained using the appropriate equations. Measurement and simulation results are presented in comparison to the SQ3 model, resulting in the following conclusions:

- The main advantage of SQ3 model is the transit time equation. Although this equation is not physical based, it allows to fit transit time characteristics of various shapes.

- The main disadvantage of SQ3 model is the lack of a quasisaturation model.

- The main advantage of the VBIC model is the quasisaturation model.

- The main disadvantage of model is the inaccurate modelling of the transit time at higher collector voltages.

Table 2: Comparison of VBIC and SQ3 for VPNP modelling

topic	VBIC	SQ3
quasisaturation	good	not available
transit time	poor at higher collector voltages	good
output conductance	good up to mid currents, poor at higher collector currents	sufficient
n-pocket current	good	good

Summarising we can say, that the VBIC model delivers better DC-results as the SQ3, at least up to mid currents, because it takes into account the quasisaturation. The AC-results, however, are bad at higher collector voltages.

Taking into account the fact, that in modern RF circuits low supply voltages (e.g. 3V) are usual, the deficiencies of the VBIC model at higher voltages are less important. From this point of view, VBIC is the preferred model for VPNP modeling.
Nevertheless an improved pnp-version of the VBIC model is necessary, including at least an improved transit time equation.

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VBIC Model

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** Hewlett Packard GmbH Böblingen

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