Parameter Extraction
for
BJT Quasisaturation Models[1]

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Abstract - A transistor acts in the state of quasisaturation (QS), if the external base - collector voltage is negative ($V_{BCx} < 0$) and the inner base - collector voltage is positive ($V_{BCi} > 0$), caused by the collector resistance voltage drop. QS is possible for each bipolar transistor, but it becomes important in practice especially for devices with lightly doped collector regions, operating at high current densities. This paper is concerned with BJT quasisaturation models. A short survey of QS models used in today's circuit simulators is presented. The parameters and their effects are discussed for the CadenceSPICE and the HSPICE model. Model parameter extraction methods are explained based on measurement results.

1 Quasisaturation Models

The widespread SPICE - Gummel Poon model (SGP - model) is using a constant collector resistance. Usually an RC value is chosen, that results in a good fit of the saturation characteristic. For QS modeling, however, this constant RC model is not useful. The collector resistance of an integrated BJT usually consists of three parts: epitaxial layer, buried layer and sinker. Buried layer and sinker resistance are constant, whereas the epi layer resistance depends on both the collector current and the BC -voltage. Therefore, a QS model must create a decreasing collector resistance at decreasing collector voltage V_{BC} . This can be realized by a controlled collector resistance (e.g. the CadenceSPICE - model) or by a controlled current source (all models based on the Kull - model).

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[1] This paper is intended for the HP User Meeting Oct. 1997 Berlin, Rev.: 6.6.97

1.1 CadenceSPICE - Model

The CadenceSPICE quasisaturation model [2] is based on a variable collector resistance, dependent both on the collector current and the internal BC - voltage[2]:

$$R_{CC} = RCM + RCV * \frac{\left[1 + \left(\frac{I_C}{CCO}\right)^{CEX}\right]}{1 + \left(\frac{n_i}{DOPE}\right)^2 \exp\left(\frac{V_{BCi}}{NR * U_T}\right)} \tag{1}$$

This equation acts as follows[3]: For a reverse biased BC - junction the exponential term is small and the denominator of eqn (1) is nearly 1. Thus R_{CC} is constant for a given collector current. For a forward biased BC - junction, however, the denominator is > 1 and the effective collector resistance is reduced. On the other hand, an increasing collector current increases the numerator and in this way the effective collector resistance is increased. The CadenceSPICE quasisaturation model does not include additional circuit elements for dynamic QS effects or substrate current simulation (Fig. 1).

Table 1: CadenceSPICE - QS model parameters [2]

Parameter	Description	Unit	Default
RCM	Minimum collector resistance	Ohm	0
RCV	Variable collector resistance	Ohm	0
CEX	Current crowding exponent	-	1
CC0	Current crowding normalization current	A	1
DOPE	Collector doping concentration	cm^{-3}	0

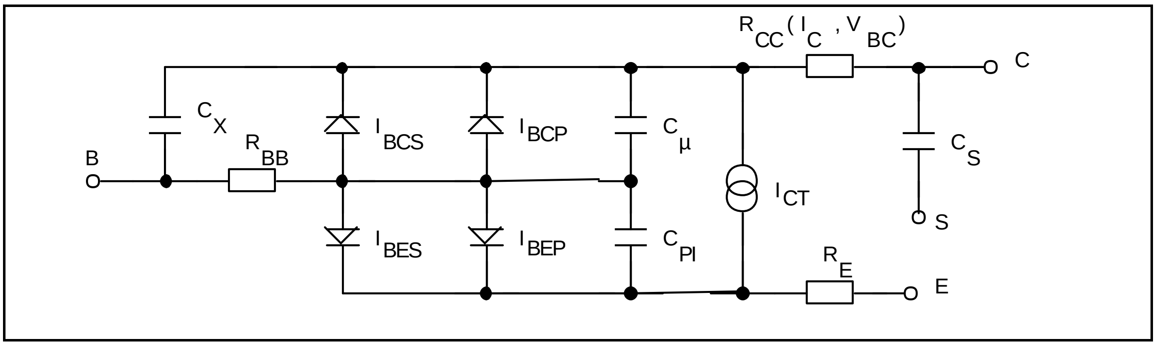

Fig. 1: CadenceSPICE - QS model equivalent circuit [2]

1.2 KULL - Model

Kull et.al. [3] have examined the QS mechanism and developed an appropriate model in 1985. It is based on an equation for the current through the epi layer. This is completed by a term for the carrier velocity dependence on the field strength. Kull has proposed the following relationship for the epi layer current as a function of the internal and external BC - voltage (V_{cbo} and V_{cbw}) and three additional model parameters R_{CO} , V_O and γ:

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[2] In this paper the original spelling is reproduced for original equations and figures. SPICE model parameters are written in capital letters

[3] The same equation (with NR = 1) is used in the SPECTRE simulator too.

$$Iepi = \frac{\left[K_1(V_{cbo}) - K_1(V_{cbw}) - \left[\ln\frac{1+K_1(V_{cbo})}{1+K_1(V_{cbw})}\right]\right] + \left[\frac{V_{cbo}-V_{cbw}}{V_t}\right]}{\frac{R_{CO}}{V_t} * \left[1+\frac{V_{cbo}-V_{cbw}}{V_o}\right]} \tag{2}$$

with

$$K_1(V) = \sqrt{1+g \ * \exp\left(\frac{V}{V_t}\right)} \tag{3}$$

$$g \ = \left[2ni / N_I\right]^2 \tag{4}$$

The charge storage elements are modelled by:

$$Qo = Qco\left[K_1(V_{cbo}) - 1 - \frac{g}{2}\right] \tag{5}$$

$$Qw = Qco\left[K_1(V_{cbw}) - 1 - \frac{g}{2}\right] \tag{6}$$

Fig. 2 shows the equivalent circuit of the Kull - model for an pnp - transistor. It includes as additional elements: the collector current source I_{EPI} , the parasitic transistor base - epi - substrate and the capacitances Q_O and Q_W , representing the additional charge stored in the epi layer.

Table 2: Kull - QS model parameters [3]

Parameter	Description	Unit	Default
R_{CO}	Epitaxial resistance under equilibrium condition	Ohm	-
V_o	Carrier saturation velocity voltage	V	-
γ	Epitaxial dope coefficient	-	-
Qco	Epitaxial charge coefficient	C	-

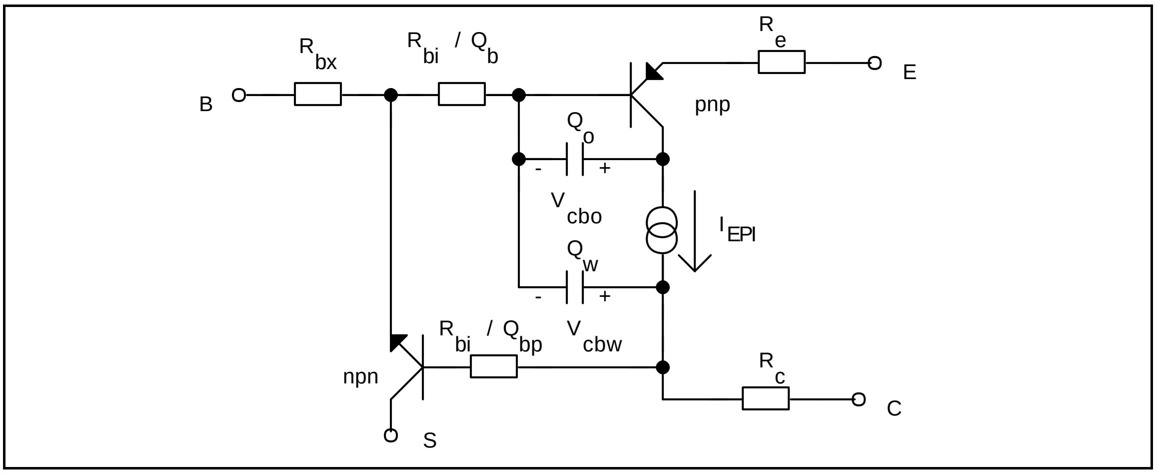


Fig. 2: Kull - QS model equivalent circuit [3]

1.3 PSPICE - Model

The PSPICE model for QS [4] is based on Kull's equation as well:

$$Iepi = \frac{VO * \left[V_t \left[k(Vbc) - k(Vbn) - \ln \left[\frac{1 + k(Vbc)}{1 + k(Vbn)} \right] \right] + Vbc - Vbn \right]}{RCO * \left[|Vbc - Vbn| + VO \right]} \tag{7}$$

with

$$K(v) = \sqrt{1 + GAMMA * \exp \frac{v}{vt}} \tag{8}$$

The storage charge is given by:

$$Qo = QCO * \left[K(Vbc) - 1 - GAMMA / 2 \right] \tag{9}$$

$$Qw = QCO * \left[K(Vbn) - 1 - GAMMA / 2 \right] \tag{10}$$

Table 3 shows the PSPICE - QS - model parameters. Additional to the usual constant collector resistance RC a QS - parameter RCO exists. This parameter acts as a switch too: for RCO > 0 the QS model is used.

Table 3: PSPICE - QS model parameters [4]

Parameter	Description	Unit	Default
GAMMA	Epitaxial region doping factor	-	1 E -11
QCO	Epitaxial region charge factor	C	0
RCO	Epitaxial region resistance	Ohm	0
VO	Carrier mobility knee voltage	V	10
ISS	Substrate saturation current	A	0
NS	Substrate emission coefficient	-	1

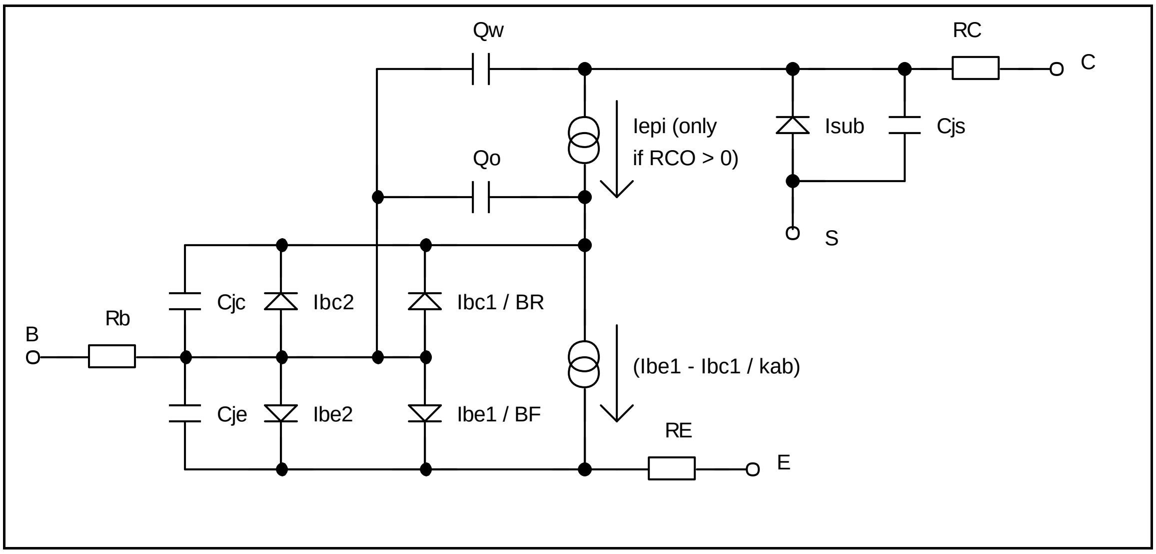


Fig. 3: PSPICE - QS model equivalent circuit [4]

1.4 HSPICE - Model

The HSPICE - QS model is also based on the Kull model. It is called the LEVEL2 - BJT model and defined as follows [5]:

$$Iepi = \frac{ki - kx - \ln\left[\frac{1+ki}{1+kx}\right] + \frac{vbc - vbcx}{NEPI * vt}}{\left[\frac{RC}{NEPI * vt}\right] * \left[1 + \frac{|vbc - vbcx|}{VO}\right]} \tag{11}$$

with

$$ki = \sqrt{1 + GAMMA * \exp\frac{vbc}{NEPI * vt}} \tag{12}$$

$$kx = \sqrt{1 + GAMMA * \exp\frac{vbcx}{NEPI * vt}} \tag{13}$$

$$GAMMA = \left[\frac{2n_i}{n_{epi}}\right]^2 \tag{14}$$

The emission coefficient NEPI is introduced here as an additional parameter (default = 1). The parasitic pnp substrate transistor is represented by the current source Ibs and the diode Isc.

$$Ibs = BRS * (Ibc - Isc) \tag{15}$$

where

Ibc = npn reverse transfer current, caused by forward biased BC - junction

and

$$Isc = ISS * \left[\exp\frac{vsc}{NS * vt} - 1\right] \tag{16}$$

From a physical point of view the current at the substrate terminal, in the npn reverse mode (forward biased BC junction), is the forward transport current of the parasitic pnp. This current Ibs is defined in the HSPICE model by eqn (15) using the parameter BRS and the currents Ibc and Isc. Note, that Ibc is the npn reverse transfer current (not the BC - diode current), caused by forward biased BC - junction. The charge storage elements qi and qx, respectively the capacitances Cx and Ci, are determined by:

$$qi = QCO * [ki - 1 - GAMMA / 2] \tag{17}$$

$$Ci = \left[\frac{GAMMA * QCO}{2 * NEPI * vt * ki}\right] * \exp\frac{vbc}{NEPI * vt} \tag{18}$$

and

$$qx = QCO * [kx - 1 - GAMMA / 2] \tag{19}$$

$$Cx = \left[\frac{GAMMA * QCO}{2 * NEPI * vt * kx}\right] * \exp\frac{vbcx}{NEPI * vt} \tag{20}$$

The HSPICE - QS parameters are summarized in Table 4. Note the difference in meaning of RC for LEVEL1 and LEVEL2. Fig. 4 shows the equivalent circuit of the model.

Table 4: HSPICE - QS model parameters [5]

Parameter	Description	Unit	Default
LEVEL	LEVEL1 = SGP - model, LEVEL2 = QS - model	-	1
GAMMA	Epitaxial doping factor	-	0
NEPI	Emission coefficient	-	1

QCO	Epitaxial charge factor	C	0
RC	LEVEL1 = ohmic collector resistance, LEVEL2 = epitaxial layer resistance under equilibrium condition	Ohm	0
VO	Carrier velocity saturation voltage (zero indicates an infinite value)	V	0
BRS	Current gain of the parasitic substrate transistor	-	1
ISS	Reverse saturation current bulk to collector	A	0
NS	Substrate emission coefficient	-	1

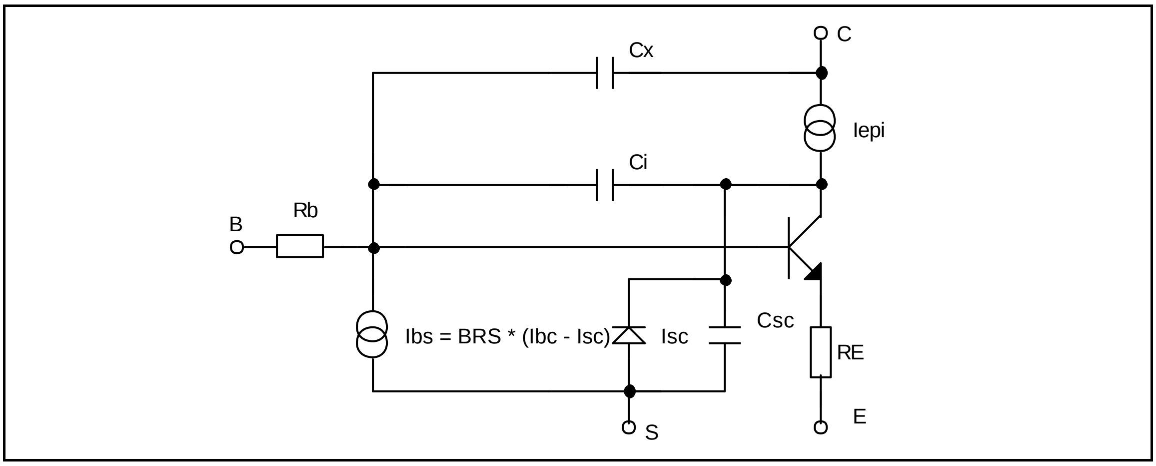


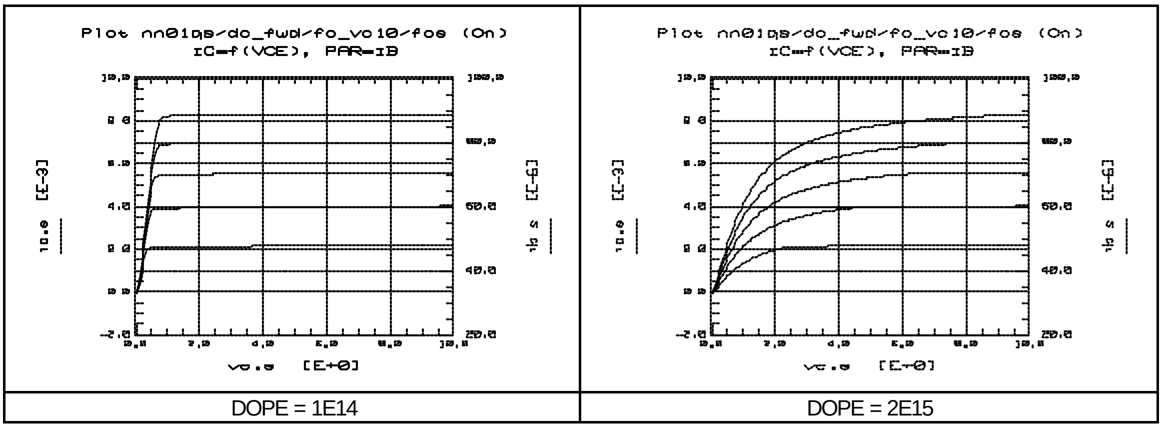
Fig. 4: HSPICE - QS model equivalent circuit [5]

2 Quasisaturation Model Parameters

In this section the effects of the QS parameters for two models, the CdsSPICE - and the HSPICE - model, on BJT characteristics are investigated.

2.1 CadenceSPICE - Model

If the parameter DOPE is specified, the output characteristics are bended, because the effective collector resistance is decreased for decreasing voltage V_{CE} (Fig. 5).

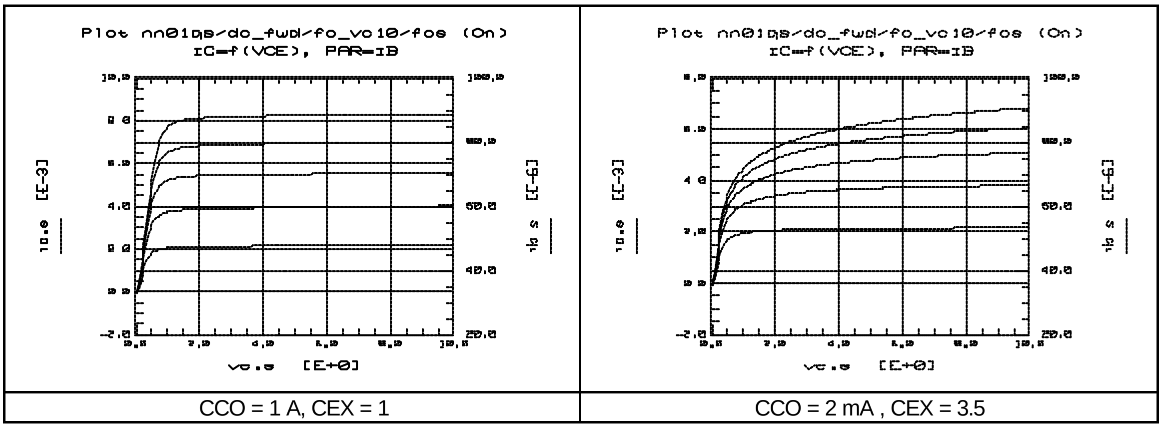


DOPE = 1E14

DOPE = 2E15

Fig. 5: Effect of DOPE on the output characteristic

The reference current CCO and the exponent CEX create a shift in the output characteristic at high currents (Fig. 6), e.g. for a decreasing value of CCO the effective collector resistance is increased and thus the collector current is reduced.



CCO = 1 A, CEX = 1

CCO = 2 mA , CEX = 3.5

Fig. 6: Effect of CCO and CEX on the output characteristic

RCV, the variable collector resistance, determines the rise of the output characteristic in the saturation mode, if the other parameters are default (Fig 7). Typical values are about 1K.

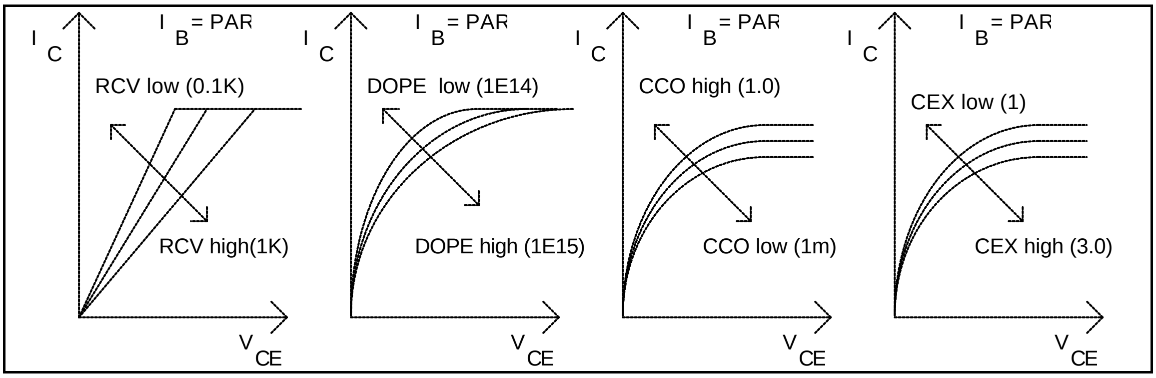


Fig 7: Effect of RCV, DOPE, CCO and CEX on the output characteristic

2.2 HSPICE - Model

Examining the effect of the HSPICE - QS parameters GAMMA, VO and RC[4] it is useful to consider eqn (11) more in detail. Rearranging this relationship for NEPI = 1, it reduces to:

$$Iepi = \frac{V_{COR} + V_{EPI}}{RC * \left[1 + \frac{V_{EPI}}{VO} \right]} \tag{21}$$

where

$$V_{COR} = V_T \left[ki - kx - \ln \left[\frac{1 + ki}{1 + kx} \right] \right] \tag{22}$$

and

$$V_{EPI} = V_{BCi} - V_{BCx} \tag{23}$$

V_{COR} here is a correction voltage, depending both on epi layer doping GAMMA and the epi layer voltage drop V_{EPI} . This correction voltage is added to V_{EPI} . In the active normal case ($V_{BCx} < 0$, $V_{BCi} < 0$) V_{COR} is zero, but in the QS V_{COR} increases. For a given RC value this results in an increasing value for the current source Iepi. In this way the effective collector resistance is reduced. The carrier velocity saturation effect is modelled using the parameter VO. If the epi layer voltage drop reaches VO, the effective collector resistance increases and Iepi is decreased. The effect of GAMMA, VO and RC on the output characteristic was examined with simulations using

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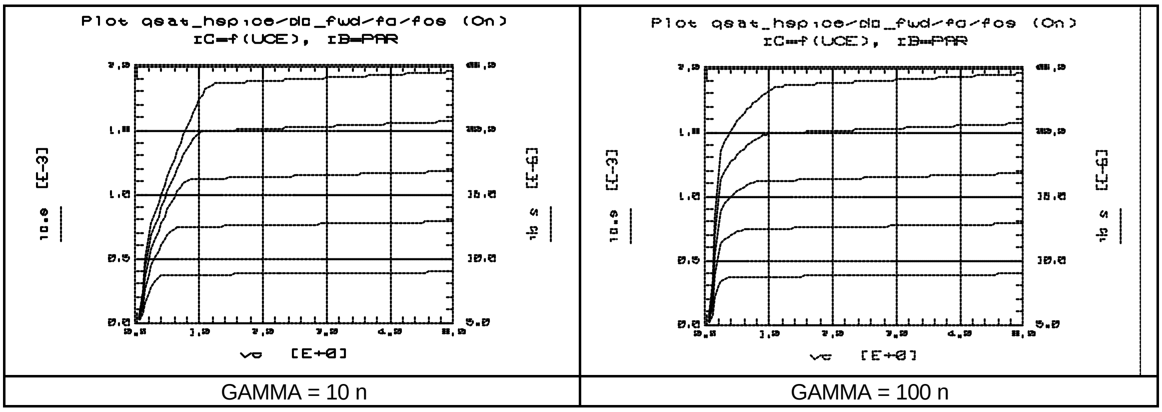
[4] Note: In the HSPICE - QS model the epi resistance parameter is unfortunately called RC instead of RCO, e.g. the LEVEL1 parameter RC is **not** identical to the LEVEL2 parameter RC.

HSPICE 93a. Each parameter was varied using the following default values for the other parameters: LEVEL = 2, RC = 500, VO = 0, QCO = 0, GAMMA = 0, NEPI = 1, ISS = 1a, IS = 400a, BF = 75, VAF = 75. The effect of QCO was investigated using the full parameter set of transistor nn01 for a transit frequency simulation. (see Fig. 8 to Fig. 12). We can summarize the effect of the HSPICE - QS parameters as following:

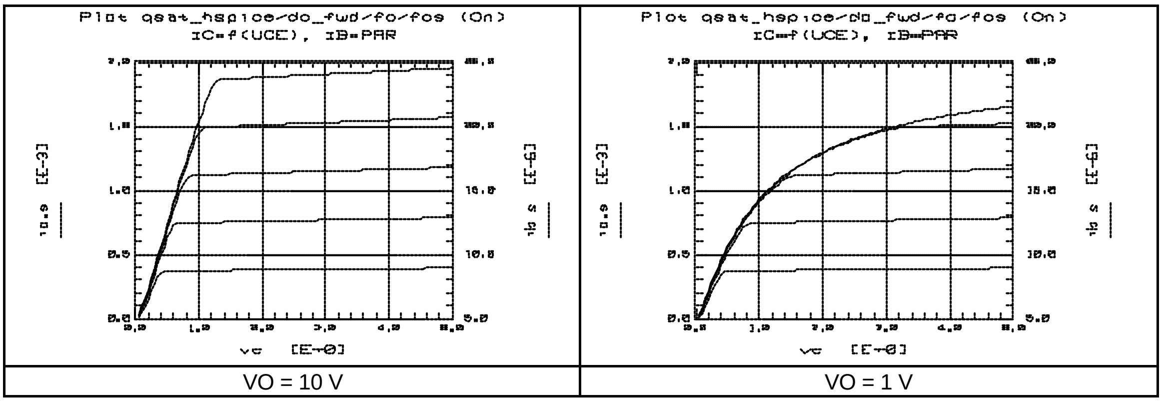
- GAMMA causes a flattening in the QS range of the output characteristic. High GAMMA values correspond to low increase of the characteristic. Meaningful GAMMA values are in the range of 1 ... 100 n.
- VO causes a rounding off in the QS range of the output characteristic. The lower the VO value, the higher the effect. Meaningful VO values are in the range of 1 ... 10 V.
- RC acts as the normal ohmic resistance for VO = ∞ and GAMMA = 0 (Fig. 10). If VO and GAMMA are specified, RC acts in the QS range as shown in Fig. 13. Changing the slope of the output characteristic in the saturation range for the HSPICE model is only possible using an additional constant collector series resistance (RCex). Note, that the value of the LEVEL2 - parameter RC is distinctly higher than the RC value of the SGP - model. Meaningful RC values are in the range of 1 ... 10 K
- QCO causes a sharp decrease of the f_T - characteristic, if the device reaches QS (Fig. 12).



GAMMA = 10 n

GAMMA = 100 n

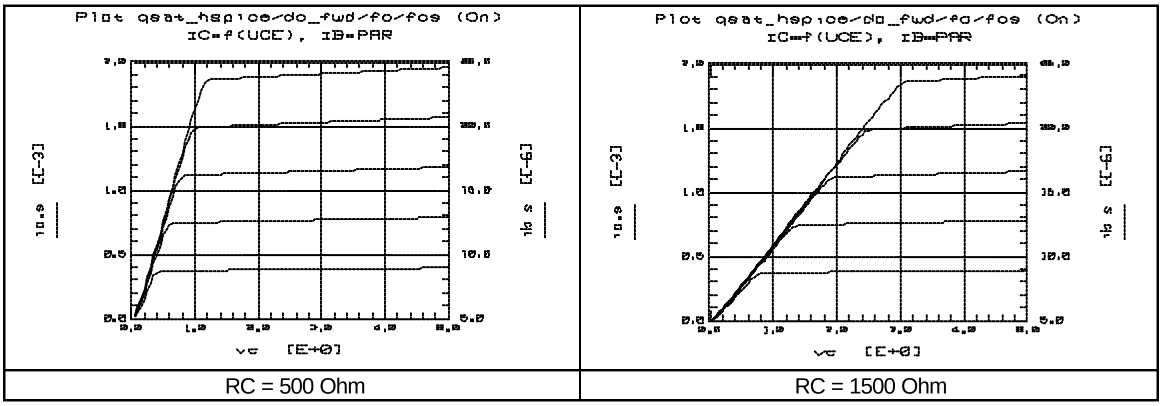
Fig. 8: Effect of GAMMA on the output characteristic



VO = 10 V

VO = 1 V

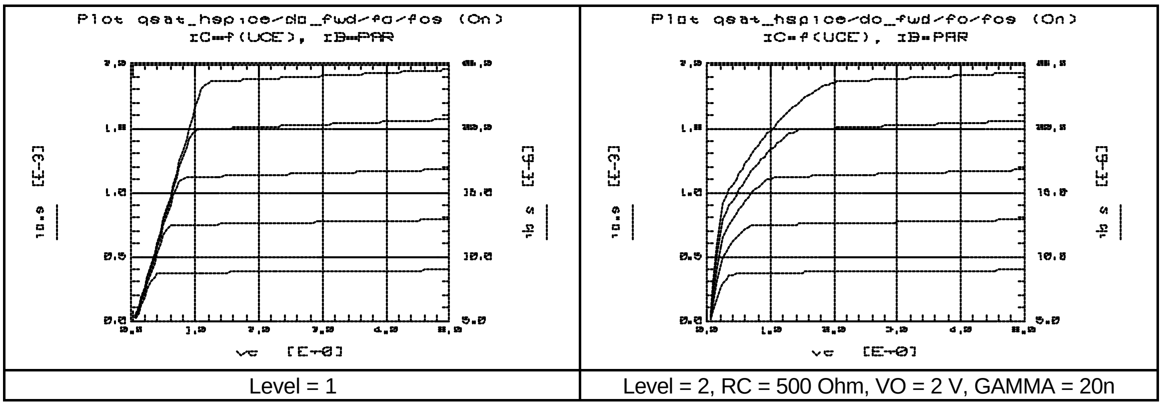
Fig. 9: Effect of VO on the output characteristic



RC = 500 Ohm

RC = 1500 Ohm

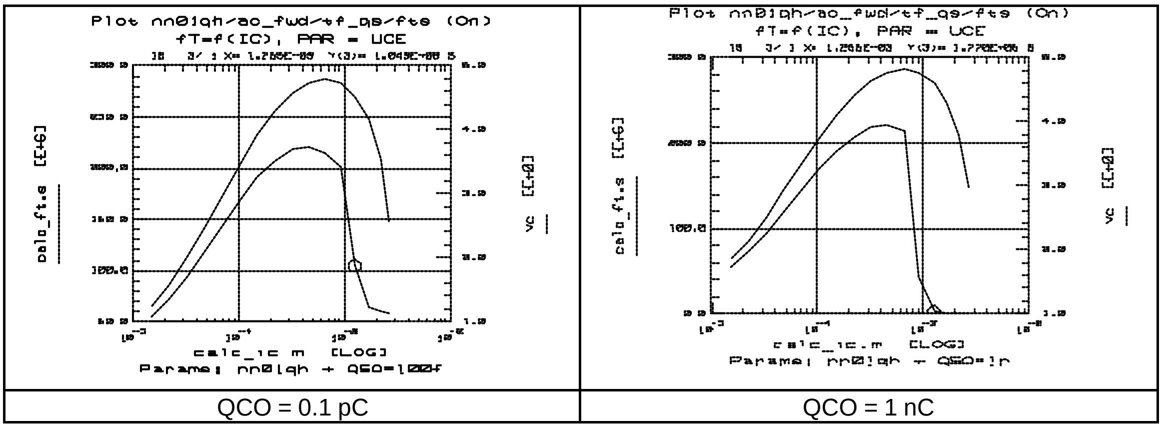
Fig. 10: Effect of RC on the output characteristic



Level = 1

Level = 2, RC = 500 Ohm, VO = 2 V, GAMMA = 20n

Fig. 11: Effect of GAMMA, VO and RC on the output characteristic



QCO = 0.1 pC

QCO = 1 nC

Fig. 12: Effect of QCO on the f_T - characteristic

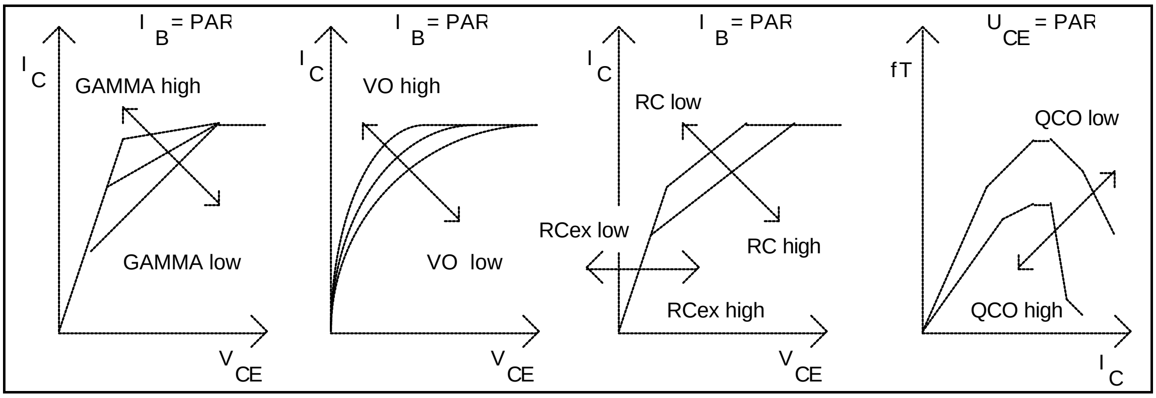


Fig. 13: Effect of the model parameters GAMMA, VO, RC, RCex and QCO

3 Quasisaturation Model Parameter Extraction

This section outlines procedures to determine the parameters for the CadenceSPICE - and the HSPICE - QS model. The DUT used for parameter extractions is an integrated npn transistor nn01, characterised by the following typical values: V_{CE} = 50 V, I_{Cmax} = 5 mA, A_E = 16 x 16 µm^2.

3.1 CadenceSPICE - Model

3.1.1 RCM, RCV, DOPE, CCO and CEX

A number of methods is available for the determination of RCM, according to the usual collector resistance determination, e.g. the forced beta method or the ΔIsub - method [6]. RCV is given by the reciprocal increase of the line between the origin and the point, where the onset of QS takes place for high currents (Fig. 14). In the next step we can determine DOPE by an optimization on a medium current output characteristic. The parameters CCO and CEX may be determined then by an optimization in the high current range of the output characteristic (Fig. 15). Useful start values are found asCCO = 10 x I_C (B_{Nmax}) and CEX = 2. Using this way, we found the following QS - parameters for nn01: RCM = 55, RCV = 1000, DOPE = 4E14, CCO = 2 m and CEX = 3.5.

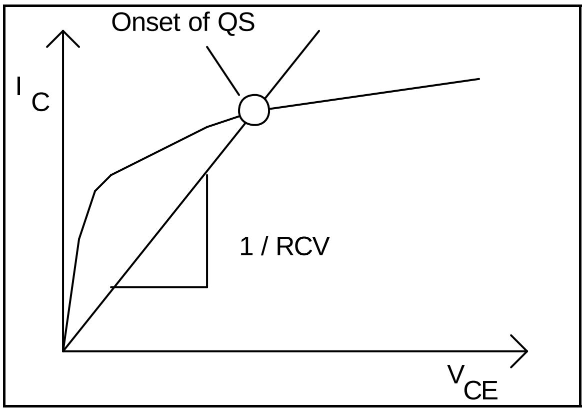


Fig. 14: Determination of RCV

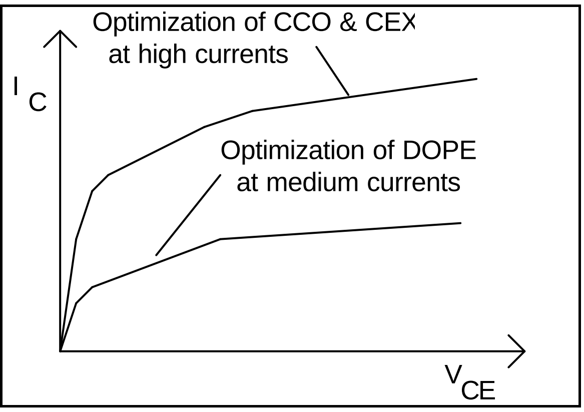


Fig. 15: Determination of DOPE, CCO, CEX

3.1.2 Model Parameter Set

The following table shows the full parameter set for nn01, used for the CadenceSPICE simulations in the section Results

Table 5: CadenceSPICE model parameter set

Parameter	Value	Parameter	Value	Parameter	Value
DOPE	4E+14	RCV	1000	RCM	55
CCO	0.002	CEX	3.5	IS	1.41E-16
BF	125	ISE	1.8E-16	IKF	0.02
NE	1.45	NF	0.986	BR	0.28
ISC	2E-16	IKR	0.1	NC	1.1
NR	0.995	VAF	244	VAR	18
RE	2.43	RB	160	RBM	40
IRB	2E-05	TF	4.6E-10	XTF	10
VTF	8	ITF	0.02	TR	2E-07
PTF	0	XCJC	1	FC	0.85
CJE	5.44E-13	VJE	0.664	MJE	0.308
CJC	4.44E-13	VJC	0.415	MJC	0.262
CJS	1.183E-12	VJS	0.415	MJS	0.318
XTB	1.043	XTI	3.244	EG	1.11

3.2 HSPICE - Model

3.2.1 VO , RC and GAMMA

For the extraction of VO we have to choose two points P1 and P2 in the output characteristic, at which the transistor nearly enters the QS range (Fig. 16) [3]. At these point we can assume V_{BCi} ≈ 0 (ki ≈ 1) and V_{BCx} << 0 (ki ≈ 1). Therefore we have for the correction voltage V_{COR} = 0 and eqn (21) reduces to:

$$Iepi = \frac{V_{BCx}}{RC * \left[1 + \frac{V_{BCx}}{VO} \right]} \tag{24}$$

Rearranging this equation and using the values of point P1 and P2, we have for VO:

$$VO = \frac{V_{BCx1} * V_{BCx2}\left(I_{C1} - I_{C2}\right)}{V_{BCx1} * I_{C2} - V_{BCx2} * I_{C1}} \tag{25}$$

Fig. 17 shows a characteristic $I_C = f(V_{CB})$, simulated using the following parameters: RC = 500, GAMMA = 20 n, VO = 1 , NEPI = 1, ISS = 1 a, IS = 400 a, BF = 75, VAF = 75.
Using the values P1: V_{CB} = 2,65 V / I_C =1,5 mA and P2: V_{CB} = 7,25 V / I_C =1,825 mA the parameters VO and RC can be re-calculated as VO = 1.03 V and RC = 496 Ohm. This result is in good agreement with the start values, but note, that the result is depending on the chosen points P1 and P2. Unfortunately it is impossible to give an exact definition for the selection of P1 and P2, but I_C and V_{BC} should be sufficient high referring to the transistors maximum ratings.
We can calculate GAMMA using eqn (14), but in practice this value is often useless. That is why we should consider GAMMA as a fitting parameter. An optimization on the output characteristic using the known values of VO and RC seems to be the best way for GAMMA determination.

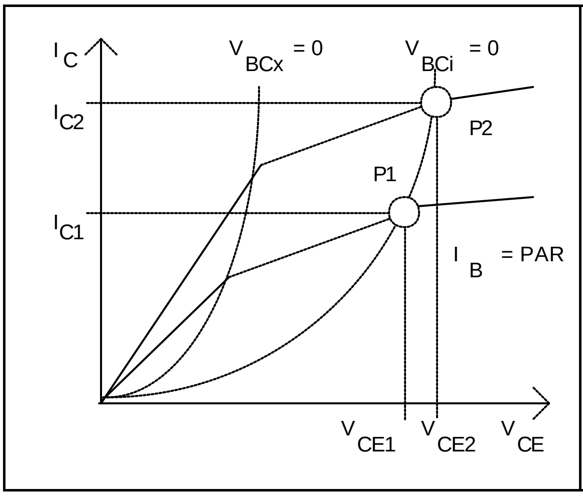


Fig. 16: Determination of VO and RC

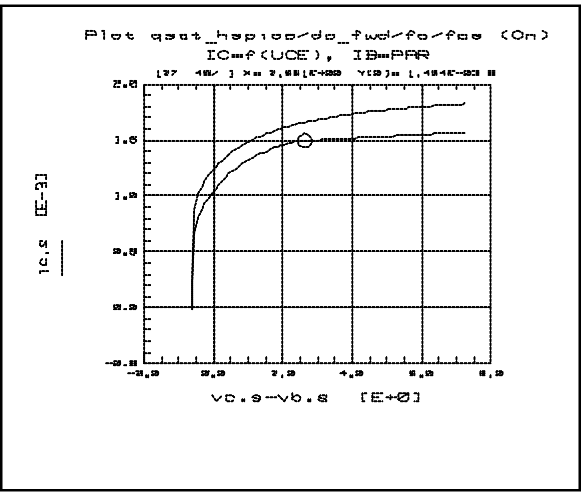


Fig. 17: Simulated characteristic
$I_C = f(V_{CB})$, I_B = 20, 30 µA

3.2.2 BRS and BR

The parameters BRS (for the parasitic pnp transistor) and BR (for the npn transistor) affect each other, so we have to consider both together. For the determination of these parameters we consider the npn transistor in the reverse mode, the parasitic pnp transistor works then under forward mode conditions (Fig. 18).

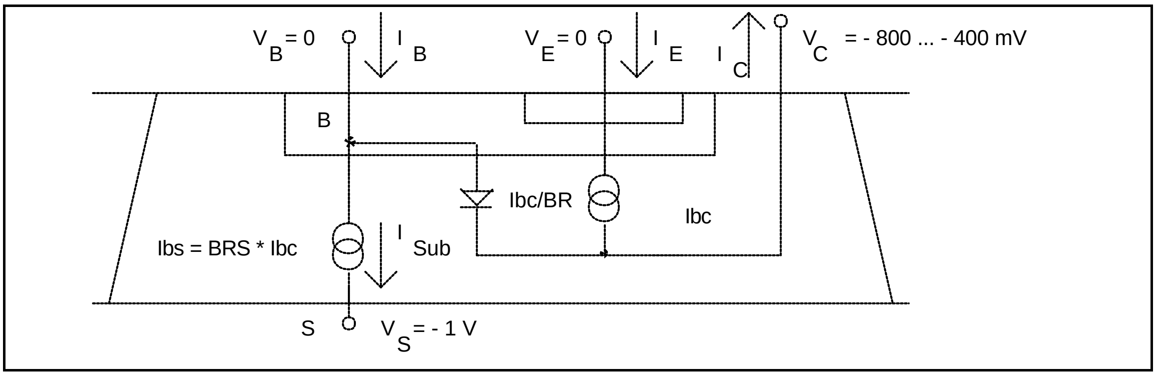


Fig. 18: Determination of BRS and BR, npn reverse mode

The epi - substrate diode current Isc is zero in this mode (Isc = 0 for $V_{sc} \le 0$) and eqn (15) for the substrate current source Ibs reduces to the form:

$$Ibs = BRS * Ibc \tag{26}$$

We can simply measure now both the currents Ibc and Ibs. Ibc, the reverse transfer current of the npn transistor, is the current at the emitter terminal (Ibc = I_E). Ibs, the forward transfer current of the parasitic pnp transistor, is the substrate terminal current (Ibs = I_{Sub}). Thus we get the parameter BRS as the ratio of two transfer currents. In practice, this ratio is current dependent and we can use the maximum value as the model parameter:

$$BRS = \left[\frac{I_{Sub}}{I_E}\right]_{MAX} \tag{27}$$

The reverse current gain of an npn - transistor for the SGP - model is usually defined as the ratio of the reverse transfer current (I_E) and the appropriate base current (I_B):

$$B_I = I_E / I_B \tag{28}$$

Using the LEVEL2 - model we have to work with a modified definition of the npn inverse current gain, because the base terminal current I_B is now divided in two parts: the substrate current I_{Sub} and the true npn base current I_{Bnpn} :

$$B_I = \frac{I_E}{I_B - |I_{Sub}|} \tag{29}$$

The ratio B_I is current dependent too. We can use the maximum B_I - value for the model parameter BR again.

$$BR = B_{\mathrm{I}max} \tag{30}$$

3.2.3 ISS and NS

Based on the definition in eqn (16) we have to measure the forward biased substrate - epi diode for the determination of the parameters ISS and NS. As shown in Fig. 19, we can measure this current at the collector terminal (Isc = I_C).

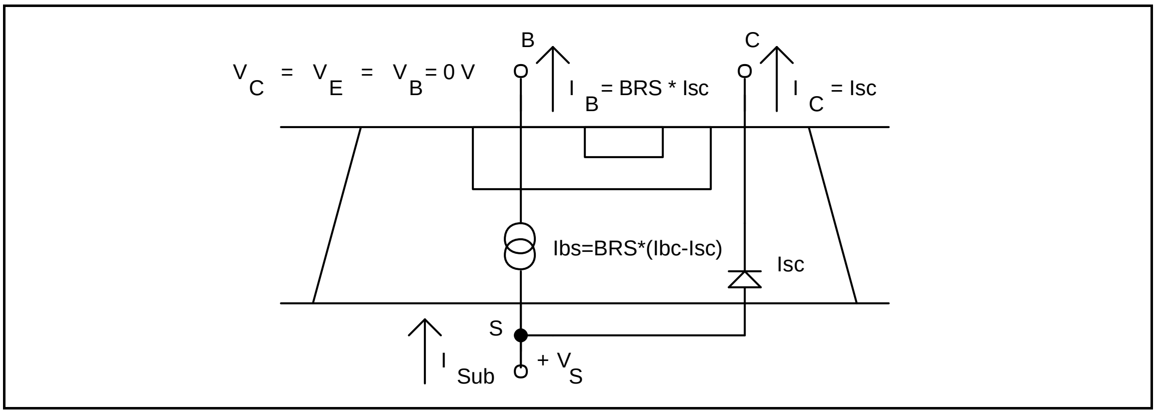


Fig. 19: Determination of ISS and NS, forward biased substrate - epi junction

We can apply the following two - point equations on the diode characteristic I_C = f (V_{SC}) with V_{BE} = V_{BC} = 0 for the calculation of ISS and NS:

$$ISS = \frac{I_{C1}^{\left[\frac{V_{SC2}}{V_{SC2}-V_{SC1}}\right]}}{I_{C2}^{\left[\frac{V_{SC1}}{V_{SC2}-V_{SC1}}\right]}} \tag{31}$$

and

$$NS = \frac{V_{SC2} - V_{SC1}}{V_T \ln \frac{I_{C2}}{I_{C1}}} \tag{32}$$

There seems to be a second possibility to determine the parameter BRS. For $V_{BC} = 0$ the substrate current source Ibs reduces to the form

$$Ibs = BRS * (-Isc) \tag{33}$$

and BRS is given, using the terminal currents I_B and I_C :

$$BRS = I_B / I_C \tag{34}$$

However, this value (typical < 1) strongly deviates from the BRS - value based on eqn (27). This contradiction is caused by the model definition itself: the substrate current source Ibs is defined referring two currents using only one gain parameter BRS. Consequently the model under this bias conditions is not able to simulate exactly the base and the substrate terminal currents.

3.2.4 QCO

The best way for determining QCO is a fit on a f_T - characteristic under QS conditions. Note, that the DC - QS parameters affect the f_T - characteristic too. Thus an additional fit of the parameters CJE, CJC, TF and ITF was necessary for transistor nn01, before QCO was fitted.

3.2.5 Extraction Strategy

The parameters BR and BRS affect the division of the base terminal current and consequently the simulated output characteristic. Therefore it is useful to determine first BR and BRS and then RC, VO and GAMMA (Table 6). In addition, it was found advantageous to use an external collector resistance RCex. Using the described extraction methods, we determined for the transistor nn01 the following QS - parameters: RCEX = 55, RC = 365, VO =2.07, GAMMA = 120 n, BRS = 10, ISS = 3.256 f, NS = 1.008, QCO = 100 f, BR = 30.

Step	Parameter	Characteristic	Condition	Mode
1	ISS, NS	$I_C = f(V_{SC})$	$V_{BE} = V_{BC} = 0$	Substrate diode = Forward biased
2	BR, BRS	$I_{Sub}, I_B, I_E = f(V_{BC})$	$V_{BE} = V_{SC} = 0$	NPN = Reverse mode
3	VO, RC	$I_C = f(V_{CB})$	I_B = PAR	NPN = QS
4	GAMMA	$I_C = f(V_{CE})$	I_B = PAR	NPN = QS
5	QCO	$f_T = f(I_C)$	V_{CE} = PAR	NPN = QS

Table 6: Extraction strategy

3.2.6 Model Parameter Set

The full parameter set, used for the simulations in the section Results is shown in Table 7.

Table 7: HSPICE - LEVEL2 model parameters for nn01

Parameter	Value	Parameter	Value	Parameter	Value
RCEX	55.00	BR	30.00	CBE	325.0f
LEVEL	2.000	NR	996.5m	CBC	197.0f
RC	365.0	VAR	18.00	CCS	952.0f
VO	2.070	IKR	100.0m	CJE	744.0f
GAMMA	120.0n	ISC	200.0a	CJC	644.0f
BRS	10.00	NC	1.100	CJS	1.183p
ISS	3.256f	RB	160.0	VJE	664.0m
NS	1.008	IRB	20.00u	VJC	415.0m
QCO	100 f	RBM	40.00	VJS	415.0m
IS	141.2a	RE	2.430	MJE	308.0m

BF	125.0	TF	320.0p	MJC	262.0m
NF	989.6m	XTF	10.00	MJS	318.0m
VAF	244.0	VTF	8.000	XCJC	1.000
IKF	20.00m	ITF	1K	FC	850.0m
ISE	180.0a	PTF	0.000	XTB	1.043
NE	1.450	TR	200.0n	XTI	3.244

4 Results

4.1 CadenceSPICE - Model

Using the CadenceSPICE - model, we simulated the output characteristic both in a low and a high current range of the DUT. As can be seen, the simulation results are sufficiently in agreement with the measurements.

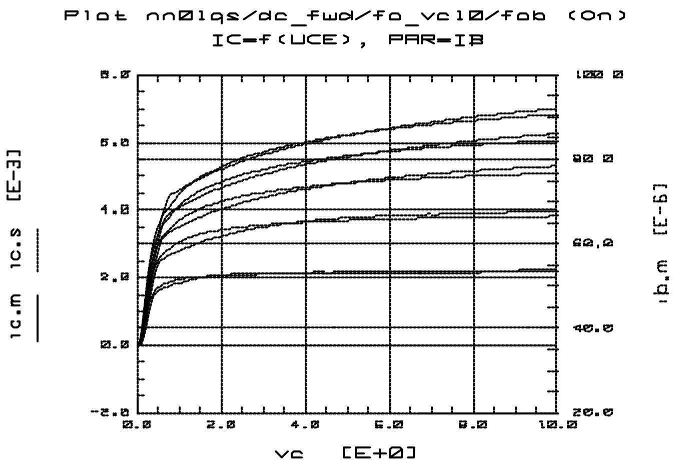


Fig 20: Output characteristic $I_C = f(V_{CE})$, I_B = 20 ... 100 µA, V_{CE} = 0 ... 10 V

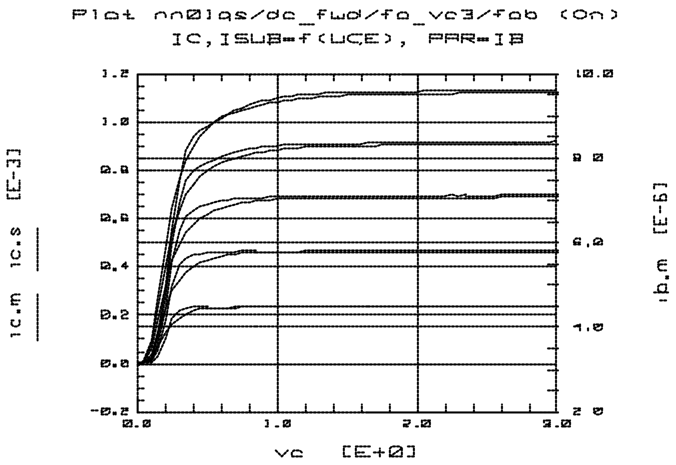


Fig 21: Output characteristic $I_C = f(V_{CE})$, I_B = 2 ... 10 µA, V_{CE} = 0 ... 3 V

4.2 HSPICE - Model

As the output characteristic (Fig. 22 and Fig. 23) shows, the typical QS flattening is well reproduced by the model. However, the substrate current, additional shown in both the characteristics, is overestimated by the simulation in the QS range. This illustrates the main disadvantage of the model, caused by the simplified definition of the substrate current source Ibs. The forward transfer characteristic demonstrates the inaccurate substrate current simulation of the HSPICE - QS - model too: while the measured current increases at $V_{BE} \approx$ 800 mV, the simulated current increases already at 740 mV (Fig. 24). The reverse mode simulation illustrates, that. I_B and I_{Sub} are nearly identical (Fig. 25). This means, the npn base current (I_B - I_{Sub}) is relative small an BR reaches a high value (about 30 for nn01). For the reverse mode of the parasitic pnp transistor (substrate - epi junction forward biased), the model simulates only the current at the collector terminal exactly (Fig. 26). Because of the substrate current source definition (only one gain BRS for both directions) the simulated currents at the substrate and the base terminal are much to high. Usually this disadvantage is not important, because this mode in most cases is not allowed for an integrated transistor.

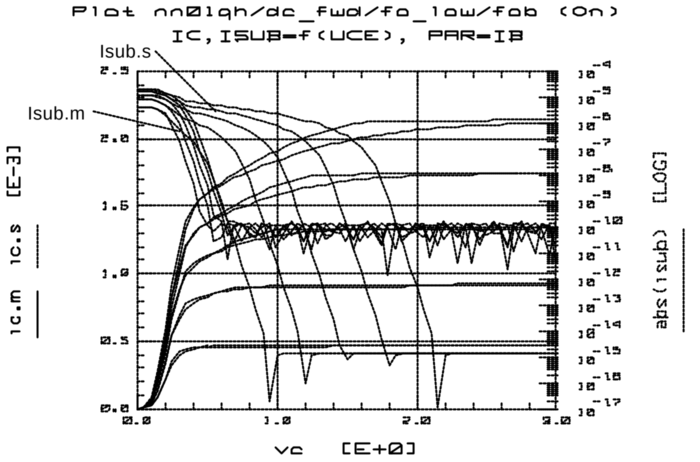


Fig. 22: Output characteristic I_C , I_{Sub} = f (V_{CE}), V_{CE} = 0 ... 3 V, I_B = 4 ... 20 µA, Step = 4 µA

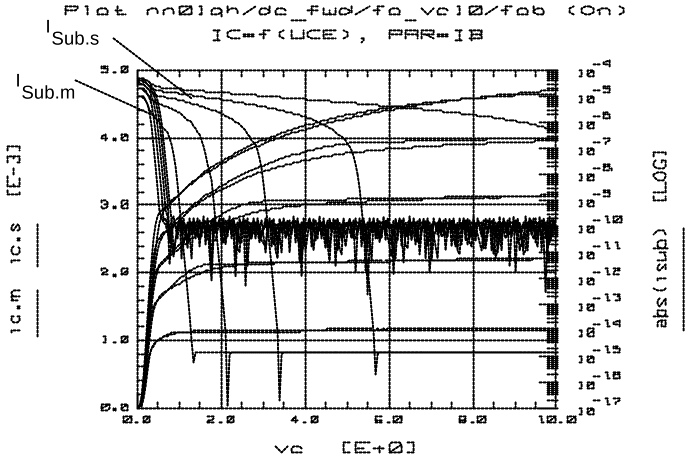


Fig. 23: Output characteristic I_C ,I_{Sub} = f (V_{CE}), V_{CE} = 0 ... 10 V, I_B = 10 ... 50 µA,
Step = 10µA

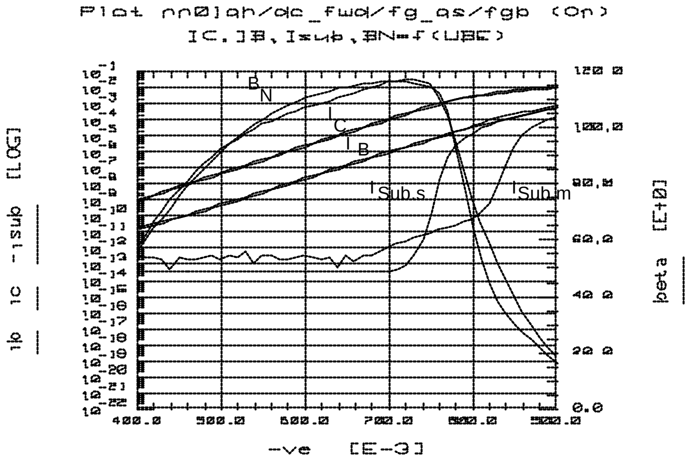


Fig. 24. Forward Gummel plot B_N , I_C , I_B , I_{Sub} = f (V_{BE}), V_{SC} = - 1V, V_{BC} = 0 V

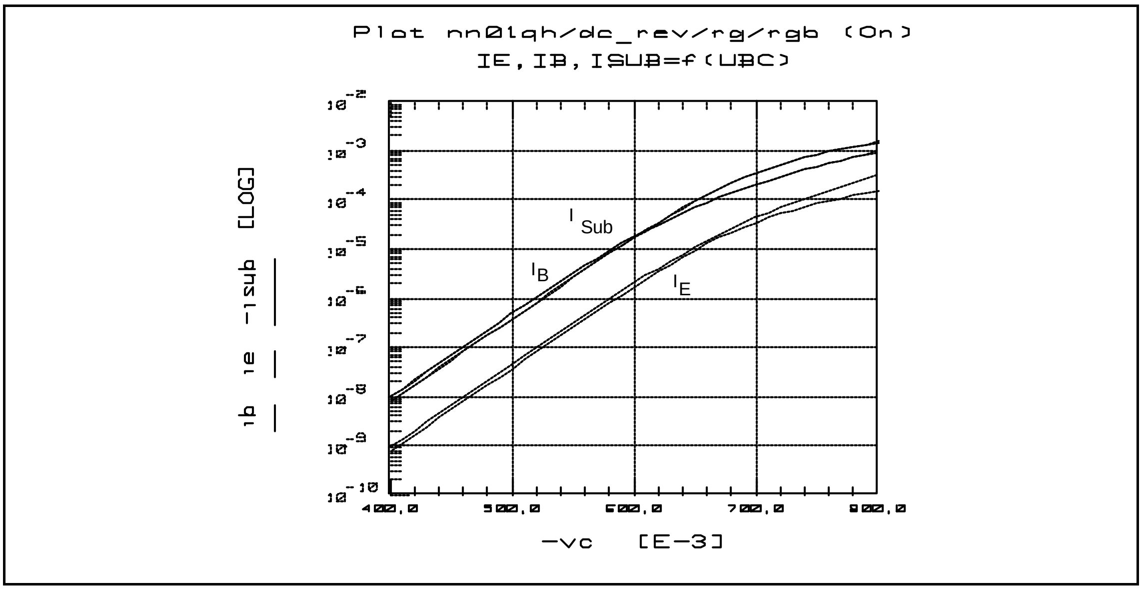


Fig. 25: Reverse Gummel plot I_E , I_B , I_{Sub} = f (V_{BC}), V_{SC} = - 1V, V_{BE} = 0 V

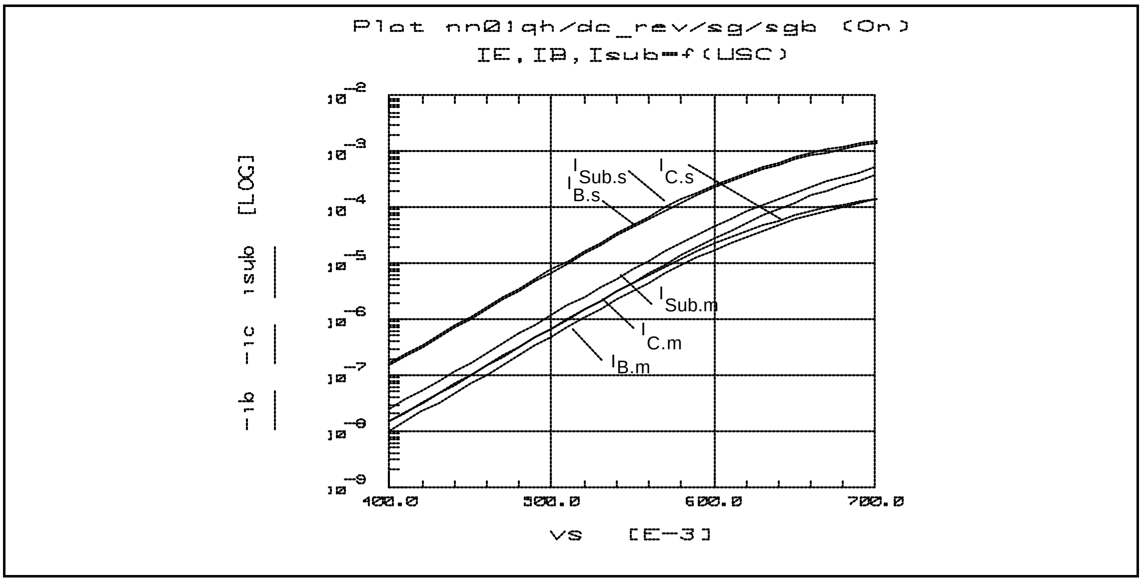


Fig. 26: PNP reverse mode Gummel plot I_C , I_B , I_{Sub} = f (V_{SC}), V_{BE} = V_{BC} = 0 V

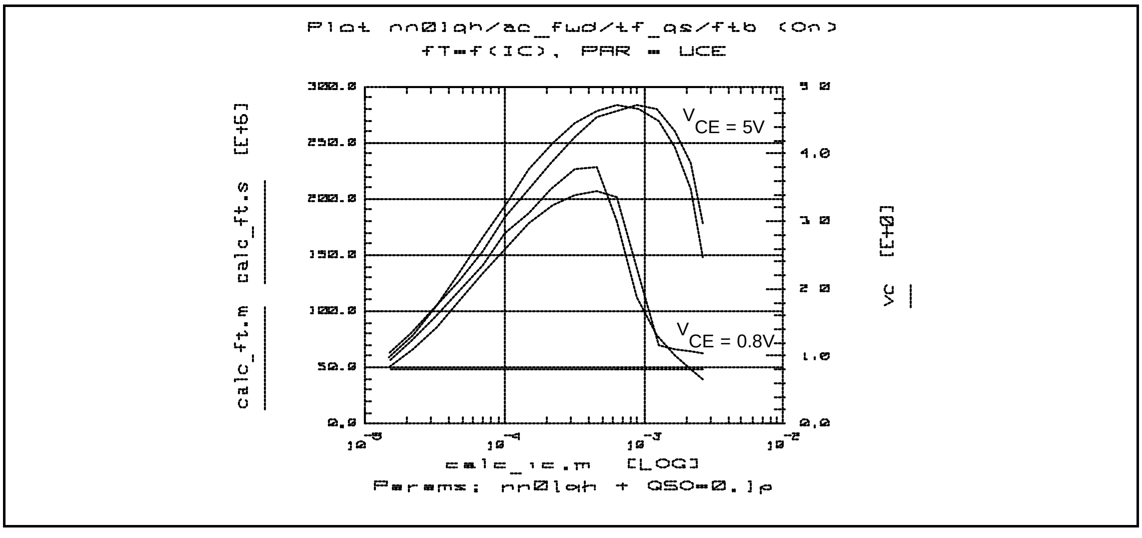


Fig. 27: Transit frequency characteristic f_T = f (I_C), V_{CE} = 0.8 V, 5V

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