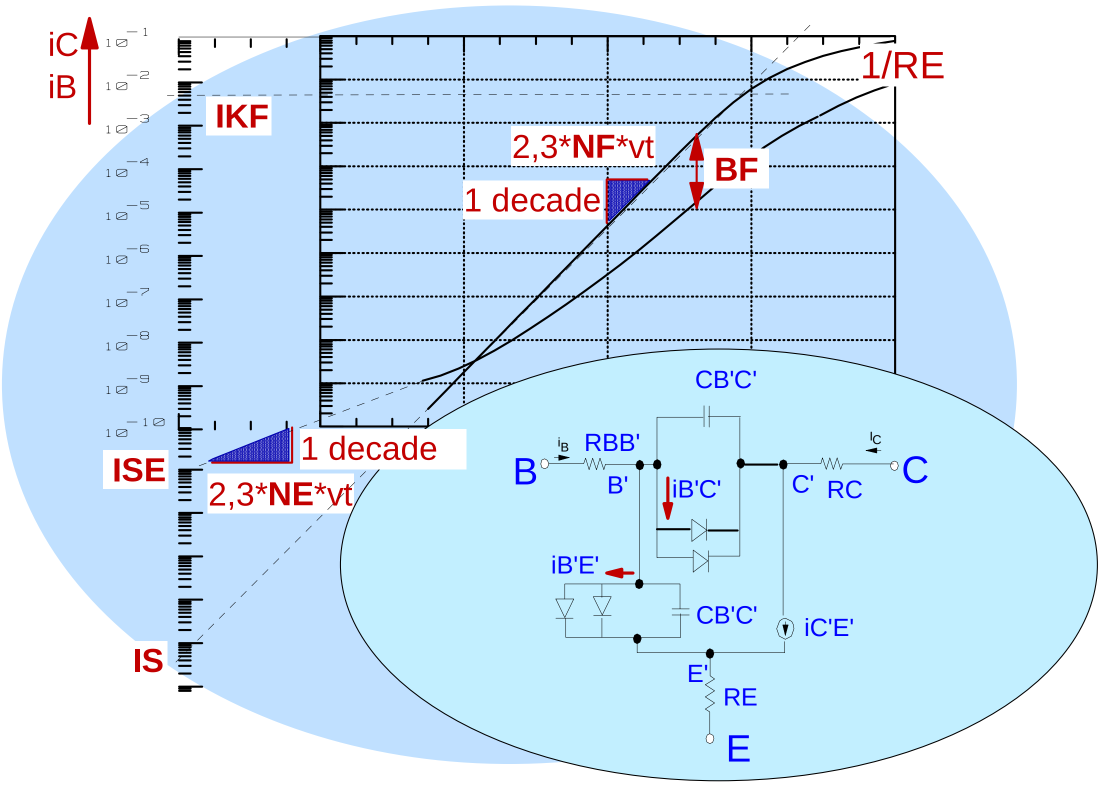
GUMMEL-POON BIPOLAR MODEL

MODEL DESCRIPTION

PARAMETER EXTRACTION



Agilent Technologies GmbH, Munich

STRUCTURE OF THIS MANUAL

This product has been developped to meet the local demands of European IC-CAP users for more technical background information on extraction techniques and for the availability of extraction source code.

Published for the first time in 1990, it has been updated since then several times.

It is part of a series of supplementary modeling toolkits for the IC-CAP users. These products feature source code and detailed technical description of the extraction routines.
Please contact the author for further information.

The author would like to thank the many users for valuable inputs, and is hoping for fruitful discussions also in the future.

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ABOUT THIS MANUAL

This manual is intended to explain the basics of modeling a bipolar transistor using the Gummel-Poon model as it is implemented in the simulation program SPICE of the University of California Berkeley (UCB) /see publication list/.
It is part of the Gummel-Poon Bipolar Model Parameter Extraction Toolkit.

This toolkit includes the IC-CAP model file

GP_CLASSIC_NPN.mdl	the MASTER model file which is described in this manual

and featuring the data management features of IC-CAP 5.x, i.e. separating measurements from extractions:

NPN_MEAS_MASTER.mdl	a master file for measurement
GP_EXTRACT_NPN.mdl	a master file for modeling

as well as manyother IC-CAP model files covering topics like:

model parameter extraction using the tuner feature
direct visual parameter extractions
alternate modeling methods for DC- CV- and RF-parameters.
bipolar transistor modeling including the parasitic transistor.

<u>Note:</u>
After you have become familiar with the modeling procedure itself, i.e. file GP_CLASSIC_NPN.mdl, you are encouraged to split the modeling into 2 parts: separate measurements and separate extraction strategy. In this case, all measurements are performed using the file NPN_MEAS_MASTER.mdl. Then, the data are exported into IC-CAP mdm files (ASCII files) and imported into the master extraction file GP_EXTRACT_NPN.mdl for extraction. This method allows to improve continuously the extraction strategy file, independent of the measurement data!

IMPORTANT NOTE:
This manual and the underlying IC-CAP model file GP_CLASSIC_NPN.mdl are intended to explain the basics of the Gummel-Poon modeling. Therefore, it covers the classical Gummel-Poon model without enhancements for also modeling the parasitic transistor.
However, as stated above, such model files are included in the file sets of this toolkit. Please see the README macros in these IC-CAP model files for more details.

You are also invited to get in contact with the author for assistance with such modeling problems.

The IC-CAP model file "GP_CLASSIC_NPN.mdl" features:

The extractions are written using PEL (parameter extraction language) and are open to the user. They can be easily modified to meet specific user needs.

Subcircuit model description, open for user enhancements (HF modeling, parasitic pnp etc.).

All transistor pins are connected to SMUs for flexible measurements

The transistor output characteristic and S-parameter measurements use a Base current stimulus rather than a Base-Emitter voltage in order to avoid 1st order thermal effects being visible. However, self-heating might be present and affect the Gummel plots in the ohmic range.
-See also the file GP_MEAS_MASTER.mdl

Organization of the chapters in this manual:
There are 5 main chapters, which explain how to determine the model parameters from CV (capacitance versus voltage), then parasitic ohmic resistors, and DC, to finally high frequency measurements using network analyzers.

More chapters cover side aspects of bipolar modeling.

The individual chapters follow always this scheme:
- explanation of the parameter-dependent measurement setup
- explanation of the mathematical basics for the parameter extraction
- explanation of the parameter extraction
- explanation about how to use the IC-CAP file.

I N T R O D U C T I O N :

C O N T E N T S:

This manual describes the modeling of a bipolar transistor using the Gummel-Poon model as implemented in the simulator SPICE.

It should be mentioned that the Gummel-Poon model itself covers only the internal part of a real-transistor. Therefore, on-wafer parasitics like a parasitic pnp transistor are not covered. Also, packaging parasitics and other non-ideal effects are not part of the model. However, they can be added by using a sub-circuit rather than just the stand-alone model.

Please check the example files included in the file directory of this toolkit for examples.

Parasitic effects is specially important for network analyzer (NWA) measurements. The modeling procedures presented in this manual refer to already de-embedded measurements.

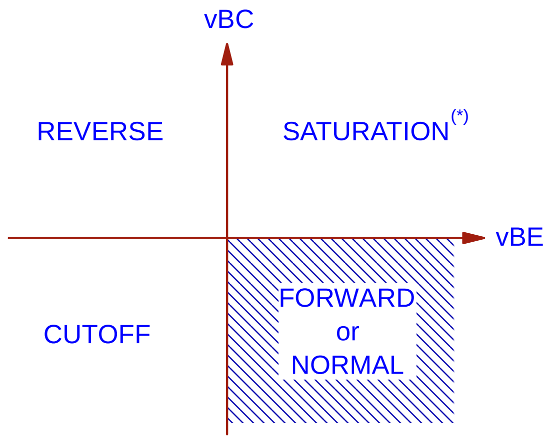
For on-wafer measurements, test probes that allow NWA calibrations down to the chip (like Cascade or Picoprobe probes) are commonly used. De-embedding means here to eliminate on-wafer parasitics, which are due to the test pads (OPEN dummy) and the lines from the test pads to the transistor itself (SHORT dummy). This is done by subtracting the Y matrix of the OPEN from the total measurement, followed -if required- by the subtraction of the Z matrix of the SHORT. It should be mentioned that in this case the SHORT itself has to be de-embedded first from the OPEN parasitics!

For packaged devices, we need to use a test fixture. In this case, the NWA has to be calibrated down to the ends of its cables using the calibration standards (SOLT) of the actual connector type. As a next step, the test fixture has to be modeled (OPEN, SHORT, THRU). Finally, the DUT (device under test) is inserted into the test fixture and measured. The now known test fixture parasitics can be de-embedded and the extraction techniques of this manual can be applied to the down-stripped inner device. A file including such a procedure is included in the toolkit filesets. See the example more_files/packaged_xtor_in_testfixture.mdl

Please contact the author if you wish more info or help on de-embedding.

Operating Modes of the Bipolar Transistor

There are four operating modes of a bipolar transistor as illustrated in figure 1. The saturation region, for example, the region vCE<0.3V in the DC output characteristics, is described by the ohmic resistors. The DC and AC extraction procedures that are proposed in this manual cover mainly the forward region. Since the model is symmetrical, the reverse parameters can be extracted following the same ideas, but applied to the reverse measurements.



(*) NOTE: in the saturation range, BE and BC layers are 'overcharged'.

Fig.1: operating modes of a bipolar npn transistor

The Gummel-Poon equivalent schematic

Fig.2b shows the large signal schematic of the Gummel-Poon model. It represents the physical transistor: a current-controlled output current sink, and two diode structures including their capacitors. This structure represents pretty much the physical situation of a bipolar transistor, see fig.2a.

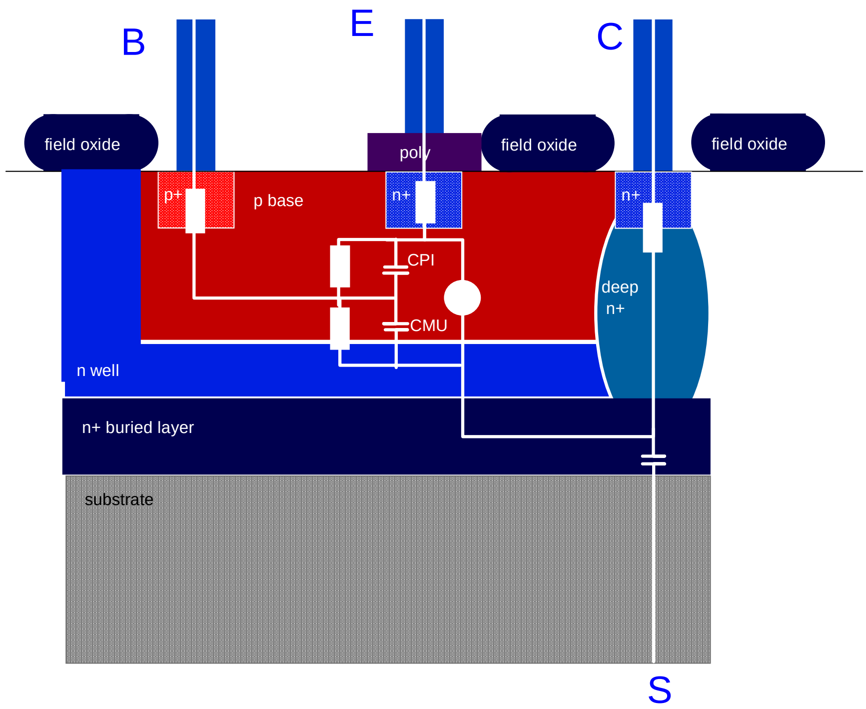


Fig.2a: physical situation for a bipolar transistor, neglecting the parasitic pnp transistor.

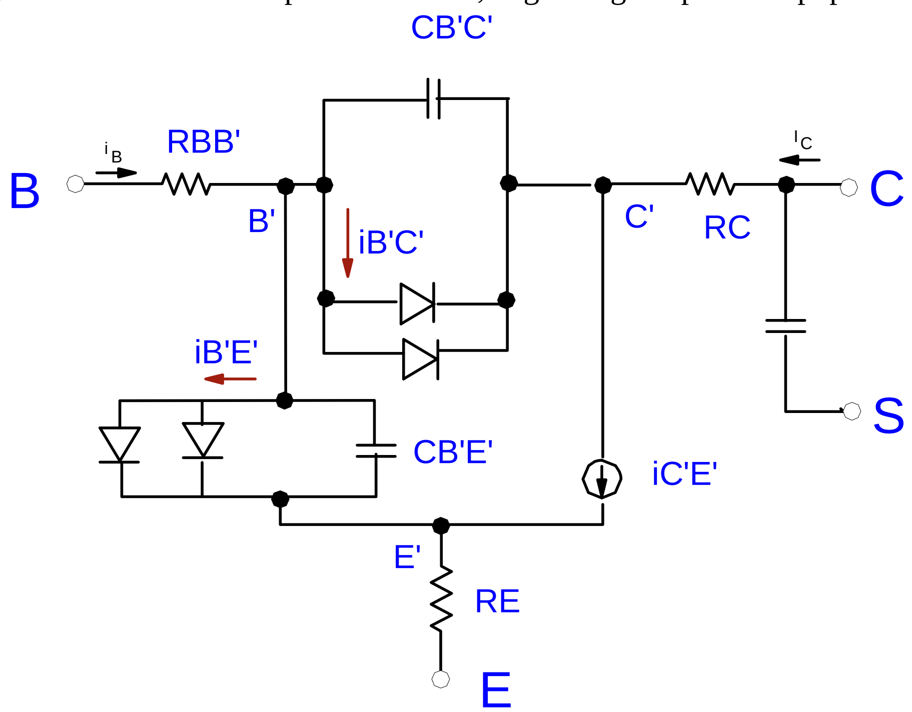


Fig.2b: Gummel-Poon large signal schematic of the bipolar transistor

From fig.2b, the small signal schematic for high frequency simulations can be derived. This means, for a given operating point, the DC currents are calculated and the model is linearized in this point (fig.2c). Such a schematic is used later for SPICE S-parameter simulations.
It must be noted that the schematic after fig.2c is a pure linear model. It cannot be used to predict non-linear high-frequency behavior of the transistor. In order to do this, RF simulators like HP_MDS or HP_ADS perform high-frequency simulations using the large signal model (harmonic balance simulations).

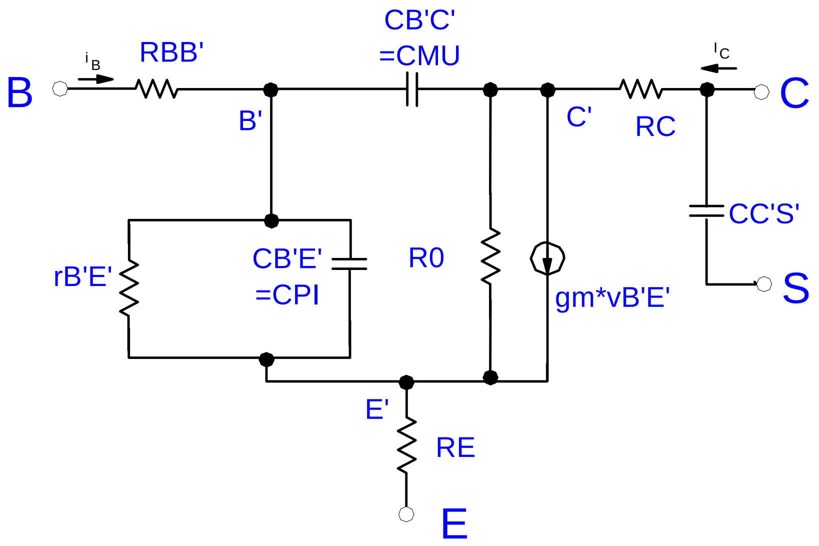


Fig.2c: AC small signal schematic of the bipolar transistor
NOTE: X_{CJC} effect neglected.

In order to make the presentations of the schematics complete, fig.2d depicts the subciruit used for modeling a npn transistor including the parasitic pnp. As said above, IC-CAP files for this type of modeling are included in the filesets of this toolkit. However, the description of this manual does not cover this. See the macros in the model files instead.

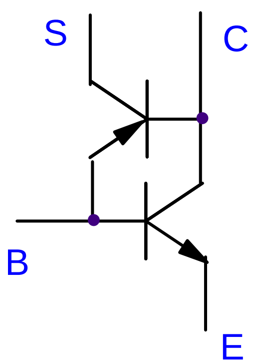


Fig.2d: subcircuit schematic when including the parasitic pnp.

The Gummel-Poon Model Equations

For the reader's convenience all the Gummel-Poon equations are presented at a glance.
In order to make them better understandable, we assume no voltage drops at RB, RE and RC, i.e. vB'E'=vBE and vB'C'=vBC.

TEMPERATURE VOLTAGE:

$$vt = \frac{k\,T}{q} = 8.6171\,E\text{-}5 * (T\,/{}'C + 273.15)$$

BASE CURRENT:

$$i_B = i_{BE} + i_{BC} \qquad (A)$$

$$i_B = i_f/BF + iBE_{rec} + i_r/BR + iBC_{rec} \qquad (B)$$

rec: recombination effect

with

ideal forward diffusion current: $i_f = IS\,\{ \exp[\frac{v_{BE}}{NF\,vt}] - 1 \}$ (C)

B-E recombination effect: $iBE_{rec} = ISE\,\{ \exp[\frac{v_{BE}}{NE\,vt}] - 1 \}$ (D)

ideal reverse diffusion current: $i_r = IS\,\{ \exp[\frac{v_{BC}}{NR\,vt}] - 1 \}$ (E)

B-C recombination effect: $iBC_{rec} = ISC\,\{ \exp[\frac{v_{BC}}{NC\,vt}] - 1 \}$ (F)

(see equiv. schematic in fig_2)

this gives:

$$iB = \frac{IS}{BF}\,\{ \exp[\frac{vBE}{NF\,vt}] - 1 \} + ISE\,\{ \exp[\frac{vBE}{NE\,vt}] - 1 \}$$

$$+ \frac{IS}{BR}\,\{ \exp[\frac{vBC}{NR\,vt}] - 1 \} + ISC\,\{ \exp[\frac{vBC}{NC\,vt}] - 1 \} \qquad (G)$$

COLLECTOR CURRENT:

$$i_C = 1/NqB\ (i_f - i_r) - i_r/BR - iBC_{rec}$$ (H)

(see definition of iB above and equiv.schematic in fig_2)

or:

$$i_c = \frac{IS}{Nqb} * \left[\left(\exp\frac{vBE}{NF * vt} - 1 \right) - \left(\exp\frac{vBC}{NR * vt} - 1 \right) ..... \right]$$

$$- \frac{IS}{Nqb} * \left[\exp\frac{vBC}{NR * vt} - 1 \right]$$

$$- ISC * \left[\exp\frac{vBC}{NC * vt} - 1 \right]$$ (I)

with the base charge equation

$$NqB = \frac{q_{1S}}{2} * \left(1 + \sqrt{1 + 4q_{2S}}\right)$$ (J)

for the modeling of non-idealities like the base-width modulation:

$$q1s = \frac{1}{1 - \frac{vBE}{VAR} - \frac{vBC}{VAF}}$$ (K)

and the hi-level injection effect:

$$q2s = \frac{IS}{IKF}\left[\exp\left(\frac{vBE}{NF * vt}\right) - 1\right] + \frac{IS}{IKR}\left[\exp\left(\frac{vBC}{NR * vt}\right) - 1\right]$$ (L)

BASE RESISTOR:

$$RBB = RBM + 3(RB - RBM)\frac{\tan(z) - z}{z * \tan^2(z)}$$ (M)

with

$$z = \frac{\sqrt{1 + \left(\frac{12}{PI}\right)^2 \frac{i_B}{I_{RB}}} - 1}{\frac{24}{PI^2}\sqrt{\frac{i_B}{I_{RB}}}}$$

PI = 3,14159 (N)

SPACE CHARGE AND DIFFUSION CAPACITORS:

$$CBC = CSBC + CDBC = \qquad (O)$$

$$= \frac{CJC}{[1 - vBC / VJC]^{MJC}} + \frac{TR}{NR\ vt} \frac{IS}{NqB} \exp\left[\frac{vBC}{NR\ vt}\right] \qquad (P)$$

and

$$CBE = CSBE + CDBE = \qquad (Q)$$

$$= \frac{CJE}{[1 - vBE / VJE]^{MJE}} + \frac{TFF}{NF\ vt} \frac{IS}{NqB} \exp\left[\frac{vBE}{NF\ vt}\right] \qquad (R)$$

with the transit time

$$TFF = TF \left\{ 1 + XTF \left[\frac{if}{if + ITF}\right]^2 \exp\left[\frac{vBC}{1{,}44\ VTF}\right] \right\} \qquad (S)$$

and the ideal forward base current i_f from the definition of i_B , i.e. equation (C).

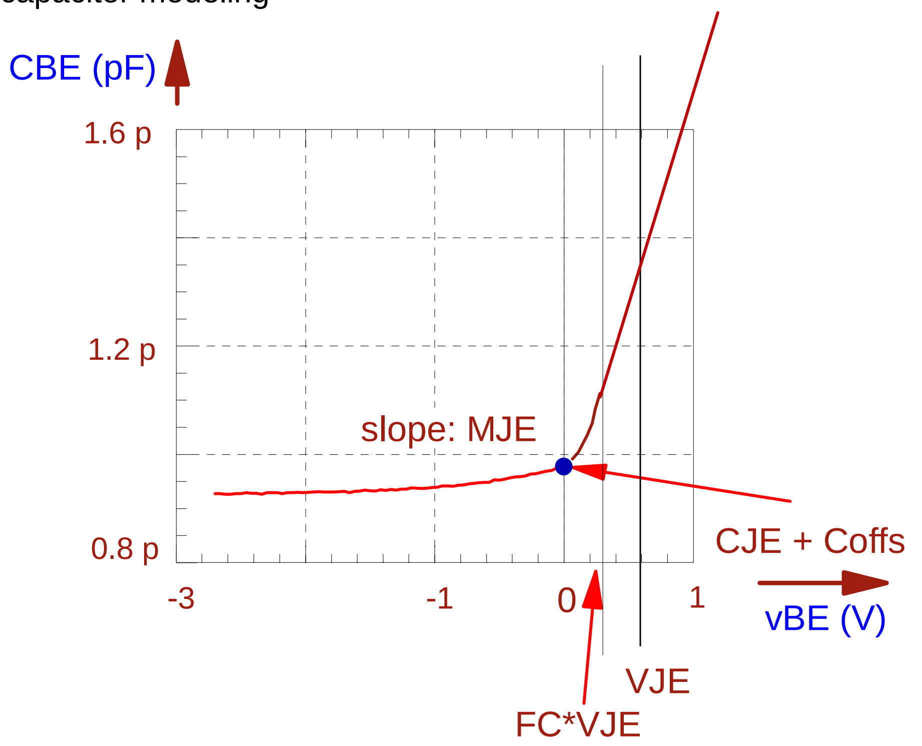
List of the SPICE Gummel-Poon Parameters

Name	Parameter explanation	SPICE default	typ.value	Unit
DC:				
IS	transport saturation current	.1E-15	1.E-15	A
XTI	temperature exponent for effect on IS	3	3	
EG	energy gap for temperature effect on IS	1.11	1.11	eV
BF	ideal forward maximum beta	100	150	
BR	ideal reverse maximum beta	1	.5	
XTB	forward & reverse beta temp.coeff.	0	2.5	
VAF	forward Early voltage	infinite	100	V
VAR	reverse Early voltage	infinite	50	V
NF	forward current emission coeff.	1	1.0	
NR	reverse current emission coeff.	1	1.0	
NE	B-E leakage emission coeff.	1.5	1.7	
NC	B-C leakage emission coeff.	2	1.3	
ISE	B-E leakage saturation current	0	.1E-12	A
ISC	B-C leakage saturation current	0	1.E-13	A
IKF	forward beta hi current roll-off	infinite	.05	A
IKR	reverse beta hi current roll-off	infinite	.3	A
OHMIC PARASITICS:				
RB	zero bias base resistance	0	100	Ohm
IRB	current at medium base resistance	infinite	.0001	A
RBM	min.base resistance at hi current	RB	25	Ohm
RE	emitter resistance	0	5	Ohm
RC	collector resistance	0	10	Ohm
CBE:				
CJE	B-E zero-bias deplet.capacitance	0	1.E-12	F
VJE	B-E built-in potential	.75	.6	V
MJE	B-E junction exponential factor	.33	.4	
CBC:				
CJC	B-C zero-bias deplet.capacitance	0	.5E-12	F
VJC	B-C built-in potential	.75	.6	V
MJC	B-C junction exponential factor	.33	.4	
XCJC	fraction of B-C capacitor connected to int.base	1	1	
CCS:				
CJS	zero-bias collector-substrate capacacitance	0	0	F
VJS	substrate junction built-in potential	.75	0	V
MJS	substrate junction exponential factor	0	0	
CAPACITOR FORWARD CHARACTERISTICS:				
FC	forward bias depletion cap.coeff.	.5	.5	
TRANSIT TIME:				
TF	ideal forward transit time	0	1.E-12	sec
XTF	coeff.for bias dependence of TF	0	10	
VTF	voltage describing VBC dependence of TF	infinite	5	V
ITF	hi-current parameter for effect on TF	0	20.E-3	A
PTF	excess phase at frequency 1/(TF*2PI)	0	0	deg
TR	ideal reverse transit time	0	50.E-12	sec
NOISE:				
KF	flicker noise coeff.	0		
AF	flicker noise exponent	1		
TEMPERATURE EFFECTS				
.TEMP	device temperature for simulation /'C	27	27	'C
.OPTIONS TNOM	device meas. and param. extraction temp	27	27	'C

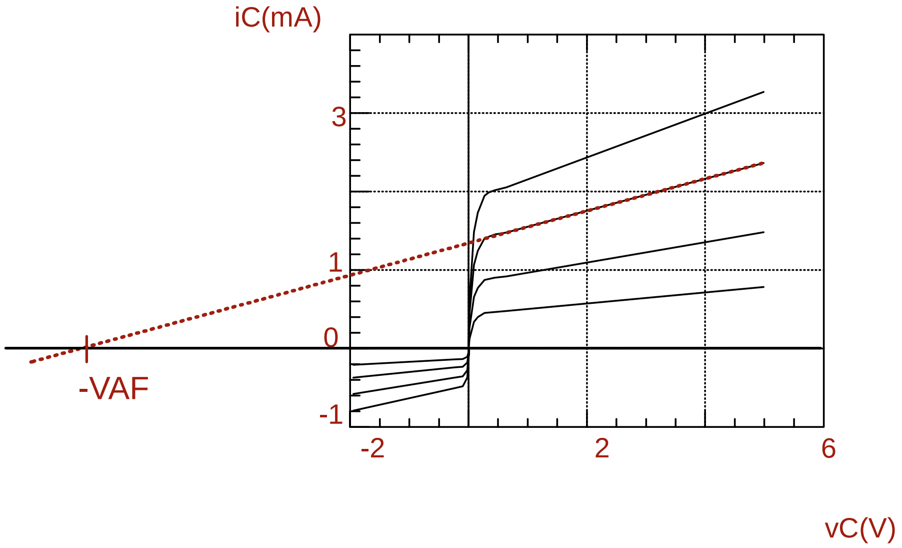
A quick Tutorial on the Gummel-Poon Parameters

Although it is recommended to go through the individual extraction steps of the corresponding sections of this manual, this chapter puts together the graphical equivalents of the parameter extraction techniques.

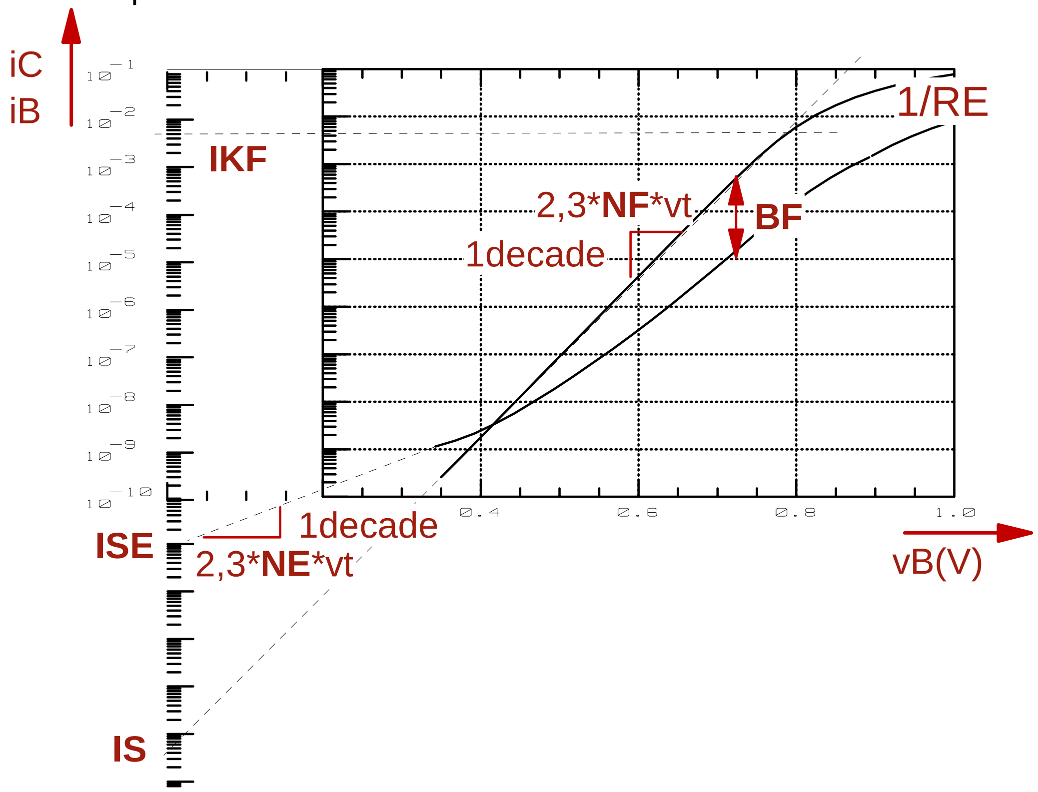
space charge capacitor modeling

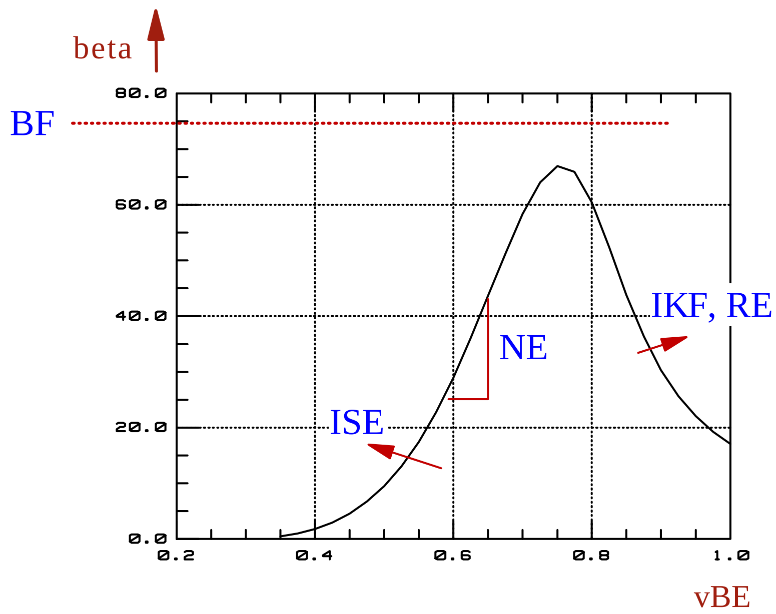


Early voltage extraction

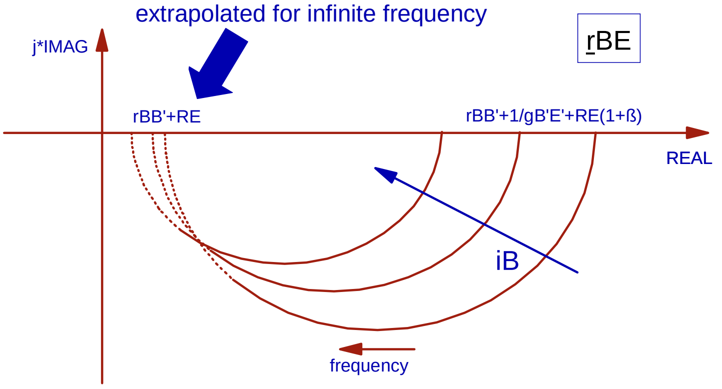


forward beta parameter extraction

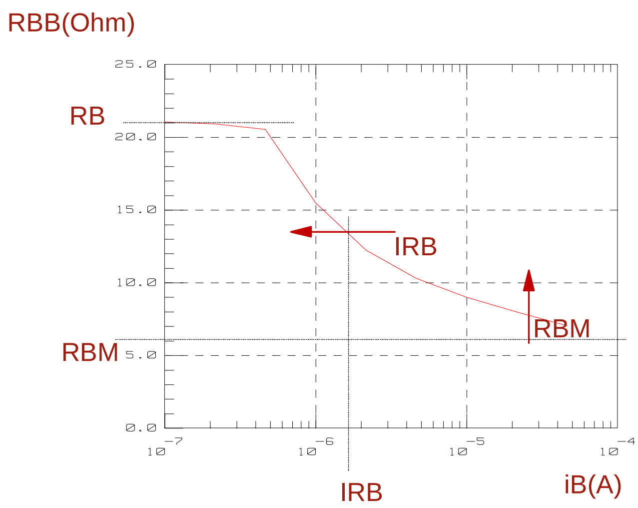




Base resistor parameter extraction

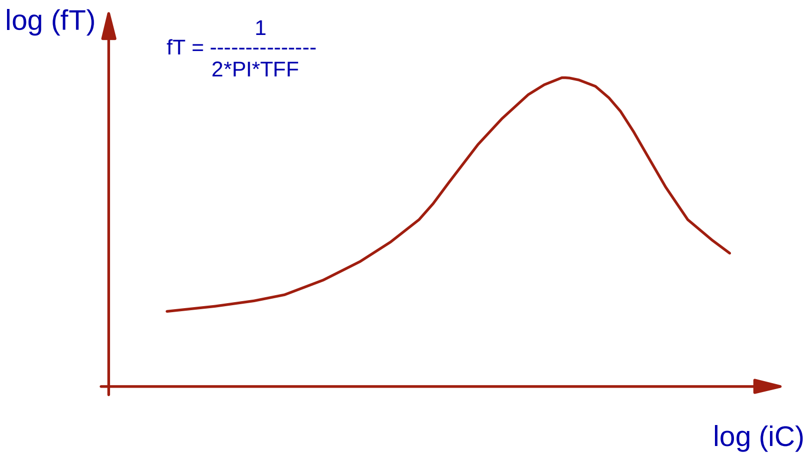


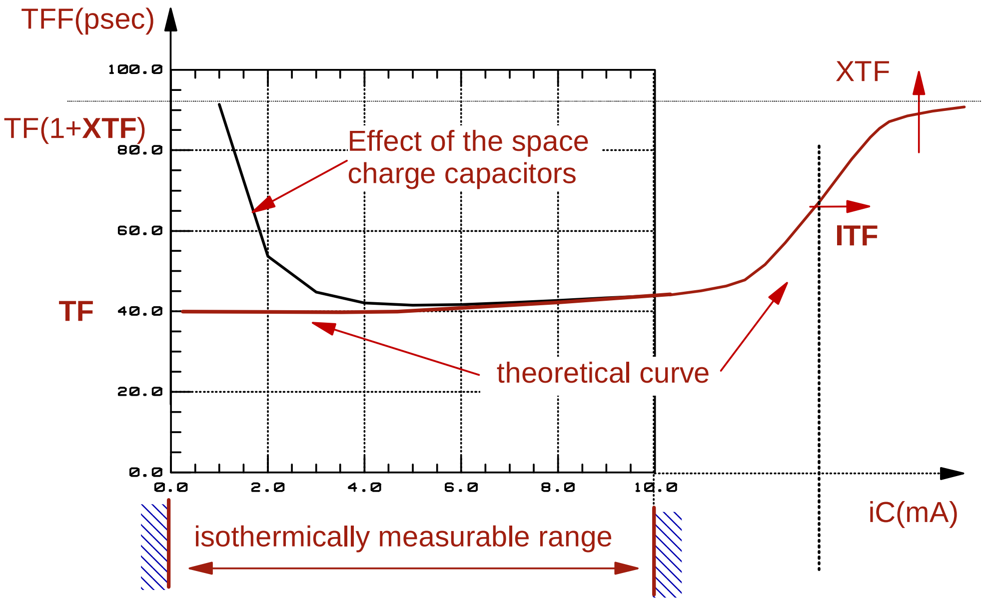
this gives:



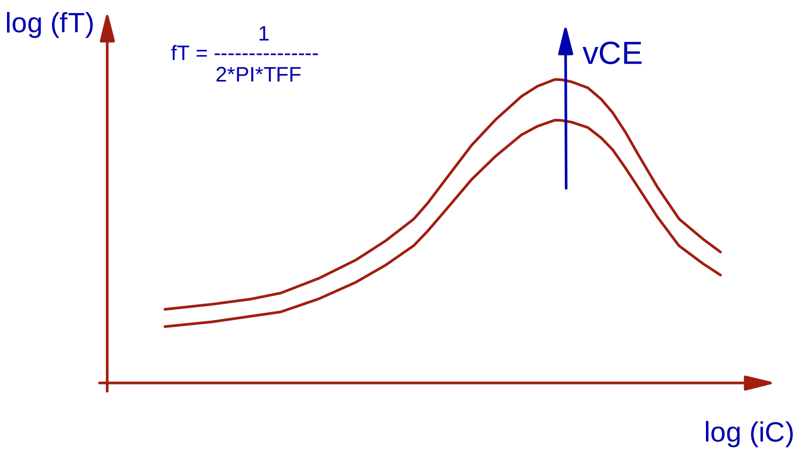
Transit time parameter determination

First model the TFF trace without V_{CE} effect





then model the dependence on V_{CE}:



Proposed Global Extraction And Optimization Strategy

First reset the model parameters to default (Window Model Parameters). This will firstly get rid of old parameter values which belong to the last modeling and not to our actual, current one, and secondly, the default parameters are those which reduce the complexity of a model completely. For example, VAF=1000 means: no Early effect, IKF=1000 no knee current, all resistors Rx=1m means no ohmic effects and so on. During the extraction process, we get more and more parameter values, and thus, the model becomes more and more complex and accurate.

cv:

Extract the CV parameters CJx, MJx and VJx for the BE- and BC-capacitance,
optional also for CCS.
optimize the CV parameters.

ohmic parasitics:

Extract the parasitic resistors RE, RBM and RC from flyback-measurements.
or:
Extract them from overdriven S-parameter measurements: high current at Base, half the current out of the Collector, Emitter grounded, frequency swept.

dc:

Extract VAR and VAF from the output characteristics.
Extract IS, NF, ISE, NE and BF from the forward Gummel-Poon plots.
Optimize the Gummel-Poon plot for IS, NF, ISE and NE, well below ohmic effects show up.
Extract IKF from the ß-curve.
Optimize RE in the upper region of the Gummel-Poon plots (i_B and i_C).
Optimize BF and IKF in the ß-curve at high bias.
Fine-optimize VAR, BR, VAF and BF in the output characteristics setup.
Fine-tune all DC parameters in all DC setups.

S-parameters:

De-embed the measurement data.
Extract the base resistor parameters RB, IRB and RBM from S11 measurements
with swept frequency and base current as a secondary sweep.
Transform S- to H-parameters and get a frequency f_{-20dB} from the -20dB/decade of h_{21}.
Measure again S-parameters, but now with the constant frequency f_{-20dB} and swept i_B and
swept v_{CE} and extract TF, XTF and ITF, as well as VTF.
Optimize S-parameter fitting of TF, XTF, ITF (lowest v_{CE}).
Optimize the S-parameter fitting of VTF (all v_{CE}).
Go back to the rBB' setup and optimize the S-parameter fitting of the RB, IRB and RBM.
Then, again in the rBB' setup, optimize the TF, XTF, ITF and VTF parameters.

Finally:

Re-simulate all setups and check the fitting quality in the verify setups.
If required, perform optimizer fine-tuning.

Macro 'extract_n_opt_ALL' in IC-CAP file gp_classic_npn.mdl contains an example for such a modeling strategy based on the measurement data included in the file.
This strategy may vary a bit depending on the actual data. In this case, simply modify the macro to meet your local requirements.

NOTE:

A smart way of defining or verifying the most appropriate extraction strategy is to synthesize quasi-measured data from simulation results, and to check the extraction routines on these data. This means to simulate all setups using a given parameter set, to transform these simulated data into measured ones and to try to get the (known) parameters back again.

In this way you are sure that your extraction strategy works well for a perfect Gummel-Poon transistor. If you have afterwards problems during the curve fitting, you might consider that your physical device under test may not be so well represented by the Gummel-Poon model!

To 'synthesize' such pseudo-measured data in IC-CAP, make sure the parameter values in the IC-CAP parameter list are all set to typical values that you will expect later for your parameter extraction, perform a simulation for every setup in your model file, change the setup output data type to 'S', hit <RETURN>, change it back to 'B' again and hit again <RETURN>. Now you have identical data in both measured and simulated arrays. Then reset the parameter values to default and try your extraction strategy.

See also the appendix.

<u>LAST NOT LEAST:</u>

Before performing your measurements, i.e. before defining the measurement ranges, contact tour design engineer colleagues and ask them about the specific operating range.

As a general rule, modeling should be done in those regions where the transistor will be operated later.

C V M O D E L I N G, P A R A M E T E R E X T R A C T I O N

C O N T E N T S:

Since the CV parameters are --like for most bipolar models-- independent of the other model parameters, they are usually extracted first.

We follow this idea and begin with the CV modeling, followed then by the parasitic resistor modeling and the non-linear DC curves. Finally, the S-parameter measurements are modeled.

The Gummel-Poon Capacitor Equations

Provided that: $v_{BE} = v_{B'E'}$ and $v_{BC} = v_{B'C'}$, the capacitors in the Gummel-Poon model given in the introduction chapter with equations (O) ... (S) are:

$$C_{BC} = C_{SBC} + C_{DBC}$$

$$= \frac{C_{jC}}{[1 - v_{BC} / V_{JC}]^{M_{JC}}} + T_R \frac{di_C}{dv_{BC}}$$

$$\stackrel{(H)}{=} \frac{C_{jC}}{[1 - v_{BC} / V_{JC}]^{M_{JC}}} + \frac{T_R}{N_R V_T} \frac{I_S}{N_{qB}} \exp\left(\frac{v_{BC}}{N_R V_T} \right) \qquad \text{(CV-1)}$$

with equation (H) from the introduction chapter.

and

$$C_{BE} = C_{SBE} + C_{DBE}$$

$$= \frac{C_{jE}}{[1 - v_{BE} / V_{JE}]^{M_{JE}}} + T_{FF} \frac{di_C}{dv_{BE}}$$

$$\stackrel{(H)}{=} \frac{C_{jE}}{[1 - v_{BE} / V_{JE}]^{M_{JE}}} + \frac{T_{FF}}{N_F V_T} \frac{I_S}{N_{qB}} \exp\left(\frac{v_{BE}}{N_F V_T} \right) \qquad \text{(CV-2)}$$

again with equation (H1) from the introduction chapter and additionally with

$$T_{FF} = T_F \left\{ 1 + X_{TF} \left[\frac{i_f}{i_f + I_{TF}} \right]^2 \exp\left[\frac{v_{BC}}{1,44\, V_{TF}} \right] \right\} \qquad \text{(CV-3)}$$

and the ideal Collector current i_f from (C)

C_{SBi} models the <u>space charge</u> and C_{DBi} the <u>diffusion capacitance</u> between Base and Emitter or base and Collector respectively.

v_{BE} and v_{BC} are the stimulating voltages.

MODELING THE SPACE CHARGE CAPACITORS:

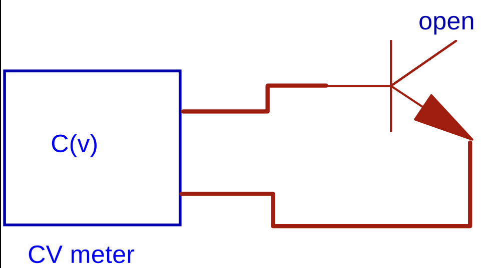
Extraction of CJE, VJE, MJE,
(CJC, VJC, MJC and CJS, VJS, MJS is the same)

C_{JE} B-E zero-bias deplet.capacitance
V_{JE} B-E built-in potential
M_{JE} B-E junction exponential factor

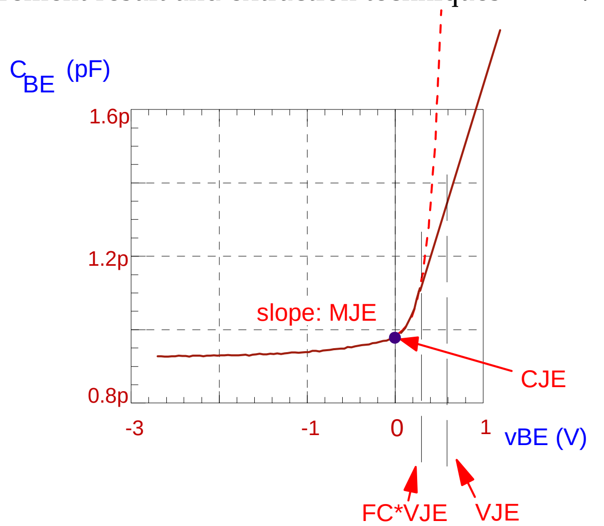
These parameters model the Base-Emitter and the Base-Collector space charge capacitance, i.e. the first term in (CV-1) and (CV-2). The second terms with the TFF and TR parameters will be modeled later by S-parameter mesurements.

For the measurement of the Base-Emitter capacitance, the Collector is left open while the Emitter is open during the measurement of the Base-Collector capacitance. In both cases, the modeling formula is the same. Therefore this chapter covers only the modeling of the Base-Emitter capacitance.

Measurement setup:

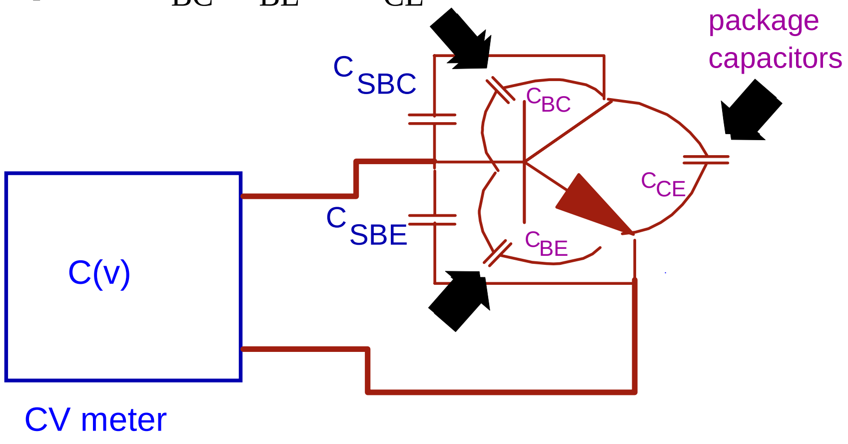


Measurement result and extraction techniques .



Mesuring and modeling the Base-Emitter capacitance

NOTE on the influence of the remaining capacitances of the open pin: as one of the transistor pins is left open, the measurements of C_{SBC} and C_{SBE} are always an overlay of the other capacitances
C_{SBi} (i = E,C) and, when measuring packaged devices, the small parasitic package capacitors C_{BC} , C_{BE} and C_{CE} .



The total measured capacitance is therefore C_{SBi} in parallel with the parasitic ones. This means that the measurement results are always too big.

When using a capacitance meter like the Agilent4284, that eliminates by its measurement principle parasitic capacitances to ground, this effect can be avoided by applying an AC short to the open transistor pin versus ground (big capacitor).

The equation:

The behavior of the space charge capacitor is given by equations (CV-4a) and (CV-4b):

for $v_{BE} < F_C * V_{JE}$:

$$C_{SBE} = \frac{C_{JE}}{\left(1 - \frac{V_{BE}}{V_{JE}}\right)^{M_{JE}}} \qquad \text{(CV-4a)}$$

and else:

$$C_{SBE} = \frac{C_{JE}}{(1 - F_C)^{(1+M_{JE})}} * \left[1 - F_C * (1 + M_{JE}) + M_{JE} * \frac{V_{BE}}{V_{JE}}\right] \qquad \text{(CV-4b)}$$

with

C_{JE} : space charge capacitance at v_{BE} = 0V

V_{JE} : built-in potential or pole voltage (typ. 0,7V)

M_{JE} : junction exponential factor, determines the slope of the cv plot
(abrupt pn junction (<0,5um): M_{JE} = 1/2)
(linear pn junction (> 5um): M_{JE} = 1/3)

F_C : forward capacitance switching coefficient, default 0,5

Determination of the CV parameters:

For simplicity, we only use the measurement data from the negative bias.
The logarithmic conversion of (CV-4a) yields:

$$\ln(C_{SBE}) = \ln(C_{JE}) - M_{JE} \ln[1 - v_{BE} / V_{JE}] \qquad \text{(CV-5)}$$

This equation can be interpreted as a linear function according to the ideas of linear regression analysis:

$$y = b + m \quad x$$

with

$$y = \ln(C_{SBE}) \qquad \text{(CV-6a)}$$

$$b = \ln(C_{JE}) \qquad \text{(CV-6b)}$$

and

$$m = - MJ_E \qquad \text{(CV-6c)}$$

$$x = \ln[1 - v_{BE} / V_{JE}] \qquad \text{(CV-6d)}$$

Linear regression means to fit a line to given measurement points. Therfore, the three main equations of a linear regression are b=f(xi,yi) and m=f(xi,yi), together with a fitting quality factor r^2=f(xi,yi,m,b). For a good fit, r^2~0.9...0.9999. See also the appendix.

How to proceed:

the measured values of C_{SBC} are logarithmically converted according to (CV-6a). Following (CV-6d), the stimuli data of the forcing voltage v_{BE} are nonlinearily converted too. This is done using a starting value for the unknown parameter V_{JE} (e.g. 0,2V). These two arrays are now introduced into the regression equations (see appendix) as corresponding y_i- resp. x_i-values. A linear curve is fitted to this transformed 'cloud' of stimulating and measured data. Thus we get the y-intersect $b(V_{JE})$ and the slope $m(V_{JE})$ for the actual value of V_{JE}. In the next step, this procedure is repeated with an incremented V_{JE}, and we get another pair of $m(V_{JE})$ and $b(V_{JEC})$. But now the regression coefficient r^2 will be different from the earlier one. I.e. depending on the actual value of V_{JE}, the regression line fits better or worse the transformed data 'cloud'. Once the best regression coefficient is found, the iteration loop is exited and we finally get V_{JE_opt} as well as the corresponding $b(V_{JE\_opt})$ and $m(V_{JE\_opt})$.

Thus we get from (CV-6c):

$$M_{JE} = - m(V_{JE\_opt})$$

and from (CV-6b):

$$C_{JE} = \exp[\, b(V_{JE\_opt}) \,]$$

Validity of this extraction: The parameter extraction for the space charge capacitor is valid only for stimulus voltages v_{BE} below $F_C * V_{JE}$, $F_{C\_default}$ = 0,5.

WHAT TO DO IN IC-CAP:

Since this is our first parameter extraction step, we first reset all parameter values to default, see IC-CAP Window: 'Model Parameters'
Otherwise, we might end up with a mix of parameter values obtained during our last transistor modeling and today!

open setup "/gp_classic_npn/cv/cbe_bhi"
(means modeling of C_{BE}, with Base contact at high voltage pin),
perform a measurement,
click a box into plot "cvsv" (capacitance vs. voltage) to select the measurement data used later for extractions. Click 'Copy to Variables' under 'Options' in that plot. This will cause IC-CAP to save the box corners in the 'cbe_bhi' Setup Variables X_LOW, X_HIGH, Y_LOW, Y_HIGH
perform transform "br_CJE_VJE_MJE" (box regression). This transform applies a data transformation and regression analysis to the data inside the box.
Then simulate with the extracted parameter values, using simulation or the substitute transform calc_cv.

do the same for the capacitor CBC in setup 'cbc_bhi'.

NOTE: try also macro 'extract_n_opt_CV'

Some comments on CV-modeling

In practice there is always an overlay of this capacitance with some parasitic ones, e.g. package or pad capacitances. If they are not known and therefore cannot be de-embedded (calculated out of the measured data), the extracted CV parameter values may have no physical meaning. This may happen especially to V_{JC} and MJ_C.

If there are resolution problems with fF-capacitances and CV meters, a network analyzer can be used instead of the CV meter as well. In this case, the Base is biased and Emitter and Collector are grounded. The measured S-parameters are deembedded, converted to Y parameters and the CV traces can be calculated out of their imaginary parts.
See IC-CAP file: 1_gummel_poon/more_files/s_to_cv.mdl for more details.

MODELING THE RESISTORS

CONTENTS:

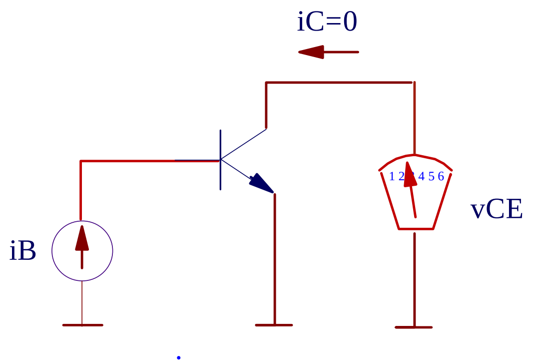
The methods given below are considered as standard extractions. But the parameter values are pretty often merely a 'first guess'. Also, the other model parameters are still not yet knwon. Therefore, no simulation or optimization is performed in the setups of DUT prdc in file gp_classic_npn.mdl

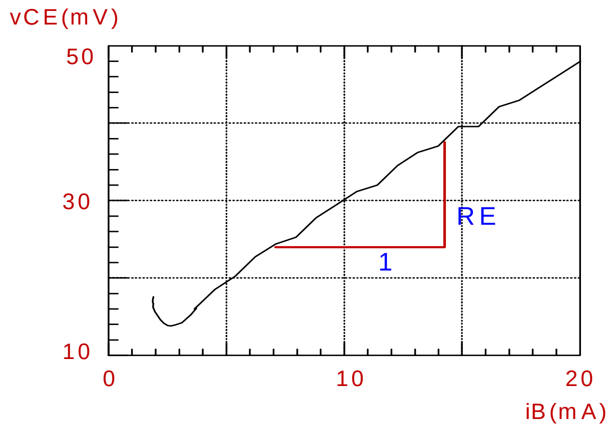
Instead, these parasitic resistor parameters are finetuned in the setups dc/fgummel and dc/rgummel. In the Gummel plots, they are tuned in order to fit the ohmic regions: RE in the forward Gummel plot (i_C and i_B vs v_{BE}) and RC in the reverse plot (i_E and i_B vs v_{BC}).

MODELING THE EMITTER RESISTOR

Extraction of RE

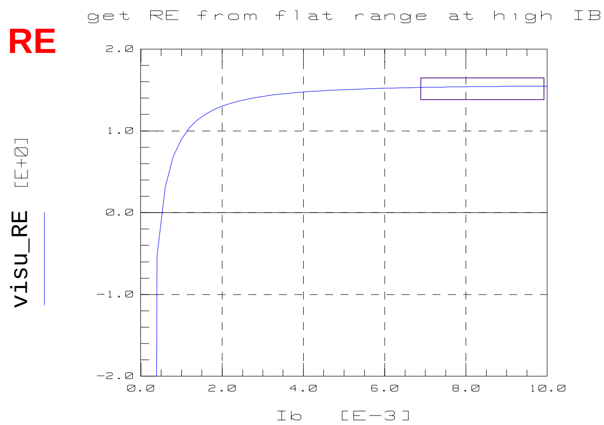
Measurement setup: Measurement result:





$$R_E = \frac{\partial v_{CE}}{\partial i_B}$$

transformed measured data:



Measurement of the open Collector voltage ('flyback method')
and the transformed measurement data in the RE domain (delta(vCE) / delta(iB))

<u>Extracting the parameters:</u>
The ohmic emitter resistor is physically located between the internal Emitter E' and the external Emitter pin E. When we apply a Base current and have the Emitter pin grounded, we get a voltage at the open Collector that is proportional to the Base current through this Emitter resistor. If we derivate vCE with respect to iB, we get the equivalent RE for each operating point. The value of RE is then the mean value of the flat range in this plot.

WHAT TO DO IN IC-CAP:

- measure the setup rb_re
- run transform visu_RE and enter '1' (data transform)
  this will derivate the measured data and display the
  calculated effective RE against the stimulus iB.

- click a box around the most constant range of measured data and click ‘Copy to Variables’
- re-execute transform visu_RE to extract the RE value (enter '1' for this operation mode).

Do not simulate or optimize this setup, since
- the other DC model parameters are not known yet
- the Gummel-Poon model cannot represent ‘unconventional’ measurement conditions like the actual flyback method. The values of the ohmic parasitics will be fine-tuned later in the Gummel-Poon plots

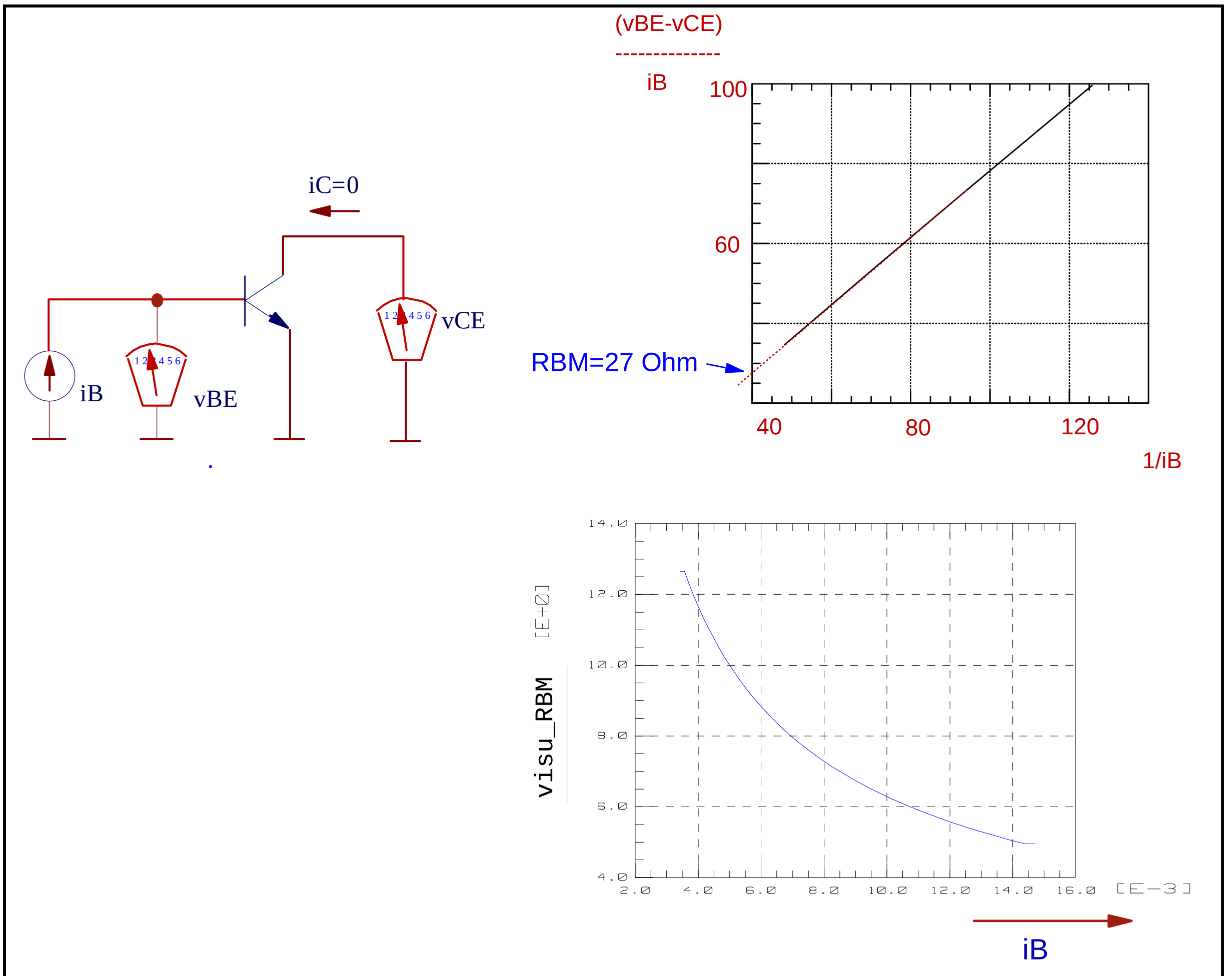
Extraction of RBM from DC measurements

RBM min.Base resistance at high current

There are several methods to determine the Base resistor: either the constant part of it (RBM) from pure DC measurements, or the non-linear RBB' = f(RB,IRB,RBM) from a s11 plot or from noise figure measurements.

Applying these three methods to the same transistor 'will generate typically three different values' for the Base resistor (!).

An interesting method to determine RBM is to use the RE-flyback method, with additionally measuring vBE /T.Zimmer/. This method is applied now.



Measurement setup and determination of RBM out of transformed measured data.

The theoretical values of the measured voltages are:

```
vCE  =  VT * ln(1/AI)  + iB * RE
```

AI: reverse current amplification in common Base

and

```
vBE  =  iB * RE  +  iB * RBM  +  vB'E'
```

Subtracting these equations and dividing by iB yields:

```
vBE  -  vCE       const
-----------   =   -----  +  RBM
     iB            iB
```

i.e. a regression analysis applied to these transformed measured data will give the y-intersect RBM.

In a final step, we then apply a loop to these data, in which a line is fitted to two adjacent points, and the y-intersect is calculated. The incremental y-intersect is then displayed against the stimulus iB.

NOTE: when RB becomes measurable DC-wise (the 'ohmic' range in the Gummel-Poon plot), its value is typically already lowered to the value of RBM. This means, parameter RB (the higher Base resistor value for lower Base bias), cannot be determined by this method. Therefore, we simply set RB=RBM.

NOTE: See also the appendix chapter 'direct visual parameter extraction'

WHAT TO DO IN IC-CAP:

- the measurement of setup rb_re is re-used
- run transform visu_RBM and enter '1' (data transform)
  this will calculate the local Base resistor for each bias point, as described above,
  and display the RBM value against the stimulus iB.

- click a box around the most constant range of measured data and click ‘Copy to Variables’
- re-execute transform visu_RBM to extract the RBM value

Again, do not simulate or optimize this setup, since the other DC model parameters are not known yet

NOTE: If a sensitivity analysis for a Gummel-Plot shows a reasonable impact of the Base resistor to the forward and reverse Base current, an optimizer run on these two curves simultaneously might make sense to obtain a guess on the actual value of RBM. However, this is usually not the case.

MODELING THE COLLECTOR RESISTOR

For the extraction of RC, the same flyback method like for RE is applied. The only difference is that the Collector pin is grounded, and the Emitter pin is left open and its voltage is measured.

WHAT TO DO IN IC-CAP:

- measure the setup rc
- run transform visu_RC and enter '1' (data transform)
  this will derivate the measured data and display the
  calculated effective RC against the stimulus iB.

- click a box around the most constant range of measured data and click ‘Copy to Variables’
- re-execute transform visu_RC to extract the RC value.

Again, do not simulate or optimize this setup

Note: Try also and study macro 'extract_resistors"

NOTES:

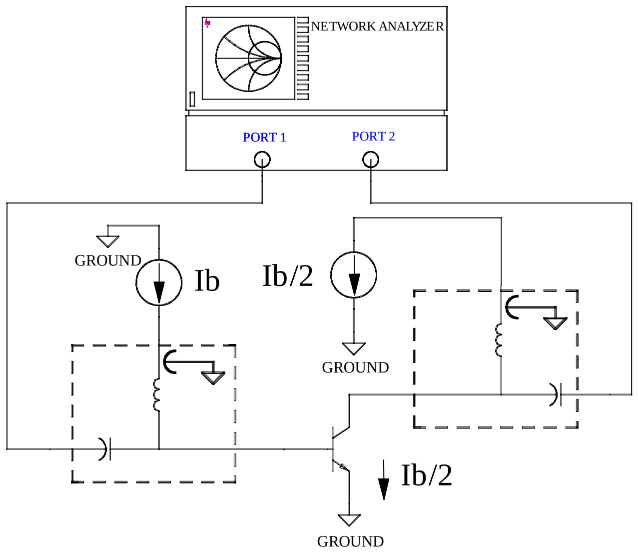
as mentioned above, these 'classical' extractions of the ohmic model parameters are used to get a good estimation about the parameter values. The values will be fine-tuned later in the setups fgummel and rgummel.

For details on alternate DC modeling methods of the parasitic resistors, see also the publications of /Berkner/ and /MacSweeny/.

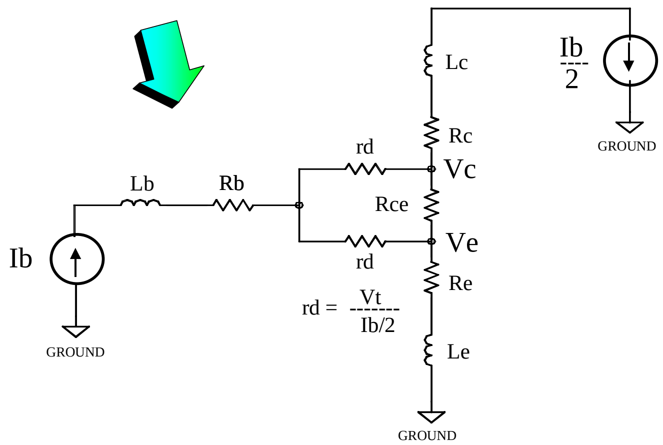
If there is a parasitic pnp transistor present, this method will not give accurate RC values. See the corresponding model file of this toolkit.

An alternate method to calculate the ohmic parasitic resistors from s-parameter measurements

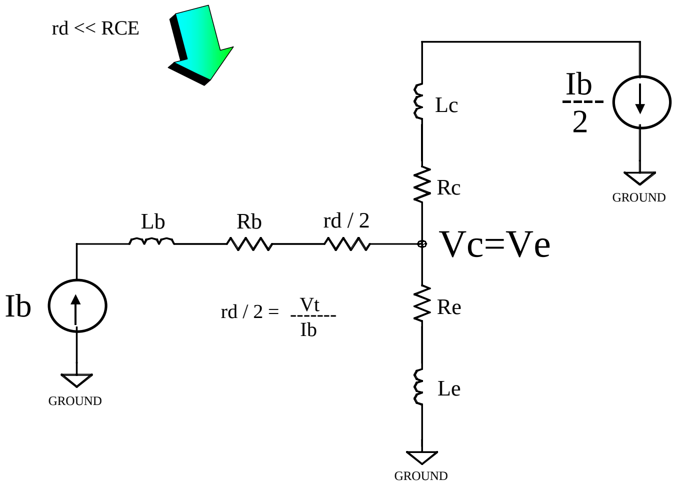
Since the fitting of the S-parameters is the goal of a good transistor modeling, it makes sense to think about extracting the ohmic parameters from S-parameter measurements also. The following figures sketch a reliable way to do that. The basic idea is to overdrive the transistor and to reduce its effect to simple diode characteristics ('*hot*' measurement). With the known value of the Base current, the remaining resistor values can be calculated easily.

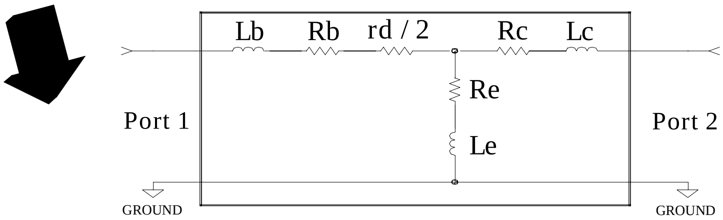


Bias for Parasitic Extraction



Bias for Parasitic Ext. continued



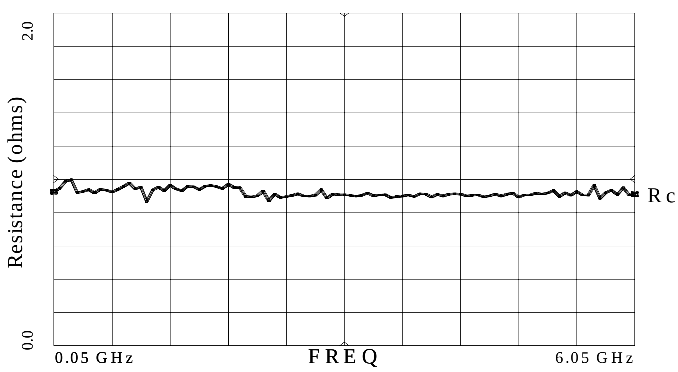


$$\begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} = \begin{bmatrix} rd/2+Rb+Re+jw(Le+Lb) & Re+jwLe \\ Re+jwLe & Rc+Re+jw(Le+Lc) \end{bmatrix}$$

$Rc = Real(Z_{22}) - Re$
$Re = Real(Z_{12})$
$Rb = Real(Z_{11}) - Re - rd / 2$ with $rd = VT / Ib$

$Lc = Imag(Z_{22})/w - Le$
$Le = Imag(Z_{12})/w$
$Lb = Imag(Z_{11})/w - Le$

The resistor values are finally displayed versus frequency and their values are obtained as a simple mean value. If there is a frequency drift, take the mean value from the lowest frequency. The plot below gives an example:



N O N L I N E A R   D C   M O D E L I N G

C O N T E N T S:

Three measurements are required in order to extract the DC parameters:
-> an output plot including both, forward and reverse operation,
-> and two so-called Gummel plots, one for forward and another for reverse mode.

These three plots have a certain context between each other. Neglecting this context can easily lead to one of the famous, so-called 'infinite modeling loops'.

This can be explained as follows:
Let's consider the forward Gummel plot. It is based on a measurement of i_B and i_C simultaneously, versus v_{BE} and is typically plotted half-logarithmically. Most often, the applied Collector-Base voltage is set to v_{BC} = 0V. The reason for this is that it simplifies the modeling equations (H)...(L) drastically. However, this approach can easily lead to the 'infinite loop' mentioned above!

IC-CAP does not need the simplified equations. The optimizer in IC-CAP always uses a true simulator like SPICE in the background that includes the complete Gummel-Poon equations. Therefore, while extracting the DC parameters or other parameters, there is no reason for having v_{BC} = 0 for the Gummel plot.

We can take a smarter approach. We first measure the forward output characteristic and extract VAR and VAF. Then, we leave this setup for the moment, and measure the forward Gummel plot. Differently from the commonly used method mentioned above, we apply a v_{CE} that is not zero, but between 2V and *half* the value of the maximum v_{CE} of the output plot. We do this for the following reason. Once the Gummel plot is fitted for this special voltage, the following output plot simulation already hits the measured curves exactly in the middle of the output characteristic. A final fine-tuning is then easily achieved by adjusting VAF and BF. Otherwise, if we use v_{BC} = 0 for the Gummel plot, it can easily happen that if the Gummel plot itself is nicely fitted, the output characteristics doesn't match and so on. Because, in this case, if the Gummel plot fits, this means that the output characteristic fits in the *saturation* range (v_{CE} ~ 0.2 ... 0.9V) and *not* int the disired linear range (v_{CE} ~ 0.5 to v_{CEmax}).

An illustration of this idea is presented below in fig.DC-1. First, the output characteristic is measured and VAR and VAF are extracted. Then, considering a cut through this plot for a fixed v_{CE} (4V in the example), and using *this* value of v_{CE} when measuring the Gummel-Poon plot, we have data points that refer directly to our previous output characteristics measurement with the corresponding v_{BE} . The relationship between i_C and i_B leads to the beta plot, also plotted against linear v_{BE} instead of the usual logarithmic i_C (which is the same for i_C below the ohmic effects in the Gummel-Poon plot) and again highlighting the corresponding output data points by buttons. Therefore, if beta fits, so does the output characteristic, which we were starting from.

Therefore, if we extract the DC forward parameters from a Gummel-Poon measurement that is biased like this, all measurements fit together.

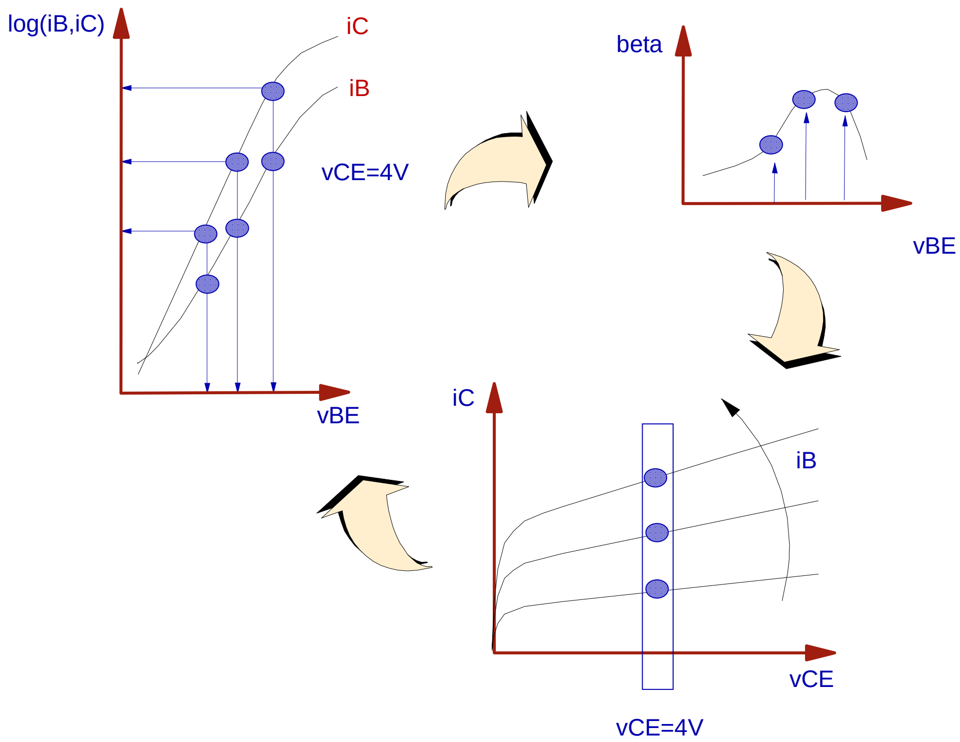


Fig.DC-1: Proposed context of the DC forward measurement setups.

For the output characteristic, by forcing i_B instead of v_{BE} , we also prevent from measuring thermal self-heating effects, which are not included in the standard Gummel-Poon model. However, we should also measure the same Collector currents values with a corresponding v_{BE} as well. This is sketched below in fig.DC-2..

Note:Such a check is implemented in file data_mgmt/BIP_MEAS_MASTER.mdl

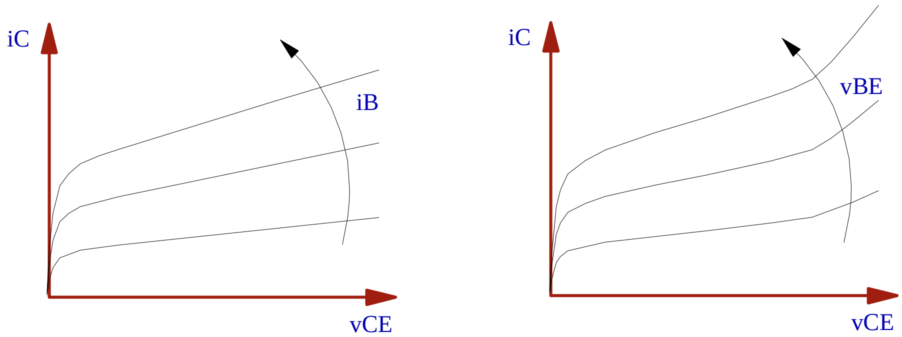


Fig.DC-2: Forcing iB rather than vBE for the output characteristics prevents from obtaining measurement curves including the self-heating effect.
Note: i_C is scaled identically for both plots!

Because, if the output characteristic drifts off when forcing v_{BE} , we should be careful when measuring the Gummel plot, because it could be affected by self-heating as well. The ohmic effects are in this case *overlaid* by the thermal self-heating, and we will either get wrong model parameters for RE and IKF, or no good fitting at all.

It is recommended in this case to apply a v_{CE} as low as possible for the Gummel plot (below the thermal runaway), but well above the saturation region of the output plot.

Note: See also IC-CAP file bip_output_char_i_or_v.mdl in the more_files directory.

MODELING THE OUTPUT CHARACTERISTIC

EXTRACTION OF VAR AND VAF

V_{AR} reverse Early voltage
V_{AF} forward Early voltage

Modern fast bipolar transistors exhibit small values of VAR. Due to the simplifications of the G-P space charge model implementation in SPICE, this may affect the other model parameters. This typically happens for VAR<5.

Therefore, the extraction of the nonlinear DC parameters is best started with the extraction of VAR, followed by VAF. As will be shown in this chapter, VAR/VAF can be determined with only little overlay of the other (actually still unknown) parameters.

After the Early voltages are extracted, and before optimizing the fit of this setup, we need to go ahead and extract the remaining DC forward model parameters from the Gummel plot. Only then, with the correct BF etc., the simulation of the output characteristic can fit the measured data (!).

Therefore, we come back to this setup and fine-tune the VAR/VAF values by optimization later.

Now, let's discuss the theoretical background of the Early voltage extraction. For an easier understanding, we consider VAF.

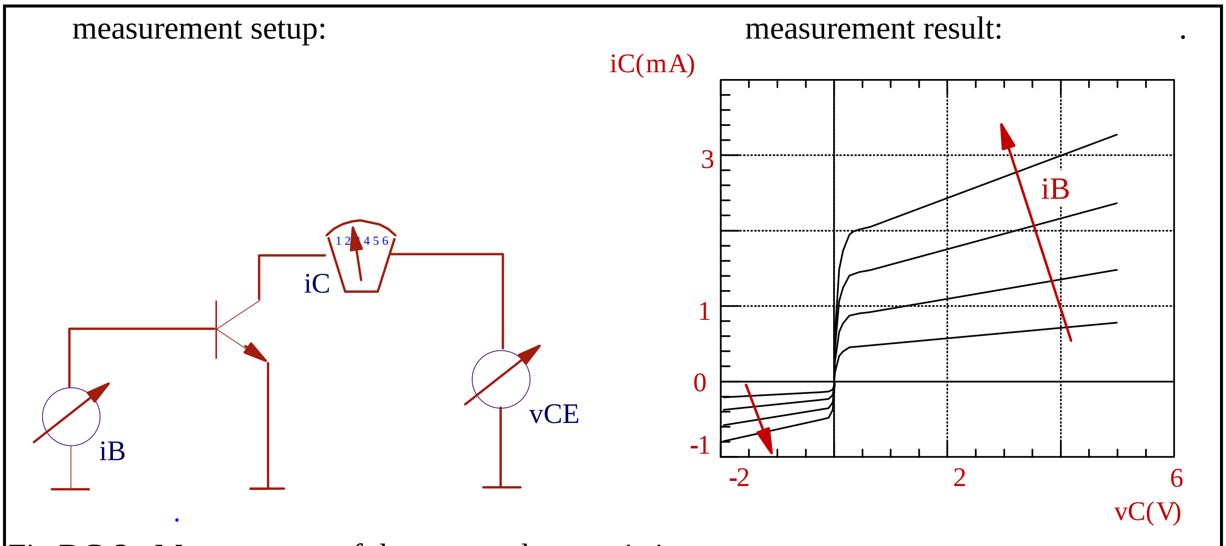


Fig.DC-3: Measurement of the output characteristic

The equation:

Provided that: $v_{B'E'}=v_{BE}$ and $v_{B'C'}=v_{BC}$, the Gummel-Poon model describes iC by

$$i_C = \frac{IS}{N_{qB}} \left\{ \left(\exp\left[\frac{v_{BE}}{NF\ vt} \right] - 1 \right) - \left(\exp\left[\frac{v_{BC}}{NR\ vt} \right] - 1 \right) \right\} - \frac{IS}{BR} \left\{ \exp\left[\frac{v_{BC}}{NR\ vt} \right] - 1 \right\} - ISC \left\{ \exp\left[\frac{v_{BC}}{NC\ vt} \right] - 1 \right\} \qquad \text{(DC-4)}$$

see equ. (H) ... (L) of the introduction chapter

with the Base charge equation

$$NqB = \frac{q_{1S}}{2} * \left(1 + \sqrt{1 + 4q_{2S}}\right) \qquad \text{(DC-5)}$$

for the modeling of non-idealities like the Base-width modulation:

$$q_{1S} = \frac{1}{1 - \frac{v_{BE}}{VAR} - \frac{v_{BC}}{VAF}} \qquad \text{(DC-3)}$$

and the hi-level injection effect:

$$q_{2S} = \frac{IS}{IKF} \left\{ \exp\left[\frac{v_{BE}}{NF\ vt} \right] - 1 \right\} + \frac{IS}{IKR} \left\{ \exp\left[\frac{v_{BC}}{NR\ vt} \right] - 1 \right\} \qquad \text{(DC-4)}$$

In order to handle this complex formula, we have to start with some simplifications:

We consider only the forward active region. Here, the Base Collector voltage is $v_{BC} < 0V$; therefore the terms

$$\left\{ \exp\left[\frac{v_{BC}}{N_i\ V_T} \right] - 1 \right\} \qquad \text{for } N_i = N_R \qquad \text{resp. } N_i = N_C$$

in equ.(DC-4) and (DC-4) may be neglected.

Thus (DC-4) becomes:

$$i_C = \frac{I_S}{N_{qB}} \exp\left[\frac{v_{BE}}{N_F\ V_T} \right] \qquad \text{(DC-5)}$$

and (DC-4) :

$$q_{2S} = \frac{I_S}{I_{KF}} \exp\left[\frac{v_{BE}}{N_F\ V_T} \right] \qquad \text{(DC-6)}$$

Equ.(DC-6) in (DC-5) yields:

$$NqB = \frac{1}{2\left(1 - \frac{v_{BE}}{V_{AR}} - \frac{V_{BC}}{V_{AF}}\right)} * \left(1 + \sqrt{1 + 4\frac{I_S}{I_{KF}}\exp\left(\frac{v_{BE}}{n_F V_T}\right)}\right) \qquad \text{(DC-7)}$$

(DC-7) in (DC-5) gives:

$$i_C = \frac{2 * I_S \exp\left(\frac{v_{BE}}{n_F V_T}\right)}{1 + \sqrt{1 + 4\frac{I_S}{I_{KF}}\exp\left(\frac{v_{BE}}{n_F V_T}\right)}} * \left(1 - \frac{v_{BE}}{V_{AR}} - \frac{v_{BC}}{V_{AF}}\right) \qquad \text{(DC-8)}$$

Fig.DC-3 showed i_C versus v_{CE}. Therefore, (DC-8) has to be re-arranged using

$$v_{BC} = v_{BE} - v_{CE} \qquad \text{(DC-9)}$$

(DC-9) in (DC-8):

$$i_C = \frac{2 * I_S \exp\left(\frac{v_{BE}}{n_F V_T}\right)}{1 + \sqrt{1 + 4\frac{I_S}{I_{KF}}\exp\left(\frac{v_{BE}}{n_F V_T}\right)}} * \left(1 - v_{BE}\left(\frac{1}{V_{AR}} + \frac{1}{V_{AF}}\right) + \frac{v_{CE}}{V_{AF}}\right) \qquad \text{(DC-10)}$$

with typically v_{BE} << VAR and v_{BE} << VAF we get:

$$i_C = \frac{2 * I_S \exp\left(\frac{v_{BE}}{n_F V_T}\right)}{1 + \sqrt{1 + 4\frac{I_S}{I_{KF}}\exp\left(\frac{v_{BE}}{n_F V_T}\right)}} * \left(1 + \frac{v_{CE}}{V_{AF}}\right)$$

or

$$i_C = \frac{2 * I_S \exp\left(\frac{v_{BE}}{n_F V_T}\right)}{1 + \sqrt{1 + 4\frac{I_S}{I_{KF}}\exp\left(\frac{v_{BE}}{n_F V_T}\right)}} * \frac{1}{V_{AF}}\left(V_{AF} + v_{CE}\right) \qquad \text{(DC-11)}$$

Thus we got $i_C = f(v_{CE}, i_B)$ as shown in fig.DC-3, with $v_{BE} = f(i_B)$.

Extracting the parameter:

We consider all assumptions from above valid(!!). This means that we should be sure that the output characteristics measurement has been taken in the linear range of the Gummel-Poon plot, i.e. with a maximum iC well below the ohmic effects. Then, the Collector current of the output characteristics measurement (equation DC-11), becomes zero for v_{CE} = -VAF.

How to proceed:

VAF is the x-axis intersect of the tangent fitted to the linear region of the output characteristics.

NOTE: As you will find out with your own measurements, VAF is rather a function of the bias current than a constant. The standard deviation of the values of VAF found by applying tangents to all slopes in the output plot is most often very big. Depending on the type of transistor, sigma(VAF) can range up to VAF/2 ! The reason is that the assumptions in equations (DC-5)..(DC-11) are pretty straight forward. Therefore an estimation of VAF by using only 1 tangent may be sufficient, when an optimizer run is performed later (after the extraction of the remaining DC forward parameters). Please note again that the IC_CAP optimizer calls the simulator which includes the full set of model equations and therefore finds the correct final value of VAF.

An alternate method could also be to determine VAF out of the delta of two Gummel plot curves $i_C(v_{BE})$ for two different Collector-Emitter bias voltages. See equation (DC-18) of the next chapter.

WHAT TO DO IN IC-CAP:

open setup "/gp_classic_npn/dc/routput",
perform a measurement,
perform transform "br_VAR" (extract VAR)
simulate with the extracted value of VAR.

Then,

open setup "/gp_classic_npn/dc/foutput",
perform a measurement,
perform transform "be_VAF" (extract VAF)
simulate with the extracted value of VAF.

Do not be confused about the simulation result, and that the curves do not match. Because all other DC parameters are still set to default, it is only important that the slopes of simulated and measured curves match!
We will have a much better fitting after the extraction of the other DC forward parameters.

Have also a look into "/gp_classic_npn/dc/routput/READ_ME".

MODELING THE COLLECTOR CURRENT:

EXTRACTION OF IS AND NF

I_S transport saturation current
N_F forward current emission coefficient

These 2 parameters, together with the already known Early voltages, are the only ones that are dominant in the measurement setup given in the figure DC-4 below. NF determines the slope and IS the y-intersect of the half-logarithmically plotted $i_C(v_{BE})$.

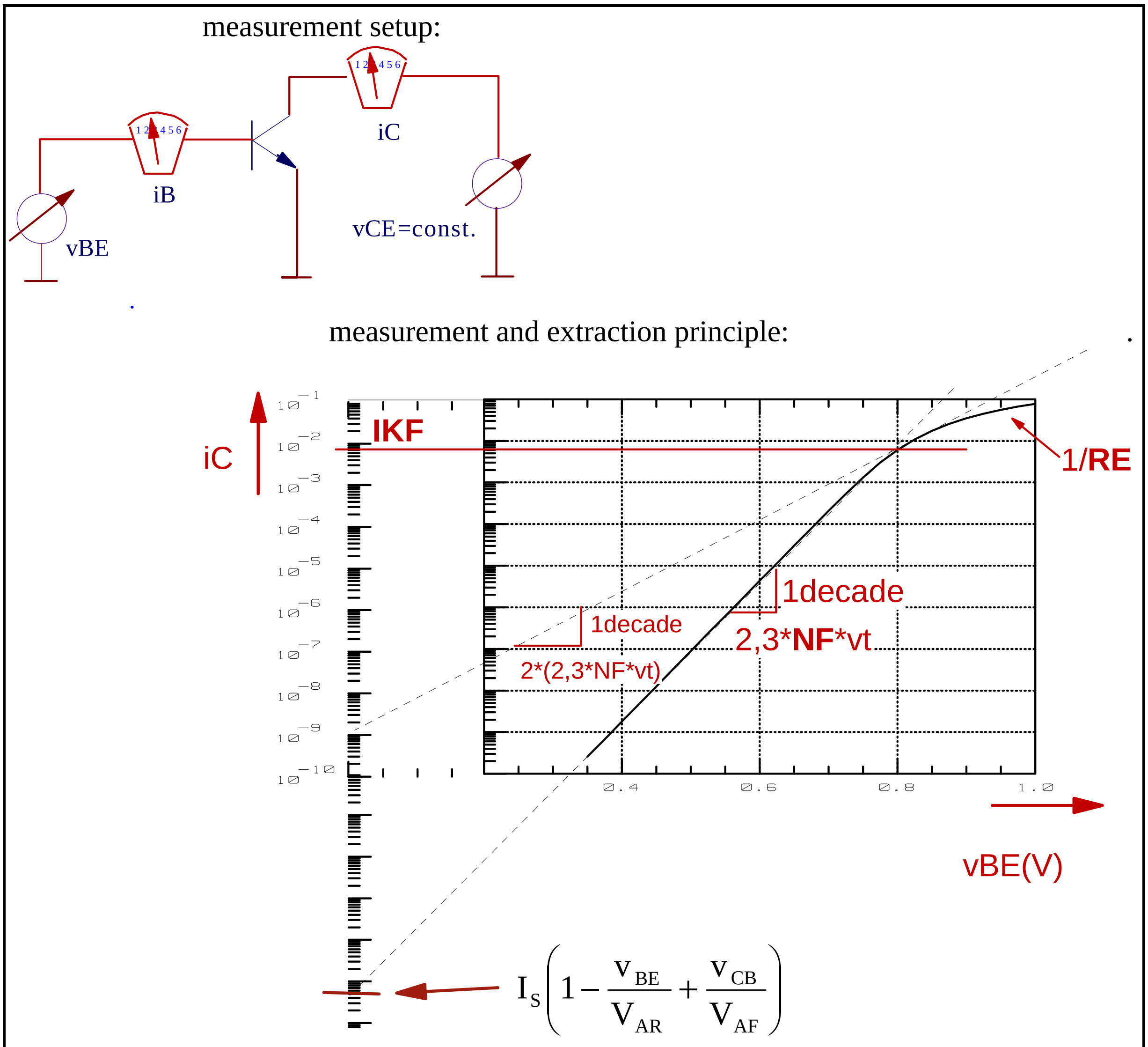


Fig.DC-4: Measurement of the Collector current vs. B-E voltage

The equation:

Provided that $v_{B'E'}=v_{BE}$ and $v_{B'C'}=v_{BC}$, we start again with the i_C formula (H) ... (L) from the introduction chapter:

$$i_C = 1/NqB\ (i_f - i_r) - i_r/BR - iBC_{rec}$$

or:

$$i_C = \frac{I_S}{N_{qB}} \left\{ \left(\exp\left[\frac{v_{BE}}{N_F V_T} \right] - 1 \right) - \left(\exp\left[\frac{v_{BC}}{N_R V_T} \right] - 1 \right) \right\} - \frac{I_S}{B_R} \left\{ \exp\left[\frac{v_{BC}}{N_R V_T} \right] - 1 \right\} - I_{SC} \left\{ \exp\left[\frac{v_{BC}}{N_C V_T} \right] - 1 \right\} \quad \text{(DC-12)}$$

with

$$NqB = \frac{q_{1S}}{2} * \left(1 + \sqrt{1 + 4q_{2S}}\right) \quad \text{(DC-13)}$$

for the modeling of the charge dependencies, especially the Base-width modulation factor

$$q_{1S} = \frac{1}{1 - \frac{v_{BE}}{V_{AR}} - \frac{v_{BC}}{V_{AF}}} \quad \text{(DC-14)}$$

and the hi-level injection effect (half the slope of log(iC) vs. vBE for high iC)

$$q_{2S} = \frac{i_f}{I_{KF}} + \frac{i_r}{I_{KR}} \quad \text{(DC-15a)}$$

or:

$$q_{2S} = \frac{I_S}{I_{KF}} \left\{ \exp\left[\frac{v_{BE}}{N_F V_T} \right] - 1 \right\} + \frac{I_S}{I_{KR}} \left\{ \exp\left[\frac{v_{BC}}{N_R V_T} \right] - 1 \right\} \quad \text{(DC-15b)}$$

In order to determine I_S from (DC-12) for small v_{BE}, i.e. no ohmic and no I_{KF} effects, we get for forward biasing

$$v_{BE} \sim 0{,}7V \ll |V_{AR}|$$

$$v_{BC} < 0\ V$$

and

$$|v_{BC}| \ll |V_{AF}|.$$

Therefore (DC-12) simplifies to:

$$i_C = \frac{I_S}{N_{qB}} \exp\left(\frac{v_{BE}}{N_F v_T} \right) \quad \text{(DC-16)}$$

Let's have a closer look to equ. (DC-13). Firstly, the formula reminds to apply the following series approach for small values of x :

$\sqrt{1+x} \approx 1+x/2$

what means for our case:

$$NqB = q_{1S} * (1 + q_{2S}) \qquad \text{(DC-17)}$$

NqB from (DC-17) is split into two parts: q_{1S} represents a lowering of the Collector current for increasing Early voltages (DC-14). This can be seen in the i_C Gummel plot as a curve shift to lower Collector currents. On the other hand, the other coefficient q_{2S} begins to contribute for high Collector currents above I_{KF} in forward operation resp. I_{KR} in reverse (DC-15a), and reduces the Collector current as well.

For the modeling of I_S and N_F, we consider the lower and medium current ranges well below the Effect of I_{KF} or the influence of the ohmic Resistor R_E . Therefore, (DC-16) simplifies to:

$$i_C = I_S \left(1 - \frac{v_{BE}}{V_{AR}} - \frac{v_{BC}}{V_{AF}}\right) \exp\left(\frac{v_{BE}}{N_F v_T}\right)$$

or, because vBC=-vCB

$$i_C = I_S \left(1 - \frac{v_{BE}}{V_{AR}} + \frac{v_{CB}}{V_{AF}}\right) \exp\left(\frac{v_{BE}}{N_F v_T}\right) \qquad \text{(DC-18)}$$

For bigger values of the Early voltages, the terms $|v_{Bx}| < |V_{Ax}|$ can be neglected and we obtain:

$$i_C = I_S \exp\left(\frac{v_{BE}}{N_F v_T}\right) \qquad \text{(DC-19)}$$

NOTE: compare (DC-19) with the measurement result given in the figure DC-4 above.

Extracting the parameters:

Following the curve fitting techniques given in the chapter on regression analysis in the appendix, a transformation can be applied to the measured data in order to obtain a linear context between the measured values of i_C and the stimulating values of v_{BE} in (DC-19):

A $\log_{10}$ conversion of (DC-19) gives:

$$\log(i_C) = \log(I_S) + \frac{v_{BE}}{N_F V_T} \log(e)$$

or:

$$\log(i_C) = \log(I_S) + \frac{1}{2{,}3026\, N_F V_T} v_{BE} \qquad \text{(DC-20a)}$$

This can be considered as a linear form:

```
                y     =     b     +        m     *   x          (DC-20b)
when setting:   |           |              |         |
                y = log (iC)               |         |
                            b = log (IS)   |         |
```

and

$$m = 1 / (2{,}3026\ N_F\ V_T) \Big|_{x = v_{BE}}$$

How to proceed:

We select a sub-range of the measured data, where the half-logarithmicly plotted data represent a straight line. Then the logarithmically converted i_{Ci} of this sub-range are interpreted as y- and the linear v_{BEi} values as x-data for the regression formula. Applying these formulas, we obtain y-intersect 'b' and the slope 'm' of the straight fitted line.

From comparing (DC-20a) with (DC-20b) we know how to re-substitute the parameters out of 'b' and 'm':

$$I_S = 10^{\,b} \qquad \text{(DC-21a)}$$

and

$$N_F = 1 / (2{,}3026\ m\ V_T) \qquad \text{(DC-21b)}$$

Validity of the extraction:

v_{BE} between 0,2V [no noise] and 0,7V [no high current effects]

WHAT TO DO IN IC-CAP:

open setup "/gp_classic_npn/dc/fgummel",
perform a measurement,
click a box into plot "ibic_vbe" around a linear range for the IS/NF extraction
click 'Copy to Variables' (check how the box bounds are exported into the
setup variables X_LOW, X_HIGH, Y_LOW, Y_HIGH)
perform transform "br_IS_NF" (box regression IS, NF), which refers to X_LOW etc.
simulate with the extracted parameter values.
optimize with transform "bo_IS_NF"

Have also a look into "/gp_classic_npn/dc/fgummel/READ_ME".

HINT:

Transforming the measured data such that the model parameter can be displayed directly against the stimulating voltage or current is another smart way to determine model parameters. In the case of NF this would mean to start with

$$i_C = I_S \exp\left(\frac{v_{BE}}{N_F v_T}\right)$$

to convert it logarithmically in order to obtain

$$\ln(i_C) = \ln(I_S) + \frac{1}{N_F v_T} * v_{BE}$$

This is the mathematical representation of the half-logarithmic Gummel plot for i_C. The parameter NF is proportional to the slope and we have therefore to differentiate $\ln(i_C)$ with respect to v_{BE} and obtain:

$$\frac{\partial \ln(i_C)}{\partial v_{BE}} = \frac{1}{N_F v_T}$$

Solved for NF gives

$$N_F = \frac{1}{V_T * \frac{\partial(\ln(i_C))}{\partial(v_{BE})}}$$

Therefore, if we display the calculated NF (what is the 'effective NF' for every measured data point) versus v_{BE}, we get

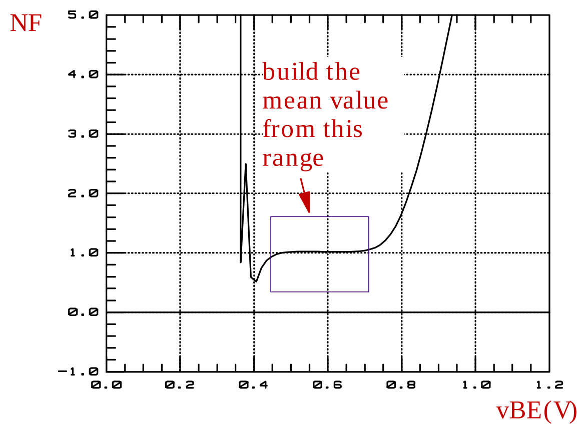


Fig.DC-5: Direct visual extraction for parameter NF

This allows us to check, if the model is able to fit the measured data at all (if there is a constantly flat range) and then to easily extract the parameter as the mean value of that flat range.

In directory "visu_n_extr" of this toolkit you will find more IC-CAP model files that follow the idea of direct visual extraction. See also appendix A for more infos.

Some comments on $i_{c\_measured}$ above IKF:

Referring to many graphical parameter extraction methods and also to fig.DC-4 above, some more background information is given for the model curve of i_C above IKF.

From (DC-17) we get neglecting the Early-effect:

$$NqB = \frac{1}{2} * \sqrt{1 + 4\frac{i_C * N_{qB}}{I_{KF}}}$$

or solved for N_{qB}:

$$NqB = 1 + \frac{i_C}{I_{KF}}$$

Let us consider the two cases:

$i_C < I_{KF}$

here is

$$N_{qB} \sim 1$$

and introducing this into (DC-16) gives finally:

$$i_C = I_S \exp\left[\frac{v_{BE}}{N_F V_T}\right] \qquad \text{(DC-22a)}$$

$i_C > I_{KF}$

and here is

$$N_{qB} \sim \frac{i_C}{I_{KF}}$$

$$i_C = \frac{I_{KF}}{i_C} I_S \exp\left[\frac{v_{BE}}{N_F V_T}\right]$$

or:

$$i_C = \sqrt{I_{KF} I_S} \exp\left[\frac{v_{BE}}{2 N_F V_T}\right] \qquad \text{(DC-22b)}$$

<u>Interpreting the result:</u>
From (DC-22b) we learn that the i_C curve has half the slope for currents above IKF (see fig.DC-4). In practice, however, there is always an overlay with the ohmic resistor RE, and therefore (DC-22b) is not so well suitable for extracting IKF.
However, the overlaying parameter RE is affecting basically both, the i_C and the i_B curve in the same way. This means that the effect of RE cancels out for the beta curve ß=i_C/i_B. On the other hand, parameter IKF affects only i_C. Therefore, IKF is commonly extracted from the beta curve of the transistor !

MODELING THE BASE CURRENT:

EXTRACTION OF BF , ISE AND NE

B_F	ideal forward maximum beta
I_{SE}	B-E leakage saturation current
N_E	B-E leakage emission coefficient

In the literature, the three parameters of this chapter are most often introduced with their corresponding influence on the different ranges of the i_B curve in DC-6.

In practice, there is most sometimes an overlay of the influences. This is especially true for BF in the beta plot (overlaid from IKF and NE). Also, modern transistors have pretty low recombination effects for the B-E diode: the 'famous knee' (see finger pointer in fig.DC-6) is not visible. Therefore we will not follow the graphical extraction method, but develop another method instead. We will derive a formula for the 3 parameters directly from measured data that has been taken from the range around the 'knee'.

same measurement setup as in fig.DC-4

extraction principle:

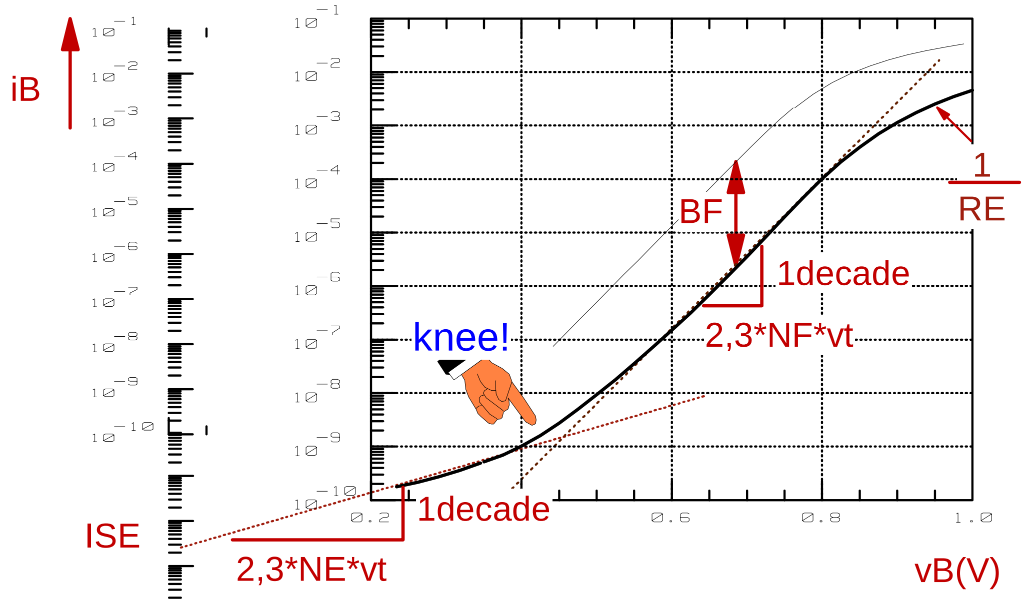


Fig.DC-6: measurement of the Base current vs. B-E voltage

The equation:

Provided that $v_{B'E'}=v_{BE}$ and $v_{B'C'}=v_{BC}$, then

$$i_B = i_{BE} + i_{BC}$$

see equu. (A) ... (G) of the introduction chapter

$$= \frac{I_S}{B_F}\left\{ \exp\left[\frac{v_{BE}}{N_F V_T}\right] - 1 \right\} + I_{SE}\left\{ \exp\left[\frac{v_{BE}}{N_E V_T}\right] - 1 \right\}$$

$$+ \frac{I_S}{B_R}\left\{ \exp\left[\frac{v_{BC}}{N_R V_T}\right] - 1 \right\} + I_{SC}\left\{ \exp\left[\frac{v_{BC}}{N_C V_T}\right] - 1 \right\} \qquad \text{(DC-23)}$$

We assume once again that:

$v_{BE} \sim 0{,}7V << |V_{AR}|$

and $v_{BC} < 0\ V$

This simplifies equ.(DC-23) to:

$$i_B = \frac{I_S}{B_F} \exp\left[\frac{v_{BE}}{N_F V_T}\right] + I_{SE} \exp\left[\frac{v_{BE}}{N_E V_T}\right] \qquad \text{(DC-24)}$$

Introducing (DC-19) -i.e. the Collector current i_C with neglected high current effects- into (DC-24) yields the pretty simple form:

$$i_B = \frac{i_C}{B_F} + I_{SE} * \exp\left(\frac{v_{BE}}{N_E * V_T}\right) \qquad \text{(DC-25)}$$

We will use both $i_C = f(v_{BE})$ and $i_B = f(v_{BE})$ from the simultaneously measured currents of the Gummel-Poon measurement of fig.DC-4.

We now have i_B as a function of v_{BE} as desired.

Extracting the parameters:

This equation (DC-25) is one of the few cases during the bipolar modeling, where a non-linear transform applied to the measured data doesn't give a straight line. (At least, the author had not sufficient intuition!). Therefore the partial derivations of the curve fitting error in (DC-25) versus B_F, I_{SE} and N_E have to be calculated and then set to zero. The solution of this system of equations finally gives these 3 parameters.

As i_B ranges from pico- to milli-Ampère, we will have to minimize the relative error between measured and fitted curve. Thus we get from (DC-25) after dividing by i_B:

$$1 = \frac{i_C}{i_B B_F} + \frac{I_{SE}}{i_B} \exp\left[\frac{v_{BE}}{N_E V_T}\right] \qquad \text{(DC-26)}$$

Equation (DC-26) is only approximately true for the real measured data i_{Bi}, i_{Ci} and v_{BEi}. Therefore it is expanded by the individual error E_{rel_i} for every data point of index i:

$$1 + E_{rel_i} = \frac{i_{Ci}}{i_{Bi} B_F} + \frac{I_{SE}}{i_{Bi}} \exp\left[\frac{v_{BEi}}{N_E V_T}\right] \qquad \text{(DC-27)}$$

or:

$$E_{rel_i} = \frac{i_{Ci}}{i_{Bi} B_F} + \frac{I_{SE}}{i_{Bi}} \exp\left[\frac{v_{BEi}}{N_E V_T}\right] - 1 \qquad \text{(DC-28)}$$

Using least means square techniques we now have:

$$E_{tot} = \sum_{i=1}^{N} E_{rel_i}^2 = \sum_{i=1}^{N} \left[\frac{i_{Ci}}{i_{Bi} B_F} + \frac{I_{SE}}{i_{Bi}} \exp\left(\frac{v_{BEi}}{N_E V_T}\right) - 1 \right]^2 \overset{!}{=} \text{Minimum} \qquad \text{(DC-29)}$$

It can be shown that the parameters B_F and I_{SE} can be separated out of the partial derivations with respect to B_F and I_{SE} with a reasonable effort. This is unfortunately not possible for N_E . This parameter has to be iterated - similar to VJ of the space charge capacitor - until the sum of individual errors according to (DC-29) is minimized.

Step by step:

The partial derivation of (DC-29) versus B_F is:

$$\frac{1}{B_F} \sum_{i=1}^{N} \frac{i_{Ci}^2}{i_{Bi}^2} + I_{SE} \sum_{i=1}^{N} \frac{i_{Ci}}{i_{Bi}^2} \exp\left(\frac{v_{BEi}}{N_E V_T}\right) - \sum_{i=1}^{N} \frac{i_{Ci}}{i_{Bi}} = 0 \qquad \text{(DC-30)}$$

and versus I_{SE}:

$$\frac{1}{B_F} \sum_{i=1}^{N} \frac{i_{Ci}}{i_{Bi}^2} \exp\left(\frac{v_{BEi}}{N_E V_T}\right) + I_{SE} \sum_{i=1}^{N} \frac{1}{i_{Bi}^2} \exp\left(\frac{2 v_{BEi}}{N_E V_T}\right) - \sum_{i=1}^{N} \frac{1}{i_{Bi}} \exp\left(\frac{v_{BEi}}{N_E V_T}\right) = 0 \qquad \text{(DC-31)}$$

(DC-30) is expanded by

$$-\sum_{i=1}^{N}\frac{i_{Ci}}{i_{Bi}^2}\exp\left(\frac{v_{BEi}}{N_E V_T}\right)$$

and (DC-31) by

$$\sum_{i=1}^{N}\frac{i_{Ci}^2}{i_{Bi}^2}$$

These two new equations are added and their sum is solved for I_{SE}:

$$I_{SE} = \frac{\sum_{i=1}^{N}\frac{1}{i_{Bi}}\exp\left(\frac{v_{BEi}}{N_E V_T}\right)\sum_{i=1}^{N}\frac{i_{Ci}^2}{i_{Bi}^2} - \sum_{i=1}^{N}\frac{i_{Ci}}{i_{Bi}}\sum_{i=1}^{N}\frac{i_{Ci}}{i_{Bi}^2}\exp\left(\frac{v_{BEi}}{N_E V_T}\right)}{\sum_{i=1}^{N}\frac{1}{i_{Bi}^2}\exp\left(\frac{2v_{BEi}}{N_E V_T}\right)\sum_{i=1}^{N}\frac{i_{Ci}^2}{i_{Bi}^2} - \left[\sum_{i=1}^{N}\frac{i_{Ci}}{i_{Bi}^2}\exp\left(\frac{v_{BEi}}{N_E V_T}\right)\right]^2} \qquad \text{(DC-32)}$$

Now we can also separate B_F from (DC-30):

$$B_F = \frac{\sum_{i=1}^{N}\frac{i_{Ci}^2}{i_{Bi}^2}}{\sum_{i=1}^{N}\frac{i_{Ci}}{i_{Bi}} - I_{SE}\sum_{i=1}^{N}\frac{i_{Ci}}{i_{Bi}^2}\exp\left(\frac{v_{BEi}}{N_E V_T}\right)} \qquad \text{(DC-33)}$$

Validity of the extraction: v_{BE} above measurement resolution and below high current effects.

How to proceed:

A subset of the measured data i_{Bi} and i_{Ci} , i.e. the range around the 'KNEE' (see fig.DC-6) are selected and introduced into equations (DC-32) and (DC-33). Next a suitable starting value for N_E is selected (e.g. $N_E = 1$) and the error according to (DC-29) is calculated. N_E is then incremented until this error becomes a minimum. The triplet of N_E, B_F and I_{SE} of this minimized error is the final parameter extraction result.

NOTE: the complexity of (DC-32) and (DC-33) illustrates that transforming measured data to a linear context and applying linear regression techniques is often a much smarter approach for parameter extraction.

WHAT TO DO IN IC-CAP:

in setup "/gp_classic_npn/dc/fgummel",
click a box into plot "ibic_vbe" around the 'knee' at low vb,
click "Copy to Variables",
perform transform "br_ISE_BF_NE" (box regression ISE, BF, NE),
simulate with the extracted parameter values.
perform transform "bo_ISE_BF_NE" (box optimization ISE, BF, NE),
you may also try the tuning function in "tune_ISE_BF_NE"

If there is no 'knee' with your measured transistor, the Base current recombination effect does not occur. In this case, switch off the Base current recombination effect in the G-P model. This can be done by setting ISE to a very small value (ISE=1E-30) and the slope parameter NE to a flat slope (NE=2).

Have also a look into "/gp_classic_npn/dc/fgummel/READ_ME".

<u>Note:</u>
For low values of VAR, the Collector current formula of (DC-19) inserted into (DC-25) is not quite correct. It would lead to a too low extracted value of BF, due to the shift of the iC Gummel plot. Equation (DC-18), without the assumption of big Early voltages, is better in this case. Therefore, correct the measured Collector current values to

$$i^*_{C\_measured} = \frac{i_{C\_measured}}{1 - \frac{v_{BE}}{V_{AR}}}$$

Note: VAR<<VAF, so this correction is sufficient

before inserting them into equations (DC-32) and (DC-33).

MODELING THE CURRENT AMPLIFICATION AT HIGH CURRENT

EXTRACTION OF IKF

I_{KF} forward beta high current roll-off

Referring to fig. DC-4, IKF models the Webster push-out effect. This effect describes a decrease of the proportionality of log(iC) versus vBE. Unfortunately, as already mentioned, this effect is also overlaid by RE. However, while RE affects mainly both, iB and iC, IKF only affects iC. Therefore, it can be best extracted out of ß=iC/iB.

From a modeling standpoint, the beta plot should not be considered isolated from its origin, the iB and iC curve. Therefore, we display it always versus the same stimulus vBE, together with iB and iC. This helps a lot in better understanding the influence of the parameters ISE and NE on the increase of beta, and of IKF for the decrease.

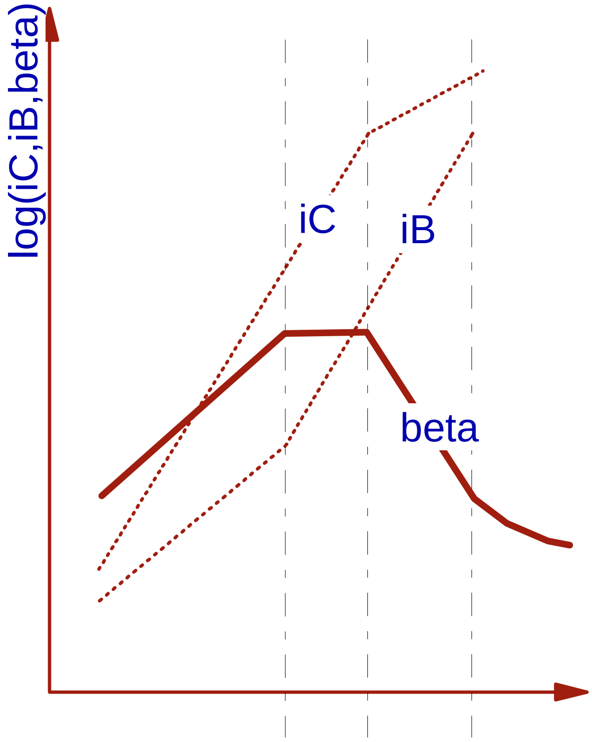


Fig.DC-7: Plotting beta together with i_C and i_B for better understanding of the context of the model parameters

It becomes also clear why for some transistors, BF seems not to affect the maximum of the beta trace at all: there is no parallel region between log(iC) and log(iB) in the Gummel plot, or referring to the beta plot, IKF reduces already beta, before it can reach the value of the BF for increasing bias. This is shown in fig. DC-8.

same measurement setup as for fig.DC-4

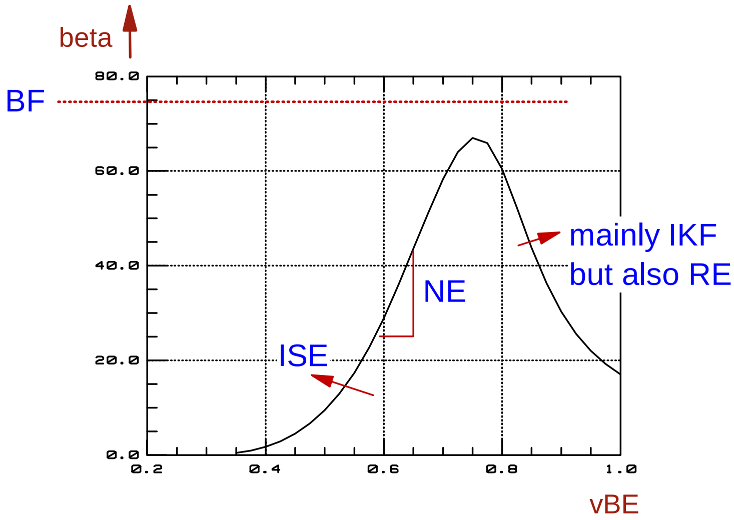


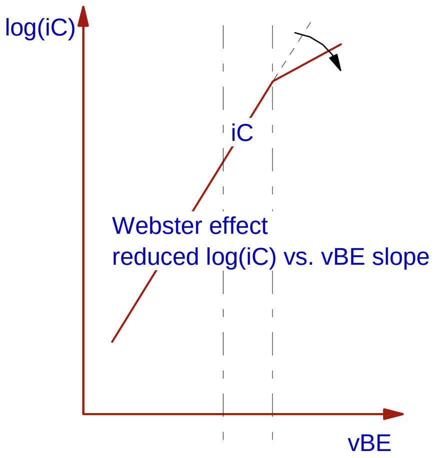
Fig.DC-8: Beta from the measurements of fig.DC-4 and fig.DC-3.

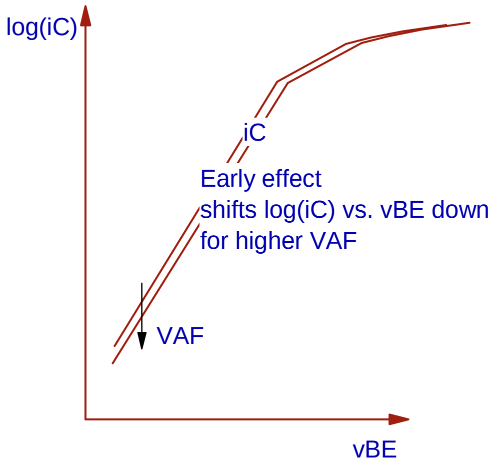
Note: log(ß) is plotted versus v_{BE} . The conventional plot using log(ß) versus log(i_C) is identical to it, provided we neglect high-current effects. But this way of plotting log(ß) is more straightforward, because it displays a measurement result versus a stimulus and <u>not</u> another measurement result.

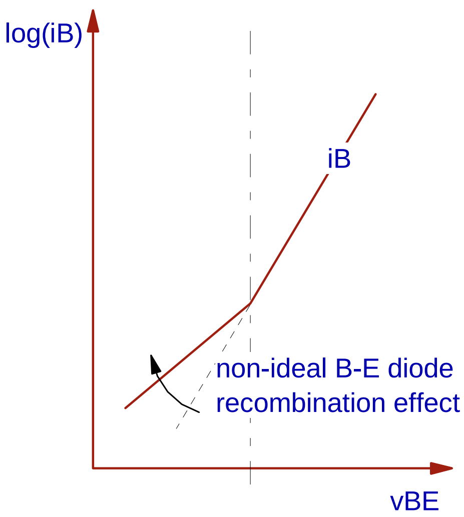
NOTE: for best results, estimate or extract R_E first!

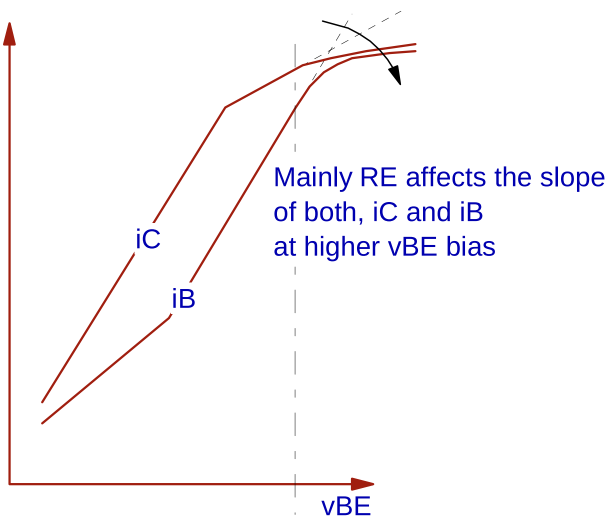
Hint:
avoid thermal self-heating effects!. For Collector currents >10...50mA, thermal self-heating has to be taken into account. This becomes visible if the beta-plot for a *forward and reverse vBE sweep looks different* at high vBE. To avoid this, DC pulsed measurements with pulse widths about 1us should be used in this case (the HP4142 offers only pulse widths ≥100us!).

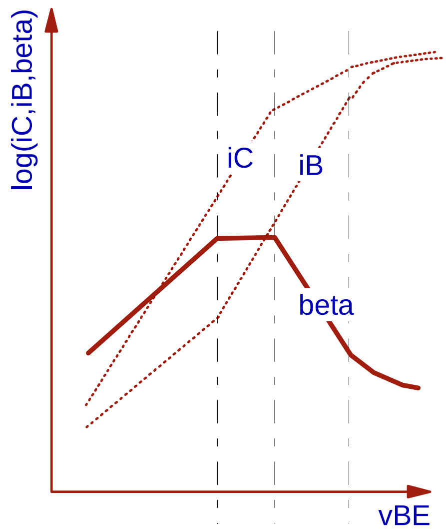
Before we start with the extraction of IKF, we are now ready to understand the following schematized Gummel plots. They characterize at a glance the different effects for the Base and Collector current in the Gummel-Poon model-











The equation:

The current amplification is defined as:

$$\beta = \frac{i_C}{i_B} \qquad \text{(DC-34)}$$

Provided that: $v_{B'E'}=v_{BE}$ and $v_{B'C'}=v_{BC}$ and further

$$v_{BE} \ll V_{AR}$$

and $$v_{BC} < 0\ \mathrm{V}$$

We introduce (DC-24) for i_B and(DC-16) for i_C into (DC-34):

$$\frac{i_C}{i_B} = \frac{\frac{I_S}{N_{qB}} \exp\left[\frac{v_{BE}}{N_F V_T}\right]}{\frac{I_S}{B_F} \exp\left[\frac{v_{BE}}{N_F V_T}\right] + I_{SE} \exp\left[\frac{v_{BE}}{N_E V_T}\right]} \qquad \text{(DC-35)}$$

We further introduce an approximation for N_{qB}, see equ.DC-18. This gives:

$$\frac{i_C}{i_B} = \frac{\left(1 - \frac{i_C}{I_{KF}} - \frac{v_{BC}}{V_{AF}}\right) I_S \exp\left[\frac{v_{BE}}{N_F V_T}\right]}{\frac{I_S}{B_F} \exp\left[\frac{v_{BE}}{N_F V_T}\right] + I_{SE} \exp\left[\frac{v_{BE}}{N_E V_T}\right]} \qquad \text{(DC-36)}$$

Divided by

$$I_S \exp\left[\frac{v_{BE}}{N_F V_T}\right]$$

we get:

$$\frac{i_C}{i_B} = \frac{\left(1 - \frac{i_C}{I_{KF}} - \frac{v_{BC}}{V_{AF}}\right)}{\frac{1}{B_F} + \frac{I_{SE}}{I_S} \exp\left\{\frac{v_{BE}}{V_T}\left[\frac{1}{N_E} - \frac{1}{N_F}\right]\right\}} \qquad \text{(DC-37)}$$

Extracting the parameters:

As we want to consider again relative errors, we proceed as in the chapter of the determination of the i_B parameters:

$$1 = \frac{i_B}{i_C} * \frac{\left(1 - \frac{i_C}{I_{KF}} - \frac{v_{BC}}{V_{AF}}\right)}{\frac{1}{B_F} + \frac{I_{SE}}{I_S} \exp\left\{ \frac{v_{BE}}{V_T} \left[\frac{1}{N_E} - \frac{1}{N_F} \right]\right\}}$$

This formula is again more ore less true for the measured data i_{Ci} and i_{Bi} with the stimulating voltage v_{BE}. Thus we have to introduce again an individual error E_{reli} for each measured data point of index i:

$$1 + E_{reli} = \frac{i_{Bi}}{i_{Ci}} * \frac{1 - \frac{i_{Ci}}{I_{KF}} - \frac{v_{BCi}}{V_{AF}}}{\frac{1}{B_F} + \frac{I_{SE}}{I_S} \exp\left\{ \frac{v_{BEi}}{V_T} \left[\frac{1}{N_E} - \frac{1}{N_F} \right]\right\}} \qquad \text{(DC-38)}$$

Finally the total error for all measured data (1 .. N) is (least means square):

$$E = \sum_{i=1}^{N} E_{reli}^2 = \sum_{i=1}^{N} \left\{ \frac{i_{Bi}}{i_{Ci}} * \frac{1 - \frac{i_{Ci}}{I_{KF}} - \frac{v_{BCi}}{V_{AF}}}{\frac{1}{B_F} + \frac{I_{SE}}{I_S} \exp\left\{ \frac{v_{BEi}}{V_T} \left[\frac{1}{N_E} - \frac{1}{N_F} \right]\right\}} - 1\right\}^2 \qquad \text{(DC-39)}$$

<u>How to proceed:</u>
In order to keep things simple, (DC-39) is solved for a best I_{KF} by iteration. Thus I_{KF} is set to a starting value, e.g. 10A, and then divided by 2 in every iteration, until the total error given in (DC-39) is minimized. Fine-tuning is then done by the optimizer.

WHAT TO DO IN IC-CAP:

open setup "/gp_classic_npn/dc/fgummel" and plot "ibic_vbe"
(beta is the right-axis data), then
perform transform "e_IKF" (extract IKF) and
check the simulation result.
run transform 'o_BF_IKF_RE' for fine-tuning the parameters of this setup.

Have also a look into "/gp_classic_npn/dc/fgummel/READ_ME".

and also in some alternate methods on the IKF-extraction in the 'direct visual parameter extraction' model file.

EXTRACTION OF THE REMAINING REVERSE PARAMETERS

EXTRACTION OF NR, ISC, NC, BR, IKR

The reverse modeling can be performed like the forward modeling. Simply exchange Emitter and Collector.

WHAT TO DO IN IC-CAP:

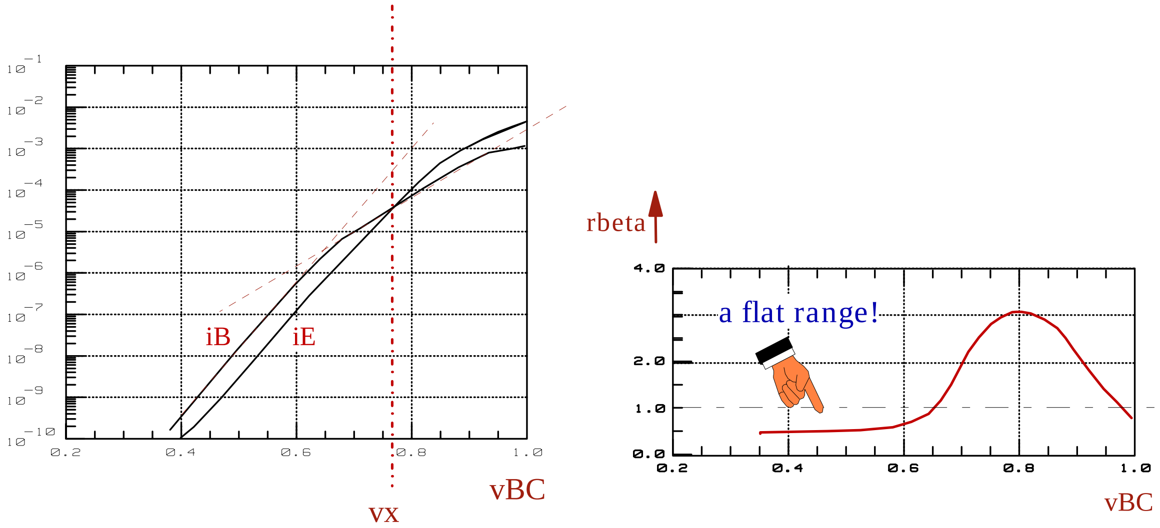
open transform README in setup "/gp_classic_npn/dc/rgummel" and follow the modeling sequence given there

Last not least, macro 'extract_n_opt_DC' includes a suitable automated modeling strategy for both DC forward and reverse.
Included in this example is also the interesting and pretty often recognizable effect, that the reverse Early voltage is affecting the forward modeling, due to its low value.
The strategy used in this macro covers that effect by looping a bit between forward and reverse extraction and optimization.

This sequence may be different for your actual transistor. Just correct the macro if required.

Note on reverse Gummel-Poon modeling

If your reverse beta curve and reverse Gummel plot look like below,



(a steeper slope of log(i_B) versus v_{BC} for $v_{BC} < v_x$, something that is not included in the Gummel-Poon model), you might consider replacing the Gummel-Poon recombination modeling (parameters ISC and NC) by a external diode with its parameters IS, N and RS.

For more details, refer to file rgummel_special.mdl under directory 'more_files' in this toolkit file collection.
See also the chapter on the limitations of the Gummel-Poon model at the end of this manual.

A C S M A L L S I G N A L M O D E L I N G O V E R V I E W

We are now ready to consider the basics of modeling for frequencies higher than 100MHz. It is assumed that the measurements have been made on the pure semiconductor device without being affected by packaging parasitics, bond pads or other parasitics. If this is not possible, de-embedding techniques have to be applied. This means to find the proper semiconductor behavior out of the distorted measurement by de-embedding.

It should also be mentioned that the probe pins have to have an excellent frequency performance within the transistor measurement frequency range. Once again the key to meaningful AC measurements and thus modeling is a good network analyzer calibration with excellent standards and a correctly defined calibration kit data in the network analyzer.

If you need additional support for de-embedding, calibration and for better understanding S-parameters, please refer to the additional toolkits for IC-CAP. Please contact the author.

Let us go first for the AC small signal equivalent schematic. It can be derived from fig.2b of the introduction chapter as a linearization at each bias point of the transistor.

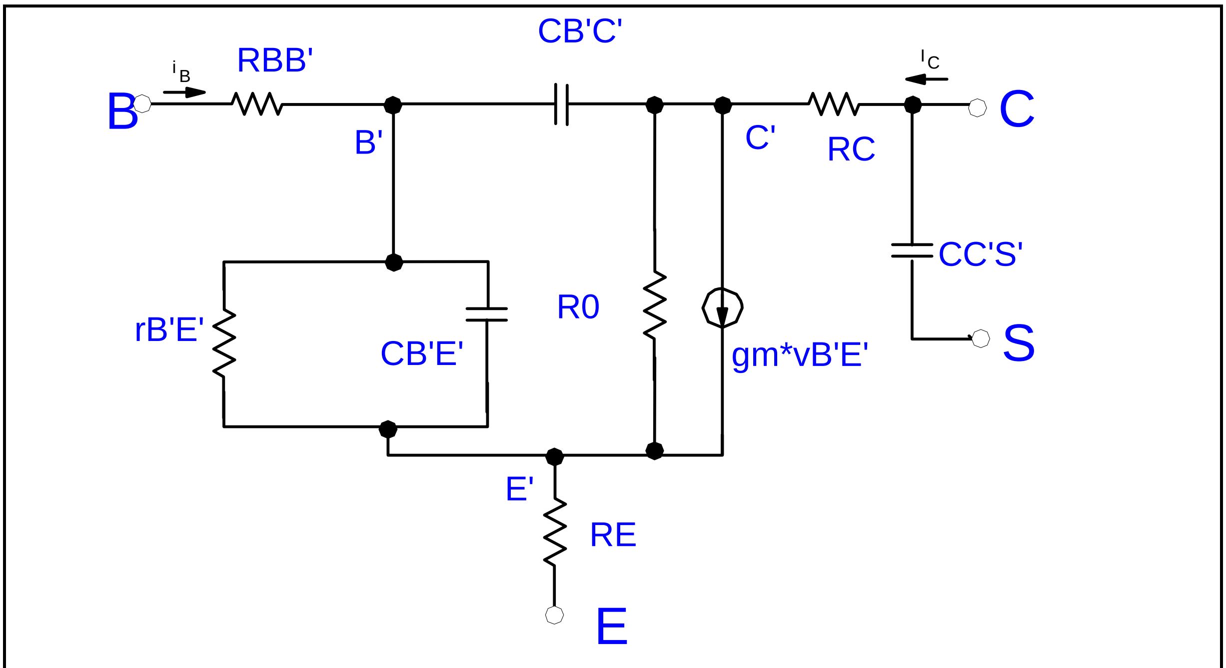


Fig.AC-1: AC small signal schematic of the bipolar transistor
NOTE: X_{CJC} effect neglected.

The following equations give the values of the internal elements in fig.AC-1.
They represent the linearized DC- and CV-equations at the DC operating point.

$g_{B'E'}$

From equ.(B) in the introductory chapter we get from the derivative of i_B versus $v_{B'E'}$:

$$g_{B'E'} = \frac{I_S}{B_F\, N_F\, V_T} \exp\left[\frac{v_{B'E'}}{N_F\, V_T} \right] + \frac{I_{SE}}{N_E\, V_T} \exp\left[\frac{v_{B'E'}}{N_E\, V_T} \right] \qquad \text{(AC-1)}$$

where the second term can most often be neglected for operating points of i_C above 1 uA.

---

g_0

The output conductance is:

$$g_0 = -di_C / dv_{B'C'} - di_B / dv_{B'C'} \qquad \text{(AC-2)}$$

$C_{B'E'}$ or C_{PI}

including the delay time effect modeled by T_{FF} is given in equ.(P) and (R3) of the introductory chapter for the particular operating point voltages.
As a first order estimation, $C_{B'E'}$ simplifies to

$$C_{B'E'} = T_{FF} \frac{d\, i_C}{d\, v_{B'E'}} \approx T_{FF}\, g_m \qquad \text{(AC-3)}$$

---

while

$C_{B'C'}$ or C_{MU} (MU or μ stands for 'mutual'

simplifies because of $v_{BC} < 0$ at forward operation to:

$$C_{B'C'} = \frac{C_{jC}}{\left[1 - v_{B'C'} / V_{JC}\right]^{m_{JC}}} \qquad \text{(AC-4)}$$

---

g_m

The transconductance g_m finally is using equ. (H)

$$\frac{d\, i_C}{d\, v_{B'E'}} + \frac{d\, i_C}{d\, v_{B'C'}}. \qquad \text{(AC-5)}$$

MODELING THE BASE RESISTOR $r_{BB'}$

Extraction of RB, IRB and RBM

R_B	zero bias Base resistance
I_{RB}	curr. at medium Base resistance
R_{BM}	min.Base resistance at hi current

It is assumed that $X_{CJC} = 1$, and that ß is the DC current amplification.

This chapter explains how to model the Base resistor from S11 data.
It is organized like this:

- derivation of the small signal schematic for the parameter extraction
- short introduction into the basics of the Smith chart
- discussing the expected frequency dependence of rBE, considering rBB' constant
- enhancing the schematic for the bias-dependent omic resistor rBB'

As an approximation to keep the equations simpler, we further assume: $v_{BE} = v_{B'E'}$ and $v_{BC} = v_{B'C'}$. Simulations and optimizer runs after the parameter estimation will eliminate this simplification.

The measurement setup for the $r_{BB'}$ characterization is given below in fig.RBB-1.

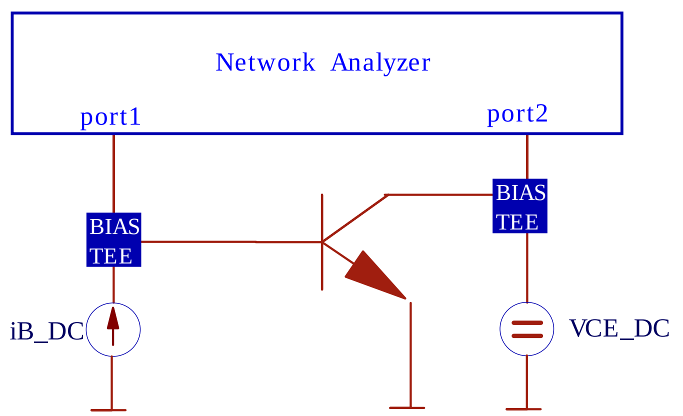


Fig.RBB-1: Measurement setup for the rBB measurement

To begin with, we refer to fig.AC-1 from the previous chapter. We simplify it to cover mainly the imput impedance. This leads to the schematic of fig.RBB-2. This figure explains the two cases: frequency -> 0 Hz and frequency -> ∞ Hz frequency -> 0 Hz, $C_{B'C'}$ and R_C omitted

frequency -> infinite, $C_{B'C'}$ and R_C omitted

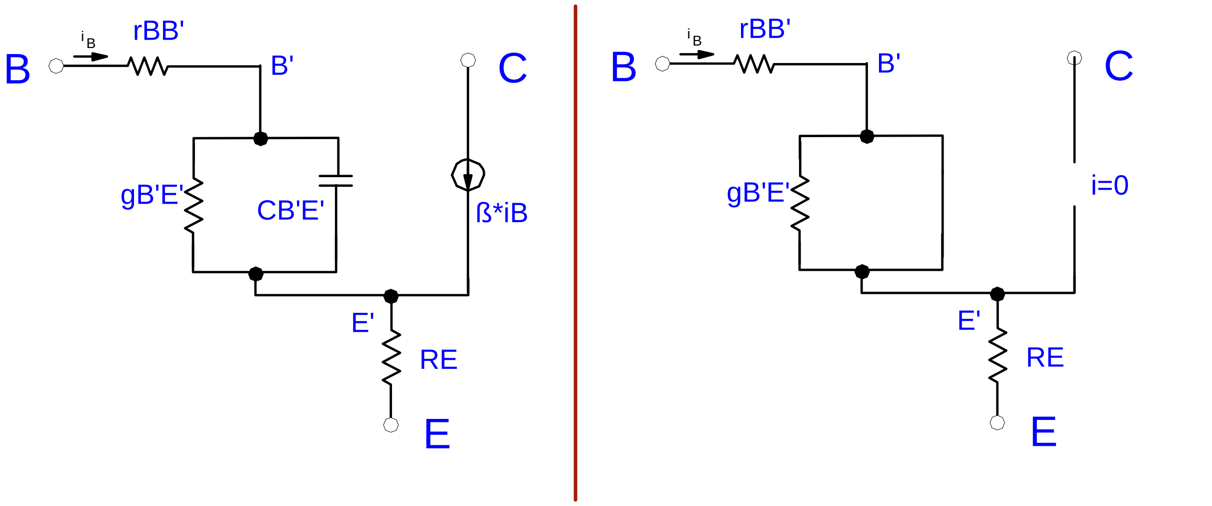


Fig.RBB-2:
Simplified AC Base-Emitter input impedance scheme for low and high frequencies

In order to evaluate the schematic and the device parameters of fig.RBB-2, we have to consider the measured S11 data. This is best done by displying them in a Smith chart.

As a reminder, a Smith chart transforms the right side of the complex resistor plane R into the area of a circle of radius '1' using the transform

$$\underline{S11} = \frac{\underline{R} - 50}{\underline{R} + 50}$$

with the NWA measurement impedance of 50 Ohm.

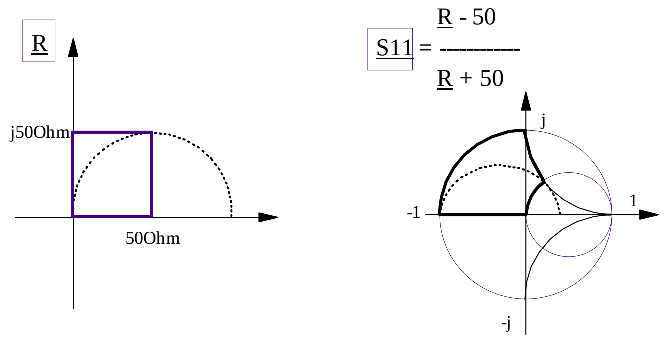
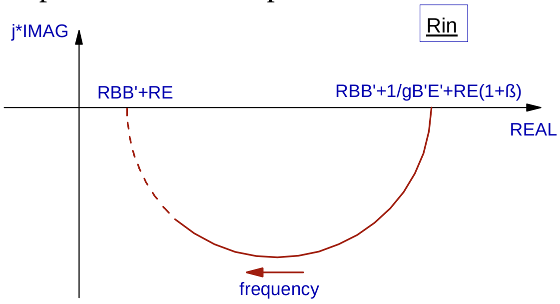


Fig.RBB-3: Note on the Smith chart transformation
(S_{11} = 1 for R = infinite , S_{11} = 0 for R = 50 Ohm , S_{11} = -1 for R = 0 Ohm)

Therefore, we can use S11 instead of H11 for the RIN modeling as well and our measurement result should look like figure RBB-4 .

Note: In order to get familiar with the problem, we consider first the hypothetical case that rBB' is *no* function of bias. In other words, the Base resistor is considered as a constant, ohmic resistor RBB'.

Complex resistor plane for the interpretation of RIN



and the same measurement result displayed in a Smith chart:

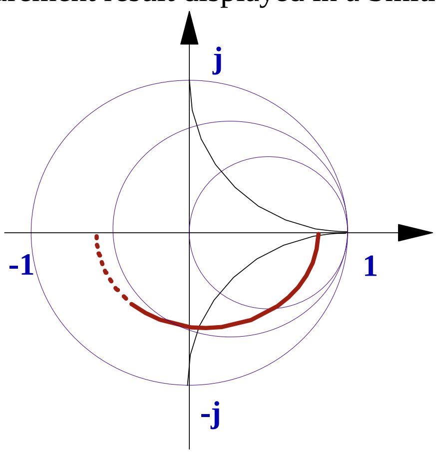


Fig.RBB-4: idealized input resistor curve, assuming an ohmic, bias-independent RBB'

Ideally, <u>R</u>IN should look like a circle.
The starting point at DC is RBB'+1/gB'E'+RE(1+ß). For higher frequencies, $C_{B'E'}$ will act more and more like a short and eliminate the influence of resistor 1/gB'E'. For infinite frequencies, <u>R</u>IN should hit the x-axis at $R_{BB'} + R_E$ (effects of $C_{B'C'}$ and R_C omitted!). Now the Base-Emitter capacitance has completely shorted $1/g_{B'E'}$ and thus the transconductance g_m became 0 as well. This means that the transistor has no beta any more.

S11 / Smith chart	=	$R_{BB'} + R_E$	(frequency -> infinite) (RBB-1)

As R_E is known from DC measurements, the value of $R_{BB'}$ can be estimated quite accurately.

This method is advantageous because the estimation of the Base resistor is affected only by the parameter R_E. Moreover, there is mostly $R_{BB'} >> R_E$, so that the influence of a uncertain value of R_E is minimized using this method.

So far we considered $r_{BB'}$ to be simply ohmic, i.e.constant. In reality, $r_{BB'}$ is modeled more complexly. One separate resistor from the outer Base contact to the inner Base contact (ohmic R_{BM}) and a bias-dependent part from the inner Base contact to inside the inner Base. This means, the higher i_B, the more i_C is extending its flow area closer to the internal Base contact due to current crowding. This means, we expect a lower Base resistor for higher bias. The sketch below depicts that. The Gummel-Poon model combines both resistors into a single, bias dependent one.

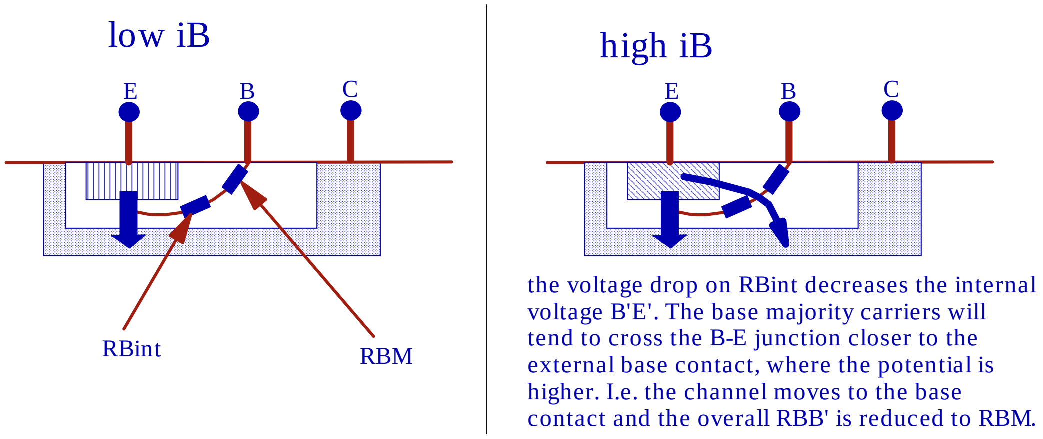


Fig.22a: current crowding leads to a bias dependent Base resistor

Now, overlying this DC bias dependency with the frequency dependance from above, we end up with S11 curves like sketched in fig.RBB-6.

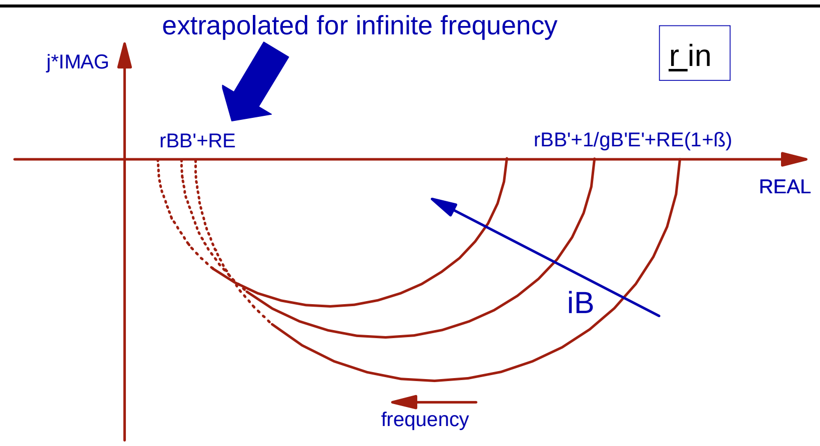


Fig.RBB-6: idealized input resistor curve with a real, bias-dependent Base resistor $r_{BB'}$, effects of $C_{B'C'}$ and R_C omitted.

Real-world measurement curves will look like these curves at low frequencies only. This is due to the overlay of more second order effects. In order to separate rBB' with the proposed method, we must fit circles to the S11 curves at low frequencies and then calculate the x-intersect from an extrapolation of the circle for infinite frequency, which is then assumed to be equal to rBB'+RE. This is shown in fig.RBB-7.

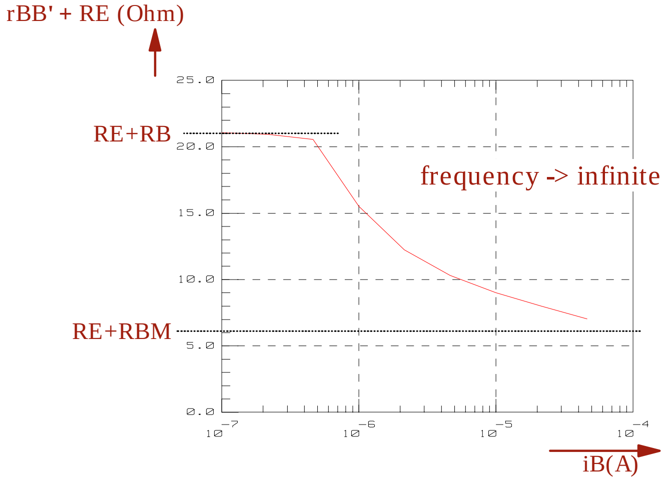


Fig.RBB-7: Idealized ohmic Base- plus Emitter-resistor versus i_B

Notes on some limitations of this extraction strategy:

Because of the influence of 1/gB'E' on the range where the circle must be fitted to, iB should be as high as possible to not dominate the r BB' - effect by 1/gB'E' . Also, to keep the rBB' influence dominant over RE (1+ß), iB should also be as high as possible, so that iC > IKF, and therefore ß, is as small as possible.

Unfortunately, the rBB' -measurement could now be dominated by thermal effects. Moreover, this range of iB typically is also not the operating one. This contradicts general rule to always concentrate on meaningful measurements close to the operating range for good parameter extractions.

Finally, the trace of Fig.RBB-7 is often overlaid by the parameters TF, ITF, XTF and VTF, which will be extracted next. Therefore, an optimization (of the S11 parameters) of this setup should only be applied after the fitting of these transit time parameters.

The extraction strategy:

Circles must be fitted to the low-frequency sections of interest. They are centered to the x-axis. The suitable circle formula is::

$$(x - x_0)^2 + y^2 = r^2 \qquad \text{(RBB-2)}$$

or

$$x^2 + y^2 = r^2 - x_0^2 + 2\, x_0\, x$$

This again can be considered as a linear form (!) with

$$y_{lin} = b + m\, x_{lin}$$

where

$$x^2 + y^2 = y_{lin} \qquad \text{(RBB-3a)}$$

$$r^2 - x_0^2 = b \qquad \text{(RBB-3b)}$$

$$2\, x_0 = m \qquad \text{(RBB-3c)}$$

and

$$x = x_{lin} \qquad \text{(RBB-3d)}$$

This means: The measured data x_i and y_i are introduced into equ.(RBB-3a). Next the $y_{lin}(i)$ are plotted versus the $x_{lin}(i)$ and a straight line regression is applied. From the slope m, using (RBB-3c), we get:

$$x_0 = m / 2$$

and from the y-intersect b using (RBB-3b):

$$r = \sqrt{(b + x_0^2)}$$

Finally the left circle intersection with the x-axis (for the frequency -> infinite) for our rBB'-extraction is:

$$r_{BB'} + R_E = x_0 - r$$

After all these pre-considerations, we are now able to generate the trace of RBB out of the measured S-parameters. This means we are now ready to consider the formula for RBB in the Gummel-Poon model:

The equation:

The nonlinear Base resistor is described in the Gummel-Poon model as:

```
                                         tan(z)  - z
rBB'  = RBM  + 3 [RB - RBM] -------------
                                           z tan^2(z)
```

$$rBB' = RBM + 3\,[RB - RBM]\,\frac{\tan(z) - z}{z \tan^2(z)} \qquad \text{(RBB-4)}$$

with

$$z = \frac{\sqrt{1 + \left(\frac{12}{PI}\right)^2 \frac{iB}{IRB}} - 1}{\left(\frac{24}{PI^2}\right)\sqrt{\frac{iB}{IRB}}}$$

see model equations (M) and (N) of the introducion chapter.

Fig.RBB-8 shows the plot of this equation:

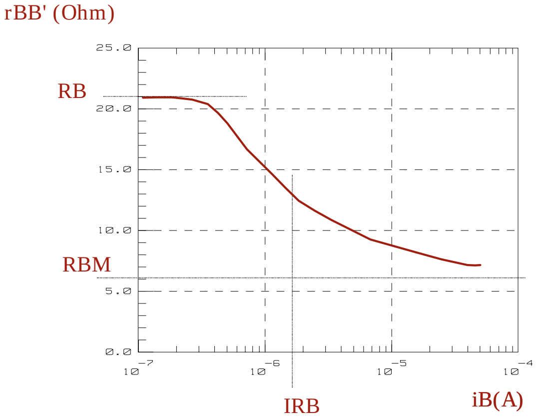


Fig.RBB-8: Base resistor curve as implemented in the model

This means: after we got $r_{BB'}$ from the measurement, we now have to fit the model curve from fig.RBB-8 to the measured data of fig.RBB-7 (after subtraction of R_E).

How to proceed:

When i_B -> 0 then z -> 0 and therefore

$$\frac{[\tan(z) - z]}{z \tan^2(z)} \rightarrow 1/3$$

We get from (53) solved for RB:

RB	$= r_{BB'}$	$[\, i_B \rightarrow 0 \,]$,	zero bias Base resistance

When i_B -> infinite then z -> PI/2 and

$$\frac{3\,[\tan(z) - z]}{z \tan^2(z)} \rightarrow 0$$

what gives from equation (RBB-4) solved for RBM:

RBM	$= r_{BB'}$	$[\, i_B \rightarrow \text{infinite} \,]$,	min.Base resistance at hi current

Finally when $i_B = I_{RB}$ then z = 1,21. Thus

$$r_{BB'} = R_{BM} + [R_B - R_{BM}]\; 0{,}51 \quad , \quad (R_B + R_{BM}) / 2$$

That's why

IRB	is the current where the Base resistor is half its max.value
	$R_{BM} + (R_B - R_{BM}) / 2$

WHAT TO DO IN IC-CAP:

first of all, the network analyzer has to be calibrated.

You should then remeasure your SHORT-OPEN-LOAD-THRU calkit standards using the IC-CAP DUTs 'CAL_xxx' for documentation.

Then, an OPEN structure on the wafer, resp. an empty package, has to be measured for the de-embedding of the inner transistor from the outer parasitics. This is both done by IC-CAP DUT 'nwa_meas'. Here, the setup 'dummy_open' is used to measure the OPEN structure, while setup 'freq_n_bias' is used to measure *all* data for the HF modeling. The de-embedding is then performed in setup 'de_embed' and transform 'S_deemb'. PLease note that the stimuli must be identical in all these setups, i.e. the same 'freq', 'ib' and 'vc' Inputs must be used.
Finally, the de-embedded data are exported from this setup and re-imported partially (depeding on the required sub-data for the individual modeling steps) into the setups of DUT 'nwa_extr'.

<u>after the de-embedding, the modeling steps are:</u>

- open setup "/gp_classic_npn/nwa_extr/rbb",
- import the data from the .mdm file,
- simulate the setup with the so far determined DC and CV parameters
- perform transform "calc_RBB" to convert the measured S-parameters to rBB'
  (set model variable DEMO=0 in order to obtain the check of the upper frequency limit for the rBB extraction)
- check the plot "rbbvsib" (rbb versus ib)
- perform transform "e_RB_IRB_RBM"
- simulate with the extracted parameters.

- optimize after you are finished with the TFF parameter extractions.
  (see proposed optimization sequence given there).

see also transform READ_ME

NOTE: As you might experience, it can be quite complex to obtain a reasonable S11 plot from which a $r_{BB'}$ curve like that one in fig.RBB-8 can be derived. If despite all of these efforts the transformed measured data do not match the curve of fig.RBB-8, set RB=RBM and model the Base resistor bias independent.

Note: avoid thermal self-heating (esp.when measuring packaged devices). This will show up if the fitting of the forward Gummel plot of iB for high vBE becomes worse when the fitting of the S11 plot is improving during fine-tuning of RE and RB. If this occurs, reduce the bias for both the rBB and forward Gummel setup. If you need these high bias values, consider using pulsed measurements (DC bias pulse width around 1us).

TRANSIT TIME MODELING

CONTENTS:

Modeling the Diffusion Capacitor CDBC

Section 1: Extraction of TF, ITF, and XTF

T_F ideal forward transit time
X_{TF} coefficient for bias dependence of T_F
I_{TF} high-current parameter for effect on T_F

For forward active operation of the transistor, the AC behavior is modeled by C_{BC} and C_{BE} (see equations O ... S in the introduction chapter). In this operating mode, the already CV-modeled C_{SBC} dominates over C_{DBC} in equ. (P), while in equ.(R), the more important term is C_{DBE}. This chapter covers the modeling of C_{DBE}.

C_{DBE} is described by the bias-dependent transit time T_{FF} in equ.(R), and T_{FF} is modeled with the formula:

$$T_{FF} = T_F \left\{ 1 + X_{TF} \left[\frac{i_f}{i_f + I_{TF}} \right]^2 \exp\left(\frac{v_{BC}}{1{,}44\, V_{TF}} \right) \right\}$$

see (S)

with the ideal forward Base current i_f from equation (C).

T_F is the ideal forward transit time modeling the 'excess charge'. The parameters X_{TF}, and I_{TF} cover the operating point dependence from the DC bias $i_C \sim i_f$, while V_{TF} describes the dependence from $v_{CB} \sim v_{CE}$.

Preconsiderations concerning the measurement:

Like in the previous chapter, the parameter estimation is again performed using a simplified model, whereas the parameter fine-tuning is finally done during an optimizer run using the full set of SPICE model equations.
Referring to appendix B, it can be shown that the transistor's h_{21}(f)-parameter behaves frequency wise like a low-pass filter with the transfer function

$$\underline{h}_{21}(f) = ß \frac{1 - p / p_{01}(i_C, v_{CE})}{1 + p / p_{p1}(i_C, v_{CE})} \quad \text{with} \quad p = j * 2PI * f$$

Typically, there is p01 > pp1. Therefore we can neglect the zero p01 against the pole pp1, and the transit frequency for $|h_{21}(f)| = 1$ is simply

$$f_{T\ 1-pole}(i_C, v_{CE}) = \frac{1}{2 * PI * T_{FF}(i_C, v_{CE})}$$

(see (11) of appendix B)

or inverted:

$$T_{FF}(i_C, v_{CE}) = \frac{1}{2 * PI * f_{T\ 1-pole}(i_C, v_{CE})} \qquad \text{(TFF-1)}$$

where $f_{T1-pole}$ is a function of the bias current i_C and the bias voltage v_{CE} .

Note:

In many publications, like e.g. /Sinnesbichler p.106/, it is mentioned that the transit time after equ.(TFF-1) is

$$T_{FF}(i_C, v_{CE}) = \frac{1}{2 * PI * f_{T\ 1-pole}(i_C, v_{CE})} - RC * CBC$$

In this case, the TFF used for modeling is RC*CBE smaller than the value converted from fT.

In some other publications, this formula is extended to

$$T_{FF}(i_C, v_{CE}) = \frac{1}{2 * PI * f_{T\ 1-pole}(i_C, v_{CE})} - (RC + RE + RB / ß) * CBC$$

or after /B.Ardouin, p.198/

$$T_{FF}(i_C, v_{CE}) = \frac{1}{2 * PI * f_{T\ 1-pole}(i_C, v_{CE})} - (CBE + CBC) \cdot \frac{v_T}{i_C} - (RCX + RE) \cdot CBC$$

In practice, however, with the goal of a direct extraction of the TFF parameters followed by a post-optimization, the additional terms can be neglected and the simple equation (TFF-1) is sufficiently correct.

Now back to the parameter extraction:
We will first consider the extraction of the parameters T_F, I_{TF} and X_{TF}. V_{TF} will be covered later. This means, a special measurement for $f_T(i_C)$ and later for $f_T(i_C, v_{CE})$ is needed. As we assume a 1-pole low-pass for h_{21}, the gain-bandwidth product is a constant. Therefore it is sufficient to measure a $h_{21}(i_C, v_{CE})$ at a fixed frequency higher than the the -3dB frequency.

In other words, this fixed frequency should be from a -20dB/decade range of MAG[h_{21}(freq, i_C, v_{CE})] . This measurement frequency can be found when transforming the measured $r_{BB'}$ S-parameters to H-parameters (using the TwoPort function). From the dB-plot of ABS(h_{21}(f)) versus log(frequency) we determine a frequency where the slope fits a -20dB/decade roll-off.

NOTE: if your MAG[h21] does not follow the -20dB/decade law, there is probably a so-called over-deembedding. This means more parasitics are subtracted than present in reality. The opposite de-embedding problem, under-deembedding, does not affect the slope, but it can show up like a second -20dB/decade slope shifted in frequency.

This frequency is now used as a fixed frequency f_{-20dB} for the S-parameter measurements of this setup. The underlying DC bias values are a swept i_C and a constant and small value of v_{CE} (to neglect the VTF effects). Then, these S-parameters are converted into H-parameters and we get for the constant gain-bandwidth product of this assumed one-pole low-pass filter:

$$1 * f_{T\,1-pole}(i_C, v_{CE}) = |\underline{h}_{21}(i_C, v_{CE})| * f_{-20dB}$$

or

$$f_{T\,1-pole}(i_C, v_{CE}) = |\underline{h}_{21}(i_C, v_{CE})| * f_{-20dB} \qquad \text{(TFF-2)}$$

$f_{T1\text{-}pole}$ after equ.TFF-2 is valid for all DC bias conditions, i.e. for the whole bias-dependent array of h_{21}. This new array $f_{T1\text{-}pole}$ is then introduced into (TFF-1), what gives the bias-dependent array of T_{FF} to be fitted.

Fig.TFF-1 shows log|h_{21}| as a two-dimensional function of the Collector current i_C and the frequency freq. It shows the transit frequency with and without simplification (Appendix B). The dependence of v_{CE} is neglected for simplification.

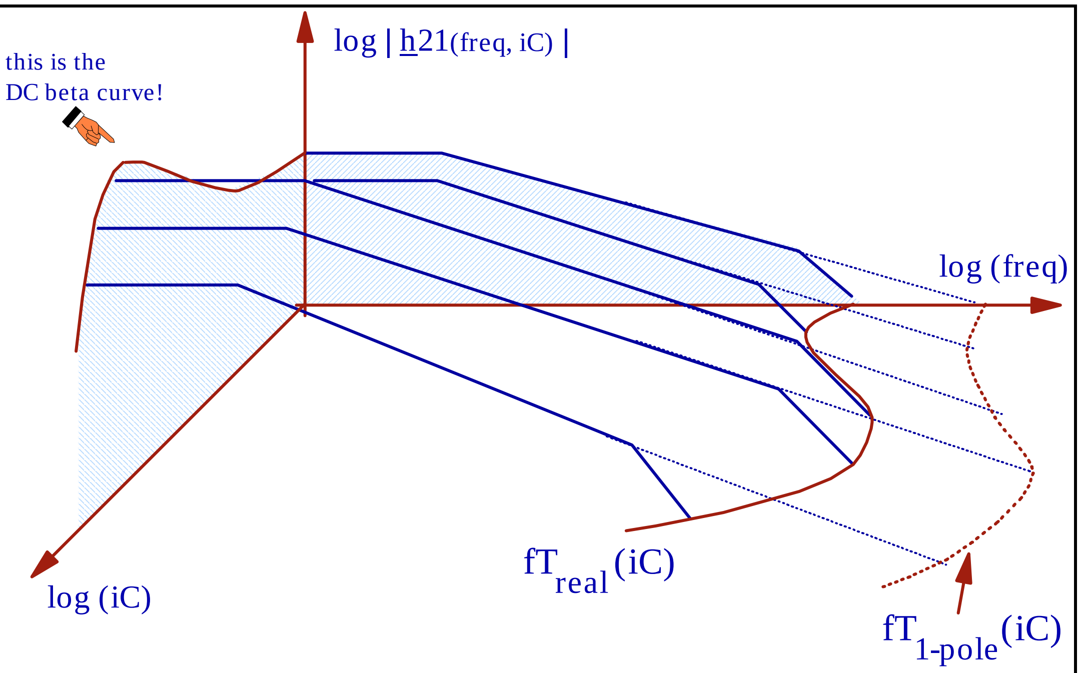


Fig.TFF-1: log |h21| as a function of log (i_C) and log(f) for v_{CE}=const.

Usually, f_T is plotted against i_C . This is the typical diagram published in many data sheets. Fig.TFF-2 shows such a curve, also including the dependence of f_T from v_{CE} .

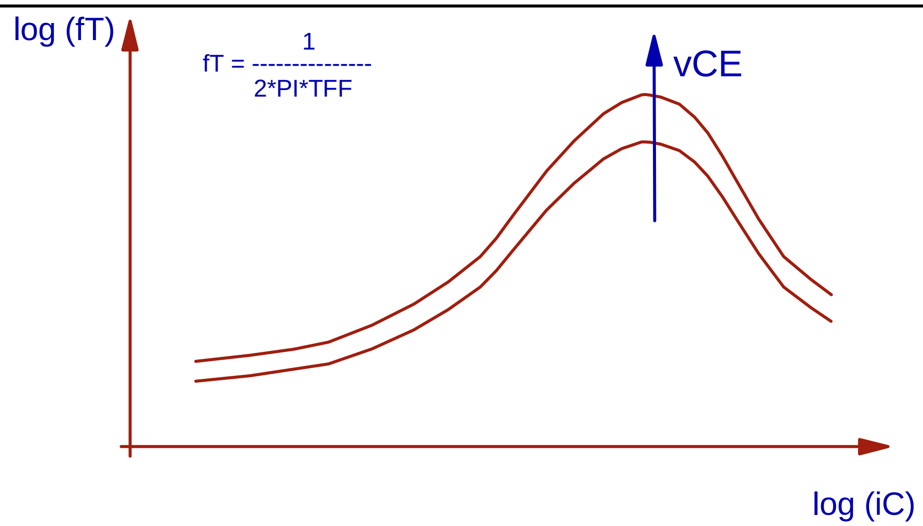


Fig.TFF-2: f_T as a function of i_C and v_{CE}

Note: for a correct modeling, check the f_T curve at low i_C for so-called self-biasing! This effect occurs if the RF signal power at the Base is in the range of the DC bias power. Under this condition and considering the non-linear diode characteristic at the Base of the transistor, the rectified AC signal will contribute to the DC bias! A flat trace of the f_T curve at low Collector current is an indicator for that effect.

For more detailed examples about how the RF power level might affect the f_T curve, see literature P.v.Wijnen, chapters 3 and 4, and the IC-CAP examples on non-linear RF modeling, available from the author.

Preconsiderations concerning the model equation:

As cited at the beginning of this section, we start with:

$$T_{FF} = T_F \left\{ 1 + X_{TF} \left[\frac{i_f}{i_f + I_{TF}} \right]^2 \exp\left[\frac{v_{BC}}{1{,}44\ V_{TF}} \right] \right\} \qquad \text{(TFF-3)}$$

see (S)

with the ideal Collector current i_f from equ.(C).
v_{BC} as well as v_{BE} in (C) are the DC bias voltages at the operating point.

In this equation, i_f is the ideal Collector current. If we consider currents below IKF ,we can set $i_f = i_C$. After the extraction of the parameters of this section, we will use a final optimization on the S-parameter curves, which will eliminate this small error.

Therefore the curve-to-be-fitted is:

$$T_{FF} = T_F \left\{ 1 + X_{TF} \left[\frac{i_C}{i_C + ITF} \right]^2 \exp\left[\frac{-v_{CB}}{1{,}44\ V_{TF}} \right] \right\}$$

If we choose vCB ~ 0, we can further simplify and get finally:

$$T_{FF} = T_F \left\{ 1 + X_{TF} \left[\frac{i_C}{i_C + ITF} \right]^2 \right\} \qquad \text{(TFF-4)}$$

Validity of (TFF-4): $i_C < I_{KF}$ and linear forward operating point with $v_{CB} \sim 0$ or as small as possible.

Performing the measurement:

Measurement setup:

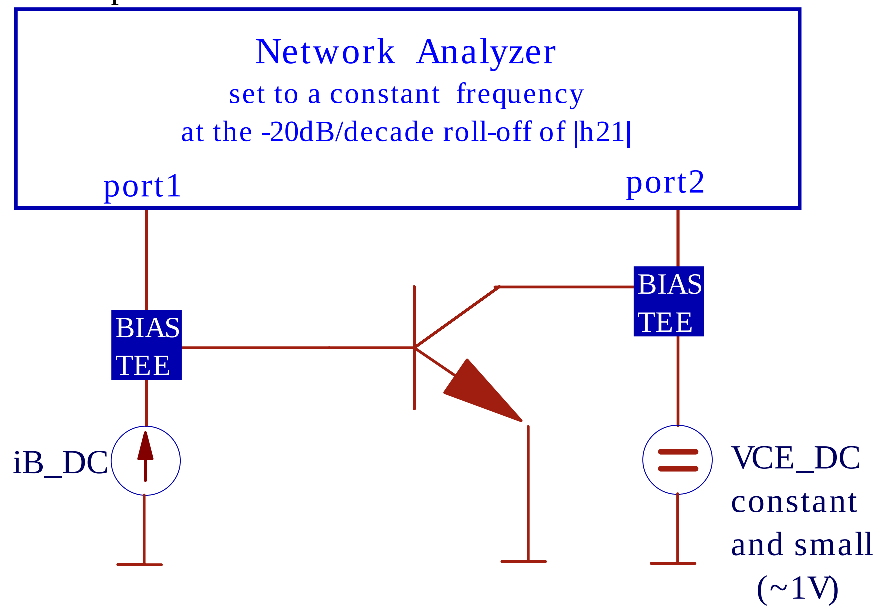


Fig.TFF-3: measurement setup for the TFF parameter measurement

First the network analyzer is set to a constant frequency on the -20dB/decade roll-off of $|\underline{h}_{21}(f)|$. The used test frequency had been estimated from the S_to_H parameters of the RBB measurement. More details were given above.

Next the transit frequency for a 1-pole low-pass model is calculated as given in (TFF-2):

$$f_{T1\text{-}pole}(i_C, v_{CE}) = |\underline{h}_{21}(i_C, v_{CE})| * f_{-20dB} \qquad \text{(see TFF-2)}$$

Then we calculate $T_{FF} = 1/(2\ PI\ f_{T1\text{-}pole})$ as the bias-dependend total transit time.

Extracting the parameters:

Figure TFF-4 highlights the theoretical trace of T_{FF} of equation (TFF-4). This figure shows the theoretical curve in addition to the measured one to make things more clear. The real measured curve is overlaid by the space charge capacitor effects for low collector currents. This can be seen also from equation (CV-1) in the chapter on CV modeling. For a detailed description of this effect, see /Berkner 1993/

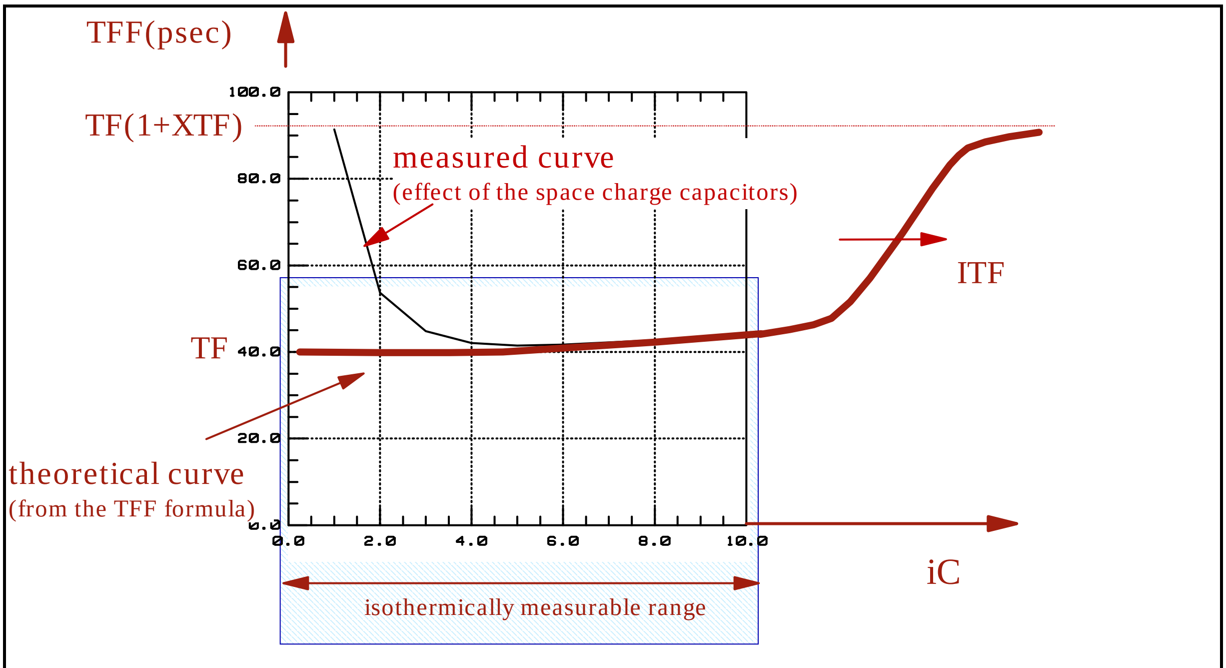


Fig.TFF-4: The theoretical transit time T_{FF} (bold) as a function of i_C for $v_{CB} = 0$ compared to the theoretical trace.

Due to these overlay and measurement problems, it had been found that a pretty simple and straight-forward extraction technique can be applied that gives nevertheless quite reasonable results. This method is explained below. There exist some more complex strategies, but the extraction results may be not much better. As sketched in fig.TFF-4, this is mainly because it is not possible to force such a high Collector current that the trace of equ.(TFF-4) can be obtained from measurements without being overlied and distorted by thermal self-heating effects.

How to proceed:

T_F

is extracted as the minimum value of T_{FF}.
Note that a prerequisite is vBC=0, i.e. select a vCE ~ 1V for the extraction, and no Collector voltage in quasi-saturation!

X_{TF}

The behavior of T_{FF} was given in fig.TFF-4. It is difficult to measure for a higher Collector current due to thermal limitations. So X_{TF} is estimated from the trace of TF at max. applicable Collector bias current under the assumption that it would be TFF at infinite current:

$$MAX(T_{FF}) = T_F (1 + X_{TF})$$

or

$$X_{TF} = \frac{MAX(T_{FF})}{T_F} - 1 \qquad \text{(TFF-5)}$$

This usually gives a pretty good first-order estimation. Due to the Collector current limitations, an estimation correction like $X_{TF}=10*X_{TF(equ.TFF-5)}$ can improve the starting conditions for the optimizer.

For more details, see the optimizer strategy at the end of this chapter.

I_{TF}

Referring to the same measurement restrictions as above, a good first-order estimation of I_{TF} is related to the max. Collector current measured:

$$I_{TF} = MAX(i_{C\_meas}) / 2 \qquad \text{(TFF-6)}$$

Again, since the end of the TFF trace is often not measurable, correct this estimation by $I_{TF}=5*I_{TF(equ.TFF-6)}$.

NOTE: this ITF extraction method follows the idea of the Base resistor IRB parameter extraction!

WHAT TO DO IN IC-CAP:

the extraction of the parameter TF, ITF and XTF is performed simultaneously with parameter VTF in a single extraction routine.

See next chapter

Part 2 - Extraction of VTF

VTF voltage describing the vBC dependence of TF

Finally, we consider also the v_{CE} sweep.

Measurement setup:

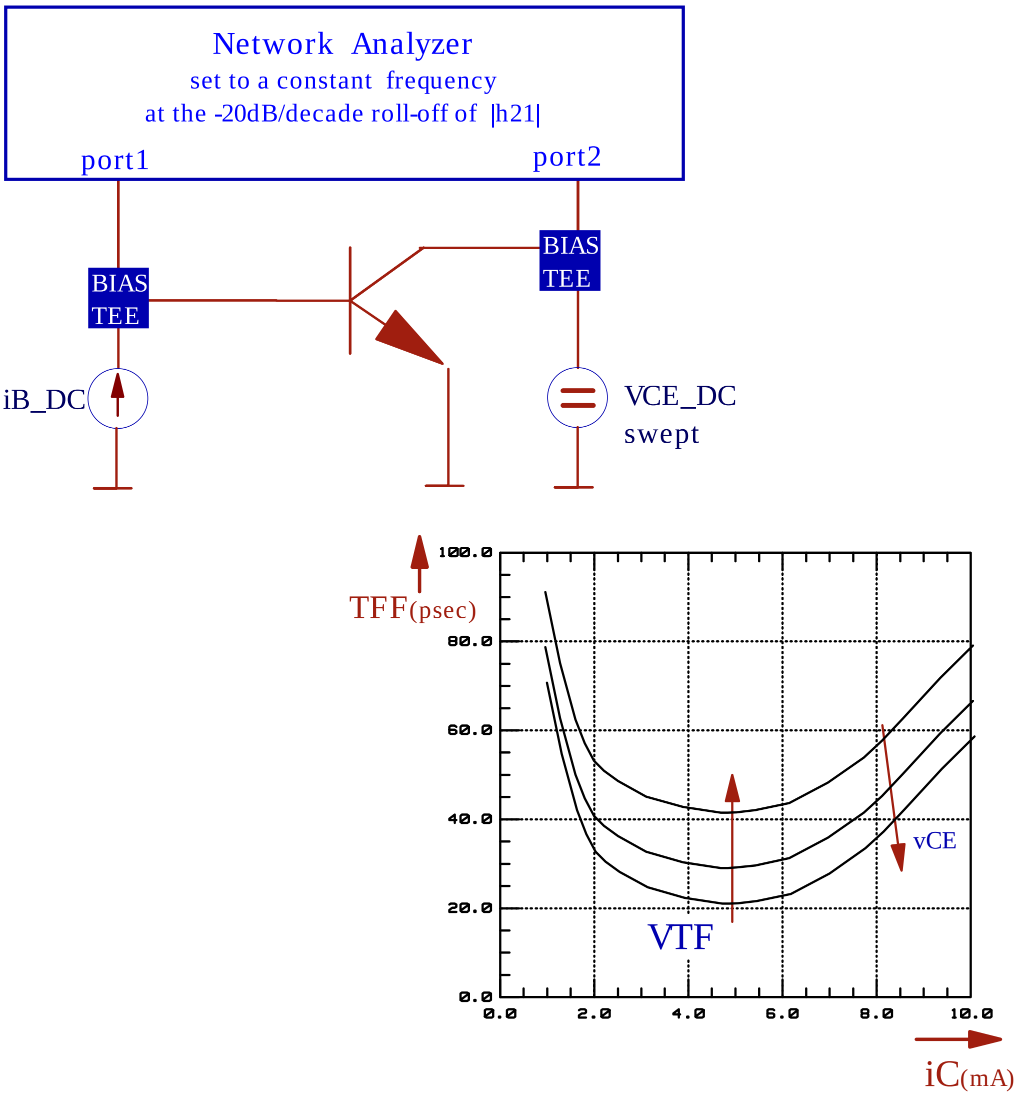


Fig.TFF-5 The transit time T_{FF} as a function of i_C and v_{CE}

V_{TF} can be obtained from (TFF-3) for a fixed value of iC:

$$TFF = \cdot const \cdot e^{\frac{-v_{CB}}{1.44 \cdot VTF}}$$

or

$$\frac{T_{FF1}}{T_{FF2}} = \frac{\exp[-v_{CB1} / 1,44\ V_{TF}]}{\exp[-v_{CB2} / 1,44\ V_{TF}]} = \exp\left[\frac{v_{CB2} - v_{CB1}}{1,44\ V_{TF}}\right]$$

This gives:

$$\ln\left[\frac{T_{FF1}}{T_{FF2}}\right] = \frac{v_{CB2} - v_{CB1}}{1,44\ V_{TF}}$$

and finally:

$$VTF = \frac{v_{CB2} - v_{CB1}}{1,44 * \ln\left(\frac{TFF_1}{TFF_2}\right)} \qquad \text{(TFF-7)}$$

After the extraction of these four parameters for C_{DBC} , we will next run an optimization to improve the fitting of the f_T plot. However, it is very important, that we do not forget to also optimize the S-parameter fittings for all bias conditions after that (setup rbb).

Notes:

From the pre-considerations given above, the plots RBB vs. iB and TFF vs. iC represent curves which had been *extrapolated* from the S-parameters. So, the S-parameter measurement is "the real world" and the fitting should be optimized in this domain !

Again, all the AC extraction methods need absolute clean measurements and elimination of parasitics by de-embedding techniques. Otherwise no curve fitting might be possible or the parameters obtained might make no physical sense.

WHAT TO DO IN IC-CAP:

open setup "/gp_classic_npn/nwa_extr/tf_ib_vcb",

select a -20dB/decade frequency from the plot 'mag_h21vsf' of setup 'rbb' (middle mouse click) and enter it to model variable F20dB

import the de-embedded data of this -20dB/decade frequency into setup 'tf_ib_vcb' from the exported .mdm file of setup 'nwa_meas/de_embed'

perform transforms "S_to_H", "calc_ft", "calc_TFF" and "ic"

check the plots "ftvsic" and "TFF" (TFF versus calculated Collector current i_C)

perform transform "e_TF_ITF_XTF_VTF"

and simulate with the extracted parameters

for optimization, see below.

HOW TO OPTIMIZE THE S-PARAMETER SETUPS IN IC-CAP:

for the data supplied with the file 'gp_classic_npn.mdl',
this S-parameter sequence is best:

1.) in setup "/gp_classic_npn/nwa/tf_ib_vcb", run the optimizer transforms "o_TF_ITF_XTF" to optimize for the *smallest* v_{CB} bias.,
then, execute "o_VTF" to opotimize the fitting for the *biggest* v_{CB} bias
finally, call "o_TF_ITF_XTF_VTF to optimize *all* bias points"

2.) go back to setup "rbb" and optimize the S-parameter fitting of the RB, IRB and RBM setup by running "o_RB_IRB_RBM"

3.) in the same setup, execute finally "o_TFF", what optimizes all TFF parameters at all bias conditions of setup "rbb".

see also macro "extract_n_opt_NWA"

NOTE: Depending on the device type, it has been observed that the f_T-fitting affects also the r_{BB}-fitting.

Extraction of PTF

PTF excess phase at frequency 1/(2PI * TF)

Implemented into the model as a 2nd order all-pass Bessel-function, this parameter can be used to add some extra phase to the RF simulation curves. It can be obtained when plotting the phase of h_{21} of the TFF measurement from above (fig.TFF-5)versus v_{BE} /T.Zimmer/. As a limitation of the method, the measurable range is again only covering a small part of the desired curve. Therefore it is advisable to use an optimizer run in order to get the value of P_{TF}.

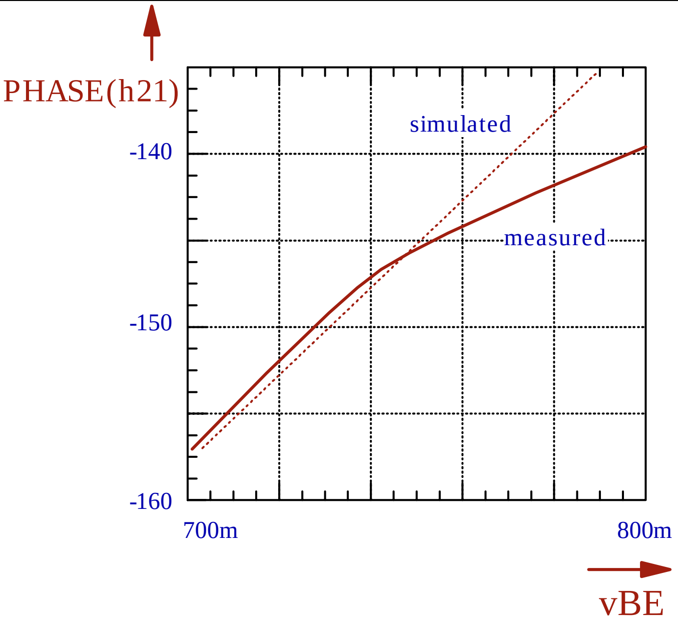


Fig.TFF-6: Measured and simulated phase(h21) before optimization.

Extraction of TR

TR ideal reverse transit time

The reverse transit time of the Gummel-Poon model is modeled by one parameter only, T_R.

T_R can be obtained from an h_{21} measurement versus frequency at a typical reverse operating point. An optimizer run on the S-parameters of this setup is used to obtain the parameter value from a typical starting value, e.g. T_F * 100.

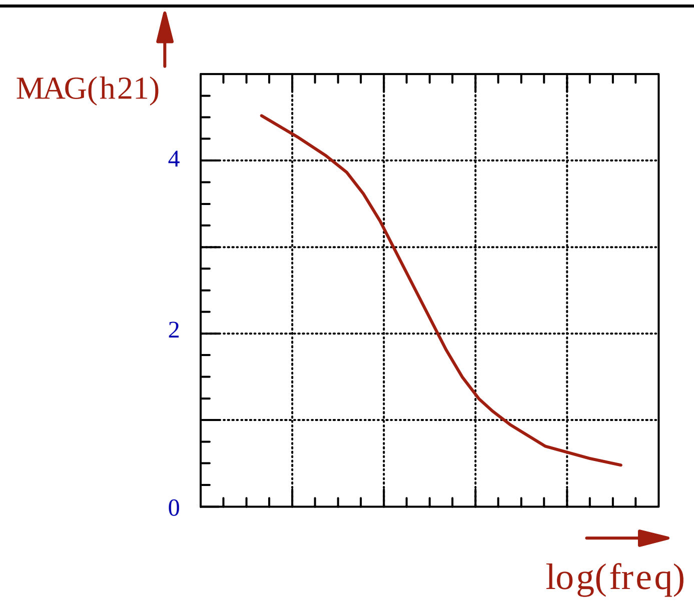


Fig.TFF-7: The reverse transit time T_R.

NOTE:
After /Sinnesbichler/, TR can be optimized in the S22 and S12 plot of the reverse biased S-parameters.

NOTE: the reverse biased S-parameter can be measured for these example DC bias levels:

1st order DC sweep: `vCE = -1 ... -5V`

2nd order DC sweep: `vBE = -0,7 ... -1V`

NOTE: T_R can also be obtained from pulse measurements using an oscilloscope. In many cases, T_R is *the* dominant parameter in such a setup, often more important than TFF and the junction capacitances. Therefore, for the modeling of transistors in digital applications, the T_R modeling is a must.

MODELING OF THE PARAMETER XCJC

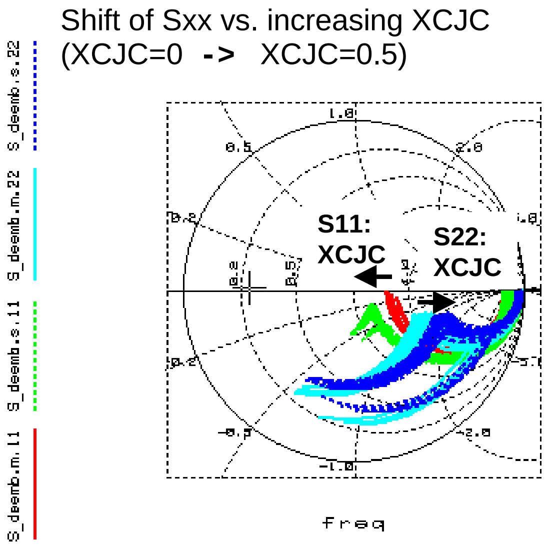
This parameter distributes the CCB junction capacitance between the inner and the outer Base contact. Its default value is XCJC=1, meaning the CCB capacitance is tied completely to the inner Base. For XCJC=0, the capacitance is between the outer Base and the Collector.

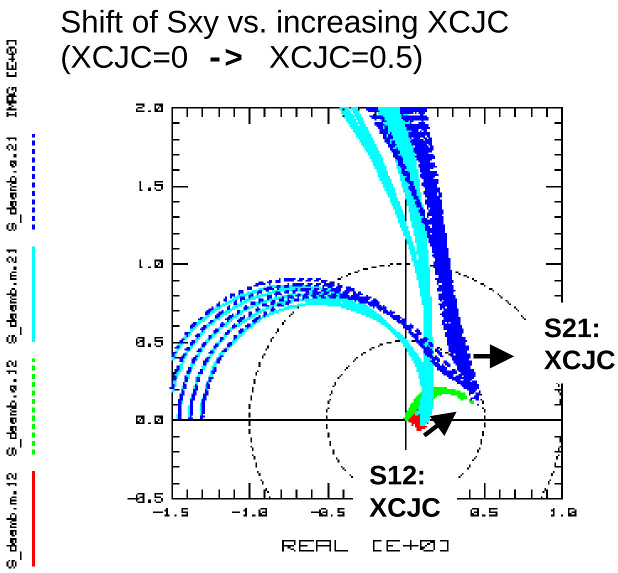
This parameter is difficult to determine from CV measurements. However, if the geometry of the device is known, it can be calculated pretty easily after

XCJC = 1 - AE/AB,

where AE is the Emitter area, and AB the total area of the Base, including the Emitter area.

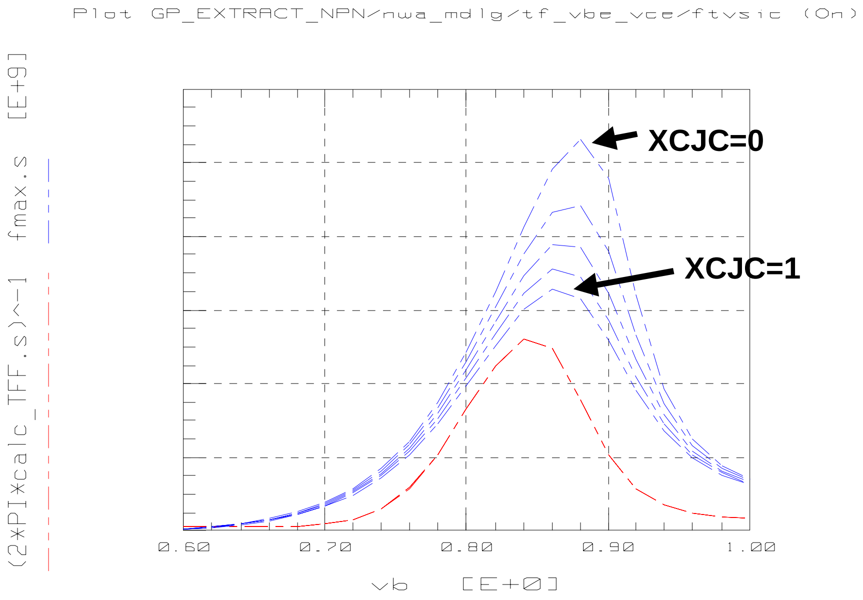
Usually, this parameter is fine-tuned after the other HF parameters have been determined. The following S-parameter figures show the effect of XCJC for device modeling.





Generally speaking, if S12 becomes 'big' for high frequency, it is either Re or XCJC!

Note: fmax may also be used to model the effect of XCJC, as depicted below:
Although there is no efect of XCJC on ft, fmax is heavaily affected.



A final remark on the S-Parameter Extraction and Optimization Strategy

When you run into problems when fitting both, the transformed rBB' and TF curves, you should try to optimize what you have measured, i.e. S-parameters rather than the rBB' or the ft and TF plots. This is the real world and the fitting there might be more important than the fitting of the rBB' plot with all its limitations (extrapolated S_{11} at infinite (!) frequency) or the f_T plot (again extrapolated from H_{21} from S-to-H parameter transformation).

Of course, the best modeling result is a good fit in all domains, the S-parameters and the transformed rBB' and TFF curves.

MODELING OF THE TEMPERATURE EFFECTS

The parameters given below are modified when the selected simulation temperature TEMP is different from the extraction temperature TNOM. (Temperatures in 'K).

used auxiliary variables:

$$V_T = \frac{k\,TEMP}{q}$$

$$E_G = 1.16 - \frac{7.02\,E\text{-}4\,TEMP^2}{TEMP + 1108}$$

$$N_i = 1.45\,E10 \left(\frac{TEMP}{TNOM}\right)^{1.5} \exp\left[\frac{q}{2k}\left(-\frac{E_G}{TEMP} + \frac{1.1151}{TNOM}\right)\right]$$

temperature dependant modeling parameters:

$$I_S(TEMP) = I_S(TNOM) \left(\frac{TEMP}{TNOM}\right)^{X_{TI}} \exp\left[\frac{E_G}{V_T}\left(\frac{TEMP}{TNOM} - 1\right)\right]$$

$$B_F(TEMP) = B_F(TNOM) \left(\frac{TEMP}{TNOM}\right)^{X_{TB}}$$

$$B_R(TEMP) = B_R(TNOM) \left(\frac{TEMP}{TNOM}\right)^{X_{TB}}$$

$$I_{SE}(TEMP) = I_{SE}(TNOM) \left(\frac{TEMP}{TNOM}\right)^{-X_{TB}} \left[I_S(TEMP) / I_S(TNOM)\right]^{1/N_E}$$

$$I_{SC}(TEMP) = I_{SC}(TNOM) \left(\frac{TEMP}{TNOM}\right)^{-X_{TB}} \left[I_S(TEMP) / I_S(TNOM)\right]^{1/N_C}$$

$$V_{JE}(TEMP) = V_{JE}(TNOM) \left(\frac{TEMP}{TNOM}\right) + 2\,V_T \ln\left[\frac{1.45\,E10}{N_i}\right]$$

$$V_{JC}(TEMP) = V_{JC}(TNOM) \left(\frac{TEMP}{TNOM}\right) + 2\,V_T \ln\left[\frac{1.45\,E10}{N_i}\right]$$

Limitations of the Gummel-Poon Model

OHMIC EFFECTS:

The Collector and Emitter resistance parameters are constant and not functions of current or voltage. They have no temperature coefficients.

FORWARD DC MODELING

The parameter I_{KF} models the begin of the decrease in beta. As a limitation of the model the slope of ß above the knee current I_{KF} has a fixed value of "-1" on a log-log scale. However, this is most often overlaid by RE.

The modeling of the saturated region in the output characteristics ($V_{CE} < 0.5V$) lacks of specific parameters. Therefore the model cannot cover modern transistors in this range (quasi-saturation).

No reverse breakdown effects are included in both Base-Collector and Base-Emitter diode.

REVERSE DC MODELING:

The reverse DC modeling suffers from a separate parameter I_S. Thus N_R sometimes has to be mis-used for better fitting the reverse i_E versus v_{BC} plot.

Like in the forward region, the slope of ß above the knee current I_{KR} has a fixed value of "-1" and also the output characteristics saturation range is modeled inflexible.

AC MODELING:

The TF modeling, especially versus vCE, is not physical and often not accurate
The TR parameter is not a function of current or voltage like TFF.

TEMPERATURE MODELING

The TNOM value of VJE, VJC and VJS must be greater than 0,4V to insure convergence for temperature analysis up to 200'C.

APPLICATIONS IN INTEGRATED CIRCUITS

There is no parasitic transistor inncluded in the model

Conclusion: disregarding these limitations, the Gummel-Poon model is a good compromise between accurate modeling and a limited amount of parameters. It is still very useful especially when enhancing it with external parasitics like inductors, parasitic diodes or lateral pnp transistors.

Currently, new models like the VBIC or the HiCum come more and more into play.
Ask the author for the corresponding toolkits.

A P P E N D I C E S

C O N T E N T S:

LINEAR CURVE FITTING: REGRESSION ANALYSIS

IC-CAP File:
$ICCAP_ROOT/examples/demo_features/4extraction/basic_PEL_routines/1fit_lin.mdl

Let's assume we made 'N' measurements y_i at the stimulating points x_i. I.e. we obtained the array $\{x_i , y_i\}$. Subsequently, these measured values were plotted.

A curve Y(x) shall be fitted to this array of measured data points using least square curve-fitting technique. Referring to an individual measurement point, the fitting error is:

$$E_i = Y(x_i) - y_i \qquad (1)$$

and for all data points:

$$E = \sum_{i=1}^{N} E_i^2 = \sum_{i=1}^{N} \left[Y(x_i) - y_i \right]^2 \qquad (2)$$

This error shall be minimized.

The fitting will be done by varying the coefficients of the fitting curve of equation (2). The minimum of the total error E depends on the values of these coefficients. This means, we have to differentiate E partially versus the curve coefficients and to set the results to zero. We obtain a system of equations, solve it, and get the values of the coefficients for a best curve fit. This is known as regression analysis.

NOTE: This regression analysis is simple for a straight line fit. But in general, measured data is non-linear. Unfortunately, a non-linear regression analysis can be quite complicated. This problem can be solved if we use a suitable transformation on the measured data. This means that the measured data is transformed to a linear context between the y_i- and the x_i-values. As will be seen in the diode example later, this is a pretty smart way to get the curve fitting parameters easily without much calculations.

Provided we have got an array of N measured data points of the form $\{x_i,y_i\}$.
A linear curve with the equation

$$y(x) = m\,x + b \qquad (3)$$

shall be fitted to these points. This situation is depicted below.

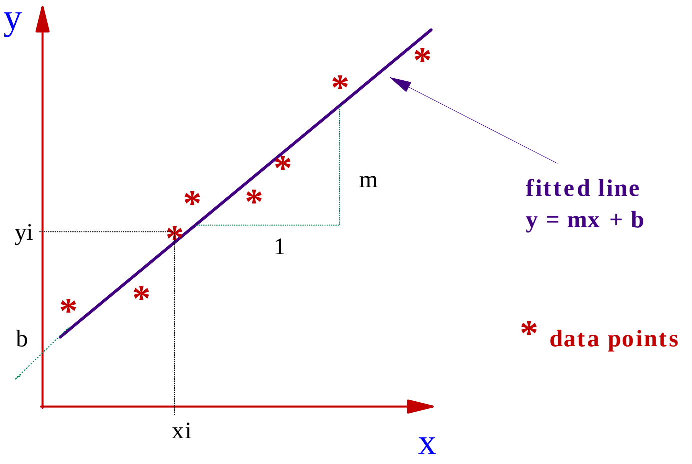


Fig_1: linear regression applied to measurement points

The error of the i-th measurement is:

$$E_i = [\ m\ x_i\ +\ b\ ]\ -\ y_i \qquad (4a)$$

Using the least means square method following equ.(2) yields:

$$E = \sum_{i=1}^{N} {E_i}^2 = \sum_{i=1}^{N} [\ m\ x_i\ +\ b\ -\ y_i\ ]^2 \overset{!}{=} \text{Minimum} \qquad (4b)$$

Partial differentiation versus slope 'm' gives:

$$2 \sum_{i=1}^{N} [\ m\ x_i\ +\ b\ -\ y_i\ ]\ x_i = 0 \qquad (5)$$

and versus y-intersect 'b':

$$2 \sum_{i=1}^{N} [\ m\ x_i\ +\ b\ -\ y_i\ ] = 0 \qquad (6)$$

We obtain from (5) after a re-arrangement:

$$m \sum_{i=1}^{N} {x_i}^2 + b \sum_{i=1}^{N} x_i = \sum_{i=1}^{N} y_i\ x_i \qquad (7)$$

and from (6):

$$m \sum_{i=1}^{N} x_i + N\ b = \sum_{i=1}^{N} y_i \qquad (8)$$

Multiplying (7) by -N and (8) by $\Sigma\ x_i$ and adding these two equations allows the elimination of the coefficient 'b', and we can separate the slope 'm':

$$m \left[\left(\sum_{i=1}^{N} x_i \right)^2 - N \sum_{i=1}^{N} x_i^2 \right] = \sum_{i=1}^{N} x_i \sum_{i=1}^{N} y_i - N \sum_{i=1}^{N} x_i y_i \qquad (9)$$

or:

$$m = \frac{\sum_{i=1}^{N} x_i \sum_{i=1}^{N} y_i - N \sum_{i=1}^{N} x_i y_i}{\left(\sum_{i=1}^{N} x_i \right)^2 - N \sum_{i=1}^{N} x_i^2} \qquad (10)$$

and from (8) for the y-intersect 'b':

$$b = \left[\sum_{i=1}^{N} y_i - m \sum_{i=1}^{N} x_i \right] / N \qquad (11)$$

with 'm' according to (10).

With equations (10) and (11), we determined the values of the two coefficients of the linear curve which fits best into the 'cloud' of measured data.

Finally, a curve fitting quality factor r^2 is defined. Its value ranges from $\{0 < r^2 < 1\}$. The closer it is to 1, the better is the fit of the linear curve.

$$r^2 = m^2 \frac{\sum_{i=1}^{N} x_i^2 - \frac{1}{N}\left(\sum_{i=1}^{N} x_i \right)^2}{\sum_{i=1}^{N} y_i^2 - \frac{1}{N}\left(\sum_{i=1}^{N} y_i \right)^2} \qquad (12)$$

with 'm' from (10)

About The Modeling Dilemma

Using IC-CAP for the extraction of model parameters offers a lot of flexibility in terms of creating user-defined models and implementing the corresponding extraction routines.

But when developping a new extraction strategy, we may run into two major problems:
do the routines extract the parameters correctly?
and is the model able to fit the measured device at all?

This appendix proposes a method that allows us to
- verify the quality of the extraction routines
- check the fitting of our model to the measured data
and to perform the parameter extraction simultaneously.

This model-fit-check method is also called **'Direct Visual Parameter Extraction'**

Both issues are pretty important in order to obtain reliable parameters and thus satisfying simulation results of the complete circuit.

Verifying the quality of extraction routines

Let's start with the first problem: assumed, we would know the parameters to-be-extracted in advance, it would be easy to check the validity of the extraction routines!

Using IC-CAP, it is simple to perform such a check. The trick is to 'synthesize' quasi-measured data out of a set of parameters and to apply then the extraction routines to these data. This can be done as follows:

1. Define a measurement setup in IC-CAP, for which the extraction routines shall be tested. Example: an output characteristic for an Early-voltage extraction.

2. Select a 'typical' set of parameters (no default values like 'zero' or 'infinite', but instead real realistic values!)

3. Change the 'Output' data type to 'S' (simulated only). The array behind that output is no longer one-dimensional, i.e.no measurement data any more, but only simulation data.

4. Simulate this setups using these parameter values.

5. Change the 'Output' data type back to type 'B'.
   IC-CAP <u>doubles</u> now the data field to measurement <u>and</u> simulation data. Thus the simulated data of step 4. is now converted to measured data!

6. Reset the model parameters by clicking 'Reset to Defaults' and simulate the setup using the default parameters.

7. Apply the extraction routine-under-test and check the quality of the extracted parameters.

Provided we get the parameter values back within a good tolerance, we can be sure that the extraction works correctly. If we now apply the extraction to real-world measured data, we should obtain the right parameters. This is true if the measured data have the same shape like the model equations! If not, we might have to choose another model or go for subcircuit modeling. And this leads us to the second part of this paper:

Direct visual parameter extraction of BF , ISE and NE

or: Checking if the model can fit the measured data at all

This task can be solved by transforming the measured data to a domain where the parameter itself can be displayed against the measurement stimulus. As an example, we know that the x-intersect of lines fitted to an output characteristic of a bipolar transistor should hit always the same point, the Early voltage. If we apply an IC-CAP PEL (parameter extraction language) program to calculcate the x-intersect of a line that is fitted to two adjacent measured points, and if we display the result of this operation versus the collector voltage (first order sweep), we will obtain a plot of the 'equivalent' Early-voltages of adjacent measurement points.

The advantage of using this method is that we can see clearly, if the model is able to fit the measured data at all. We only have to check if there is a flat region in the transformed data domain or not. If it is there, we can extract the parameter very simply by calculyting the mean value of the flat region. And we know at the same time, in which range the parameter is dominant and can therefore be used for fine-tuning with the optimizer. If there is no flat range, the model cannot fit the measured data. We could vary the parameter as much as we like and would not achieve a fit of the simulated to the measured data!

For the application of this method, we start with some basic equations that refer to figure 1:

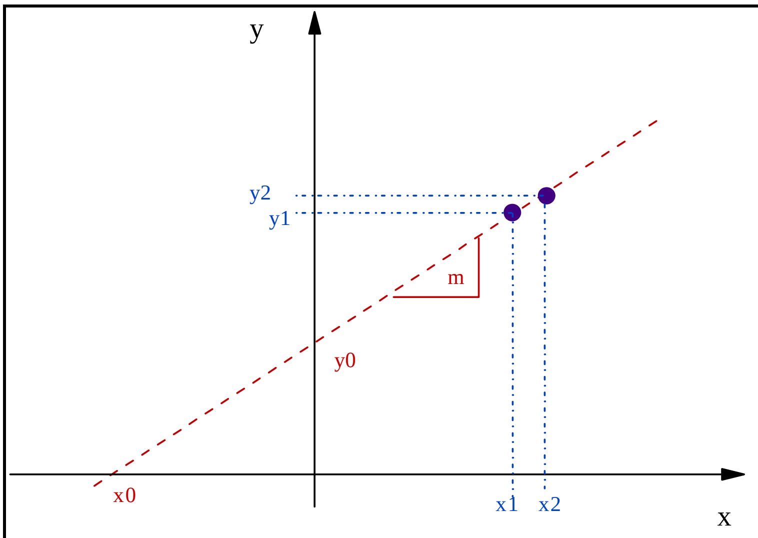


Fig_1: on the definition of some basic equations for the direct visual parameter extraction

Assumed we have

$$y = m*x + y0$$

where m: slope
y0: y-intersect

Then it is:

$$x0 = -b/m$$

and:

$$m = \frac{y2 - y1}{x2 - x1}$$

Some more usefull formulas to calculate

<u>x-intersect x0</u>

<u>y-intersect y0</u>:

starting with

$$\frac{y2}{-x0+x2} = \frac{y1}{-x0+x1} \qquad \frac{y2-y0}{x2} = \frac{y1-y0}{x1}$$

we get:

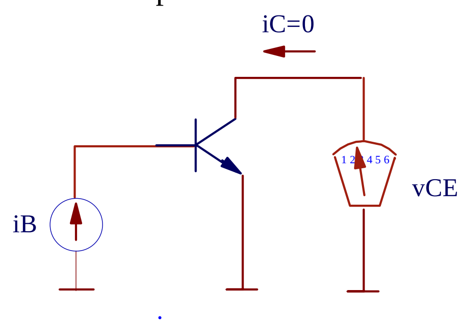
$$x0 = \frac{x1y2 - x2y1}{y2 - y1} \qquad y0 = \frac{x2y1 - x1y2}{x2 - x1}$$

These equations are implemented to the model files of directory "visu_n_extr".

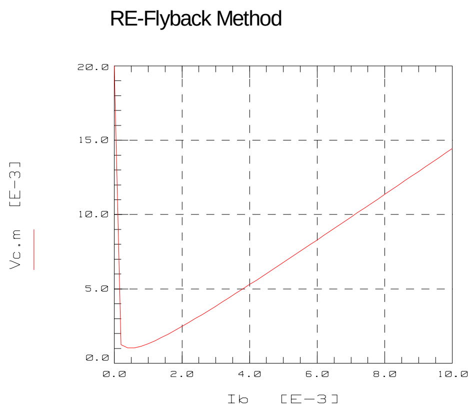
The following plots give some examples on how to apply this idea to the parameter extraction of a bipolar transistor using the Gummel-Poon model. It should be mentioned that this method can be applied to all the parameters of this model, as well as to other models like Statz, Curtice, BSIM etc.

RE

measurement setup:



$$RE = \frac{d(vC)}{d(iB)}$$



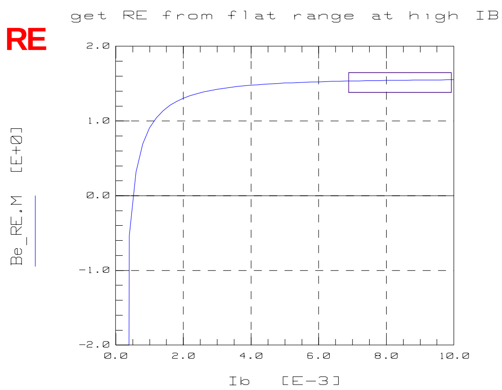
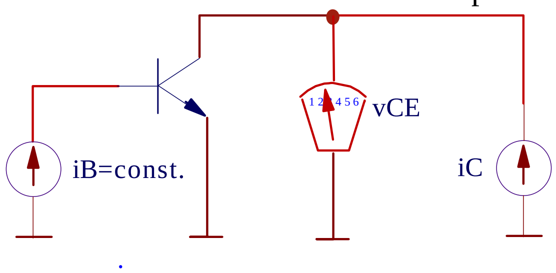


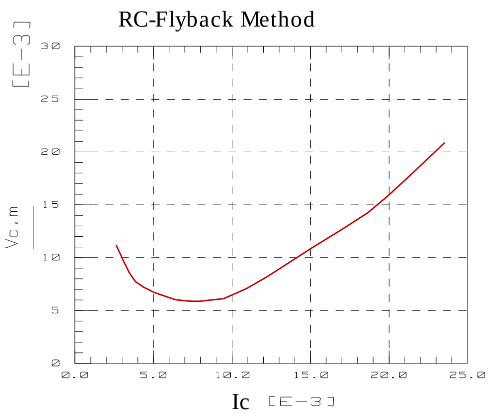
Fig_2: RE is the derivative of the flyback measurement
the "visualized" RE is the y-intersect of two adjacent data points from that flyback plot.

RC

measurement setup:



$$RC = \frac{\partial v_{CE}}{\partial i_C} - R_E$$



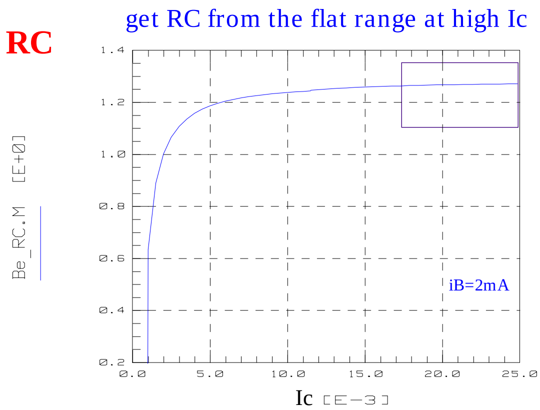


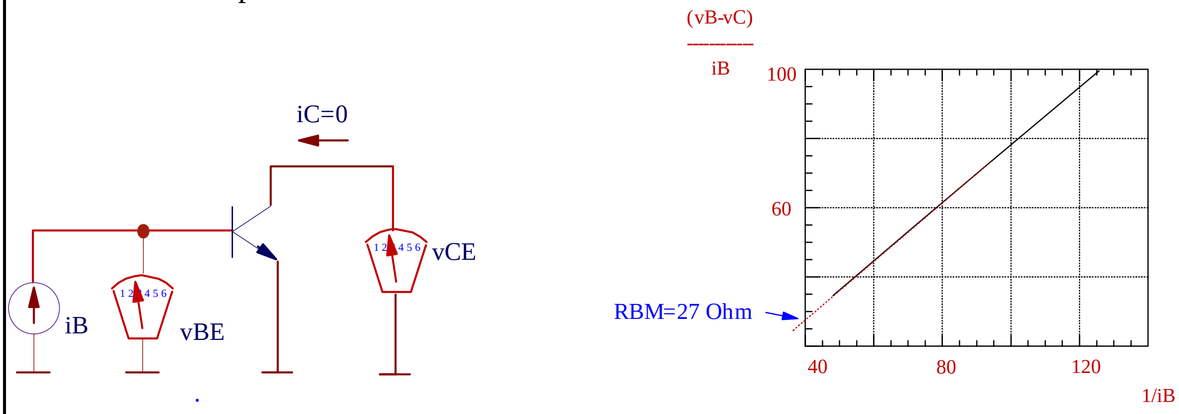
Fig_3: RC can be calculated from measured vC at stimulated iB and iC
the "visualized" RC is the y-intersect of two adjacent data points

RBM

(after the method of /Zimmer/, see fig.10 of the chapter on the DC resistor modeling):

measurement setup

transformed measurement result:



Next, the y-intersects of two adjacent data points in this transformed data plot are drawn versus 1/iB again. From the curve above, we might expect a flat range, but in reality we get a curve like the left one below: In our example, RBM ranges from 5 to 12 Ohm. In this case, we might display the visualized data versus iB, and interprete the result simular to fig.24 in the rBB chapter in order to estimate not only RBM but also IRB and RB (see plot on the right).

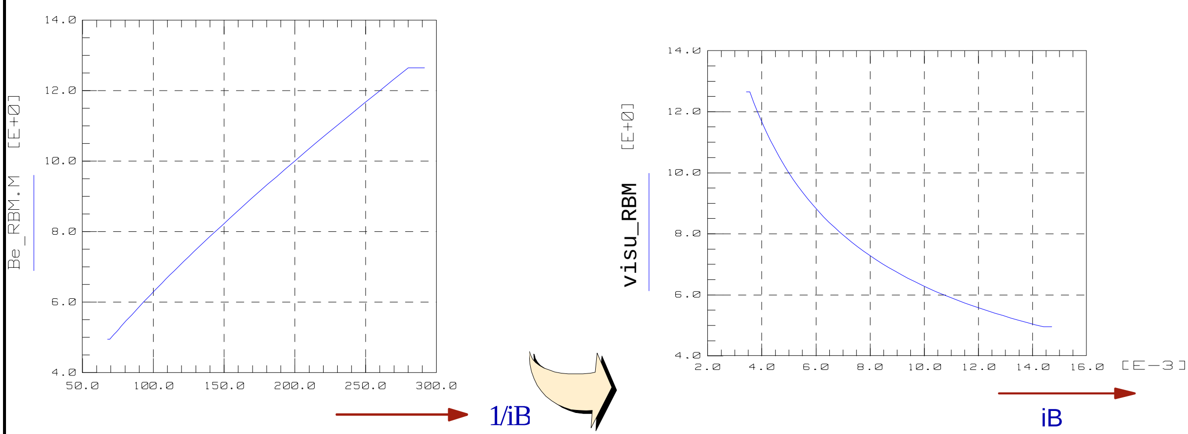
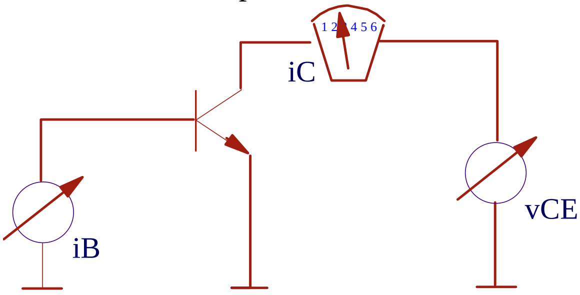


Fig_4: visual extraction of RBM and to estimate IRB and RB possibly.

VAF

measurement setup:



! Determine VAF out of the x-intersect of a line through two adjacent measurement points:

```
X=vc
Y=ic.M

i=1
WHILE i < SIZE(Y)-1
  VAF[i]=ABS(X[i+1]*Y[i-1]-X[i-1]*Y[i+1])//(Y[i+1]-Y[i-1])
  i = i + 1
END WHILE
```

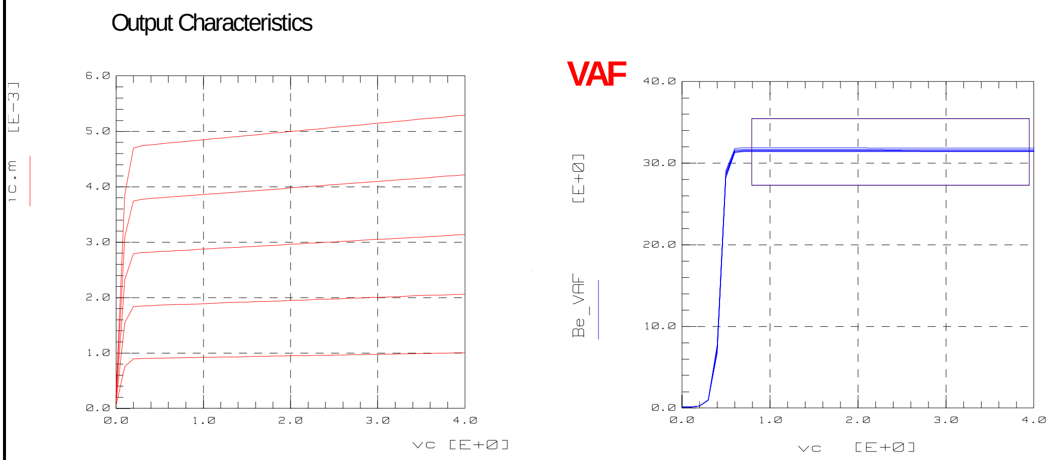
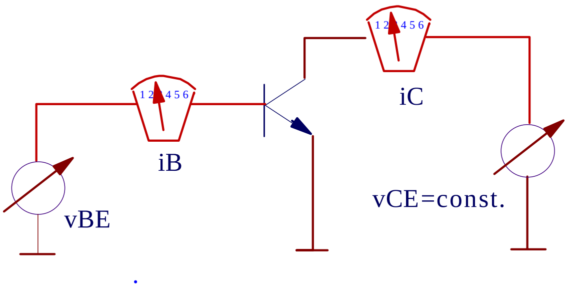


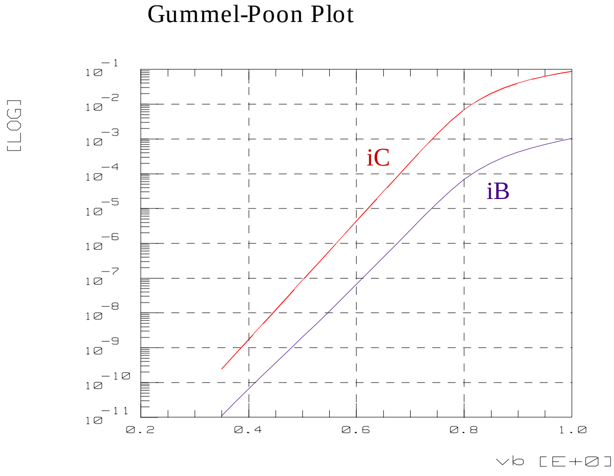
Fig_5: the Early voltage is calculated from the x-intersect of a line through two adjacent data points

Let's study also the application of the "visual" method for the Gummel-Poon plot:

measurement setup:



measurement result:



interpretation of the Gummel-Poon plot for parameter extraction:

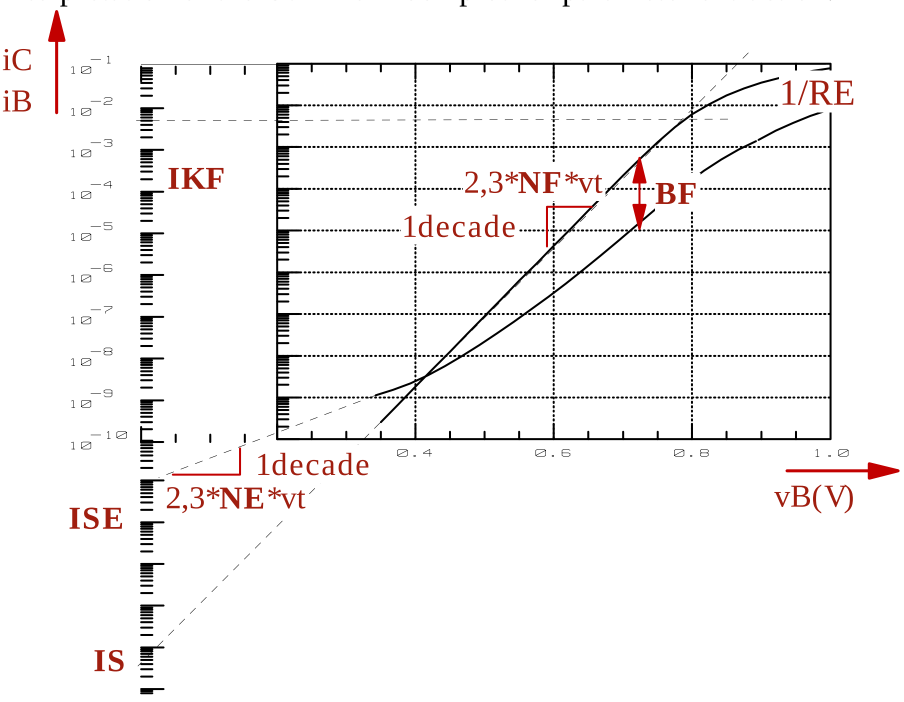


Fig_6: a typical Gummel-Poon measurement and its interpretation.

NF

$$NF = \frac{1}{VT * \frac{\partial(\ln(i_C))}{\partial(v_{BE})}}$$

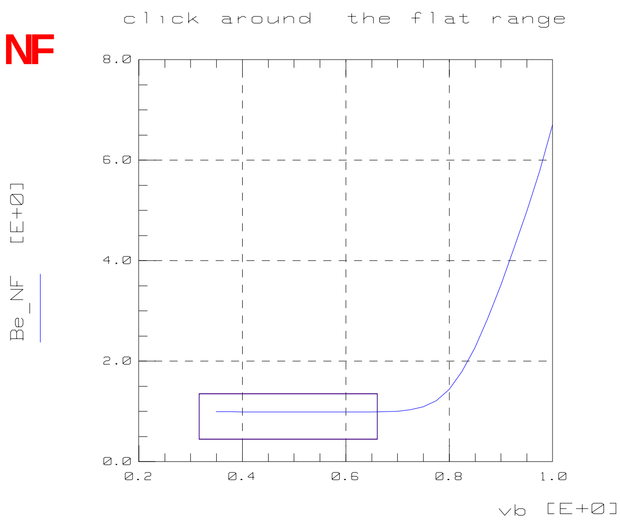


Fig_7: NF can be obtained from the inverse derivative of the iC-Gummel-plot

IS

!Determine IS out of the y-intersect of a line through two adjacent measurement points:

```
X=vb
Y=log10(ic.M)

i=1
WHILE i < SIZE(Y)-1
   ISE[i]=(X[i+1]*Y[i-1]-X[i-1]*Y[i+1])//(X[i+1]-X[i-1])
   i = i + 1
END WHILE
```

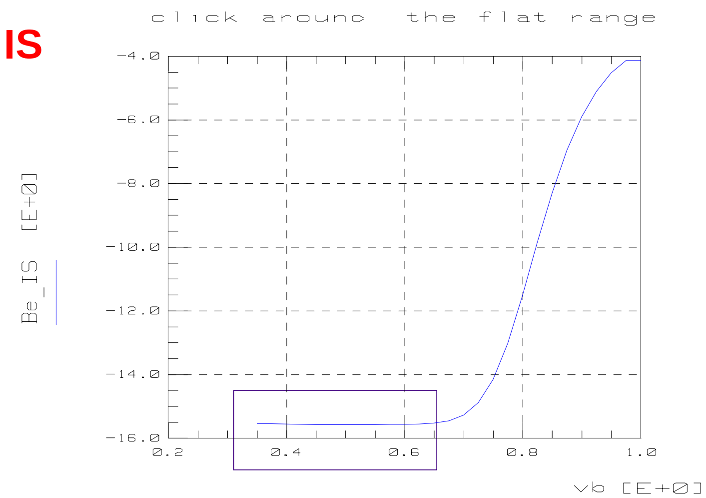
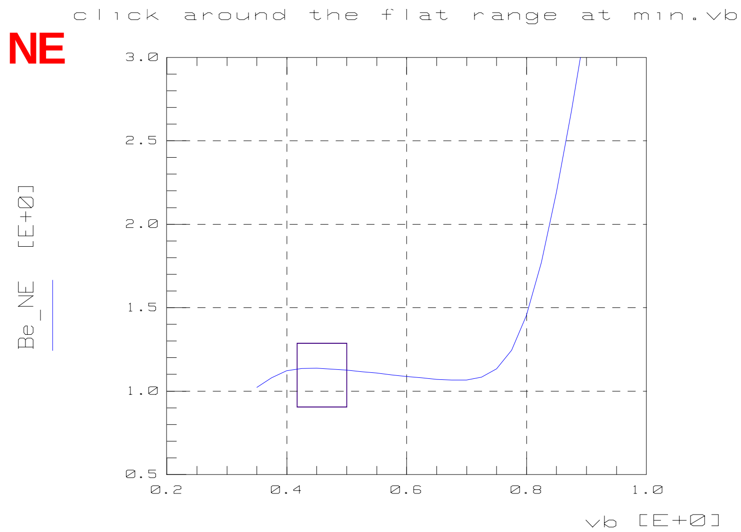


Fig_8: IS from the y-intersect of a line through two adjacent measurement points

NE



$$NE = \frac{1}{VT * \frac{\partial(\ln(i_B))}{\partial(v_{BE})}}$$

Fig_9: NE as obtained from the inverse derivative of the iB-Gummel-plot

ISE

Determine ISE out of the y-intersect of a line through two adjacent measurement points:

```
X=vb
Y=log10(ib.M)

i=1
WHILE i < SIZE(Y)-1
   ISE[i]=(X[i+1]*Y[i-1]-X[i-1]*Y[i+1])//(X[i+1]-X[i-1])
   i = i + 1
END WHILE
```

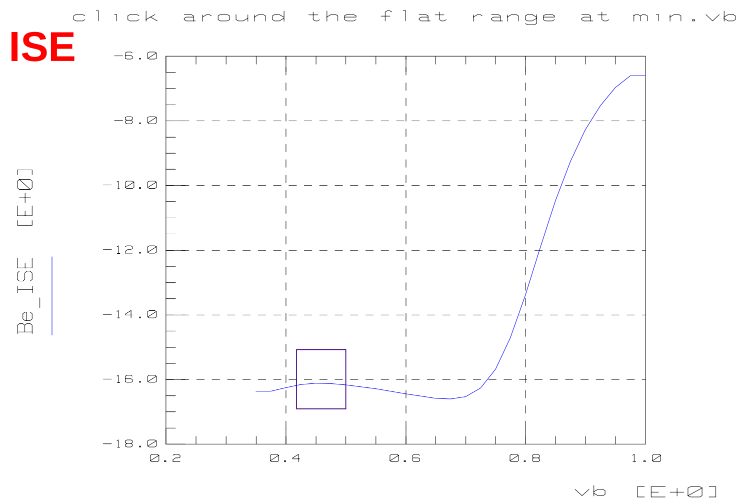


Fig_10: ISE from the y-intersect of a line through two adjacent measurement points

As can be expected from an inspection of fig_6, the measured data does not show a big recombination effect on the ib(vbe) curve. This means that the parameters ISE and NE will not contribute a lot to the curve fitting and may be difficult to extract. This is exactly the meaning of the transformation results in fig_9 and fig_10!

IKF

In order to be able to visualize the effect of IKF (see fig.6 in the introduction chapter), we have to 'strip-off' the effect of RE on the Gummel-Poon iC curve:

```
 ! this extraction assumes that RE is already extracted properly. It
eliminates the effect of RE on  the Gummel-Plot, such that IKF can be
extracted from the 'knee' of the 'stripped-off' ic-Gummel-plot!

!strip-off the effect of RE on the Gummel-plot
!iCmeas = IS*exp(vBEint/(vt*NF) = IS*exp((vBEext-(iC+iB)*RE-iB*RB)/(vt*NF))
! i.e. multiplying iCmeas by exp((vBEext-(iC+iB)*RE-iB*RB)/(vt*NF))
! will give iC without the influence of RE and RB !!

 X  =ABS(vb)
 Yic=ABS(ic.M)
 Yib=ABS(ib.M)

! calculate the stripped-off Gummel Plot
 tmp = Yic*(exp((MAIN.RE*(Yic+Yib)+MAIN.RBM*Yib)//(MAIN.NF*VT)))

! calculate the ideal ic Gummel curve
!  as a reference
 tmp1= MAIN.IS*(exp(X//(MAIN.NF*VT)))

 RETURN tmp+j*tmp1
```

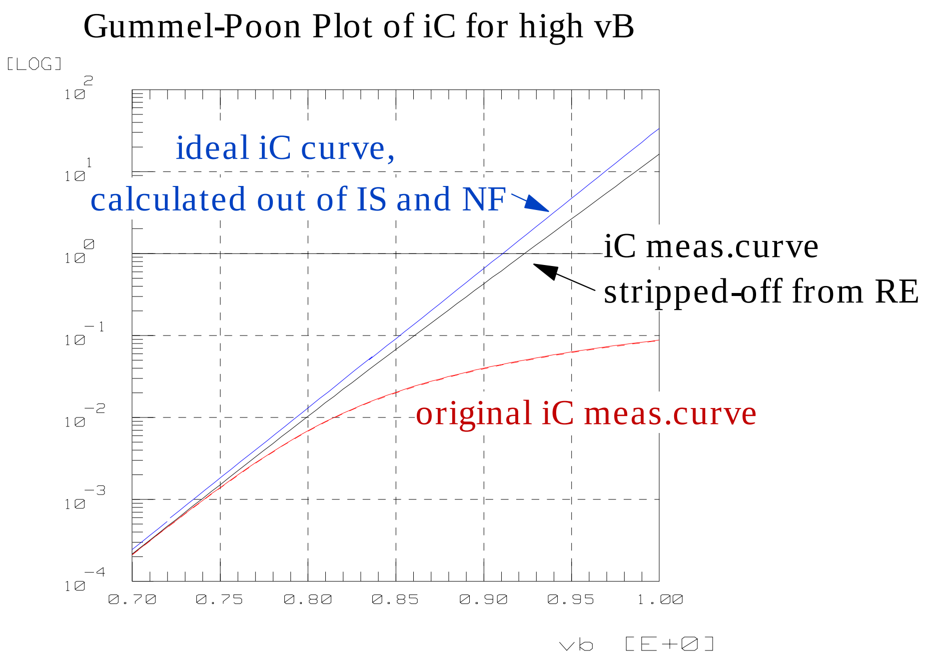


Fig_11a: off-stripped effect of RE on the Gummel-Poon iC curve

NOW CALCULATING IKF

```
X = ABS(vb)
Y = IMAG(strip_off_RE)*(REAL(strip_off_RE))^-1
Y1 = REAL(strip_off_RE)

i=SIZE(Y)-1
index = 0
WHILE i > 0
  IF Y[i] < SQRT(2) THEN
    index = i
    i = 0
  END IF
  i = i - 1
END WHILE

PRINT index

! calculuate IKF out of iC(vBE) at that index
MAIN.IKF = Y1[index]
```

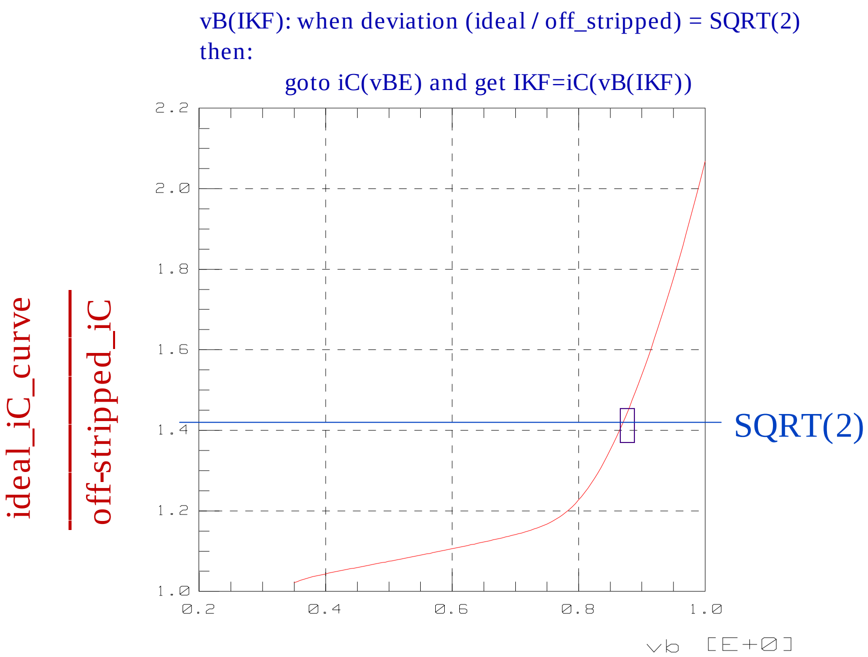


Fig_11b: calculating IKF from the ratio ideal_iC_curve / stripped-off iC

BF

the idea is to strip-off the base charge effect (NqB) from the beta plot and then apply the often cited "max. beta" extraction for BF:

```
!Calculation  of  ß = iC/iB

 vbe  = vb-ve
 vbc  = vb-vc
 beta = ABS(ic) / ABS(ib)


 ! with some simplification is:
 !      beta ~ BF/nqb
 !           nqb  = q1//2*(1+SQRT(1+4*q2))
 !    with    q1  ~ (1-vbe//VAR-vbc/VAF))^-1
 !            q2  ~ IS//IKF*exp(vbe//(NF*VT))
 !            VT :  temp.voltage (a model variable)

 q1 = (1-vbe//MAIN.VAR-vbc//MAIN.VAF)^-1
 q2 = MAIN.IS//MAIN.IKF*exp(vbe//(MAIN.NF*VT))
 nqb=q1//2*(1+SQRT(1+4*q2))

 BF=ic//ib*nqb

RETURN BF
```

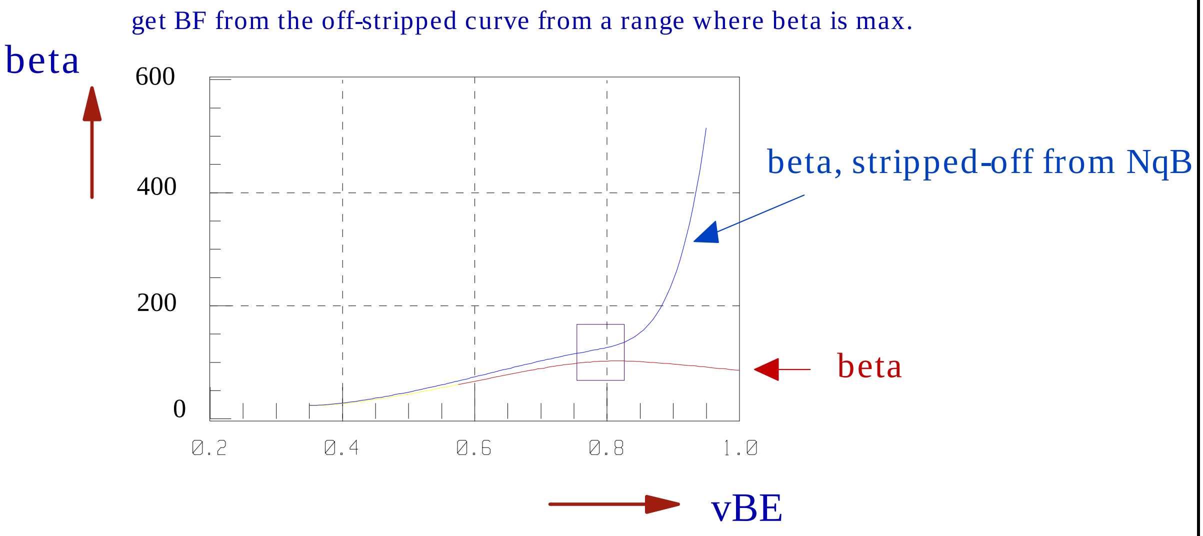


Fig.12: beta plot and beta off-stripped from base charge effects NqB for BF extraction

This method of direct visual parameter extraction can be applied to all Gummel-Poon parameters. For more examples see the IC-CAP files in directory "visu_n_extr".

Let us conclude with an example of a PEL program used for this method. It is the VAF transform of fig_5. The program performs either the data transformation (`SWITCH=1`) or performs the parameter extraction (`SWITCH=-1`):

```
X=vc                                              ! link to stimulus data
Y=ic.M                                            ! link to measured data

Y = SMOOTH3(Y)                                    ! smooth measured data

SWITCH = -SWITCH                                  ! a model variable
LINPUT "data transform (1) or extraction(-1) ?",SWITCH,dummy
SWITCH = dummy

PRINT "calculating the x_intersect ..."

tmp = Y ! a lousy array declaration

i=1
WHILE i < SIZE(Y)-1
  tmp[i]=ABS(X[i+1]*Y[i-1]-X[i-1]*Y[i+1])//(Y[i+1]-Y[i-1])
  i = i + 1
END WHILE

tmp[0] = tmp[1]                               ! watch-out for the array bounds
tmp[SIZE(tmp)-1] = tmp[SIZE(tmp)-2]

IF SWITCH == -1 THEN                    ! sum-up all parameters within the box
  i = 0
  N = 0
  result = 0
  WHILE i < SIZE(X)
     IF X[i] > X_LOW AND X[i] < X_HIGH THEN
       result = result + tmp.M[i]
       N = N+1
     END IF
       i = i+1
  END WHILE
  MAIN.VAF = result//N                                ! export parameter value
  PRINT "MAIN.VAF = ",MAIN.VAF
ELSE
  MENU_FUNC("Visu_va","Display Plot")
  LINPUT "click a box and re-run this transform to extract VAF",dummy
  RETURN tmp
END IF
```

NOTE:

Included in your "bipolar toolkit" is a directory called "visu_n_extract" that contains IC-CAP model files with suggestions on direct visual extraction for most of the Gummel-Poon parameters.

Literature:

HP-EESOF, Characterization Solutions Journal, Spring 1996.

Calculation of h21 of the Gummel-Poon Model

We start from the small signal equival.schematic of fig.9, using some small simplifications:

```
                                                ||
                             ---------------||---------
                            |               ||CB'C    |     iC
          B       ____      |B'     ----              |  <----  C
     ------o----|____|----o--------   |iB'            o---------o-----
       |         rBB'      |         |  V             |               |
      _|_                  |         |               _|_              |
     /   \ ^             -----       -              /   \ ¦          ---
     \   / |       CB'E  -----      | |rB'E         \   / ¦          ---
     /   \ |               |         -              /   \ ¦ iC'=ß iB' |  CAC
     \___/ |iB             |         |              \___/ V           |
       |                   |         |                |               |
       |                   |         |                |               |
      ---                 ---       ---              ---             ---
```

Note: Collector is AC-wise shorted, since h21 = iC / iB for RL = 0

The equations:

Input:

$$vB = iB\ rBB' + vB' \quad -- \quad iB = vB / rBB' - vB' / rBB' \qquad (1)$$

Output:

$$0 = (iC - ß\ iB') / pCB'C + vB' \qquad (2)$$

Internal 1:

$$vB' = (iC - ß\ iB' + iB)\ \underline{r}B'E \qquad (3)$$

$$\text{using} \quad \underline{r}B'E = rB'E\ //\ CB'E$$

Internal 2:

$$iB' = vB' / rB'E \qquad (4)$$

This gives the matrix

	iC	vB	iB'	vB'	
(1):	0	1 / rBB'	0	- 1 / rBB'	iB
(2):	1 / pCB'C	0	- ß / pCB'C	1	0
(3):	$-\underline{r}B'E$	0	ß $\underline{r}B'E$	1	iB $\underline{r}B'E$
(4):	0	0	-1	1 / rB'E	0

Solved for iC:

```
           | iB      1 / rBB'         0         -  1 / rBB'|
           |                                               |
           |  0          0       - ß / pCB'C        1      |
           |                                               |
           |iB rB'E      0           ß rB'E         1      |
           |                                               |
           |  0          0             -1        1 / rB'E |
iC  =  ------------------------------------------------------
          |   0       1 / rBB'         0         -  1 / rBB'|
          |                                                 |
          |1 / pCB'C      0       - ß / pCB'C        1      |
          |                                                 |
          | -rB'E         0           ß rB'E         1      |
          |                                                 |
          |   0           0             -1        1 / rB'E |
```

Solving for the 2nd column of the nominator and the 2nd column of the denominator:

```
                        |  0        - ß / pCB'C        1     |
                        |                                    |
           -  1 / rBB'  |iB rB'E       ß rB'E          1     |
                        |                                    |
                        |  0              -1        1 / rB'E |
iC  =  ------------------------------------------------------
                      |1 / pCB'C      - ß / pCB'C        1     |
                      |                                        |
         -  1 / rBB'  | -rB'E             ß rB'E         1     |
                      |                                        |
                      |  0                  -1        1 / rB'E |
```

Now solving for the 1st column of the nominator and the 3rd row of the denominator:

```
                        |  - ß / pCB'C       1      |
             - iB rB'E  |                           |
                        |       -1        1 / rB'E  |
iC  =  -----------------------------------------------------------
             |1 / pCB'C   1|           |1 / pCB'C   - ß / pCB'C|
       -(-1) |             |  + 1 / rB'E |                     |
             | -rB'E      1|           | -rB'E          ß rB'E |
```

gives for h21:

```
                                     ß
                 - rB'E { - --------------  - (-1)}
       iC                      pCB'C rB'E
h21 = ----   =  --------------------------------------------------
       iB          1                 1       ß rB'E    ß rB'E
                 ------  + rB'E  + ------- { -------- - -------- }
                  pCB'C               rB'E     pCB'C      pCB'C


                 ß / rB'E  -  pCB'C
             =  ---------------------
                 1 / rB'E  +  pCB'C
```

Finally re-substituting

$$1 / \underline{r}B'E = 1 / rB'E + pCB'E$$

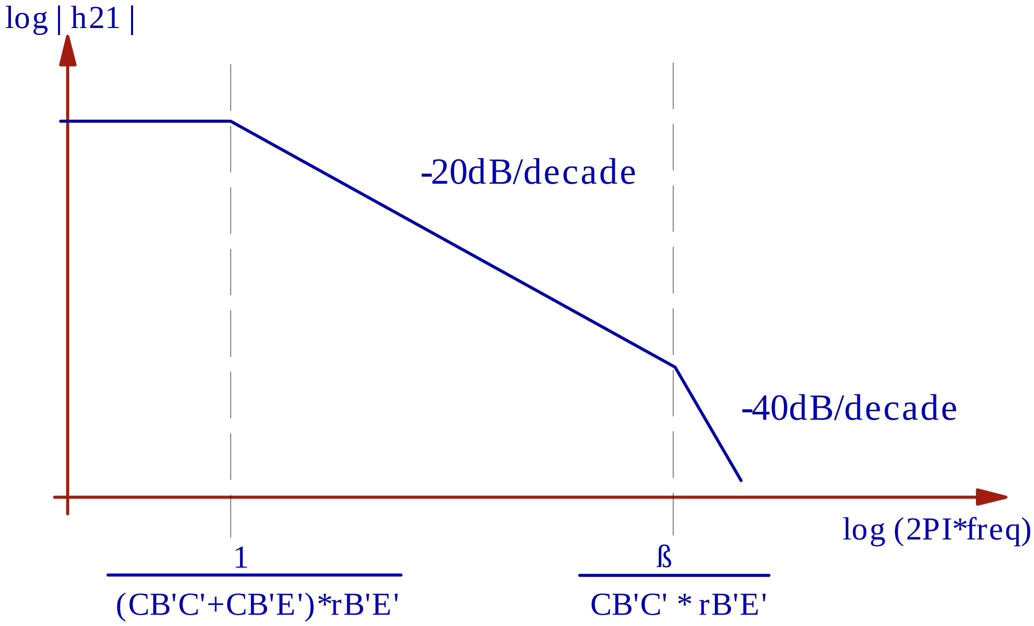
yields

$$h21 = \frac{ß / rB'E \quad - \quad pCB'C}{1 / rB'E \quad + \quad p(CB'C + CB'E)}$$

or

$$h21 = \frac{ß \quad - \quad pCB'C \; rB'E}{1 \; + \; p(CB'C + CB'E) \; rB'E} \qquad (5)$$

what is depicted below:



Now we are ready to calculate the transit frequency fT:

from (5) follows for $h21 = 1$:

$$1 + 4\,PI^2\,fT^2\,(CB'C + CB'E)^2\,r^2B'E \;=\; ß^2 \;-\; 4\,PI^2\,fT^2\,C^2B'C\,r^2B'E$$

or:

$$f_T = \frac{g_{B'E}}{2PI}\sqrt{\frac{ß^2 - 1}{\left(C_{B'C} + C_{B'E}\right)^2 - C_{B'C}^2}} \qquad (6)$$

with:

$$gB'E' = \frac{diB'}{dvB'E} \sim \frac{iB'}{NF\ VT} \qquad (7)$$

$$CB'C \sim CSBC(vBC) \qquad (8)$$

and:

$$CB'E \sim \frac{TFF}{NF\ VT}\ iC \qquad (9)$$

SIMPLIFICATION:

In order to keep things a little simpler for parameter extraction, equ.(5) is modified a bit, neglecting the zero (at high frequencies) against the pole(low frequencies):

$$h21 \sim \frac{ß}{1 + p(CB'C + CB'E)\ rB'E} \qquad (10)$$

Calculating again the transit frequency for this simplified h21 yields:

$$ß^2 = 1 + 4\ PI^2\ f^2 T1\text{-pole}\ (CB'C + CB'E)^2\ r^2 B'E$$

$$f_{T1-pole} = \frac{1}{2PI}\sqrt{\frac{ß^2 - 1}{\left(C_{B'C} + C_{B'E}\right)^2\ r_{B'E}^2}}$$

$$\begin{matrix} CB'E > CB'C \\ \sim \\ ß > 1 \end{matrix} \qquad \frac{ß}{2\ PI\ CB'E\ rB'E}$$

$$\begin{matrix} (9) \\ \sim \\ (7) \end{matrix} \qquad \frac{ß}{2\ PI\ \dfrac{TFF}{NF\ VT}\ iC\ \dfrac{NF\ VT}{iB'}}$$

since ß = iC / iB', we get the pretty simple form

$$fT1\text{-}pole \sim \frac{1}{2\,PI\,TFF} \qquad (11)$$

Or solved for TFF:

$$TFF = \frac{1}{2\,PI\,fT1\text{-}pole} \qquad (12)$$

what is the well-known formula for the TFF parameter extraction.

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For providing valuable feedback on the extraction methods presented in this document, the author would like to especially acknowledge

Mr. Jörg Berkner, Infineon, Munich, Germany

Mr.Dermot MacSweeny from the NMRC in Cork, Ireland,

