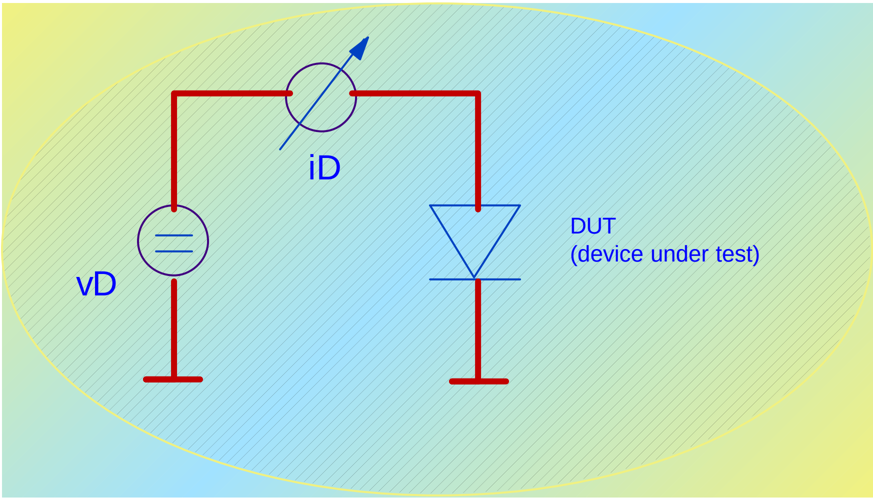
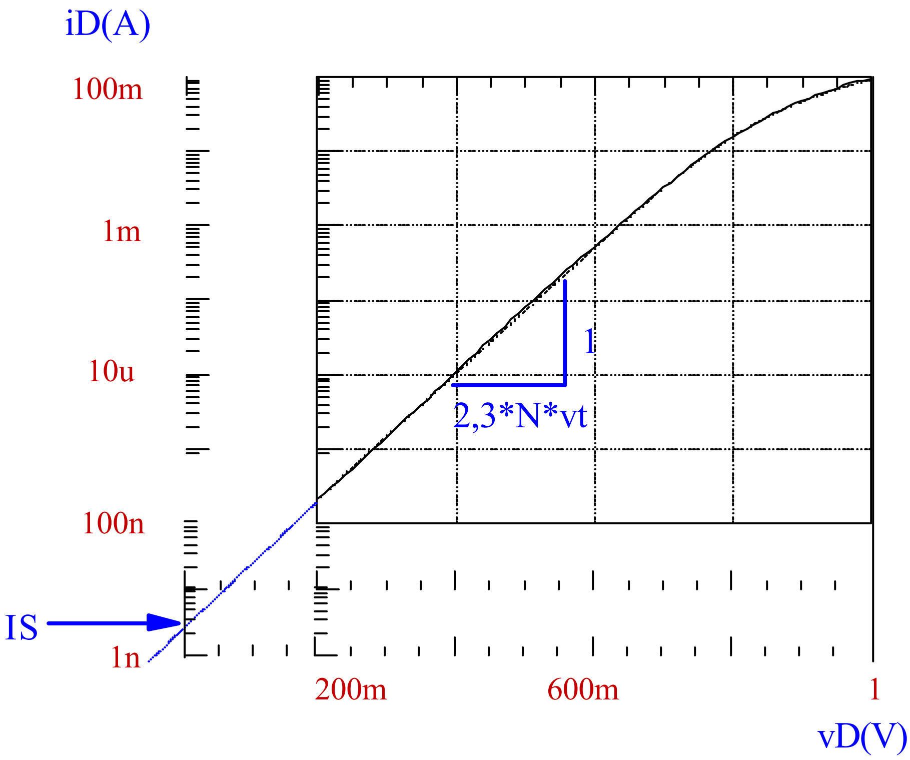
MODELING A DIODE





INTRODUCTION

Referring to the Characterization Handbook's chapter on curve fitting, regression analysis was introduced as a method for linear curve fitting. It was a pretty simple and straight -forward method. We had to so lve the partially differentiated equations for the parameters m and b of the equation

```
y = m*x + b
```

It was mentioned that fitting more complex measurement curves leads to much more complex problems for solving the set of equations for the parameters. And pretty often, this set of non-linear equations cannot be solved without numerical methods.

But there is a way around:
A simple way to fit non-linear curves to measured non-linear data is to transform them to a linear world. But the question is, how to do it. A look at the 'target function' of the model equation gives a hint. An example: in case of an exponential function like

$$i = I_S * e^{\frac{v}{N*vt}},$$

the transform needed is a simple logarithmic conversion. This automatically gives the non-linear transformat ion for the measured data. Once the measured data are transformed to this linear range, a linear regression analysis is applied. And so we yield the slope and y -intersect of the fitted line. The model parameters are finally calculated out of these two valu es.

This will become much more transparent in the following diode modeling example.

The SPICE equivalent schematic for a diode is shown in figure 1. It consists of the ideal diode D representing its non-linear DC characteristic plus two voltage dependent capacitors for taking care of the space charge (CS) and delay effects (CD) as well as a series resistor RS for the high-current effects. The series inductor (bonding effect ...) is neglected as well as a parasitic capacitor (housing effects ...).

```
         iD
         --->
                ____
       o-------|____|--------
  |            RS           |
  |               ---------o--------          |
  |              |          |        |          |
  |            -----        |        |          | vD_internal
  |            \   /       ---      ---         |
vD|             \ /        --- CD   --- CS      |
  |            ----- D      |        |          |
  |              |          |        |          |
  |               ---------o--------          |
  V                         |                   V
       o---------------------
```

D	Diode with its nonlinear DC characteristic	
RS	Series resistor	
CS	**<u>S</u>**pace charge capacitance,	the hyperbolic function
CD	**<u>D</u>**iffusion capacitance,	the transit time effect

Fig.1: Equivalent schematic of the diode

DC CHARACTERIZATION: PARAMETERS IS, N AND RS

Neglecting high current effects, i.e. RS=0 or vD = vD internal, and also neglecting recombination effects for low biasing voltage, the diode current in the forward active region is modeled using:

DC:

$$i_D = I_S * e^{\frac{v_D}{N*vt}} - 1$$

(1)

with

```
Is :       saturation current   (leakage current, typical fA)
N  :       emission coefficient (ideal diode:     N=1)
VT :       temperature voltage   27mV    at 25°C
           or   VT   =  k*T/q  =  8.6171 E-5 * (T /'C  +  273.15)
           (1a)
```

Measurement setup:

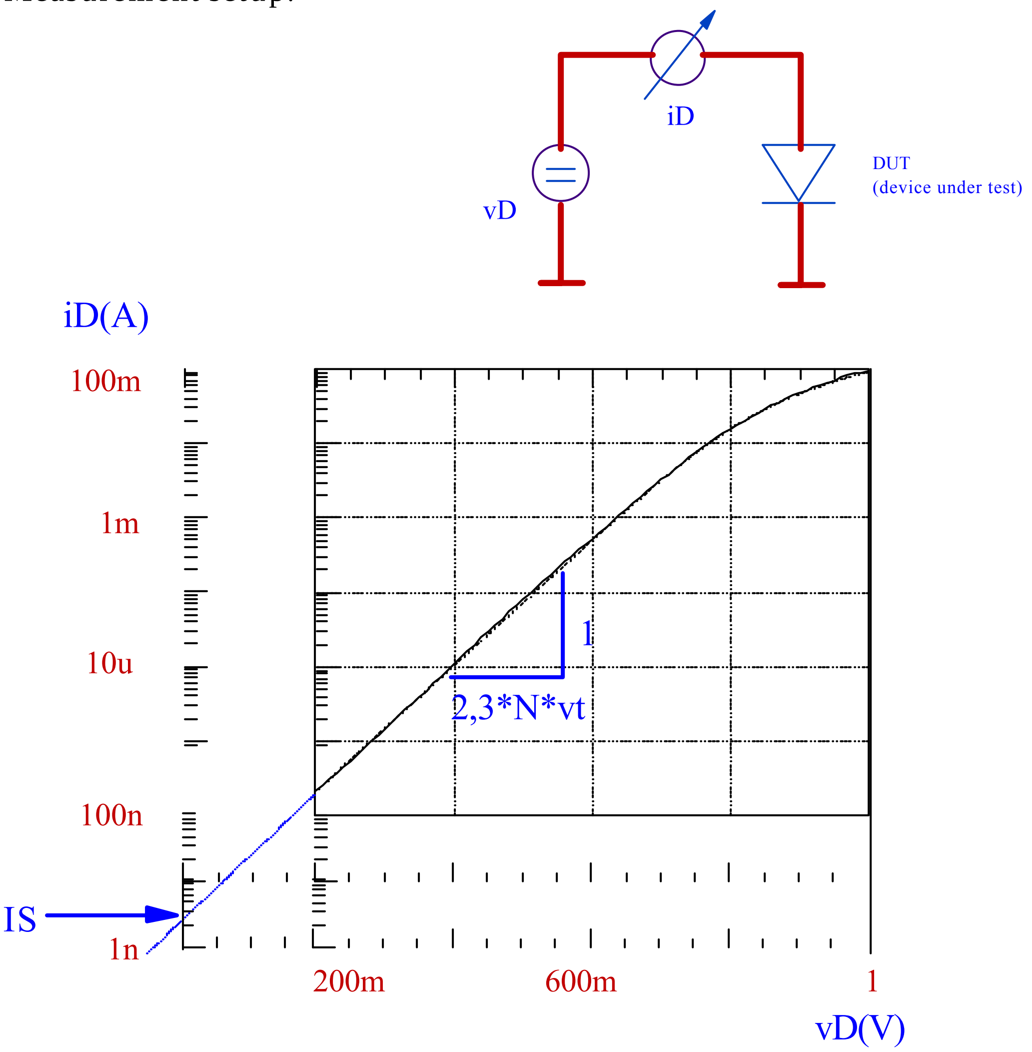


Fig.2: DC characteristic of the diode under forward bias

Determination of the DC parameters IS and N:

Provided vD >0, i.e. neglecting the term (-1) in (1), and applying a logarithmic conversion yields:

$$\log(iD) = \log(Is) + \frac{vD}{N\ VT} \log(e)$$

$$= \log(Is) + [1 / (2{,}3\ N\ VT)]\ vD \qquad (2a)$$

This is an equation of the form:

$$y = b + m \quad x \qquad (2b)$$

In order to interpret (2b) linearly, we have to substitute:

$$y = \log(iD) \qquad (2c)$$

$$b = \log(Is) \qquad (2d)$$

$$m = [1 / (2{,}3\ N\ VT)] \qquad (2e)$$

$$x = vD \qquad (2f)$$

This explains how to manipulate the measured data: after the logarithmic conversion of the measured values of i_D (2c), they are introduced with the still linear values of vD (2f) into the regression equations (10) and (11) of the previous regression analysis chapter 4.1 as yi- and xi-values. We obtain the y-intersect b and the slope m of the linear regression function. Solving (2d) for Is and (2e) for N we finally are able to calculate these two parameters out of b and m as follows:

$$Is = 10^{b} \qquad (3)$$

and

$$N = 1 / (2{,}3\ m\ VT) \quad \text{with VT from (1a)} \qquad (4)$$

Validity of this extraction:

The parameter extraction described above is valid only in that range of measured data, where the assumptions are true. This means: eq.(3) and (4) are valid for vD > 0 (data above the measurement resolution (non-noisy data), typ. >0,2 V). The diode current should not be dominated by recombination effects (the weaker slope at low bias voltages) but also below high-current effects (no ohmic effects, the famous knee in the half-logarithmic diode characteristic above typically 0.7 V)

PARAMETER RS

After the parameters Is and N are extracted, the value of RS can be found from the two highest bias points of index n and index (n-1) as follows:

```
                vD(n) - vD(n-1)
RS_start = -------------------                                 (5)
                iD(n) - iD(n-1)
```

Another method to determine the ohmic part of a diode characteristic is to consider the voltage drop between the ideal diode characteristics its shift due to the ohmic effect. This is done by firstly determining the maximum current from the sweep by

```
i_RS=ia.m[max_index]
```

and then by calculating that voltage drop following

v_RS = measured_voltage - ideal_diode_voltage

or

```
v_RS =  va.m[max_index] - vt*N*ln(i_RS / IS)
```

what finally gives

```
RS = v_RS / i_RS
```

Of course, the diode DC parameters IS and N have to be determined first.

<u>Pre-requisite for a good RS extraction:</u>

the ohmic effect dominates the diode characteristics. Referring to fig.2, the decline for high bias voltage must be clearly visible in the extraction range.

CV CHARACTERIZATION

The frequency behavior of a semiconductor can be modeled by a space charge capacitance (dominant at reverse biasing) and a diffusion capacitance (dominant at forward bias) that models the time delay effects. The first capacitance is typically measured with a negative bias using a CV meter (capacitance versus voltage) while the latter is commonly measured using a network analyzer.

This chapter covers the modeling of the space charge capacitor. Another method is using a network analyzer, measuring the s11 curve with a negative bias. This is not covered here.

Space charge capacitance in general, extracting parameters Cj0, Vj, m

The behavior of the space charge capacitor is given by:

CV:

for vD > FC * VJ :

$$C_S = \frac{C_{JO}}{\left(1 - \frac{v_D}{V_J}\right)^M}$$

(6a)

and else:

$$C_S = \frac{C_{JO}}{(1 - F_C)^{(1+M)}} * \left[1 - F_C * (1 + M) + M * \frac{v_D}{V_J}\right]$$

(6b)

with

CJO space charge capacitance at vD = 0V

VJ built-in potential or pole voltage (typ. 0,7V)

M junction exponential factor, determines the slope of the CV plot

(abrupt pn junction (<0,5um): M = 1/2)

(linear pn junction (> 5um): M = 1/3)

FC : forward capacitance switching coefficient, default 0,5

For more details see /Antognetti/.

Measurement setup:

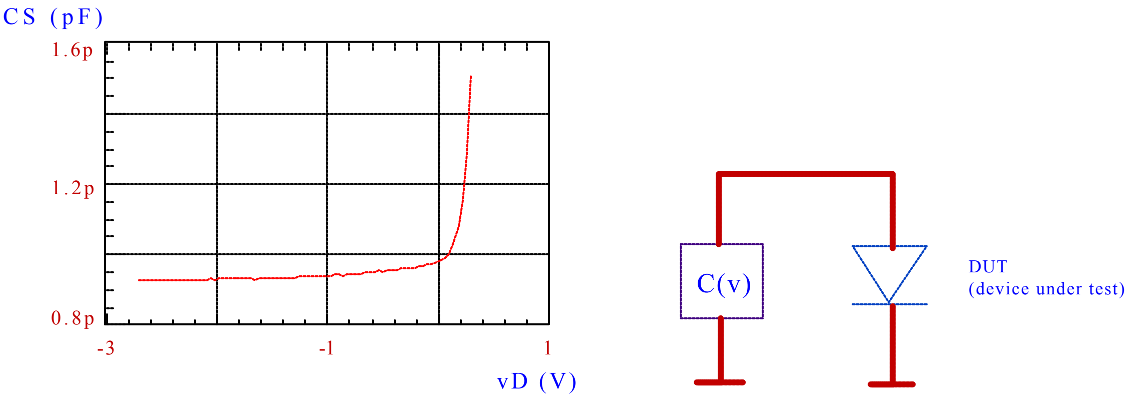


Fig 3: CV-characteristics of the diode

Determination of the CV parameters:

We only use the negative bias region for parameter extraction. The logarithmic conversion of (6a) gives:

$$\ln\left(C_S\right) = \ln\left(C_{J0}\right) - M * \ln\left(1 - \frac{v_D}{V_J}\right) \qquad (7)$$

This equation can be interpreted again as a linear function:

```
                y    =     b     +  m    *    x
with            y = ln(CS)                                  (8a)
                           b = ln(CJO)                      (8b)
                                    m =  - M                (8c)
and                                          x = ln[1 - vD / VJ ]     (8d)
```

How to proceed: the measured values of CS are converted logarithmically according to (8a). Then, following (8d), the stimulating data of the voltage sweep vD are converted too. Since the parameter V_J has a physical meaning, its value should be in t he range of 0,2 .. 1V.

Therefore we select 0,2V as a starting value for VJ. These two arrays are now introduced into equations (10) and (11) of the chapter on regression analysis as yi - resp. xi-values. A linear curve is fitted to this transformed 'cloud' of stimulating and measured data and we get the y-intersect b and the slope m for the actual value of VJ. These two values are the best choice for the given VJ. In the next step, this procedure is repeated with an incremented VJ, and we get another pair of m(VJ) and b(VJ) . But now the regression coefficient r^2 will be different from the earlier one: depending on the actual value of VJ, the regression line fits the transformed data 'cloud' better or worse. Once the best regression coefficient is found, the iteration loop is stopped and we get VJ_opt as well as the corresponding b(VJ_opt) and m(VJ_opt).

The final parameter values are then from (8c):

```
        M   =  - m(VJ_opt)                                               (9a)
```

and from (8b):

```
        CJO  =  exp [ b(VJ_opt) ]                                        (9b)
```

<u>Validity of this extraction:</u>
The parameter extraction for the space charge capacitor is valid only for stimulus voltages below `FC * VJ` , `FC_default = 0,5`.

In practice, there is always an overlay of this capacitance with some parasitic ones, e.g. packaging or bond pads. If they are not known and therefore cannot be de -embedded (eliminated from the measured data by calculations), the three modeling parameters may have values that have no physical meaning. This is especially true for VJ and M.
Nevertheless the fitting of the proposed method is generally very good and pretty easy.

In order to also determine the parasitic offset capacitance, see the examples in the chapter about 'regression analysis applications'.

HF MODELING: PARAMETER TT

The small signal equivalent schematic of the diode for high frequencies is given in fig.4. When comparing it to fig.1, it can be seen that the element D, representing the non -linear DC transfer curve of the diode has been replaced by the linearized small signal conductance gD.

```
          iD
         --->     ____
       o-------|____|--------
  |            RS            |
  |                 ---------o--------          |
  |                |         |        |         |
  |                |         |        |         |vD_internal
  |                -        ---      ---        |
vD|               | |       --- CD   --- CS     |
  |                -   gD    |        |         |
  |                |         |        |         |
  |                 ---------o--------          |
  V                          |                  V
       o---------------------

gD         small signal conductance of the given operating point
RS         ohmic series resistor
CS         space charge capacitance,       the hyperbolic function
CD         diffusion capacitance,          the transit time effect
```

Fig.4: small signal equivalent schematic of the diode for high frequencies

Let's start with gD , the slope to the DC diode characteristics at the operating bias point.

$$g_D = \frac{\partial i_D}{\partial v_D} \overset{(1)}{=} \frac{IS}{N\times vt}\times\left(\exp\left(\frac{v_D}{N\times vt}\right)-1\right) = \frac{1}{N\times vt}\times i_D \qquad (18)$$

The diffusion capacitance in the operating point is given by /Antognetti/:

$$C_D = T_T * g_D = T_T * \frac{1}{N\times vt}\times i_D \qquad (19)$$

CD is typically overlying the space charge capacitance.
When performing a 2-port measurement of a diode with a network analyzer, we can calculate the total diode capacitance by converting the S-parameters to Y-parameters and calculating Cdiode= IMAG(-Y12) / (2 PI freq).
The resulting CV curve is depicted below in fig.5:

Converting S-parameters to CV plots:
The influence of the diode transit time TT to the CV curve

Diffusion Capacitance:

$$C_D = T_T * g_D$$

with

$$g_D = \frac{\partial i_D}{\partial v_D}$$

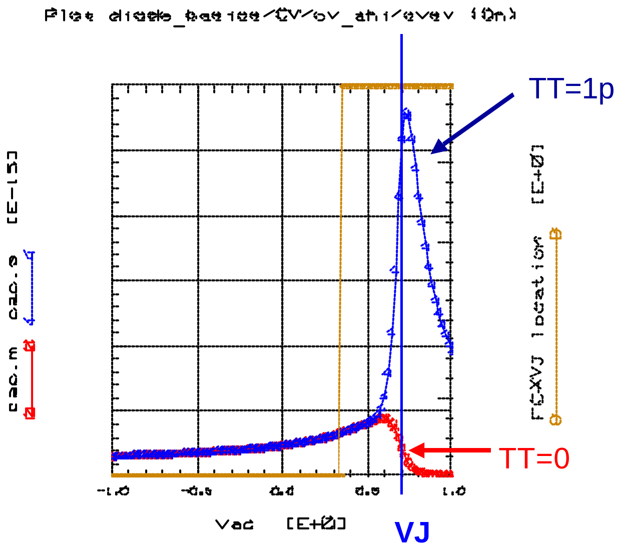


Fig.5: the diffusion capacitance overlays the space charge capacitance for high DC bias.

When applying a NWA, compared to CV meter measurements, there are no restrictions related to positive diode DC biasing and measurement instrument restrictions. Therefore, we can easily forward bias the diode and study the transition from the space charge capacitance to the diffusion capacitance. This gives the diffusion capacitance.

NOTE: in practice, especially for packaged devices, the diffusion capacitance is overlaid by the package inductor! See further below!

<u>Determining TT:</u>

As can be seen in fig. 6, CD can also be optimized in the S -parameters for medium DC biases, below the influence of RS. In other words, r elated to S-parameters, TT shifts the Sxx and Sxy traces (adds phase). The effect is dominant for medium and higher DC biases below take -over of RS.

NOTE:

When RS begins dominating the diode DC trace, think of the 'inner' diode as a resistor with 1/gD in series with a voltage source of e.g. 0,7V. Therefore, the capacitances CS and CD are shortened by this decreasing diode resistor, and therefore, TT is shortened by this resistor!!

The influence of the diode transit time TT to S-parameters

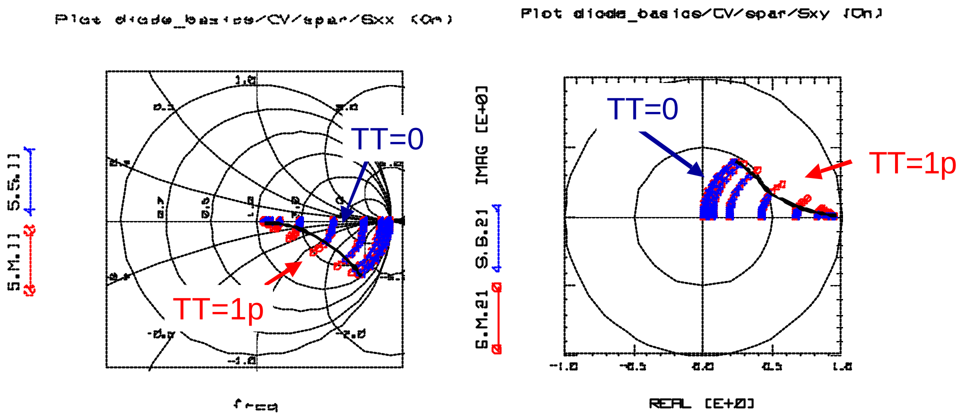


Fig.6: How the transit time TT influences the S -parameters

Finally, when taking also the series inductance into account, which is a typical first -order model of the diode package, we get S -parameters like shown in fig.7:

S-parameter Modeling (package included)

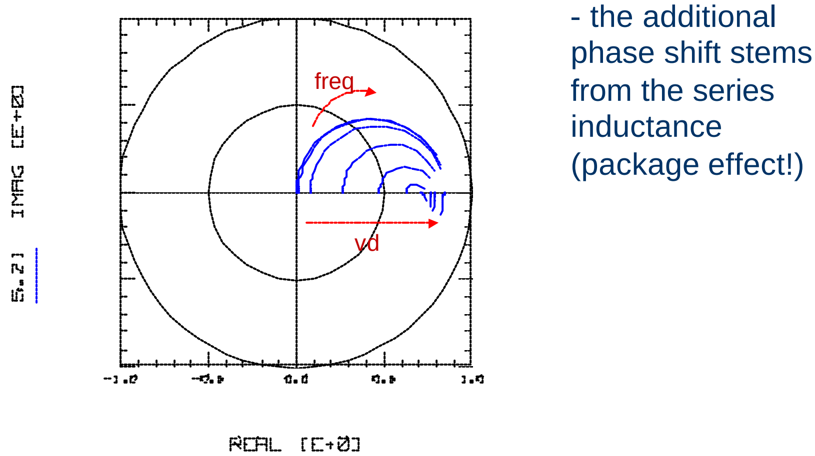


Fig.7: Diode S21-Parameters including the package inductor

The series inductor overlays the so far discussed S -parameters of the inner diode. It basically adds phase to the inner diode S -parameters, and for high DC biasing (where RS domiinates), the inductance affects the diode S21 trace considerable: S21 now turns downwards, tending towards S21->0 for infinite frequency.

<u>Related to modeling, LS can be obtained from optimization of high DC biases.</u>

MODEL LIMITATIONS

In order to keep the models simple and usable and to have reasonable simulation times, they might suffer from some limitations:

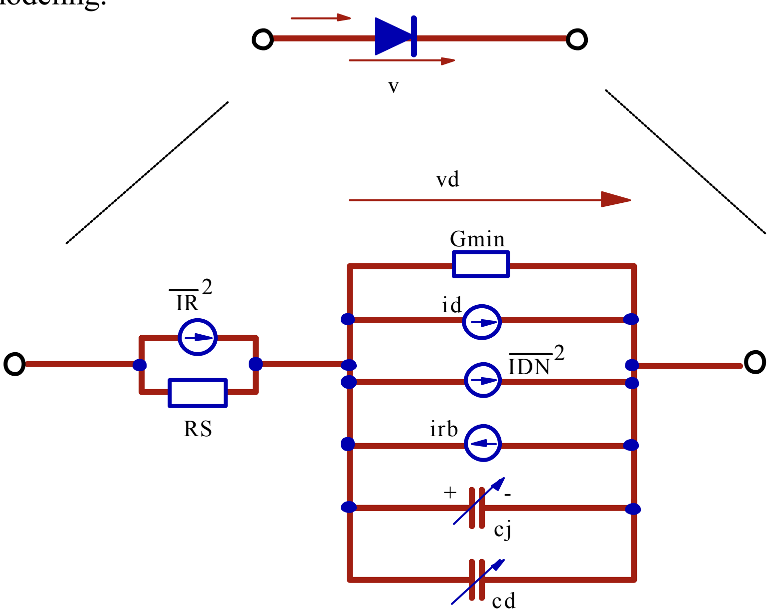
DC: diodes may show recombination effects at low forward bias voltages. This shows-up as a lower slope on a half-logarithmic scale. In order to cover this effect, the diode model is replaced by a subcircuit, consisting of a diode for the recombination effect, another one in parallel for the upper voltage area and a resistance in series with both diodes. (Both diodes have RS=1e-6 Ohm).

CV: any parasitic capacitance is not included in the diode model. Using again a sub-circuit, a 2nd parasitic capacitance can easily be added.

RF: the parasitic series inductor is not included. Again, a subcircuit could be used for modeling.

DETAILS OF THE BERKELEY SPICE DIODE MODELING FOR EXPERTS

This section gives detailed information about the Berkeley Spice model including thermal and noise modeling.



The complete SPICE parameter list:

parameter	description	units	default	example .
RS	ohmic resistance	Ohm	0	10
IS	saturation current	A	1.0E-14	1.0E-14
N	emission coefficient	-	1	1
BV	reverse breakdown voltage	V	infinite	40
IBV	current at breakdown voltage	A	1.E-3	
CJO	zero bias junction capacitance	F	0	2.0E-12
VJ	junction potential	V	1	0.6
M	grading coefficient	-	0.5	0.5
FC	coeff.for forward-bias deplet.cap.	-	0.5	0.5
TT	transit time	sec	0	1E-10
EG	activation energy	eV	1.11	1.11 for Si 0.67 for Ge
XTI	saturation current temp.exp.	-	3.0	3
KF	flicker noise coefficient	-	0	
AF	flicker noise exponent	-	1	
TNOM	parameter measurement temp.	'C	27	
GMIN	min.SPICE conductance (a SPICE convergence parameter)		1E-12	

MODEL EQUATIONS:

DC model:

forward:

$$id = IS\left(e^{\frac{vd}{N*vt}} - 1\right) + G\min * vd \qquad \text{with} \qquad vt = \frac{k * TEMP}{q}$$

reverse:

$$irb = IBV\left(e^{-\frac{vd+BV}{vt}} - 1\right)$$

AC model:

Junction capacitance

$$cj = \frac{CJO}{\left(1 - \frac{vd}{VJ}\right)^{M}} \qquad \text{for } vd<FC*VJ$$

$$cj = \frac{CJO}{(1 - FC)^{M}}\left[1 + \frac{M}{VJ(1 - FC)}(vd - FC * VJ)\right] \qquad \text{for } vd>FC*VJ.$$

Diffusion capacitance:

$$cd = TT\left(IS * \frac{1}{vt * N} * e^{\frac{vd}{N*vt}} + G\min\right)$$

Noise model (used only in AC analysis)

$$\overline{IRS}^{2} = \frac{4 * k * TEMP}{RS} * \Delta f \qquad \text{thermal noise}$$

$$\overline{IDN}^{2} = \underbrace{2 * q * id * \Delta f}_{\text{shot noise}} + \underbrace{\frac{KF * id^{AF}}{f}\Delta f}_{\text{flicker noise}}$$

Thermal model:

$$IS_{TEMP} = IS_{TNOM}\left(\frac{TEMP}{TNOM}\right)^{\frac{XTI}{N}} e^{\frac{q*EG}{k*N}\left(\frac{TEMP-TNOM}{TEMP*TNOM}\right)}$$

$$VJ_{TEMP} = VJ_{TNOM}\left(\frac{TEMP}{TNOM}\right) + 2 * vt * \log\frac{n_i}{n_{i\_TEMP}}$$

with $n_i = 1.45E\text{-}10$

$$n_{i\_TEMP} = n_i * \left(\frac{TEMP}{TNOM}\right) e^{\frac{q}{2k}\left(\frac{-EG}{TEMP} + \frac{1.15}{TNOM}\right)}$$

REFERENCE

Diode modeling and modeling in general:

P.Antognetti, G.Massobrio, Semiconductor Device Modeling with SPICE, McGraw-Hill, 1988, ISBN 0-07-002107-4