A Physically Based Parameter Extraction Scheme for SCR Models

L. Göhler, T. Langer*, J. Sigg*

Universität der Bundeswehr München, Werner-Heisenberg-Weg 39, D-85577 Neubiberg, Germany
Tel.: +49-89-6004-3665, Fax: +49-89-6004-2223, email: goehler@e4s21.et.unibw-muenchen.de
*Siemens AG, Corporate Technology, Munich, Germany

Abstract—This paper presents a physically based parameter extraction scheme for SCR models. The methods are discussed and demonstrated with an example. The comparison between simulated and measured device behaviour shows agreement within 15% tolerance.

I. INTRODUCTION

In recent years research has been concentrated on the development of power device models, but very often parameter determination was not considered sufficiently. These values were assumed to be given, although even the manufacturer can not provide all required information. Therefore the focus of this paper is directed towards this deficit.

Pure curve-fitting (automated or by trial and error) of a given model to measured transients and DC-characteristics may lead to satisfying results, but there is still an uncertainty, whether the initial values of that optimization are precise enough to obtain a parameter set, which allows the model to reproduce the whole range of operation.

For comparatively simple models with only a few not very complex equations one can use derivations of these model equations for extraction [1]. Normally there is no further refinement of the parameter set necessary. In contrast, most more accurate models consist of a large number of equations (explicit and implicit) or are completely numerical. In this case it is of utmost importance to derive appropriate measurement methods for parameter determination.

All extraction methods described in the following are non-destructive and require electrical measurements only.

II. PARAMETER EXTRACTION SCHEME

A. PARAMETER SET

The parameter set, consisting of 15 parameters, corresponds to the model of a cathode-shorted SCR in [2] (see Fig. 1 and Table I).

Additional values, such as thermal resistances and capacities, are taken from the data book.

B. EXTRACTION OF DOPING PARAMETERS

By measuring the capacitance \( C_j \) of the reverse biased middle junction the uniform n-base doping and the doping profile parameters of the p-base can be extracted (Fig. 2).
Since the positive anode-gate bias voltage \( V_r \) forces the \( p \)-emitter-\( n \)-base junction in forward mode its capacity has a considerably larger value than \( C_j \). Consequently, the resulting capacity equals \( C_j \) in good approximation.

Fig. 2. Junction capacitance measurement.

Measured \( C_j \) (\( V_r \)) data are transferred to a device simulator with a built-in optimizer, such as MEDICI, to get the doping parameters \( N_{b1}, N_{b2} \) (\( p \)-doping at peak emitter doping), \( A \) and the sum of \( p \)-base and \( n \)-emitter width \( w_{b2} + w_{e2} \). As the SCR model presupposes a Gaussian \( p \)-base doping profile this profile type should be selected in the device simulator, too. The less dimensions optimization has the easier a solution is found. Assuming that \( N_{b1} \) dominates \( C_j \) at the maximum \( V_r \) gives the opportunity to save one dimension:

\[
A \approx C_{j\text{min}} \sqrt{\frac{2V_{r\text{max}}}{q\varepsilon_0\varepsilon_r N_{b1}}}.
\] (1)

Due to the shorts \( w_{b2} + w_{e2} \) can not be splitted into \( w_{e2} \) and \( w_{b2} \), but usually \( w_{e2} \) is smaller than \( w_{b2} \) and not wider than 20\( \mu \)m. \( N_{e2} \) may follow from estimations. Obviously, the same procedure works for the not shorted \( p \)-emitter-\( n \)-base junction to get \( N_{e1} \) and \( w_{e1} \). The method described becomes more complicated with increasing device dimensions. This means no disadvantage in practice, because in case of high power devices the doping profile is very well known.

As an alternative to device simulations the approximation

\[
\exp \left( -\frac{x^2}{\lambda^2} \right) \approx \frac{a_1}{1 + \exp \left( \frac{a_2}{a_3} x + a_3 \right)}
\] (2)

allows to calculate the capacity of a diffused layer junction (\( a_1, a_2, a_3 = \text{const.} \)). Based on that, a suitable optimizer, such as CYNOSURE, performs the least-square fit.

C. Determination of the Breakdown Voltage

The middle junction is reverse biased again (Fig.3). \( V_{\text{brga}} \) equals that voltage the anode current starts to increase rapidly for.

Fig. 3. Breakdown voltage measurement.

D. Determination of the \( n \)-Base Lifetime

In forward conduction mode electrons and holes are stored in the \( n \)- and \( p \)-base (Fig.4, \( Q_{b1} \gg Q_{b2} \)). During turn-off extraction and recombination reduce excess carrier concentration until the device blocks. In case of no injection into the emitters charge control equation describes the charge decrease in the form:

\[
i_a = \frac{Q_{b1}}{\tau_{an}} + \frac{dQ_{b1}}{dt}.
\] (3)

The variable \( i_a \) denotes the anode current and \( \tau_{an} \) stands for the current-dependent high-level injection lifetime. Applying active or passive commutation (Fig.5) the current waveform can be approximated by linear and sinusoidal functions (Fig.6), so that (3) becomes integrable.

Fig. 4. Charge in the \( n \)-base and the \( p \)-base under static conditions (\( Q_{b1} \gg Q_{b2} \)).

Fig. 5. Measuring circuits for passive (a) and active (b) commutation.
With $Q_{\text{a}1} (t = 0) = I_f \tau_{an}^*$ and $Q_{\text{b}1} (t = t_b) = 0$ the value of $\tau_{an}^*$ follows from active
\[ 0 = \left( \frac{I_f}{I_r} - 1 \right) \left( 1 - \exp \left( -\frac{t_a}{\tau_{an}^*} \right) \right) \times \left( \frac{t_b - t_a}{t_a} - 1 \right) + \exp \left( \frac{t_b - t_a}{\tau_{an}^*} \right) - 1 \] for passive commutation:
\[ 0 = -1 - \left( \frac{\tau_{an}^*}{t_a - t_a} + 1 \right) \left( \exp \left( -\frac{t_b - t_a}{\tau_{an}^*} \right) - 1 \right) + \frac{I_f \exp \left( -\frac{t_b}{\tau_{an}^*} \right)}{I_r - \frac{1}{1 + \omega^2 \tau_{an}^*}} \left( \omega^2 \tau_{an}^* + \exp \left( \frac{t_a}{\tau_{an}^*} \right) \right) \times \left( \cos (\omega t_a) + \omega \tau_{an}^* \sin (\omega t_a) \right). \] Due to emitter backinjection $\tau_{an}^*$ depends on the anode current. Hence, $\tau_{\text{b}1}$ equals the limit of $\tau_{an}^*$ for $I_f \to 0$:
\[ \tau_{\text{b}1} \approx 0.5 \lim_{I_f \to 0} \tau_{an}^*. \] The factor 0.5 derives from Shockley-Read-Hall recombination formula for equal minority carrier lifetime in n- and p-silicon. Commutation speed has minor influence on active commutation only. For passive commutation the assumption of static conditions at $I(t) = I_f$ is not tenable at high frequencies.

\section*{E. Determination of the n-Base Width}

The small signal behaviour of the gate-anode pnp-transistor offers a comfortable way to obtain $w_{\text{b}1}$. Fig.7 presents the equivalent circuit under these conditions. Both junctions are modelled by a capacity and a resistor connected by the n-base resistance $R_{\text{b}1}$.

For a neglectable small value of the parasitic capacity $C_e$ one gets:
\[ R_{\text{b}1} = \frac{1}{qN_{\text{b}1} \mu n} \frac{w_{\text{b}1}}{A} \approx \lim_{\omega \to \infty} R_{\text{meas}e} (\omega), \] otherwise $R_{\text{b}1}$ results from a comparison of the resistance measured at different frequencies with the frequency behaviour according to Fig.7.

\begin{figure}[h]
\centering
\includegraphics[width=0.8\textwidth]{fig7.png}
\caption{Small signal model of the gate-anode structure ($C_d$ diffusion capacity $p^+ n^-$-junction, $R_a$ differential resistance $p^+ n^-$-junction, $R_{\text{b}1}$ n-base resistance, $C_j$ junction capacity $n^- p$-junction, $R_{\text{m}}$ differential resistance $n^- p$-junction, $C_c$ parasitic capacity).}
\end{figure}

As a high frequency LRC meter provides only very low currents this procedure is limited to smaller devices.

Another method considers rapid active commutation ($t_a < \frac{\tau_{an}^*}{\omega}$, Fig.6b) at low currents. In a first step (4) provides a value for the current-dependent lifetime $\tau_{an}^*$. One can then assume the charge at reverse peak current to be
\[ Q (t = t_a) = qAw_{\text{b}1} p_e, \] where $p_e$ denotes the DC-excess carrier density in the n-base at the middle junction. Using (4) again, (9) gives an approximate value for $w_{\text{b}1}$ ($I_{an} = \sqrt{2D_{an} \tau_{\text{b}1}}; D_{an}$ ambipolar diffusion constant in the n-base):
\[ w_{\text{b}1} \approx 4L_{an} \left( \frac{I_f}{I_f - 1} \right) \times \left( \frac{I_f}{I_f - 1} \right). \] In some cases $w_{\text{b}1}$ corresponds to the measured breakdown voltage $V_{br}$. If $V_{br}$ is smaller than the value expected from n-base doping, the current increases due to punch trough:
\[ w_{\text{b}1} = \sqrt{\frac{2e\varepsilon r V_{br}}{qN_{\text{b}1}}}. \]  

\section*{F. Determination of the p-Base Lifetime}

Normally the high-level injection lifetime in the n-base $(2 \cdot \tau_{\text{b}1})$ equals that in the p-base near the middle junction, but doping dependence of lifetime should not be neglected. Using the results by Roulston [4] and Scharfetter, $\tau_{\text{b}2} (x)$ becomes (Fig.8):
\[ \tau_{\text{b}2} (x) = \frac{\tau_{\text{b}20}}{1 + \frac{x_{\text{b}20}}{x_{\text{b}2}}}. \]
with \( \tau_{n2} \approx \tau_{p1} \) and \( N_{ref} \approx 3 \times 10^{15} \text{cm}^{-3} \). The identity of \( \tau_{n1} \) and the maximum minority carrier lifetime \( \tau_{n2} \) explains again from the assumption of equal lifetime in \( n \)- and \( p \)-type silicon.

Fig. 8. Minority carrier lifetime in the \( n \)-base and the \( p \)-base.

A common approach for the effective \( p \)-base lifetime evaluates the integral of \( 1 / \tau_{n2}(x) \) over \( w_{b2} \):
\[
\frac{1}{\tau_{n2}} = \frac{1}{w_{b2}} \int_{w_{b2}} w_{b2} \frac{dx}{\tau_{n2}(x)},
\]
expressing, that the recombination term in the ambipolar diffusion equation does not change when using the effective lifetime instead of the doping dependent one. Inserting (2) into (12) leads to:
\[
\frac{1}{\tau_{n2}} = 1 + \frac{N_{n1}}{N_{ref}} + \frac{a_1 A N_{n2}^2}{a_2 w_{b2} \tau_{n1} N_{ref}} \ln (f_1),
\]

\[
f_1 = \frac{1 + \exp \left( -\frac{a_2}{a_3} (w_{b2} - a_3) \right)}{1 + \exp \left( -\frac{a_2}{a_3} (w_{b2} + w_{b3}) - a_3 \right)}.
\]

G. Determination of Sat. Currents and Short Resistance

With known emitter doping profiles device simulations of the interesting \( p-n \)-junctions are an elegant way to obtain the saturation currents.

Fig. 9. Measuring the anode current as a function of gate-cathode voltage.

Fig. 10. Thyristor structure and equivalent circuit for \( I_a = f (V_{ge}) \) measurement.

Alternatively, the anode current of the ignited thyristor is plotted versus the gate-cathode voltage (Fig. 9, 10). The saturation currents of back injection are obtained by fitting this curve to the model equations in static case for both the \( p^+ n^- \) and \( n^- p^- \)-junction. This method facilitates to ignore the base resistance and forces the current to flow along the main current path. Measuring the characteristic dynamically limits self-heating of the device. Fig. 10 shows the SCR's cathode region and the corresponding circuit.

Equation (14) results for the short resistance at low anode currents, because \( R_s \gg R_p \).

\[
R_s \approx \frac{V_{ge}}{I_a}
\]

(14)

The \( v-i \)-characteristic of the \( n \)-emitter-\( p \)-base-junction has the following form (\( D_n \) diffusion constant for electrons, \( L_{ap} \) ambipolar diffusion length in the \( p \)-base):
\[
I_a = \left( \frac{2qAD_n n_i^2}{N_{b2} L_{ap} \tanh \left( \frac{n_{b2}}{L_{ap}} \right)} + I_{s2} \right) \times \\
\left( \exp \left( \frac{V_{ge}}{mVT} \right) - 1 \right) + \left( 1 + \frac{\mu_n}{\mu_p} \right) I_{s2} \frac{n_i^2}{N_{b2}} \times \\
\left( \exp \left( \frac{2V_{ge}}{mVT} \right) - 1 \right),
\]

(15)

which can be transformed to take into account the parasitic resistance \( R_p \) (\( V_T \) thermal voltage):
\[
V_{ge} = mVT \ln (f_{th}) + I_a R_p,
\]

(16)

\[
f_{th} = \frac{f_1}{2f_2} \left( \sqrt{1 + \frac{4f_2}{f_1}} + \frac{f_1}{f_2} + 1 \right) - 1,
\]

\[
f_1 = \frac{2qAD_n n_i^2}{N_{b2} L_{ap} \tanh \left( \frac{n_{b2}}{L_{ap}} \right)} + I_{s2},
\]

\[
f_2 = \left( 1 + \frac{\mu_n}{\mu_p} \right) I_{s2} \frac{n_i^2}{N_{b2}}.
\]

This leads to \( I_{s2}, R_p \) and \( m \) (in case of a low short resistance it should be subtracted before). The function
of $m$ is to incorporate the remaining self-heating. With the eliminated voltage drop due to $R_p$ a least-square fit of that curve to the anode current as a function of $V_{am}$ (effective voltage across $p$-emitter-$n$-base junction) yields $I_{sc1}$:

$$I_a = \frac{2qAD_y n_i^2}{N_{e1} L_{an} \tanh \left( \frac{u_{sc}}{E_{sc}} \right)} \left( \exp \left( \frac{V_{am}}{mVT} \right) - 1 \right) +$$

$$\left( 1 + \frac{\mu_n}{\mu_p} \right) I_{sc1} \frac{n_i^2}{N_{e1}} \times \left( \exp \left( \frac{2V_{am}}{mVT} \right) - 1 \right).$$

(17)

The difference between $V_{gc}^*$ and $V_{am}$ may be seen as constant:

$$V_{am} \approx V_{gc}^* - V_d$$

(18)

### III. Results

Fig. 11. Comparison between measurement (—) and simulation (- -) of passive commutation with manufacturer/fitted data.

![Fig. 11](image1.png)

Fig. 12. Comparison between measurement (—) and simulation (- -) of passive commutation with extracted parameters.

![Fig. 12](image2.png)

Table II

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Manufacturer/Fit</th>
<th>Extracted</th>
</tr>
</thead>
<tbody>
<tr>
<td>$A$ [$cm^2$]</td>
<td>0.03</td>
<td>0.037</td>
</tr>
<tr>
<td>$N_{e1}$ [$cm^{-2}$]</td>
<td>$2 \cdot 10^{14}$</td>
<td>$1.7 \cdot 10^{14}$</td>
</tr>
<tr>
<td>$N_{e2}$ [$cm^{-2}$]</td>
<td>$4 \cdot 10^{16}$</td>
<td>$2.7 \cdot 10^{16}$</td>
</tr>
<tr>
<td>$N_{e1}$ [$cm^{-2}$]</td>
<td>$1 \cdot 10^{17}$</td>
<td>$9 \cdot 10^{16}$</td>
</tr>
<tr>
<td>$N_{e2}$ [$cm^{-2}$]</td>
<td>$5 \cdot 10^{19}$</td>
<td>$1 \cdot 10^{19}$</td>
</tr>
<tr>
<td>$w_{e1}$ [$\mu m$]</td>
<td>156</td>
<td>173</td>
</tr>
<tr>
<td>$w_{e2}$ [$\mu m$]</td>
<td>18</td>
<td>28</td>
</tr>
<tr>
<td>$w_{c1}$ [$\mu m$]</td>
<td>38</td>
<td>49</td>
</tr>
<tr>
<td>$w_{c2}$ [$\mu m$]</td>
<td>18</td>
<td>20</td>
</tr>
<tr>
<td>$\tau_{e1}$ [$\mu s$]</td>
<td>4</td>
<td>3.5</td>
</tr>
<tr>
<td>$\tau_{e2}$ [$\mu s$]</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>$I_{sc1}$ [$A$]</td>
<td>$0.6 \cdot 10^{-15}$</td>
<td>$1.8 \cdot 10^{-15}$</td>
</tr>
<tr>
<td>$I_{sc2}$ [$A$]</td>
<td>$9 \cdot 10^{-14}$</td>
<td>$9 \cdot 10^{-14}$</td>
</tr>
<tr>
<td>$R_s$ [$\Omega$]</td>
<td>10</td>
<td>6</td>
</tr>
<tr>
<td>$V_{brad}$ [$V$]</td>
<td>1000</td>
<td>1060</td>
</tr>
</tbody>
</table>

Fig. 13. Comparison between measurement (—) and simulation (- -) of DC-characteristic with manufacturer/fitted data.

![Fig. 13](image3.png)

Fig. 14. Comparison between measurement (—) and simulation (- -) of DC-characteristic with extracted parameters.

![Fig. 14](image4.png)
All extraction routines were performed for a small thyristor \((I_{\text{TAV}} = 7.5\, \text{A})\) to prove the concept. In Table II a comparison is displayed between data given by the manufacturer or obtained from fitting the model to best agreement with measurement and extracted parameters.

In transient simulation (Fig. 11, 12) a reverse peak current deviation of approximately 15% occurs. The DC-current difference (Fig. 13, 14) at \(I = I_{\text{TAV}}\) amounts 14%. These deviations are mainly determined by the different values of \(\tau_1\) and \(I_{\text{st}}\), which lead to an increased device resistance and a reduction of the stored charge in the model compared to the device. It is clearly to be seen, that the extracted parameters meet manufacturer data.

IV. Conclusion

A physically based parameter extraction scheme for SCR models is presented. The comparison with manufacturer data proves a good agreement within 15% tolerance. Some of the procedures, such as junction capacity measurement, minority carrier lifetime determination and the extraction of \(n\)-base width can be applied to other power devices, too. All methods are destruction-free.

References


