Physics-Based Models of Power Semiconductor Devices for the Circuit Simulator SPICE

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ABSTRACT

Models of power semiconductor devices are implemented in the circuit simulator PSpice. The combination of subcircuits and mathematical functions enables very compact solutions. High accuracy and validity in a wide operation range are obtained due to the derivation from device physics. Models of the power diode and the IGBT are presented as examples.

INTRODUCTION

Many designers of power electronic circuits still hesitate to use the tool of circuit simulations, although adequate models of power devices are now available for some simulation programs. To overcome this reluctance an inexpensive simulation software is desired and it should be easy to use on Personal Computers. The manufacturers of power semiconductor devices should provide parameterized models of their products which can supplement or even replace data sheets since much more information and accuracy would be available. A progress in achieving such an aim can be observed already, but some further steps to improve the situation are still necessary.

The most advanced power device models with the best accuracy [1-5] are incorporated in simulation programs which enable an easy insertion of model equations (like Saber, Eldo). These programs are very powerful but also very expensive. The standard simulation program with the widest distribution, however, is Spice with its numerous derivatives. Since this software is relatively inexpensive and can be run on PCs there is a great demand for adequate device models. But the way to incorporate new models in Spice is restricted and not as straightforward as e.g. in Saber. Power device models for Spice are mainly realized by subcircuits which, however, can become very complex and can require much computation time. Many Spice models of power devices have been developed during the last years [6-8] but yet they are not as accurate and efficient as the mathematical models in other simulation programs.

This paper shows the implementation of very efficient power semiconductor device models in the simulation program PSpice. Although these models are relatively simple, they are not purely empirical behavioral models, but they are derived from device physics. Therefore high accuracy and validitiy in a wide range of operation are obtained.

MODEL DESCRIPTIONS

The models are realized as subcircuits consisting of generic Spice models of semiconductor devices (diode, MOSFET) combined with controlled current and voltage sources and some passive elements. For the controlled sources mathematical functions are used which are available in PSpice from version 6.2 up. Due to this combination very compact models could be obtained. This paper shows the models of the power diode and the (NPT-) IGBT as examples of this modeling method. The models are based on the approaches which have been described in [3] and [4]. The mathematical expressions resulting from these approaches have partly been simplified further in order to make the implementation in PSpice easier and to improve the simulation speed.

<u>Power diode:</u>

The p-ν-n structure of a power diode is shown in fig. 1, and the subcircuit which represents the power diode model can be seen in fig. 2. Few elements are used to describe the steady-state and dynamic behavior.

Two standard diodes reproduce the DC-characteristics; D_o takes into account recombination in the lightly doped middle region, D_E contributes the current components of the emitter regions. The currents of D_o and D_E have different dependences on voltage (different emission coefficients). Junction capacitance and breakdown are also considered with the diodes.

The current source I^*_Q supplies a current in the dynamic case. This current is caused by the diode charge and mainly determines the diode switching behavior. The current source I_R represents the diode resistance which is modulated by the diode charge.

The charge is determined with the help of an additional subcircuit which is shown in fig. 3. This circuit is a representation of the charge control equation.

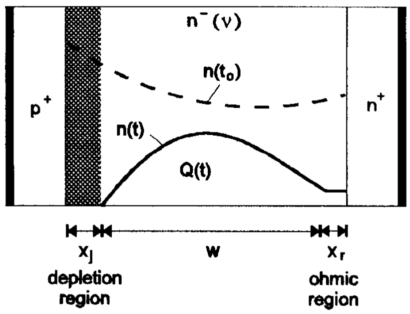

Figure 1: Structure of a p-v-n diode and charge carrier distribution in lightly-doped region during forward operation (n(t$_o$)) and during turn-off (n(t))

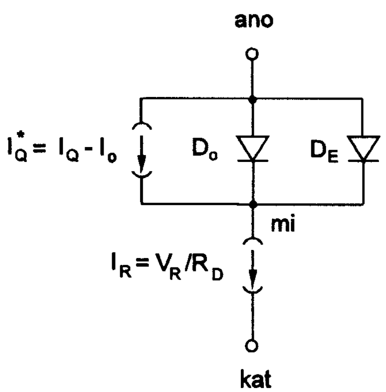

Figure 2: Subcircuit of the power diode

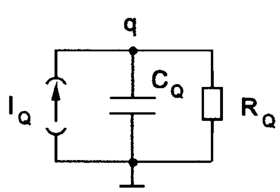

Figure 3: Subcircuit for stored charge

$$I_Q = \frac{Q}{\tau} + \frac{dQ}{dt} \qquad (1)$$

The capacitor C_Q is the charge storage element, the resistor R_Q takes into account the recombination of electrons and holes ($R_Q = \tau/C_Q$, τ is the charge carrier lifetime). The current component I^*_Q is the difference between I_Q and I_o which is determined by the following relation between charge and current:

$$I_Q - I_o = \frac{Q_o - Q}{T_d} \qquad (2)$$

$$with \quad T_d = \frac{a_1 \frac{w^2}{D}}{1 + \frac{w}{12D}\frac{dx_j}{dt}}$$

This relation is obtained with the help of an approximate solution of the ambipolar diffusion equation. I_o is the current of D_o and Q_o is the corresponding charge in the equilibrium state ($Q_o = \tau I_o$). I_Q - I_o is unequal to zero if the charge Q differs from the equilibrium value. I_Q can be positive or negative. The amount of this current depends not only on the charge difference but also on T_d which is not a constant but a strongly varying factor. w is the width of the charge distribution and x_j is the width of the depletion region at the pn-junction. x_j is calculated as a function of the junction voltage (V_{ano}-V_{mi}). When the excess charge is axtracted during turn-off the depletion region expands into the middle region and a neutral region where the electron density has dropped to the doping concentration builds up on the side towards the n^+ end region (see fig. 1). The width x_r of this ohmic region is approximated with $a_2 x_j$. a_1 and a_2 are constant factors which depend on the diode parameters, and they influence the turn-off behavior (the softness of reverse recovery).
The factor T_d becomes smaller if w is reduced and if the decrease of w with time is fast (large value of dx_j/dt). A smaller T_d results in a higher current. This description considers therefore that a fast turn-off of the diode causes corresponding high reverse currents. This is a significant difference between this power diode model and other (mainly empirical) models which use a constant factor for the reverse current characteristics.

IGBT:

Fig. 4 shows the structure of the NPT-IGBT (one half elementary cell). The IGBT can be described as the combination of a MOSFET and a Bipolar Transistor. Accordingly the subcircuit which is used to model the IGBT consists of a MOSFET part and a bipolar part. This subcircuit is shown in fig. 5.
For the MOSFET part a standard Spice MOSFET model is used, for the bipolar part, however, a standard Spice model is not suitable since it cannot correctly reproduce the characteristics due to high-level injection and the non-quasi-static effects during transients. Therefore a special equivalent circuit is used to model the typical device behavior.
The MOSFET model is a simple level-1 Spice model which is extended by resistors and capacitors in order to take into account the specific characteristics of the MOSFET structure.

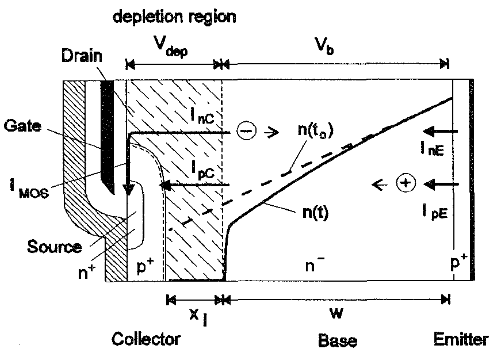

Figure 4: Structure of a NPT-IGBT and charge carrier distribution in base during stationary on-state ($n(t_o)$) and during turn-off ($n(t)$)

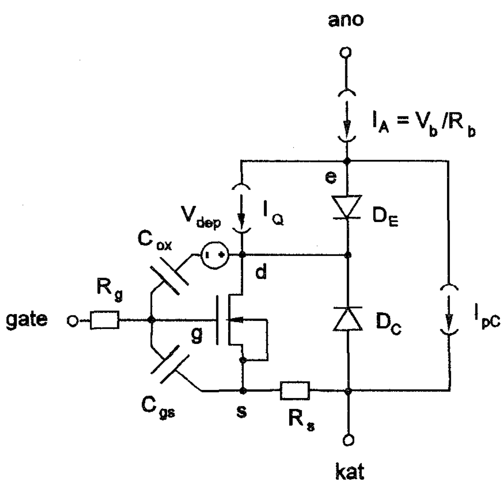

Figure 5: Subcircuit of the NPT-IGBT

Deviations of the real device from the ideal current equations of the level-1 model are considered with the source resistance R_s in a simple but effective way. The resistor R_g represents the internal gate resistance of the device. The capacitances of the MOS-structure consist of a constant capacitance C_{gs} between gate and source and a voltage-dependent capacitance between gate and drain which is described by the capacitor C_{ox} and a voltage source V_{dep} in series. V_{dep} considers the voltage drop in the case of depletion below the gate (V_n is a parameter-dependent constant factor):

$$V_{dep} = V_{dg} + V_n\left(1-\sqrt{1+\frac{V_{dg}}{V_n}}\right) \qquad (3)$$
$$\text{if} \quad V_{dg} \geq 0 \qquad \text{else} \quad V_{dep} = 0$$

The bipolar part of the IGBT is modelled with two diodes and three current sources. D_E reproduces the electron current at the base-emitter junction (I_{nE}) which is the base current of the Bipolar Transistor. The diode D_C represents the base-collector junction; during normal operation only the junction capacitance is effective.

The current source I_Q takes into account how the components of electron and hole currents change across the base region; the DC-component is due to recombination, and the dynamic contribution is caused by variations of the base charge. The current components at the borders of the neutral base region are indicated in the schematic of fig. 4. The electron current I_{nC} is composed of the MOSFET current and the displacement current of the base-collector junction capacitance (diode D_C). The relations between the current components are:

$$\begin{aligned} &I_{nE} = I_{nC} - I_Q \\ &\text{with} \quad I_{nE} = I(D_E)\,,\, I_{nC} = I_{MOS} + I(D_C) \\ &I_{pE} = I_{pC} + I_Q \end{aligned} \qquad (4)$$

I_{pC} is the collector hole current which is represented by a current source in the subcircuit model. According to the transport equation in the case of high-level injection I_{pC} is proportional to I_{nC} and to the slope of the charge carrier distribution at the boundary between collector depletion region and the stored base charge.

$$I_{pC} = \frac{\mu_p}{\mu_n} I_{nC} + \left(1+\frac{\mu_p}{\mu_n}\right) qAD\frac{dn}{dx}\bigg|_{xj} \qquad (5)$$

μ_p and μ_n are the mobilities of holes and electrons, respectively. The shape of the stored charge can be seen in fig. 4 for the two cases of stationary forward operation and turn-off. The slope of the charge carriers is determined with the help of the same approach which is applied to obtain the stored charge. The result is:

$$\begin{aligned} &I_{pC} = \frac{\mu_p}{\mu_n} I_{nC} + \left(1+\frac{\mu_p}{\mu_n}\right)\left(F_1 I_o + F_2(I_o - I_Q)\right) \\ &\text{with} \quad F_1 = \left(\cosh\left(\frac{w}{L}\right)-1\right)^{-1} \\ &\qquad\quad F_2 = \frac{1}{2}\left(1+\tanh\left(\frac{w}{6D}\frac{dx_j}{dt}\right)\right) \end{aligned} \qquad (6)$$

w is the neutral base width, L and D are diffusion length and diffusion coefficient, respectively, x_j is the width of the collector depletion region. The current I_o is determined by the equilibrium charge Q_o ($I_o = Q_o/\tau_b$) which can be calculated from the current I_{nE} (with the model parameter I_{sE}):

$$Q_o = qAn_i \tanh\left(\frac{w}{L}\right)\sqrt{\frac{I_{nE}}{I_{sE}}} \qquad (7)$$

The term $F_2(I_o - I_Q)$ in equ. (6) is the important contribution to the IGBT current in the dynamic case; its amount depends strongly on the speed of the moving charge boundary x_j which is a function of the anode-cathode voltage.
The current I_Q and the charge in the base are determined with the subcircuit shown in fig. 3 in the same way as has been described for of the power diode. The base resistance which is modulated by the injected excess charge is considered with the current source I_A.

MODEL PARAMETERS

The parameters which are required for the diode and IGBT models are listed in table 1 and table 2, respectively. Additionally to these parameters several physical quantities are used which have fixed values and need not to be extracted from device characteristics. The number of parameters is relatively small which makes the application of the models easier.

Table 1: Parameters of diode model

Symbol	Unit	Parameter
A	cm^2	area
w_o	cm	width of middle region
N_d	cm^{-3}	doping of middle region
τ	s	charge carrier lifetime
I_{sE}	A	saturation current for D_E
I_{sM}	A	saturation current for D_o
C_{jo}	F	depletion capacitance
BV	V	breakdown voltage
R_c	Ω	contact resistance
SF	-	softness factor

Table 2: Parameters of IGBT model

Symbol	Unit	Parameter
A	cm^2	area
A_{gd}	cm^2	gate-drain overlap
w_b	cm	base width
N_b	cm^{-3}	base doping
τ_b	s	lifetime in base
I_{sE}	A	saturation current for D_E
C_{jE}	F	B-E depletion capacitance
C_{jC}	F	B-C depletion capacitance
V_{th}	V	threshold voltage
K_p	A/V^2	transconductance
C_{ox}	F	oxide capacitance
C_{gs}	F	gate-source capacitance
R_s	Ω	source resistance
R_g	Ω	gate resistance
L_c	cm	size of elementary cell

Due to the physical origin of the models some data of the device geometry and the doping concentrations are required. It is a little complicated (but not impossible) to obtain these parameters from electrical measurements; therefore they should preferably be provided by the device manufacturers. Most of the parameters, however, can easily be extracted from measurements or taken from data sheets. But anyway the best solution would be to get the complete parameter set from the manufacturers; they should deliver parameterized models of their products in addition to data sheets. (The presented models will be available from the power semiconductor division of the Siemens semiconductor group.)

RESULTS

The models have been validated by many comparisons of simulations with experiments. A few examples are demonstrated in this paper. Fig. 6 shows the basic test circuit which has been used to measure and simulate the dynamic behavior of IGBT and power diode. The parasitic inductances of the wires are included in the netlist for simulation since they are not neglectable in the case of fast transients. The inductance at the IGBT emitter e.g. influences the turn-on behavior of the transistor.

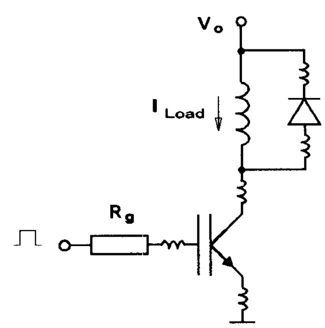

Figure 6: Test circuit

Fig. 7 shows the reverse recovery of the diode for two different values of the total parasitic inductance L_σ in the circuit branch where the current is commutated.
This inductance determines the slope of the current if the IGBT is switched sufficiently fast. The resulting maximum reverse current, the reverse recovery time, and the voltage overshoot are well reproduced by the diode model.
The switching waveforms of the IGBT can be seen in fig. 9. Two different resistance values have been used in the gate drive for these examples. Good agreement between measurements and simulations is obtained. The model correctly predicts the dependence of the switching delay on the gate resistance or the current and voltage curves at the anode which determine the power loss during switching. The remaining differences are mainly due to the fact, that the gate

drive of the IGBT has just been simulated with a simple pulsed voltage source and a resistor R_g (see fig. 6).

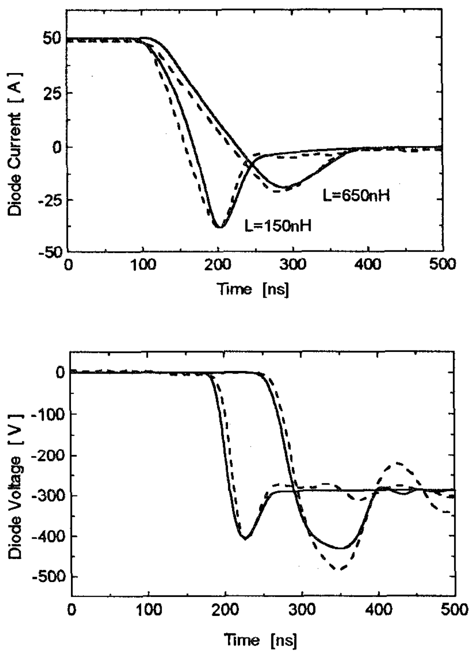

Figure 7: Reverse recovery of the power diode, comparison of simulations (—) and measurements (- -)

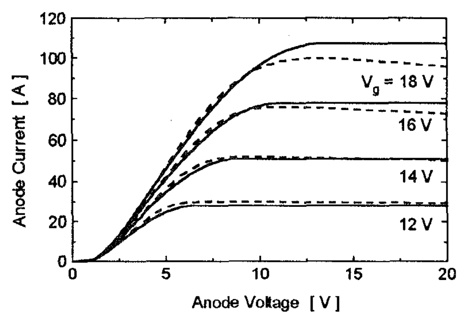

Figure 8: DC-characteristics of the IGBT, (—) simulations, (- -) measurements

The DC-characteristics of the IGBT are shown in fig. 8. At the measured curves a decrease of current with increasing voltage can be observed in the saturation region. This effect is due to the self-heating of the IGBT which is not yet included in the PSpice models.

CONCLUSION

Models of power semiconductor devices are incorporated in the circuit simulator PSpice as subcircuits in combination with mathematical functions. This method enables very compact models with short calculation times. The subcircuits are easy to understand and the mathematical functions guarantee high accuracy since they are obtained by derivations from device physics. These models are probably close to the optimum compromise between simplicity and accuracy.

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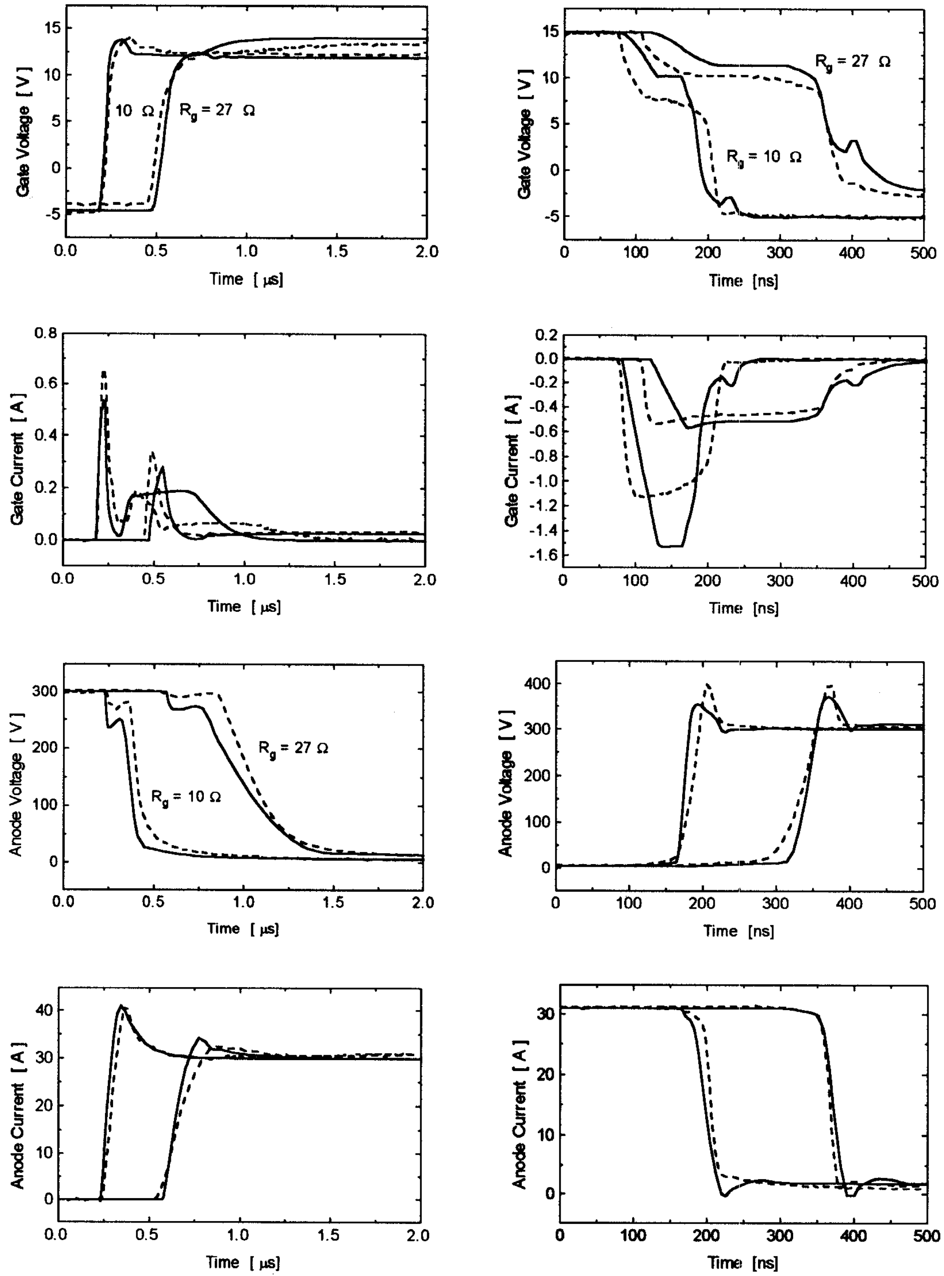

Figure 9: Turn-on (left column) and turn-off (right column) of an IGBT with inductive load and free-wheeling diode, comparison of simulations (—) and measurements (- -)