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STATE OF THE ART OF DV/DT AND DI/DT CONTROL OF INSULATED GATE POWER SWITCHES

Pierre LEFRANC¹, Dominique BERGOGNE²

(1) SUPELEC Département Energie Plateau du moulon - 3 rue Joliot Curie 91192 Gif-sur-Yvette - France +33 (0)1.69.85.15.08 pierre.lefranc@supelec.fr
(2) Laboratoire AMPERE - INSA de Lyon Bat Leonard de Vinci, Avenue Capelle 69621 Villeurbanne - France +33 (0)4.72.43.82.38 dominique.bergogne@insa-lyon.fr

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1. INTRODUCTION

IGBT and MOSFET power switches are commonly used in power converters. MOSFET for middle power and IGBT for high power converters : gate drive circuits are used to drive such components. Some applications need to control dv/dt and di/dt from power switches to reduce electromagnetic emissions. New solutions have been introduced and tested since 1990 (corresponding to the maturation of IGBT technology) [1,2].

We expose first consequences due to dv/dt and di/dt of power switches such as EMC problems. A theoretical approach highlights coupling phenomena between the power side and signal electronics. Conduction and emission aspects are discussed to show that dv/dt and di/dt must be reduced for some applications.

Solutions to these problem cannot be proposed without understanding power switch structure and current-voltage transients. Therefore, focus is made on a theoretical approach on IGBTs (Insulated Gate Bipolar Transistor). MOSFETs (Metal Oxyde Field Effect Transistor) are considered in this part too but not described in detail. First a physical approach is proposed (structure of IGBT cell) that leads to an IGBT model. A simplified electrical model is considered that facilitates explanations on transient voltage and current. Some considerations are outlined in order to introduce proposed solutions.

In proposed solutions, the authors underline solutions published in international papers and personal one. Main ideas are developed to give an overview of state of the art of dv/dt and di/dt control. Current ideas are made on inductance estimation of IGBT power modules necessary for di/dt control : experimental and simulation results are compared for 1200A-3300V IGBT modules.

2. CONSEQUENCES OF TRANSIENT VOLTAGES AND CURRENTS

Power converters are well known to be the source of electromagnetic disturbances. Consequences can be found for electrical machines (problem of wires insulation), for analogic and digital electronics, telecommunication problems, radio systems [3], etc...

In this section, the main phenomena that must be taken into account to understand the associated technical problems is discussed in order to focus later on dv/dt and di/dt control.

Transient problems are due to di/dt and dv/dt from power switches. Current and voltage vary with different speeds depending on gate drivers. Gate drivers are close to switches. They send turn-on and turn-off signals from control-side to gate power switches, they feed power and pulsed currents to gate resistors. Security measures are implemented to protect IGBT and MOSFET from destruction due to short-circuit and impedance faults for example. Complementary functions are also implemented for dv/dt and di/dt control so as to reduce electromagnetic emissions with a minimum increase of power losses. In Fig1, an example of an inverter leg (two IGBT and load represented) with control and gate drivers is shown. The secondary side of the gate driver is shown in Fig2 : a power supply insulates the secondary side from the primary side and a bidirectional transmission exchanges information from primary to secondary. To drive the power switch, the "gate drive" function sends turn-on and turn-off signals close to the isolated gate. An impulsed current is also generated during turn-on and turn-off so as to charge and discharge the gate. The "security and transient control" function supervises switching conditions, detects and reacts in case of disfunctioning [4].

In classical configurations and applications, the gate drive circuit consists of a voltage source ±15V that is switched to feed a gate resistor and a gate power switch. It follows that transient voltage and current can not be adapted during the functioning of the converter.

Current and voltage transients cause electromagnetic emissions that disturb surrounding electronics. The emission can also be conduction current or electromagnetic field. The main drawback of voltage is well known common conduction current. In Fig3, we represent the common mode parasitic capacitors. They model global
capacitance between power switches and cooling systems (on the right of the figure) and global capacitance between control hardware and mechanical ground. The voltage transients generate a current through parasitic capacitances. Even with a low value of capacitance, around 100pF for example, with a classical voltage transient of 1kV/µs, we have $i_{cm} = C_p \frac{dV}{dt} = 100 \times 10^{-12} \times 1 \times 10^{-6} = 100 \text{mA}$ current. This current can easily perturb electronic components such as operational amplifiers for measurements or microcontroller and DSP for the converter control.

The second problem comes from electromagnetic field emission. Voltage and current transients generate electromagnetic fields that can generate parasitic current and voltage in surrounding electronics. Unfortunately, it is difficult to quantify the electromagnetic field emitted from power switches, power converter and coupling effects. An example of disfunctioning due to voltage transients and electromagnetic fields is that of an integrated photodiode receiver [5]. The photodiode-current generated by the photodiode is amplified by a transimpedance-amplifier and the logical information is transmitted to the "gate drive" circuit : Fig4. Under normal conditions, a current is generated in the photodiode due to light (around 10µA). Because of switchings in the converter, there are some perturbations that can be observed at the output of the receiver (digital). The electromagnetic field generates a current in the connections that are in series with the photodiode and parasitic signals are amplified by the transimpedance amplifier. To reduce these parasitic effects, a copper shield (100µm thick) is applied to the reciever and connected to ground : Fig5. This solution gives good results and has been tested at 150kV/µs without parasitic effects.

Sometimes, it is impossible to reduce the effects of common current and electromagnetic field. Therefore, voltage and/or current transients must be decreased by the drivers. Solutions to this problem are discussed in the following section. Physical mechanisms must first be highlighted in order to understand the way to control and decrease transients. So, in the next part, we study simple modellings of a well known power switch : the IGBT.

## 3. PHYSICAL APPROACH OF IGBT BEHAVIOR

The IGBT power component was developed in the early 80’s. Its development led to many improvements and typical structures. We can cite some of them that are well known : PT-IGBT (Punch Through IGBT), NPT-IGBT (No Punch Through), FS-IGBT (Field Stop), LPT-IGBT (Light Punch Through) and SPT-IGBT (Soft Punch Through). FS, LPT and SPT technologies come from different manufacturers but the main idea is the same : optimize the doping profile of the IGBT cell to reduce conduction and commutation losses.

The basic IGBT cell structure is similar to that of MOSFET : a P⁺ layer is added at the drain side of the MOSFET to create an IGBT, see Fig6. The purpose of this layer is to inject holes into N⁻ Substrat layer during conduction to decrease voltage drop ($V_{ceas}$) and therefore conduction losses (compared to MOSFET). The drawback is the increase of commutation time at turn-off : minority carriers are injected and the time for recombination increases. That is why the IGBT has a long turn-off time compared to MOSFET : the "current tail" is the well known electrical consequence of this phenomenon.

In [1] and [2], the IGBT structure is modeled by an associa-
tion of a MOSFET and pnp-bipolar transistors : Fig7. Equations developed in [2] consider the physical dimensions of IGBT cell. The IGBT model consists of three state equations considering the base-collector $v_{bc}$, the base charge $Q$ (base charge of the bipolar transistor : charge accumulation of the N$^-$ substrat layer modeled by the $R_n$ resistor) and the gate source voltage $v_{gs}$. These equations can be used in a circuit simulation software. Unfortunately, it is difficult to use them to obtain an analytical solution. The model considered must be simplified so as to lead to a simplified model.

Fig. 7. Equivalent model of IGBT cell : association of MOSFET and bipolar transistors

First, the static characteristic of IGBT cell are considered : Fig8. Two regions are identified : saturation and ohmic zones. In the ohmic zone, the equivalent model is a voltage source in series with a resistor :

$$V_{ce} = V_0 + R_c I_c$$

In the saturation zone, we consider the simple equation :

$$I_c = K \cdot f(v_{ge})$$

Different f function are commonly used :

$$f(v_{ge}) = (v_{ge} - v_{th}) [6]$$
$$f(v_{ge}) = (v_{ge} - v_{th})^2 [7]$$

Fig. 8. Static characteristics of IGBT cells

Dynamical properties of the IGBT behavior are modeled by equivalent capacitors : Fig9. Some of these present non-linear aspects and depend on $V_{bc}$ and $V_{ce}$ :

- $C_1$ : capacitance between gate and emitter, depends on oxyde thickness near gate and emitter contacts and cell shape.
- $C_2 - C_4$ : capacitances between gate and emitter (of P$^+$ layer), $C_2$ depends on oxyde thickness and cell shape, $C_4$ depends on depletion zone of N$^+$P junction.
- $C_3 - C_5$ : capacitances between gate and collector, $C_3$ depends on oxyde thickness and cell shape, $C_3$ depends on depletion zone of P/P$^+$N$^-$ junction.
- $C_6$ : capacitance between emitter and collector, depends on depletion zone of P/P$^+$N$^-$ junction.

So as to simplify transient behavior of IGBT, we consider that $C_{ge}$ is constant and does not depend of any voltage (we suppose that variations of $C_4$ are negligible). $C_{ce}$ and $C_{ge}$ equivalent capacitance depend on $V_{ce}$ and $V_{ge}$ respectively : $V_{ce}$ and $V_{ge}$ have an influence on the depletion zone. So as to simplify transient behavior of IGBT, we consider that $C_{ge}$ is constant and does not depend of any voltage (we suppose that variations of $C_4$ are negligible). $C_{ce}$ and $C_{ge}$ equivalent capacitance depend on $V_{ce}$ and $V_{ge}$ respectively : $V_{ce}$ and $V_{ge}$ have an influence on the depletion zone of P/P$^+$N$^-$ junction : Fig10.

The model proposed here, static characteristics and non-linear effects on equivalent capacitances, leads to easy understanding of transient characteristics of IGBT and requirements for drive circuits of IGBT. We consider this model and a buck converter with inductor load : Fig11.

In [8], equations for $di_c/dt$ and $dv_{ce}/dt$ are given for turn-off
and turn-on at the beginning of the commutations:

\[
\begin{align*}
\frac{dv_{ce,off}}{dt}(t = 0) &= \frac{V_{th} - I_L/g_m - v_{ee}}{R_e C_{gc}} \\
\frac{dv_{ce,on}}{dt}(t = 0) &= \frac{V_{th} + I_L/g_m - v_{ee}}{R_e C_{gc}} \\
\frac{di_{c,off}}{dt}(t = 0) &= \frac{v_{ee} - (V_{th} + I_L/g_m)}{R_e (C_{gc} + C_{ge})} + g_m L_s \\
\frac{di_{c,on}}{dt}(t = 0) &= \frac{v_{ee} - V_{th}}{R_e (C_{gc} + C_{ge})} + g_m L_s 
\end{align*}
\]

Where:

- \(R_e\) : gate resistor [ohm]
- \(V_{th}\) : gate threshold voltage [V]
- \(g_m\) : device transconductance [A/V]
- \(I_L\) : load current [A]
- \(C_{gc}\) : device gate-collector capacitance [F]
- \(C_{ge}\) : device gate-emitter capacitance [F]
- \(L_s\) : equivalent inductor at the emitter side [H]

Note that transients depend on:

- driver characteristics : \(v_{ee}\)
- power side : \(I_L\)
- device characteristics : \(V_{th}, g_m, C_{gc}, C_{ge}\) and \(L_s\)

So, to control transients, driver characteristics can be adapted. Device characteristics and IGBT current can not be adapted so as to control transients. In the next part, we expose published solutions in academic and industrial works.

4. PROPOSED SOLUTIONS

In order to control transients in power switches, the first idea would be to measure \(v_{ce}\) and \(i_c\) transients and create a local feedback as in classical automatic systems. This idea is developed in [9] to lead to two principles: \(di/dt\) and \(dv/dt\) controls (see Fig12).

For \(di/dt\) control, the device is controlled owing to \(v_{lee}\). The voltage across \(L_{ee}\) gives the \(di\)/\(dt\) information [3,8,10]. The \(L_{ee}\) inductance value depends on IGBT module structure and may be difficult to be measured. For \(dv/dt\) control, the measurement is made with a capacitor of few pico-Farad connected directly to the collector of the module. The current in the capacitor gives the image of \(dv_{ce}/dt\) [8].

In [8], the authors use a current generator so as to control transients. They use classical bipolar transistor to perform such current generator and implement these solutions on 1200V-70A IGBTs. The synopsis is given Fig13. \(dv/dt\) can be adapted from 5000V/\(\mu\)s to 1000V/\(\mu\)s at turn-on and turn-off with a 1200V-70A IGBT. Experimental and simulations results are compared. The method is also applicable to a wide range of \(di/dt\) : from 10A/\(\mu\)s to 500A/\(\mu\)s.

In [11], the turn-on controller is given in Fig14. The voltage \(v_g\) rises to \(u_{bias}\) that is near the threshold voltage of the IGBT. The voltage ramp \(\Delta U\) between \(t_1\) and \(t_2\) influences the turn-on transient of the IGBT. The current slope \(di_c/dt\) and \(di_d/dt\) are reduced. According to the authors, the turn-off passive or active diode snubber can be minimized or even omitted. The parameters \(\Delta U\), \(\Delta T\) and \(u_{bias}\) can be adjusted depending on the IGBT and diode. But as there is not feedback, the \(di_c/dt\) only depends on parameters that are set by users.
In [12], the authors consider the turn-on of IGBT on an inductive load. The aim is to reduce power dissipation between IGBT and the power diode. The IGBT current can be optimized by the control of gate resistance during switching operations. The optimization has been performed on 1200A-3300V IGBT power modules. Simulations and experimental results are performed. With the help of three resistor values, the turn-on is controlled so as to reduce the diode recovery current and so the power dissipation in IGBT and diode without decreasing the voltage slope. A energy reduction of about 20% is accomplished using this solution. In Fig15, the principle of the three gate resistor method is given. To start the switching turn-on, a middle resistor value is used to perform the beginning of the collector current increase (consequently the diode current decrease). To slow down the diode current transient in order to limit the recovery current, the gate resistor value is increased. After the maximum IGBT current, the gate resistance is reduced to accelerate the IGBT $V_{ce}$ voltage to reduce power loss.

In [9], the authors introduce the $di/dt$ measurement with the help of emitter inductance. The voltage across the power emitter connection and emitter signal give the image of $di/dt$: see Fig12. In Fig16, the model introduced in [5] is given. The problem is to estimate the emitter inductance value: $L_7$. For example, we consider the IGBT power module Eupec FZ1200R33KF2 in Fig17.

Two solutions to estimate $L_7$ are proposed [5]. The first was done using the sofware InCa$^\text{TM}$. The 3D structure has been des-
Inductance modelling of IGBT power modules, example of Eupec FZ1200R33KL2C 3300V-1200A

Internal structure of IGBT power module FZ200R33KF2 described with simplifications: only one power emitter connexion is considered, see Fig18 and Fig19. InCa™ software is based on the PEEC (Partial Element Equivalent Circuit) method to calculate inductance. Here, another solution to estimate \( L_7 \) is proposed: a measurement of \( V_{ce} \) and \( I_c \) during turn-on leads to the identification of \( L_7 \) by the formula: \( L_7 = \frac{V_{ce}}{I_c} \). In Fig20, switching waveforms are given for a FZ1200R33KF2 Eupec module at turn-on. The inductance \( L_7 \) is estimated at about 3nH. Results with InCa™ are different from experimental results. The following table summarizes the results.

<table>
<thead>
<tr>
<th>Module Type</th>
<th>InCa™</th>
<th>Experimental</th>
</tr>
</thead>
<tbody>
<tr>
<td>FZ1200R33KF2</td>
<td>6.5nH</td>
<td>3nH</td>
</tr>
<tr>
<td>Results for a CM1200HB66H IGBT power module taken from Mitsubishi:</td>
<td></td>
<td></td>
</tr>
<tr>
<td>InCa™</td>
<td>9nH</td>
<td>5nH</td>
</tr>
<tr>
<td>Experimental</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Note that simulation results are quite different from experimental ones. InCa software gives approximated values due to the simplifications explained before.

Estimation of \( L_7 \) from switching gives a more realistic value. \( L_7 \) depends on the module type and manufacturer. A comparison of two 1200A-3300V IGBT power modules CM1200HB66H (Mitsubishi) and FZ1200R33KF2 (Eupec) gives quite different values: 5nH and 3nH respectively. Configurations of IGBT drivers that control \( di/dt \) must be adapted to the IGBT power module. The estimation of \( L_7 \) must be updated often. IGBT module manufacturers can modify internal structures without notifying consumers. Value for \( L_7 \) would certainly change and \( di/dt \) control would not be op-
timed and could be unusable. This is the main drawback of di/dt control.

5. CONCLUSION

An original approach to technical problems concerning electromagnetic perturbations generated by power converters and coupling effects on surrounding electronics is proposed. The problem originates from voltage and current transients and the main idea is to control dv/dt and di/dt on power switches. In order to do so, modelling of gate power switches and more specifically IGBT is proposed.

The description of a classical IGBT cell is given to describe both its static and dynamic behavior. Beginning with physical models [1,2], some simplifications are considered that lead to an easy to use IGBT model. Equations for dv/dt and di/dt are given for the case of a buck converter with inductive load. These equations highlight the way to control voltage and current transients. Focus is made on drive techniques based on voltage and current feedback for gate control, and on methods based on open loop drive techniques (voltage ramp on $v_{ge}$ and three gate resistors technique).

To finish, this article is about techniques to estimate emitter inductance used in di/dt control. Comparison is made between software and switching methods. Some results are given for two IGBT power modules.

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