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

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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 (conresponding 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. Focus is made on inductance estimation of IGBT power modules necessary for di/dt control : expermental 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 clos 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 impuled 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 $\pm 15V$ 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.

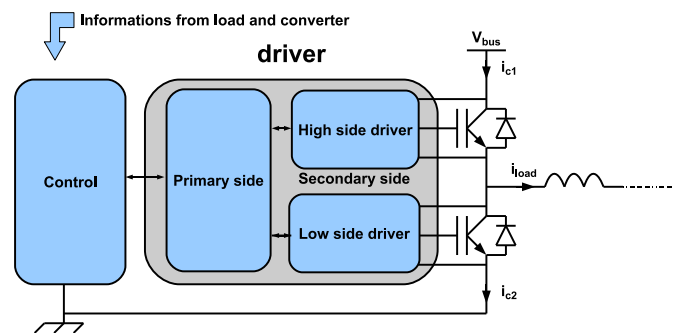


Fig. 1. Inverter leg with control and driver

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

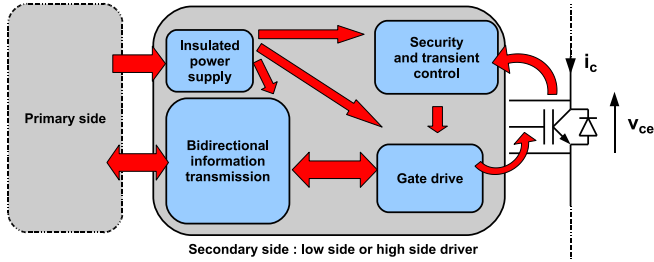


Fig. 2. Detail of secondary side of gate driver

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 $1\text{kV}/\mu\text{s}$, we have $i_{cm} = C_p \cdot \frac{dV}{dt} = 100e^{-12} \cdot 1e^3 / 1e^{-6} = 100\text{mA}$ current. This current can easily perturb electronic components such as operational amplifiers for measurements or micro-controller and DSP for the converter control.

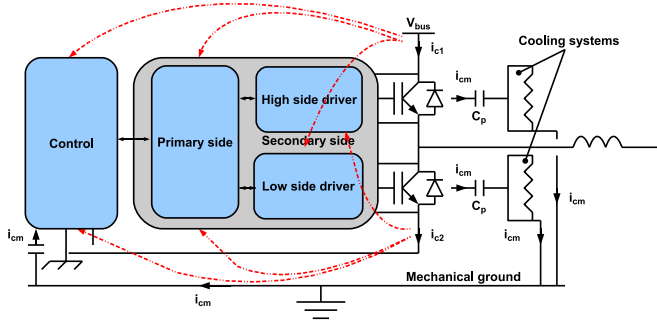


Fig. 3. Identification of problems in power converters, common mode current, electromagnetic field emissions and couplings

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 $10\mu\text{A}$). 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 ($100\mu\text{m}$ thick) is applied to the receiver and connected to ground : Fig5. This solution gives good results and has been tested at $150\text{kV}/\mu\text{s}$ 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

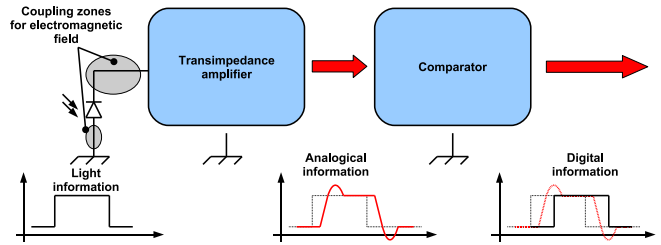


Fig. 4. Photodiode, transimpedance amplifier and comparator

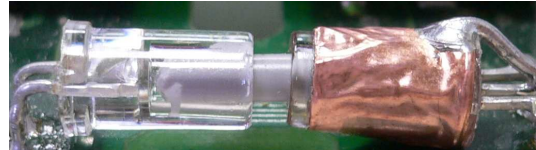


Fig. 5. Electromagnetic shield of photodiode receiver

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_{cesat}) 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.

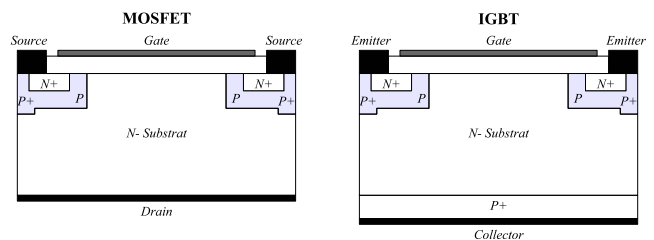


Fig. 6. MOSFET and IGBT cells

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^- substrate layer modeled by the $R_n(mod)$ 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.

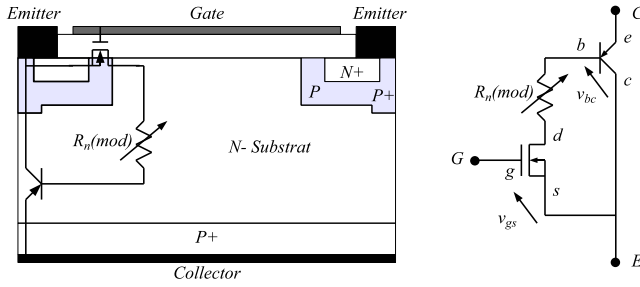


Fig. 7. Equivalent model of IGBT cell : association of MOSFET and bipolar transistors

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 :

$$V_{ce} = V_0 + R.I_c$$

In the saturation zone, we consider the simple equation :

$$I_c = K.f(V_{ge})$$

Different f function are commonly used :

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

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

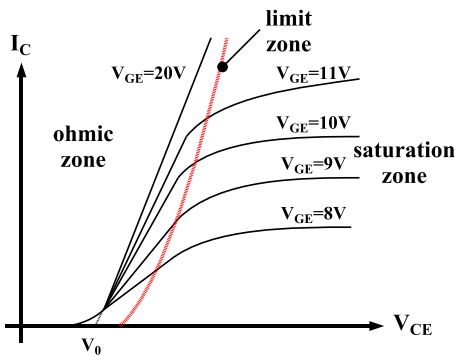


Fig. 8. Static characteristics of IGBT cells

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

- C_1 : capacitance between gate and emitter, depends on oxide 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 oxide 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 oxide thickness and cell shape, C_5 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.

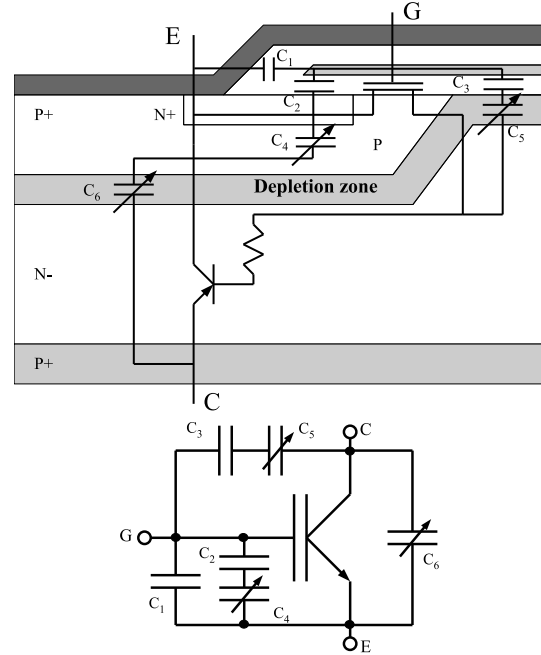


Fig. 9. Equivalent capacitors of IGBT cell with non-linear effects

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_{gc} equivalent capacitance depend on V_{ce} and V_{gc} respectively : V_{ce} and V_{gc} 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_{gc} equivalent capacitance depend on V_{ce} and V_{gc} respectively : V_{ce} and V_{gc} 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

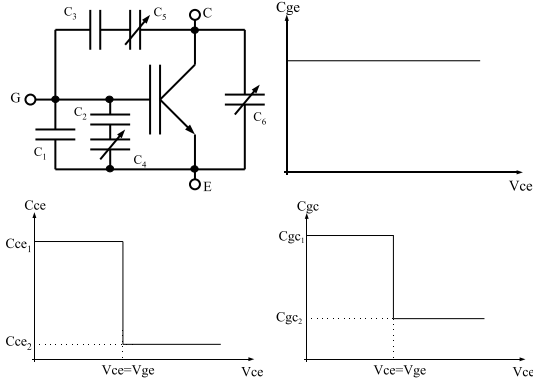


Fig. 10. Equivalent circuit of IGBT cell, non linear effects and simplifications, evolutions of C_{gc} , C_{ge} and C_{ce}

and turn-on at the beginning of the commutations :

$$\begin{aligned} \frac{dv_{ce,off}}{dt}(t=0) &= \frac{V_{th} - I_L/g_m - v_{ee}}{R_g \cdot C_{gc}} \\ \frac{dv_{ce,on}}{dt}(t=0) &= \frac{V_{th} + I_L/g_m - v_{cc}}{R_g \cdot C_{gc}} \\ \frac{di_{c,off}}{dt}(t=0) &= g_m \cdot \frac{v_{ee} - (V_{th} + I_L/g_m)}{R_g \cdot (C_{gc} + C_{ge}) + g_m \cdot L_s} \\ \frac{di_{c,on}}{dt}(t=0) &= g_m \cdot \frac{v_{cc} - V_{th}}{R_g \cdot (C_{gc} + C_{ge}) + g_m \cdot L_s} \end{aligned}$$

Where :

- R_g : 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_{lee2} . The voltage across L_{ee2} gives the di_c/dt information [3,8,10]. The L_{ee2} 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/ μ s to 1000V/ μ 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/ μ s to 500A/ μ 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 ΔU between t_1 and t_2 influences the turn-on transient of the IGBT. The current slope di_c/dt and di_{d1}/dt are reduced. According to the authors, the turn-off passive or active diode snubber can be minimized or even omitted. The parameters ΔU , ΔT 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.

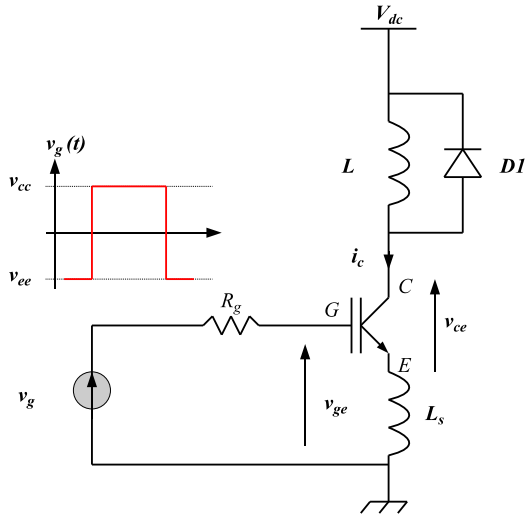


Fig. 11. Buck converter, IGBT and diode commutation cell

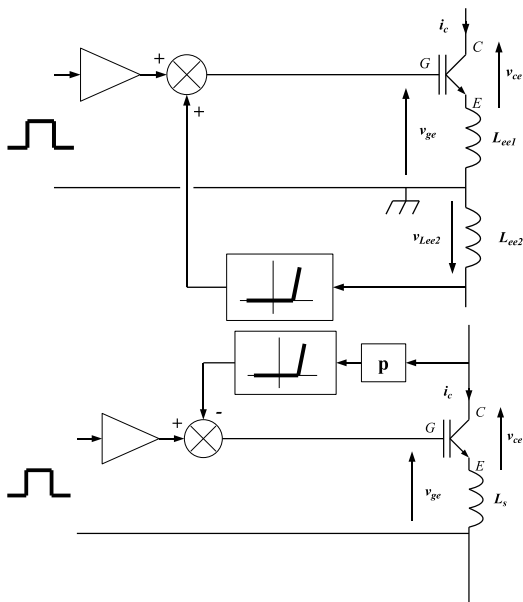


Fig. 12. Principle of di/dt and dv/dt controls

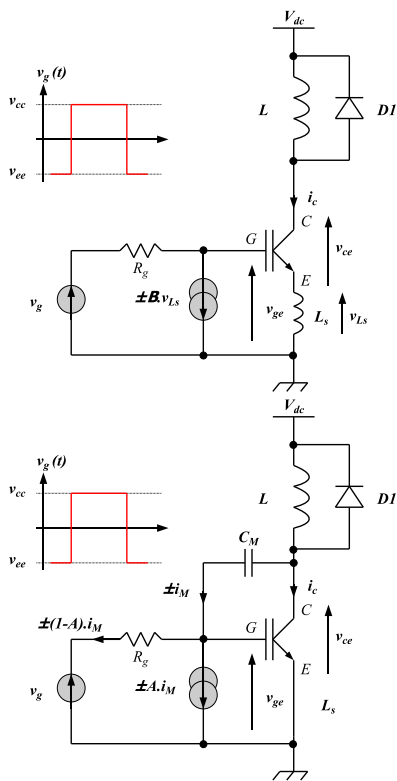


Fig. 13. di/dt and dv/dt controls with current generators

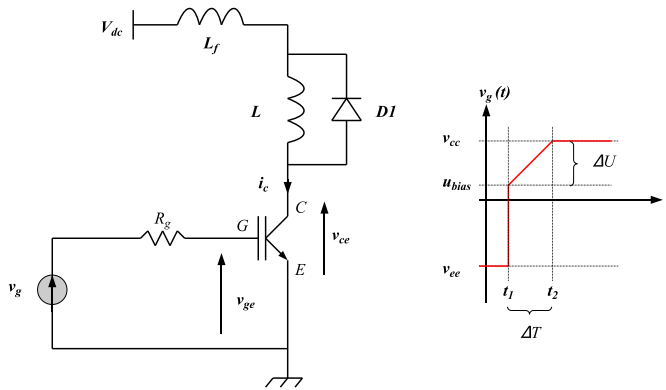


Fig. 14. Turn-on control with a voltage ramp

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.

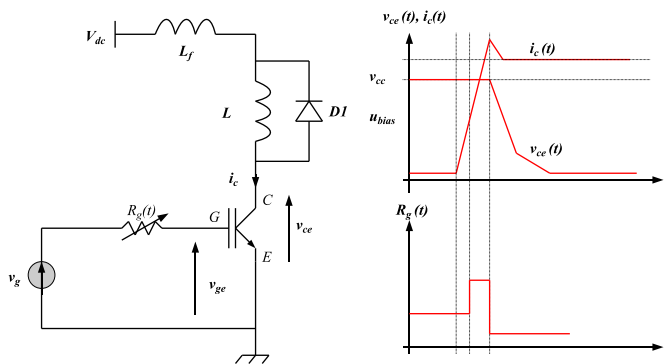


Fig. 15. Gate driver with three resistor value at turn-on to reduce diode recovery current and accelerate v_{ce} decrease

In [9], the authors introduce the di_c/dt measurement with the help of emitter inductance. The voltage across the power emitter connection and emitter signal give the image of di_c/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 software InCaTM. The 3D structure has been des-

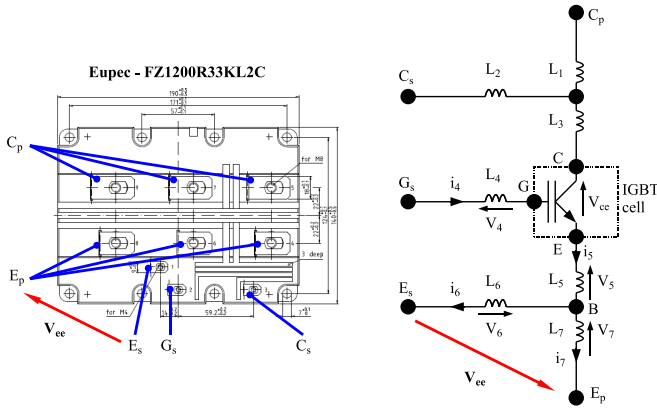


Fig. 16. Inductance modelling of IGBT power modules, example of Eupec FZ1200R33KL2C 3300V-1200A

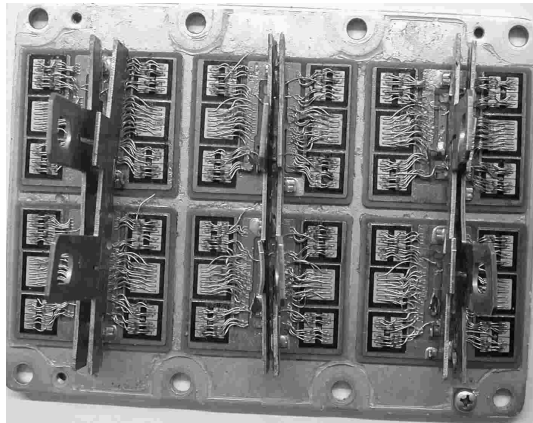


Fig. 17. Internal structure of IGBT power module FZ200R33KF2

cribed with simplifications : only one power emitter connexion is considered, see Fig18 and Fig19. InCaTM software is based on the PEEC (Partial Element Equivalent Circuit) method to calculate inductance. Here, an other solution to estimate L_7 is proposed : a measurement of V_{ee} and I_c during turn-on leads to the identification of L_7 by the formula : $L_7 = \frac{V_{ee}}{di_c/dt}$. 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 InCaTM are different from experimental results. The following table summarizes the results.

FZ1200R33KF2 :
 InCa : 6.5nH
 Experimental : 3nH

Results for a CM1200HB66H IGBT power module taken from Mitsubishi :
 InCa : 9nH
 Experimental : 5nH

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

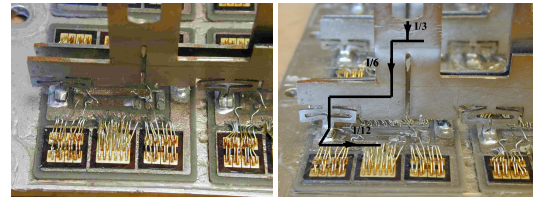
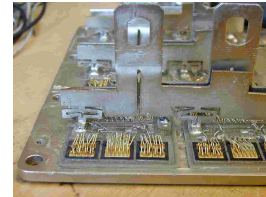
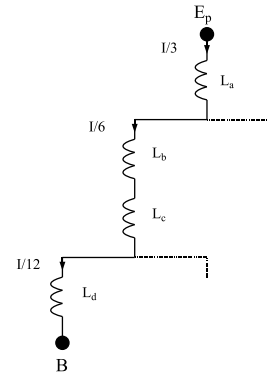


Fig. 18. Simplification of module structure

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-

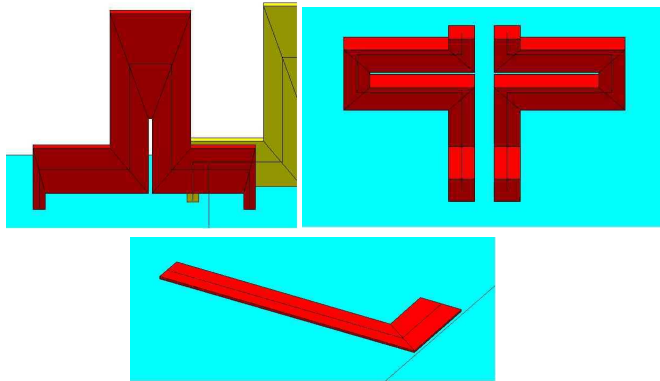


Fig. 19. InCa description of simplified power module

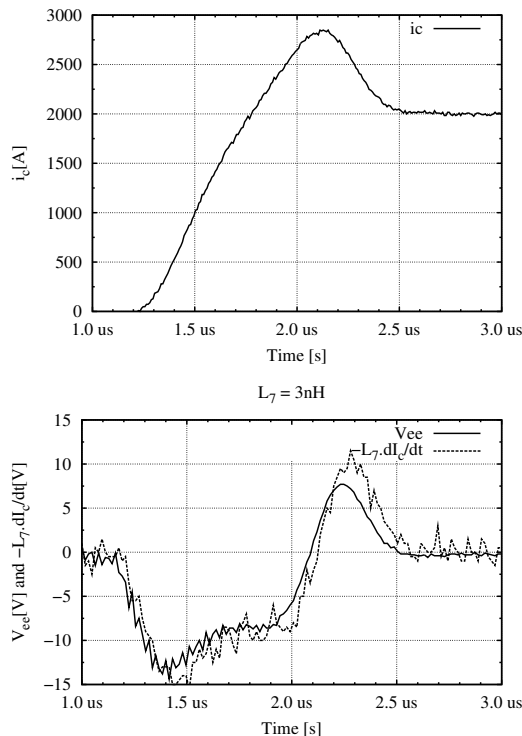


Fig. 20. IGBT power module FZ1200R33KF2, estimation of L_7 at turn-on

timized 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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