PCIM -93

COMPARISON OF ELECTROMAGNETIC PERTURBATIONS AND SWITCHING LOSSES BETWEEN SELF-SWITCHING AND PWM MODE OF CURRENT CONTROL IN A SWITCHED RELUCTANCE MOTOR.

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Abstract

In a previous paper [1] we have shown the interest of the bifilar winding in doubly salient reluctance motor feed by low voltage (automotive application). The interest is the minimization of switches number and of silicium area. In this paper, we propose to continue the study of a new control method with this feeding process structure (bifilar winding with one switch per phase [2]).
This method minimizes electromagnetic perturbations (by reducing current gradient), commutation losses of switches, and allows the use of standard diodes instead of fast recovery diodes.
We will compare this new control mode with the PWM one. We will show, for example, at the same average torque value, that switching commutation losses are notably lower than in PWM control and that total losses of the motor+converter set remains constant.

1. INTRODUCTION

The aim of this study is to design an economical static converter for a low power switched reluctance motor (0.1Nm 3000 rpm) in an automotive application. The following constraint had to be considered:

- low level of conducted electromagnetic perturbations.
- low level of switch losses.
- simple and low cost electronic circuits.

1.1. MACHINE AND CONVERTER CHARACTERISTICS

The studied machine is doubly salient structure having 3 phases, 6 stator poles and 4 rotor poles (figure 1):

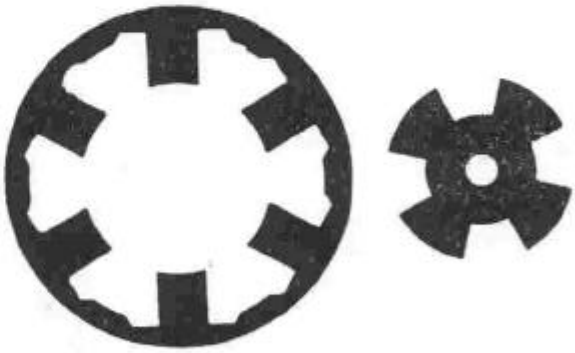


figure 1 : motor structure: 6 stator poles, 4 rotor poles

In our application, the air gap length is an important parameter and its value is large. Therefore the magnetic circuit is never saturated (I sat≈20A).

So, the motor torque for one phase can be written as follow:

$$C(\theta) = \frac{1}{2} \cdot \frac{L(\theta)}{d\theta} \cdot I_1^2$$

where θ is the electric angle, $L(\theta)$ is the inductance versus angle and I1 is the phase current.

Then, two machine-converter structures are possible:

The first one is to associate each motor phase with an asymmetric half bridge (figure 2):

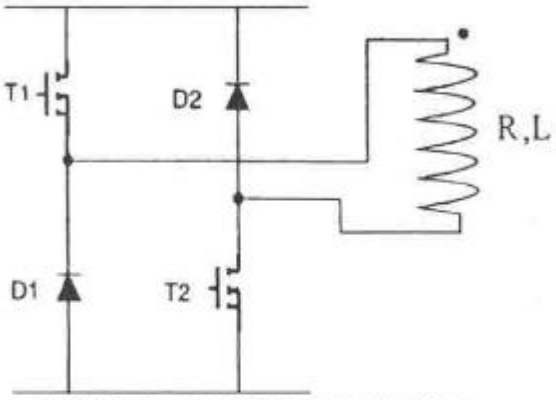


figure 2: example of power converter with an asymmetric half bridge per phase.

The second one is to associate the power converter with a bifilar winding machine (figure 3).

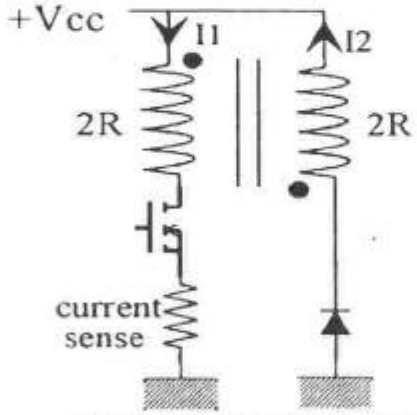


figure 3: example of bifilar machine with it's power converter.

This economical power converter has only one switch per phase but at the turning off, the voltage across the MOS transistor may rise to over twice the supply voltage plus an overvoltage due to leakeage inductance of coil. The main disadvantage of this structure lies in the inefficient use of the copper in the machine. Only one of the bifilar windings carries current at any time (magnetizing or demagnetizing moment). We will compare these two structures considering that copper volume and silicium are stay constant. An other disadvantage is the increasing of switching losses because of storage energy in leakage inductance of phases.

1.2. COPPER LOSSES IN THE MACHINE

At first, we consider the copper volume is constant. R designates the total winding resistor for a single coil phase. 2R will designates one winding resistor in case of a bifilar one (figure 3). (the two machine have the same flux ampere.turn characteristics).

1.2.1. ASYMMETRICAL HALF BRIDGE

To compare copper losses, we will consider the nominal feeding mode corresponding to a full wave voltage. (usually adopted in high speed conditions). The corresponding idealised phase current wave-forms are presented in figure 4.

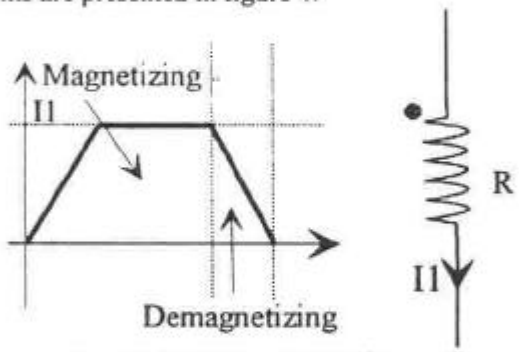


figure 4: current modelling. Machine fed by a full wave voltage.

The sum of copper losses is:

$$P_{ju} = q.R.I_{eff}^2$$

where q is the phase number.

1.2.2. BIFILAR WINDINGS

The corresponding current wave-forms are presented in figure 5 (same supply conditions than figure 4).

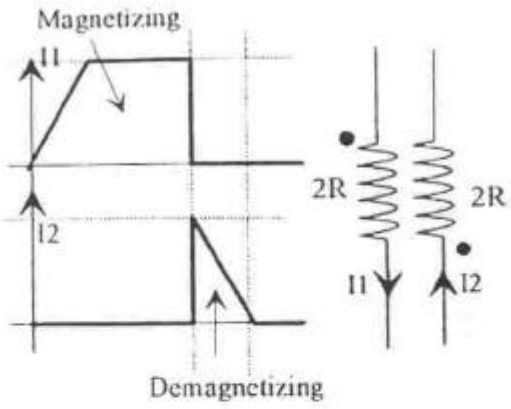


figure 5: current modelling (bifilar windings). Machine fed by full wave of voltage.

This time the sum of copper losses

$$P_{jbif} = q.2.R.(I_{1eff}^2 + I_{2eff}^2)$$

is:

$$P_{jbif} = 2.Pju$$

In bifilar structure, copper losses are doubled in comparison with a single coil machine.

1.3. MOS CONDUCTION LOSSES

1.3.1. ASYMETRICAL HALF BRIDGE

In our automotive application, we have a low voltage power supply (12V). Transistor are in MOS technology. In the case where BVds<60V and if the silicium area is the same in the two structures, we can write the MOS conduction losses as follow:

$$P_{mos\ uni} = 2.q.(2.R_{ds\ on}).I_{magnetisationeff}^2$$

where R_{dson} is the MOS conducting resistor for silicium used inone phase converter (each transistor has one half of total area).

1.3.2. BIFILAR WINDINGS

$$P_{mos\ bif} = q.R_{ds\ on}.I_{magnetisationeff}^2$$

$$P_{mos\ uni} = 4.P_{mos\ bif}$$

So in asymetrical half bridge the conduction losses are multiplied by 4.

1.3.3. CONCLUSION

For the same copper volume and the same silicium area in the switching circuit, none of the above circuits could be considered ideal. In our automotive application, it's important to minimize the conduction losses because of the low voltage feeding (12V) and to minimize the number of switches for economical reasons.

At nominal speed (full wave contrql), we have computed copper and conduction losses at the same global efficiency in the twice structure. At rated point speed is 3000 rpm and torque is 0.1Nm . MOS are IRFZ 44 (60V, R_{dson}=28mΩ) and diodes are BYW81 PI 200 (200V, 15A).

Despite of the inefficiency use of copper in the bifilar windings, we choose this last structure for economical and reliability reasons.

We obtain the converter of figure 6. For economical reasons, the current measure is performed by means of a 0.05Ω resistor connected to the magnetizing circuit.

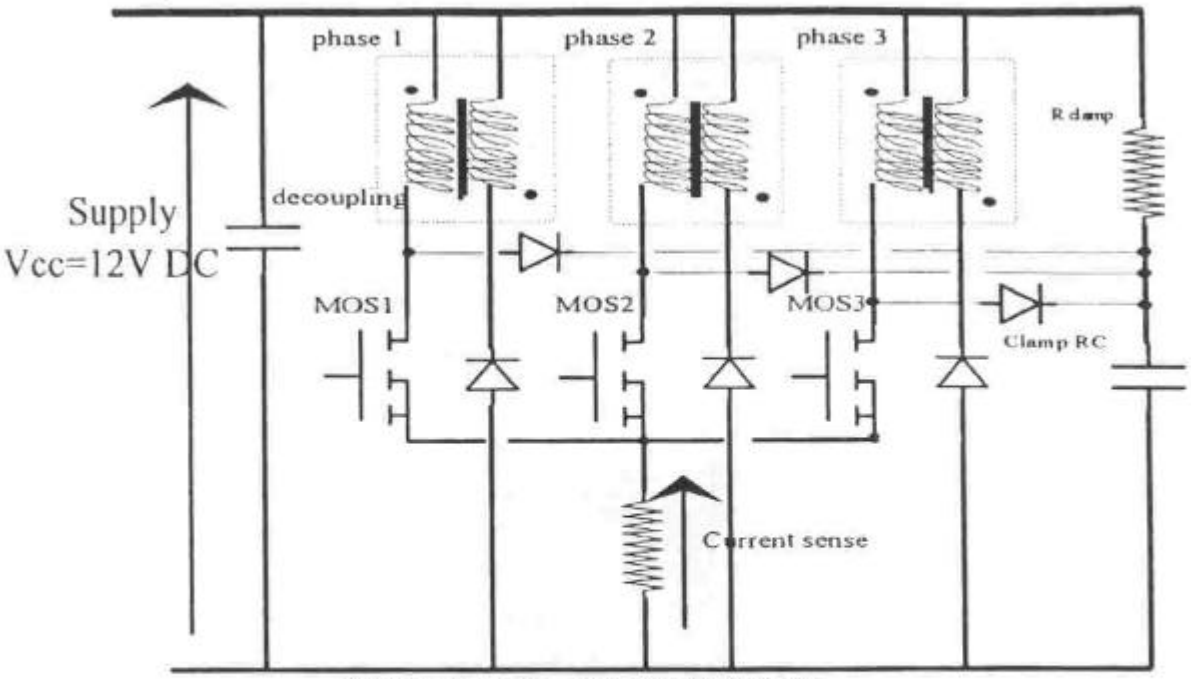


figure 6 : scheme of the power converter finely used.

2. CONTROL METHOD

2.1. PWM MODE

At low speed, to control the torque, we have to control the current magnitude in the phase winding.

If we use a PWM method because of the leakage flux in the bifilar windings, switching losses may be important. To avoid, in a first time, the transistor breakdown we use a clamp circuit (resistor, capacitor diode common to three phase. In fact, at the MOS turn off instant, the electrical equivalent circuit can be represented on figure 7.

In any case we observe two supply conditions:

2.1.1. Energizing coil:

$$V_{cc} = (R_{ds\,on} + R_{sense} + 2R).I_1 + L(\theta).\frac{dI_1}{dt} + I_1.\omega.\frac{dL}{d\theta}$$

2.1.2. Unernergizing coil:

$$-V_{cc} = (2R).I_2 + L(\theta).\frac{dI_2}{dt} + I_2.\omega.\frac{dL}{d\theta} + V_d$$

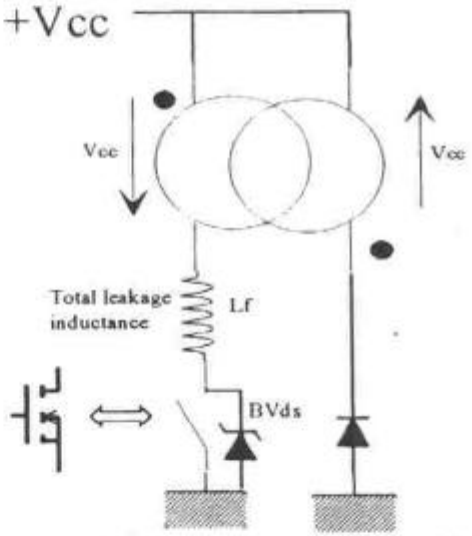


figure 7: model of MOS tuning off configuration. The snubber is designed by a zener diode.

The snubber limits the MOS turning off voltage to V_{br} value ($Vbr<BV_{ds}$) The snubber is so calculated to obtain:

$$V_{br} = 2.V_{cc} + \Delta V$$

this leads to

$$R_{clamp} = \frac{2.\Delta V.(V_{cc} + \Delta V)}{f.I_{max}.L_f}$$

$$C_{clamp} = \frac{I_{max}^2 . L_f}{a . \Delta V^2}$$

Lf is the total leakage inductance and a.ΔV the voltage ripple across C.

The corresponding loss power in the resistor clamp is:

$$P_{off} = \frac{1}{2} . L_f . I_{max}^2 . F . \frac{1}{1 - \frac{2 V_{cc}}{V_{bl}}}$$

2.2. SELF SWITCHING MODE

The problem of the PWM mode is the excess of switching losses in transistors and diodes because of hard commutation conditions. Also, the commutation generates conducted electromagnetic perturbations.

The new control method has to realise a soft diode turn off. Indeed, if the diode turn off is performed with a low $\frac{dI_f}{dt}$, the recovery charge is reduced.

This can be performed by scanning the zero voltage switching of the diode (example of control scheme is presented on figure 8). We obtain in fact a ZVS commutation.

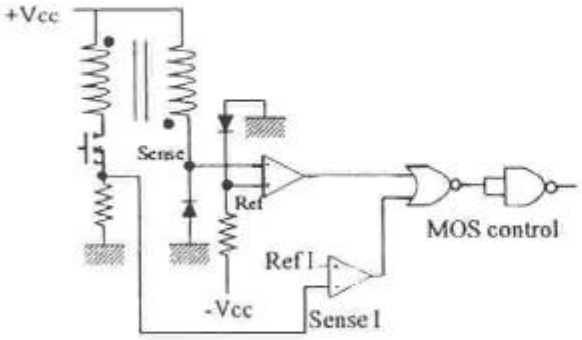


figure 8 : example of control circuit for a ZVS commutation in the diode

3. DIODES SWITCHING LOSSES

3.1. PWM MODE

In this control mode, the diode turn off can be modelled by the following wave-forms (figure 9).

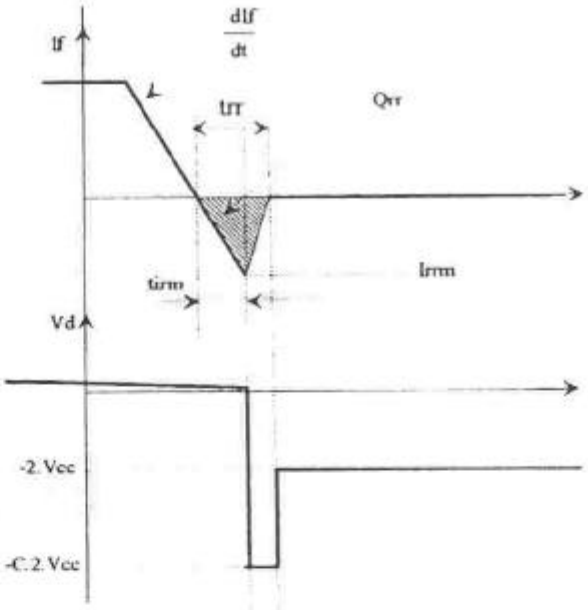


figure 9: model of diode turn off (unernegizing phase). C is the over voltage coefficient. (currently C=3). Here we suppose that 2.C.Vcc<Vrrm.

The magnetic energy is instantaneously transferred from the secondary to the primary coil. So the diode parameter for its commutation are:

$$-(2.V_{cc} + Vd) \approx L_f . \frac{dI_f}{dt}$$

$$\frac{dI_f}{dt} = -\frac{(2.V_{cc} + Vd)}{L_f} \approx -12A/\mu s$$

(Lf=2μH)

This leads to a reverse recovery charge: Qrr=18nC at Tj=100°C.

In case of PWM, the diode turn off energy is

$$W_{doff} = 2.Vcc.Q_{rr}$$

3.2. SELF SWITCHING MODE

If we decide to hold off the MOS until the complete demagetization, the $\frac{dI_f}{dt}$ term can be term can be modelled by:

$$\frac{dI_f}{dt} = -\frac{2.Vcc + Vd}{L(\theta) + L_f}$$

where L(θ) is the main inductive parameter of the machine. In the worst case, (unaligned position L(θ)=Lo=0.3mH) this leads to:

$$\frac{dI_f}{dt} = -80\ 10^{-3}\ A/\mu s$$

Then we can reasonably consider that Qrr can be missed comparing to the above case (12A/µs).
So we can project the use of standard diodes (low cost). The second interest of this control mode is to limit the switching frequency. However, the low frequency range of commutations leads to an audible noise witch can be a nuisance and the ripple torque is increased.

We present a computing simulation for the current wave-forms in the case of PWM and self switching control. The control parameter is the current reference applied (figures 10 and 11). In order to simplify the PWM control, we choose to hold off the MOS during a constant time (Toff=10µs) and the current magnitude is controlled by scanning the magnetizing current (voltage across Rsense figure 6). The angular supply window is 120° for each phase.

3.3. CURRENT WAVE-FORMS AT LOW SPEED 500rpm

3.3.1. PWM MODE

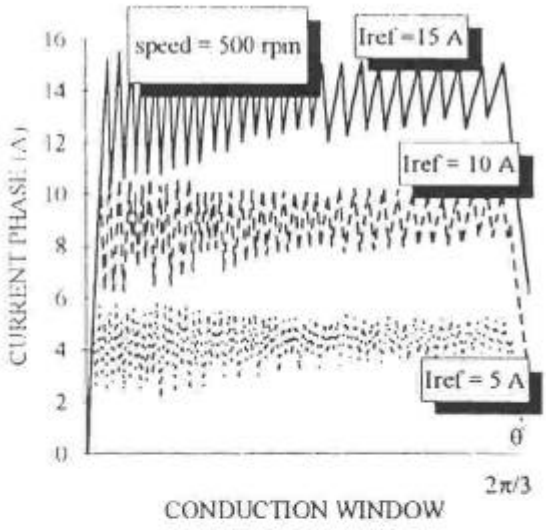


figure 10 : computing simulation of current wave-forms in PWM mode with Toff=cst=10µs and speed=500rpm; different current references from 5A to 15A.

3.3.2. SELF SWITCHING MODE

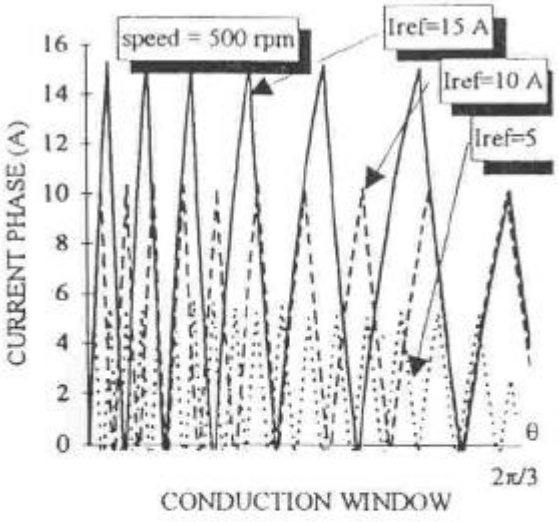


figure 11: computing simulation of current wave-forms in SELF SWITCHING mode with speed=500rpm; different current references from 5A to 15A.

3.4. CURRENT WAVE-FORMS AT HIGH SPEED :3000rpm

3.4.1. PWM MODE

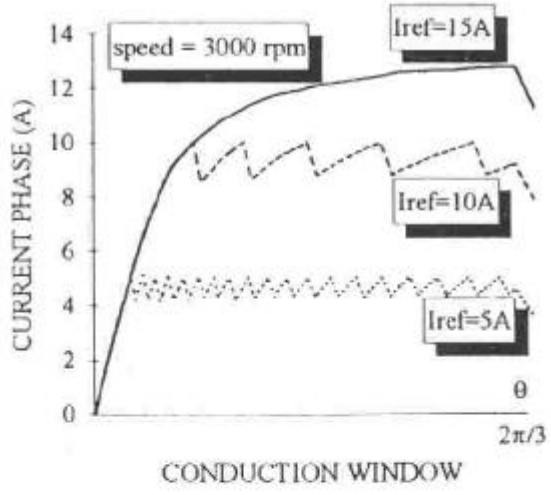


figure 12 :computing simulation of current wave-forms in PWM mode with Toff=cst and speed=3000rpm; different current references from 5A to 15A.

3.4.2. SELF SWITCHING MODE

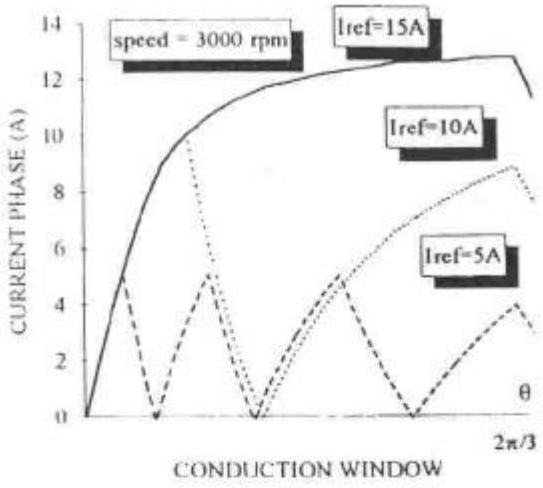


figure 13: computing simulation of current wave-forms in SELF SWITCHING mode and speed=3000rpm; different current references from 5A to 15A.

We can observe on the above figures (12 and 13) that the nominal steady point (full wave of voltage) is reached near a current reference of 12A. Then, the two current controls become similar. This can be shown if we use a trapeizoidal wave model for the growth of the inductive parameter we obtain the following formulae:

$$\frac{dL(\theta)}{d\theta} \approx \frac{\Delta L(\theta)}{\Delta\theta} = \frac{L_{aligned} - L_{unaligned}}{\frac{2.\pi}{3}}$$

in our case:

$L_{aligned}$=1.2mH and $L_{unaligned}$=0.3mH

So at N=3000 rpm (Ω=314 rad/s et Imax=12A)

$$R_\omega = \omega . \frac{dL(\theta)}{d\theta} \approx 0.537\Omega$$

$$I_M = \frac{U_m}{R_\omega + R_s}$$

where Rs designate the stator coil resistor (0.3Ω).

4. COMPARISON OF SWITCHING LOSSES.

4.1. DIODES

4.1.1. Switching on losses.

For the two supply mode we can consider that switching on commutation losses in the diode can be missed. Indeed, in PWM mode (worst case of losses) the $\frac{dI_f}{dt}$ term is low enough to reasonably missed this kind of losses. This can be twice verified in the self switching mode.

4.1.2. Switching off losses

PWM: we obtain in that case the following formula:

$$W_{doff} = 2.Vcc.Q_{rr}$$

SELF SWITCHING:

$$W_{doff} \approx 0$$

In fact, the comparison seems to be very simple: Self switching mode eliminate switching off losses thanks to the low rate of the corresponding Qrr. We can observe this results on the real commutations wave-forms in the two control mode on figure 14, PWM and figure 15 SELF-SWITCHING.

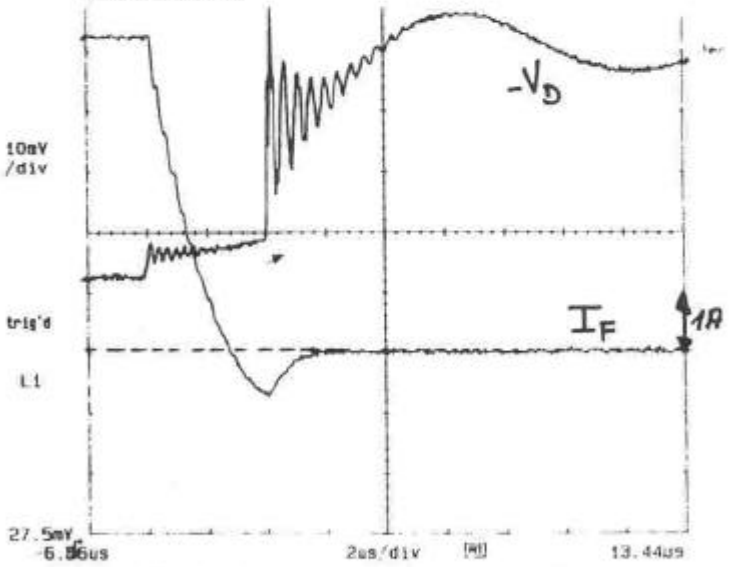


figure 14: real commutation wave-forms in the PWM mode. (fast recovery diode BYW81PI 200). Average current=5A.

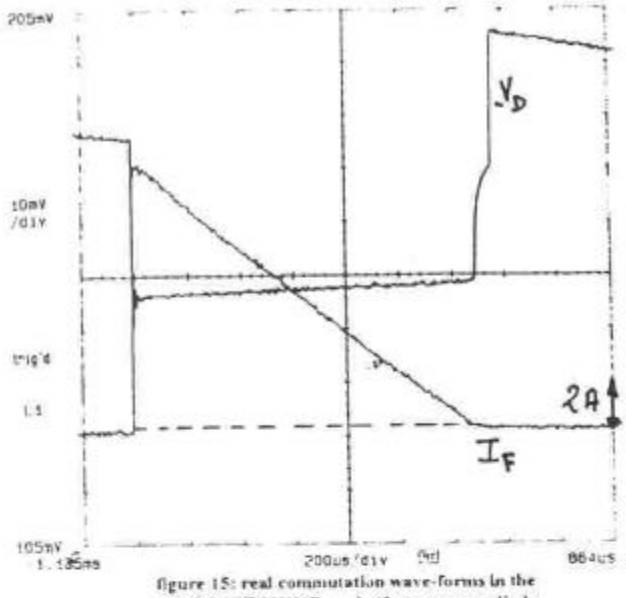
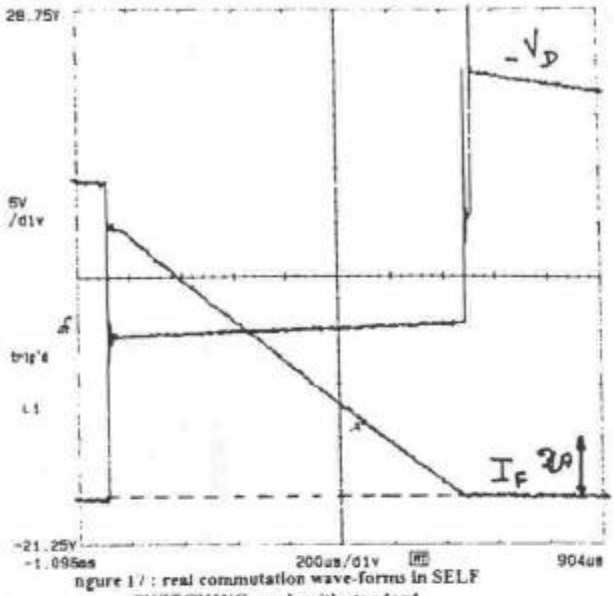


figure 15: real commutation wave-forms in the SELF SWITCHING mode (fast recovery diode BYW 81 PI 200). Average current=5A.



ngure 1 / : real commutation wave-forms in SELF SWITCHING mode with standard diodes(1N5408); Average current=5A.

We can observe that the reverse recovery charge decrease thanks to the low dIf/dt in the self-switching mode. In that case we can use standard diode instead of fast recovery one (1N5408 for example). We also present on figures 16 and 17 the same commutation wave-forms using standard diodes instead of fast one. In classical PWM mode, switching with standard diodes is very bad, the losses are excessive.

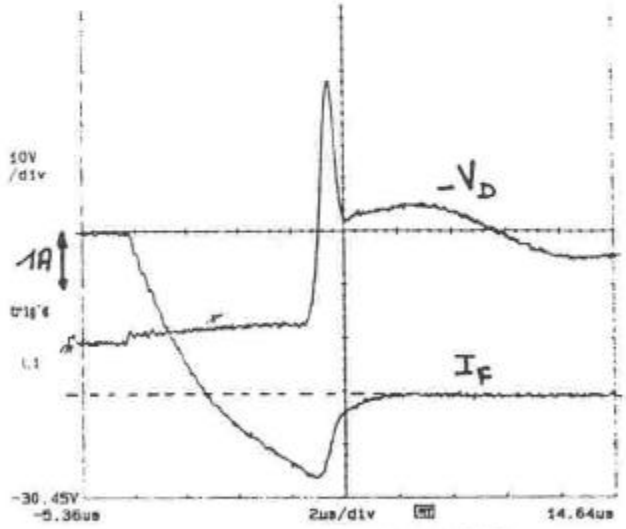


figure 16 : real commutation wave-forms in PWM mode with standard diodes (1N5408). Average current=5A.

Now the difference between the two reverse recovery charge is more evident. This difference is meanly dependant of the $\frac{dI_f}{dt}$ value witch is limited by Lf in PWM and Lo in self switching worst case.

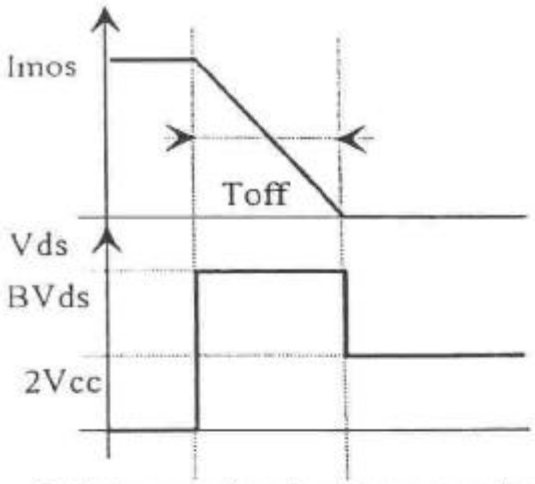
4.2. MOS

4.2.1. Switching on losses

We also consider that this kind of losses can be missed in the two mode of supply because of the quickness of the MOS switching on (less than 300ns) and the serial inductanec Lf. They will be lower in the Self switching mode because of the ZVS commutation. In fact the current increasing is limited by the mean of the inductive parameter of the stator coil corresponding to a very low $\frac{dI}{dt}$ rate.

4.2.2. Switching off losses

Without clamping circuit, a breakdown voltage appears across the MOS (corresponding to BVds). If we want to study this kind of losses, we adopt the following model:



In the two case of supply mode we can write:

$$t_{off} = \frac{L_f \cdot I_{ds\,max}}{BV_{ds} - 2.Vcc}$$

$$W_{off} = V_{ds} \cdot I_{ds} \cdot t_{off} = \frac{1}{2} \cdot B \cdot V_{ds} \cdot I_{ds\,max} \cdot t_{off}$$

$$W_{off} = \frac{1}{2} L_f \cdot I_{ds\,max}^2 \cdot \frac{1}{1 - \frac{2.Vcc}{BV_{ds}}}$$

An example of the real MOS commutation is shown on figure 18.

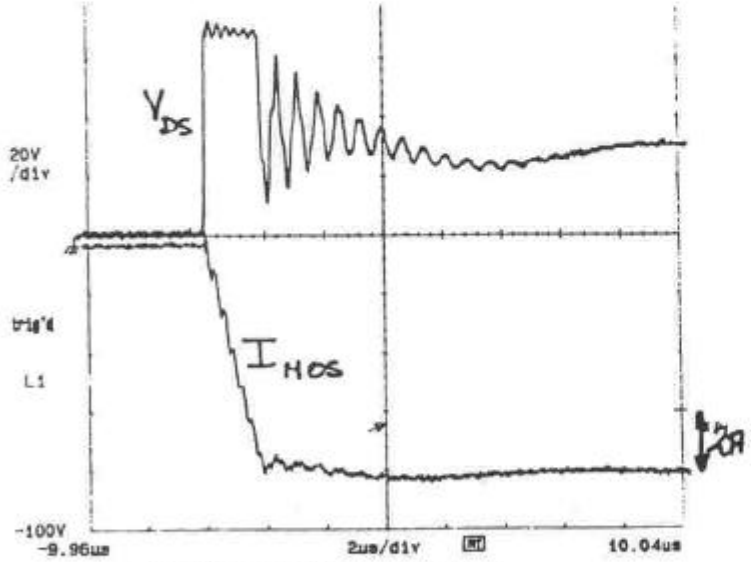


figure 18: real commutation wave-forms commutation for the MOS; Average current=5A

The same commutation wave forms can be observed in the Self switching mode. The number of commutations between the two supply mode is lower in the self switching one at the same rms phase current.

This can verified on figures 19 and 20 where we compare the sum of the MOS switching losses for the two sort of supply for different average torque value (figure 19 refers to 500 rpm and figure 20 refers to 3000 rpm).

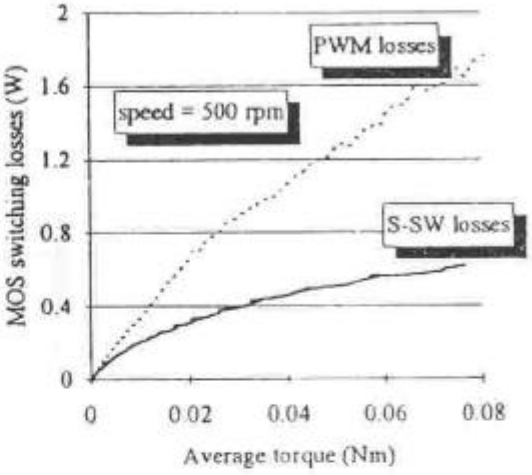


figure 19 : comparison of MOS switching losses between PWM and SELF SWITCHING mode (with fast recovery diodes) at 500 rpm.

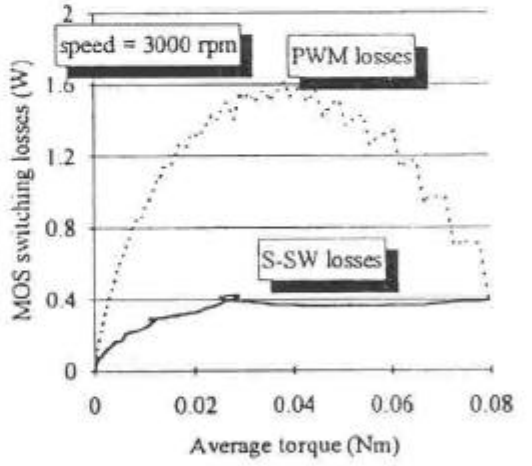


figure 20 : comparison of the MOS commutation losses between PWM and SELF SWITCHING mode (with fast recovery diodes) at 3000 rpm.

It is interesting to remark the junction of the two curves on figure 20. This conveys the full wave voltage mode in high speed. Then the two supply mode become equivalent and so there is no more difference of commutation losses in the MOS.

5. TORQUE CHARACTERISTIC

In the case of our low saturated machine we can consider that the torque value is expressed by the following formula:

$$C_{em}(\theta) \approx \frac{1}{2} \cdot \frac{dL(\theta)}{d\theta} \cdot I^2$$

A numerical simulation shows the corresponding average torque versus current reference for the two case of power supply (figure 21).

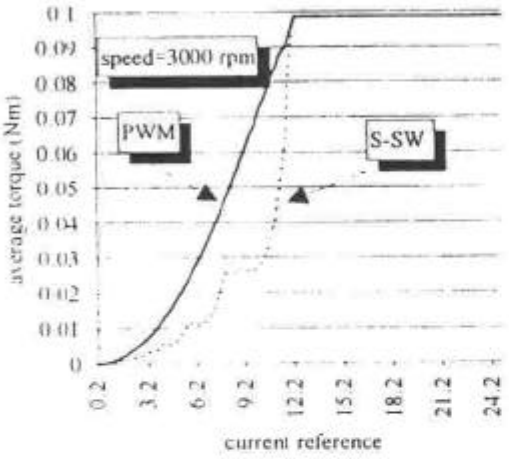


figure 21 : comparison of average torque characteristic between PWM and SELF SWITCHING mode versus current reference. speed=3000 rpm.

For the same average torque we have to double the current reference in Self Switching supply. Despite this maximun current increasing, the total joules losses in the machine and the conducted losses in the semi-conductors remain equal for the two supply. Indeed, this kind of losses is only dependant of $\frac{dL}{d\theta} \approx cst$ which remain constant, so the average torque is proportional to the RMS current square. This lead to write that the ration of Copper losses/torque is not dependant of the supply mode.

But we can remark too that the Self switching torque characteristic is not a monotonous one versus the current reference. This could be a problem concerning the torque regulation or modelisation (the equivalent gain of torque versus current reference may present sign invertions). This can be eliminate by using the advancen angle value of the supply window We show on figure 22 the influence of this angle value on the average torque characteristic for the same speed. The angle is expressed in electrical degrees.

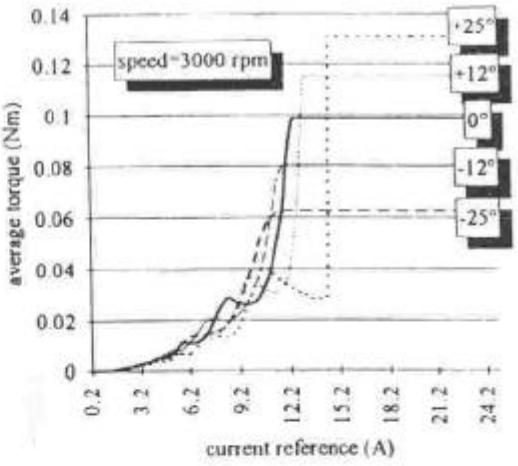


figure 22 : influence of advance angle of the supply window on the average torque characteristic. speed =3000 rpm and angle value in electrical degrees.

The extremes value of the curve envelops could give us a monotonous characteristic of the torque versus the current reference. So this could be a method by associating variables such as current reference, speed value and advance angle to obtain the same characteristic than PWM one concerning torque value.

6. ELECTROMAGNETIC PERTURBATIONS

We have studied the forms of conducted electromagnetic perturbations usually found [5]: differential mode and common mode . The experimental measurement scheme is presented on figure 23.

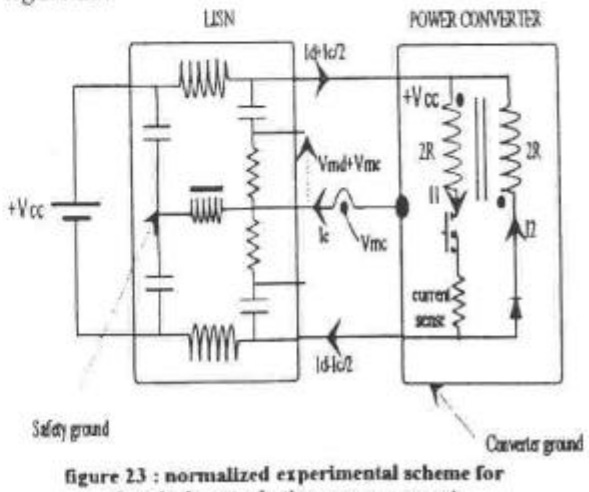


figure 23 : normalized experimental scheme for electrical perturbations measurement.

6.1. DIFFERENTIAL MODE

We obtain the two following wave-forms (PWM on figure 24 and Self switching on figure 25) for the same rms current (10A rms) in the phase stator coil.

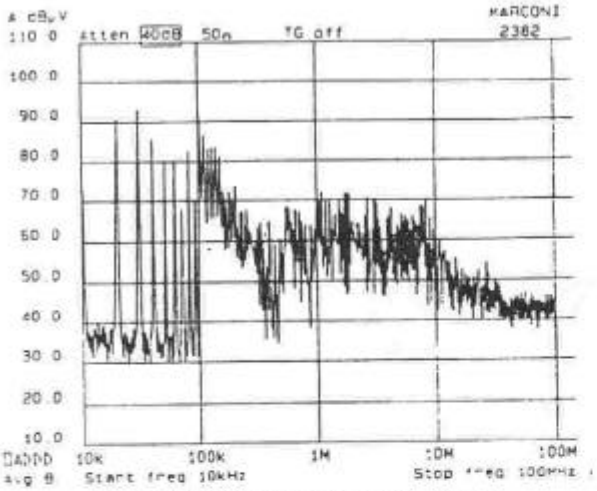


figure 24 : PWM conducted perturbations in differential mode measurements.

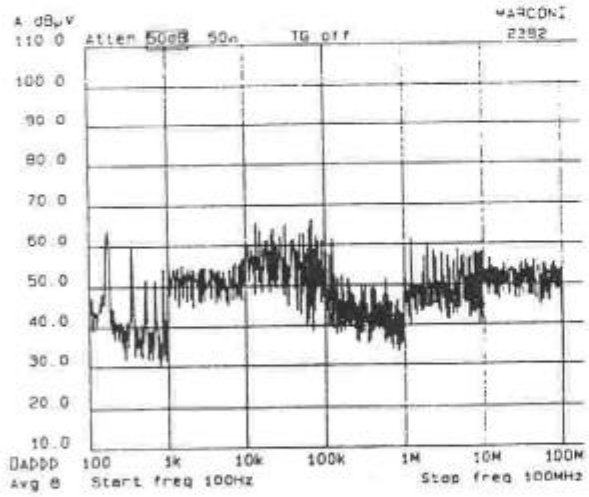


figure 25 :Self switching conducted perturbations in differential mode measurements.

We observe that electrical perturbations are notably lower in Self switching mode (figure 25) (65dB µV) than PWM one (figure 24) (95 dB µV) from 10kHz. We can find the Self switching spectral perturbation from 100Hz to 1kHz due to increased current level. If necessary this low frequency perturbation seems to be easy to filter thanks to a classic decoupling method.

6.2. COMMON MODE

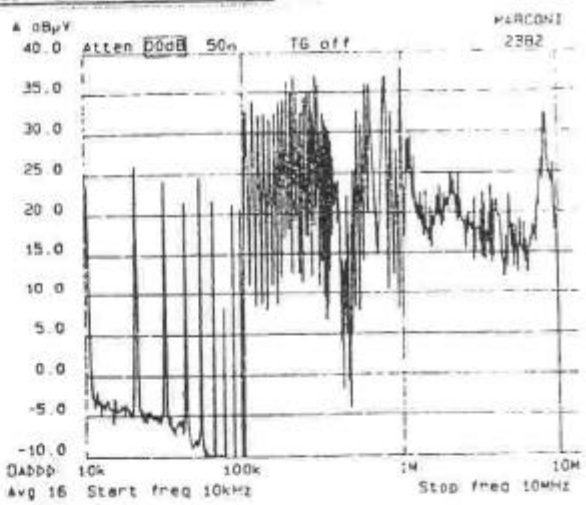


figure 26 : PWM conducted perturbations in common mode measurements.

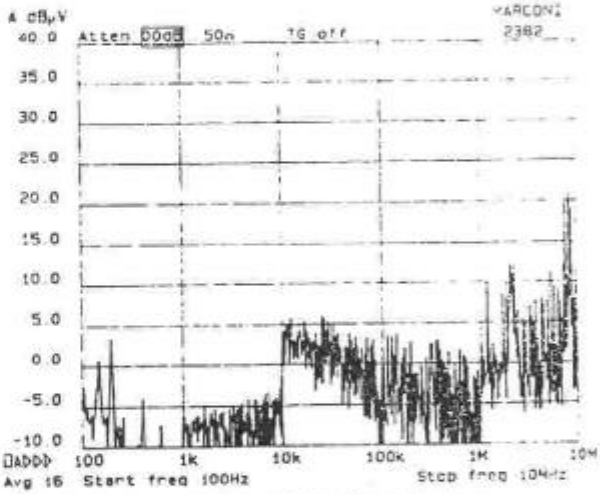


figure 27 : Self switching conducted perturbations in common mode measurements.

We can see that the Self switching perturbations are also notably lower than the PWM one at the same current reference from 10kHz (Self switching on figure 27 gives us 10dB µV and PWM on figure 26 with 35dB µV). Its interesting because the common mode perturbations is hard to be filtered by electronic method (physical contact between component and converter ground). Then we can also observe the low frequency spectral perturbations (starting from 100 Hz) of our Self switching mode even in the common mode instead of the PWM one witch starts from 10kHz.

7. CONCLUSION

We have seen that the main advantage of our new supply mode was leing in the decreasing of switching losses in the MOS and in the demagnetizing diodes. This leads to reduce the corresponding conducted perturbations either in the differential and the common mode (important point of our automotive application). But we have seen that it was necessary to hardly increase the current reference for the same average torque value than the PWM mode. Despite this current ioncreasing, motor and converter losses remain at the same level.

The second problem was leing in the non monotonous torque characteristic versus the current reference (figure 21). The solution may lie in the use of the advance angle value for each variables (current, speed, angle) set. Another limit lie in the saturation level of the machine; that is not very important because of the low saturation level of our machine (large air gap length) however the motor and switches losses may increase in an saturated machine.

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