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An Inverter/Motor Drive with Common Mode Voltage Elimination

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Abstract *- This paper presents an inverter/motor drive topology and control strategy, which serves to virtually eliminate common mode voltage in addition to reducing torque pulsations existing in typical three phase motor drives. The topology includes a six pole inverter, each pole having half the current rating of the equivalent three-phase inverter, to control a six phase machine. The proposed drive can be implemented with only a slight modification to existing equipment. A new space vector modulation strategy is presented, which allows open loop voltage control and elimination of the common mode voltage produced by the inverter.*

I. INTRODUCTION

Common mode voltage produced by the modulation of three phase power inverters creates significant common mode conducted current in motor drives. Whereas the ideal model of a motor drive has an infinite common mode impedance, in reality the common mode impedance due to capacitive coupling through the machine to earth ground can be relatively small at the modulation frequency creating significant common mode currents. These currents contribute to many unwanted problems such as coupling to nearby systems creating EMI, damaging the machine bearings and the bearing lubrication and heating the conduit carrying the three phase conductors.

As an alternative to inserting a large common mode choke to attenuate the common mode current of the inverter, a modification to the inverter topology and the modulation strategy is proposed that can eliminate the modulation component of common mode voltage produced by the inverter. The authors have recently proposed that a voltage source inverter topology with an even number of inverter poles can be modified and controlled to eliminate the common mode voltage applied to the load. This feature has been demonstrated with a four-phase inverter topology applied to a three-phase load [1]. The four-phase topology however requires additional unused hardware to make the load appear to be a balanced four-phase load. In [1] a four-phase filter was inserted between the three phase motor load and the four-phase inverter.

Alternatively it is shown in this paper that a six-phase inverter can be controlled to eliminate the common mode voltage at the switching frequency, enormously reducing the common mode filter size. Splitting each phase into two, each with half the current rating, does not change the total IGBT kVA rating of the inverter.

A six-phase machine has several advantages over a three-phase machine. The MMF in the air gap can rotate more smoothly resulting in reduced torque pulsations. Also, if a single phase fails the machine can continue to operate as a three-phase machine. If done in the factory, modification of existing three phase machines to form a six phase machine is not typically a costly modification to the machine. Hence, a six phase inverter/motor drive offers improvements in the electromechanical conversion process through reduced EMI, reduced torque ripple and improved fault tolerance.

II. SIX PHASE MOTOR

In most applications, a six-phase machine can be easily constructed by 'splitting' the standard 60° phase belt into two portions each spanning 30° [2],[3]. In Fig. 1 the resulting schematic of the stator windings is shown. It can be noted that the two sets of three phase windings are now spatially phase-shifted by 30 electrical degrees. Since the phase displacement between all six phases is not uniform,

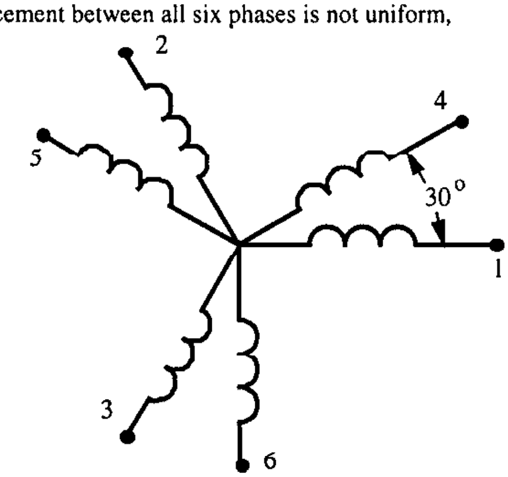

Fig. 1 – Schematic of the six phase stator windings

this type of stator configuration is sometimes referred to as an "asymmetrical six phase" or "dual three phase" connection.

Since the three phase winding has been split into two halves, it is clear that if the voltage of the machine is to remain the same the windings must be replaced with coils of double the number of turns and half the cross section. The machine can be modeled in d-q form in much the same manner as for a conventional three phase machine. That portion of the resulting equivalent circuit of the motor represented in the d-q plane is shown in Fig. 2 [2]. Note that

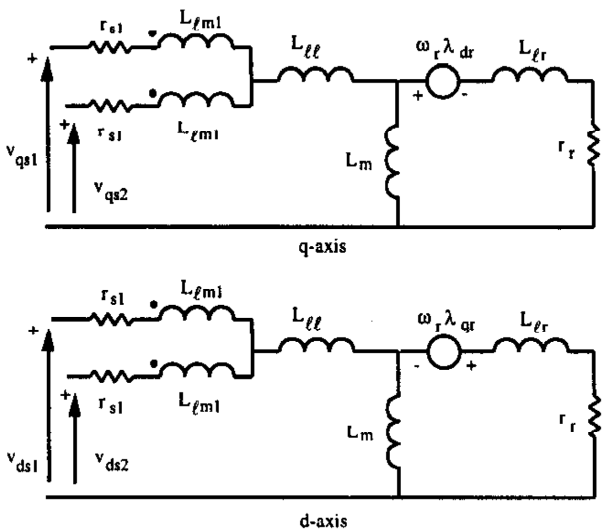

Fig. 2 – Equivalent circuit of six phase induction machine in d-q plane (stationary reference frame)

the major difference between the three phase and six phase motor is the coupling of the stator "leakage" flux which occurs when windings of the dual three phase groups occupy the same stator slots.

III. SIX PHASE VSI TOPOLOGY

The proposed inverter topology, configured to drive a six phase motor, is shown in Fig. 3. Six inverter poles are used, each having half the current rating of the equivalent three-phase inverter, to control a six-phase machine.

The authors have previously proved the advantage of having an even number of phases for common mode voltage reduction. Compared to other active solutions [1],[4], the advantage of this proposed topology is that no extra hardware, such as a second order filter, is required. This is possible because all six phases of the inverter carry load currents. The proposed drive can be implemented with only a slight modification to existing equipment. Twice as many switches with half the current rating keeps the converter kVA rating the same but requires more gate drivers.

The necessary condition to achieve zero common mode voltage is that the pole voltages of the six phases add to zero:

$$V_1 + V_2 + V_3 + V_4 + V_5 + V_6 = 0 \tag{1}$$

Equation (1) places a constraint on the allowable switch states. The voltages of three phases must be equal to $+V_{dc}/2$, at all times while the voltages of the other three phases must be equal to $-V_{dc}/2$. This means also that three upper and three lower switches are turned on in the six phases at all times. As a consequence, the common mode voltage caused by modulation of the six-phase drive can be eliminated with any modulation strategy provided that the constraint in equation (1) is satisfied.

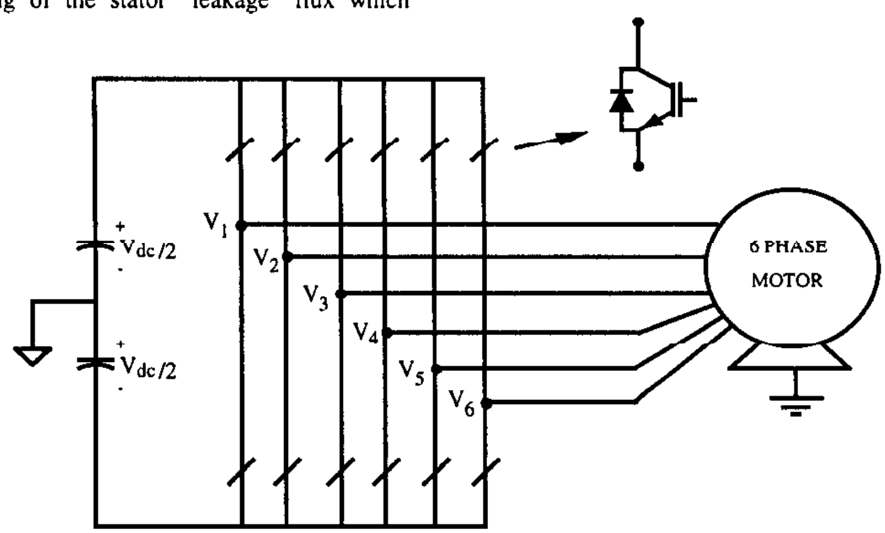

Fig. 3 - Schematic of the six phase inverter/motor drive

IV. SPACE VECTOR MODULATION STRATEGY

Recently, a control technique utilizing space vector decomposition has been proposed to control the currents of a six phase induction machine [5]. In this work it was shown that the machine current and voltage vectors might be mapped by vector space decomposition into three two-dimensional orthogonal subspaces. In fact, since the six-phase induction machine is a six-dimensional system, its control must be addressed from the point of view of a six-dimensional space.

According to [5], the real variables of the six-phase machine can be transformed to three orthogonal two-dimensional spaces d-q, z_1-z_2, o_1-o_2 by means of:

$$\begin{bmatrix} x_d \\ x_q \\ x_{z1} \\ x_{z2} \\ x_{o1} \\ x_{o2} \end{bmatrix} = T \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \\ x_5 \\ x_6 \end{bmatrix} \tag{2}$$

where T is the transformation matrix:

$$T = \frac{2}{3}\begin{bmatrix} 1 & \frac{1}{2} & \frac{1}{2} & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & 0 \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\ 1 & \frac{1}{2} & \frac{1}{2} & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & 0 \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} & 0 & 0 & 0 \\ 0 & 0 & 0 & \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \tag{3}$$

Note that the numbers 1...6 are assigned to the six phases of the inverter/motor system according to the schematic shown in Fig.1.

The transformation described by the matrix T is voltage invariant. It is interesting to note that the d-q axes coincide with the axes of the air gap flux of the machine. As the z_1-z_2 and o_1-o_2 subspaces are orthogonal to the d-q subspace, the variables on these two planes will not generate any rotating MMF in the airgap and thus they are non-electromagnetic energy conversion related. So the harmonics mapped into the z_1-z_2 and o_1-o_2 can be classified as a new type of zero sequence components.

In this paper we analyze the case of a six-phase motor with the two three-phase stator winding neutrals separated. For this case it can be shown [5] that the system becomes a four dimensional system and all the state vectors in the o_1-o_2 plane map into the origin of the o_1-o_2 plane.

The projections of the switching state vectors into the d-q, z_1-z_2 planes will be considered for our analysis. Figs. 4 and 5 show the inverter voltage vectors projected on the d-q and z_1-z_2 planes respectively.

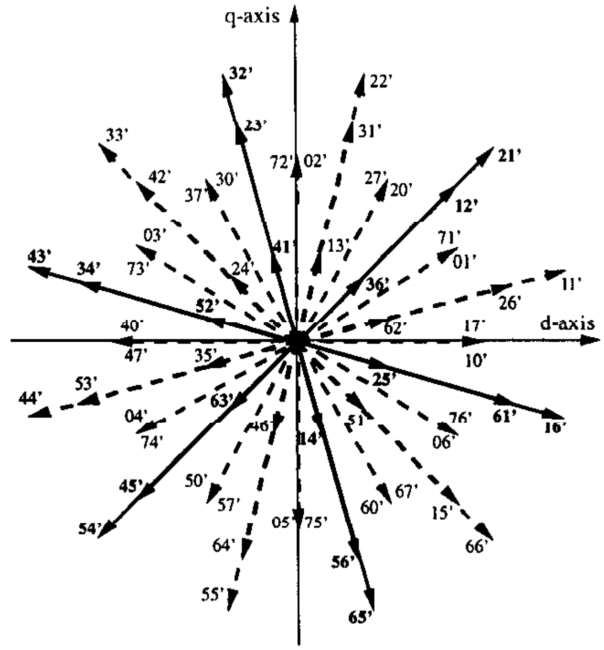

Fig.4 – Inverter voltage vectors projected on d-q plane. The states at the origin are: **70'**, **07'** (allowed), and 00',77'.

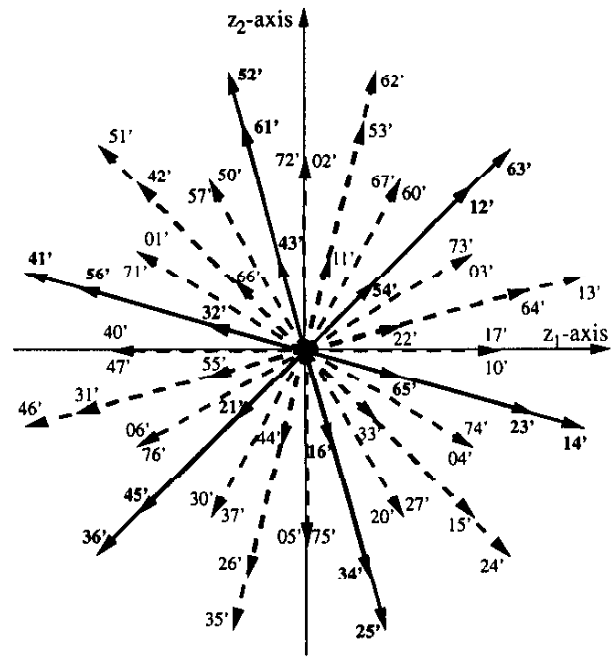

Fig. 5 - Inverter voltage vectors projected on the z_1-z_2 plane. The states at the origin are: **70'**, **07'** (allowed), and 00',77'.

Each vector is described with a number xy' where x and y' are numbers between 0 and 7 representing all the possible states of each three-phase set. In particular x represents the state of the three phases of the inverter driving phase 1,2 and 3 of the motor, while y' represents the state of the other three inverter legs, driving phases 4, 5 and 6 of the motor.

All the possible states of the two three-phase sets are shown in Table I.

TABLE I
STATES OF THE INVERTER

State x / y'	*Phase 1 / 4*	*Phase 2 / 5*	*Phase 3 / 6*
0	0	0	0
1	1	0	0
2	1	1	0
3	0	1	0
4	0	1	1
5	0	0	1
6	1	0	1
7	1	1	1

As is common, state 0 means that the bottom switch is on, while state 1 means that the top switch is on in that inverter phase leg.

Note that a typical six-phase inverter has 64 total switching states, while the modulation strategy of this paper may use only 20 of these states, due to the constraint of equation (1). This constraint forces the modulator to have three switches high and three low at all times, thus eliminating 44 states out of 64. The allowable states are represented in Figs. 4 and 5 by full line vectors, while all other states are represented by dashed vectors.

The transformations that have been described provide the mathematical framework for controlling the machine voltage. A circular trajectory in the d-q plane produces two sets of three-phase sinusoidal voltages, phase-displaced by 30^o. In the z_1-z_2 subspace a 'zero trajectory' is maintained in order to control the harmonics mapped in the z_1-z_2 plane to zero [5]. This means that in order to produce sinusoidal voltage the reference vector should be:

$$\overline{V*} = \begin{bmatrix} v^*_d \\ v^*_q \\ v^*_{z1} \\ v^*_{z2} \end{bmatrix} = \begin{bmatrix} V^* \sin(\omega t) \\ V^* \cos(\omega t) \\ 0 \\ 0 \end{bmatrix} \quad (4)$$

where V* is the amplitude of the sinewave to be produced in the d-q plane. Therefore $0 < V_{an} = V^*/2\cos 15^o < 1/\sqrt{3}V_{dc}$ is the amplitude range of the six line to neutral reference voltages, since we are using a voltage invariant transformation. This becomes clear as we note that the reference vectors on the d-q plane are obtained by summing two vectors produced by each three-phase set of voltages. For example the vector 11' on the d-q plane is the sum of vectors 1 and 1', which are the shifted by 30 degrees. The 11' state is created when the two three phase sets are in the switching state 1.

In each sector, 5 out of the 20 available switching states are chosen so that their average value during the time $T_s/2$ is equal to the reference.

$$\overline{V*}\cdot\frac{T_s}{2} = \overline{V1}\cdot T_1 + \overline{V2}\cdot T_2 + \overline{V3}\cdot T_3 + \overline{V4}\cdot T_4 + \overline{V5}\cdot T_5 \quad (5)$$

where:

$$\frac{T_s}{2} = T_1 + T_2 + T_3 + T_4 + T_5 \quad (6)$$

T_s is the switching period and $V_1 \ldots V_5$ are the five allowable switching space vectors. Eqs. (5) and (6) clearly form a five dimensional vector equation. The vectors $V_1 \ldots V_5$ must be linearly independent in order for a solution to exist to Eqs. (5),(6). Also, the time durations, $T_1 \ldots T_5$ must be positive.

A set of vectors that satisfies all the constraints can be chosen as follows:

TABLE II
CHOSEN SWITCHING PATTERN

Reference Angle	*T1, T2, T3, T4, 2T5, T4, T3, T2, T1*
$-15^o < \theta < 45^o$	16', 12', 21', 61', 70', 61', 21', 12', 16'
$45^o < \theta < 105^o$	21', 23', 32', 12', 07', 12', 32', 23', 21'
$105^o < \theta < 165^o$	32', 34', 43', 23', 70', 23', 43', 34', 32'
$165^o < \theta < 225^o$	43', 45', 54', 34', 07', 34', 54', 45', 43'
$225^o < \theta < 285^o$	54', 56', 65', 45', 70', 45', 65', 56', 54'
$285^o < \theta < -15^o$	65', 61', 16', 56', 07', 56', 16', 61', 65'

As can be noted from Table II, four active vectors and one zero vector are chosen in each 60^o sector. In Fig. 6 are shown the space vectors chosen for the case $45^o < \theta < 105^o$ in the d-q and in the z_1-z_2 subspaces.

In terms of output phase voltage, the performance of this new modulation strategy is the same as a typical three phase inverter, since the maximum achievable output voltage is 0.577 V_{dc}, where V_{dc} is the DC voltage of the inverter. With respect to the maximum achievable output voltage in a typical six phase inverter wherein all the 64 states are used, the maximum voltage obtained by this new modulation strategy using only 20 switch states is 12% smaller.

V. SIMULATION RESULTS

The six-phase system shown in Fig. 7 was simulated utilizing the ACSL digital simulation program in order to demonstrate the validity of the above analysis. This model includes a balanced RL six-phase load connected to two separate neutrals, as shown in Fig. 7. Each neutral is connected to ground through parasitic capacitors, which in the simulation are assumed to be equal with C_g=17.59 nF.

The results of these simulations are shown in Figs. 8-10. Fig. 8 shows the common mode voltage and current, respectively, i.e. $V_g = V_{g1}+V_{g2}$ and I_g of Fig. 7. The total common mode voltage is the sum of the voltages measured between the neutrals of the load and ground. In a typical three-phase inverter the common mode voltage is in the hundreds of volts range and therefore the current can be as large as tens of amperes [1],[4]. It can be noted that both the current and the voltage in Fig. 8 are very small in comparison.

In Fig. 9 is shown the phase current of phase 1. It can be noted that the phase current produced by the inverter is a sinusoidal with modulation ripple added. This is possible because the reference vector produces sinusoidal components on the d-q plane, but it is constrained to be zero on the z_1-z_2 plane. This fact is clear by looking at Fig. 10 where the currents on the d and z_1 axis are plotted. The current on the z_1 axis only the ripple current due to modulation. The slight distortion existing on the phase current is due to the fact that the z1 current component is not exactly zero. It can be concluded that the output voltage control achieved with the proposed modulation strategy allows a good control of the current harmonics mapped on the z1-z2 subspace, resulting in sinusoidal output current waveforms.

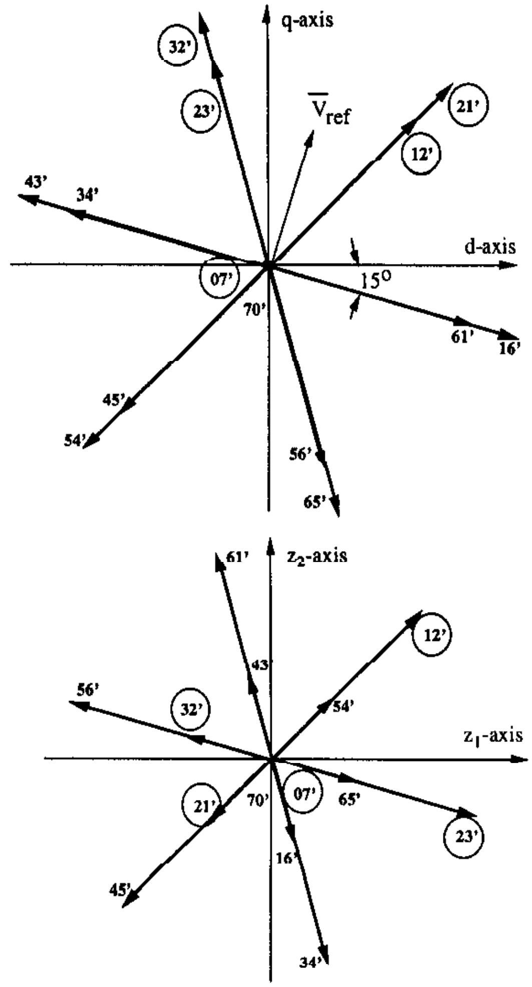

Fig. 6 – Selected vectors in the case $45^{\circ} < \theta < 105^{\circ}$

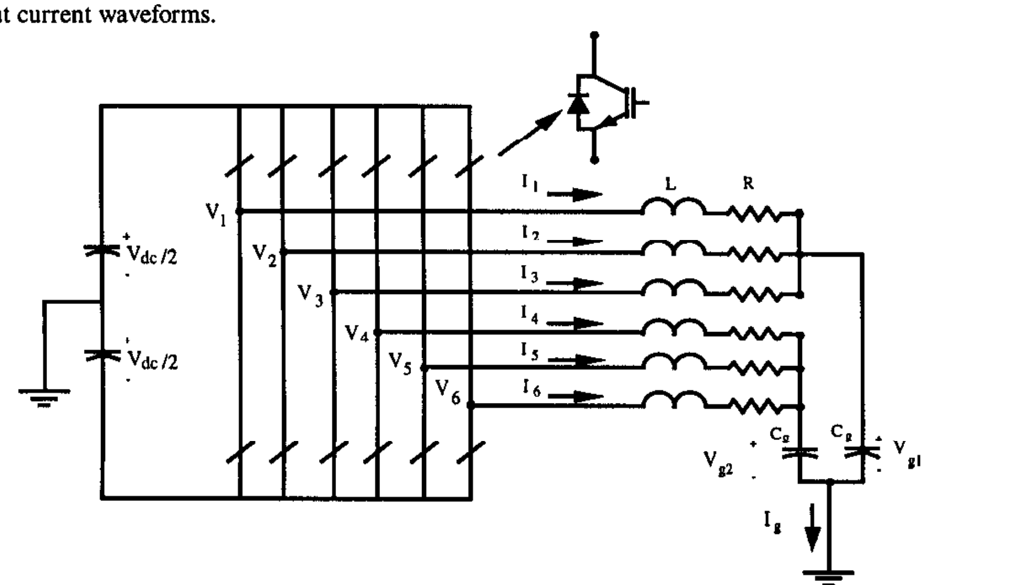

Fig. 7 - Six phase inverter with six phase RL load

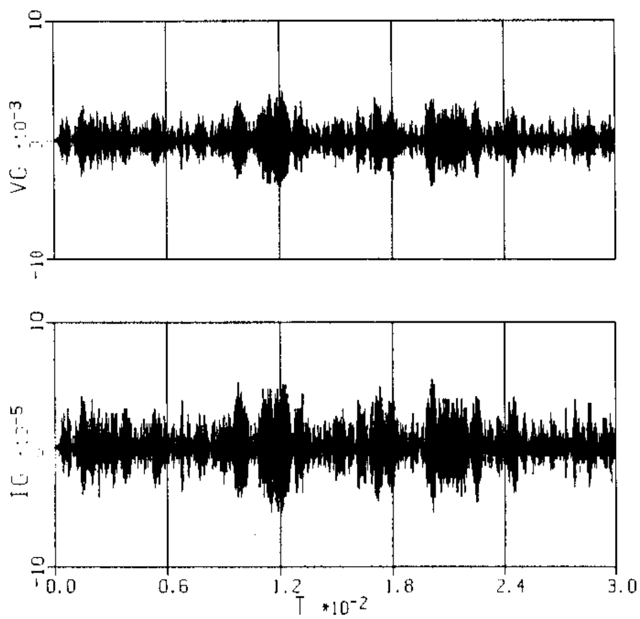

Fig. 8 – Common mode voltage (top 10 mV/div) and current (bottom 0.1mV/div).

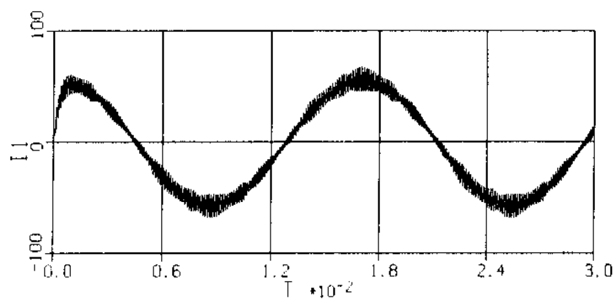

Fig. 9 – Machine phase current

VI. CONCLUSION

A six phase inverter voltage transformation has been developed to understand and control the currents in a six phase machine [5]. In this paper this transformation has been used to define a new modulation strategy for a six phase inverter/machine which reduces the common mode current created by PWM modulation when the load is capacitively coupled to ground as for the case of an electrical machine. Using a six phase drive improves the overall system performance by allowing the common mode voltage to be controlled to essentially zero and by reducing the torque pulsation components in the machine caused by a non sinusoidal MMF. Presently a six phase inverter is being built and a six phase motor is being wound in order to proof experimentally what has been proved by simulations in this paper.

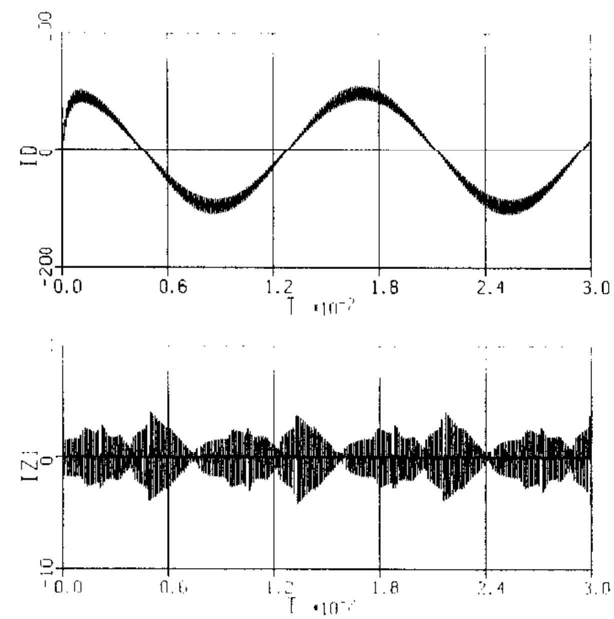

Fig. 10 – Currents on the d (top) and z1 axis (bottom)

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