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Abstract: *Applications of concentrated winding machines prompt new solutions in designing the power converter which supplies the machine, since the machine average torque can be increased significantly by supplying three-phase current waveforms matching in full the trapezoidal back EMF. However, such current supply arrangements involve a 3rd harmonic current flowing into the machine, and this paper presents an innovative converter topology to overcome the problem. This new configuration extends the standard layout of conventional PWM-VSI by adding a fourth branch devoted to control the machine neutral voltage. Computer simulations are used to describe the inverter modes of operation, as well as the paper reports experimental results from a converter prototype which has been constructed using IGBTs.*

INTRODUCTION

Applications of concentrated winding machines in electrical drives have prompted a number of significant problems which concern the design of both the machine and the power converter supplying the machine. As an example, for low-speed high-torque applications an axial-flux PM motor has been proposed [1]. This machine is characterized by a high number of poles, and designing the machine with a torque/weight ratio as higher as possible is mandatory in order to increase the machine power density and efficiency, as well as to reduce the costs. Further advantages can be achieved from a suitable arrangement of the machine supply. In fact, the performance of axial-flux PM drives can be improved greatly in terms of average output torque if the machine is fed by means of three-phase current waveforms which allow full contribution of the machine back EMF waveform in producing torque, such as in the case of a supply arrangement using either full square-wave or trapezoidal current waveforms. Such an approach, however, requires a suitable power converter configuration, because of the large 3rd harmonic current flowing into the machine neutral.

This paper deals with an innovative converter topology which permits such shaping of the machine current supply, so that the converter output current waveform can be adjusted suitably with respect to the machine back EMF. This new configuration extends the standard layout of PWM-Voltage Stiff Inverters by adding a fourth branch, which is devoted to control the voltage of the neutral resulting from a star connection of the machine phases. The paper describes the inverter modes of operation by means of computer simulation, and also reports experimental results from a converter prototype which has been constructed using IGBTs rated 500 V, 25 A.

MACHINE CURRENT SUPPLY

Benefits resulting from different current supply arrangements of a concentrated winding PM motor can be evaluated simply by assuming square-wave air-gap flux density, idealized current control, and machine operation with a large inertia load. Although in the following the analysis is specialized to axial-flux PM motors the issues lead to general conclusions, which thereby can be extended to conventional radial-flux PM motors. In developing the analysis fixed machine design characteristics are taken into account, such as the PMs geometry and the machine total number of turns (N). Hence, the machine average torque (T_{ave}) is evaluated considering different current supply arrangements, which result from both the number (m) of the machine phases and the waveform of the commanded phase current.

Machine EMF waveform

Since in PM motors the phase back EMF is a key factor in producing the machine torque, an analysis devoted to determine the most suitable current supply arrangement has to begin by examining the back EMF waveform. With the assumptions indicated above, the EMF waveform can be derived easily from either the flux cutting or the flux linking method. Considering an axial-flux motor having PMs over all the pole pitch of its stator winding, the peak value of the phase back EMF can be written as

$$E_{pk} = E_{max} / m \qquad (1)$$

where

$$E_{max} = N \ \omega_r \ B_{ave} \ (R_o^2 - R_i^2) \qquad (2)$$

represents the maximum open-circuit peak voltage achievable from connecting in series all the machine conductors, and

ω_r	mechanical rotor speed
B_{ave}	average air-gap flux density
R_o	outside radius of the magnet
R_i	inside radius of the magnet

Because each machine phase is realized by connecting in series concentrated coils equal to the number of the machine poles, the analysis can be restricted to a pair of poles and two coils per phase spaced out of a pole pitch. Such a machine configuration results in a trapezoidal waveform of the phase back EMF, which remains constant at E_{pk} along a portion (m-1)/m of the pole pitch, whereas it changes linearly when the phase coils are commutating through the interpolar axis. Considering an abscissa ϑ, which represents the angular shift of the axis of a given machine phase with respect to the pole axis, the phase EMF waveform during commutation of the phase coils can described by an expression

$$e_{ph}(\vartheta) = \frac{2 \ E_{max}}{\pi} \ \vartheta \qquad (3)$$

where ϑ ranges linearly (i.e. ω_r a constant) from $-\pi/2m$ to $\pi/2m$. Results given by eqs. (1) and (3) will prove useful later on. They clearly indicate that for fixed machine design characteristics both the EMF peak value and the commutation angle π/m decrease as the machine number of phases increases, whereas the slope of the EMF trapezoidal waveform does not depend on the value of m. However, the portion of the pole pitch in which the EMF is at its peak value increases in accordance with the machine number of phases.

Machine average torque

Assuming idealized current control allows a great insight into the machine torque production process. Whatever current control of a concentrated winding PM motor is performed, it can be recognized easily that the desired peak current (I_{pk}) should be supplied to a given machine phase during the time interval in which the phase coils are completely under the action of the poles, and thereby the phase EMF waveform is at the peak value E_{pk}. By such an approach, it clearly appears that the maximum instantaneous torque is achieved at the time instant when the coils of all the machine phases are completely under the poles (i.e. at either $\vartheta = -\pi/2m$ or $\vartheta = \pi/2m$) so that all phases may give the maximum contribution in producing torque, whereas the minimum instantaneous torque occurs whenever the coils of one of the m machine phases cross the interpolar axis (i.e. at $\vartheta = 0$), and thereby the relative instantaneous phase EMF becomes zero. Hence, above considerations suggest that the machine instantaneous torque can be described by means of a periodical time function whose period corresponds to changes of the abscissa

ϑ from $-\pi/2m$ to $\pi/2m$. Thereby, the period of the torque waveform depends on the machine number of phases, whereas the shape of such a waveform is fixed by the current control strategy. Considering that at any given time instant there is always one of the machine phases commutating gives an expression of the air-gap instantaneous power related to the machine torque as

$$P_{ag}(\vartheta) = (m-1)\, E_{pk}\, I_{pk} + e_{ph}(\vartheta)\, i_{ph}(\vartheta) \qquad (4)$$

where $e_{ph}(\vartheta)$ is given by eq. (3), and

$$i_{ph}(\vartheta) = K(\vartheta)\, I_{pk} \qquad (5)$$

represents the current that may be commanded to flow in the machine phase under commutation, being $K(\vartheta)$ a function introduced in order to take into account the shape of the current waveform.

Using eqs. (1), (3), and (5) into eq. (4), the air-gap instantaneous power can be expressed in per-unit of the power $E_{max} I_{pk}$, which represents the instantaneous air-gap power of a PM machine having an infinite number of phases (i.e. an equivalent PM d.c. motor). Hence, the per-unit instantaneous torque can be written as

$$T(\vartheta) = \frac{m-1}{m} + \frac{2}{\pi}\,\vartheta\, K(\vartheta) \qquad (6)$$

whereas averaging from $-\pi/2m$ to $\pi/2m$ gives the machine average torque

$$T_{ave} = \frac{m-1}{m} + \frac{2m}{\pi^2}\int_{-\pi/2m}^{\pi/2m} \vartheta\, K(\vartheta)\, d\vartheta \qquad (7)$$

Eq. (7) puts into evidence the possible methods of increasing the machine average torque by means of a suitable current supply arrangement. Conventional solutions for supplying trapezoidal EMF PM motors consider $K(\vartheta) = 0$, which means that each machine phase is fed by a current of a constant value I_{pk} only when the EMF is at its peak value E_{pk}. In this case the phase current waveform results in a square wave of width 180 (m-1)/m electrical degrees. It is clearly shown from eq. (7) that benefits over the average torque can be achieved by a machine number of phases higher than in conventional three-phase arrangements, as was proposed in reference [2]. However, from considering the second term of the eq. (7) it can be recognized that an extra average torque may be also produced if at any given time instant current is allowed to flow in the machine phase under commutation, so that the relative EMF can contribute in producing torque before reaching its peak value. Such an approach requires an innovative configuration of the power converter supplying the machine, since it can be easily demonstrated that in this case a m-th harmonic current has to be allowed to flow in the machine neutral.

In the following, this paper discusses an innovative power converter topology to overcome the problem, so that the machine current control may be arranged considering a phase current having either a 180 degree square waveform or a waveform matching the machine trapezoidal EMF. The first case corresponds to $K(\vartheta) = 1$, whereas $K(\vartheta) = 2\,\vartheta / \pi$ represents a machine current supply by means of a trapezoidal current waveform. Hence, the average torque given by eq. (7) becomes

$$\left[T_{ave}\right]_{fsw} = \frac{m - 1/2}{m} \qquad (8)$$

for a machine current supply using a 180 degree square wave, whereas it yields

$$\left[T_{ave}\right]_{tw} = \frac{m - 2/3}{m} \qquad (9)$$

in the case of a trapezoidal phase current waveform. The above written eqs. (8) and (9) permit evaluation of the average torque gain with respect to the conventional current supply arrangement which assumes $K(\vartheta) = 0$. It is determined that a 180 degree square wave current yields an extra average torque of 1/[2(m-1)] p.u., whereas a trapezoidal current waveform allows a torque gain of 1/[3(m-1)] p.u.. The analysis shows that benefits over the machine average torque decrease as the number of

Table I - Average output torque of a given axial-flux PM motor for different current supply arrangements.

Current supply arrangement	T_{ave}	I_{rms}	T_{ave}/I_{rms}	$T_{ave}/(I_{rms})^2$
3 ϕ - 180 degree sq. wave	0.8334	1.	0.8334	0.8334
3 ϕ - 120 degree trap. wave	0.7778	0.882	0.882	1.
3 ϕ - 120 degree sq. wave	0.6667	0.8165	0.8165	1.
5 ϕ - 144 degree sq. wave [2]	0.8	0.8944	0.8944	1.

the machine phases increases. Thereby, in the following only current supply arrangements devoted to three-phase PM machines are discussed, since they seem those ones that result in the most significant torque relative gain.

Looking at the issues of the analysis developed above, it should be noted that comparable torque benefits can be achieved from either increasing the machine number of phases or allowing a current harmonic component in the machine neutral. To this purpose, the average torque values resulting from different current supply arrangements of an axial-flux PM motor are indicated in Table I, where the rms values of the considered current supply waveforms are reported also for comparison purposes in per-unit of the peak current value. For three-phase current supply arrangements Table I shows that a 25% extra torque is achieved from a 180 square wave current, whereas in the case of a 120 degree trapezoidal current waveform the torque gain is of about 16%, if compared with the conventional supply arrangement using a 120 degree square wave. Although a trapezoidal wave current having same peak value as a 180 square wave current seems to result in a lower torque benefit, it can be noted from Table I that the trapezoidal waveform produces a larger average torque per rms amp. Hence, for applications where increased efficiency is important, the trapezoidal wave should be considered, since the resistive losses will be less for the same output power [3].

Effects of increasing the machine number of phases are also shown in Table I, which indicates that a five-phase axial-flux PM motor supplied by means of a 144 degree square wave current waveform exhibits a higher average torque/rms amp ratio if compared with any three-phase current supply arrangement. However, it will be recognized more clearly later on that such a five-phase current supply requires a power converter of higher cost, which may not be payed back completely from the resulting extra torque of the machine.

PWM CONVERTER CONFIGURATION

As mentioned above, a large 3rd harmonic component has to be allowed to flow in the machine neutral if a three-phase PM motor is supplied by means of either 180 degree square wave or trapezoidal phase current waveforms. Hence, the machine neutral has to be connected to the inverter d.c. link, and, furthermore, the control of the current waveform in the machine phases requires that the average value of the neutral voltage must be kept at half the value of the d.c. link voltage.

These problems can be resolved by the PWM converter topology shown in Figure 1. This new configuration results from the standard layout of a three-phase VSI by adding a fourth branch devoted to control suitably the neutral voltage. The inverter branches connected to the machine input terminals are operated in either six-step or PWM mode, whereas the fourth branch acts to modulate the neutral voltage so that the required average value is achieved. In order to reduce the current ripple resulting from the modulation, the switching frequency of the neutral-connected branch should be chosen several times higher than the inverter output frequency. Considering the converter modes

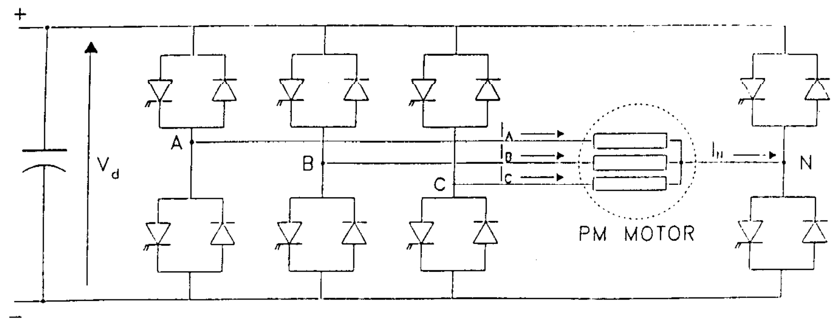

Figure 1 PWM-VSI configuration including modulation of the machine neutral voltage.

of operation, it appears that adding the fourth branch splits up each mode related to operations of conventional square-wave VSI or PWM-VSI into two extra modes, which are due to connecting the neutral alternately to the positive and negative bus of the d.c. link.

COMPUTER SIMULATION RESULTS

Above-indicated converter topology has been investigated by simulating the inverter modes of operation on a digital computer. As an example, Figure 2a and Figure 2b show simulation results related to the converter supplying the machine by means of square wave and trapezoidal current waveforms respectively. The machine is modeled by means of its R, L stator winding parameters and three-phase back EMF waveform resulting from the assumptions of a square-wave air-gap flux density and large inertia load. The model utilizes per-unit quantities. The inverter switches are operated by a constant switching frequency of 20 kHz: the neutral voltage is modulated considering a duty cycle of

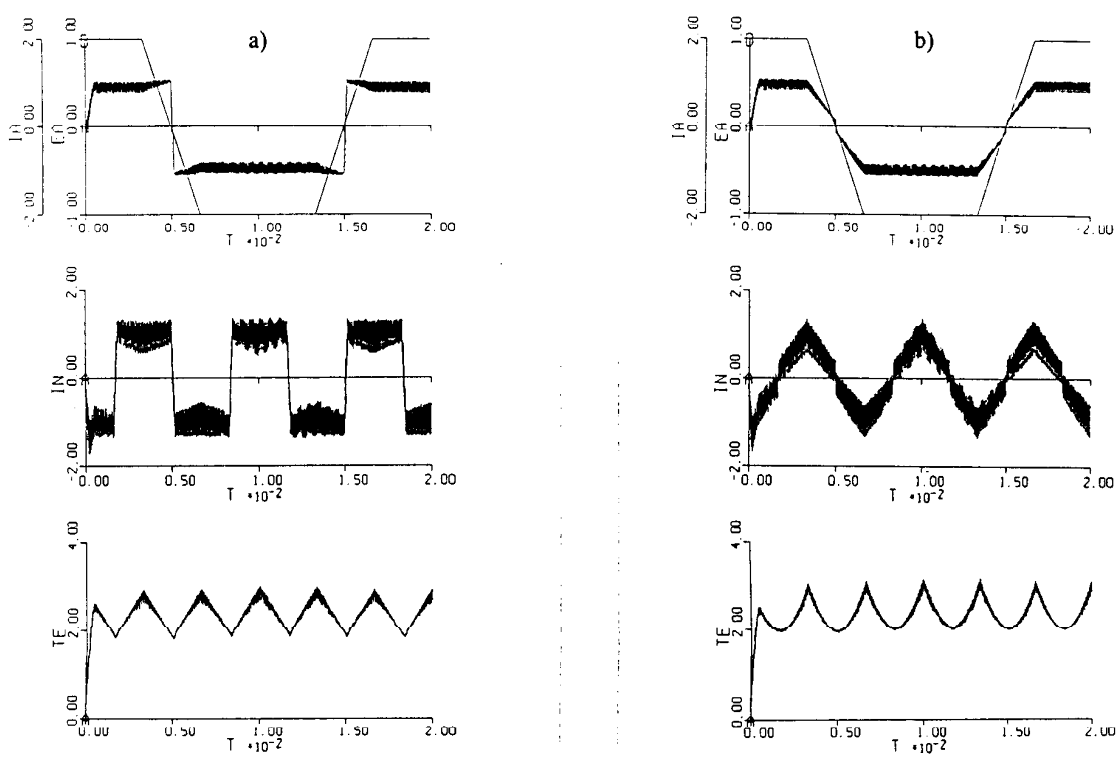

Figure 2 Computer results: E_A, machine EMF; I_A, phase current; I_N, neutral current; T_E, machine torque.

0.5, whereas the inverter branches connected to the load input terminals act as in a standard PWM-VSI having a constant switching frequency peak current controller. Computer simulations refer to an inverter output frequency of 50 Hz and a commanded peak current of 1.05 p.u.. Figures 2 and 3 show the waveforms of the current and back EMF related to one of the machine phases, as well as they indicate the current flowing in the neutral and the machine output torque.

These computer simulations demonstrate that a suitable current supply can be arranged for concentrated winding PM motors by means of the converter topology indicated in Figure 1. It can be noted that the inverter output current waveform includes a current ripple at the modulation frequency, which, however, has a negligible effect on the machine torque. The current ripple amplitude depends on both the value of the machine leakage inductance and the inverter switching frequency. Hence, for a given machine an acceptable current ripple amplitude can be achieved by setting a suitable value of the inverter switching frequency.

EXPERIMENTAL RESULTS

In order to demonstrate the feasibility of the approach, a PWM inverter having the topology shown in Figure 1 was assembled using IGBTs rated 500 V, 25 A. This converter prototype was used to supply a 16 pole axial-flux PM motor rated 32 N m [1], which has the open-circuit voltage waveform shown in Figure 3. The machine peak EMF is approximately 28 V at the rated speed of 375 rpm, and the EMF waveform results from Nd-Fe-B PMs distributed along 2/3 of the stator winding pole pitch.

For purposes of motor control an INTEL 80C196KC microcontroller is employed together with a 5000 ppr encoder and Hall-effect transducers for sensing the rotor speed and the phase currents respectively. The control algorithm implements the conventional PWM strategy, by using a 15 kHz saw-tooth carrier and control signals resulting

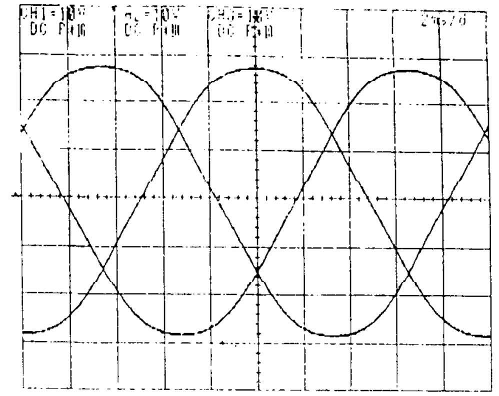

Figure 3 Machine prototype EMF waveforms.

from the error between the controller commanded phase currents and the actual motor currents. The neutral-connected inverter branch is switched at the same frequency of the saw-tooth carrier, by using a fixed duty-cycle value of 0.5 as required.

Because the machine prototype was constructed with PMs over 2/3 of the stator winding pole pitch and, thereby, the EMF waveform remains approximately constant along 60 electrical degree intervals, it was decided to carry out the experimental study by considering commanded three-phase current waveforms as follows:

a) 60 degree trapezoidal phase current waveform matching the machine EMF waveform;
b) 120 degree trapezoidal phase current waveform which changes by a slope twice that one of the EMF waveform;
c) 180 degree square wave phase current waveform;
d) 120 degree square wave phase current waveform, which approximates the current supply of conventional three-phase CRPWM-VSI: in this case the modulation strategy was operated in order to achieve a zero average current during the 60 degree commutation angle.

With reference to above-indicated current supply arrangements a), b), c) and d), oscilloscope traces resulting from the experimental study are shown in Figures 4, 5, 6 and 7. In each of such figures the waveform of both the phase current and the neutral current is indicated, being 10 A/div the current scale and 10 ms/div the time scale.

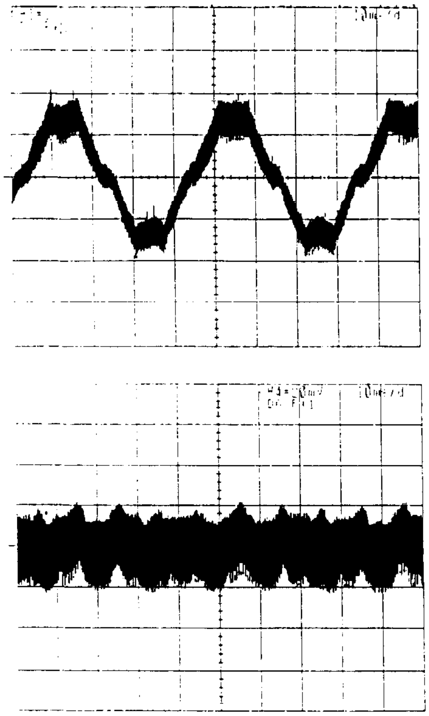

Figure 4 Machine phase and neutral currents due to a 60 degree trapezoidal commanded current.

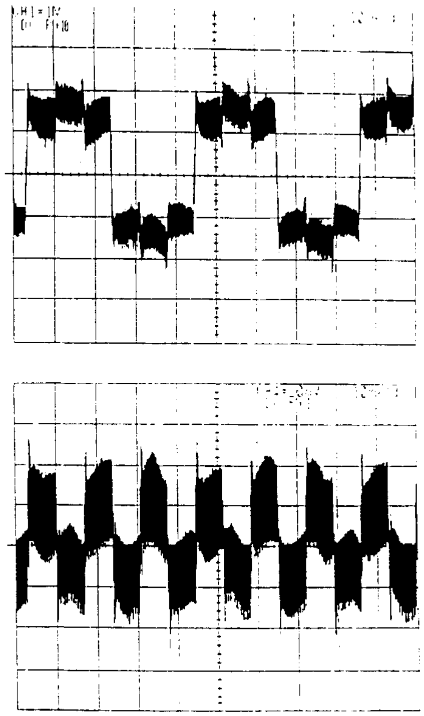

Figure 6 Machine phase and neutral currents due to a 180 degree square wave commanded current.

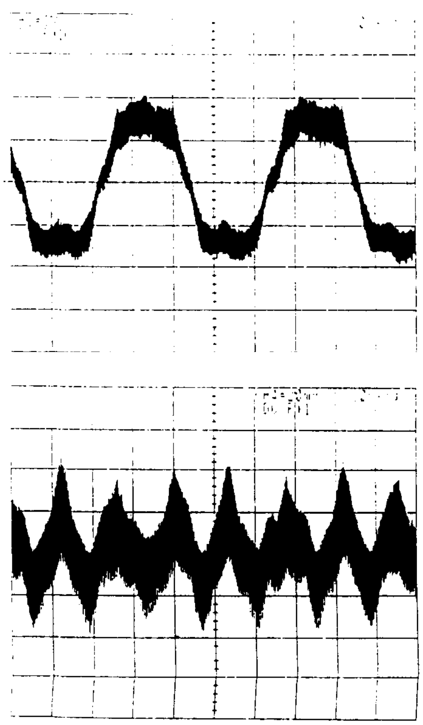

Figure 5 Machine phase and neutral currents due to a 120 degree trapezoidal commanded current.

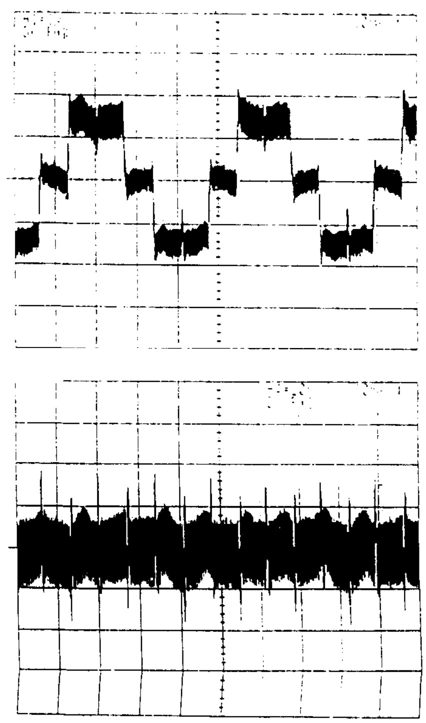

Figure 7 Machine phase and neutral currents due to a 120 degree square wave commanded current.

These results were achieved running the machine at 180 rpm under a shaft load due to a controllable d.c. brake, and all them refer to a commanded peak current of 14 A.

Experimental current waveforms show a significant phase current ripple which reflects on a ripple of the neutral current greater than expected. This appears to be the result of the low value of the machine leakage inductance, but it can also be a result of the microcontroller execution time, which would not be reduced below about 100 μs/cycle. Inspection of experimental waveforms also reveals the influence of the mutual inductance between machine phases on the phase current waveform. This is due to the current flowing in the machine neutral, and the phenomenon is particularly prominent in the case of supply currents having square wave waveform because of the resulting high value of the di/dt during current commutations.

Experimental data was additionally obtained for several load conditions of the PM drive prototype. Results are synthesized in Figure 9, which illustrates the steady state average torque vs. rotor speed characteristics related to each of the current supply arrangements indicated above. Benefits over the machine average torque resulting from allowing a current harmonic component in the machine neutral clearly appear. The experimental study is still under way in order to enhance the supply current waveforms, as well as to evaluate the drive overall efficiency for different current supply arrangements.

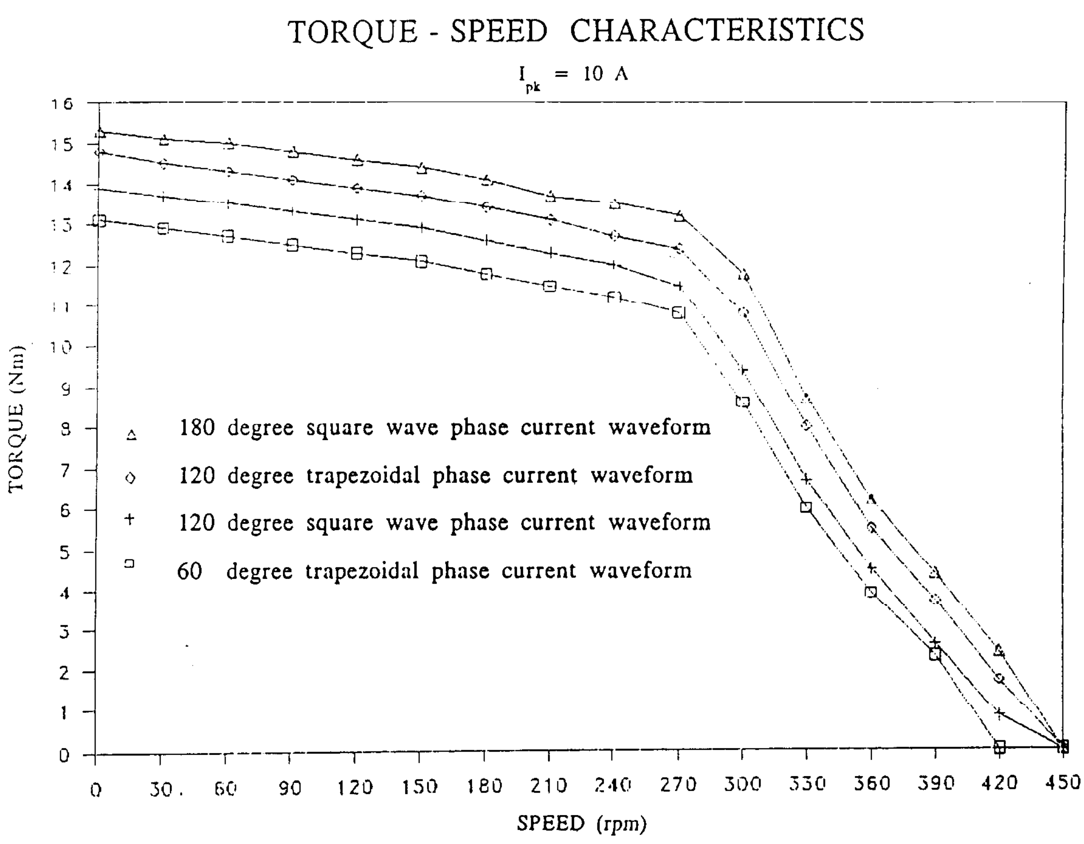

Figure 8 Average torque vs. rotor speed characteristics of the prototype of axial-flux PM motor drive resulting from different current supply arrangements.

CONCLUSIONS

In comparison with the conventional solution used for PM motor drives, this paper has discussed different current supply arrangements of a PM machine in terms of machine number of phases and commanded phase current waveform. It has been shown that benefits over the machine average torque can be achieved if the motor is fed by means of phase current waveforms which allow full contribution of the machine back EMF waveform in producing torque, such as in the case of machine supply arrangements resulting in a large current harmonic component in the machine neutral. Such an approach, however, requires a suitable power converter configuration, and this paper has presented an innovative converter topology which permits adjusting suitably of the converter output current with respect to the machine back EMF. This new configuration extends the standard layout of PWM-Voltage Stiff Inverters by adding a fourth branch, which is devoted to control of the voltage of the machine neutral.

The inverter modes of operation has been investigated by means of computer simulations, and an experimental study has been carried out driving an axial-flux PM motor by means of a converter prototype which has been constructed using IGBTs rated 500 V, 25 A. Conclusions prove feasibility of shaping suitably the three-phase supply current waveform of an axial-flux PM motor. While benefits over the machine average torque have been experimented from the converter topology discussed in this paper, the experience gained from the construction of the PM drive prototype indicates that the drive overall cost is determined mainly from the cost of the rare-earth PMs, so that the converter extra cost due to adding a fourth inverter branch is a relatively negligible cost increment.

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