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# ROBUST SPACE–VECTOR CURRENT CONTROL FOR INDUCTION MOTOR DRIVES

**Elwy E. El-kholy** * — **Ralph Kennel** ** — **Abdo El-refaei** ** — **Sabry Abd El-Latif** * — **Farok Elkady** ***

In this paper, analysis and implementation of two schemes for current control based on space vector modulation are proposed. A novel current control for vector control of the induction motor is presented in which the set of space voltage vectors are determined according to the stator current error in a stator reference frame. An off line switching table based on a software implemented comparator for minimizing this error has been implemented. Hence, optimal voltage vectors can be obtained from the inverter by using a switching table based on the angular coordinate. On the other hand, a modified predictive current controller based space vector modulation is proposed. In this technique, the space voltage vectors which are required to minimize the motor current error are calculated using a discrete motor model every sampling time. The proposed approach incorporates a modified conventional predictive current control with a new approach for reducing current ripples and decreasing the switching frequency. Simulation and experimental set-up results show the advantages and good performance of the two new modulation methods.

K e y w o r d s: current control, AC drives, PWM, space vector modulation, predictive current control

## 1 INTRODUCTION

Pulse Width Modulation variable speed drives are increasingly applied in many new industrial applications that require superior performance [1–4]. In such applications, the current controller is the most fundamental and important control loop for any variable speed drive systems [5–6]. The current controlled voltage-source inverters are usually preferred for quick response and accurate control [1–8].

The common strategies for current controllers can be classified as linear, hysteresis, and predictive controllers [7–13]. Each of these methods has individual properties and can be appropriate for certain applications. In the linear one, a constant switching frequency can be obtained but the controller parameters must be adjusted to optimize the system transient response and to minimize the magnitude and phase error in the line currents [5]. In the basic hysteresis current control, the motor currents are sensed and compared with their references using hysteresis comparators and the output signals are used to drive the power devices directly. Excellent dynamic performance can be obtained; however, the main disadvantage is the variation of the switching frequency during the fundamental period of currents [8–9]. Improved versions of hysteresis controllers that use an adaptive approach to produce a constant or nearly constant switching frequency have been presented; but with some more complexity than the basic ones [9]. However, a high switching frequency at a lower modulation index may be achieved. A predictive current controller that gives optimized performance in steady state by predicting the voltage vector that keeps the current in its hysteresis band for the longest time has been introduced [11–13]. The constant switching frequency predictive current controller calculates the voltage vector command every sampling period, which will force the current vector to its command value [11]. However, it gives optimum performance in terms of both accuracy and response time. It takes more calculations and requires a good knowledge of the load parameters. A simple current control was introduced which works in the stationary coordinate and selects the inverter voltage vector using a switching table and an electrically programmable read only memory (EPROM) [10]. In such methods possible error trajectories are predicted for the complete set of available switching state vectors of the inverter. The voltage vectors which satisfy the optimization criteria are selected when the current error exceeds the limit. Minimization of the switching frequency or maximization of the time between two switching states is the main objective of such methods. Constant switching frequency predictive current control predicts the error current at the beginning of each modulation period and evaluates the voltage vector to be generated by the PWM during the next modulation cycle so as to minimize the forecast error [9].

In this paper, a new space vector based current controller for induction motor drive is proposed. In this technique, the set of space voltage vectors (including the zero vectors) which are required to minimize the difference

* Electrical Engineering Department, Faculty of Engineering, Menoufiya University, Egypt. E-mail: eelkholy@yahoo.com

** Electrical Machines and Drives, Wuppertal University, D-42097 Wuppertal, Germany. E-mail: kennel@emad-uni-wuppertal.de, elref2002@emad-uni-wuppertal.de

*** Electrical Technology Department, Industrial College (IEC), Cairo, Egypt

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between the command current and the motor current are predicted according to the space error magnitude between them, and the position of the stator current in the $\alpha-\beta$ plane. On the other hand, a new real time predictive current control in which a design procedure giving the active voltage vectors that can lead to a minimum current error every sampling time will be described. The power spectrum of the current waveforms will be shown using Fast Fourier Transform (FFT). Implementation of the two current controller methods was made by means of a rapid prototype system using the main processor of a standard computer and some additional circuits.

## 2 CURRENT CONTROL BASED ON SPACE VECTOR MODULATION

### Principles of space vector based PWM inverter

To describe the inverter output voltages and to analyze the motor current control method the concept of complex space vector is applied. For an inverter feeding a symmetrical three-phase induction motor without neutral connection, the instantaneous voltages generated by the inverter should satisfy the following expression [2]:

$$v_a(t) + v_b(t) + v_c(t) = 0\,. \tag{1}$$

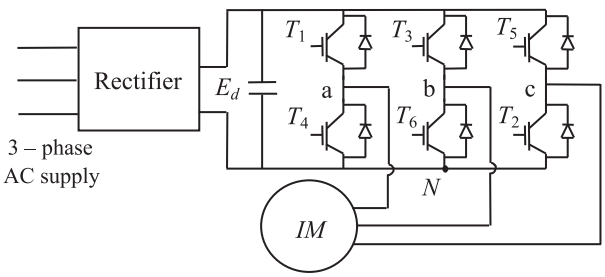



**Fig. 1.** PWM inverter fed induction motor drive.

These three space voltages can be represented as vector $[V_a,\ 0,\ 0]$ placed along the horizontal axis, while vectors $[0,\ V_b,\ 0]$ and $[0,\ 0,\ V_c]$ are delayed by $120^{-\circ}$ and $240^\circ$ from phase-a, respectively. On the other hand, the three phases of the induction motor may be connected to either the negative or positive rail of the DC supply, as shown in Fig. 1. Therefore, there are eight possible inverter configurations. By using these switching functions the stator space voltage vector can be expressed as [2]:

$$V_s(S_a, S_b, S_c) = \sqrt{\frac{2}{3}} E_d \Big( S_a + S_b e^{j\frac{2\pi}{3}} + S_c e^{j\frac{4\pi}{3}} \Big) \tag{2}$$

where, $S_a$, $S_b$ and $S_c$ are the states of the upper switches, and $E_d$ is the dc voltage.

According to the combinations of switching modes, the primary space voltage vector $V_s(S_a, S_b, S_c)$ is specified for eight kinds of vectors. Here, $V_7(0,0,0)$ and $V_8(1,1,1)$ are the space zero voltage vectors and the others are the space nonzero active voltage vectors as shown in Fig. 2.

The inverter output voltage vector is kept constant during the switching period, so the inverter current and, hence, the motor currents can be controlled by choosing the appropriate voltage vector.

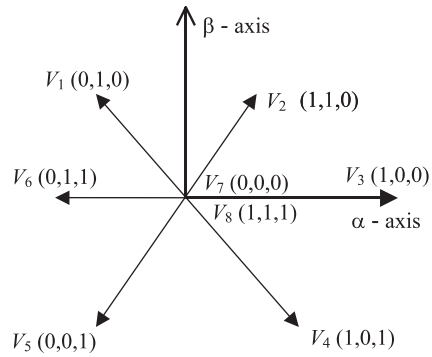



**Fig. 2.** Representation of primary space voltage vectors.

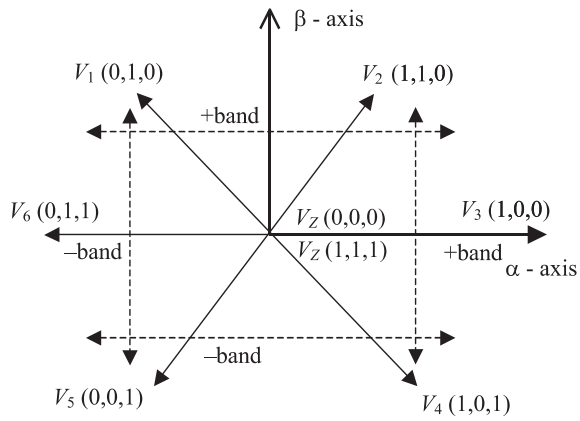



**Fig. 3.** Possible 9 regions.

### Conventional current control based on space vector modulation

In this approach, the inverter is controlled according to the vector control concept. The currentcontrol vector controller working in $\alpha-\beta$-coordinate as shown in Fig. 3. The system forms an output current space vector according to $i^*_{s\alpha}$ and $i^*_{s\beta}$ commands. The accuracy of current formation is determined by the width of the hysteresis zone of the three-level comparators [10].

The errors between the command and the actual stator currents are computed as:

$$\Delta i_{s\alpha} = i^*_{s\alpha} - i_{s\alpha}\,, \tag{3}$$

$$\Delta i_{s\beta} = i^*_{s\beta} - i_{s\beta}\,, \tag{4}$$

were, $*$ is a symbol that denotes the command value, $i_s$ are stator currents, $\alpha-\beta$-components in stator coordinates. The digitalized output signals for the a -component of the current controller are as follows:

$$\Delta i_{s\alpha} = 1 \qquad \text{for} \qquad i^*_{s\alpha} > i_{s\alpha} + \text{band}\,, \tag{5}$$

$$\Delta i_{s\alpha} = 0 \qquad \text{for} \qquad i^*_{s\alpha} = i_{s\alpha}\,, \tag{6}$$

$$\Delta i_{s\alpha} = -1 \qquad \text{for} \qquad i^*_{s\alpha} < i_{s\alpha} - \text{band}\,, \tag{7}$$

were, the band is the hysteresis band. Similarly for the $\beta$-component:

$$\Delta i_{s\beta} = 1 \qquad \text{for} \qquad i^*_{s\beta} > i_{s\beta} + \text{band}\,, \tag{8}$$

$$\Delta i_{s\beta} = 0 \qquad \text{for} \qquad i^*_{s\beta} = i_{s\beta}\,, \tag{9}$$

$$\Delta i_{s\beta} = -1 \qquad \text{for} \qquad i^*_{s\beta} < i_{s\beta} - \text{band}\,. \tag{10}$$

In this conventional method, the foregoing values have been fed into two three-level comparators to achieve the current control loop. Consequently, the error plane can be divided into nine regions as shown in Fig. 3. The voltage vector which forces the error back to zero can be easily selected using Table 1. The zero states are applied systematically, which helps to reduce the inverter switching frequency [10]. The current control in the coordinate system of $\alpha$–$\beta$-components gives the possibility to form the inverter output currents in phases $a$, $b$, and $c$ on the basis of the well known three to two phase transformation [10].

This method is simple and gives good results but the rectangular coordinate is not convenient to precisely select the inverter voltage vector. Also, commutation possibility for two IGBTs may occur and this will increase the commutation losses.

**Table 1.** Rectangular co-ordinate based switching

| | $\Delta i_{s\alpha} = -1$ | $\Delta i_{s\alpha} = 0$ | $\Delta i_{s\alpha} = 1$ |
|---|---|---|---|
| $\Delta i_{s\beta} = -1$ | $V_5$ | $V_5$ or $V_6$ | $V_6$ |
| $\Delta i_{s\beta} = 0$ | $V_1$ | $V_0$ | $V_4$ |
| $\Delta i_{s\beta} = 1$ | $V_2$ | $V_2$ or $V_3$ | $V_3$ |

**Table 2.** Angular co-ordinate based switching

| | $\phi 1$ | $\phi 2$ | $\phi 3$ | $\phi 4$ | $\phi 5$ | $\phi 6$ |
|---|---|---|---|---|---|---|
| $\Delta i_s = 1$ | $V_2$ | $V_3$ | $V_4$ | $V_5$ | $V_6$ | $V_1$ |
| $\Delta i_s = 0$ | $V_8$ | $V_7$ | $V_8$ | $V_7$ | $V_8$ | $V_7$ |
| $\Delta i_s = -1$ | $V_4$ | $V_5$ | $V_6$ | $V_1$ | $V_2$ | $V_3$ |

**Proposed current control based on space vector modulation**

In the proposed method, the complex plane can be divided into six regions and the magnitude of the current error is divided into three regions. Then, the error plane can be raised to 18 regions (*ie* double of the conventional method). The digitalized output signals for the stator current are as follow:

$$\Delta i_s = 1 \qquad \text{for} \qquad i^*_s > i_s + \text{band}\,, \tag{11}$$

$$\Delta i_s = 0 \qquad \text{for} \qquad i^*_s = i_s\,, \tag{12}$$

$$\Delta i_s = -1 \qquad \text{for} \qquad i^*_s < i_s - \text{band}\,. \tag{13}$$

Using the sector number and the stator current error the voltage vector which is required to reduce the current error to zero can be obtained from the proposed Table 2.

In the proposed method the full utilization of the possible voltage vector is more sequential and stable (*ie* more periodic times for the IGBTs) and less commutation losses may be obtained.

## 3 PROPOSED PREDICTIVE CURRENT CONTROL

The constant switching frequency predictive current controller calculates the voltage vector command every sampling period, which will force the current vector to its command value. A new real time predictive current control, which will give the active voltage vectors that can give the minimum current error every sampling time, will be described. Now, the problem is to find the appropriate voltage vector which will decrease the current error to zero as fast as possible.

The operation of the system (motor model) is described by the following equation:

$$v_s(t) = R_s i_s(t) + \sigma L_s \frac{\mathrm{d}i_s(t)}{\mathrm{d}t} + e_m(t)\,. \tag{14}$$

Here; $e_m(t)$ (the counter e. m. f vector) is expressed by

$$e_m(t) = \frac{L_m}{L_r}\frac{\mathrm{d}\psi_r}{\mathrm{d}t}\,. \tag{15}$$

It is desirable to estimate the rotor flux because of the difficulty in measuring it directly and Eq. 15 can be written in a differential form as follows:

$$v_s(k) = R_s i_s(k) + \frac{\sigma L_s}{T_s}\left[i_s(k+1) - i_s(k)\right] + e_m(k) \tag{16}$$

where $T_s$ is the sampling period.

The reference inverter voltage vector that will force the current vector to follow its command value is calculated every sampling period. It is assumed that the inverter voltage and counter e. m. f vector of the motor are assumed to be constant over the sampling period. The voltage vector $v^*_s(k)$ is calculated by changing the current $i_s(k+1)$ to follow the reference value $i^*_s(k+1)$ as given by the following equation:

$$v^*_s(k) = R_s i_s(k) + \frac{\sigma L_s}{T_s}\left[i^*_s(k+1) - i_s(k)\right] + e_m(k)\,. \tag{17}$$

From Eq. (18) the angle of the command voltage vector can be calculated from the following relation.

$$\alpha = \tan^{-1}\frac{v^*_{sq}}{v^*_{sd}}\,. \tag{18}$$

Using the voltage angle the location of the voltage vector can be determined with respect to the sector number. Now, three possibilities are considered. The first one is to use the location of the voltage vector and the nearest voltage vector for the whole modulation period, and this

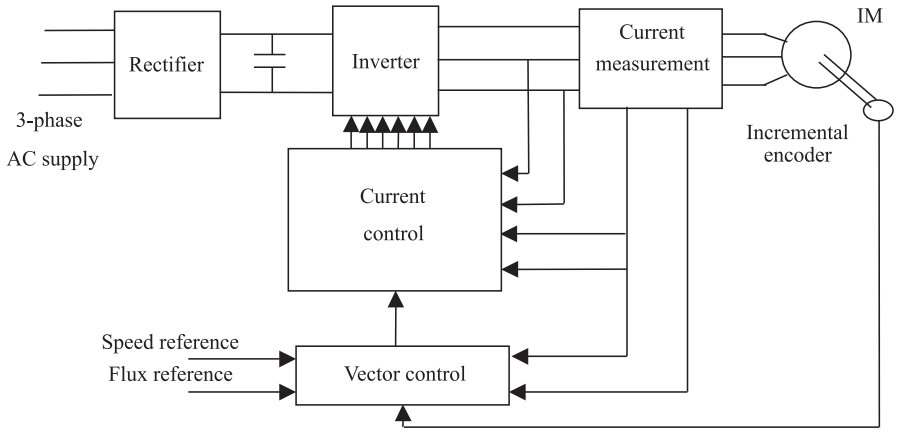



**Fig. 4.** Block diagram of the vector controlled induction motor drive

can be easily determined with respect to the sector number. The second one is to calculate the switching times for the power devices using space vector modulation. The third one is to use the direction of the command voltage itself and to insert its magnitude multiplied by the modulation index and a simple ramp modulator. The control method described until now does not solve the problem of current ripple. Now, the new proposed method with decreasing stator current ripples is explained.

**Table 3.** Line current total harmonic distortion: THD %

| | THD % |
|---|---|
| Conventional current controller based space vector modulation | 10.58 |
| Proposed current controller based space vector modulation | 8.01 |
| Conventional predictive current controller | 8.98 |
| Proposed predictive current controller | 5.24 |

A new method for reducing the stator current ripples has been presented. Our goal is to make the average output current for one period equal to the current command. The tolerance band around the current command which gives indication about current ripples can be calculated. So, the areas between actual current oscillations and its command are minimized to reduce the current ripple. If the derivative of these areas with respect to the switching time is set equal to zero, then the switching instants of the power devices are obtained. For one sample time, the stator current discrete equation can be expressed as follows:

$$i_{sd}(k+1) = i_{sd}(k) + \Big(\frac{1}{\sigma L_s} v_s(k) - \frac{R_s}{\sigma L_s} i_{sd}(k) + \frac{L_m^2 p\omega_r}{\sigma L_s L_r} i_{sd}(k)\Big) T_s\,, \tag{19}$$

$$i_{sq}(k+1) = i_{sq}(k) + \Big(\frac{1}{\sigma L_s} v_s(k) - \frac{R_s}{\sigma L_s} i_{sq}(k) - \frac{L_m^2 p\omega_r}{\sigma L_s L_r} i_{sq}(k)\Big) T_s\,. \tag{20}$$

The foregoing equation includes the influence of the applied voltage vector on the current variation, taking the operating condition into considerations.

The ripple areas can be calculated based on the values of the sampling time, and the positive and negative slopes of current equations. The on-time of the active voltage vector can be obtained according to minimization of ripple areas by the following equation [14]:

$$t_{sw} = \frac{2dI - (M^+ T_s)}{2M^+ - M^-}\,. \tag{21}$$

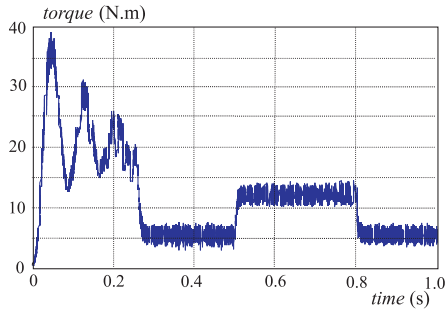
where $dI$ is the maximum difference value between the actual and command current, $M^+$, and $M^-$ are the positive and negative s lopes of the stator current wave. In abbreviation, the on durations of the active voltage vector and the zero vectors are $t_{sw}$ and $(T_s - t_{sw})$ respectively and for every new cycle the time must be reset to zero. It should be noted that if the time$<0$ or $>T_s$ the active voltage vector must be turned on during the whole sampling period.

## 4 SIMULATION RESULTS

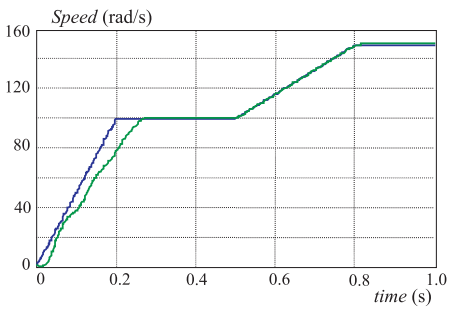
Figure 4 shows the complete diagram for vector control of induction motor drives. A Matlab/Simulink model was developed to examine the performance of current controllers. The simulations use the parameters of the experimental laboratory prototype listed in the appendix.

### Conventional current control based on space vector modulation

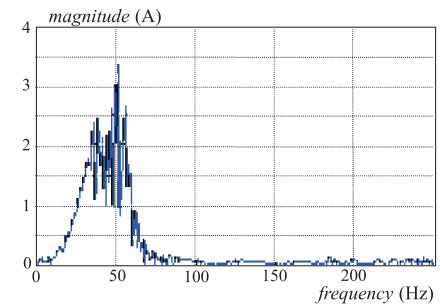
The performance of the induction motor drives using the conventional method has been studied utilizing the parameters of the actual prototype system. The research has been carried out to explain the performance of a whole vector control system including the performance of the current controller. Digital simulation results of the conventional method are shown in Fig. 5. The results show that the torque and speed responses are fast and highly dynamic.




(a) Torque response

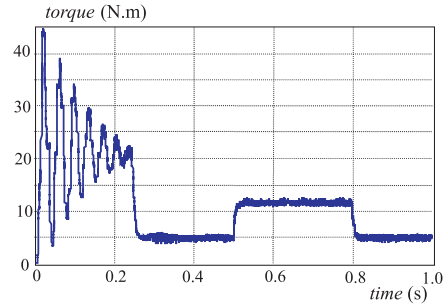



(b) Speed response

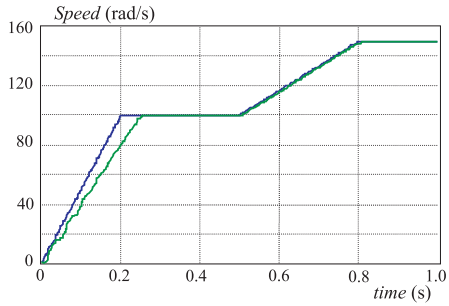



(c) THD for steady state stator current

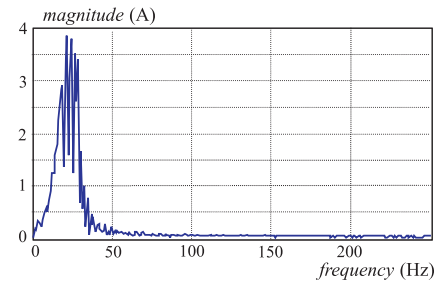
**Fig. 5.** Conventional current control based space vector modulation responses




(a) Torque response




(b) Speed response




(c) THD for steady state stator current

**Fig. 6.** Proposed current control based space vector modulation responses

### Proposed current control based on space vector modulation

Simulation of the proposed method is shown in Fig. 6, where the torque and speed response are as fast as the conventional one (*ie* good dynamic response). It is to be noted that the torque ripples have been decreased, moreover, the total harmonic distortion is less than in the conventional method (see Table 3).

### Conventional predictive current control

Simulation of the conventional method using space vector modulation to obtain the switching times for the power devices are shown in Fig. 7. It is clear that the torque and current responses are fast. Moreover, the motor torque follows the load torque nearly and this means good steady state performance.

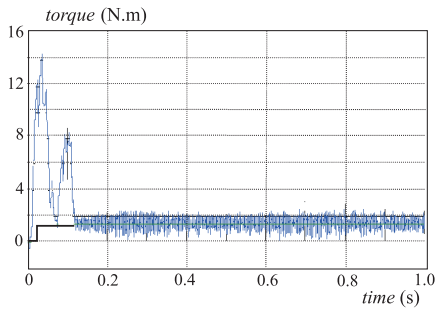
### Proposed predictive current control

Simulations of the proposed method using the magnitude of the dc link voltage and the command voltage angle in order to obtain the $\alpha-\beta$ voltage required for space vector modulation to yield the switching times for the power devices or to modulate this voltage using ramp modulator are shown in Fig. 8. It is clear that better torque performance and current response can be obtained. Figure 9 shows the torque comparison. It can be concluded that the proposed method gives better dynamic and steady state performance.

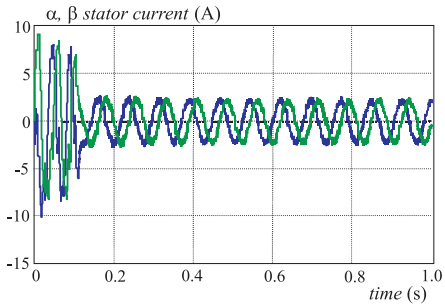
## 5 EXPERIMENTAL MODELLING AND RESULTS

### Hardware description

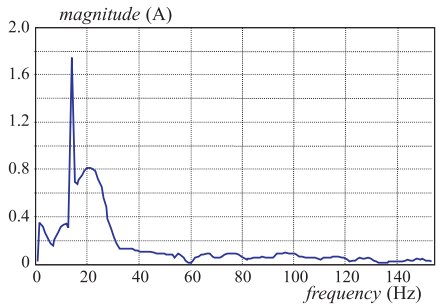
Today the development of drive control requires some important capabilities like high calculating speed, low cost, options for increasing performance, environment of PC and data connection and software estimation. An interface board inside the PC connects the parallel bus to the main board (Extension Kit). This interface board is located in the ISA-slot of the main board. Fast communication between the main processor and the I/O boards is guaranteed by using a wide parallel 16 bit data bus. The Extension Kit includes a board to send interrupt to




(a) Torque response

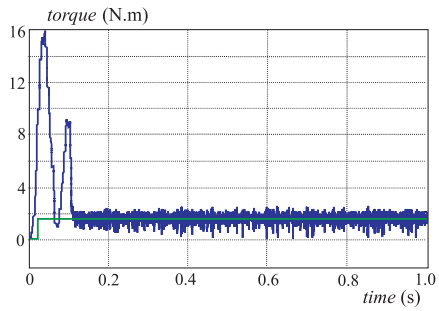



(b) Speed response

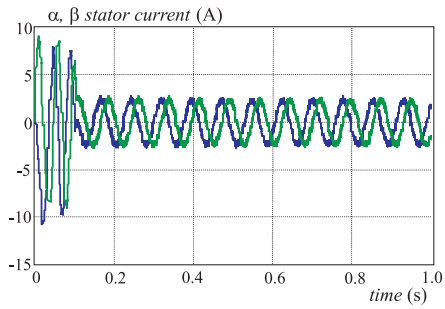



(c) THD for steady state stator current

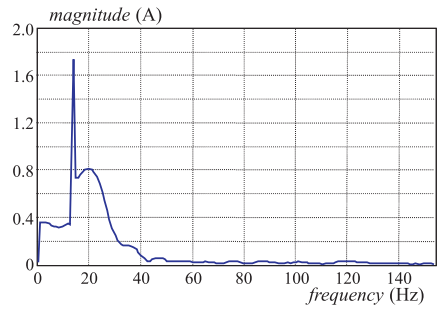
**Fig. 7.** Conventional predictive current control




(a) Torque response




(b) Speed response




(c) THD for steady state stator current

**Fig. 8.** Proposed predictive current control

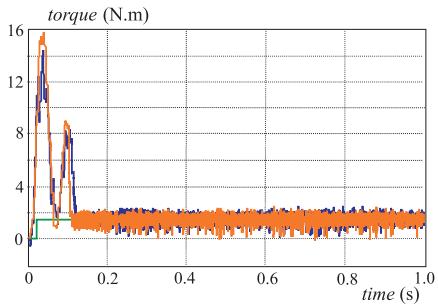


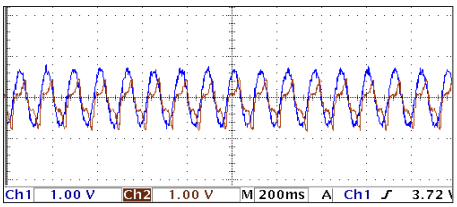

**Fig. 9.** Torque response for proposed and conventional current control methods.

the PC to synchronize the processing of the real time task with the controlled system. The main program of the software deals with initialization of I/O boards and the control algorithm is executed within interrupt service routine in which the A/D conversions for motor currents, voltages and a speed card for motor speed and the real motor are included.

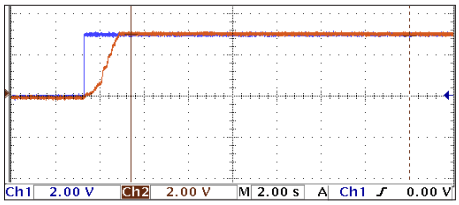
A driving circuit is used to transmit the six base signals required to turn-on or off the power transistors of the inverter, and an opto coupler is used to provide isolation between the power and the control potential by giving a suitable dead time between the two transistors in the same inverter branch, and also to prevent a short circuit on the dc supply when two transistors in the same branch are conducting at the same time [14]. The real time task is executed under MS-DOS and the real time program is executed using C language and the control algorithm is executed and the output pulses will be sent to the inverter through the drive board.

## Proposed current control based on space vector modulation

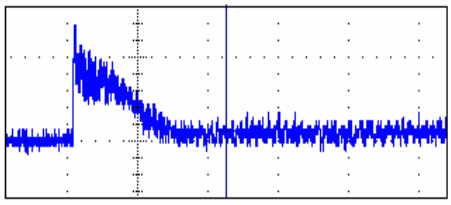
Experimental implementation of the whole vector control system including the current controller algorithm was performed to verify the behaviour of the current controller. In Fig. 10a, it is obvious that the stator current can be controlled and the speed response is shown in Fig. 10b. The torque response of the control system is shown in Fig. 10c.




(a) Actual and command stator current response (200 ms/div)

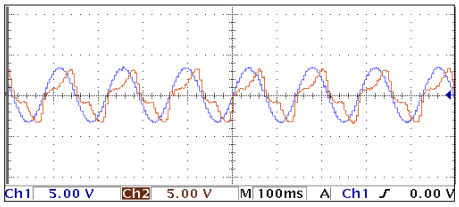



(b) Speed response (2 s/div, 100 rad/s/div)

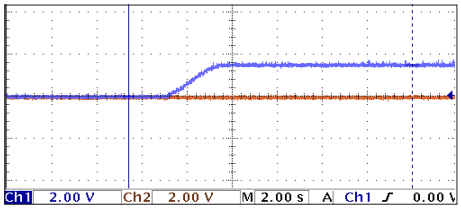


(c) THD for steady state stator current (1 s/div, 5 Nm/div)

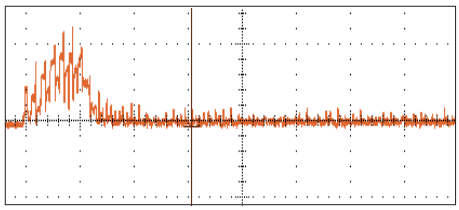
**Fig. 10.** Proposed current control based space vector modulation responses




(a) Actual and command stator current responses (100 ms/div)




(b) Speed response (2 s/div, 200 rad/s/div)



(c) Torque response (2 s/div, 5 Nm/div)

**Fig. 11.** Proposed predictive current control responses

**Proposed predictive current control**

Experimental implementation of the whole vector control system was performed to verify the behaviour of the predictive current controller. It is obvious that the stator current can follow the command current as shown in Fig. 11a. Figure 11b shows the speed response, which is rather fast. Figure 11c shows the torque dynamic performance for both transient and steady state.

## 6 CONCLUSIONS

The presented paper introduced two schemes for induction motor drive current control based on space vector modulation. The essential features of the first scheme is that the error plane raised to 18 regions, therefore, the full utilization of the possible voltage vector is more sequential and stable and less commutation losses may be obtained. The second scheme modified the predictive current control method to calculate the voltage vector command every sampling period, which will force the current vector to its command value. The two presented schemes have various merits such as fast response, less current harmonics, and low switching noises. The induction motor drive controlled by the two proposed current control techniques was simulated using Matlab/Simulink package programs. Moreover, its experimental set-up was build. To validate the effectiveness of the two proposed current control methods, the induction motor drive was subjected to starting and reference motor speed disturbances. Digital simulation and experimental results in terms of motor stator current, speed, and torque responses were plotted and obtained. The results prove the powerfulness and effectiveness of the proposed control approach in sense of fast response with less settling time and less overshoot/undershoot. Further, a comparison between the conventional current control methods and the proposed techniques was reported. The comparison results show the superiority of the proposed control methods via the high dynamic responses of induction motor drives. Also, the proposed control methods investigated the harmonics and induction motor losses values. It was found that the present control method has the ability to reduce the harmonics value via THD and where upon the induction motor losses are reduced.

## Appendix

The induction motor is a squirrel cage and has the following data:

| | |
|---|---|
| Rated power | 1.5 KW |
| Rated line voltage | 380 V |
| No. of pole pair | 2.0 |
| Stator resistance | 7.4826 Ω |
| Rotor resistance | 3.6840 Ω |
| Stator leakage inductance | 0.0221 H |
| Rotor leakage inductance | 0.0221 H |
| Mutual inductance | 0.4114 H |

| | |
|---|---|
| Moment of inertia | 0.04 |
| Supply frequency | 50.0 |
| Motor speed | 1400 rpm |
| Rated load torque | 11.0 Nm |

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**Dr Elwy E. El-kholy** was born in Menoufiya, Egypt in 1963. He received his BSc and MSc degrees in electrical engineering from the Faculty of Engineering, Menoufiya University in 1986 and 1992, respectively. From January 1993 to January 1996 he was PhD student at the INPL, Nancy University, France, under an academic channel exchange program. He received his PhD degree in 1996. He was a visiting research fellow to INPL, Nancy University, France during summer 1997. He was a visiting professor to Electrical Machines and Drives Dept., Wuppertal University, Germany from August 2001 to January 2002. He joined the Department of Electrical Engineering, Faculty of Engineering, Menoufiya University as a demonstrator in 1986. Then, he became an assistant lecturer in 1992 and was appointed as an assistant professor in 1996. In July 2002, he was appointed as an associate professor. He is now on a sabbatical leave to Aljouf College of Technology, Saudi Arabia. His main professional interests include ac and dc drives, direct torque and field oriented control techniques, DSP control and power electronics. He has published several technical papers in these areas.

**Ralph Kennel** (Prof, Dr-Ing) born in Wuppertal, Germany, studied in 1974-1979 Electrical engineering at the University of Kaiserslautern, Power engineering (especially electrical machines and drives) Diploma Thesis: Field Calculation with Finite Elements for a Fractional Horsepower Motor, PhD thesis (1984): Predictive Control for Converters, ETG Award of VDE: 1985 (German Association of Electrical Engineers) 1979–1983: Scientific staff at the Institute of Electronics/Power Electronics of the University of Kaiserslautern, 1983–1985: Development engineer for servo drives with Robert BOSCH GmbH at Erbach/Odenwald, 1986–1987: Group manager for spindle drives with microprocessor control with Robert BOSCH GmbH at Erbach/Odenwald, 1988–1997: Department manager for servo drives with Robert BOSCH GmbH at Erbach/Odenwald, 1997–1999: Department manager "Advanced and Product Development of Fractional Horsepower Motors" for automotive applications with Robert BOSCH GmbH at Bühlertal/Baden, since 1994: Visiting professor at the University of Newcastle-upon-Tyne, since 1999: Professor, Head of the Electrical Machines and Drives Laboratory at Wuppertal University General Activities:

**Sabry Abd El-Latif** (Prof, Dr) was born in Dakahlia, Egypt in 1945. He received his BSc degree in electrical engineering from El-Mansoura University in 1969. He was awarded the DEA degree and the PhD degree in power electronics from the INPL, Nancy, France, in 1977 and 1979, respectively. He joined the Higher Institute of Electronics, Menoufiya, Egypt in 1969 as a demonstrator. From 1976 to 1979 he was on a governmental mission at the INPL, Nancy, France. He joined the Department of Electrical Engineering, Menoufiya University as a lecturer in 1979, promoted to associate professor in 1983 and to full professor in 1987 at the same Department. He joined the INPL, Nancy, France from Sept. 1981 to Augusts 1982 on a post doctoral fellowship. From 1985 to 1990 he was on sabbatical leave from Menoufiya University to work at the United Arab Emirates (UAE) University. He was appointed Head of Electrical Engineering Department from 1990 to 1993 and from 1997 to 2000. He is currently the Dean of the Faculty of Engineering, Menoufiya University. He has 40 years of teaching experience and 20 years of consultant experience. He supervised a number of MSc and PhD students and examined a number of MSc and PhD theses at Menoufiya University and other universities. His current interests are power electronics, control of electrical machines and analysis of machine dynamics.