کارگاه‌های آموزشی مرکز اطلاعات علمی

مقاله نویسی علوم انسانی

آموزش مهارت های کاربردی در تدوین و چاپ مقاله

اصول تنظیم قراردادها
Stator Resistance Voltage Drop and Load Torque Compensations for PWM Inverter Fed Induction Motor Drive

Mohamed M. Negm, Osama S. Ebrahim, and Mohamed F. Salem

Abstract—The paper proposes a modified constant voltage/frequency (V/f) control algorithm for a 3-phase induction motor (IM) fed from a space vector pulse width modulated (SVPWM) inverter. This algorithm is experimentally verified using ordinary data acquisition card (DAC). A compensation method utilizes the stator resistance voltage drop space vector, in the stationary frame, is synthesized to correct the on-line stator voltage space vector and maintain the stator flux constant during the steady state as well as transient conditions. Moreover, an estimated value of the load torque is used to correct the command angular frequency of the IM and improve the performance of the PI-controller with load torque perturbations. The SVPWM switching patterns, which represent the reference stator voltage space vector, are realized by generating dependent digital words on the DAC digital lines. Accordingly, the computer storage memory is minimized and a saw tooth carrier generator is eliminated. Experimental results have been carried out to verify the effectiveness and applicability of the proposed control algorithm.

Index Terms—Data acquisition card, optimal preview controller, induction motor, SVPWM.

I. INTRODUCTION

Owing to the great advances in power electronics and microprocessor technologies, digital control of variable speed induction motor (IM) drive has become more and more popular in wide range of application ranging from moderate to high performance systems. The field-oriented control (FOC) has been recognized as the algorithm, which gives the IM drive the best dynamic performance [1]. However, constant voltage/frequency (V/f) control method still has gained wide acceptance in industrial applications especially, in those loads where the transient performance characteristics are undemanding. The simplicity of this control technique makes it easy and fast to perform with little calculation capability and best adequate for low cost general-purpose industrial drives [2]. The V/f control method is based on the steady state characteristics of the machine where the stator voltages are varied in proportional to frequency to keep the stator flux and hence the pullout torque constants. This is of course true as long as the voltage drop across the stator resistance can be neglected. However, large errors would be occurring at low speeds and high loads, where the resistive voltage component may exceed the induced voltage component and hence, a correction of the command stator voltage is required (voltage boost) [1]-[3]. One possible solution is to change the command stator voltage according to the steady state machine characteristics taking the stator resistance into consideration. Family of pre-calculated plots covering the whole control range is required to be stored in the computer memory. The solution is sensitive to the machine parameter variations that are not accounted for and adds to the system complexity [1]. In [4] two compensation methods for the stator resistance voltage drop that maintain the stator flux linkage approximately constant are introduced. The methods based on the IM model in a rotating reference frame hence, a coordinate transformation is needed. Another alternative is to convert from impressed voltage source to current source in order to eliminate the effects of stator resistance [5]. Although, CSI has inherent capability for regeneration and short circuit protection, there are serious problems associated with its application, such as: bad input power factor, high stator current harmonic content, high torque pulsation, low utilization of the DC bus voltage, variable switching frequency and ill defined harmonic spectrum. Most of these limitations need additional external circuits and more complex PWM techniques to be solved. Therefore, in today's industrial world, pulse width modulated voltage source inverter (PWM-VSI) is preferred over CSI, which requires a better match between the inverter and the driven motor [6]. Different PWM techniques such as; sinusoidal PWM, hysteric PWM and the relatively new space vector pulse width modulation (SVPWM) technique, have been proposed. The SVPWM technique is based on the space vector representation of the command voltage. In comparison with the sinusoidal PWM technique, the space vector PWM technique generates less harmonic distortion in the output voltages and currents, provides more efficient use for the DC bus voltage and requires less execution time and storage memory program for digital implementation [7], [8]. Application of digital signal processor (DSP) to realize the SVPWM technique for 3-phase voltage source inverter has been proposed [8], where an external saw tooth carrier generator is required. There are many different controllers based on the modern control theories have been proposed for control of the AC drive systems, such as: fuzzy logic control, neural network, adaptive neuro-fuzzy inference system, adaptive control and other optimal preview control [9], [10]. However, the conventional proportional-integral (PI) controller still has gained the widest acceptance. This is because the fact that the conventional PI-controller is quite simple and easy to implement and tuned. Unfortunately, the conventional
The SVPWM technique is to instantaneously realize the reference voltage vector $v^*$ by controlling the duration of the switching states, which are corresponding to the basic space vectors such that in one switching period, the average output voltage can be written as:

$$v^*(t) = \frac{t_0}{T_c}v_0 + \frac{t_1}{T_c}v_1 + \frac{t_2}{T_c}v_2 + \frac{t_3}{T_c}v_3 + \frac{t_4}{T_c}v_4 + \frac{t_5}{T_c}v_5 + \frac{t_6}{T_c}v_6 + \frac{t_7}{T_c}v_7,$$

(1)

where, $t_0, t_1, ... , t_7$ are the turn ON time of the vectors $v_0, v_1, ..., v_7$, in an arbitrary sector. For example, in sector-3 (S3), between $v_1$ and $v_2$ on Fig. 1, in one switching period, the reference stator voltage vector can be expressed as

$$v^*(t) = \frac{t_0}{T_c}v_0 + \frac{t_1}{T_c}v_1 + \frac{t_2}{T_c}v_2 + \frac{t_7}{T_c}v_7,$$

(2)

symmetrical SVPWM implies that

$$t_0 = t_2 = \frac{1}{2}(T_c - t_1 - t_2).$$

(3)

Fig. 2 shows the successive digital words, represented in binary notation, to be generated for a reference voltage lies in sector-3 (S3). The necessary delay time for each digital word to be resident on the digital lines of the converter card is computed as

$$\text{Delay}_0 = 0.25t_0; \text{Delay}_1 = 0.50t_1$$

$$\text{Delay}_2 = 0.50t_2; \text{Delay}_3 = 0.50t_0$$

(4)

The flowchart in Fig. 3 indicates the on-line implementation of the proposed symmetrical regular sampling SVPWM algorithm using PCI-6014NI DAC [12]. From the direct and quadrature components $(v_{sd}, v_{sq})$ of the reference stator voltage vector, we can determine the sector number in which this voltage vector lays and to make full use of active turn-on time of the space vectors, the reference voltage is normally split into two nearest adjacent voltage vectors and two zero-vectors $v_0$ and $v_7$ in an arbitrary sector. However in order to reduce the switching actions and to heavy computation process needed on-line to eliminate the effects of the load torque perturbation. In this paper, a modified V/f control scheme for IM drive fed from SVPWM inverter is realized using PCI-6014-NI ordinary multifunction data acquisition card (DAC). The SVPWM switching patterns, which represent the stator voltage space vector, are generated on the digital lines of the DAC through sector dependent digital words that leads to minimize the computer storage area and eliminate the saw tooth carrier generator. Simple and effective stator resistance voltage drop compensation method is proposed. The method depends on monitoring the stator resistance voltage drop space vector in the stationary reference frame to change the reference stator voltage space vector. In addition, an estimated value of the load torque, which is applied on the motor shaft, is used to change the command synchronous speed. This speed is the output signal of the conventional PI-controller, which will compensate for the load torque perturbations. The practical implementation of the proposed control algorithm requires the values of the stator resistance as well as the measured instantaneous values of two phases of the stator current to be known. Verification of the proposed algorithm is illustrated from the theoretical and experimental studies.

II. SVPWM IMPLEMENTATION

The SVPWM technique is a highly efficient way to generate the six pulse signals necessary at the power stage. Fig. 1 shows the six sectors limited by eight discrete voltage vectors, which are corresponding to the possible combinations of ON and OFF states of the power transistors in two levels three-phase inverter. The length of the nonzero vectors is two thirds of the DC bus voltage while the angle between any two adjacent non-zero vectors is 60 degrees. The zero vectors are at the origin and apply zero voltage to a three phase loads. The eight vectors are called the basic space vectors and are denoted by $v_0 = [000], v_1 = [100], ... ,v_7 = [111]$. The objective of the SVPWM technique is to generate on the digital lines of the switching patterns, which represent the stator ordinary multifunction data acquisition card (DAC). The fed from SVPWM inverter is realized using PC6014-NI.
The space vector notation of the stator voltage equation, expressed in the stationary reference frame, of the IM can be written as [10]

$$v_s = R_s i_s + j \omega_s \psi_s$$

(7)

where $v_s$, $i_s$, and $\psi_s$ denote the space vectors of the stator voltage, current and flux linkage, respectively, while $R_s$ is the stator phase resistance. At steady state operation, (7) can be rewritten as

$$v_s = R_s i_s + j \omega_s \psi_s$$

(8)

where $\omega_s$ is the synchronous angular speed.

Referring to Fig. 4, if the command voltage $v^* = C \omega_s e^{j \theta_s}$, is attained at any operating instant $\theta_s$, where $\omega_s^*$ denotes the command synchronous angular speed. Then the reference stator voltage space vector $v_s^*$ will be given by

$$v_s^* = v_{dq}^* = C \omega_s^* e^{j \theta_s} + R_s i_s$$

(9)

Equating (8) and (9), owing to $v_{dq}^* = R_s i_s + j \omega_s^* \psi_s$, gives the voltage to frequency ratio

$$C = j \omega_s^* \psi_s / \omega_s e^{j \theta_s},$$

which is considered constant at the nominal operating condition. Accordingly, the control law will maintain the magnitude of the stator flux linkage to maintain constant over the entire operating conditions. The on-line implementation of this compensation method requires determination of the stator resistance voltage drop space vector ($R_s i_s$). That means the value of $R_s$ must be estimated (or measured if applicable) on-line, in addition to measure the instantaneous values of two phases of the stator current $i_a$ and $i_b$ to evaluate the stator current space vector $i_s$ as

$$i_s = \frac{2}{3} (i_a + i_b e^{j2\theta} + i_b e^{j240})$$

(10)

where for balance condition: $i_c = - (i_a + i_b)$.

**IV. PI-SPEED CONTROLLER DESIGN**

There are several methods to determine the PI-controller’s gains, which depend on the nature and complexity of the process. The analytical approach, which needs known of the transfer function of the process, can be used to determine the proportional gain ($K_p$) and the integral gain ($K_i$) of the PI-speed controller depending on the required steady state and transient specifications of the closed loop control system.

The electromechanical equation of the IM drive is

$$T_{em} = J \frac{d \omega_r}{dt} + F \omega_r + T_L$$

(11)

where $p$ stands for $d / dt$ and $T_{em}$, $T_L$, $J$ and $F$ are the developed electromagnetic torque, load torque, inertia and friction coefficient constants, respectively. In designing the conventional PI-controller, the load torque is considered as a disturbance and omitted from (11). Assuming the electromagnetic torque of the IM is limited bellow its pullout value, its expression can be approximated as [14]

$$T_{em} \approx \left( \frac{T_{cmp}}{\omega_{op}} \right) (\omega_s - \omega_r) \approx K (\omega_s - \omega_r)$$

(12)

where $T_{cmp}$ and $\omega_{op}$ are the pullout torque and the slip frequency at which it occurs. For PI-speed controller, the command stator frequency is given by

$$\omega_s^* = K_p (\omega_s^d - \omega_r) + \frac{K_i}{p} (\omega_s^d - \omega_r).$$

(13)

By manipulating (11), (12), and (13), the following transfer function between the rotor speed $\omega_r$ and the desired rotor speed $\omega_s^d$, is obtained
where \( P \) is the number of pole pairs and the symbol “\( \times \)” denotes the cross-vectorial product. It is evident that, differentiation of the rotor speed will be associated with amplification of noise ranging from low to high frequency components, so it cannot be eliminated by conventional filters, which leads to a considerable delay time. Therefore, a variable gain digital filter is employed at the output stage of the load torque estimator. The state equation of the variable gain filter is [13],

\[
q(k+1) = q(k) + K_f(k)e_f(k) \\
e_f(k) = p(k) - q(k)
\]

where \( q, p, e_f, \text{ and } K_f \) are the filter output, input, error, and gain, respectively. The filter gain equals minimum value when the filter error, within a certain hysteresis band, corresponds the expected noise bandwidth and it increases linearly to one if the difference between the filter output and input increases.

VII. STATOR FLUX ESTIMATOR

The stator flux space vector, in the stationary reference frame, can be estimated from the stator voltage and current space vectors as [1]

\[
\hat{\psi}_s = \frac{v_s - R_s i_s}{p}
\]

The integration drift problem due to the dc offset and measurement noise can be avoided by using high pass filter, hence (19) becomes

\[
\hat{\psi}_s = \frac{v_s - R_s i_s}{p + \omega_c}
\]

The cut-off frequency \( \omega_c \) is selected fairly small compared to the working frequency to reduce the phase and magnitude errors. In this work, the fundamental component of the stator voltage provided by the inverter is assumed to be
be equal to the reference voltage. The errors caused by this assumption are often small in medium and high-speed ranges. While at low speeds, the errors caused due to voltage drops and dead times in the switching devices become more prominent in the output voltage. Further precise results can be achieved by using an inverter modulator as proposed in [3].

VIII. EXPERIMENTAL SET UP

Fig. 4 shows the IM drive system schematic diagram based on the proposed control algorithm. In this figure all blocks inside the dotted line box represent software programming functions, which are written in C++ language and executed on Pentium-III PC computer with 750 MHz clock and 250 MB RAM. Euler's backward method is used to map (17), (18) and (20) into the discrete form. Some hardware blocks in the schematic diagram can be realized by using application specific IC (ASIC) to give more compact packing and enhanced performance. However for the stage of prototype verification it was decided to avoid controller windup, an anti-windup flag is activated when the inverter is being saturated and this in turn will stop updating the integral part of the PI-control law of (16) till normal operation retained again. This was necessary to mitigate noise and improve system performance. The controller gains are set at $K_p = 3$ and $K_i = 30$ (depend on the required control system performance). The direct and quadrature components of the reference stator voltage are processed through the SVPWM algorithm where 1.25 kHz switching frequency is used. The switching patterns ($S_a, S_b, S_c$) are generated on the digital lines (D0, D1, and D2) of the DAC with a duration controlled by the computed switching intervals as explained in Section II. The generated PWM switching signals are then fed to a dead time and driving circuit (D.T. & D.C.), where the 3-PWM signals are complemented to get 6-PWM signals and a proper blanking time of 40 μs for the high and low side signals of the same phase arm is provided. The outputs of the dead time circuit are isolated and amplified in the driving circuit to provide adequate PWM gating signals for the 3-phase MOSFET inverter bridge (MOSFET Inverter). A DC bus voltage of 190 V is obtained using a 3-phase diode bridge rectifier and filtered with 2000 μF capacitor filter such that it could be considered constant. The time needed to perform the necessary calculations and data storage is found to be 0.4 ms, therefore a 1.2 sampling time period is selected.

IX. EXPERIMENTAL RESULTS

The horizontal lines of the experimental results illustrated in Figs. 5-10, denote the time in seconds, while the desired rotor speed is demonstrated by solid line while its response is indicated by dot line. Comparisons are made between the proposed stator IR compensation method and V/f control using 5% voltage boost. First the rotor speed is ramped from 0 to 80 r/s in 0.6 s, using a linear ramp while the motor operates under an increasing load up to 50% of rated torque. The rotor speed ($\omega_r$), stator flux ($\psi_s$), and stator phase current ($I_a$) responses in both cases are shown in Figs. 5-6, respectively. Using the proposed stator voltage drop compensation method, the flux has constant rated level with lower ripples and current drawn from the supply. This means that lower iron and copper losses, which leads to better efficiency. An important feature of these test results is the comparison between the rotor-speed responses shown in Fig. 7. With the proposed stator resistance voltage drop compensation method the response was faster with lower overshoot. This improvement was achieved because of the availability of greater torque from the motor as a result of the near nominal flux. The initial acceleration is similar in both cases. During the first tenth seconds, the flux level rises and little or no torque developed. The steady state is reached after approximately 1.5 s. Fig. 8 shows comparisons of the rotor speeds under two extreme cases of stator resistance mismatch. In curve (a) and (b) the estimated stator resistance is 1.5 and 0.5 from its nominal
X. CONCLUSION

In this paper an open loop stator flux control using stator resistance voltage drop compensation method in the stationary frame is introduced. Laboratory experiments have been carried-out, and showed that the compensation method is efficient to maintain constant stator flux, which keeps the motor torque producing its capability especially at low speeds. Furthermore, load torque compensation method is incorporated with the PI-speed controller to improve its performance against load torque perturbation. Experimental and theoretical analysis does not put additional constraints on the system stability rather than conversion of the load torque estimator. As a further development, implementation of these compensation methods in speed sensorless IM drive with on-line stator resistance estimation (or measurement) is currently in progress.

REFERENCES

Mohamed M. Negm was born in Cairo, Egypt, in 1956. He received the B.Sc., M.Sc., and Ph.D. degrees in electrical engineering from Ain-Shams University, Cairo, Egypt in 1979, 1983 and 1990, respectively. Since 1979, he joined the Department of Electrical Engineering, Faculty of Engineering, Ain-Shams University. During the interval 1985-1989, he was granted a Japanese scholarship at the Department of Electrical Engineering, Faculty of Engineering, Hokkaido University, Japan. During the interval 1992-2002, he was on a loan with the College of Technology at Dammam, Saudi Arabia. Since 2000 he has been with the Department of Electrical Engineering, Faculty of Engineering, Ain-Shams University, Egypt, where he is currently a Professor of control of power and electrical machines. Since 2004, he has been with the Electrical and Electronics Engineering Technology Department, Yanbu Industrial College, Royal Commission for Jubail and Yanbu, Saudi-Arabia. His research interest includes interdisciplinary area of optimal preview, VSS, ANN, and adaptive control system theories and their applications, control of power systems, control of electrical machines and robotics, control of power electronic systems, advanced process control design, digital control systems, sensorless control and applications of PLC and microprocessors in industry. Professor Negm is a senior member of IEEE.

Mohamed F. Salem was born in Cairo, Egypt, in 1948. He received the B.Sc. and M.Sc. degrees in electrical engineering from Ain-Shams University, Cairo, Egypt in 1971 and 1974, respectively. He received the Ph.D. degree in electrical engineering from Polytechnical Institute, Leningrad, USSR in 1979. During the interval 1980-1984, he was an Associated Professor in the Electrical Department, Ain Shams University. During the interval 1984-1985, he was a visitor Dr. in the Polytechnical Institute, Leningrad, USSR. During the interval 1990-2004, he was a Professor of electrical machines in the Electrical Department, Ain-Shams University. His research interest covers fields analysis and design of linear induction machines, linear actuators, induction generators, and their control.

Osama S. Ebrahim was born in Cairo, Egypt, in 1970. He received the B.Sc. and M.Sc. degrees in electrical engineering from Ain-Shams University, Cairo, Egypt in 1994 and 1999, respectively. Since 2004, he has been a Ph.D. staff member in the Electrical Power Department, Ain-Shams University, Cairo, Egypt. His research interest includes analysis and design of electrical machines, modern control theories and their applications, sensorless control of electrical machines, power electronic systems design, and microcontroller applications.
کارگاه‌های آموزشی مرکز اطلاعات علمی

مقاله نویسی علوم انسانی

اصول تنظیم قراردادها

آموزش مهارت های کاربردی در تدوین و چاپ مقاله