DSP-BASED POSITION CONTROL APPLIED TO SQUIRREL-CAGE INDUCTION MOTOR USING VECTOR CONTROL AND SPACE-VECTOR PWM MODULATION

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***Abstract* – The principle of vector control of AC machine enables the dynamic control of AC motors, and induction motors in particular to a level comparable to that of a DC machine. The vector control of currents and voltages results in control of spatial orientation of the electromagnetic fields in the machine and has led to term field orientation [1]. Field oriented control schemes provide significant improvement to the dynamic performance of ac motors. The usual method of induction motor position and torque control, which is becoming an industrial standard [2], uses the indirect field orientation principle in which the rotor speed is sensed or estimated by rotor position and slip frequency is added to form the stator impressed frequency. This paper proposes an indirect field oriented control applied to a small squirrel-cage induction motor using space vector pulse width modulation(SVPWM) for position and torque control using a digital signal processor and relay feedback method to evaluate the controller parameters, in such a way to show that AC motor can have similar performance in servo systems compared to DC motors.**

***Keywords* – Vector Control, Space Vector, Digital Signal Processor, PID Controllers, Relay Feedback, Modified Ziegler-Nichols Method..**

I. INTRODUCTION

Modern industrial processes place stringent requirements on industrial drives by way of efficiency, dynamic performance, flexible operating characteristics, ease of diagnostics and communication with a central computer. These coupled with the developments in micro-electronics and power devices have led to a firm trend towards digital control of drives. There is a wide variety of applications such as machine tools, elevators, mill drives, robots, etc., where quick control over the torque and position of the motor is essential. Such applications are dominated by DC drives and cannot be satisfactorily operated by an induction motor drive with constant volt/hertz (v/f) scheme. Over the last two decades the principle of vector control of AC machines has evolved, by means of which AC motors and induction motors in particular, can be controlled to give dynamic performance comparable to what is achievable in a separately excited DC drive. These controllers are called vector controllers because they control both amplitude and phase of the ac excitation. The vector control of currents and voltages results in control of the spatial orientation of the electromagnetic fields in the machine and has led to the term field orientation. Usually, this term is reserved for controllers which maintain a 90^o spatial orientation between critical field components and hence the term field angle control was adopted for systems which depart from the 90^o orientation. Indirect field orientation, used in this paper, makes use of the fact that satisfying the slip relation is a necessary and sufficient condition to produce field orientation [4]. On the other hand, the advantage of direct field orientation is the elimination of the rotor position encoder. Unfortunately, this eliminates the direct knowledge of a significant disturbance to the system, i.e., the motor speed and rotor position. While there are schemes which can successfully overcome this loss of information, they require new sensors which are in many ways less desirable than the encoder, such as hall elements in the air gap. Other elements which only require voltage and current measurements are also feasible, but they are severely limited at low speed and introduce new and very troublesome parameter dependencies [4]. These are the main reasons for the chosen of indirect field oriented control. SVPWM refers to a special technique of determining the switching sequence of the upper three power transistors of a three-phase voltage source inverter. It has been shown to generate less harmonic distortion in the output voltages or current in the windings of the motor load [3]. SVPWM provides more efficient use of the dc bus voltage, in comparison with the direct sinusoidal modulation technique. Also, it has lower base band harmonics than regular PWM or other sine based modulation methods, or otherwise optimizes harmonics, prevents un-necessary switching hence less commutation losses and has a different approach to PWM modulation based on space vector representation of voltages in *d-q* plane. [3]. The controller used was a PID, whose parameters where calculated using modified Ziegler-Nichols method, which has many advantages compared to other PID parameter tuning techniques, and has became an industrial standard [10]. This paper proposes a digital position and torque control of a squirrel-cage induction motor using indirect field oriented control and SVPWM using a 32 bit DSP TMS320F2812, , which enable enhanced real-time algorithm and cost-effective design of intelligent controllers of induction motors.

II. DYNAMIC MODELING OF THE INDIRECT FIELD-ORIENTED INDUCTION MOTOR SERVO DRIVE

The block diagram of the experimental indirect field-orientated induction motor position servo drive is shown in reference [5]. The drive mainly consists of an induction servo motor, a ramp comparison current SVPWM inverter, a field orientation mechanism, a coordinate translator, an inner speed control loop and an outer position control loop. The induction servo motor used in this drive system is a three-phase Y-connected, 4-pole, ¼ HP, 60 Hz, 220V 0.66 A. The state equations of an induction motor in the synchronously rotating reference frame can be written as follows [5]:

$$\frac{d}{dt}\begin{bmatrix} i_{ds} \\ i_{qs} \\ \lambda_{dr} \\ \lambda_{qr} \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{\sigma L_s}-\frac{R_r(1-\sigma)}{\sigma L_r} & \omega_e & \frac{L_m R_r}{\sigma L_s L_r{}^2} & \frac{P\omega_r L_m}{2\sigma L_s L_r{}^2} \\ \omega_e & -\frac{R_s}{\sigma L_s}-\frac{R_r(1-\sigma)}{\sigma L_r} & \frac{-P\omega_r L_m}{2\sigma L_s L_r{}^2} & \frac{L_m R_r}{\sigma L_s L_r{}^2} \\ \frac{L_m R_r}{L_r} & 0 & -\frac{R_r}{L_r} & \omega_e-\frac{P}{2}\omega_r \\ 0 & \frac{L_m R_r}{L_r} & -(\omega_e-\frac{P}{2}\omega_r) & -\frac{R_r}{L_r} \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \\ \lambda_{dr} \\ \lambda_{qr} \end{bmatrix} \quad (1)$$

$$+\frac{1}{\sigma L_s}\begin{bmatrix} v_{ds} \\ v_{qs} \\ 0 \\ 0 \end{bmatrix}$$

$$T_e = \frac{3P}{4}\frac{L_m}{L_r}(i_{qs}\lambda_{dr} - i_{ds}\lambda_{qr}) \quad (2)$$

where:

R_s : Stator resistence per phase L_s : Stator magnetising inductance per phase

R_r : Rotor resistence per phase referred to stator

L_r : Rotor magnetising inductance per phase referred to stator

L_m : Magnetising inductance per phase P : Number of Poles

ω_e : electrical angular speed ω_r : slip angular speed

v_{ds} : d-axis stator voltage v_{qs} : q-axis stator voltage

i_{ds} : d-axis stator current i_{qs} : q-axis stator current

$$\sigma = 1 - \frac{L_m{}^2}{L_s L_r} \quad \lambda_{qr} = L_m i_{qs} + L_r i_{dr} \quad \lambda_{dr} = L_m i_{ds} + L_r i_{qr}$$

The dynamic model of the induction motor and the whole drive system can be much simplified by using field-orientated control as shown in Fig. 1 [2]. In an ideal field-orientated induction motor, decoupling between d and q-axes is achieved, and the total rotor flux linkage is forced to align in the d-axis. Accordingly, the flux linkage and its derivative in the q-axis are set to zero:

$$\lambda_{qr} = 0 \text{ and } \frac{d\lambda_{qr}}{dt} = 0 \quad (3)$$

The rotor flux linkage can be found from the third row of equation 1 and from using equation 3 as:

$$\lambda_{dr} = \frac{L_m i_{ds}}{1 + s\frac{L_r}{R_r}} \quad (4)$$

Compared with the time constant of the mechanical system, the time constant in equation 4 is assumed negligible and i_{ds} is set constant ($i_{ds} = i^*_{ds}$) for the desired constant rated rotor flux. Then equation 4 becomes:

$$\lambda_{dr} = L_m i^*_{ds} \quad (5)$$

From equations (3) and (5) the torque equation (2) is simplified to:

$$T_e = \frac{3P}{4}\frac{L_m^2}{L_r} i^*_{ds} \quad (6)$$

where i^*_{qs} denotes the torque current command generated from the torque controller $G_c(s)$. Indirect field orientation, the slip angular is needed to calculate the unit vector for coordinate translation. Using the forth row of equation 1 and equation 3, the slip angular frequency ω_{sl} can be estimated by:

$$\omega_{sl} = \frac{L_m R_r i^*_{qs}}{L_r \lambda_{dr}} = \frac{R_r i^*_{qs}}{L_r i^*_{ds}} \quad (7)$$

The generated torque, rotor speed and rotor angular position θ_r are related by:

$$\omega_r = s\theta_r = \frac{1/J}{s + B/J}[T_e(s) - T_L(s)] \quad (8)$$

where B denotes the viscous damping frequency and J denotes the inertia constant.

III. SPACE VECTOR PULSE WIDTH MODULATION (SVPWM)

SVPWM technique has become a popular PWM technique for three-phase voltage source inverters in applications such as control of AC induction and magnet and permanent-magnet synchronous motors [6]. It refers to a special switching scheme of six power transistors of a 3 phase power converter and generates minimum harmonic distortion to the current in the windings of a 3-phase induction motor. It also provides a more efficient use of supply voltage in comparison with the sinusoidal modulation method [7]. First one must convert the equivalent orthogonal projection of an

ABC system onto the two dimensional plane using d-q transformation, resulting in diagram showed in Figure 2:

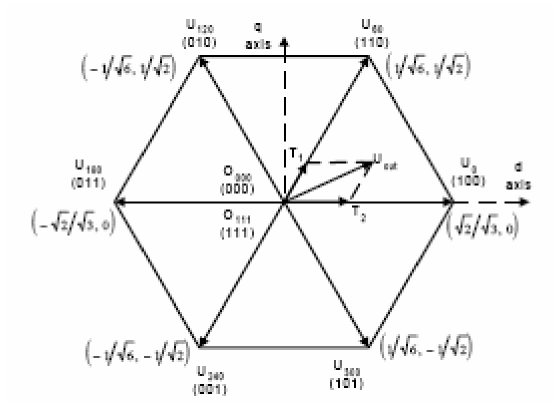

Fig. 1 – Space Vector Diagram

The objective of SVPWM is to approximate the reference voltage U_{out} instantaneously by combination of switching states corresponding to basic space vectors, which are based in the switching patterns of 3-upper power transistors in the converter. One way to achieve this is to require, for any small period of time T, the average inverter output be the same as the average reference voltage U_{out} as shown in equation (9):

$$U_{out}(nT) = \frac{1}{T}(T_1 U_x + T_2 U_{x+60}) \tag{9}$$

Which means that for every PWM period, U_{out} can be approximated by having the inverter in switching state U_x and U_{x+60} (or U_{x-60}) for T_1 and T_2 duration of time respectively. Since the sum of T_1 and T_2 should be less than or equal to T_{pwm}, the inverter needs to be in O_{000} or O_{111} state for the rest of the period. Therefore, equation (9) becomes:

$$T_{pwm} U_{out} = T_1 U_x + T_2 U_{x+60} + T_0 (0_{000}\, or\, 0_{111}) \tag{10}$$

where $T_0 = T_{pwm} - T_1 - T_2$.

IV. CONVERTING CONTROL VECTOR'S CURRENT COMMAND INTO SPACE VECTOR VOLTAGE COMMAND

The main problem using control vector algorithm into DSP is that, for space vector modulation, one must convert the current command calculated by control vector into voltage command used by space vector. To accomplished this task it is necessary to decouple the voltage equation so as to be able to independently control the two components of stator current.

The development of this decoupling can be seen in [4], which gives:

$$v_{qs}^e = (r_s + L_s^{'} s) i_{qs}^e + \omega_e L_s i_{ds}^e \tag{11}$$

$$v_{ds}^e = r_s i_{ds}^e - \omega_e L_s^{'} i_{qs}^e \tag{12}$$

Where:

$$L_s^{'} = L_s - \frac{L_m^2}{L_r} \tag{13}$$

V. MODIFIED ZIEGLER-NICHOLS METHOD

If an arbitrary point on the Nyquist curve of open-loop system is chosen, the parameters of a PI controller can be calculated in way that this point can be allocated in a suitable position [8]. If the chosen point is described by polar coordinates:

$$A = G_P(i\omega_0) = r_a e^{i(\pi+\phi_a)} \tag{14}$$

which must be reallocated, using a controller, to:

$$B = G_t(i\omega_0) = r_b e^{i(\pi+\phi_b)} \tag{15}$$

Writing the frequency response of the controller as $C = r_c e^{i(\phi_c)}$ and using equations (11) and (12):

$$r_b e^{i(\pi+\phi_b)} = r_a r_c e^{i(\pi+\phi_a+\phi_c)} \tag{16}$$

The controller should thus be chosen so that:

$$r_b = \frac{r_c}{r_a} \tag{17}$$

$$\phi_c = \phi_b - \phi_a \tag{18}$$

For a PI controller this implies:

$$K = \frac{r_b \cos(\phi_b - \phi_a)}{r_a} \tag{19}$$

$$T_i = \frac{1}{\omega_0 \tan(\phi_a - \phi_b)} \tag{20}$$

The point to be moved is usually the ultimate point, which can be determined by relay feedback method [8]. Has been suggested by Pessen [9] to move the ultimate point to $r_b = 0.41$ and $\phi_b = 61^o$.

VI. SIMULATION RESULTS

Applying the relay feedback method for the first controller(i.e., torque command controller [5]), the result can be seen in Figure 3. Using the equation which determines the ultimate point [8]:

$$G(i\omega_u) = -\frac{\pi a}{4d} \tag{21}$$

where d denotes the relay amplitude and a is the output amplitude of the system, the point can be located at $r_a = 0.0628$ and $\phi_a = 0^o$. Using the Pessen suggestion, the following torque controller parameters are found: $K_p = 3.3616$ and $T_i = 0.0044$. The result is shown in Figure 2. Repeating the same method, but with the first controller connected(i.e., position command controller), one gets the result showed in Figure 3.

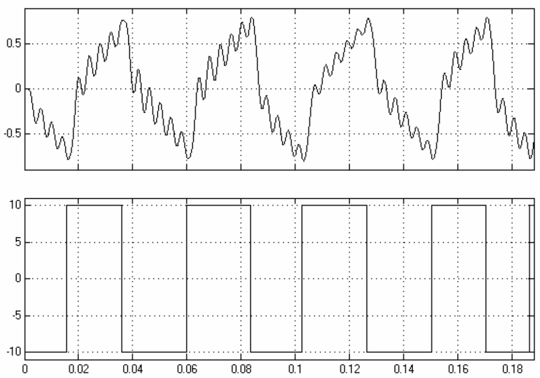

Fig. 2- Relay Feedback Applied to Torque

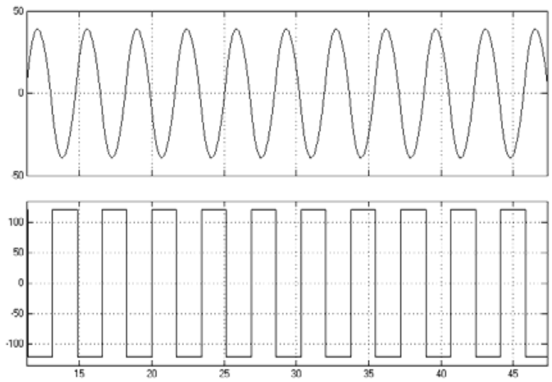

Fig. 3 – Relay Feedback Applied to Position Controller

the ultimate point in Nyquist curve for the system with torque PI controller is at $r_a = 0.2389$ and $\phi_a = 0^o$. Calculating the PI parameters using the same Pessen's criteria gives $K_p = 7.3432$ and $T_i = 0.1591$. Having proportional gain and integral time for both controllers, a reference position of 4 radians was applied to the whole system, and after 3 seconds this reference was changed to 6 radians. Results are showed in Figure 4.

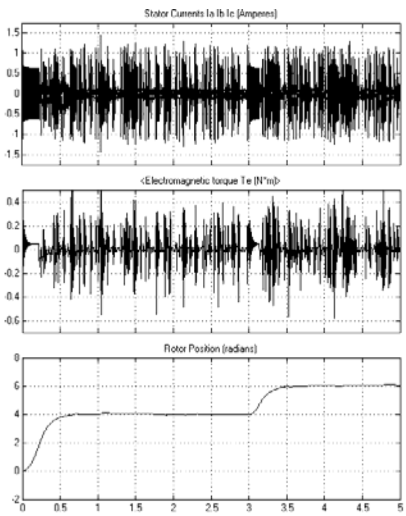

Fig. 4 – Stator Currents, Electromagnetic Torque and Rotor Position results from simulation tests.

An error of 0.5% was obtained at steady state. It's particularly hard to follow the position reference for this machine because of its low moment of inertia and damping coefficient. This can be clearly seen analyzing the stator currents, which are considered the control effort of the system.

VII. CONCLUSION

This paper proposes an algorithm to simulate control vector and space vector applied to a DSP, which uses voltage command instead of current command seen in control vector theory. Modified Ziegler-Nichols method was used to calculate the PID controller parameters, showing satisfactory results in simulation tests. Using the equations proposed in this paper one can build a block diagram in Simulink®, which generates the code to be applied to a DSP.

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