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Advances in vector control of *ac* motor drives – A review

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Abstract. This paper attempts to present a comprehensive review of the advances made in vector control or field orientation as applied to high performance *ac* motor drives. Brief application survey, machine models in *d-q* representation, implementation issues with inverters and cycloconverters, parameter effects etc for both induction and synchronous motor vector control are dealt with and sample results from studies on them are presented. The latest advance on this control like direct torque control (DTC) has been briefly discussed. A substantial updated bibliography, though by no means complete, is included for those who are interested in keeping track of the present state-of-the-art and working further in this area.

Keywords. Vector control; field orientation; *ac* motor drives; high performance drives; induction motor; synchronous motor; direct torque control.

1. Introduction

Electric drives for motion control must have a fast torque response, four quadrant operation capability and controllability of torque and speed over a wide range of operating conditions. A separately excited *dc* motor, earlier used as the primary machine and later with simple power electronic controllers and current feedback, provides direct control of the magnitude of armature current and, in proportion, the torque, and has been the most popular choice for many industrial drives for such requirements in spite of its inherent drawback of the bulky, expensive and maintenance-prone commutator. On the other hand, *ac* motors, specially induction motors with their simple, less expensive, and more robust structures are more suitable for industrial environments though their control is quite complex. This is due to the fact that the rotor current in an induction motor which is responsible for the torque production owes its origin to the stator current which also contributes to the air-gap flux resulting in a coupling between the torque- and flux-producing mechanisms. In the *dc* machine, the field current in the stationary poles producing the magnetising flux and the armature current directly controlling the torque are independently accessible. Moreover, for a fully compensated *dc* motor, the spatial angle between the flux and the armature mmf is held at 90° with respect to each other, independent of the load, by the commutator and the brushes whereas in an *ac* motor (both induction and synchronous), the spatial angle between the rotating stator and rotor fields varies with the load and gives rise to oscillatory dynamic response. Control methods for *ac* motors that emulate the *dc* motor control by orienting the stator current so as to attain independent and 'decoupled' control of flux and torque are known as 'field orientation' control and require control of both the magnitude and phase of *ac* quantities and thus are referred to as 'vector control methods'.

Early conceptual works in vector control were by Blaschke (1972) and Hasse (1969), which were translated into practical implementation later by Gabriel *et al* (1980), Leonhard (1985) and many others with the advances in microprocessors and microcomputers along with power electronics. Now, it has been established as a powerful technique in the field of *ac* motor drives and adopted worldwide. An exhaustive list of publications has been reported in this topic, which includes an IEEE Tutorial Course (Novotny & Lipo 1985) and two exclusive books (Vas 1990; Boldea & Nasar 1992). Work has continued unabated in this field and several issues like simplification of practical system with advanced microprocessors, design of current regulators/flux observers, reliability enhancement, performance improvement, parameter adaptation etc. are still attracting the researchers in this field. This paper attempts to make a summary review of the progress in vector control as applied to both induction and synchronous motor drives highlighting some typical results from the drives developed by the author and his research students at the Indian Institute of Technology, Kharagpur.

2. Vector control of induction motors

2.1 *Brief application survey*

The principle of vector control is used in current regulated PWM inverter (CRPWM), CSI, VSI, and cycloconverter-fed induction motor drives. The controlled current operation of the motor results in simpler implementation. The CRPWM inverter is common for high performance servo drives while CSI and cycloconverters are used for larger drives. High frequency PWM transistor inverters (10 kHz), developed around 1979, made it possible to use vector controllers in various kinds of industries including pinch roll drives of continuous casting plates, machine-tool drives and gear-less servo drives as reported by Kume & Iwakane (1987). The control method was applied to a large-scale paper mill (Tanaka *et al* 1983) with induction motors of 300–560 kW rating using CSI. Application of vector controlled induction motors for high performance servo drives has been brilliantly surveyed by Leonhard (1986). High horsepower vector controlled induction motor servo drive using adaptive rotor flux observer has been recently developed with improved steady state and dynamic response (Huang *et al* 1994). The recent trend is to eliminate the speed and position sensors in high performance vector controlled induction motor drives (Okuyama *et al* 1990; Onishi *et al* 1994; Tajima *et al* 1995). Very high power (MW) range cycloconverter-fed induction motors, with vector control for steel mill drive are mature drive systems in Japan (Sugi *et al* 1983; Saito *et al* 1987) and Germany (Timpe 1982; Hasse 1977). *Siemens* has recently announced optimised vector controlled SIMOVERT master drives for elevator applications (Scheirling & Schonherr 1995) having many important features.

2.2 *Induction motor model and basic vector control equations*

2.2a *Dynamic model:* A dynamic model developed either with the concept of space phasors (Leonhard 1985; Murphy & Turnbull 1988) or d-q representations (Novotny & Lipo 1985; Bose 1986) may be utilised to develop the basic machine equations for implementation of vector control. We like to use the latter for convenience and familiarity. The d-q axes model of an induction motor with reference axis rotating at synchronous speed w_e is

$$\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + \sigma L_s p & -\sigma L_s \omega_e & \frac{L_m}{L_r'} p & -\frac{L_m}{L_r'} w_e \\ \sigma L_s \omega_e & R_s + \sigma L_s p & \frac{L_m}{L_r'} w_e & \frac{L_m}{L_r'} p \\ -L_m \frac{R_r}{L_r'} & 0 & \frac{R_r'}{L_r'} + p & -\omega_{sl} \\ 0 & -L_m \frac{R_r'}{L_r'} & \omega_{sl} & \frac{R_r'}{L_r'} + p \end{bmatrix} \begin{bmatrix} i_{ds}^e \\ i_{qs}^e \\ \psi_{dr}^{e'} \\ \psi_{qr}^{e'} \end{bmatrix}, \tag{1}$$

where

$$\sigma = 1 - \frac{L_m^2}{L_s L_r'}, \quad p = \frac{d}{dt}, \quad \omega_{sl} = (\omega_e - \omega_r).$$

The electromagnetic torque developed by a 3-phase, P-pole, induction motor is

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r'}(i_{qs}^e \psi_{dr}^{e'} - i_{ds}^e \psi_{qr}^{e'}), \tag{2}$$

where

$$\psi_{dr}^{e'} = L_m i_{ds}^e + L_r' i_{dr}^{e'}, \tag{3}$$

$$\psi_{qr}^{e'} = L_m i_{qs}^e + L_r' i_{qr}^{e'}. \tag{4}$$

The field orientation implies that the stator current components obtained be oriented in phase (flux component) and in quadrature (torque component) to the flux vector which can be either stator flux (ψ_s), airgap or mutual or magnetising flux (ψ_m), or rotor flux (ψ_r) as shown in the equivalent circuit in figure 1 (Sathiakumar *et al* 1986). The orientation of the stator current with respect to the stator, rotor and airgap flux has been examined and the relative merits and developments of the schemes have been reported (Bayer & Blaschke 1977; Sathiakumar *et al* 1986; Ho & Sen 1988; Erdman & Hoft 1990). It has been shown that the rotor flux orientation alone provides natural decoupling, fast torque response and all round stability. The stator flux and airgap flux orientation, however, are attractive due to ease of flux computation and for the purpose of wide range of field weakening operation (Xu & Novotny 1992) but need decoupler network (De Doncker & Novotny 1988). A new strategy called the 'Universal field oriented controller' has been developed by De Doncker & Novotny (1988) which decouples flux and torque in an arbitrary flux reference frame.

Rewriting the rotor voltage equations in (1)

$$p\psi_{dr}^{e'} + \frac{R_r'}{L_r'}\psi_{dr}^{e'} - \frac{L_m}{L_r'}R_r' i_{ds}^e - \omega_{sl}\psi_{qr}^{e'} = 0, \tag{5}$$

$$p\psi_{qr}^{e'} + \frac{R_r'}{L_r'}\psi_{qr}^{e'} - \frac{L_m}{L_r'}R_r' i_{qs}^e + \omega_{sl}\psi_{dr}^{e'} = 0. \tag{6}$$

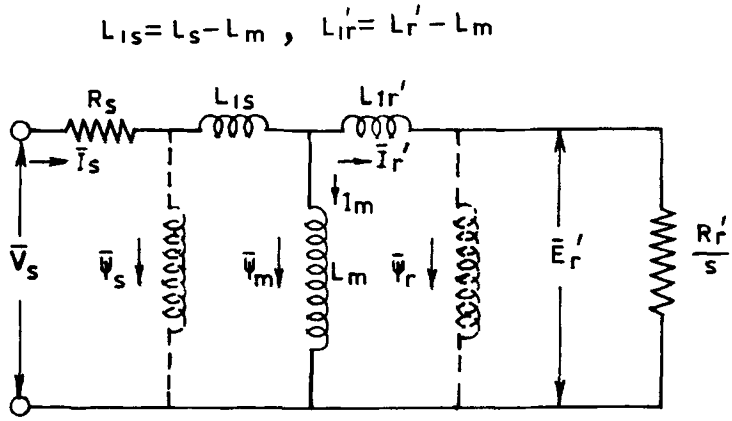


Figure 1. Conventional stator referred induction motor equivalent circuit showing different flux vectors.

For rotor flux orientation control, the rotor flux axes are locked with the synchronously rotating reference system such that the rotor flux is entirely in the d-axis,

$$\psi_r^{e'} = \psi_{dr}^{e\,'}, \tag{7}$$

$$\psi_{qr}^{e'} = 0. \tag{8}$$

Substituting (7) & (8) in (5) & (6) yields

$$\omega_{sl} = \frac{L_m}{\psi_r'^e}\left(\frac{R_r'}{L_r'}\right) i_{qs}^e, \tag{9}$$

$$\frac{L_r'}{R_r'} p\psi_r'^e + \psi_r'^e = L_m i_{ds}^e. \tag{10}$$

For the range of operation below the base speed, the flux $\psi_r^{e'} = \psi_{dr}^{e'}$ is kept constant, when

$$p\psi_{dr}^{e'} = 0. \tag{11}$$

From (4) & (8),

$$i_{qs}^e = -\frac{L_r'}{L_m} i_{qr}', \tag{12}$$

which shows a direct equilibrium relation between the torque component current i_{qs}^e and the rotor current $i_{qr}'^e$. The torque equation is

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m}{L_r'} i_{qs}^e \psi_r'^e, \tag{13}$$

which shows the desired property of providing a torque proportional to the torque command i_{qs}^e.

During flux changes in the transient, $p\psi_{dr} \neq 0$ and from (10)

$$i_{ds}^e = \frac{\psi_{dr}'^e - L_m i_{ds}^e}{L_r'}. \tag{14}$$

Combining (14), (4) & (8) to eliminate $i_{dr}^{e'}$, yields the equation relating i_{ds}^{e} and $\psi_{dr}^{e'}$ (flux command and the flux),

$$(R_r' + L_r' p)\psi_{dr}^{e'} = R_r' L_m i_{ds}^{e}, \tag{15}$$

which in the steady state is

$$\psi_{dr}^{\prime e} = L_m i_{ds}^{e}. \tag{16}$$

The close parallel to the *dc* machine is now clearly visible. With the flux command held constant, a change in i_{qs}^{e} is followed instantly by corresponding change in $i_{qr}^{\prime e}$. While with a change in flux command, a transient rotor current is induced which subsequently decays with the rotor open circuit time constant L_r'/R_r' as shown in (15).

2.2b *Steady state model:* A convenient steady-state equivalent circuit model of the field oriented induction motor as shown in figure 2 can be obtained from the conventional equivalent circuit (figure 1) by using a referral ratio $a = L_m/L_r'$ in lieu of the common choice of the stator to rotor turns ratio (Novotny & Lipo 1985). With adoption of this ratio, the stator current is seen to be subdivided into the orthogonal components $I_{s\psi}$ (flux component) and I_{sT} (torque component), equivalent to i_{ds}^{e} and i_{qs}^{e} referred in the dynamic model, and the slip relation (9) can be obtained by equating the voltages across the parallel branches as

$$\omega_{sl} = s\omega_e = \frac{R_r'}{L_r'}\frac{I_{sT}}{I_{s\psi}}. \tag{17}$$

Equation (17) expresses the co-ordination between the slip and the current components required to attain correct field orientation as relevant for indirect vector control discussed later. The torque expression is obtained from the airgap power as

$$T_e = \frac{3}{2}\frac{P}{2}\frac{L_m^2}{L_r'} I_{s\psi} I_{sT}, \tag{18}$$

which shows the desired torque control via current components $I_{s\psi}$ and I_{sT}.

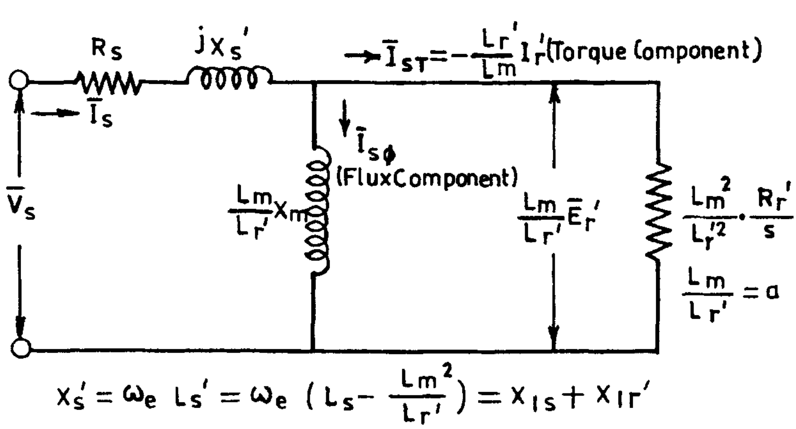


Figure 2. Derived steady state equivalent circuit for rotor flux orientation scheme.

2.3 *Induction motor vector control implementation*

The implementation of vector control requires information regarding the magnitude and the position of the flux vector (stator, rotor or mutual, as the case may be) and fast control of stator current in both magnitude and phase. Depending upon the method of flux acquisition, the vector control can be direct (Blaschke 1972) or indirect (Hasse 1969). The universal field oriented controller developed by De Doncker & Novotny (1988) is applicable to both these field orientation schemes and the generalised approach by Ogaswara *et al* (1988) to indirect control of induction and synchronous motors.

2.3a *Direct field orientation:* In the direct method, also known as flux feedback method, the airgap flux is directly measured with the help of sensors such as Hall probes, search coils or tapped stator windings (Zinger *et al* 1990) or estimated/observed from machine terminal variables such as stator voltage, current and speed (Jansen *et al* 1994). Since it is not possible to directly sense rotor flux, it is synthesised from the directly sensed airgap flux using the following equations

$$\psi_{dr}^{s'} = \frac{L_r'}{L_m}\psi_{dm} - L_r' i_{ds}^{s}, \tag{19}$$

$$\psi_{qr}^{s'} = \frac{L_r'}{L_m}\psi_{qm} - L_r' i_{qs}^{s}. \tag{20}$$

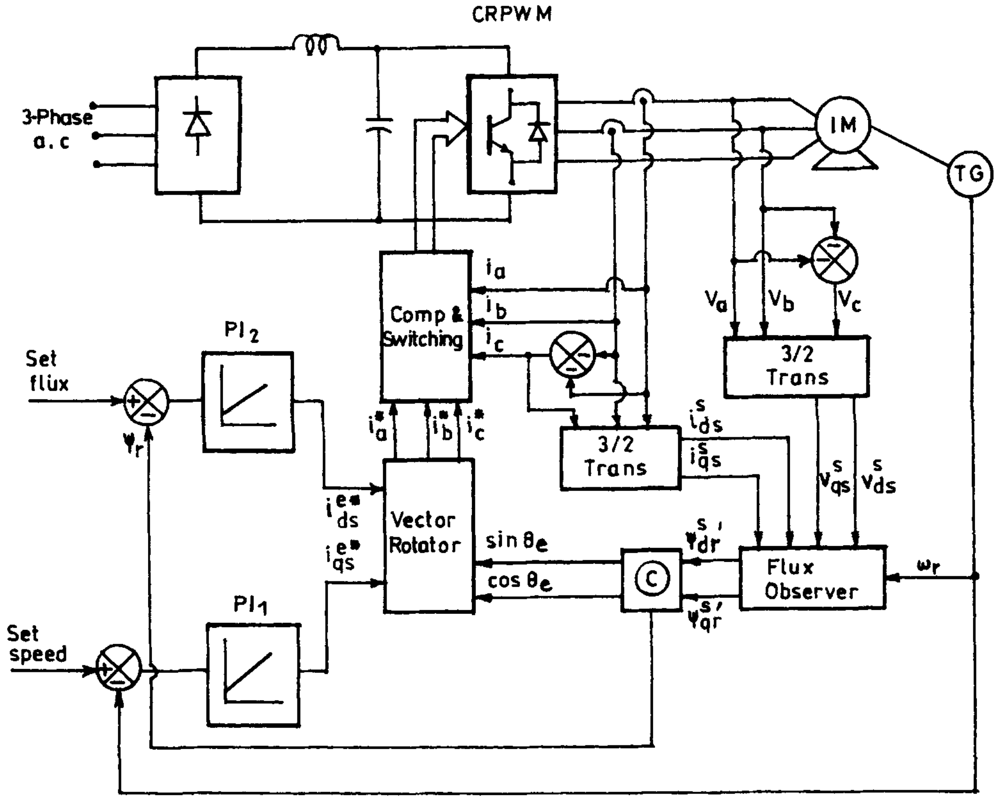
A variety of flux observers can be employed to estimate and improve the flux response with less sensitivity to machine parameters as detailed by Verghese & Sanders (1988) and Atkinson *et al* (1991). A major drawback with the direct orientation schemes is their inherent problem at very low speeds when the machine IR drop dominates and the required integration of the signals to measure the airgap flux is difficult. Closed-loop stator flux observers based on the motor current, voltage and the measured rotor position have been found to obviate this difficulty (Jansen *et al* 1993; Lorenz *et al* 1994).

A rotor flux observer based direct vector control scheme as implemented in the laboratory is shown in figure 3 (Chattopadhyay *et al* 1992; Thakur 1996) using a CRPWM inverter with flux and torque regulating loops. The vector rotator block implements the transformation from rotating to stationary axes followed by a 2/3 phase transformation resulting in the following expressions

$$\left.\begin{aligned} i_a^* &= i_{qs}^{*e}\cos\theta_e + i_{ds}^{*e}\sin\theta_e \\ i_b^* &= (-\tfrac{1}{2}i_{qs}^{*e} - \tfrac{\sqrt{3}}{2}i_{ds}^{*e})\cos\theta_e + (\tfrac{\sqrt{3}}{2}i_{qs}^{*e} - \tfrac{1}{2}i_{ds}^{*e})\sin\theta_e \\ i_c^* &= (-\tfrac{1}{2}i_{qs}^{*e} + \tfrac{\sqrt{3}}{2}i_{ds}^{*e})\cos\theta_e - (\tfrac{\sqrt{3}}{2}i_{qs}^{*e} + \tfrac{1}{2}i_{ds}^{*e})\sin\theta_e \end{aligned}\right\}. \tag{21}$$

The speed loop control provides the torque command whereas the flux command is selected according to the operating requirements in either constant torque or constant horsepower region. For CRPWM inverter, line currents are controlled in such a way as to follow the reference current commands generated from the vector rotator.

2.3b *Indirect vector control:* An alternative to direct measurement or estimation of the flux position for application of vector control to the induction motor without flux sensors is



$\boxed{C}$ $\quad |\hat{\Psi}_r| = \sqrt{(\hat{\Psi}_{dr}^{s\prime})^2 + (\hat{\Psi}_{qr}^{s\prime})^2}$

$$\cos\theta_e = \frac{\hat{\Psi}_{dr}^{s\prime}}{|\hat{\Psi}_r|}, \quad \sin\theta_e = \frac{\hat{\Psi}_{qr}^{s\prime}}{|\hat{\Psi}_r|}$$

Figure 3. A rotor flux observer based direct vector control scheme for an induction motor with a CRPWM inverter.

to employ the slip relation (9) to compute the flux position relative to the rotor by summing a sensed rotor position signal with a commanded slip position signal

$$\theta_e^* = \theta_{sl}^* + \theta_r. \tag{22}$$

Figure 4 illustrates the basic structure of an indirect field orientation scheme using a CRPWM inverter (Thakur *et al* 1993; Thakur 1996). The commanded currents i_{qs}^{e*} and i_{ds}^{e*} are converted to stator referred reference currents by rotating to stationary and 2/3 phase transformations as in the case of direct field orientation. i_{qs}^{e*} is controlled according to the desired torque and constant rotor flux. i_{ds}^{e*} is obtained from (16) in the steady state.

Indirect field orientation, also known as flux feed-forward control, does not have inherent low speed problems and is preferred in most systems which must have zero speed. However, the inherent limitation is in the slip calculation which depends on the commanded machine parameters that may differ from the actual values during running condition of the drive.

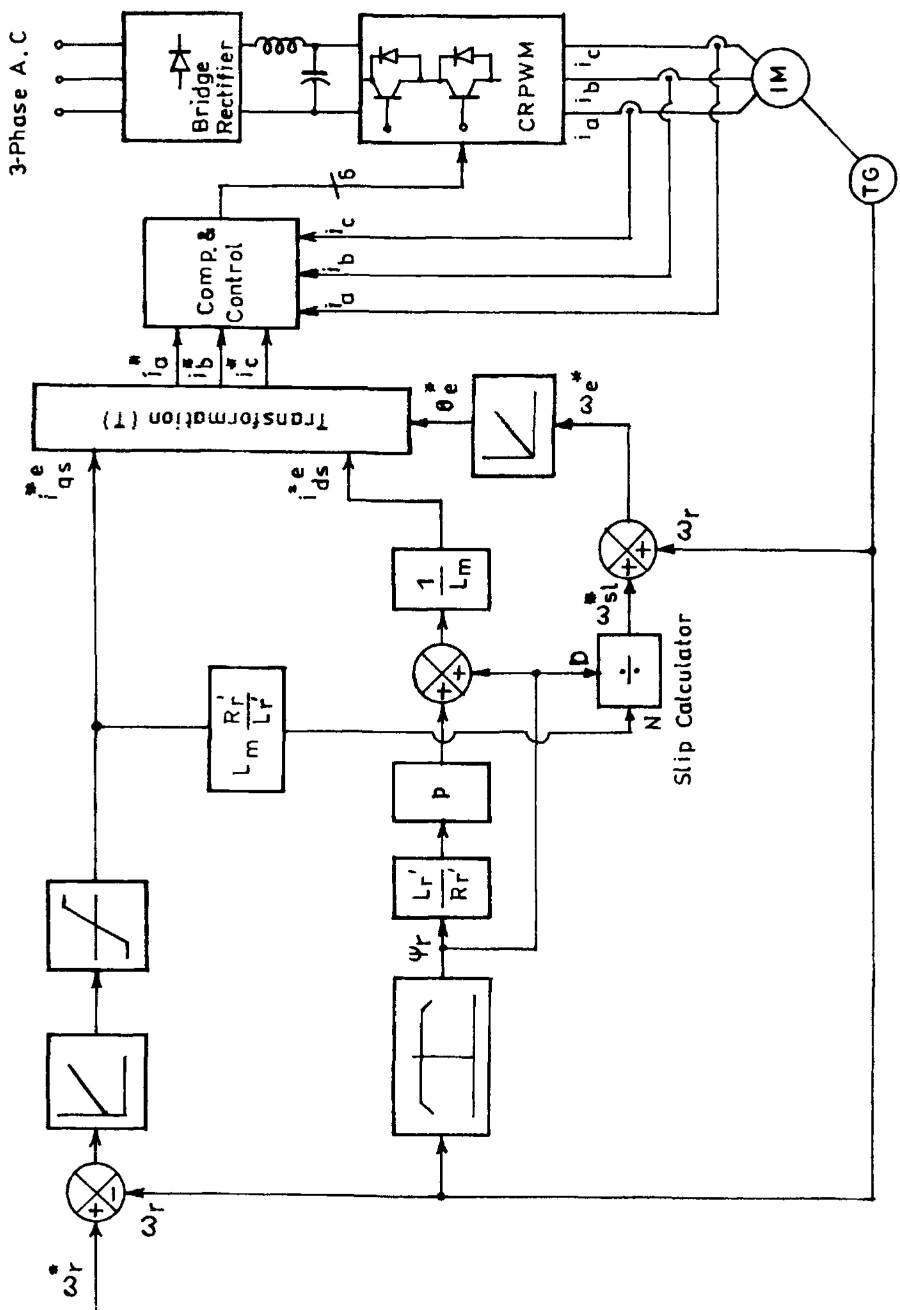


Figure 4. A rotor flux oriented indirect vector control scheme for an induction motor with a CRPWM inverter.

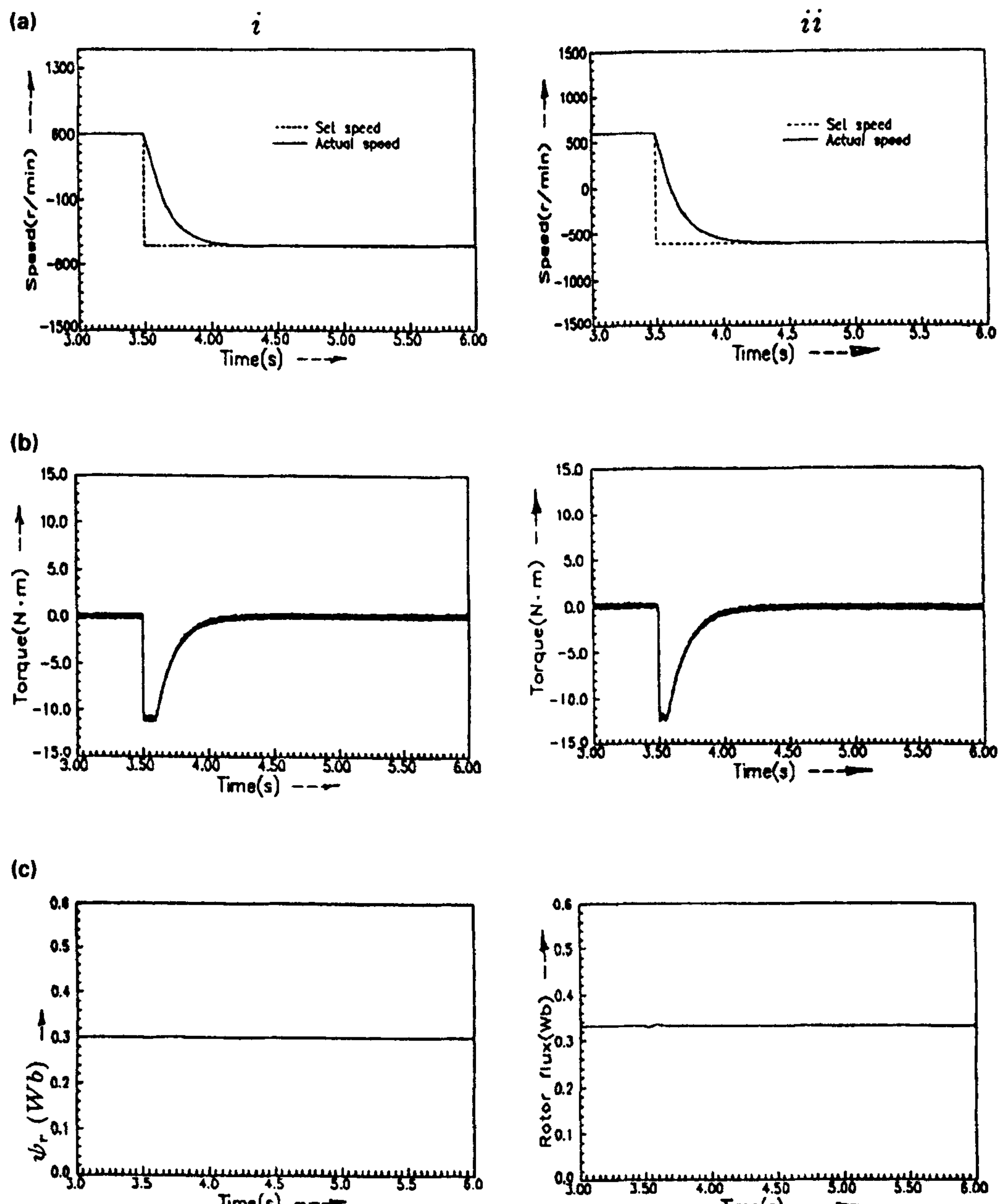


Figure 5. Simulation results showing (a) speed (b) torque, and (c) rotor flux of a vector controlled induction motor drive for speed reversal (600 to −600 rpm): (i) direct vector control, (ii) indirect vector control.

2.3c *Microprocessor-based controller and typical results:* With the availability of the advanced microprocessors, the implementation of vector control schemes has become simpler and cost effective as the differential equations involved are readily solved in real time. Beginning with an 8-bit microprocessor, a 16-bit or a 32-bit or now a DSP, a transputer or a custom-made LSI chip has become a part of the vector control hardware (Gabriel *et al* 1980; Sathiakumar *et al* 1986; Mingbao *et al* 1987; Wu & Strangas 1988; Asher & Sumner 1990; Ho & Sen 1990; Xu & Novotny 1991; Kao & Lin 1992; Lakaparampil 1994).

Multi-microprocessor configuration has also been used to implement sophisticated control structure (Harashima *et al* 1985; Saito *et al* 1987; Tzou & Wu 1990). The limitation in microprocessor application due to its finite word length, execution time and the operational instructions must be taken into account in designing a processor based system as they affect significantly the performance of the system (Dote 1988; Jelassi *et al* 1992). A faster operation may be obtained in a hybrid scheme (Thakur *et al* 1993; Thakur 1996) using both analog hardware and microprocessor based controller where tasks such as 2/3 phase transformation and PWM switching signal generation are achieved with hardware and the PI controller/observer design and implementation by a microprocessor with a PC-XT. Digital computer simulation technique is preferred to optimise the effects of various factors before implementation. Few typical simulation results as obtained and experimentally verified by Thakur (1996) are shown in figure 5 for both indirect and direct vector control. It is seen as expected that the performance of the latter is somewhat superior.

2.4 *Effects of motor parameter variations and adaptation*

Both the schemes described used machine parameters either in the calculation for the slip command for implementing the indirect vector control or to synthesise the flux vector to implement the direct vector control. In the indirect control the main problem is the rotor circuit time constant L_r'/R_r' which is sensitive to both temperature and flux level (Nordin *et al* 1985; Krishnan & Bharadwaj 1991; De Doncker 1994). Direct field orientation systems are sensitive to stator resistance and total leakage inductance but, typically, the parameter sensitivity is less here than that with the indirect control, specially because of the flux regulation through feedback. With deviation of parameters, the field orientation is not perfect and the controller should track the machine parameters. Several methods of parameter adaptation have been attempted (Garces 1980; Matsuo & Lipo 1985; Dalal & Krishnan 1987; Krishnan & Doran 1987; Nilsen & Kazmeirkowski 1989; Bal & Grant 1992; Ghosh & Bhadra 1992), along with a number of identification schemes including Model Reference Adaptive control (Ohnishi *et al* 1986; Holtz & Thimm 1989; Vas 1990; Bal & Grant 1992; Moriera & Lipo 1993). Automated initial tuning in the form of self-commissioning technologies has also been developed (Khambadkone & Holtz 1991; Lorenz *et al* 1994; Borgard *et al* 1995; Yanagawa *et al* 1995). Recent work on the on-line tuning to improve the robustness of vector control induction motor has been reported using special torque control strategy (Noguchi *et al* 1997; Tadakuma *et al* 1997) and feed forward/feedback control with neural network. A new flux and stator resistance identifier for *ac* drives has been proposed by Kerkman *et al* (1996).

Two new approaches to induction motor field-orientation are presented in Matsuo *et al* (1994) which employ rotor end ring current phase detection to make the controller independent of rotor time constant variations. However, it has been reported that the control performance is adequate within the normal operating temperature for most of the high performance applications and the parameter adaptation may be essential only in the case of critical applications. The parameter sensitivity in small machines is low enough to cause serious problems (Nordin *et al* 1985).

Issues regarding field-oriented controller for induction motors with double cage and deep bar rotor are discussed in Vas (1990). For these motors, the angular slip torque has to be calculated in such a way that it contains the effects of the deep bar or the double cage. Improved cage rotor models are developed by Healey *et al* (1995).

2.5 *Effects of magnetic saturation and core loss*

The flux level in an induction machine is a function of both the stator and the rotor currents. Both the performance and the losses are effected by its selection (Khater *et al* 1987). Normal modelling of the machine will not remain valid under magnetic saturation, particularly so under dynamic condition. The saturation effects for vector-controlled machines have been considered by Lorenz & Novotny (1990) and Vas & Alakul (1990). Under saturation conditions, the peak torque per ampere is best produced by increasing the torque producing current command in proportion to the total stator current. The sensitivity of rotor flux estimation depends on the selection of the machine model (Levi & Vuckovic 1989, 1990). The load torque condition has been observed to play an important role in machine saturation (Ohm 1989).

Vector control principles have been traditionally derived on the assumption that the iron core loss may be neglected. However, recently, it has been shown (Levi 1995; Levi *et al* 1996) that the core loss introduces unwanted cross coupling leading to detuning and for compensation, a decoupling circuit for indirect rotor flux oriented control is suggested, which makes the controller more complex.

2.6 *Current, flux and torque regulators*

Current regulators for vector controlled *ac* drives are more complex than those for *dc* drives as both amplitude and phase of the stator current are to be controlled. Both CSI and PWM converters with current regulation are used. The current regulators classified into three groups, hysteresis, PI with ramp comparison PWM and predictive (optimal) voltage vector location have been adequately discussed by Lorenz *et al* (1994) and Lee *et al* (1994). The various solutions differ in implementation costs, robustness with respect to parameter variation and their ability to track current commands with high fidelity and low distortion.

Regulation of flux is limited by the estimation of the flux magnitude and angle in direct vector control. Both open loop and closed loop flux observers have been used for direct and indirect field orientation (Hillenbrand 1977; Bouch *et al* 1992; Jansen *et al* 1993; Lorenz *et al* 1994). It was shown that a position sensor along with a current sensor will facilitate a simple open-loop observer for rotor flux. The closed-loop observer with motor current, voltage and rotor position measurement using the best features of both the current model and the voltage model open-loop observers will give better performance – flux regulation as well as flux estimation.

Recently, fuzzy and neural network-based estimators of feedback signals such as rotor flux, unit vectors and torque for indirect and direct vector control schemes have been reported (Miki *et al* 1991; Sousa & Bose 1993; Simoes & Bose 1995). These have the advantages of faster execution speed, harmonic ripple immunity and fault tolerance characteristics compared to a DSP based estimation.

2.7 *Direct torque control (DTC)*

The latest control method developed and commercialised by ABB, Sweden from the concept of the field-oriented or vector control is direct torque control (DTC), a patented concept developed again in Germany by Depenbrock (1988). The basic control scheme is shown in figure 6 when both the flux and torque are controlled by a hysteresis controller (Tiitinen *et al* 1996; Nash 1997). The delays associated with the PWM stage are eliminated since the PWM modulation is replaced by an optimal switching (Space PWM) logic. The adaptive motor model estimates the actual torque, stator flux and shaft speed as well as the frequency. The flux and torque are calculated every 25 μs and the speed and the frequency once per millisecond. The input to the motor model includes the motor current for two stator phases, line voltage and power switch positions. The optimal switching logic is realised by ASIC hardware (ACS 600). The switch information for the power module is utilised in the calculation of the appropriate voltage vector which will satisfy both the torque status and flux status outputs. This method results in a better torque response than the flux vector control and, in addition, assuming moderate speed accuracy is acceptable (typically 0.1–0.3%), the need for a pulse encoder is eliminated. Implementations of special functions

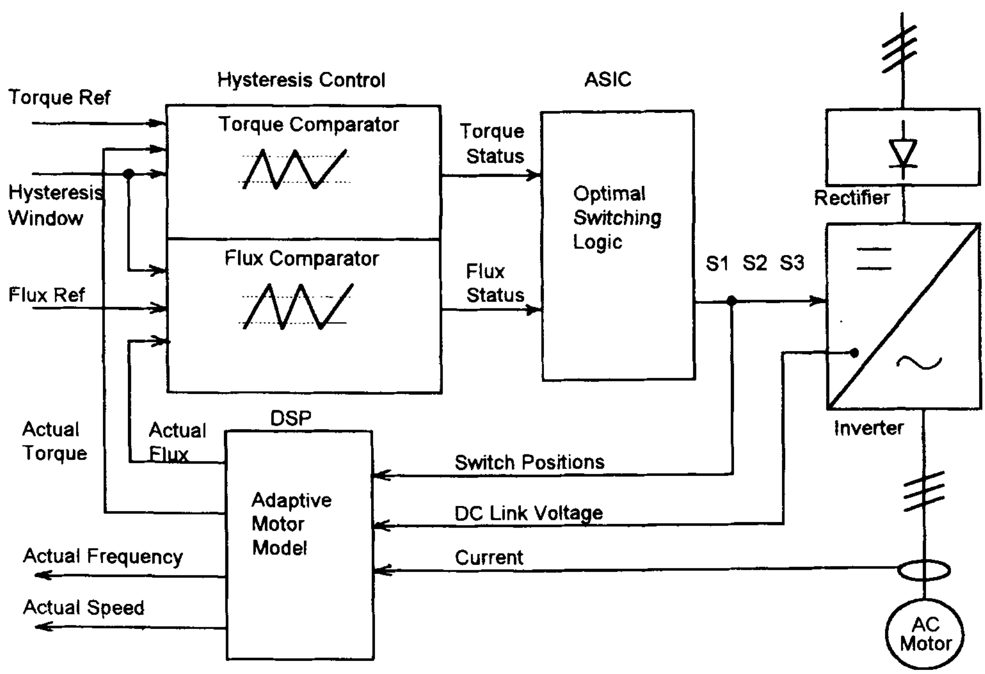


Figure 6. Direct torque control (DTC) scheme.

like flying start, flux braking, flux optimisation and powerloss ride through are all made easier with this control approach, as claimed.

2.8 *Doubly-fed and multiphase induction motor control*

The vector control of a doubly-fed slip ring induction motor in a Scherbius scheme as used in high power pump drives with a current controlled cycloconverter in the rotor side is amply described by Vas (1990) and Bose (1986) for super/sub-synchronous speed control. The same system can be used for VSCF generation systems, where the control strategy remains the same except that the active and the reactive currents of the cycloconverter are controlled to control real and reactive powers, respectively, at the stator terminals by the feedback method. A novel control strategy to realise torque and reactive power control of a doubly excited induction machine with position sensorless scheme using rotor voltage and currents as feedback signals has been proposed recently by Xu & Cheng (1995).

A strategy for improvement of the reliability for vector-controlled induction motor drive with a modified topology where the neutral point is returned to the midpoint on the *dc* link is proposed by Liu *et al* (1993). This allows for continuous disturbance-free operation of the drive even with complete loss of one leg of the inverter or motor phase. This method has been extended to field-oriented control for a multiphase induction machine with an unbalanced stator winding structure (Zhao & Lipo 1996)

3. Vector control of synchronous motors

3.1 *Brief application survey*

While vector-controlled induction motor drives have been used mostly in the industry for medium power ranges, vector-controlled synchronous motor drives are either in the very high power range (1–10 MW) with wound-field machines fed from cycloconverters or in the few kilowatt range with permanent magnet synchronous motors (PMSM) or synchronous reluctance motors for servo drives. The control of synchronous motors is different from that of induction motors primarily due to the fact that in the former, the magnetising current can be supplied from the field side independently of the armature current and the space position of the field is located by the position of the rotor. Additionally, the steady state slip between the rotor (which usually carries the field winding) and the controlled flux vector vanishes in the steady state. Therefore, the indirect or the feed-forward type of vector control as used extensively for the induction motor drives does not apparently seem obvious for a synchronous machine. The 'transvector control' as applied to a synchronous motor by Bayer *et al* (1972) is essentially a direct type of flux feedback control where the stator current is orthogonally oriented with respect to the stator flux vector to achieve unity steady state power factor. The decoupling is achieved by a closed loop flux feedback in addition to feeding a part of the magnetising current from the stator during the transient to compensate for a sluggish field current change. Siemens has reported (Timpe 1982; Pallmann 1992) the development of vector controlled cycloconverter-fed subsynchronous motor drives for use in reversing rolling mills to achieve high dynamical control response. Brown Boveri reported the development of the first gearless tube mill (Blauenstein 1970;

Stemmler 1970) using flux feed-forward control scheme. Terens *et al* (1982) used both static and dynamic flux models to control a similar drive. Nakano *et al* (1984) reported the development of a high performance synchronous motor drive for a rolling mill, with an open-loop flux estimator and PI current controller. An airgap flux oriented vector controlled cycloconverter drive was developed by Hill *et al* (1987) for an icebreaker. A very good survey of field-oriented control of synchronous machines including various applications has been made by Novotny & Jansen (1991) with a discussion on the difference between the 'space angle control' (SAC) relevant to self-synchronous commutatorless motor (CLM) and the true field-oriented (FO) or vector control. While in the former the angle of the armature current vector with respect to the field axis may be other than 90°, in the latter it is strictly restricted to 90°. Earlier, high power drives using CSI converters and wound field-synchronous motors with load commutation for fan and compressor drives were SAC systems utilising a rotor position detector to cause the power converter to supply stator excitation in synchronism with the induced voltage from the field excitation. PMSM motors operated in true FO system as used for servo drives and machine tools are reported by Kaufman *et al* (1982) and Wescheta (1983).

3.2 *Synchronous machine model and vector control implementation*

3.2a *Wound field synchronous motor model:* The d-q model of a wound field salient pole synchronous machine with damper windings in Park (rotor) reference frame is

$$\begin{bmatrix} v_{qs} \\ v_{ds} \\ 0 \\ 0 \\ v'_{fr} \end{bmatrix} = \begin{bmatrix} R_s + pL_{qs} & \omega_r L_{ds} & pL_{qm} & \omega_r L_{dm} & \omega_r L_{dm} \\ -\omega_r L_{qs} & R_s + pL_{ds} & -\omega_r L_{qm} & pL_{dm} & pL_{dm} \\ pL_{qm} & 0 & R'_{qr} + L'_{qr} & 0 & 0 \\ 0 & pL_{dm} & 0 & R'_{dr} + pL'_{dr} & pL_{dm} \\ 0 & pL_{dm} & 0 & pL_{dm} & R'_{fr} \\ & & & & +p(L'_{lfr} + L_{dm}) \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i'_{qr} \\ i'_{dr} \\ i'_{fr} \end{bmatrix}, \tag{23}$$

$$T_e = \frac{3}{2}\frac{P}{2}(\psi_{ds} i_{qs} - \psi_{qs} i_{ds}) \tag{24}$$

$$= T_L + \frac{2}{P} J \frac{d\omega_r}{dt}. \tag{25}$$

The field circuit parameters $v'_{fr}, i'_{fr}, R'_{fr}, L'_{lfr}$ refer to the stator. For steady state, p-terms vanish,

$$i_{qs} = I_{qs}, \quad i_{ds} = I_{ds}, \quad i'_{dr} = i'_{qr} = 0, \quad i'_{fr} = I_f.$$

3.2b *Vector control and angle control:* The rotor position feedback and vector control of the motor stator current to maintain the space angle between the field winding and the stator mmf results in stator currents that translate to set values of i_{qs} and i_{ds} in the rotor reference frame. This is due to the instantanenous control of the phase of the stator current to always maintain the same orientation of the stator mmf vector with respect to the field winding in the d-axis of the d-q model. The resulting axes current are shown in

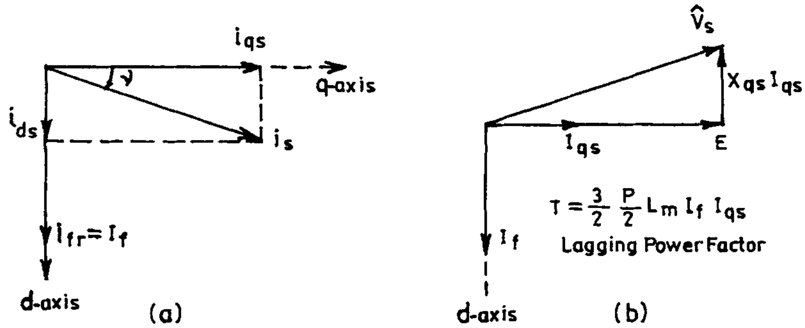


Figure 7. Phasor diagram of synchronous machine currents in d-q axes: (**a**) space angle control, (**b**) field orientation ($\gamma = 0$, $I_{ds} = 0$).

figures 7a and b (Novotny & Lipo 1985) for space angle control and the field orientation (when $\gamma = 0$, $I_{ds} = 0$). Note that for field orientation the field current in the d-axis and the stator current in the q-axis are 90° apart.

3.2c *Implementation with CSI and CRPWM inverter:* The implementation calls for control of magnitude and phase of the stator current with respect to the location of the field winding axis. Figure 8 shows a direct implementation ($\gamma = 0$) using absolute rotor position sensing and a CSI. With $\gamma = 0$, the stator current is entirely q-axis current and is equivalent to a torque command. The γ^* command is entered in the 'phase regulator' block and the drive can be operated at other than $\gamma^* = 0$. Figure 9 shows a simple means for implementing torque control with independent q-axis and d-axis currents using a CRPWM. The absolute rotor position information is used to convert the i_{qs}^{e*} and i_{ds}^{e*} commands in the rotor reference frame to a stator reference frame – which become the current commands for the CRPWM. Normal field orientation is obtained by setting $i_{ds}^{e*} = 0$. Varying i_{ds}^{e*} provides control of power factor and other varying performances. The 'rotor to stator transformation' block in figure 9 implements the same equations as in (21).

3.2d *Implementation with cycloconverter:* Cycloconverter-fed synchronous motors have been preferred for low speed large power drives e.g. mine hoist winders, gearless

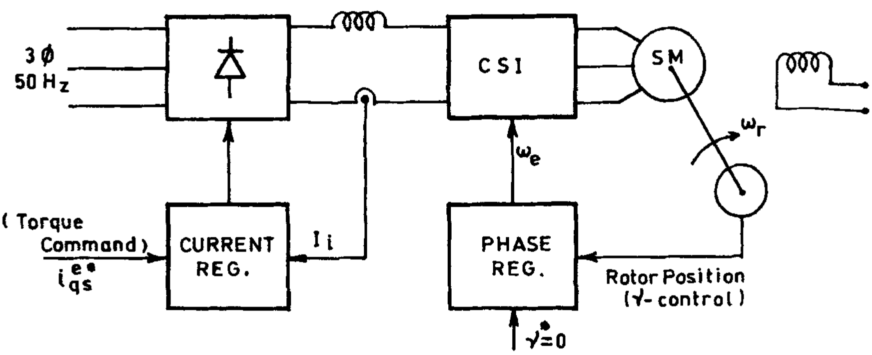


Figure 8. A synchronous motor vector control scheme using a CSI ($\gamma = 0$, $I_{ds}^{*e} = 0$).

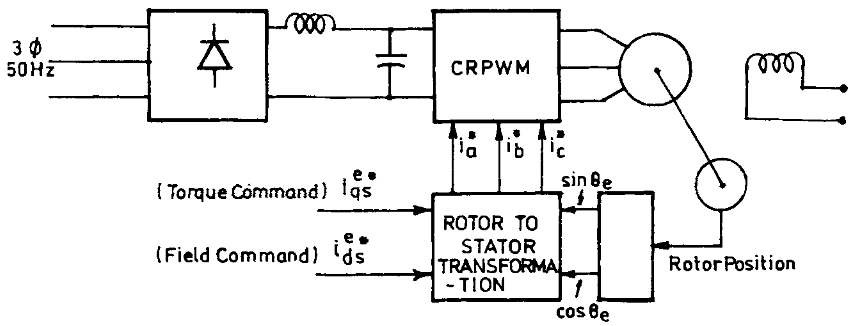


Figure 9. A synchronous motor vector control scheme using a CRPWM ($\gamma = 0$, $i_{ds}^{*e} = 0$).

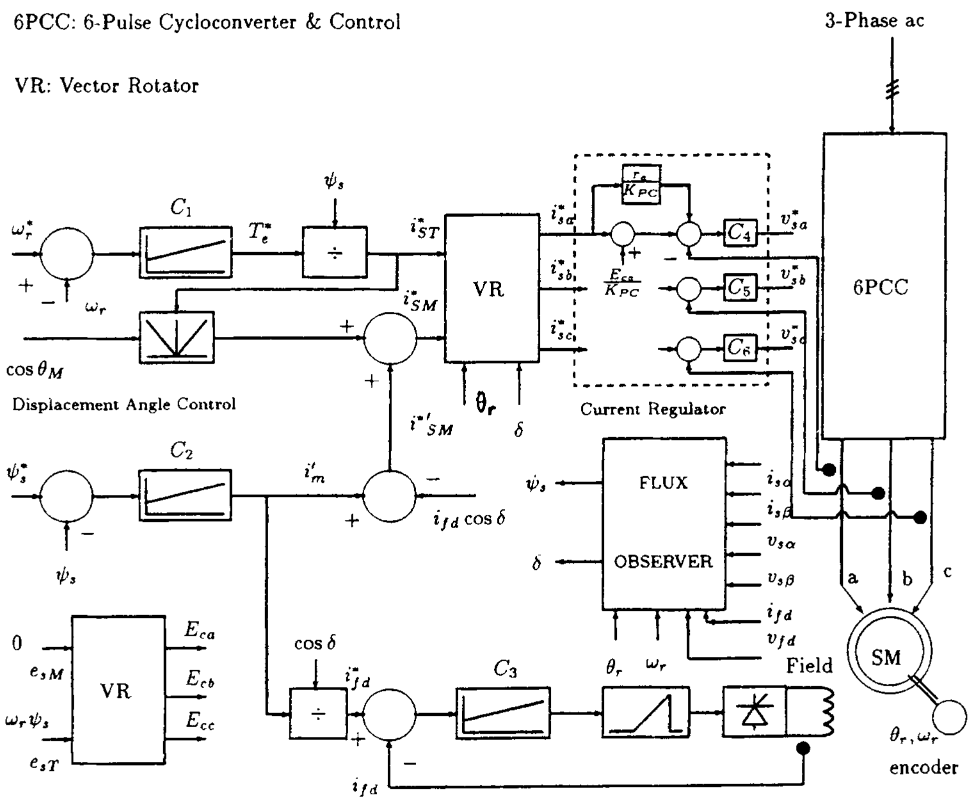


Figure 10. A stator flux oriented vector control scheme for a cycloconverter fed synchronous motor with a flux observer.

cement mill drives, rolling mill drives, ship propulsion drives etc. Synchronous motors have been preferred in these drives rather than induction motors because of their power factor and large torque capability at low speeds. Furthermore, a naturally commutated cycloconverter, compared to an inverter, provides a near-sinusoidal current excitation resulting in negligible torque ripple, inherent four quadrant capability, robustness and large power handling capability.

A stator flux-oriented vector control scheme which is an improvement of that in Bayer *et al* (1972) and Nakano *et al* (1984) for a 6-pulse non-circulating current cycloconverter-fed synchronous drive with a flux observer has been developed recently by Das (1996) and Das & Chattopadhyay (1997) for a rolling mill drive. Figure 10 shows the implementation scheme which aims at a control that maintains a spatial orthogonality between the flux vector ψ_s and the armature current vector i_a as shown in the space phasor diagram of figure 11. The reference speed and reference flux commands are given to the vector controller that generates the reference analog voltages for the cycloconverter (through the current controller) and the field converter. The stator flux is estimated by a closed-loop reduced order observer. Referring to figure 10, C_1 is the speed controller that generates the torque command which is divided by the stator flux to generate the torque command of current i_{sT} . The magnetisation current along the flux axis (i'_m) is obtained from a flux controller C_2 . The transient stator flux component of current i_{sm} is obtained from the relationship, $i^{*'}_{sm} = i'_m - i_{fd} \cos \delta$, which decays down to zero in the steady state. The steady state displacement angle is decided by the displacement angle controller. The set value of the field current is obtained from the relation, $i^*_{fd} = i'_M / \cos \delta$. C_3 is the field current controller that generates the control voltage for triggering the field converter. The vector rotator (VR) transforms the vector from two axes flux – torque reference frame to *abc* stationary reference frame. The observer and the control circuit design aspects together

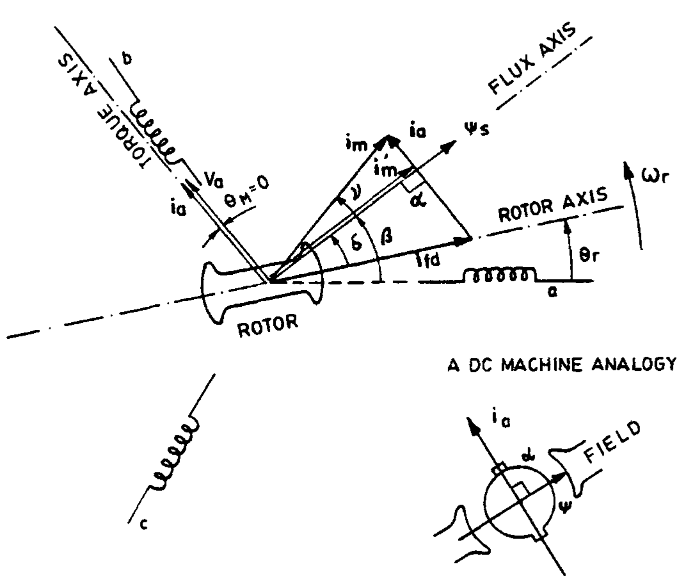


Figure 11. Space phasor diagram for vector controlled synchronous motor.

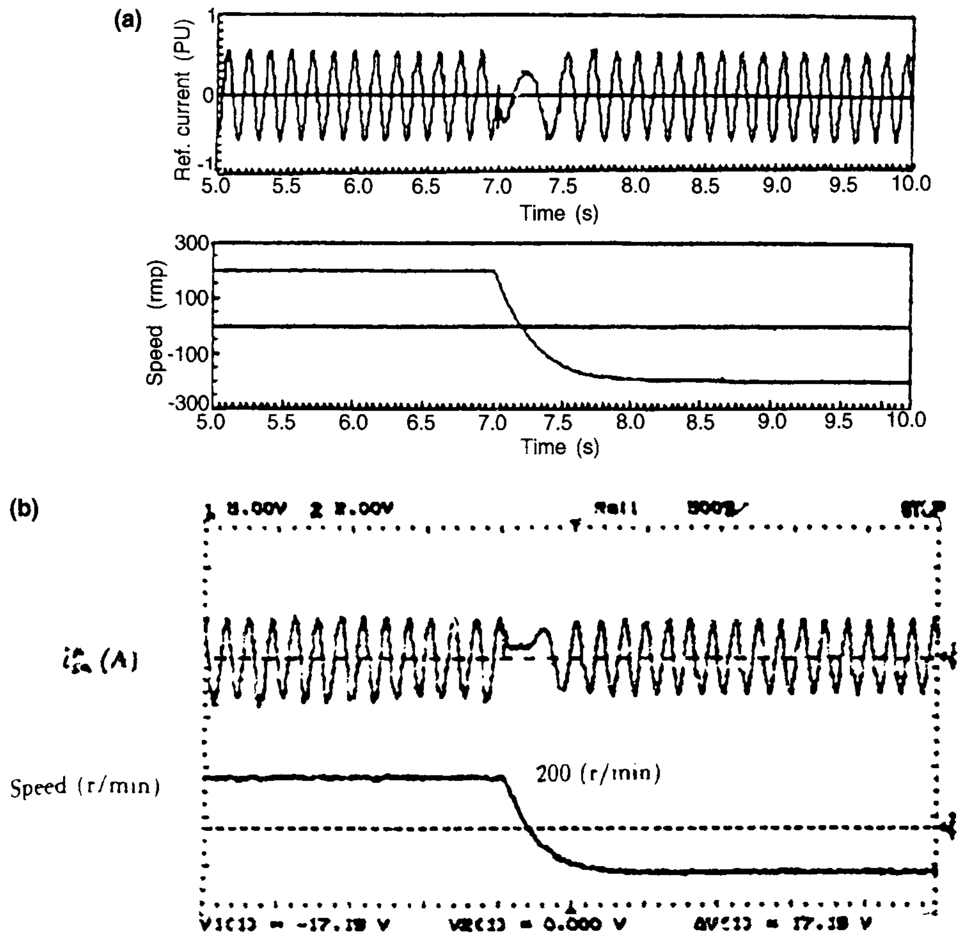


Figure 12. Speed and current responses to a step reversal (200 to −200 rpm) of a vector controlled cycloconverter fed synchronous motor drive: (**a**) simulation results, (**b**) experimental results.

with the PC-based implementation are detailed in Das (1996) and Das & Chattopadhyay (1997) and typical results obtained are shown in figure 12.

Vector control of a synchronous motor can be made with respect to three flux/mmf vectors, namely, the stator flux, the damper flux and the field mmf. A unified analysis of these three schemes by Das (1996) shows that the stator flux orientation results in a unity power factor which is not the case with the other schemes. A damper flux orientation scheme which is comparable to rotor flux orientation in the induction motor drive has been recently reported by Chongjium *et al* (1995) without any detailed analysis. Orientation with field mmf results in a lagging motor terminal power factor and is not expedient for high power drives.

3.2e *Saturation and damper effects:* Magnetic saturation effects of the d-axis and q-axis for a damperless, salient pole stator flux-oriented wound rotor synchronous motor drive have been studied by Brass & Mecrow (1992) by developing a saturated flux model. In a separate paper (Brass & Mecrow 1993), the effect of damper windings on field-oriented

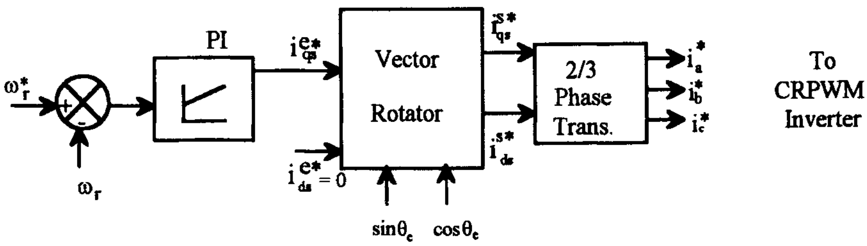


Figure 13. Vector control scheme for a permanent magnet synchronous motor.

synchronous motor has been studied and it is shown that the presence of damper windings improves the torque dynamics.

3.2f *Vector control of permanent magnet synchronous and reluctance motors:* Permanent magnet machines, both surface-mounted and interior-magnet types, are extensively used in servo drives and robotic applications, and vector control provides smooth torque operation of these motors through the entire speed range, including zero, with high power factor. The control schemes for these machines have been extensively discussed in Vas (1990) and so are not detailed here. In a permanent magnet synchronous motor, the rotor field flux ψ_f and the corresponding equivalent field current I_f can be considered as constant. For surface-mounted machines, the saliency and armature reaction is negligible. Therefore, $\psi_f = \psi_m$ and for maximum torque sensitivity with stator current $I_{ds}^{e*} = 0$, and $I_s^* = i_{qs}^e$. Figure 13 shows the vector control principle for the PMSM derived from the induction motor control diagram with the modifications (Bose 1986), $w_{sl} = 0$, $\theta_r = \theta_e$. A microprocessor-based field-oriented control scheme for a permanent magnet hysteresis synchronous motor is presented in Qian & Rahaman (1993).

Synchronous reluctance motor drives have recently received renewed attention due to the application of field oriented control to these motors (Boldea *et al* 1991; Xu *et al* 1991; Matsuo & Lipo 1993). Excellent control performance of the drive systems has been obtained though there exists a limitation in the field weakening range.

4. Conclusions

The vector control of *ac* drives in which there have been a spurt of activities, has by now gained maturity but still continues to provide interesting and challenging scope for innovations to researchers and application engineers. This paper has made an attempt to make a summary review of the activities on various aspects in this important field for control of both induction and synchronous machines till date with the informations available in the published literature. It is expected that the review will help those interested in the development of efficient and high performance drives of the future.

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