Basic Schemes for Vector Control of AC Induction Motors

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Abstract: In this paper, various strategies for controlling AC induction motors are presented. First, the basic scheme of motor control for steady state applications is discussed. Then, the Field Oriented or Vector Control approach is presented, two variations of which are discussed: Direct Vector Control and Indirect Vector Control. Finally, the complete general schemes for both types of AC motor control are analysed.

1. Introduction

Often in motion control a question arises about what type of a drive has to be used. Basically, when it is necessary to use electrical drives, there are two main options: AC or DC drives. Many factors, of both complexity and cost and performance and maintenance, are to be taken into consideration when making a choice. The importance of such a selection makes it necessary to carry out a comparative study on the control-AC motor system and its counterpart, the control-DC motor system. The present work covers the schemes and basic theory for controlling an AC motor, making at the same time comparisons with the corresponding DC motor control characteristics.

2. Control of the AC Motor

The model of an AC induction motor with its electrical (Eqs.1 and 2) and mechanical (Eqs.3 and 4) equations , is presented. The equivalent equations for a DC motor (electrical Eqs.5 and 6, and mechanical Eqs.7 and 8) are introduced in order to put both models into contrastive evaluation. By particularly comparing Eqs.3 and 7, the greater complexity of the AC motor model is apparent.

Basically, magnetic excitation $\phi_e(t)$ and armature $\phi_a(t)$ fluxes in DC motors are orthogonal and fixed-in-space vectors and, thereby, the currents generating them can be independently controlled.
The variable asignment of Eqs.1 to 8 is as follows:

$$u_s(t) = i_s(t)R_s + L_s\frac{di_s(t)}{dt} + M\frac{d[i_R(t)e^{j\varepsilon}]}{dt} \quad (1)$$

$$0 = i_R(t)R_R + L_R\frac{di_R(t)}{dt} + M\frac{d[i_s(t)e^{-j\varepsilon}]}{dt} \quad (2)$$

$$J\frac{d\omega(t)}{dt} = \frac{2}{3}M[i_s(t)] \wedge [i_R(t)e^{j\varepsilon(t)}] - m_L(t) \quad (3)$$

$$\frac{d\varepsilon(t)}{dt} = \omega(t) \quad (4)$$

$$u_a = R_a i_a + L_a\frac{di_a}{dt} + c_1\phi_e\omega \quad (5)$$

$$u_e = R_e i_e + L_e\frac{d i_e}{dt} \quad (6)$$

$$J\frac{d\omega}{dt} = m_d - m_L = c_2\phi_e i_a - m_L \quad (7)$$

$$\frac{d\varepsilon}{dt} = \omega \quad (8)$$

For the AC machine

$u_s(t)$: stator voltage	$i_s(t)$: stator current
$i_R(t)$: rotor current	R_s: stator resistance
R_R: rotor resistance	L_s: stator inductance
L_R: rotor inductance	M: mutual inductance

For the DC machine

u_a: armature voltage	i_a: armature current
u_e: field voltage	i_e: field current
ϕ_e: magnetic flux	R_a: armature resistance
L_a: armature inductance	R_e: field resistance
L_e: field inductance	

For both machines

J : inertial moment — ε : angle of the motor shaft

ω: angular speed — m_d : motor torque

m_L : load torque

On the other hand, the magnetic flux and the rotoric current $i_R(t)$ in the AC motor are strongly and mutually dependent variables, and the vectors representing them move in relation to the stator and the rotor as well. The magnetic flux and rotoric current i_R are determined in a complex way by means of the amplitude, frequency and phase of the statoric currents, two of which are independent variables, since the summation of the three statoric currents is zero on account of the insulated neutre.

Besides, an extra complexity in the AC induction motor shows up, making it radically different from the DC motor, consisting in the impossibility of measuring rotoric currents in squirrel-cage AC motors. Therefore, it may be concluded that the AC motor is quite a more complex to control machine than the DC motor, being the model for the former highly non-linear, multivariable and with coupled variables, as shown in Figure 1.

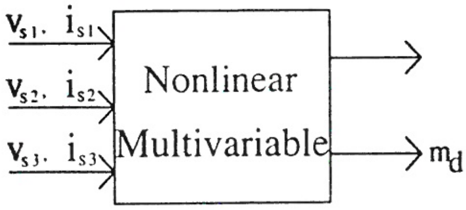


Figure 1. Model for the AC Induction Motor To Be Controlled.

3. AC Motor Control with Open Loop Flux Control

Though the AC motor model is very complex, a simplification can intervene for the particular case of feeding the motor with sine three-phase statoric input currents. By using Eq.3 for this case of sine three-phase inputs, with load torque $m_L = 0$ and $R_s = 0$, there results,

$$m_d = \frac{3}{2}\frac{1-\sigma}{\sigma}\frac{1}{L_s}\left(\frac{U_s}{\omega_1}\right)^2 \frac{2}{\frac{s}{s_p}+\frac{s_p}{s}} \qquad (9)$$

$$\frac{s}{s_p} = \frac{\sigma(\omega_1-\omega)L_R}{R_R} = \sigma\omega_2 T_R \qquad (10)$$

where U_s is the vector module of statoric voltages, $\boldsymbol{\omega_1}$ the statoric frequency, $\boldsymbol{\omega}$ the motor angular speed $\boldsymbol{\omega_2}$ the rotoric current frequency, $\boldsymbol{\sigma}$ the leakage factor and s, s_p the actual and the pull-out frequency slip respectively whose relationship is defined by Eq.10.

Eq.9 can be simplified whenever the next relationship is kept (equivalent to nominal flux)

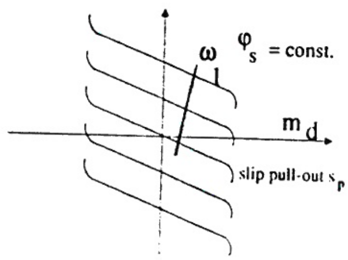
$$\frac{U_1}{\omega_1} = K = \text{cnst.} \qquad \mathbf{(11)}$$

therefore,

$$m_d = m_{po}\frac{2}{\frac{s}{s_p}+\frac{s_p}{s}} \qquad \mathbf{(12)}$$

where

$$m_{po} = \frac{3}{2}\frac{1-\sigma}{\sigma}\frac{1}{L_s} \qquad \mathbf{(13)}$$



Figuren 2.Angular Velocity vs. Electric Torque under Steady State Conditions.

Figure 2 depicts the angular velocity vs. motor torque for the motor fed with sinusoidal three-phase inputs. These characteristics resemble those equivalent ones for a DC motor. It should be noted that both characteristics are valid only under steady state conditions and, therefore, cannot be regarded for transitory state conditions.

For the above stated model, the statoric resistance is zero. Consequently, the proposed control scheme is no longer valid at low speeds because the voltage drop through R_s is no longer negligible when compared with the inductive voltage drop. Therefore, a correction factor should be introduced to account for this voltage drop, when operating at low speed.

$$U_1 = |R_s I_s + j\omega_1 K| \qquad \mathbf{(14)}$$

This concept is dealt with in Figure 3.

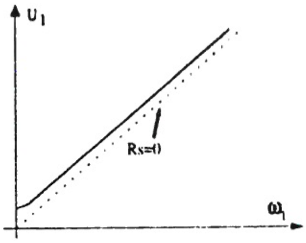


Figure 3.Normalized Statoric Voltage vs. Normalized Velocity for Constant Magnetic Flux

The deflections of Eq.11 turn to be an important factor only at low frequencies. For $\omega_1 = 0$, the necessary voltage is indicated only by the statoric resistance. In order to get free from the variations on R_s, L_s, it is more convenient to work with imposed statoric currents. This way the motor's dynamics is greatly simplified because the statoric electrical equations will be cancelled out from the mathematical model. Furthermore, a more realistic criterion is to assume sinusoidal currents instead of sinusoidal voltages, particularly if Pulse Width Modulated Inverters (PWMs) are used. The capability of controlling stator currents is thus obtained.

A further improvement is obtained by imposing a constant magnetic flux in the air -gap instead of setting it on the stator. This implies not taking into account the stator leakage inductances, that is keeping constant the magnetization current module I_m in Figure 4. The required current I_m is then,

$$I_m = \frac{j\omega_1 \sigma_R L_0 + \frac{R_R}{s}}{j\omega_1 (1+\sigma_R) L_0 + \frac{R_R}{s}} I_s \qquad (15)$$

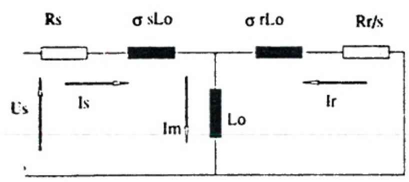


Figure 4. Equivalent Circuit for One Phase of the Motor under Steady State Conditions

or, conversely,

$$I_s = \frac{1 + j(\omega_1 - \omega)\tau_R}{1 + j(\omega_1 - \omega)\frac{\sigma_R}{1+\sigma_R}\tau_R} I_m \qquad (16)$$

where

$$\tau_R = \frac{L_R}{R_R} \qquad (17)$$

and is the motor's mechanical speed. Therefore,

$$I_s = I_m \sqrt{\frac{1 + (\omega_2 \tau_R)^2}{1 + \left(\frac{\sigma_R}{1+\sigma_R}\omega_2 \tau_R\right)^2}} \qquad (18)$$

Figure 5 depicts the characteristics of I_s in terms of ω_2 for obtaining the constant flux in the air-gap.

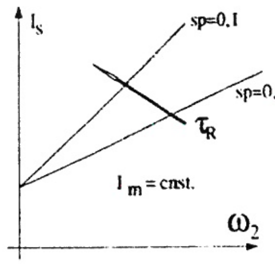


Figure 5.Normalized Stator Current vs. Normalized Velocity for Constant Flux in the Air- Gap.

The problem now resides on the dependence of I_s on τ_R, a parameter liable to drift with temperature changes, or with the magnetic saturation of the rotoric inductance. This method of keeping U_1 / ω_1 constant, is the most appropriate choice for speed control of an AC induction motor under conditions of steady-state, open-loop flux and sine three-phase inputs in the stator.

4. AC Motor Field Oriented Control with Imposed Statoric Currents

The control strategies developed in the previous section are not applicable when fast dynamic responses are needed, because they are based on a steady state model. Besides, it is extremely difficult to operate a motor at low speed and high motor torque. Thus, it is nearly impossible, for instance, to perform a positioning control procedure with such control strategies. Then, it is worthwhile working with some alternative control procedures which take into account dynamic conditions for the AC motor model other than the specific case for steady state conditions and three-phase sinusoidal input currents.

One such alternative is the Field Oriented Control, which is also known as Vector Control (**VC**). This theory assumes both the statoric and rotoric currents as imposed variables, that is, there exists a current feedback loop having a fairly good performance so as to ensure that the motor statoric currents be equal to the corresponding reference currents. Hence, the necessity for a power inverter having an appropriately high commutation frequency and a DC supply voltage high enough as to prevent inverter saturation.

Basically, VC [1][2][3] involves a set of mathematical transformations which modifies the AC motor dynamic model and turns it into a less complex one, much easier to control with a procedure analogous to that used for DC machines. The decrease in complexity is most notable in the expression for motor torque.

For this reason, the theory henceforth presented will start from the motor torque equation for the AC motor dynamic model,

$$m_d(t) = \frac{2}{3} M \,\mathrm{Im}[i_s(t)[i_R(t)e^{j\epsilon(t)}]^*] \quad (19)$$

$$m_d(t) = \frac{2}{3} M[i_s(t)] \wedge [i_R(t)e^{j\epsilon(t)}] \quad (20)$$

These equations should undergo some kind of mathematical tranformation so as to decrease their complexity and to cancel out the rotoric current term, a variable impossible to be measured in conventional squirrel-cage motors. Such transformation entails the angular shift of the reference axis. In the dynamic model described at point 2, the reference system was set on the fixed or static part of the motor, that is, the stator. The VC theory now sets a new reference system oriented on the rotoric flux direction, defined in statoric co-ordinates.

Figure 6a shows both the statoric current vector i_s and the angular position of the rotor and flux or rotoric field, all of them referred to the statoric reference axis. The current i_{mR} is assumed to be an equivalent magnetizing current capable of generating such flux.

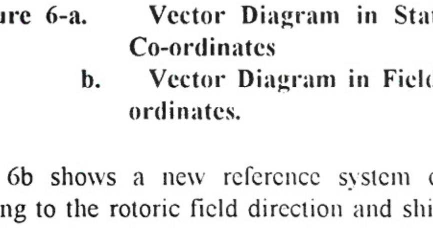


Figure 6-a. Vector Diagram in Statoric Co-ordinates
b. Vector Diagram in Field co-ordinates.

Figure 6b shows a new reference system oriented according to the rotoric field direction and shifted as an angle ρ from the former statoric axis. The statoric current i_s can be broken down into i_{sd}, the component along the rotoric field axis, and i_{sq}, the component in quadrature.

Going back to the motor torque equation, which is still expressed in statoric co-ordinates, it can be expressed in terms of i_{mR} instead of the rotoric current i_R. The resulting equation is,

$$m_d(t) = \frac{2}{3} \frac{M}{1+\sigma_R} [i_s(t)] \wedge [i_{mR}(t)] \quad (21)$$

By expressing the above equation into field co-ordinates and expliciting the vector product, it becomes,

$$m_d(t) = \frac{2}{3} \frac{M}{1+\sigma_R} [|i_{sd}(t)| \; |i_{mR}(t)| \sin 0^\circ + \\ + |i_{sp}| \; |i_{mR}| \sin 90^\circ] \quad (22)$$

and, this way, the fundamental equation for VC is reached.

$$m_d(t) = \frac{2}{3} \frac{M}{1+\sigma_R} |i_{sp}(t)| \; |i_{mR}(t) \quad (23)$$

where vectors $i_{sq}(t), i_{mR}(t)$ are orthogonal and, therefore, mutually independent. Quite apparently, however, this equation closely resembles the equivalent one for DC drive motor torque, $i_{mR}(t)$ being analogous to the main flux $\phi_e(t)$, and $i_{sq}(t)$ analogous to the DC motor armature current, i_a.

In Figure 6b it is also noticed that $i_{sd}(t)$ has the same direction as that of $i_{mR}(t)$, thus becoming a simple task to control the magnitude of the latter variable by means of $i_{sd}(t)$. Indeed, it can be inferred that these two vectors behave in like manner as the magnetic flux $\phi_e(t)$ and the corresponding excitation voltage $U_e(t)$ in the DC motor, being confirmed therefrom the common feature of a considerable time delay due to the magnetic equations in both motors. Consequently, $i_{sd}(t)$ and $i_{sq}(t)$ can be regarded as the two control actions for the new model modified by VC transformations.
The new d-q co-ordinate system, set along the direction of i_{mR}, differs from that used for synchronous drives, where the transformation is based on rotor positioning (angle $\varepsilon(t)$). In VC, the reference angular position is the rotoric flux moving an angle $\rho(t) - \varepsilon(t)$ respecting the rotor. The remaining equations of the vector field co-ordinate dynamic model can be obtained by introducing i_{mR} into the equations for the statoric co-ordinate dynamic model [1].

Next, the equations conforming the dynamic model in field co-ordinates follow,

$$\tau_R \frac{di_{mR}}{dt} + i_{mR} = i_{sd} \quad (24)$$

$$\frac{d\rho}{dt} = \omega + \frac{i_{sp}}{\tau_R i_{mR}} \quad (25)$$

$$J\frac{d\omega}{dt} = m_d(t) - m_L(t) =$$
$$= \frac{2}{3}\frac{M}{1+\sigma_R}|i_{sp}(t)||i_{mR}(t)| - m_L(t) \quad (26)$$

$$\frac{d\varepsilon}{dt} = \omega \quad (27)$$

The VC principle is an alternative to uncoupling the variables from the multivariable model representing the AC induction motor. In this model, the statoric circuit electric equations governing the voltage-current dynamic relationships have not been considered, because it was assumed that the input variables were statoric currents themselves. Therefore, it is worth recalling that the equations have been obtained in a model where the electric torque and the magnetic flux are determined by currents. Now, the relationships existing between the statoric currents and their components along the d-q field co-ordinate system should be determined, beginning with i_s which is broken down into its real, i_a, and imaginary, i_b, components on the statoric reference system, as shown in Figure 7.

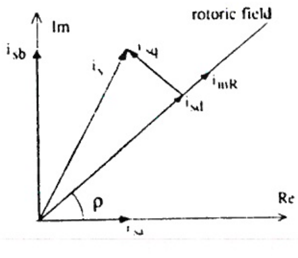


Figure 7. Statoric Current Broken Down into Components in Field Co-ordinates

Therefore,

$$i_s(t) = i_{s1}(t) + i_{s2}(t)e^{j\gamma} + i_{s3}(t)e^{j2\gamma} =$$
$$= i_{sa} + ji_{sb} \quad (28)$$

where $\gamma = 120°$. And as,

$$i_{s1}(t) + i_{s2}(t) + i_{s3}(t) = 0 \quad (29)$$

it follows that,

$$\begin{bmatrix} i_{sa}(t) \\ i_{sb}(t) \end{bmatrix} = \begin{bmatrix} \frac{3}{2} & 0 & 0 \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_{s1}(t) \\ i_{s2}(t) \\ i_{s3}(t) \end{bmatrix} \quad (30)$$

The co-ordinate transformation of the statoric current to the rotoric field reference axis is,

$$i_s(t)e^{-j\rho} = i_{sd} + ji_{sq} \quad (31)$$

which can also be expressed as

$$i_s(t)e^{-j\rho} = (i_{sa} + ji_{sb})(\cos\rho - j\,\mathrm{sen}\rho) \quad (32)$$

By equalizing second terms of eqs. (31) and (32) the expressions for $i_{sd}(t), i_{sq}(t)$ in terms of $i_s(t)$ and angle $\rho(t)$ are obtained

$$\begin{bmatrix} i_{sd}(t) \\ i_{sq}(t) \end{bmatrix} = \begin{bmatrix} \cos\rho & \mathrm{sen}\,\rho \\ \cos\rho & -\mathrm{sen}\,\rho \end{bmatrix} \begin{bmatrix} i_{sa}(t) \\ i_{sb}(t) \end{bmatrix} \quad (33)$$

It can be concluded that Eq.(30) is a trigonometric transformation of a three-component into a two-component vector. Likewise, Eq.(33) is the co-ordinate transformation from a statoric reference system into another one having as reference the vector representing the rotoric magnetic field. This new system is called field co-ordinate system.

Up to now, the AC motor models have only been presented, but they have not been used yet on the motor control problem itself. In this respect, the VC theory allows for carrying out a new control scheme based on the motor dynamic model in field co-ordinates. Such a control should perform transformations which are inverse to the above-mentioned ones. Figure 8 shows both the AC motor model expressed in field co-ordinates and the corresponding inverse transformations necessary for the controller to act upon the motor.

The field co-ordinate control acts upon a model which behaves like a DC motor and, therefore, very easy to control with mutually independent control variables, as they have already been uncoupled by VC transformations.

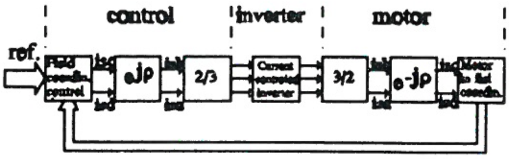


Figure 8.Simplified Scheme of Vector Control

The attainment of the angle $\rho(t)$ is a critical aspect for operating with vector control and will be further discussed.

By assuming that the angle obtained of the rotoric flux orientation $\rho(t)$ matches perfectly well the real angle of the rotoric flux in the motor, the field into rotoric co-ordinates inverse transformation can be expressed by the following equation,

$$\begin{bmatrix} i_{sa_{ref}} \\ i_{sb_{ref}} \end{bmatrix} = \begin{bmatrix} \cos\rho & -\operatorname{sen}\rho \\ \operatorname{sen}\rho & \cos\rho \end{bmatrix} \begin{bmatrix} i_{sd_{ref}} \\ i_{sq_{ref}} \end{bmatrix} \qquad (34)$$

The three-into-two current component transformation is written in Eq. 35. It should be recalled that only currents are regarded as the motor input variables, thereby greatly simplifying its model. This fact imposes a power actuator between the control action and the motor, which should behave as a very efficient controlled current supply.

For a correct operation of the field co-ordinate control, it is necessary to know in advance the rotoric flux orientation angle ρ, which can be found by two methods, namely, Direct Vector Control (DVC) and Indirect Vector Control (IVC).

The original VC proposal is based on the direct measurement of the rotoric flux [2][3] by means of special sensors like Hall-effect sensors. Although this Direct Vector Control scheme gives a correct measurement of the angle ρ, its implementation encounters serious difficulties with the mechanical installation of the sensors.

A second approach to DVC is the estimation of angle ρ through the measurement of statoric currents and angular speed, using the model of Eqs.24 and 25. This alternative is shown in Figure 9. The model of Eqs.24 and 25 is shown in the block diagram of Figure 10.

This alternative presents greater advantages over the previous method in that its flux signals are not based on flux measurements carried out on the motor itself. Therefore, they are not affected by harmonics generated by spire slots in the motor windings, as it is the case with the sensors placed inside the motor.

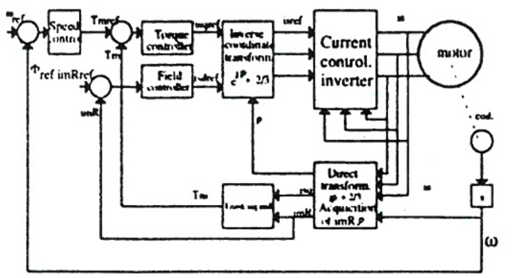


Figure 9.General Model of a Direct Vector Control (DVC)

Besides, there is no need to modify the motor's mechanical construction, and the magnetic flux acquisition is operative at very low frequencies, even at frequency zero.

The main question in obtaining data on the motor's flux is how to find the rotoric time constant τ_R which is prone to largely drifting with temperature variations. Eqs. 24 and 25 include τ_R as a coefficient parameter; hence, some sort of rotoric parameter estimation procedure is recommended. Some schemes based on the evaluation of τ_R are dealt with in the literature [4],[5], but they will not be covered here.

A simpler approach is to assume that the real statoric currents are equal to the reference currents supplied by the control system. In such an instance, the same

$$\begin{bmatrix} i_{s}1_{ref} \\ i_{s}2_{ref} \\ i_{s}3_{ref} \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & 0 \\ -\frac{1}{3} & \frac{1}{\sqrt{3}} \\ -\frac{1}{3} & -\frac{1}{\sqrt{3}} \end{bmatrix} \begin{bmatrix} i_{sa_{ref}} \\ i_{sb_{ref}} \end{bmatrix} \qquad (35)$$

model as expressed by Eqs. 24 and 25 is used for computing $\rho(t)$, including the magnitudes for the reference current i_s. This control scheme is called Indirect Vector Control (**IVC**).

IVC is based on the assumption of a current-controlled inverter inserted in the power circuit, whose characteristics enable the stator currents to be considered almost equal to the reference currents coming into this inverter. This is a great advantage, for it will only do to carry out a co-ordinate transformation.

With both vector control methods, the main objective of field oriented control is reached, which is to uncouple the currents that generate motor torque and magnetic flux. Nevertheless, IVC wins over DVC due to its simplicity for implementation, as there is no need to carry out any co-ordinate transformation for the measured currents. Figure 11 depicts the typical IVC scheme in block diagram.

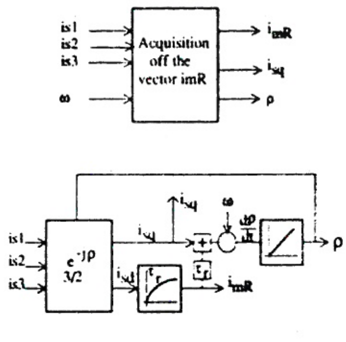


Figure 10.Block Diagram for Obtaining the Rotoric Flux.

In both VC models, i_{ds} is quasi constant and it is supplied by the magnetic flux controller in order to keep constant the magnetic flux or, otherwise, to control both velocity and position by decreasing the magnetic flux. This control method is equivalent to that used for DC motors and it is mostly used in the above-rated velocity range. The electric torque current i_{qs} is controlled in the main loop of velocity or position control, so as to obtain the required torque for reaching the desired velocity or position.

The performance limitations of this VC are largely due to its sensitivity to motor parameter inaccuracies arising when computing $\rho(t)$. Besides, the drift of the rotoric resistance severely affects the control performance. Unfortunately, the rotoric resistance changes with temperature variations in the rotor, thus offsetting the ideal situation where motor torque and magnetic flux are not coupled.

The hindrances above stated can adversely affect the performance of an AC motor vector control, and they can be overcome by using adequate methods for motor parameter identification.

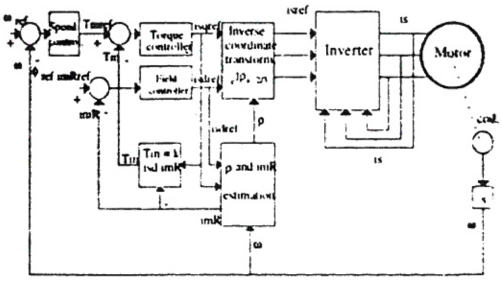


Figure 11. Block Diagram of an IVC.

5. Control of Statoric Currents

In the AC motor VC models outlined in 4, the statoric currents are considered as control variables or control actions. In order to implement the VC, a current-controlled power inverter needs be included, otherwise the AC motor dynamic model in field co-ordinates would become more complex and, therefore, more difficult to put into practice.

An alternative to statoric co-ordinate current control is the field co-ordinate current control. The same VC outline as expressed in 4. is kept but, instead of controlling statoric currents, the control is performed directly on i_{sq} and i_{sd}. Better dynamic characteristics for the control system are obtained this way. Figure12 shows the block diagram of this alternative for DVC only, as this type of current control is not practicable in IVC.

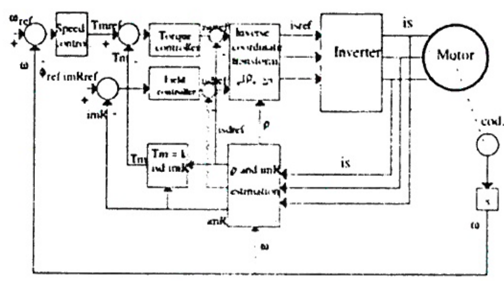


Figure 12.Block Diagram of an DVC with Current Control in Field Co-ordinates.

6. Advantages of Vector Control

The main advantages found in VC can be summarized as follows:

- VC enables to independently control both the magnetic flux and the motor torque, thus virtually transforming the behavior of an AC motor into that of a DC motor.

- By meeting the condition that an obtained angle $\rho(t)$ be precise enough, an AC motor can behave in such a way that under different load torques the motor torque vs. velocity function moves along the velocity axis, so the operating point can never reach the unstable region. Therefore, even when the motor may turn overloaded, it will stop upon reaching maximum torque.

- VC is a technique that uses the motor dynamic model, so it is adequate not only for steady state applications but also for fast dynamic purposes.

- Under steady state conditions, when DC-type variables are processed, the control is much less sensible to unavoidable motor phase shiftings.

- Due to the excellent dynamic characteristics obtained with VC, the AC motor becomes a useful alternative to positioning and fast dynamic applications, such as numeric-controlled tool-machines and robotics.

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