



## A Survey of Speed and Flux Control Structures of Squirrel-Cage Induction Motor Drives

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**Abstract:** The paper presents an overview of the adjustable speed induction motors with short-circuited rotor from the classical V-Hz open-loop- to the field-oriented closed-loop methods. Scalar-control structures are presented based on direct and indirect flux regulation versus vector-control strategies with direct and indirect field-orientation, for voltage- and current-source frequency converter fed drives. Synthesis about DC-link frequency converters, pulse modulation procedures and mechanical characteristics of the flux-controlled machine are included. Details regarding generation, computation and identification of feedforward and feedback control variables are treated. A new vector control structure is proposed for voltage-controlled drives, which combines the advantages of rotor- and stator-field-orientation procedures.

**Keywords:** Vector and scalar control, direct and indirect flux control, stator- and rotor-orientation field, direct and indirect field-orientation.

### ***Dedication:***

*In memoriam Professor Arpad Kelemen, my mentor in power electronics and electrical drives.*

## **1. Introduction**

It is estimated that more than 75% of all electrical drive applications require adjustable-speed. Nowadays the most wide-spread electric drive is based on the squirrel-cage short-circuited-rotor induction machine (SqC-IM) due to its low cost, robustness, reduced size and simple maintenance (it is in fact a brushless machine). An AC drive with rotating magnetic field, due to its nonlinear and highly interacting multivariable mathematical model, as an actuator, behaves considerably with more difficulty in a control structure compared to a compensated DC machine, where the torque is controlled directly with the

armature current, while the motor resultant flux is kept constant by means of the field winding. In case of the induction motor (IM), such a decoupling inherently can not exist. Electro-magnetically the induction motor was the electric machine (EM) most difficult to analyze. Nevertheless, the SqC-IM drives are employed in various industrial fields and dominate the power range from hundred Watts to tens of MWs. This is due to the innovation in power electronic converters (PEC), namely such as new topologies realized with high frequency power electronic devices, and the evolution of the drive-dedicated signal processing equipments, which permit the implementation of advanced control theories leading to a highly improved performance of economical industrial AC drives.

## 2. Generalities about control strategies of the cage induction motors

The area of adjustable AC drives is complex and multi-disciplinary, involving both, power- and signal-electronics, PEC circuits, microprocessors, EMs, various control procedures, system theory, measurement technique, etc. In the control of a technological process, the EM is the actuator of the mechanical load, but in the control of the EM the PEC will be the actuator. There are natural intrinsic feedback effects, as the effect of the mechanical load upon the motor (e.g. the load-torque dependence on the speed value and sense and/or on position), and the motor effect upon the PEC. Such a system in AC drives usually can not operate in feedforward control, due to its multivariable, parameter-varying and nonlinear structure with coupling effects of the state variables. Both the steady state characteristics and the dynamic behavior of the system may be analyzed only by means of a mathematical model based on the space-phasor theory [1].

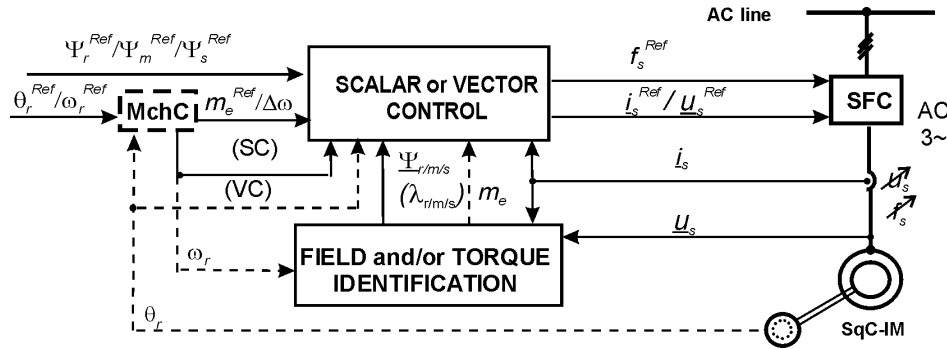


Figure 1: The general block diagram of a controlled cage-rotor induction motor drive.

In advanced AC drives a static frequency converter (block SFC in Fig. 1) controls the motor by means of two input variables; these are the amplitude and

frequency of the supply voltage. Consequently, mathematically it is possible to impose also *two reference values* in the control system of the SqC-IM. In a motor control system usually a loop is dedicated to the mechanical quantities, such as position ( $\theta_m$ ), speed ( $\omega_m$ ) and torque ( $m_e$ ), and another to the magnetic quantities, which may be one of the resultant fields, i.e. belonging to the stator  $\Psi_s$ , air-gap  $\Psi_m$  or rotor  $\Psi_r$  [2], [24].

The motor control system needs information about the technological process and about the drive. That means it requires sensing and computing of the mechanical, electrical and magnetic quantities. In general, the rotor position or speed, the stator currents and voltages are measured, while the torque and magnetic field may be only identified, using estimators or observers.

The control structure depends on the procedures of flux control, field-orientation and identification of the feedback quantities, the PEC type, including its pulse modulation method, and the character of the mechanical load.

The generation of the control quantities for the static frequency converter may be based on scalar control (SC) or vector control (VC) principle.

### 3. Comparison of the scalar and vector control procedures

A scalar AC-drive system controls only the magnitude of the prescribed quantities, without taking into account the relative position (phase shift) of the current-, voltage- and flux space phasors, which correspond to the three-phase variables of each quantity. Consequently, only the module of the controlled flux vector should be identified. In SC schemes the two control-loops work independently as is shown in *Table 1*:

*Table 1:* Scalar control (SC) strategy of induction motor drives

Imposed reference quantities	Decoupled control loops	Intermediate control variables	Decoupled control loops	Motor control variables
<b>Position / Speed / Torque</b>	→	<b>Absolute Slip</b>	→	<b>Frequency – <math>f_s</math></b>
<b>Stator / Air-gap/Rotor Flux</b>	→	<b>Stator Current</b>	→	<b>Voltage – <math>U_s</math></b>

The mechanical loop generates the actual working frequency usually by means of slip compensation, while the reactive loop provides the amplitude or r.m.s. value of the control variable for the stator current (neglecting its torque-producing active character), resulting directly or inherently the stator voltage.

The VC procedure is based on the field-orientation principle, and it needs the identification of a resultant flux as a vector. That means not only its magnitude (amplitude), but also the position (phase) angle ( $\lambda$ ), because the stator-current control variable (with its active-/torque producing and reactive-/field producing components) will be generated in field-oriented (FO) axis frame. In VC

structures, as it is presented in *Table 2*, the two intermediate control variables are different from those of a SC scheme, because they are special ones, i.e. the decoupled FO components of the stator-current space-phaser (SPh).

Usually the stator-current FO components are directly generated from the controllers in the decoupled control loops. In order to generate the SFC control variables, the two control loops will be re-coupled due to the reverse transformation of the FO components into natural (i.e. stator-oriented) ones, which needs the feedback variable  $\lambda$  (the position angle of the rotating orientation flux). In fact it realizes the self-commutation of the IM by means of the current- or voltage- space phasor with respect to the orientation flux one. As a consequence, a VC system achieves high performance considering its static stability and dynamic behavior.

*Table 2:* Vector control (VC) strategy of induction motor drives

Imposed reference quantities	Decoupled control loops	Intermediate field-oriented control variables	Re-coupled control loops	Motor control variables
<b>Position / Speed / Torque</b>	→	<b>Stator Current Active Component</b>	<b>X</b>	<b>Voltage vector</b>
<b>Stator / Air-gap/Rotor flux</b>	→	<b>Stator Current Reactive Component</b>		$\underline{u}_s$ ( $U_s$ & $\gamma_s$ )

The control variable of the supplying stator-voltage ( $\underline{u}_s = U_s e^{j\gamma_s}$ ) contains information inherently about the actual working frequency  $f_s$ , because the synchronous speed – at last in steady state – will be  $\omega_{sy} = 2\pi f_s = d\gamma_s/dt = d\lambda/dt$  and it is determined by the re-orientation angle  $\lambda$ .

In the VC structures the natural behavior of the IM is taken into account by means of the FO state-space (dynamic) model of the SqC-IM, based on the space-phaser theory, in contrast with the SC ones, where this aspect is ignored.

#### 4. Comparison of direct and indirect torque control

The direct control of the variables is made with proper controllers. Due to their natural correlation, the position, speed and torque are controlled in the same *active control loop*, as it is shown in *Fig. 2* [2], [24]. It has the output reference variables according to *Table 1* and 2:  $\Delta\omega$  the absolute slip in SC and  $i_A$  the active component of the current in VC, respectively, which are proportional to the torque, as it is shown in the next sections.

If the position control is not required, the proportional (P) position controller may be disconnected and the speed reference value will be prescribed. The torque (PI) controller may be also missing in case of regulation by the speed

loop error or if the torque is controlled indirectly (see *Figure 3*) by dividing the reference torque value with the flux amplitude (in VC) or its square value (in SC), according to the expressions given in the next sections.

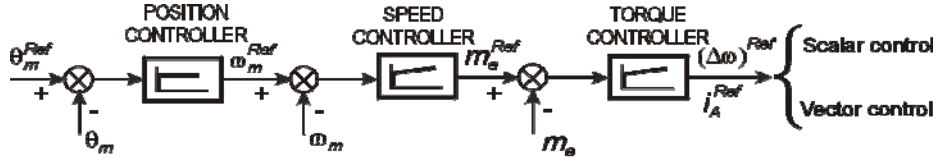


Figure 2: The complete active control loop of the electromechanical variables with direct control of the induction motor torque.

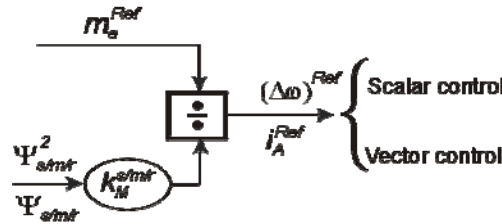


Figure 3: Indirect torque control of the cage-rotor induction motor.

In the middle of the '80s – fifteen years after Germany had developed the classical field-oriented VC – the so-called *direct torque (& flux) control* (DTC) was introduced, which needs both stator-flux and motor torque identification. It is a very simple and robust VC method for AC motors supplied from PWM-VSI with space-vector modulation (SVM) method, based on SPh theory.

## 5. Comparison of direct and indirect flux control procedures

The indirect flux control (IFC) is made without any controller and mainly it is characteristic to SC structures. The r.m.s. values of the control variables – like stator current or stator voltage – are computed based on the steady-state mathematical model of the SqC-IM (i.e. the classical time-phasor equations), the desired flux value resulting inherently. In *Fig. 4* are shown two basic procedures of indirect flux control, which are achieved by means of function generators. One of them has the input the absolute slip and the other the synchronous speed, before and after the slip compensation, respectively.

The stator flux may be kept at a quasi-constant value by the so-called control at  $U/f = \text{ct.}$  (denoted as V-Hz procedure as well). It results from the stator-voltage equation. Near the rated working point (i.e.  $U_{sN} - f_{sN}$ ) the stator resistance may be neglected, and this leads to a linear expression:

$$U_s^{Ref} = \frac{U_{sN}}{f_{sN}} f_s^{Ref} . \quad (1)$$

For lower speed range of the drive, the stator resistance  $R_s$  has to be taken into account. A possible solution is as follows:

$$U_s^{Ref} = R_s I_s + \frac{U_{sN} - R_s I_s}{f_{sN}} f_s^{Ref} , \quad (2)$$

where  $I_s$  is the measured feedback stator current [2], [24]. It is an  $R_s$ -compensation procedure with variable slope characteristic.

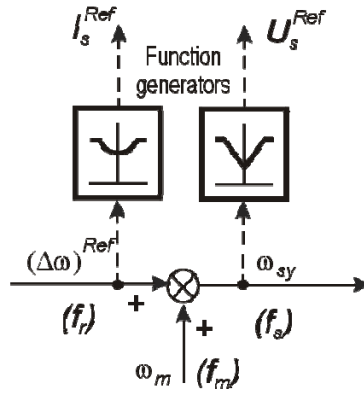


Figure 4: Indirect rotor- and stator-flux control, before and after slip-compensation in scalar control structures.

In V-Hz control, according to (1) and (2), the reference of the stator-voltage r.m.s. value (or amplitude) is generated depending on the frequency reference. This procedure may be applied to all types of AC machines, including synchronous ones, too.

The rotor flux of the SqC-IM may be also controlled indirectly, based on the rotor-voltage equation, by computing the stator-current, depending on  $\Delta\omega$  the absolute slip. Considering the rated value of the rotor-flux-based magnetizing current  $I_{mrN}$ , the reference value of the stator current may be generated according to the following expression:

$$I_s^{Ref} = I_{mrN}^{Ref} \sqrt{(\tau_r * \Delta\omega)^2 + 1} , \text{ where } I_{mrN}^{Ref} = \frac{\Psi_{rN}^{Ref}}{L_m} \quad (3)$$

and  $\tau_r = L_r / R_r$  is the rotor time constant. The angular speeds of the IM may be expressed with the corresponding frequencies as follows:

– the synchronous angular speed

$$\omega_{sy} = 2\pi f_s, \quad (4)$$

where  $f_s$  is the frequency of the stator quantities (voltages, currents, etc.);

– the absolute angular slip

$$\Delta\omega = 2\pi f_r, \quad (5)$$

where  $f_r$  is the frequency of the rotor quantities (EMFs, currents, etc.);

– the rotor electrical angular speed (usually it is measured):

$$\omega_m = 2\pi f_m, \quad (6)$$

where  $f_m$  is the mechanically rotating rotor frequency for  $z_p = 1$  pole-pairs. Consequently, the slip compensation may be written as follows:

$$\omega_{sy} = \Delta\omega + \omega_m \quad \text{or} \quad f_s = f_r + f_m, \quad (7)$$

The direct flux control (DFC) may be achieved by using a controller, which is basically of PI-type, and it needs an imposed reference value i.e. one of the resultant field values: stator-, air-gap or rotor-flux, as it is shown in *Table 1*, *Table 2* and also in *Fig. 5* ( $\Psi_{s/m/r}$ ).

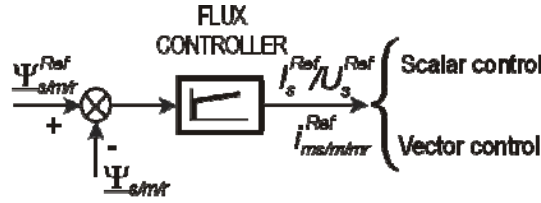


Figure 5: Direct flux control with PI controller in the reactive loop.

The flux controller in SC schemes generates the reference value of the stator voltage or current in amplitude or r.m.s. value. In VC structures, the flux controller will provide the corresponding magnetizing current, ( $i_{ms/m/r}$ ), which may be equal or not to the reactive component of the field-oriented stator-current space phasor.

## 6. Comparison of the stator- and rotor-flux control

The well-known *Kloss*'s equation gives the analytical expression of the IM static mechanical characteristics (SMC) at constant stator voltage  $U_s$  and frequency  $f_s$ . If the pull-out critical torque is kept at constant value by adjusting the value of  $U_s$  according to the actual value of  $f_s$ , the SMCs have different

shapes. In Fig. 6 two  $U_s = \text{ct}$  characteristics are represented – torque versus the absolute slip  $\Delta\Omega$  (measured in electrical rad/s) – for two supplying frequencies, i.e.  $f_{sN}$  at  $U_{sN}$  (both rated values) and  $f_s = 0$  at  $U_{s0}$ , which provide the same breakdown torque. The zero value of the rotor speed corresponds to  $\Delta\Omega = 314$  rad/s. For different frequencies at  $U_s = \text{ct}$ , the speed-torque SMCs – due to the different feature of the slip curves – are not parallel, but at constant resultant flux they will become parallel [3].

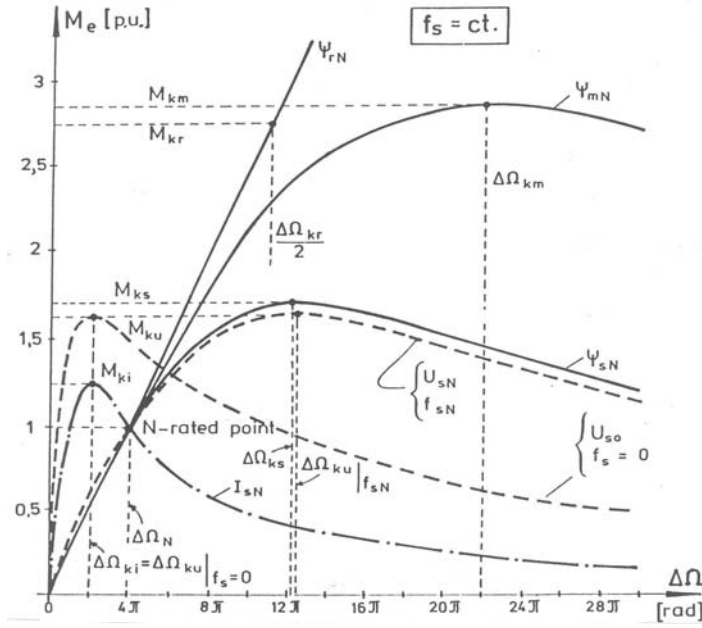


Figure 6: Stator-voltage, stator- and rotor-flux-controlled mechanical characteristics: the electromagnetic torque versus the absolute slip.

Fig. 6 also presents the mechanical characteristics for  $I_s = I_{sN}$ ,  $\Psi_s = \Psi_{sN}$ ,  $\Psi_m = \Psi_{mN}$  and  $\Psi_r = \Psi_{rN}$ , which are useful in FO structures. In a VC scheme the magnitude (module) of the orientation flux vector is usually controlled directly.

#### A. Stator-flux controlled (SFC) characteristics

For constant stator flux it results a simplified Kloss's type expression that is no more depending on the frequency, as follows [3]:

$$M_e = 2M_{k_s} \left( \frac{\Delta\Omega_{k_s}}{\Delta\Omega} + \frac{\Delta\Omega}{\Delta\Omega_{k_s}} \right)^{-1}, \text{ where } M_{k_s} = k_M \frac{\Psi_s^2}{2L_m} \cdot \frac{1-\sigma}{\sigma(1+\sigma_s)}; \quad \Delta\Omega_{k_s} = \frac{1}{\sigma\tau_r} \quad (8)$$



are the critical (pull-out) torque and slip,  $\sigma_s = L_{\sigma s} / L_m$  is the stator leakage coefficient and  $\sigma$  is the resultant one. The self-cyclic inductance  $L_m$  corresponds also to the three-phase mutual magnetic effect between the stator and rotor, and it gives the useful resultant field in the air-gap:

$$\underline{\Psi}_m = L_m \underline{i}_m, \text{ where } \underline{i}_m = \underline{i}_s + \underline{i}_r \quad (9)$$

is the conventional magnetizing current.

The torque-slip SMCs at  $\Psi_s = \text{ct}$  from Fig. 6 are valid for any stator frequency. In spite of being a combination of a linear- and a hyperbolic shape, it leads to parallel speed-torque characteristics for different stator frequencies, excepting flux-weakening region [3], as is shown in Fig. 7.

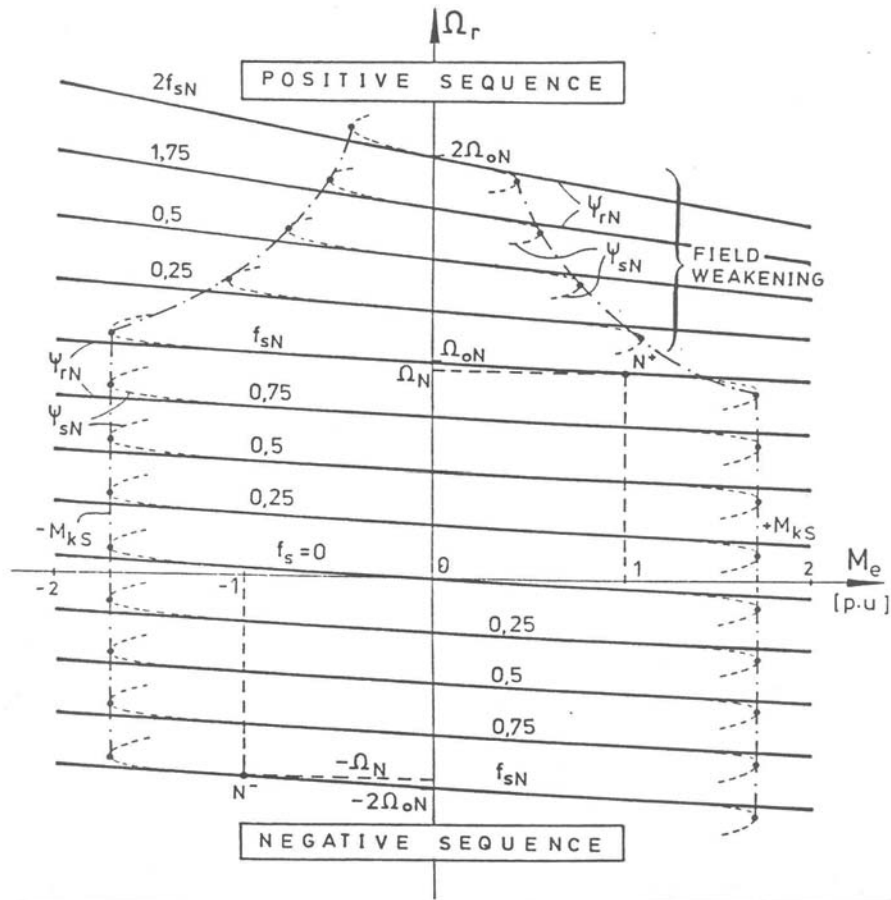


Figure 7: Stator- and rotor-flux-controlled mechanical characteristics: the angular speed versus the electromagnetic torque.

### B. Rotor-flux controlled (RFC) characteristics

At constant rotor-flux, the SMC becomes linear without any hyperbolic effect, as in case of a compensated separately excited DC machine. The torque-slip characteristic is given by the following expression:

$$M_e = 2M_{k_r} \frac{\Delta\Omega}{\Delta\Omega_{k_r}} = \frac{k_M \Psi_r^2}{R_r} \Delta\Omega \quad (10)$$

and they are represented in *Fig. 7*, too.

Due to the linearity of the SMCs, the RFC ensures more stability in behavior of the IM with respect to SFC-ed drives, where the SMCs present pull-out critical torque, due to the so called “*Kloss*” feature given by a hyperbolic shape.

The torque coefficient in (8) and (10) is  $k_M = z_p \cdot 3/2$ , if the flux is corresponding to its peak value, or  $k_M = 3 z_p$ , if the flux is expressed by the r.m.s. value of  $\Psi_s$  or  $\Psi_r$ , respectively.

## 7. Comparison of the rotor- and stator-field orientation

The FO principle was initially proposed by *Blaschke* in 1971 [4], and it referred to the decoupled control of the mechanical and magnetic phenomena of the short-circuited rotor IM by means of the stator-current rotor-field-oriented components. In fact, field-orientation means change of variables corresponding to phase- (3/2) and coordinate- (complex plane) transformations of the control and feedback variables in a VC structure [6].

### A. Rotor-field orientation (RFO)

The classical RFO is usually applied for SqC-IM drives. That means that the direct axis of the complex plane, denoted with  $d\lambda_r$ , is oriented in the direction of the resultant rotor-flux  $\underline{\Psi}_r$ , as it is shown in *Fig. 8*.

As a consequence, the flux components result according to (11.1) and (12.1) from *Table 3*.

In case of the SqC-IM ( $u_r = 0$ ), if  $\Psi_r$  may be considered at a constant value (that means steady-state or  $\Psi_r$  is a controlled variable), the rotor-current  $\underline{i}_r$  and rotor-flux  $\underline{\Psi}_r$  space phasors are perpendicular one to other. This property led to the idea of the original FO principle based on the rotor-flux-oriented axis frame, in which the stator-current space phasor may be split into two components, as in (13.1), where the RFO components of the stator-current SPh result according to (14.1), (15.1) and (16.1) in *Table 3*. Consequently, the rotor-flux controller may generate directly the field-producing (reactive) component ( $i_{sd\lambda_r}$ ) in a control

structure, because this component is equal to the rotor-flux-based magnetizing current ( $i_{mr}$ ); the speed or torque controller will generate the torque producing (active) quadrature component ( $i_{sq\lambda_r}$ ) of the stator current.

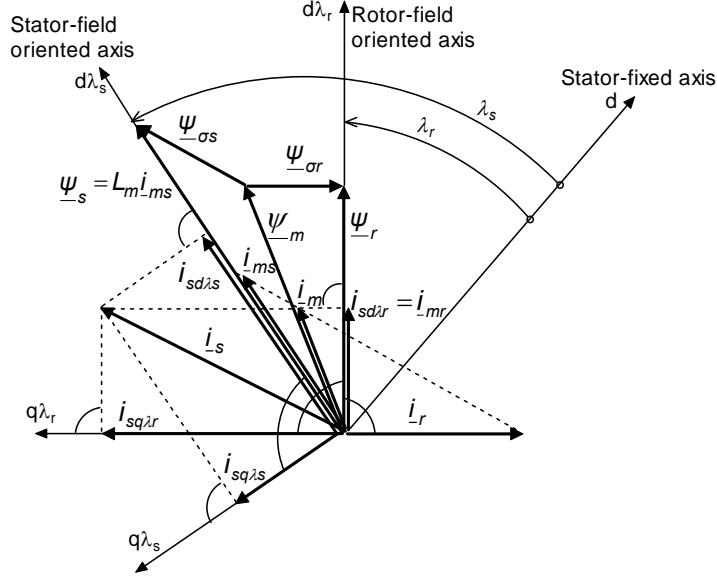


Figure 8: Phasor diagram of the magnetizing currents, fluxes and stator-current field-orientated components.

If the frequency converter is controlled in current, there is no need for model-based computation of the control variables, because they are generated directly by the controllers. If the IM is controlled in voltage, the computation of the stator-voltage components based on the RFO model is highly complex and motor parameter dependent [1], [7], [8], [9], [10].

### B. Stator-field orientation (SFO)

SFO means that the direct axis (denoted  $d\lambda_s$ ) of the coordinate frame is oriented in the direction of the resultant stator flux vector  $\underline{\psi}_s$  (see Fig. 8), therefore its components are according to (11.2) and (12.2) in Table 3.

In the stator-field-oriented (SFO-ed) axes frame the stator-current SPh components from (13.2) can be expressed using (14.2) and (15.2).

Comparing the SFO-ed components with RFO-ed ones, it must be remarked that although the active component in both cases is proportional to the electromagnetic torque, according to (17.1) and (17.2), the reactive one in SFO

is no more equal to the stator-flux-based magnetizing current ( $i_{ms}$ ), according to (16.2), as Fig. 8 also shows.

Table 3: Comparison of rotor- and stator-field orientation

Rotor-Field-Orientation (RFO)	Stator-Field-Orientation (SFO)
$\Psi_{rd\lambda r} = \underline{\Psi}_r =  \underline{\Psi}_r  = \Psi_r \quad (11.1)$ $\Psi_{rq\lambda r} = 0 \quad (12.1)$ <u>Rotor-field-oriented</u> <u>stator-current components:</u> $\underline{i}_{s\lambda r} = i_{sd\lambda r} + \underline{j} i_{sq\lambda r} \quad (13.1)$ where $i_{sd\lambda r} = i_{mr} = \Psi_r / L_m \quad (14.1)$ $i_{sq\lambda r} = m_e / K_{Mr} \Psi_r = -(1+\sigma_r) i_r \quad (15.1)$ <u>Rotor-flux-based</u> <u>magnetizing current:</u> $i_{mr} = \underline{\Psi}_r / L_m = i_{sd\lambda r} \quad (16.1)$ <u>Electromagnetic torque:</u> $m_e = K_{Mr} \Psi_r i_{sq\lambda r} = -K_M \Psi_r i_r \quad (17.1)$ where the torque coefficient is $K_{Mr} = K_M / (1+\sigma_r) \quad (18.1)$ <u>Synchronous angular speed:</u> $\omega_{\lambda r} = d\lambda_r / dt \quad (19.1)$ <u>Orientation-field angle:</u> $\lambda_r = \int \omega_{\lambda r} dt \quad (20.1)$	$\Psi_{sd\lambda s} = \underline{\Psi}_s =  \underline{\Psi}_s  = \Psi_s \quad (11.2)$ $\Psi_{sq\lambda s} = 0 \quad (12.2)$ <u>Stator-field-oriented</u> <u>stator-current components:</u> $\underline{i}_{s\lambda s} = i_{sd\lambda s} + \underline{j} i_{sq\lambda s} \quad (13.2)$ where $i_{sd\lambda s} \neq i_{ms} = \Psi_s / L_m \quad (14.2)$ $i_{sq\lambda s} = m_e / K_M \Psi_s \quad (15.2)$ <u>Stator-flux-based</u> <u>magnetizing current:</u> $i_{ms} = \underline{\Psi}_s / L_m \neq i_{sd\lambda s} \quad (16.2)$ <u>Electromagnetic torque:</u> $m_e = K_M \Psi_s i_{sq\lambda s} \quad (17.2)$ where the torque coefficient is $K_M = (3/2) z_p \quad (18.2)$ <u>Synchronous angular speed:</u> $\omega_{\lambda s} = d\lambda_s / dt \quad (19.2)$ <u>Orientation-field angle:</u> $\lambda_s = \int \omega_{\lambda s} dt \quad (20.2)$

On the other hand, in SFO schemes the stator-voltage equation provides a more simple computation of the voltage control variables for a voltage-source inverter (VSI) compared to RFO, because in SFO axis frame, the stator-flux SPh has only one component (the direct one), which is equal to its module [1], [7], [11]:

$$u_{sd\lambda s} = R_s i_{sd\lambda s} + e_{sd\lambda s} \text{ and } u_{sq\lambda s} = R_s i_{sq\lambda s} + e_{sq\lambda s}, \quad (21)$$

where EMFs are

$$e_{sd\lambda s} = d\Psi_s/dt \text{ and } e_{sq\lambda s} = \omega_{\lambda s} \Psi_s. \quad (22)$$

The direct component ( $e_{sd\lambda s}$ ) is the self-induced EMF, which becomes zero in steady state. This is due to the variation in magnitude of the  $\Psi_s$ . The quadrature component ( $e_{sq\lambda s}$ ) is generated by the rotation of the stator field with the synchronous speed  $\omega_{\lambda s}$  – given by (19.1) –, and it can be computed from the SPh components of the identified orientation field.

For voltage-PWM-VSI-fed drives – due to a simpler voltage model –, SFO is recommended [1], [7], [12]. The computation of the control variables can be made based on expressions (21) and (22). These expressions are affected only by the stator resistance  $R_s$ , which may be identified online as well.

SFO was extended also to the synchronous motor drives [5].

## 8. Comparison of stator and rotor field identification

Because the initially proposed direct flux sensing (see [13]) is no more recommended, nowadays the indirect flux sensing is applied almost exclusively, which is based on the computation of the orientation field from other measured variables. There are two basic field-identification procedures of the field: the so called I- $\Omega$  (stator-current & rotor-speed) method for rotor-flux identification and the integration of the stator-voltage equation for stator-flux computation.

### A. Stator-flux identification (SFI)

In the ‘70s and ‘80s this flux identification method could be applied only for AC drives supplied from a current-source inverter (CSI), which operates with full-wave currents and quasi-sine-wave terminal voltages, determined by the freely induced rotating EMFs [1], [14], [15], [23], [25]. However, in the last two decades it became a possible method for PWM-inverter-fed drives as well, which are operating with relatively high sampling frequency. Nowadays this procedure seems to be the simplest one for the calculation of the resultant stator flux.

This flux identification procedure is based on the stator-voltage model, written with natural two-phase components in the stator-fixed axis frame. First, the stator EMFs are computed according to equations:

$$d\Psi_{sd}/dt = e_{sd} = u_{sd} - R_s i_{sd} \text{ and } d\Psi_{sq}/dt = e_{sq} = u_{sq} - R_s i_{sq}, \quad (23)$$

and then it is followed by the direct integration of them, obtaining the flux components:

$$\Psi_{sd} = \int e_{sd} dt \text{ and } \Psi_{sq} = \int e_{sq} dt. \quad (24)$$

The inputs of the EMF computation block are the two-phase feedback variables of the measured stator-currents and the identified stator voltages computed from the measured DC-link voltage at the input of the inverter and the PWM logic signals generated by the inverter control block.

Today it seems that this method is the most preferable for field identification, due to the fact that it is not affected by the motor parameters, excepting  $R_s$ . If it is necessary, the stator resistance may be measured on-line. The applicability of this flux identification depends first of all on the quality of the integration procedure [16].

### B. Rotor-flux identification (RFI)

Still in the '80s, the rotor-model-based I- $\Omega$  flux identification procedure was preferable for IM drives supplied from PWM-inverters. It was introduced by Hasse in 1969 [7]. According to this procedure, there are two possibilities to perform RFI: either with natural (stator-fixed) stator-current components or with RFO ones. The latter procedure needs slip compensation [1], [8], [9]. Both I- $\Omega$  methods are strongly affected by the rotor parameters.

Nowadays it is preferable the rotor-flux computation by compensation of the identified stator-flux, using the expressions of the leakage fluxes depending on the measured stator currents. The unknown rotor current from the expression of the rotor leakage flux is eliminated based on the magnetizing current equation (9). The compensation is made without any cross effect between the d-q components, which in synthesized form yields to:

$$\Psi_{rd/q} = (1 + \sigma_r) \Psi_{sd/q} - (\sigma_s + \sigma_s \sigma_r + \sigma_r) L_m i_{sd/q}, \quad (25)$$

where  $\sigma_r = L_{\sigma r} / L_m$  is the rotor leakage coefficient. The coefficient of  $L_m$  is equal to  $(1 - \sigma)^{-1}$ , where  $\sigma$  is the resultant leakage coefficient. The stator-flux components are obtained based on the direct integration of the stator-voltage equation according to the procedure, which was presented in the previous *A* subheading.

## 9. Comparison of direct (DFO) and indirect (IFO) field orientation

In VC structures the recoupling of the active and reactive control loops is made by means of a reverse coordinate transformation (CooT), which calculates the natural two-phase components of the stator current from the input field-oriented ones. This needs as input variable also the angle of the orientation-field ( $\lambda_r$  or  $\lambda_s$ , respectively, according to *Fig. 8* and *Table 3*). The identification procedure of the orientation angle determines the field-orientation character, which may be direct or indirect.

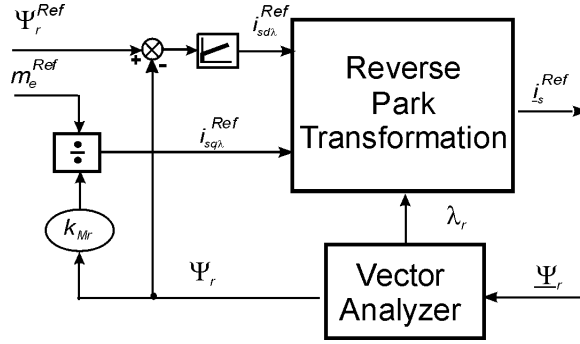


Figure 9: Direct field-orientation (DFO) realized with field-orientation angle computed in a vector analyzer.

#### A. Direct field-orientation (DFO) procedure

Fig. 9 shows the simplest recoupling of the rotor-field-oriented IM supplied from a current-controlled PEC. It needs the three-phase stator-currents as references (represented symbolically with the space-phasor  $\underline{i}_s^{Ref}$ ), which are computed in the reverse Park transformation (combined from a coordinate- and phase transformation) block.

The orientation-field angle is identified in a vector analyzer (VA), which has as inputs the stator-fixed / stator-oriented two-phase coordinates of the rotor flux (in Fig. 9 it is represented symbolically by the space-phasor  $\underline{\Psi}_r$ ). This is the direct field-orientation (DFO) procedure, where the orientation-field angle is computed based on the stator-fixed axis frame.

Flux identification procedures based on the stator-oriented axis frame lead to DFO, i.e. the direct integration of the stator-voltage equation and the I-Ω procedure calculated with the rotor-voltage equation written with stator-oriented coordinates.

#### B. Indirect field-orientation (IFO) procedure

Slip compensation is used not only in SC (see Fig.4), but also in VC structures. The indirect field-orientation (IFO) procedure means, that the field-orientation angle is computed by integration of the synchronous speed, which is usually obtained by slip compensation, as in (7) [1], [7], [8], [9]:

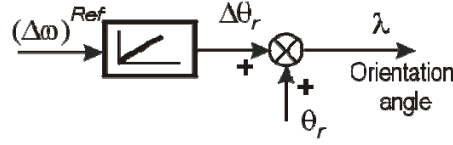
$$\omega_{\lambda_r} = d\lambda_r/dt = \Delta\omega + \omega_m, \quad (26)$$

The absolute slip of the IM with short-circuited rotor windings ( $\underline{u}_r = 0$ ) is computed from the rotor-field-oriented voltage equation for  $d\Psi_r/dt = 0$  (i.e. steady state or controlled rotor flux), as follows [1], [8]:

$$\Delta\omega = \tau_r^{-1} i_{sq\lambda r} / i_{sd\lambda r} \quad (27)$$

The I- $\Omega$  flux identification procedure based on the rotor-voltage equation written with RFO components leads to IFO, where the current components ( $i_{sq\lambda r}$  and  $i_{sd\lambda r}$ ) are computed from the measured stator currents on the feedback side.

In rotor- or stator-flux-oriented VC structures for IFO the orientation angle ( $\lambda_r$  or  $\lambda_s$ ) may be also calculated after the integration of the absolute slip, as is presented in *Fig. 10*.

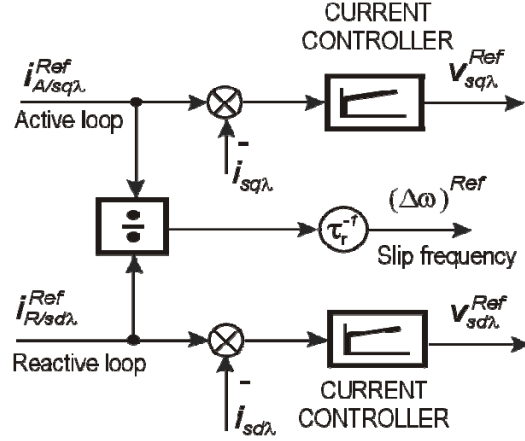


*Figure 10:* Absolute slip computation in the control loop for indirect field-orientation (IFO).

First it is obtained the slip angle  $\Delta\theta_r$ , to that is added the measured or estimated rotor position  $\theta_r$ , then achieving the orientation angle, as follows:

$$\lambda = \Delta\theta + \theta_r. \quad (28)$$

In VC systems, the field-oriented current components generated in the active and reactive control loops may also serve for the computation of the absolute slip according to expression (27), as it is shown in *Fig. 11*.



*Figure 11:* Absolute slip computation in the control loop for indirect field-orientation (IFO).



In case of a voltage controlled PEC (like VSI with feed-forward voltage-PWM), current controllers are recommended, which control the field-oriented stator-current components before the computation of the stator-voltage control variables. These will provide the re-coupling of the two control loops by means of a coordinate transformation block.

The synchronous speed  $\omega_{sr}$  is usually needed as input variable in the computation block of the stator-voltage references. It may be also generated from the absolute slip (computed in *Fig. 11*) by slip compensation, according to (26).

## 10. Computation of the PEC actuator control variables

The computation of the actuator control variables depends on the PEC type and, above all, its pulse modulation procedure, which can be one of two fundamental ones: pulse amplitude modulation (PAM) and pulse-width modulation (PWM). In *Table 4* the control variables of the DC-link frequency converters are given, which are depending on the inverter (voltage- or current-source) type and its pulse modulation method.

*Table 4:* Inverter types, pulse modulation procedures and control variables of the DC-link frequency converters

Converter control type	VECTOR CONTROL				SCALAR CONTROL
Inverter type	MOS-FET/BT/IGBT-VSI			GTO-CSI	Thy-CSI
Converter output	Voltage-source character		Current-source character		
Pulse modulation method	Open-loop feedforward voltage-PWM		Closed-loop current-PWM	PAM	PAM
Inverter control procedure	Carrier-wave modulation	Space-vector modulation SVM	Bang-bang current control	DC-link current control	
Converter control variables	Instantaneous three-phase voltages: $u_{a,b,c}$	Voltage-amplitude $U$ and $\gamma$ phase angle	Instantaneous three-phase currents: $i_{a,b,c}$	Current-amplitude $I$ and $\varepsilon$ phase-angle	Current: amplitude $I$ and $f$ frequency

In VC structures they can be computed in four ways, according to the current- or voltage space-phaser expression with polar- or with three-phase coordinates (which inherently keep the vector character) as follows:

where  $i_{a,b,c}$  and  $u_{a,b,c}$  are the instantaneous values of the three-phase currents and voltages, respectively. Angles  $\varepsilon$  and  $\gamma$  are the electrical phase positions of the respective SPhs, which inherently contain “information” about the imposed motor supply frequency, because the inverter output frequency is equal to the derivative of these position angles. The space-phasor coefficient is usually  $k_{p\bar{h}}=2/3$ . In this case the module of the space phasor will be equal to the amplitude (peak value) of the sine-wave phase variables [1], [8].

## 11. Scalar control structures

Figure 12: Synthesis of scalar control structures with DFC and IFC of the SqC-IM controlled in current or voltage.

The simplest VC structure of the SqC-IM presented in *Fig. 14* is achieved by current controlled static frequency converter, rotor-field orientation (RFO) and rotor flux control (RFC). In comparison with structure from *Fig. 13*, it has in addition flux compensation of the identified stator flux, according to (25).

The reverse Park transformation is in fact a sine-wave generator with vectorial character. This structure is not affected by the motor parameters (excepting field identification and controller tuning). Furthermore, such a control system presents the best performance compared with the schemes with stator-field orientation (SFO), stator-flux control (SFC) drives and/or IM supplied from voltage-mode controlled inverter [1], [2], [11], [17].

Some motor-control-oriented digital signal processing (DSP) equipments present on the market do not dispose of implementation possibility of the current-feedback PWM, suitable for current-controlled VSIs, only of possibility of the voltage-feedforward ones, like carrier-wave and SVM. That means the IM can be supplied only by a voltage-source inverter (VSI) with voltage-control [26], [27].

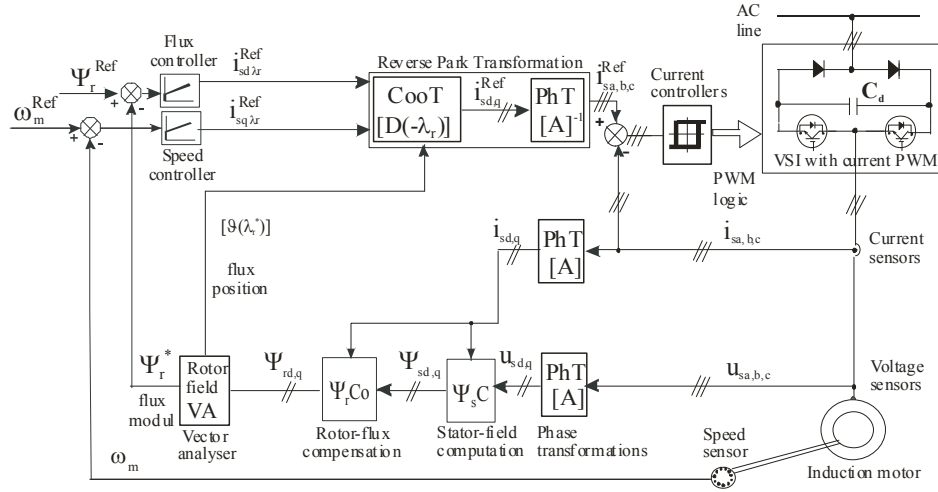


Figure 14: The simplest vector control structure with rotor-field-orientation of the short-circuited IM drive supplied from a current-feedback PWM controlled VSI.

In RFO schemes the computation of the voltage control variables is sophisticated and affected by the motor parameters such as rotor resistance ( $R_r$ ), rotor time constant  $\tau_r$ , leakage coefficients and others. Consequently, the drive control performance may be lightly damaged. This problem is usually solved by renouncing the RFC and applying SFO, which leads to a much simpler stator-voltage computation, dependent only on the stator resistance ( $R_s$ ).

Fig. 15 presents the simplest VC structure for voltage-controlled IM with stator-field-orientation, which is less affected by motor parameter than the RFO one. The stator-flux-based magnetizing current  $i_{ms}$  may be also generated at the output of the flux controller, as Table 3 shows in Section 7. This structure has a somewhat sluggish response to speed reversal, torque command and



$[o(\lambda_s)]$  and  $[o(\lambda_r)]$ , resulting from the VAs of the stator- and rotor-fluxes. The trigonometric functions required for the CooT blocks are symbolized with an “oscillatory” matrix containing two elements:  $[o(\lambda)] = [\cos(\lambda), \sin(\lambda)]^T$  [1].

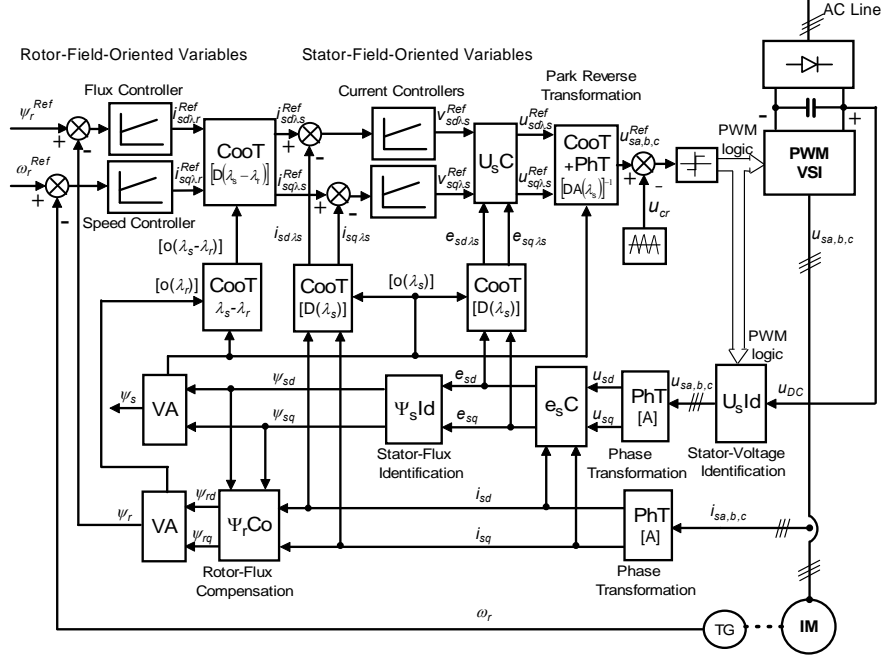


Figure 16: Vector control structure with dual-field-orientation of the short-circuited IM drive supplied from a voltage-feedforward PWM controlled VSI.

The stator-voltage control variables are computed in the  $U_sC$  block based on equations (9), where the input EMFs result from the feedback side and the *Ohm*'s law voltage drops are generated by the controllers of the SFO-ed current components. This structure eliminates the influence of the rotor parameters.

### 13. Conclusion

The RFC-ed IM with RFO-ed structure behaves similarly to a DC machine, both from point of view of dynamics and stability, due to the linear mechanical characteristics. The best control scheme seems to be a RFO with RFC and current-controlled converter as actuator. Compared to other structures, its dynamic response is superior, the computation requirements are reduced, and it is less dependent on the motor parameters. But the implementation of the current-feedback PWM presents difficulties.

Voltage-controlled VSI-fed drives (usually with SFC, either SC or VC structure), generally can not ensure the same performance, neither regarding stability and torque ripple, nor dynamics compared to RFC achieved by current-controlled VSI. This is due to the natural behavior of the IM, considering the magnetizing and torque producing phenomena.

The RFO with RFC for voltage-controlled converter-fed drives requires the highest computational capacity of the DSP, and – in addition –, the quality of the operation may suffer from the sensitivity to motor parameters, especially the coefficients of leakage and rotor time constant.

The SFO with SFC, especially used for voltage controlled converter-fed drives, is less computationally demanding and more robust, but the reaction to torque commands in low-inertia drives can lead to stability problems.

The DFO combines the advantages of the two types of field-orientation procedures for voltage-controlled IM drives, on the one hand of the RFC and RFO and on the other hand of the SFO. This combination ensures reduced computational demand, increased stability, a good dynamic and robustness, avoiding the influence of the rotor parameters.

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