Vector Control of the Cage Induction Motor with Dual Field Orientation

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Abstract: A new vector control structure is proposed for the short-circuited induction motor fed by voltage-source inverter with voltage-feedforward (carrier-wave or space-vector) pulse-width modulation, which combines the advantages of two types of field-oriented procedure: rotor-field orientation and rotor-flux control for generation the stator-current control variables and stator-field orientation for computation of the stator-voltage command variables for the voltage controlled inverter.

Keywords: rotor-field orientation, stator-field orientation, flux identification, vector control, space-vector modulation, voltage-source inverter

1 Introduction

The simplest vector control (VC) structure for induction motor (IM) drives is achieved by current controlled static frequency converter (like VSI with current feedback PWM), rotor-field orientation (RFO) and rotor flux control (RFC). It is not affected by the motor parameters (excepting field identification and controller tuning). Furthermore, such a control system presents the best performances with respect to schemes with stator-field orientation (SFO), stator-flux control (SFC) and/or the proper voltage control of the IM drive [1], [2], [3], [5], [6].

Some motor-control-oriented digital signal processing (DSP) equipments present on the market don't dispose over implementation possibility of the current-feedback PWM, suitable for current-controlled VSIs, only the voltage-feedforward ones, like carrier-wave (CWM) and space-vector modulation (SVM). That means the IM may be supplied by a proper voltage-source converter with voltage-control PWM. In RFO schemes the computation of the voltage control variables is sophisticated and affected by the motor parameters like as rotor resistance (R_r), rotor time constant $\boldsymbol{\tau}_r$, leakage coefficients and others. Consequently, the drive control performance may be lightly damaged. Usually this problem is solved by renouncing on the RFO and RFC and applying SFO with SFC.

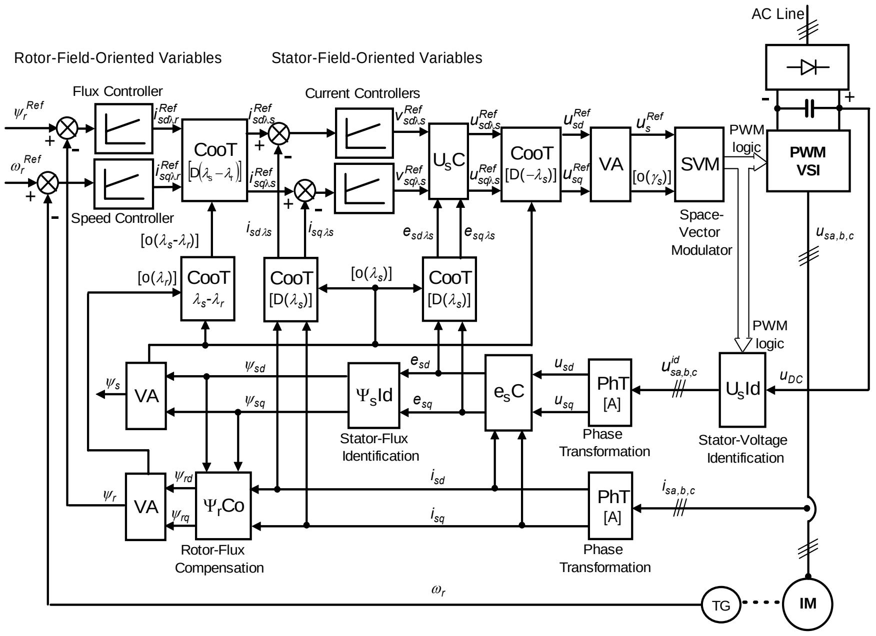

Figure 1

Vector control structure of the cage induction motor drive with dual field orientation: RFO-ed variables at the decoupled control side and SFO at the re-coupling of the two control loops

SFO leads to a much simpler stator-voltage computation and dependent only on the stator resistance (R_s). For the above presented reasons in Figure 1 a new vector control structure is proposed with dual field orientation (FO), i.e. there are applied both, SFO and RFO, too.

2 Comparison of Rotor- and Stator-Field Orientation

In classical field-orientation of the IM drives proposed initially in 1971 by *Blaschke* [7], usually the rotor flux (Ψ_r) is controlled and the stator-current space-phasor (SPh) is oriented according to the resultant rotor field.

2.1 Rotor-Field Orientation (RFO) Procedure

RFO means that the coordinate frame direct axis (i.e. the real axis of the complex plane), denoted with $d\lambda_r$, is oriented in the direction of the resultant rotor-flux vector $\underline{\Psi}_r$, as is shown in Figure 2 and the flux components are [1]:

$$\Psi_{rd\lambda r} = \Psi_r = |\underline{\boldsymbol{\Psi}}_r| \quad \text{and} \quad \Psi_{rq\lambda r} = 0 \tag{1}$$

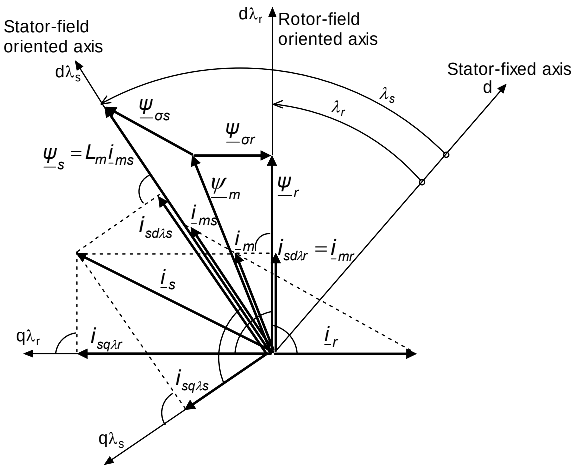

Figure 2
Space-phasor diagram
of the magnetizing currents,
fluxes and the field-oriented
components of the stator-
current space phasor;
at $u_r = 0$ and Ψ_r = ct.
(controlled rotor fux):
$i_{mr} = i_{sd\lambda r}$
but
$i_{sd\lambda s} \neq i_{ms}$

In the case of the IM with cage- or short-circuited rotor ($u_r = 0$), if the Ψ_r may be considered at constant value (that means steady-state or Ψ_r is a controlled variable with constant reference), the SPhs of the rotor-current $\underline{i}_r$ and rotor-flux $\underline{\Psi}_r$ are perpendicular each to other. This property led to the idea of the original field-orientation principle based on the RFO-ed coordinate axes, in which the stator-current SPh $\underline{i}_{s\lambda r} = i_{sd\lambda r} + \mathbf{j}\, i_{sq\lambda r}$ is decomposed into two components, as follows

$$i_{sd\lambda r} = i_{mr} = \Psi_r/L_m \;\; and \;\; i_{sq\lambda r} = m_e/k_{Mr}\Psi_r \; = \; -(1+\sigma_r)\, i_r. \tag{2}$$

Above i_{mr} is the rotor-flux-based magnetizing current; it results from the rotor-flux controller in the magnetic control loop and corresponds to the direct stator-current component $i_{sd\lambda r}$, i.e. the field-producing (reactive) one. In the mechanical control loop the speed or torque (m_e) controller generates the torque producing (active) $i_{sq\lambda r}$ quadrature component, as is shown in Figure 1.

If the frequency converter is controlled in current, the control variables are directly generated by the flux and speed/torque controllers. If the IM is controlled in voltage, the computation of the stator-voltage control variables is based on the RFO-ed model, which is highly complex and motor parameter dependent, as was mentioned before [1], [4], [8], [9], [10]. Furthermore, for the computation of the induced rotating EMFs there was adopted the absolute slip$\Delta\omega$ compensation (with the rotor actual speed ω_r) in order to determine the synchronous angular speed of the rotating magnetic field $\omega_{\lambda r}$ and then the position angle λ_r of it, by integration of the synchronous speed, as follows:

$$\omega_{\lambda r} = \mathrm{d}\lambda_r/\mathrm{d}t = \Delta\omega + \omega_r \text{ and } \lambda_r = \int \omega_{\lambda r}\, \mathrm{d}t. \tag{3}$$

Because the initial position of the rotor field usually is not known, this procedure leads to the so called "*indirect field-orientation*" (IFO) [1], [2], [4], [5], [8], [9]. The computation of the absolute slip may be made according to the expression $\Delta\omega = \tau_r^{-1}\, i_{sq\lambda r}/i_{sd\lambda r}$, which is also rotor-parameter dependent [3].

2.2 Stator-Field Orientation (SFO) Procedure

SFO means that the direct axis of the coordinate frame, denoted dλ_s is oriented in the direction of the resultant stator-flux vector $\underline{\Psi}_s$, therefore [1]:

$$\Psi_{sd\lambda s} = \Psi_s = |\underline{\boldsymbol{\Psi}}_s| \quad \text{and} \quad \Psi_{sq\lambda s} = 0\,. \tag{4}$$

In the stator-field-oriented (SFO-ed) axes frame the stator-current SPh may be written with components as $\underline{\boldsymbol{i}}_{s\lambda s} = i_{sd\lambda s} + \mathbf{j}\, i_{sq\lambda s}$, where

$$i_{sd\lambda s} \neq i_{ms} = \Psi_s / L_m \quad and \quad i_{sq\lambda s} = m_e / \, k_M \, \Psi_s\,. \tag{5}$$

Comparing (5) with (2), it must be remarked that nevertheless the active component is also here proportional to the electromagnetic torque, but the reactive one is no more equal to the stator-flux-based magnetizing current i_{ms} (see Figure 2).

On the other hand in SFO-ed axis frame the stator-flux SPh has only one component (the direct one), which is equal to its module. As a consequence, in comparison with RFO, in SFO schemes the stator-voltage equation gives a more simple computation of the VSI control variables, as follows [1], [6], [8]:

$$u_{sd\lambda s} = R_s\, i_{sd\lambda s} + e_{sd\lambda s} \qquad \text{and} \qquad u_{sq\lambda s} = R_s\, i_{sq\lambda s} + e_{sq\lambda s}\,. \tag{6}$$

Above the direct component $e_{sd\lambda s}$ = dΨ_s/dt is the self-induced EMF due to the variation in magnitude of the Ψ_s . It is zero in steady state or at Ψ_s = ct. The quadrature component $e_{sq\lambda s} = \omega_{\lambda s}\, \Psi_s$ is generated by the rotation of the stator field with the speed $\omega_{\lambda s} = \mathrm{d}\lambda_s/\mathrm{d}t$, which integrated gives the position angle $\lambda_s = \int \omega_{\lambda s}\, \mathrm{d}t$ of the $\underline{\Psi}_s$. SPh.

For voltage-PWM-VSI-fed drives SFO is recommended, due to a simpler voltage model [1], [8], [11], [12]. The computation of the control variables, based on expressions (6) is affected only by the stator resistance R_s , which – if it is necessary – may be identified also online.

3 Comparison of the Stator- and Rotor-Flux Control

The flux control of an IM may be made directly by means of a proper flux controller imposing the reference value or indirectly, controlling other quantities and resulting inherently the desired flux value. It is well known the *Kloss*'s equation, which gives the analytical expression of the IM static mechanical characteristics at constant U_s voltage and constant f_s frequency. If the pull-out critical torque is kept at constant value by adjusting the stator voltage, for different frequencies the characteristics have different shapes [13].

In Figure 3 there are represented two U_s = ct characteristics – torque versus the absolute slip $\Delta\Omega$ (measured in electrical rad/sec) – for the rated frequency f_{sN} at rated voltage U_{sN} and for $f_s = 0$ at U_{so} , which gives the same break-down torque. The U_s = ct speed-torque characteristics at different frequencies are not parallel, due to the different feature of the slip curves, which are found between the two extreme ones presented in Figure 3 [13].

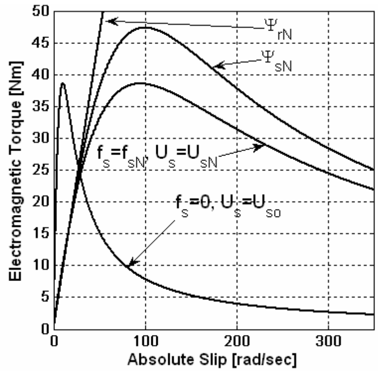

Figure 3
Stator-voltage-, stator-flux- and rotor-flux-controlled mechanical characteristics: electromagnetic torque versus absolute slip $\Delta\Omega$, measured in electrical angle (at zero rotor speed the absolute slip is 314 rad/sec)

In a field-oriented VC scheme usually it is controlled the module of the orientation flux SPh. In the next there will be compared the static mechanical characteristics at constant flux, which are also shown in figure 3 at rated values for SFO ($\Psi_s = \Psi_{sN}$) and RFO ($\Psi_r = \Psi_{rN}$).

3.1 Stator-Flux Controlled (SFC) Mechanical Characteristics

If instead of the stator voltage the stator flux is kept at constant value, results a simplified *Kloss*'s expression, where the coefficient, that is depending on the stator frequency f_s (a typical feature the of the mechanical characteristics at U_s = ct.) is disappeared. In such conditions the following analytical expression results [13]:

$$M_e = 2M_{k_s}\left(\frac{\Delta\Omega_{k_s}}{\Delta\Omega}+\frac{\Delta\Omega}{\Delta\Omega_{k_s}}\right)^{-1};\; M_{k_s} = k_M\,\frac{\Psi_s^2}{2L_m}\cdot\frac{1-\sigma}{\sigma(1+\sigma_s)};\quad \Delta\Omega_{k_s} = \frac{1}{\sigma\tau_r}. \tag{7}$$

M_{ks} is the pull-out torque, $\Delta\Omega_{ks}$ – the critical slip (at Ψ_s = ct.), σ – the resultant leakage coefficient: $\sigma = 1 - 1/(1+\sigma_s)(1+\sigma_r)$, where $\sigma_{s,r} = L_{\sigma s,r}/L_m$ is the stator and rotor leakage coefficient, respectively. The self-cyclic inductance L_m corresponds to the three-phase useful field. The torque coefficient k_M is z_p 3/2 or 3 z_p , depending on the torque calculation based on the flux peak-value or r.m.s. one, respectively [1].

The torque-slip characteristic at Ψ_s = ct from Figure 3, in spite of the fact it is a combination of a linear- and a hyperbolic shape, it is valid for any stator frequency. Consequently, the speed-torque characteristics at different stator frequencies will be parallel, excepting the flux-weakening region, only [13].

3.2 Rotor-Flux Controlled (RFC) Mechanical Characteristics

Particularly in the case of the resultant rotor-flux-control the characteristics become linear without any hyperbolic effect, according to the expression [13]:

$$M_e = 2M_{k_r} \frac{\Delta\Omega}{\Delta\Omega_{k_r}} = \frac{k_M \Psi_r^2}{R_r} \Delta\Omega \; . \tag{8}$$

The RFC – due to the linearity of the static mechanical characteristics at Ψ_r= ct. – ensures more stability at speed reference and torque perturbation of the IM with respect to SFC-ed drives, where the mechanical characteristics present pull-out critic torque due to the so called “*Kloss*” feature given by a hyperbolic shape [3].

4 Comparison of Stator- and Rotor-Flux Identification

There are two basic procedures for flux identification: the so called I-Ω (stator-current & rotor speed) method for Ψ_r and the integration of the stator-voltage equation for $\Psi_{s.}$

4.1 Stator-Flux Identification (SFI) Methods

The simplest procedure for the calculus of the stator flux is based on the stator-voltage model, written with natural two-phase components in the stator-fixed axis frame. In Figure 1 it is made in two steps. First in block e_sC are computed the stator EMFs according to equations:

$$\mathrm{d}\Psi_{sd}/\mathrm{d}t = e_{sd} = u_{sd} - R_s\, i_{sd} \qquad \text{and} \qquad \mathrm{d}\Psi_{sq}/\mathrm{d}t = e_{sq} = u_{sq} - R_s\, i_{sq}\,, \tag{9}$$

followed by the direct integration of them in the identification block Ψ_sId:

$$\underline{\boldsymbol{\Psi}}_s = \Psi_{sq} + \mathbf{j}\,\Psi_{sq}\,, \text{ where } \Psi_{sd} = \int e_{sd}\,\mathrm{d}t \qquad \text{and} \qquad \Psi_{sq} = \int e_{sq}\,\mathrm{d}t\,. \tag{10}$$

The e_sC block has inputs the two-phase reaction variables, resulting from the two phase-transformation (PhT) blocks: one of the measured stator-currents and the other of the stator-voltages, identified in block U_sId. The voltage identification is made using the measured DC-link voltage and the PWM logic signals generated by the SVM modulator block.

This flux identification procedure leads to the “*direct field-orientation*” (DFO), where the position λ_s of the flux SPh may be directly identified without integration by means of a vector analyzer (VA). In the ‘70s this flux identification method could be applied only for the current-source inverter (CSI) fed drives operating with full-wave currents and quasi-sine-wave terminal voltages, due to the freely induced rotating EMFs [1]. In the last two decades it became possible also for VSI-fed drives operating with relatively high PWM sampling frequency. It seems it is the most preferable method for the stator-flux identification, due to the fact it leads to the DFO and it is not affected by the motor parameters, excepting R_s . Its applicability depends first of all on the quality of the integration procedure [14].

The Ψ_s may be also calculated with leakage flux compensation, from the rotor- or air-gap flux, if one of them is identified based on another method [1].

4.2 Rotor-Flux Identification (RFI) Methods

Still in the ‘80s for PWM-VSI-fed drives, the rotor-model-based I-Ω -flux identification procedures were preferable. It was introduced by *Hasse* in 1969 [8]. There are two possibilities to perform it: with natural (stator-fixed) stator-current components, that leads to DFO or with RFO-ed ones, method that needs slip compensation, according (3) and therefore may permit only IFO [1], [4], [9]. Both I-Ω methods are strongly affected by the rotor parameters. Nowadays it is preferable the compensation of the stator-flux obtained with SFI based on direct integration presented before in 4.1. Using the measured stator currents the compensation may be realized in SPh-form, as follows:

$$\underline{\boldsymbol{\Psi}}_r = \Psi_{rq} + \mathbf{j}\ \Psi_{rq} = (1+\sigma_r)\ \underline{\boldsymbol{\Psi}}_s - (1-\sigma)^{-1}\ \sigma\ L_m\, \underline{\boldsymbol{i}}_s; \tag{11}$$

In Figure 1 the compensation is made in block Ψ_sCo without any cross effect between the d-q components, according to expressions (11).

5 Control Scheme with Dual-Field Orientation

Figure 1 presents the control scheme in which are applied the both field-orientation procedures: RFC and RFO of the stator-current decoupled components, generated by the speed and flux controllers, and then transformed into SFO-ed variables for the control of the stator-currents and then computation of the stator-voltage at the re-coupling side of the IM control scheme in order to command the VSI with voltage controlled SVM. This structure combines the advantages offered by each field-orientation procedure of the IM drive fed by a voltage-controlled frequency converter, as follows:

a) The direct RFC ensures a good static stability and overloading capacity due to the linearity of the mechanical characteristics at Ψ_r = ct.;

b) The decoupled control of the mechanical and magnetic phenomena realized by means of the RFO-ed components of the stator-current ensures a good dynamic to the IM drive;

c) Based on SFO-ed two-phase model, the computation of the stator-voltage control variables is made in the simplest manner – realized by the separation of the two kinds of EMFs – according equations (6). It eliminates the influence of the rotor parameters and provides the control structure with robustness;

The RFO-ed stator-current control variables $i_{sd\lambda r} - i_{sq\lambda r}$ are generated by the flux and speed controllers, and then the SFO-ed ones $i_{sd\lambda s} - i_{sq\lambda s}$ are computed by means of a coordinate transformation block CooT, indicated with matrix operator $[D(\lambda_s - \lambda_r)]$. The deviation angle $\lambda_s - \lambda_r$ between the two orientation fluxes (shown in Figure 2) may be computed also in a CooT block. Its inputs are the "oscillatory" matrices symbolized with $[o(\lambda_s)]$ and $[o(\lambda_r)]$, resulting from the VAs of the orientation fluxes. Matrix $[o(\lambda)] = [\cos(\lambda), \sin(\lambda)]^t$ – required for the CooT blocks – is a transposed, i.e.column matrix, which is containing two trigonometrical functions of the flux position angle. A VA usually computes the polar coordinates of a SPh from the two-phase coordinates: its module (equal to the amplitude or r.m.s. value of the sine wave variables) and angular position [1].

The stator-voltage control variables are computed in the U_sC block based on equations (6), where the input EMFs are coming from the feedback side and the *Ohm*'s law voltage drops are generated by the controllers of the SFO-ed current components. The SFO-ed stator-voltage control variables have to be transformed into natural (stator-fixed) two-phase coordinates (by means of the CooT marked with the matrix operator $[D(-\lambda_s)]$) and then into polar coordinates (using a VA), where $u_s = |\underline{\boldsymbol{u}}_s|$ is the module of the stator-voltage SPh $\underline{\boldsymbol{u}}_s$ and γ_s its position angle with respect to the fixed reference axis, both necessary as reference for generation the PWM logic in the SVM block [17].

6 Simulation results

Based on structure from Figure 1 simulations were performed in MATLAB-Simulink® environment. The name-plate data of the simulated and experimented cage IM are: $P_N = 2.2$ kW, $f_{sN} = 50$ Hz, $n_N = 1420$ rpm, 2 pole-pairs, $U_{sN} = 220\ \mathrm{V}^{\mathrm{r.m.s}}$, $I_{sN} = 4.7\ \mathrm{A}^{\mathrm{r.m.s.}}$, $\cos\varphi_N = 0.82$. In figure 3 at $f_s = 0$ $U_{so} = 21.42$ V.

The load torque of the drive is linear speed-dependent with reactive character. At the rated speed (ω_r^{el} = 297 rad/s) its value is equal to the IM electromagnetic torque (m_{eN} = 18.02 Nm) operating with rated data. The perturbation of the drive is achieved by changing the sense of the speed reference, imposed at rated value after starting at t = 1.5 s, when the drive achieved already the steady-state. Consequently, the steady-state of the induction motor is at 50 Hz for both rotational directions. The simulation results of the dual field-oriented control scheme of the IM are presented in Figure 4-14. Comparing them with those of the SFO-ed ones from [2], [3], [5], [6], the drive presents improved behavior.

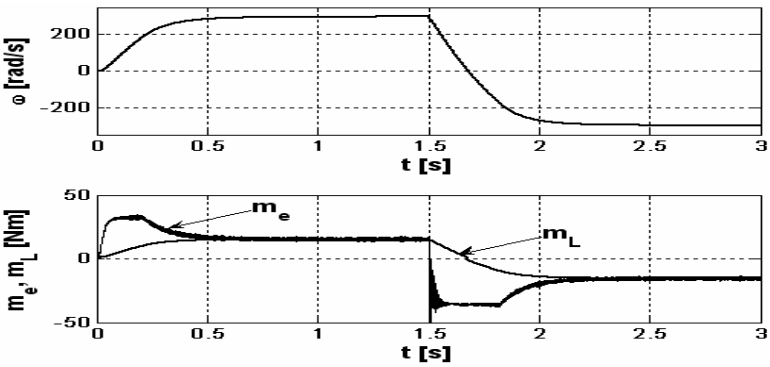

Figure 4
The rotor electrical angular speed (ω), electromagnetic torque (m_e) and load torque (m_L) versus time

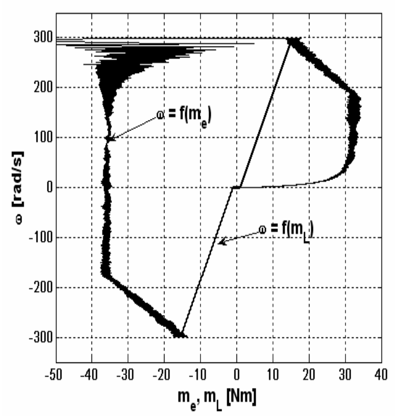

Figure 5
Mechanical characteristics ω = f(m_e) of the IM and ω=f(m_L) of the mechanical load

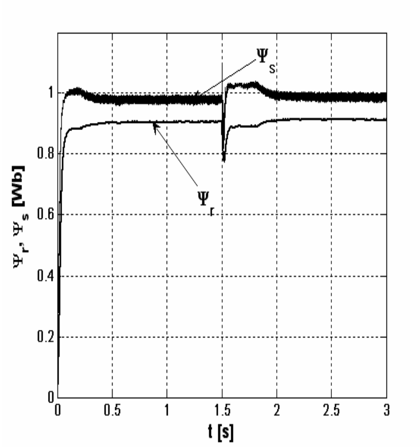

Figure 6
The amplitude of the stator-flux and the controlled rotor-flux versus time

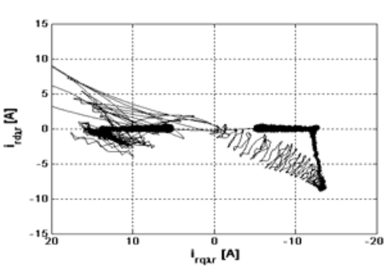

Figure 7
The trajectory of the rotor-current space-phasor in rotor-flux-oriented reference frame

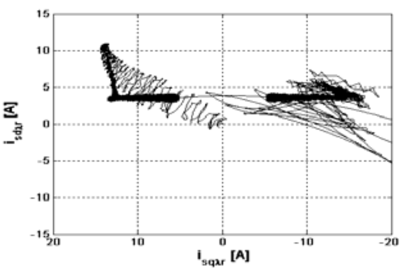

Figure 8
The trajectory of the stator-current space-phasor in rotor-flux-oriented reference frame

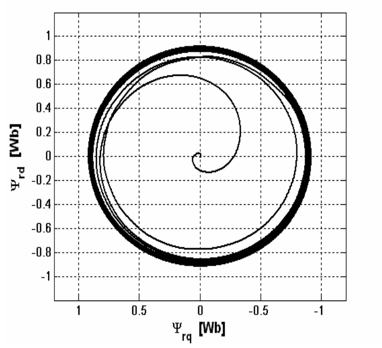

Figure 9
The trajectory of the rotor-flux space-phasor
in stator-fixed coordinate frame

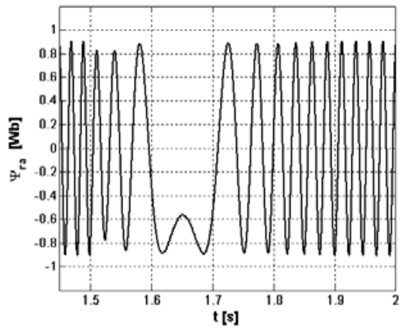

Figure 10
Rotor-flux Ψ_{ra} in phase *a* versus time
at speed reversal

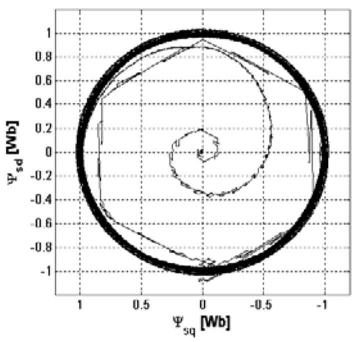

Figure 11
The trajectory of the stator-flux space-phasor
in stator-fixed coordinate frame

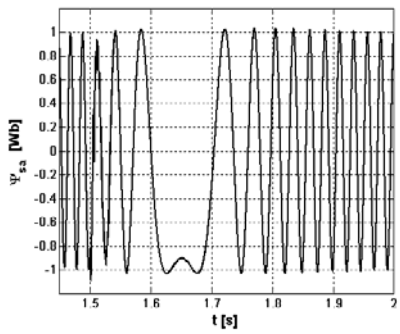

Figure 12
Stator-flux Ψ_{sa} in phase *a* versus time
at speed reversal

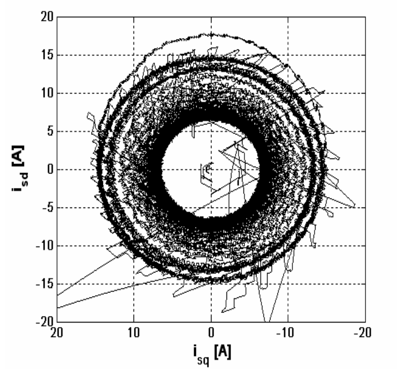

Figure 13
The trajectory of the stator-current space-phasor
in stator-fixed coordinate frame

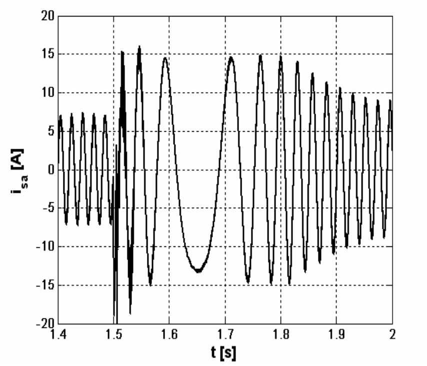

Figure 14
Stator-current i_{sa} in phase *a* versus time
at speed reversal

Conclusions

The RFC-ed IM with RFO-ed structure has a similar behavior as a DC machine, not only as dynamics, but also as static stability due to the linear mechanical characteristics. 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 running may suffer from the sensitivity to motor parameters, like the rotor resistance, rotor time constant and leakage coefficients.

The SFO with SFC, especially used for voltage controlled converter-fed drives, are less computationally demanding and more robust, but the reaction to torque commands is somewhat sluggish, which, in low-inertia drives, could lead to stability problems. The best control scheme seems to be that with RFO with RFC and current-controlled converter as actuator. In comparison with the above two systems, 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 [16]. Voltage-feedforward VSI-fed drives usually with SFC, either scalar or vector structure, generally can not ensure the same performance, either in stability and torque ripple, nor in dynamics (both in reversal process and at torque step perturbation) in comparison with RFC achieved by current-controlled VSI, due to the natural behavior of the IM, considering the magnetizing and torque producing phenomenon. For voltage-controlled IM drives the dual field-orientation may combine the advantages of the two types of field-orientation procedure, on the one hand of the RFC and RFO and on the other hand of the SFO, in order to ensure reduced computational demand, increased static stability and overloading capacity, a good dynamic and robustness, avoiding the influence of the rotor parameters.

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