DOUBLE FIELD ORIENTATED VECTOR CONTROL STRUCTURE FOR CAGE INDUCTION MOTOR DRIVE

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Abstract. *There is proposed a new structure of vector control for the short-circuited induction motor drives fed by voltage-source inverter (VSI) with voltage-feedforward pulse-width modulation (PWM), which combines the advantages of two types of field-orientated procedure, i.e. rotor-field orientation and control for generating stator-current control variables by controllers and stator-field orientation, for computation of the stator-voltage ones.*

Keywords: *vector control, stator- and rotor-field orientation, flux identification, voltage-source inverter, carrier-wave PWM.*

1. Introduction

The simplest vector control (VC) structure for induction motor (IM) drives is achieved by current controlled static frequency converter, 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], [4], [5].

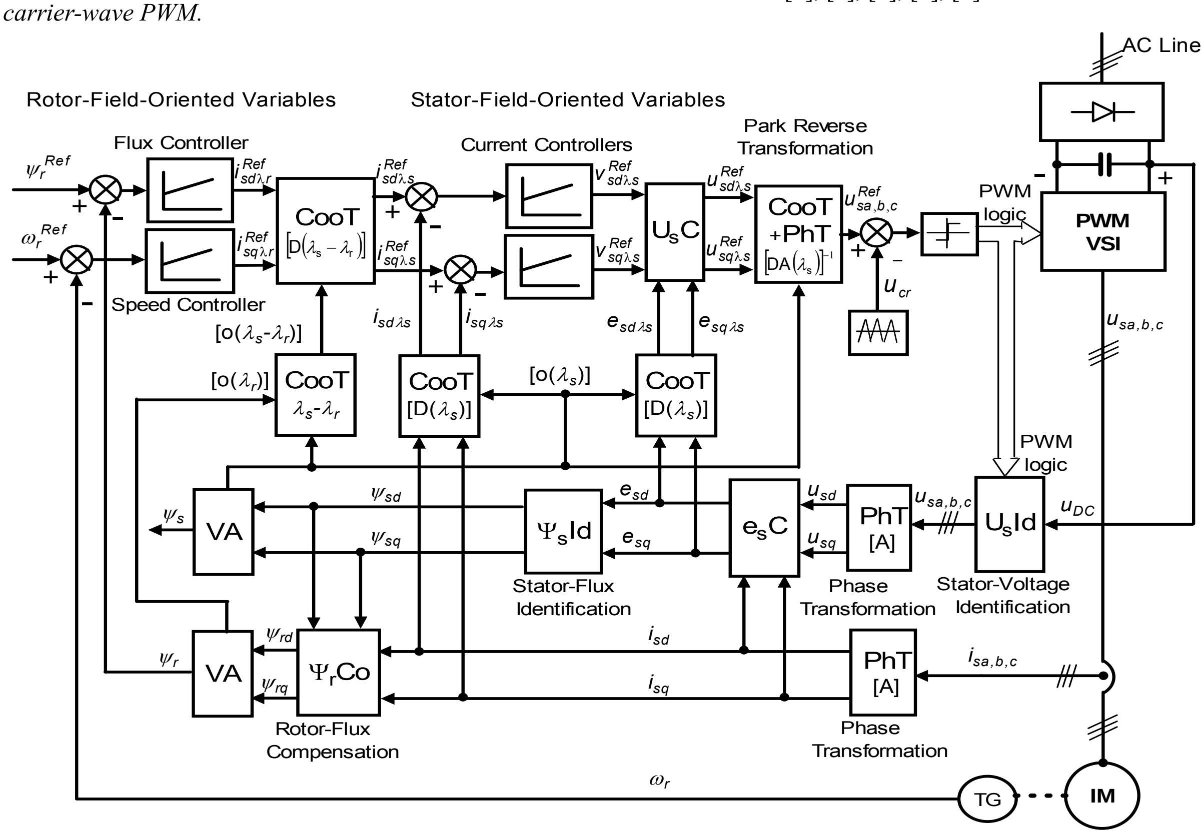

Fig. 1. Vector control structure of the cage induction motor drive with double field orientation: rotor-field-orientated variables at the decoupled control side and stator-field-orientation at the re-coupling of the two control loops.

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 (CW) and space-vector modulation (SVM). That means the IM may be supplied by a proper voltage-source converter with voltage-control. 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 τ_r, leakage coefficients and others.

Consequently, the drive control performance may be lightly damaged. Usually this problem is solved by renouncing on the RFC and applying SFO, which leads to a much simpler stator-voltage computation and dependent only on the stator resistance (R_s).

For the above presented reasons a new vector control structure is proposed with double field orientation (DFO), as follows: RFC and RFO of the stator-current components generated by the speed and flux controllers at the decoupled control side, and then transformed into SFO variables for stator-current control and stator-voltage computation at the re-coupling side of the control scheme, as is shown in *Fig. 1*.

2. Comparison of rotor - and stator-field orientation

In classical field-orientation of the IM drives with short-circuited rotor, proposed initially in 1971 by *Blaschke* [6], usually the rotor flux (Ψ_r) is controlled and the stator-current space-phasor (SPh) or the so called Park vector is oriented according to the resultant rotor field.

2.1. Rotor-field orientation (RFO)

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 *Fig. 2*, and the flux components are:

$$\Psi_{rd\lambda r} = \Psi_r = |\underline{\Psi}_r| \quad \text{and} \quad \Psi_{rq\lambda r} = 0 . \qquad (1)$$

In the case of the IM with 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), the rotor-current $\underline{i}_r$ and rotor-flux $\underline{\Psi}_r$ SPhs are perpendicular each to other. This property led to the idea of the original field-orientation principle based on the rotor-field-oriented (RFO-ed) coordinate axes, in which the stator-current SPh may be decomposed into two components:

$$\underline{i}_{s\lambda r} = i_{sd\lambda r} + \mathrm{j}\, i_{sq\lambda r} , \qquad (2)$$

where the RFO components of the stator-current SPh are

$$i_{sd\lambda r} = i_{mr} = \Psi_r / L_m \ \text{ and } \ i_{sq\lambda r} = m_e / \Psi_r (1+\sigma_r) = -i_r . \ (3)$$

Consequently, in the magnetic control loop from the rotor-flux controller results $i_{sd\lambda r}$ the field-producing (reactive) component (it is equal to i_{mr} the rotor-flux-based magnetizing current), and in the mechanical control loop the speed or torque (m_e) controller will generate $i_{sq\lambda r}$ the torque producing (active) one, as is shown in *Fig. 1*.

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 from the controllers. If the IM is controlled in voltage, the computation of the stator-voltage components based on the RFO-ed model is highly complex and motor parameter depending, as was mentioned before [1], [7], [8], [9], [10].

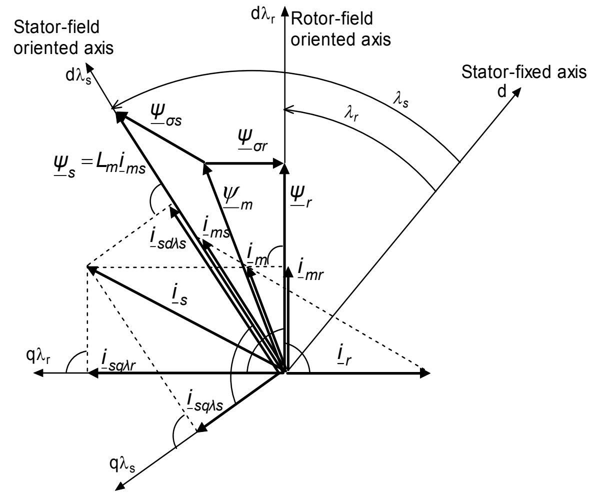

Fig. 2. Phasor diagram of magnetizing currents, fluxes and the stator-current field-orientated components.

Furthermore, for the computation of the induced rotating EMFs there was adopted the absolute slip compensation (with the rotor actual speed ω_r) in order to determine the angular speed of the rotating magnetic field as follows [1], [2], [4], [7], [8], [9]:

$$\omega_{\lambda r} = d\lambda_r / dt = \Delta\omega + \omega_r , \qquad (4)$$

and then the position angle of it by integration of the synchronous speed $\lambda = \int \omega_\lambda \, dt$. Because the initial position usually is not known, this procedure leads to the so called "*indirect field-orientation*" (IFO).

The computation of the absolute slip may be made as

$$\Delta\omega = \tau_r^{-1} \, i_{sq\lambda r} / i_{sd\lambda r} \qquad (5)$$

which is also rotor-parameter dependent [3],.

2.2. Stator-field orientation (SFO)

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

$$\Psi_{sd\lambda s} = \Psi_s = |\underline{\Psi}_s| \quad \text{and} \quad \Psi_{sq\lambda s} = 0 . \qquad (6)$$

In the stator-field-oriented (SFO-ed) axes frame the stator-current SPh may be written with components as:

$$\underline{i}_{s\lambda s} = i_{sd\lambda s} + \mathrm{j}\, i_{sq\lambda s} , \qquad (7)$$

where the SFO-ed stator-current components are

$$i_{sd\lambda s} \neq i_{ms} = \Psi_s / L_m \quad \text{and} \quad i_{sq\lambda s} = m_e / \Psi_s . \qquad (8)$$

Comparing (8) with (3), 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}, as is observable also in *Fig. 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 [1], [5], [7]:

$$u_{sd\lambda s} = R_s\, i_{sd\lambda s} + e_{sd\lambda s} \text{ and } u_{sq\lambda s} = R_s\, i_{sq\lambda s} + e_{sq\lambda s}\,. \quad (9)$$

where EMFs may be written

$$e_{sd\lambda s} = \mathrm{d}\Psi_s/\mathrm{d}t \quad \text{and} \quad e_{sq\lambda s} = \omega_{\lambda s}\, \Psi_s\,. \quad (10)$$

Above the direct component is the self-induced EMF due to the variation in magnitude of the Ψ_s . It is zero in steady state. The quadrature component is generated by the rotation of the stator field with the speed $\omega_{\lambda s} = \mathrm{d}\lambda_s/\mathrm{d}t$.

For voltage-PWM-VSI-fed drives – due to a simpler voltage model – stator-field orientation is recommended [1], [7], [11], [12]. The computation of the control variables, based on expressions (9) is affected only by the stator resistance R_s , which 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 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. In *Fig. 3* there are represented two U_s = ct characteristics – torque versus the absolute slip ΔΩ (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 has the same break-down torque. To the zero rotor speed corresponds ΔΩ = 314 rad/sec, measured in electrical angle.

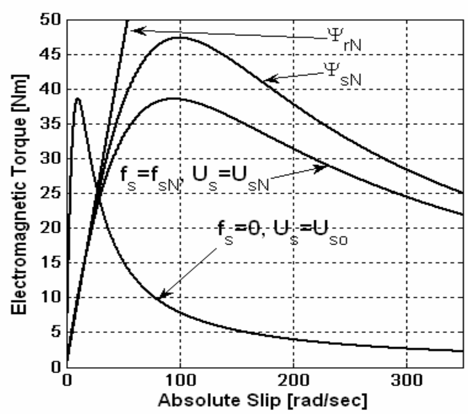

Fig. 3. Stator-voltage-, stator-flux- and rotor-flux-controlled mechanical characteristics: torque versus absolute slip.

The U_s = ct speed-torque characteristics – due to the different feature of the slip curves from *Fig. 3* – at different frequencies are not parallel [13].

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

3.1. Stator-flux controlled (SFC) characteristics

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

$$M_e = 2M_{k_s}\left(\frac{\Delta\Omega_{k_s}}{\Delta\Omega} + \frac{\Delta\Omega}{\Delta\Omega_{k_s}}\right)^{-1}, \quad (11)$$

with the pull-out torque and critical slip

$$M_{k_s} = k_M \frac{\Psi_s^2}{2L_m}\cdot\frac{1-\sigma}{\sigma(1+\sigma_s)}; \quad \Delta\Omega_{k_s} = \frac{1}{\sigma\tau_r}\,, \quad (12)$$

where $\sigma_s = L_{\sigma s}/L_m$ is the stator leakage coefficient and σ the resultant one. The self-cyclic inductance L_m corresponds to the three-phase useful field.

The torque coefficient $k_M = z_p\, 3/2$ if Ψ_s and Ψ_r are the peak value or $k_M = 3\, z_p$, if they are r.m.s. values.

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

3.2. Rotor-flux controlled (RFC) characteristics

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

$$M_e = 2M_{k_r}\frac{\Delta\Omega}{\Delta\Omega_{k_r}} = \frac{k_M\Psi_r^2}{R_r}\Delta\Omega\,. \quad (13)$$

The RFC due to the linearity of the static mechanical characteristics at Ψ_r= ct ensures more stability in behavior of the IM with respect to SFC-ed drives, where the characteristics present pull-out critic torque due to the so called "*Kloss*" feature given by a hyperbolic shape.

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 Ψ_s.

4.1. Stator-flux identification (SFI)

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 *Fig. 1* it is made in two steps. First in block **e_sC** are computed the stator EMFs according to equations:

$$e_{sd} = u_{sd} - R_s\, i_{sd} \text{ and } e_{sq} = u_{sq} - R_s\, i_{sq}\,, \quad (14)$$

followed by the direct integration of them in the identification block **Ψ$_s$Id**, as follows:

$$\Psi_{sd} = \int e_{sd}\,\mathrm{d}t \quad \text{and} \quad \Psi_{sq} = \int e_{sq}\,\mathrm{d}t\,. \quad (15)$$

The **e$_s$C** 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$_s$Id**. The voltage identification is made using the measured DC-link voltage and the PWM logic signals generated by the bi-level voltage controllers.

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 direct field-orientation 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].

4.2. Rotor-flux identification (RFI)

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 [7]. There are two possibilities to perform it: with natural (stator-fixed) stator-current components, that leads to DFO or with RFO-ed ones, which needs slip compensation and therefore may offer only IFO [1], [8], [9]. Both I-Ω methods are strongly affected by the rotor parameters.

Nowadays it is preferable the stator-flux compensation obtained with the before presented SFI, using the measured stator currents, in synthesized form, as:

$$\Psi_{rd/q} = (1+\sigma_r)\,\Psi_{sd/q} - (1-\sigma)^{-1}\,\sigma\,L_m\,i_{sd/q}; \qquad (16)$$

where $\sigma_r = L_{\sigma r}/L_m$ is the rotor leakage coefficient and the coefficient of L_m may be written as $(\sigma_s + \sigma_s\sigma_r + \sigma_r)$.

In *Fig. 1* the compensation is made in block **ΨsCo** without any cross effect between the d-q components, according to expressions (16).

5. Control scheme with double-field orientation (DFO)

Fig. 1 presents the control scheme in which are applied the both field-orientation procedures RFO and SFO, combining the advantages offered by each of them for the induction motor drive fed by a voltage-controlled static frequency converter:

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

b) The decoupling control of the mechanical and magnetic phenomena realized by means of the RFO-ed components of the stator-current presses 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 (9) and (10). It eliminates the influence of the rotor parameters and provides the control structure with robustness;

The RFO-ed variables are generated by the flux and speed controllers, but the SFO-ed ones are computed by means of coordinate transformation blocks (**CooT**).

The two kind of control variables are coupled with the field-oriented stator-current components, by means of a **CooT** block indicated with matrix operator $[D(\lambda_s\text{-}\lambda_r)]$.It has at the inputs the RFO-ed components $i_{sd\lambda r} - i_{sq\lambda r}$ and at the output the SFO-ed ones $i_{sd\lambda s} - i_{sq\lambda s}$. The deviation angle λ_s-λ_r between the two orientation fluxes (see *Fig. 2*) is computed in another **CooT** block, with inputs $[o(\lambda_s)]$ and $[o(\lambda_r)]$, resulting from the **VA**s of the stator- and rotor-fluxes [1]. A VA usually computes the polar coordinates of a SPh from the two-phase coordinates, that means the module (amplitude or r.m.s. value of the sine wave variables) and its angular position.

The trigonometry functions required for the **CooT** blocks are symbolized with an “oscillatory” matrix containing two elements:

$$[o(\lambda)] = [\cos(\lambda), \sin(\lambda)\,]^t\,. \qquad (17)$$

The stator-voltage control variables are computed in the **UsC** block based on equations (9), 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. At the output the SFO-ed control variables of the stator-voltage have to be transformed into natural (stator-fixed) two-phase coordinates (CooT with the matrix operator $[D(-\lambda_s)]$) and then in three-phase components (PhT with a matrix operator $[A]^{-1}$). The two transformations (**CooT**+**PhT**) may be realized in one step by means of a reverse Park transformation block, marked with operator $[DA(\lambda_s)]^{-1}$.

6. Simulation results

Based on structure from *Fig. 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 V$^{\text{r.m.s}}$, I_{sN} = 4.7 A$^{\text{r.m.s.}}$, cos φ_N = 0.82. In *Fig. 3* U_{s0} = 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 = 1.5s, 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 DFO-ed control scheme are presented in *Fig. 4-14*. Comparing them with those of the SFO-ed ones from [2], [3], [4], [5], it present improved behavior.

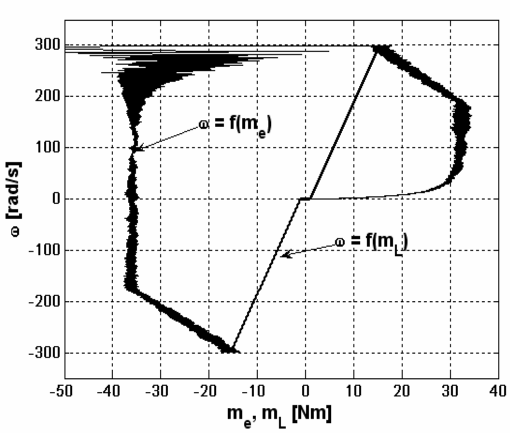

Fig. 4. Mechanical characteristics ω=f(m_e) *of the induction motor and* ω=f(m_L) *of the mechanical load.*

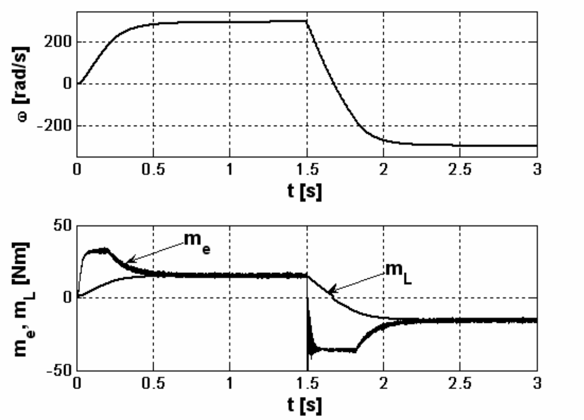

Fig. 5. The rotor electrical angular speed (ω)*, electromagnetic torque* (m_e) *and load torque* (m_L) *versus time.*

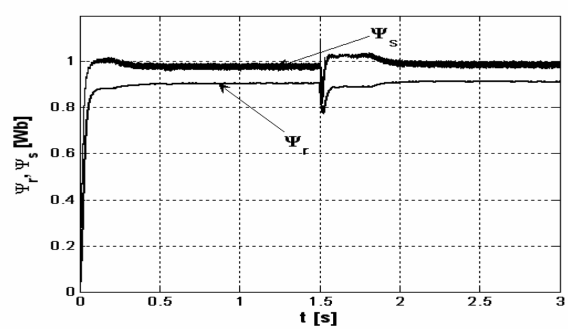

Fig. 6. The rotor-flux and stator-flux amplitude versus time.

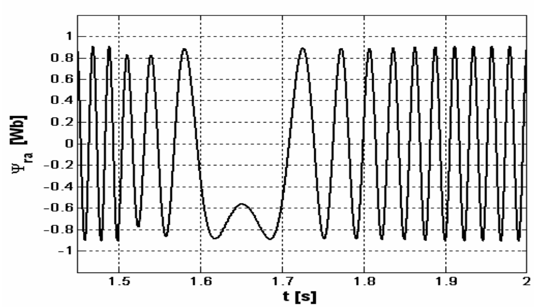

Fig. 7. Rotor-flux Ψ_{ra} *in phase* a *versus time at speed reversal.*

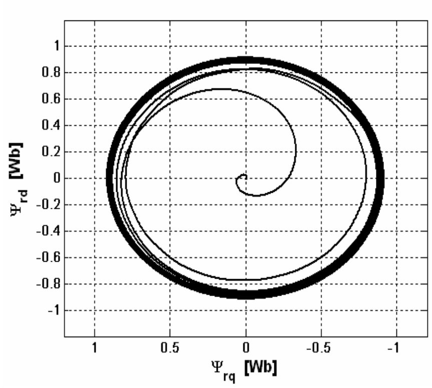

Fig. 8. The trajectory of the rotor-flux space-phasor in stator-fixed coordinate frame.

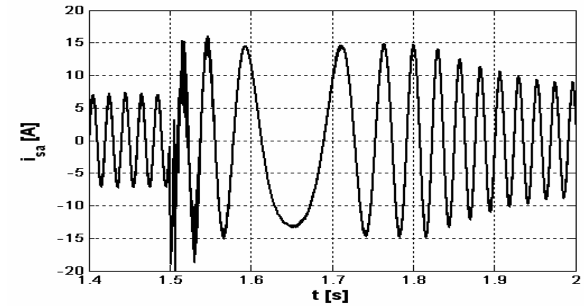

Fig. 9. Stator-current i_{sa} *in phase* a *versus time at speed reversal.*

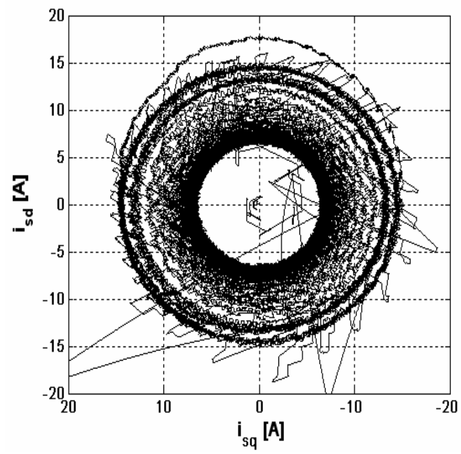

Fig. 10. The trajectory of the stator-current space-phasor in stator-fixed coordinate frame.

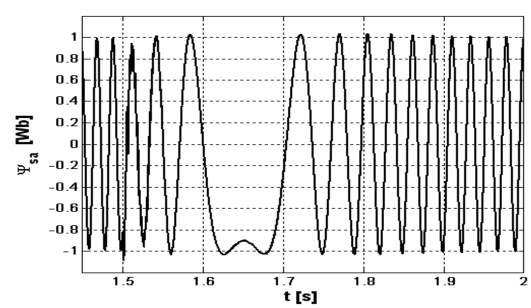

Fig. 11. Stator-flux Ψ_{sa} *in phase* a *versus time at speed reversal.*

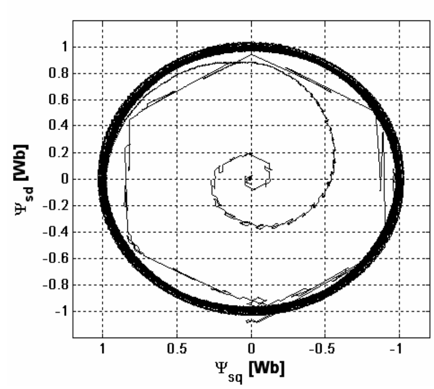

Fig. 12. The trajectory of the stator-flux space-phasor in stator-fixed coordinate frame.

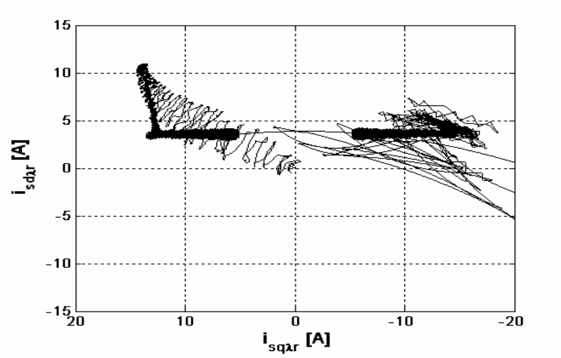

Fig. 13. The trajectory of the stator-current space-phasor in rotor-flux-oriented reference frame.

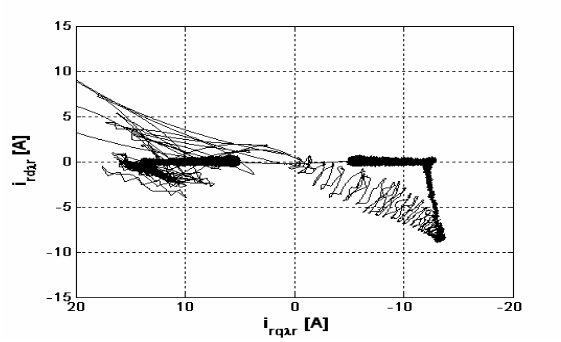

Fig. 14. The trajectory of the rotor-current space-phasor in rotor-flux-oriented reference frame.

7. Conclusions

The RFC-ed IM with RFO-ed structure has a similar behavior as a DC machine, not only as dynamics, but also as 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, especially the coefficients of leakage and rotor time constant.

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.

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 DFO 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 stability, a good dynamic and robustness, avoiding the influence of the rotor parameters.

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