Direct vector control of induction motors based on rotor resistance-invariant rotor flux observer

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***Abstract*— A novel speed-flux tracking controller for induction motors has been developed and experimentally verified. Direct rotor flux field oriented controller is designed for current-fed induction motor model on the base of full order hybrid continuous time-sliding mode flux observer. Controller guarantees local asymptotic speed-flux tracking and asymptotic direct field-orientation under condition of unknown constant load torque. The flux subsystem is invariant with respect to limited rotor resistance variations due to special structure of the flux observer. The efficiency of the proposed solution is confirmed by the results of experimental studies, which demonstrate the improved robustness properties in all motor operating conditions including nearby zero speeds.**

Keywords—induction motor, field-oriented control, sliding mode flux observer.

I. INTRODUCTION

Vector controlled Induction Motor (IM) drives are widely used electromechanical systems suitable for medium and high performance applications including general industrial and electric traction fields. Vector Field Oriented Control (FOC) [1] of IM established a de-facto industrial standard for such applications. Different modifications of indirect and direct field oriented controllers (IFOC, DFOC) have been developed during last few decades to improve dynamic performance and efficiency of the drive systems. The efficiency improvement techniques typically reported in publications adjust the flux level as a function of the electromagnetic torque using various optimization procedures [2]. The accurate flux regulation is required to achieve both control objectives.

In all solutions, based on FOC concept, the flux control subsystem is sensitive to IM parameters variation, especially the rotor resistance (see [3] and references included). As a consequence of rotor resistance perturbations the field orientation is missed, leading to errors in rotor flux vector control (modulus and angle). The wrong field orientation causes degradation of torque and speed transient performance and reduced efficiency of the electromechanical energy conversion.

The parameter sensitivity problem of a standard IM vector control algorithms [1] is well stated and variety of solutions have been considered to improve the flux vector control accuracy. Regarding to flux estimation for DFOC schemes, it is proposed to use different modifications of the full order flux observers [4]. Designed as the closed loop systems, the full order flux observers potentially may deliver improved robustness properties with respect to rotor parameters variations as compared with open loop flux estimators. However, as it is shown in [4], the majority of DFOC controllers for IM do not provide accurate flux control at near zero speed, since robustification is achieved by the action of speed dependent correction terms [5]. Being observerless solutions, the IFOC controllers have less degrees of freedom in control design in order to improve robustness. An improved IFOC controller is designed and experimentally evaluated in [6], [7], where, in contrast to existing solutions, the flux control subsystem has closed loop properties, providing robustness with respect of rotor resistance variation if IM speed is not zero.

Starting from the pioneering works of V.Utkin [8], [9] the variable structure system theory established a new direction to solve the IM control problems. Compared to the ordinary control methods, the variable structure control, operating in sliding mode, has attractive advantages of robustness to external disturbances and low sensitivity to the system parameter variations [9]. Based on the general theoretical result, given in [10], a variety of the sliding mode flux observers has been proposed for IM control. In [11] a general class of manifolds on which sliding mode flux observation and control are achieved is considered. System performance evaluation and robustness properties are studied by simulation. The dynamics of the sliding mode observer, designed in [12], is independent from the rotor resistance, however observer requires zero initial conditions for flux vector and therefore is not asymptotic. From the analysis of available solutions for sliding mode observers it follows that their robustness properties, similarly to full order flux observers, are mainly based on action of the speed dependent correction terms. As result, the development of robust DFOC of IM in full speed range, including nearby zero speed, is still an open research problem.

The aim of this paper is to design a new speed-flux tracking controller for IM based on rotor resistance invariant hybrid continuous time-sliding mode rotor flux observer.

II. PROBLEM STATEMENT

The equivalent two-phase model of the symmetrical IM, under assumptions of linear magnetic circuits and balanced operating conditions, is expressed in an arbitrary rotating reference frame (d-q) as

$$\begin{aligned} &\dot{\omega} = T - T_L / J, T = \mu(\psi_d i_q - \psi_q i_d) \\ &\dot{i}_d = -\gamma_n i_d + \omega_0 i_q + \alpha_n \beta \psi_d + \beta\omega\psi_q + \\ &\quad + \sigma^{-1} u_d + \Delta\alpha\beta(\psi_d - L_m i_d), \\ &\dot{i}_q = -\gamma_n i_q - \omega_0 i_d + \alpha_n \beta \psi_q - \beta\omega\psi_d + \\ &\quad + \sigma^{-1} u_q + \Delta\alpha\beta(\psi_q - L_m i_q), \end{aligned} \tag{1}$$

$$\dot{\psi}_d = -\alpha_n\psi_d + \omega_2\psi_q + \alpha_n L_m i_d - \Delta\alpha(\psi_d - L_m i_d),$$
$$\dot{\psi}_q = -\alpha_n\psi_q - \omega_2\psi_d + \alpha_n L_m i_q - \Delta\alpha(\psi_q - L_m i_q), \qquad (2)$$
$$\dot{\varepsilon}_0 = \omega_0,\ \varepsilon_0(0) = 0,$$

where $(i_d,\ i_q)^T$, $(\psi_d,\ \psi_q)^T$, $(u_d,\ u_q)^T$ denote stator current, rotor flux, and stator voltage vectors, T_L – is the load torque, ω – is the rotor speed, $\omega_2 = \omega_0 - \omega$ – sleep frequency, ε_0 – angular position of the reference frame (d-q) with respect to stationary reference frame (a-b). Positive constants related to electrical and mechanical parameters of the IM are defined as $\alpha = \left(R_{2n}/L_2 + \Delta R_2/L_2\right) \triangleq \alpha_n + \Delta\alpha > 0;\ \beta = L_m/\sigma L_2$; $\gamma_n = R_1\sigma^{-1} + \alpha_n L_m \beta;\ \sigma = L_1 - L_m^2/L_2$; $\mu = 3L_m/2L_2 J$, where L_m – is the magnetizing inductance, R_1, L_1 – stator resistance and inductance, L_2 – rotor inductance, R_{2n}, ΔR_2 – rated value and variation of the rotor resistance, so that $R_2 = R_{2n} + \Delta R_2 > 0$ is an actual value of rotor resistance, J – is the total rotor inertia. One pole pair is assumed without loss of generality.

Transformed variables in (1), (2) are given

$$\mathbf{x}_{(d-q)} = e^{-\mathbf{J}\varepsilon_0}\mathbf{x}_{(a-b)},\quad \mathbf{x}_{(a-b)} = e^{\mathbf{J}\varepsilon_0}\mathbf{x}_{(d-q)},\quad e^{-\mathbf{J}\varepsilon_0} = \begin{bmatrix} \cos\varepsilon_0 & \sin\varepsilon_0 \\ -\sin\varepsilon_0 & \cos\varepsilon_0 \end{bmatrix}, \qquad (3)$$

where $\mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$, $\mathbf{x}_{(y-z)}$ – denote 2-D vectors of voltages, currents and fluxes.

In DFOC systems the angular position ε_0 of reference frame (d-q) in transformation (3) is given by the flux observer.

Considering speed-flux control task let define: vector of controlled variables $\mathbf{y} = \left(\omega, \sqrt{\psi_d^2 + \psi_q^2}\right)^T \triangleq \left(\omega, |\psi|\right)^T$; reference vector $\mathbf{y}^* = \left(\omega^*, \psi^*\right)^T$, where ω^* and $\psi^* > 0$ – are speed and flux reference trajectories; vector of tracking errors $\tilde{\mathbf{y}} = \mathbf{y} - \mathbf{y}^* \triangleq \left(\tilde{\omega}, \tilde{\psi}\right)^T$.

Consider the IM model (2) and assume that: speed and flux reference trajectories ω^*, $\psi^* > 0$ are smooth functions with known and bounded first and second time derivatives; stator currents $(i_d,\ i_q)$ and angular speed ω are available for measurements; load torque T_L is unknown, bounded, and constant; motor parameters are exactly known and constant; variation ΔR_2 is bounded and unknown.

Under these assumptions, it is required to design a controller which guarantees:

O.1. Asymptotic speed and rotor flux tracking and asymptotic field orientation, i.e. $\lim_{t\to\infty}\tilde{\omega} = 0,\ \lim_{t\to\infty}\tilde{\psi} = 0,\ \lim_{t\to\infty}\psi_q = 0$, with all signals bounded.

O.2. Asymptotic flux estimation, i.e. $\lim_{t\to\infty}\tilde{\psi}_d = 0$, $\lim_{t\to\infty}\tilde{\psi}_q = 0$, where $\tilde{\psi}_d = \psi_d - |\hat{\psi}|$, $\tilde{\psi}_q = \psi_q$, are the estimation errors, $|\hat{\psi}|$ – estimate of rotor flux magnitude.

O.3. Invariance with respect to limited rotor resistance variations ΔR_2.

III. Controller Design

A. Flux observer

Let us first consider the following flux observer:

$$\dot{\hat{i}}_d = -\gamma_n\hat{i}_d + \omega_0 i_q + \alpha_n\beta|\hat{\psi}| + \sigma^{-1}u_d + k_{ed1}e_d,$$
$$\dot{\hat{i}}_q = -\gamma_n\hat{i}_q - \omega_0 i_d - \beta\omega|\hat{\psi}| + \sigma^{-1}u_q + \delta\,\text{sign}(e_q),$$
$$|\dot{\hat{\psi}}| = -\alpha_n|\hat{\psi}| + \alpha_n L_m\hat{i}_d,\ |\hat{\psi}| > 0, \qquad (4)$$
$$\dot{\varepsilon}_0 = \omega_0 = \omega + \frac{\alpha_n L_m\hat{i}_q - \beta^{-1}\delta\,\text{sign}(e_q) + e_d\beta^{-1}(\omega_0 + \gamma_1\omega)}{|\hat{\psi}|},$$

where $e_d = i_d - \hat{i}_d$, $e_q = i_q - \hat{i}_q$ – current estimation errors, $(\hat{i}_d,\ \hat{i}_q)$ – estimates of $(i_d,\ i_q)$; $\gamma_1 = (R_1/\sigma + k_{ed1})/\alpha_n > 0$, $(k_{ed1}, \delta) > 0$.

From (1), (2) and (4) the estimation error dynamics is given by

$$\dot{e}_d = -k_{ed}e_d + \alpha_n\beta\tilde{\psi}_d + \beta\omega\tilde{\psi}_q + \Delta\alpha\beta(\psi_d - L_m i_d),$$
$$\dot{e}_q = -\gamma_n e_q + \alpha_n\beta\tilde{\psi}_q - \beta\omega\tilde{\psi}_d + \Delta\alpha\beta(\psi_q - L_m i_q) - $$
$$-\delta\,\text{sign}(e_q), \qquad (5)$$
$$\dot{\tilde{\psi}}_d = -\alpha_n\tilde{\psi}_d + \omega_2\tilde{\psi}_q + \alpha_n L_m e_d - \Delta\alpha(\psi_d - L_m i_d),$$
$$\dot{\tilde{\psi}}_q = -\alpha_n\tilde{\psi}_q - \omega_2\tilde{\psi}_d + \alpha_n L_m e_q - \Delta\alpha(\psi_q - L_m i_q) +$$
$$+\beta^{-1}\delta\,\text{sign}(e_q) - e_d\beta^{-1}(\omega_0 + \gamma_1\omega),$$

where $k_{ed} = \gamma_n + k_{ed1} > 0$.

According to variable system theory [9] if $\delta > \max\left\{|\,\alpha_n\beta\tilde{\psi}_q - \beta\omega\tilde{\psi}_d + \Delta\alpha\beta(\psi_q - L_m i_q)\,|\right\}$ then in second equation of (4) the conditions for sling mode are satisfied on the manifold $e_q \equiv de_q/dt \equiv 0$ with equivalent control, equal to

$$\left[\delta\,\text{sign}(e_q)\right]_{eq} = \alpha_n\beta\tilde{\psi}_q - \beta\omega\tilde{\psi}_d + \Delta\alpha\beta(\psi_q - L_m i_q). \qquad (6)$$

“Fast” movement of variable $e_q(t)$ on the surface $e_q = 0$ reduces the order of system (5) and provides equivalent control (6), which contains the information about flux estimation errors and disturbance, caused by the variations of $\Delta\alpha$.

Taking into consideration the equivalent control (6), system (5) is reduced to

$$\dot{e}_d = -k_{ed}e_d + \alpha_n\beta\tilde{\psi}_d + \beta\omega\tilde{\psi}_q + \Delta\alpha\beta(\psi_d - L_m i_d),$$
$$\dot{\tilde{\psi}}_d = -\alpha_n\tilde{\psi}_d + (\omega_0 - \omega)\tilde{\psi}_q + \alpha_n L_m e_d - \Delta\alpha(\psi_d - L_m i_d), \qquad (7)$$
$$\dot{\tilde{\psi}}_q = -\omega_0\tilde{\psi}_d - e_d\beta^{-1}(\omega_0 + \gamma_1\omega).$$

In order to provide further analysis, let consider a disturbance $(\psi_d - L_m i_d)$ in (7) under conditions of direct field orientation. Standard structure of DFOC uses the PI flux controller in order to provide asymptotic regulation of the estimated flux. In this case $|\hat{\psi}|$ asymptotically converges to ψ^*, and therefore:

$$\psi_d - L_m i_d = \tilde{\psi}_d + |\hat{\psi}| - L_m i_d \simeq \tilde{\psi}_d + \psi^* - L_m i_d^* = \tilde{\psi}_d, \quad (8)$$

i.e. after motor excitation $\psi^* = L_m i_d^*$, when $\psi^* = \text{const}$.

Under condition (8) a reduced order dynamics (7) is independent from the perturbation $\Delta\alpha$ and given by

$$\begin{aligned} \dot{e}_d &= -k_{ed} e_d + \alpha\beta\tilde{\psi}_d + \beta\omega\tilde{\psi}_q, \\ \dot{\tilde{\psi}}_d &= -\alpha\tilde{\psi}_d + (\omega_0 - \omega)\tilde{\psi}_q + \alpha_n L_m e_d, \\ \dot{\tilde{\psi}}_q &= -\omega_0\tilde{\psi}_d - e_d\beta^{-1}(\omega_0 + \gamma_1\omega). \end{aligned} \quad (9)$$

In order to prove the stability of system (9) let introduce a linear coordinate transformation

$$z_d = e_d + \beta\tilde{\psi}_d,\ z_q = \beta\tilde{\psi}_q. \quad (10)$$

Dynamics (9) in new coordinates (10) becomes

$$\begin{aligned} \dot{e}_d &= -(k_{ed} + \alpha)e_d + \alpha_n z_d + \omega z_q + \Delta\alpha z_d, \\ \dot{z}_d &= -\gamma_1\alpha_n e_d + \omega_0 z_q, \\ \dot{z}_q &= -\gamma_1\omega e_d - \omega_0 z_d. \end{aligned} \quad (11)$$

Let consider for system (11) with $\Delta\alpha = 0$ the following Lyapunov's function:

$$V = 0.5\left[e_d^2 + (z_d^2 + z_q^2)\gamma_1^{-1}\right] > 0. \quad (12)$$

Time derivative of (12) along the trajectories of (11) with $\Delta\alpha = 0$ can be derived as follows:

$$\dot{V} = -(k_{ed} + \alpha)e_d^2 < 0. \quad (13)$$

From (12), (13) it can be concluded that signals (e_d, z_d, z_q) are bounded for all $t \geq 0$. Due to

$$V(t) \leq V(0)/(k_{ed} + \alpha_n)$$

the signal e_d is a square-integrable. Applying the Barbalat's Lemma [13], we have $\lim_{t\to\infty} e_d = 0$. From the other hand, system (11) with $\Delta\alpha = 0$, has standard form for adaptive systems

$$\begin{aligned} \dot{e}_d &= -(\alpha + k_{ed})e_d + \mathbf{\Gamma}^T(t)\mathbf{z}_{(d-q)}, \\ \dot{\mathbf{z}}_{(d-q)} &= -\gamma_1\mathbf{\Gamma}(t)e_d + \mathbf{S}(t)\mathbf{z}_{(d-q)}, \end{aligned} \quad (14)$$

where $\mathbf{S}(t) = -\mathbf{S}^T(t) = \begin{bmatrix} 0 & \omega_0 \\ -\omega_0 & 0 \end{bmatrix}$; $\mathbf{\Gamma}(t) = (\alpha_n, \omega)^T$, $\mathbf{z}_{(d-q)} = (z_d, z_q)^T$.

If for system in the standard form (14) the persistency of excitation conditions are satisfied [13] then an equilibrium $(e_d, z_a, z_b)^T = 0$ is globally exponentially stable. From (10) it follows that $\lim_{x\to\infty}(e_d, \tilde{\psi}_d, \tilde{\psi}_q)^T = 0$ as well.

Remark 1. In [14] the persistency of excitation conditions are investigated for similar to (11) structure. It is shown that persistency of excitation conditions are satisfied for all operation modes, excluding direct current excitation mode, when $\omega_0 = 0$. From the exponential stability of (11) with $\Delta\alpha = 0$ it follows that original perturbed system (11) is locally asymptotically stable for bounded rotor resistance variations, when $\Delta\alpha \neq 0$.

B. Flux controller.

Let us define the estimated flux tracking error as $\tilde{\tilde{\psi}} = |\hat{\psi}| - \psi^*$, and rewrite the third equation in (4) in the form

$$\dot{\tilde{\tilde{\psi}}} = -\alpha_n\tilde{\tilde{\psi}} - \alpha_n L_m e_d - \alpha_n\psi^* - \dot{\psi}^* + \alpha_n L_m i_d. \quad (15)$$

For system (15) the following proportional-integral flux controller is designed:

$$\begin{aligned} i_d &= \left(\alpha_n\psi^* + \dot{\psi}^* - k_\psi\tilde{\tilde{\psi}} - x_\psi\right)\big/\alpha_n L_m, \\ \dot{x}_\psi &= k_{\psi i}\tilde{\tilde{\psi}}, \end{aligned} \quad (16)$$

where $k_\psi, k_{\psi i} > 0$ – proportional and integral gains of flux controller.

From (15) and (16), the flux tracking error dynamic can be derived as

$$\begin{aligned} \dot{x}_\psi &= k_{\psi i}\tilde{\tilde{\psi}}, \\ \dot{\tilde{\tilde{\psi}}} &= -(\alpha_n + k_\psi)\tilde{\tilde{\psi}} - \alpha_n L_m e_d - x_\psi. \end{aligned} \quad (17)$$

So far as exponentially stable subsystems (17) and (11) are connected in series, the composite system (11), (17) is exponentially stable as well for any $k_\psi > 0$ and $k_{\psi i} > 0$, i.e. variables $\tilde{\tilde{\psi}}, x_\psi$ converge to zero.

From stability properties of the equilibrium $(e_d, e_q, \tilde{\psi}_d, \tilde{\psi}_q, \tilde{\tilde{\psi}})^T = 0$ it follows: asymptotic estimation of rotor flux vector (objective O.2), because $\lim_{t\to\infty}\left(\tilde{\psi}_d, \tilde{\psi}_q\right) = 0$; asymptotic tracking of the flux reference trajectories, because from condition $\lim_{t\to\infty}\tilde{\tilde{\psi}} = 0$ it follows that $\lim_{t\to\infty}\tilde{\psi} = 0$, since $\tilde{\psi} = \tilde{\psi}_d + \tilde{\tilde{\psi}}$; asymptotic field orientation $\lim_{t\to\infty}\psi_q = 0$.

C. Speed controller

According to [6], nonlinear speed controller is given by

$$\begin{aligned} i_q &= \left(-k_\omega\tilde{\omega} + \hat{T}_L + \dot{\omega}^*\right)\big/\mu\psi^*, \\ \dot{\hat{T}}_L &= -k_{\omega i}\tilde{\omega}, \end{aligned} \quad (18)$$

where $\hat{T}_L$ – integral estimation of constant T_L/J, which is proportional to load torque; $k_\omega, k_{\omega i} > 0$ are speed controller proportional and integral gains.

From (1) and (19) the speed tracking error dynamics becomes

$$\begin{aligned} \dot{\tilde{T}}_L &= k_{\omega i}\tilde{\omega}, \\ \dot{\tilde{\omega}} &= -k_\omega\tilde{\omega} - \tilde{T}_L + \mu\left[(\tilde{\psi}_d + \tilde{\tilde{\psi}})i_q - \tilde{\psi}_q i_d\right], \end{aligned} \quad (19)$$

where $\tilde{T}_L = T_L/J - \hat{T}_L$ – load estimation error.

If $(\tilde{\psi}_d, \tilde{\psi}_q, \tilde{\tilde{\psi}})^T = 0$, then system (19) is linear and asymptotically exponentially stable for any $k_\omega > 0$ and $k_{\omega i} > 0$, i.e. variables $\tilde{\omega}, \tilde{T}_L$ converges to zero, correspondingly control objective O.1 is satisfied.

The complete equations of the proposed controller inclclude: flux observer (4); flux controller (16); speed controller (18); and

d-axis PI current controller

$$u_d = \sigma\left(\gamma_n i_d^* - \omega_0 i_q - \alpha_n \beta \psi^* - k_i \tilde{i}_d - x_d\right),$$
$$\dot{x}_d = k_{ii} \tilde{i}_d; \qquad (20)$$

q-axis PI current controller

$$u_q = \sigma\left(\gamma_n i_q^* + \omega_0 i_d + \beta \omega \psi^* - k_i \tilde{i}_q - x_q\right),$$
$$\dot{x}_q = k_{ii} \tilde{i}_q, \qquad (21)$$

where $\tilde{i}_d = i_d - i_d^*, \tilde{i}_q = i_q - i_q^*$ are current tracking errors, (i_d^*, i_q^*) – current references generated by flux controller (16) and speed controller (18) correspondingly; $k_i, k_{ii} > 0$ are the current controller's proportional and integral gains. The block diagram of the speed-flux control system is shown in Fig.1.

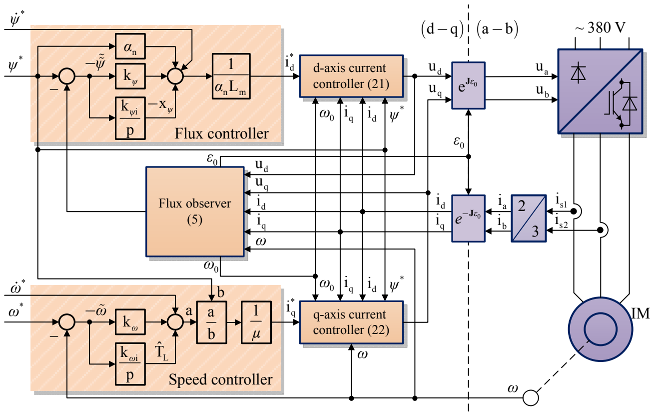


Fig. 1. Controller's block-diagram.

IV. EXPERIMENTAL RESULTS

The experiments were performed in order to compare dynamic performances and efficiency of the two algorithms: proposed direct field oriented control with rotor resistance invariant observer (4) (I-DFOC) and robust indirect field oriented control [6] (R-IFOC).

The experimental tests were carried out using a Rapid Prototyping Station (RPS), whose block diagram is shown in Fig. 2. The RPS includes: a personal computer acting as the operator interface for programming, debugging, virtual oscilloscope function; a custom floating-point Digital Signal Processor board (TMS320F28335); 20A/380V three-phase inverter, f_{PWM}=10 kHz during experiments; two 2.2 kW induction motors, whose rated data are listed in the Appendix. The motor speed is measured by means of a 1024 ppr incremental encoder. The sampling time is set to 200 μs.

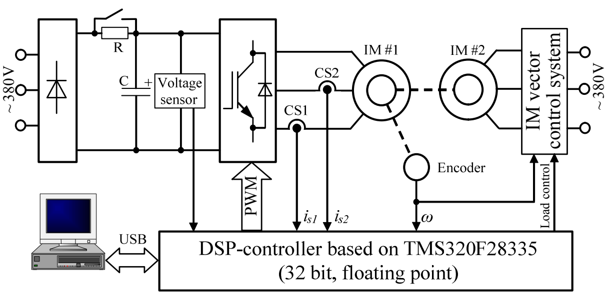


Fig. 2. Experimental set-up of electromechanical system.

The controller's parameters were set at: $k_\omega = 100$, $k_{\omega i} = 5000$; $k_\psi = 100$, $k_{\psi i} = 2500$; $k_i = 700$, $k_{ii} = 245000$; tuning parameters of the flux observer (4) are $\delta = 700$, $k_{ed1} = 0$, $|\hat{\psi}(0)| = 0.02$ Wb.

The operating sequence, reported in Fig. 3, is following: the machine is excited during the initial time interval 0÷0.25 s using a flux reference trajectory starting at 0.02 Wb and reaching the motor rated value of 0.96 Wb; the unloaded motor is required to track the speed reference trajectory, starting at t = 0.6 s from zero initial value and reaching the value of 50 rad/s (33.8 % of rated speed); at time t = 1.2 s a constant load torque, equal the motor rated value, is applied; at time t = 2 s load torque is set to zero.

During tests with rotor resistance variations the rated load torque is applied at time t = 1.2 s and remains constant to the end of test time interval. Rotor resistance variations were introduced in control algorithm, while physical value of R_2 remains unchanged.

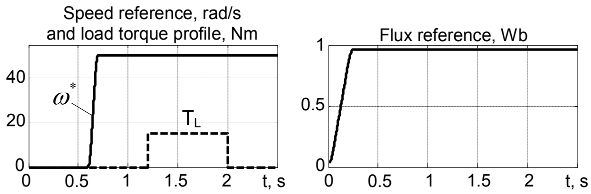


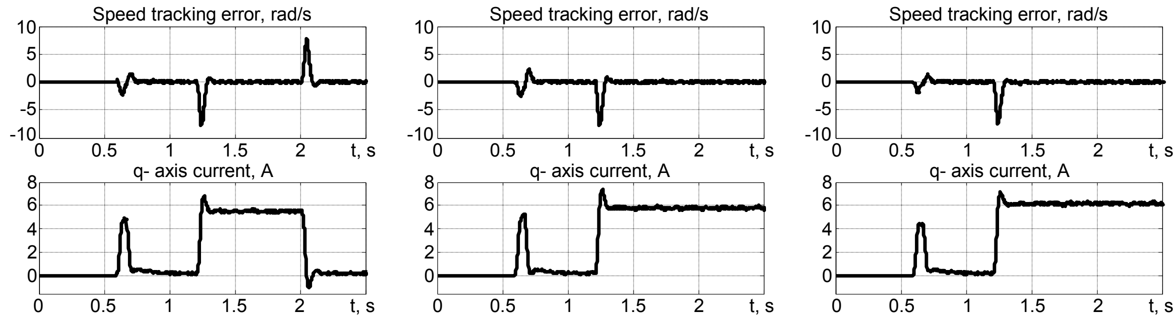
Fig. 3. Speed, flux references and load torque profile.

A first set of experiments, reported in Fig. 4 for R-IFOC and Fig. 5 for I-DFOC, was performed to compare the dynamic performances of the two algorithms when motor operates in medium speed range ($\omega^* = 50$ rad/s). In figures, $\hat{R}_2$ denotes a value of rotor resistance used in control algorithm. As it follows from Fig. 4 and Fig. 5, during operation at medium speed both controllers provides high performance speed trajectory tracking with known R_2 as well as under rotor variation conditions.

Note, that robust properties of R-IFOC are provided by the speed depended correction term $\gamma_2 \beta \omega \tilde{i}_d$ in flux subsystem, where γ_2 is a tuning gain [6]. At low speeds the equivalent gain $\gamma_2 \beta \omega$ is significantly reduced and robustness properties of R-IFOC are strongly reduced as well. This effect has been investigated at the next stage of experiments, when motor operates at low speed $\omega^* = 5$ rad/s (3.3% of rated). Transients for this case are depicted in Fig. 6 and Fig. 7 for R-IFOC and I-DFOC correspondingly.

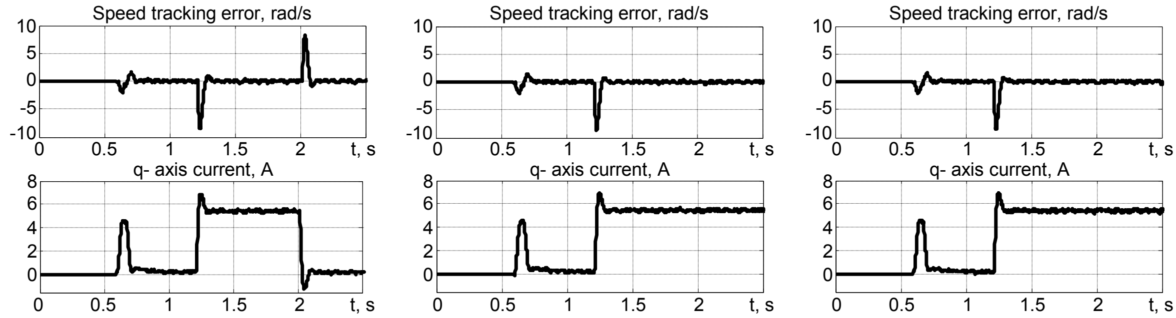
From Fig. 6.b we can conclude that in case $\hat{R}_2 / R_2 = 0.5$ no significant difference is present in speed dynamics as compared with rated transients for $\hat{R}_2 / R_2 = 1$. However higher torque current (up to 40 %) is required to produce the same motor torque. As it is shown in Fig. 6.c the significant degradation of the transient performance occurs for variation $\hat{R}_2 / R_2 = 1.7$. The steady state value of current i_q is almost 45% bigger than in nominal regime.

The results of the same tests for I-DFOC control are reported in Fig. 7. Comparison of the transients, reported in Fig. 7.a, 7.b and 7.c, shows that I-DFOC controller



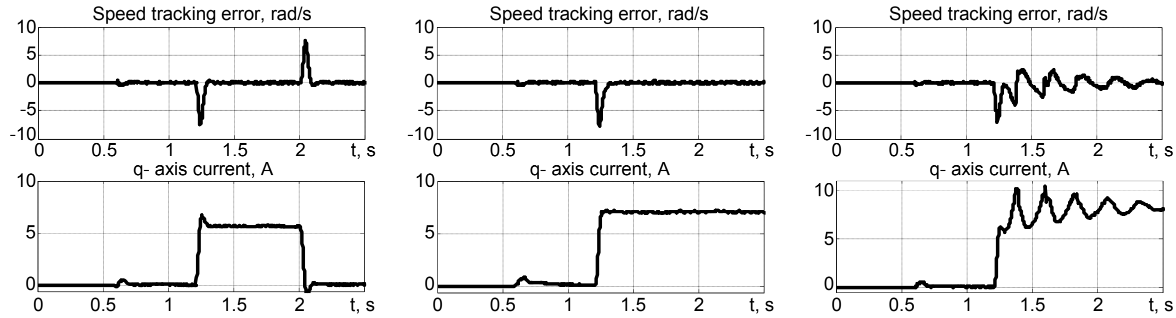
a) $\hat{R}_2/R_2 = 1$ b) $\hat{R}_2/R_2 = 0.5$ c) $\hat{R}_2/R_2 = 1.7$

Fig. 4. Transients for R-IFOC at $\omega^* = 50$ rad/s .



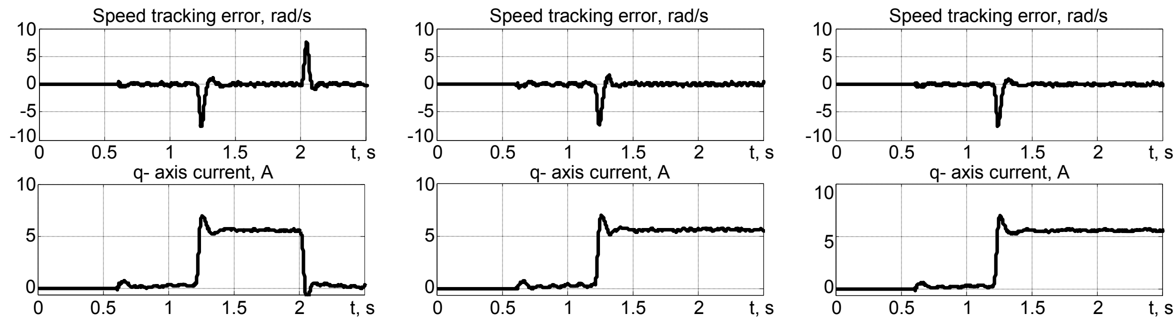
a) $\hat{R}_2/R_2 = 1$ b) $\hat{R}_2/R_2 = 0.5$ c) $\hat{R}_2/R_2 = 1.7$

Fig. 5. Transients for I-DFOC at $\omega^* = 50$ rad/s .



a) $\hat{R}_2/R_2 = 1$ b) $\hat{R}_2/R_2 = 0.5$ c) $\hat{R}_2/R_2 = 1.7$

Fig. 6. Transients for R-IFOC at $\omega^* = 5$ rad/s .



a) $\hat{R}_2/R_2 = 1$ b) $\hat{R}_2/R_2 = 0.5$ c) $\hat{R}_2/R_2 = 1.7$

Fig. 7. Transients for I-DFOC at $\omega^* = 5$ rad/s .

demonstrates invariant properties with respect of rotor resistance variations at low speed as well. No significant difference can be noted in transients for nominal (Fig. 7a) and perturbed (Fig. 7.b and Fig. 7.c) conditions. Behavior of I-DFOC at low speed practically the same as reported in Fig. 5 for medium speed.

A third set of experiments was carried out to compare the system efficiency under steady-state condition of operation. During investigation of steady state performance, a fixed reference speed of 5 rad/s is imposed. During the experiments the i_d (flux current) is set to rated value for R-IFOC, in order to get the nominal rotor flux. The steady-state regulation errors $\tilde{\omega}$ and $\tilde{i}_d$ are negligible for both algorithms.

The experimental test, whose results are reported in Fig. 8, is executed in the following way. The values of constant load torque have been set at:rated torque T_N, $0.75T_N$, $0.5T_N$, fixing the output mechanical power on the level of 75, 56 and 38 W correspondingly. Different values of parameter $\hat{R}_2$ are used in both control algorithms. For each value of $\hat{R}_2$ the steady state i_q current and output inverter active power have been recorded. The experimental results, reported in Fig. 8.a, show that the I-DFOC controller is able to keep almost a constant torque current even with large $\hat{R}_2$ - parameter error. The i_q current imposed by the R-IFOC controller considerably increases when a wrong $\hat{R}_2$ is used. This means that, as compared to the R-IFOC controller, the I-DFOC controller provides the rotor resistance invariant stabilization of the rotor flux vector (both angular position and amplitude). Therefore, in most of the operating conditions, a reduced i_q current is required to compensate for a given load torque, thus considerably increasing the system overall energy efficiency. As it is reported in Fig. 8.b the efficiency of the electromechanical energy conversion process considerably degrades for R-IFOC system: power losses increase up to 45% when $\hat{R}_2/R_2 = 0.5$ and increase up to 80% when $\hat{R}_2/R_2 = 1.7$. In the same cases I-DFOC system guarantees the robust stabilization of the efficiency approximately on the rated level.

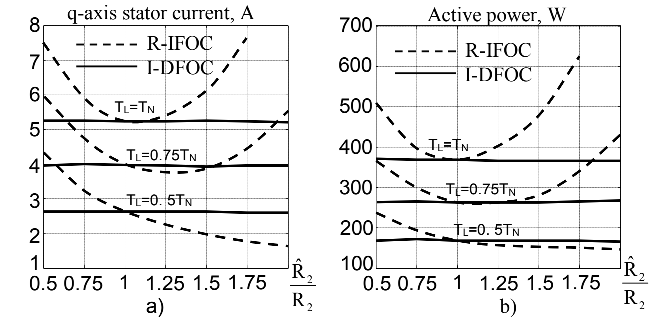


Fig. 8. Results of efficiency investigation.

Nevertheless, from Fig. 8 it follows, that at light loads, rotor resistance variations in R-IFOC tend to improve system efficiency when $\hat{R}_2 > R_2$. Such behaviour of the R-IFOC system can be explained in terms of maximum torque per Ampere (MTPA) optimization. It is known that, in order to achieve MTPA optimization at light loads, the flux level should be reduced. Since condition $\hat{R}_2 > R_2$ leads to rotor flux reduction [7], the MTPA optimization effect occurs at light loads. However for I-DFOC the energy efficient control strategies can be applied more effectively with robustly controlled flux vector.

V. Conclusions

A novel speed-flux tracking controller for induction motors has been developed and experimentally verified. Direct rotor flux field oriented controller is based on full order hybrid continuous time-sliding mode flux observer. Controller guarantees local asymptotic speed-flux tracking and asymptotic direct field-orientation under condition of unknown constant load torque. The flux subsystem is invariant with respect to limited rotor resistance variations due to special structure of the flux observer. The high performance speed tracking and efficiency of the proposed solution are confirmed by the results of experimental studies. It is shown by experiment, that in contrast to existing robust controllers, developed one provides an improved robustness properties with respect to rotor resistance variations in all motor operating conditions including nearby zero speeds.

Appendix

IM parameters: rated power 2.2 kW; rated speed 1410 rpm; rated torque 15 Nm; number of poles $2p = 4$; rated current 5 A_{RMS}; rated voltage 380 V_{RMS}; $R_1 = 4.1$ Ω; $R_2 = 1.975$ Ω; $L_m = 0.2515$ H; $L_1 = 0.264$ H; $L_2 = 0.264$ H; $J = 0.016$ kgm^2.

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