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Direct speed-flux vector control of induction motors: controller design and experimental robustness evaluation

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ABSTRACT

The paper presents a general theoretical solution of the induction motor speed-flux direct vector control, which is based on the second Lyapunov method application. The structure of the direct vector control algorithm involves the use of any asymptotic flux observer with exponential stability properties. Such approach led to improvement of the vector control system robustness properties. A constructive procedure for the design of correction terms of the rotor flux observer is proposed. Designed family of the flux observers guarantees the exponential stability and robustification with respect of parametric disturbances. It is shown that the proposed solution guarantees: global exponential tracking of the speed-flux reference trajectories together with asymptotic field orientation, asymptotic exponential estimation of the rotor flux, as well as asymptotic decoupling of torque (speed) and flux control. Comparative experimental study shows that new controller provides stabilization of the control performances as well as efficiency at nominal level when the rotor active resistance changes. The proposed direct vector control structures can be used for the development of energy-efficient high performance induction motor drives for metalworking, packaging equipment, modern electric transport and special equipment.

Keywords: Induction motor; robust vector control; flux observer; software implementation of control algorithms

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INTRODUCTION

Vector controlled Induction Motors (IM) are widely used in industry, transportation and other fields, where high dynamic performance and accuracy of speed tracking is required [1, 2], [3, 4], [5, 6]. The concept of field orientation [7], [8] in vector-controlled drives is a basic principle of IM control and can be achieved using direct and indirect methods. The modern theory of induction motor control has been developed in 1990-th and 2000-th and was summarized in excellent monograph [9] (see references overview in [9]). Both indirect and direct field-oriented controllers have been implemented in industrial drives demonstrating performances suitable for a wide spectrum of applications. Accurate field orientation and high-performance control of IM is only achieved if all electrical parameters on induction motor model are exactly known. Parameters uncertainties or variations in classical vector control schemes lead to performance and efficiency degradation and define the well-known and the most significant problem of IM vector control [9].

LITERATURE REVIEW

The field-orientation concept is the model-based approach; all IM vector control algorithms use five electrical motor parameters – stator and rotor active resistances, leakage inductances and magnetizing inductance [4, 5], [9]. During motor operation the only leakage inductances can be considered as constant parameters [2, 9]. Stator and especially rotor resistance can vary significantly due to motor heating. Magnetizing inductance due to presence of real magnetizing curve also not constant when motor flux is changing [10]. The most critical parameter, which has significant influence on control performances and efficiency, is rotor resistance. Due to motor heating, this parameter can vary up to 200 % during motor operation leading to drive performance and efficiency degradation [4, 5], [9]. The problem of the sensitivity to rotor resistance variations is not a new one and well known starting from the 1980-th [11, 12], [13, 14], [15], quite before the modern theory of nonlinear feedback control was applied for control of electric machines.

Different approaches are used to overcome this problem. The most practically reasonable are robust vector control schemes, due to their simplicity. In papers [16], [17] authors present technology of high-performance indirect field-oriented control, which provide not only high-performance flux-speed tracking, but also high robustness properties confirmed by results of intensive experimental testing. However, despite the simplicity of indirect control schemes, they have some disadvantages. First of all, there are no degrees of freedom for robustification terms design and robustness properties are weak at near zero speed. The second disadvantage appears when we consider flux weakening region of operation. In this case no information about rotor flux due to absence of measurement or estimation. In contrast to indirect control scheme, the direct field-oriented controllers use flux observer, which provides information about rotor flux vector and define the robustness properties of the control algorithms [11, 12], [13, 14], [15]. The large numbers of possible observer's structures have been proposed with some degrees of freedom for correction terms design.

An alternative approach to solve the problem of IM parameters variations is application of adaptive controllers [4], [5], [9], [18]. Nevertheless, existing adaptive controllers are quite complex, can be sensitive to variation of non-adapting parameters, require high quality inverter and measurement subsystem and hardly implementable in real industrial drives.

The main aim of this paper is to present the modified design of the generalized direct field-oriented control algorithm [19], which leads to simplification of the controller structure. The family of the flux observers for IM electromagnetic subsystem allows selecting the observer's feedbacks to guarantee not only asymptotic stability but robustness properties as well. Comparative experimental study of the dynamic and robustness properties of the three control schemes: standard and improved indirect vector controls [16], [17] and developed direct speed-flux controller have been performed using concept of rapid prototyping [20] for both hardware and software parts of experimental installation.

Induction motor model

The equivalent two-phase model of a symmetric IM in synchronous (d-q) reference frame, rotating at an arbitrary angular speed ω_0, is given by [1]

$$\begin{aligned}
&\dot{\omega} = \mu\left(\psi_{2d} i_{1q} - \psi_{2q} i_{1d}\right) - \frac{T_L}{J} - \nu\omega,\\
&\dot{i}_{1d} = -\gamma i_{1d} + \omega_0 i_{1q} + \alpha\beta\psi_{2d} + \beta p_n \omega \psi_{2q} + \frac{1}{\sigma} u_{1d},\\
&\dot{i}_{1q} = -\gamma i_{1q} - \omega_0 i_{1d} + \alpha\beta\psi_{2q} - \beta p_n \omega \psi_{2d} + \frac{1}{\sigma} u_{1q},\\
&\dot{\psi}_{2d} = -\alpha\psi_{2d} + \omega_2 \psi_{2q} + \alpha L_m i_{1d},\\
&\dot{\psi}_{2q} = -\alpha\psi_{2q} - \omega_2 \psi_{2d} + \alpha L_m i_{1q},\\
&\dot{\varepsilon}_0 = \omega_0, \varepsilon_0(0) = 0,
\end{aligned} \tag{1}$$

where: $\mathbf{u}_1 = \left(u_{1d}, u_{1q}\right)^T$, $\mathbf{i}_1 = \left(i_{1d}, i_{1q}\right)^T$ are vectors of stator voltages and currents, $\boldsymbol{\psi}_2 = \left(\psi_{2d}, \psi_{2q}\right)^T$ is vector of rotor flux, ω is the rotor angular speed, T_L is load torque, $\omega_2 = \omega_0 - p_n\omega$ is slip frequency, p_n is the number of pole pairs , ε_0 is the angular position of the synchronous reference frame (d-q) with respect to the stationary reference frame (a-b).

The constant parameters in (1) associated with the electrical and mechanical parameters of the IM are defined as follows:

$$\alpha = \frac{R_2}{L_2}, \gamma = \frac{R_1}{\sigma} + \alpha L_m \beta, \sigma = L_1\left(1 - \frac{L_m^2}{L_1 L_2}\right),$$

$$\beta = \frac{L_m}{L_2 \sigma}, \mu = \frac{1}{J}\frac{3}{2} p_n \frac{L_m}{L_2},$$

where R_1, R_2 are active stator and rotor resistances; L_1, L_2 are stator and rotor inductances, L_m is the magnetizing inductance, J is the total moment of inertia, $\nu = \frac{\nu_1}{J}; \nu_1 > 0$ is the viscous friction coefficient.

The transformed variables in (1) are given by

$$\begin{aligned}
&x^{(d-q)} = e^{-\mathbf{J}\varepsilon_0} x^{(a-b)}\\
&x^{(a-b)} = e^{\mathbf{J}\varepsilon_0} x^{(d-q)},\\
&e^{-\mathbf{J}\varepsilon_0} = \begin{bmatrix} \cos\varepsilon_0 & \sin\varepsilon_0 \\ -\sin\varepsilon_0 & \cos\varepsilon_0 \end{bmatrix},
\end{aligned} \tag{2}$$

where $x^{(y-z)}$ defines two-dimensional vectors of voltages, currents and fluxes.

In direct field orientated control the angular position ε_0 of synchronous reference frame (d-q) is provided by the flux observer.

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Control problem formulation

Let us define a vector of controlled variables

$$\mathbf{y}=\begin{bmatrix}\omega \\ \left(\psi_{2d}^2+\psi_{2q}^2\right)^{\frac{1}{2}}\end{bmatrix} \equiv \begin{bmatrix}\omega \\ |\psi_2|\end{bmatrix}, \tag{3}$$

and reference trajectories ω^* and $\psi^*>0$ for motor speed and rotor flux vector magnitude. The speed and flux tracking errors are defined as

$$\tilde{\omega}=\omega-\omega^*,\ \tilde{\psi}=|\psi_2|-\psi^*. \tag{4}$$

Assume that:

A.1. The angular speed ω and stator currents i_{1d}, i_{1q} are available for measurement.

A.2. Reference trajectories ω^* and $\psi^*>0$ are bounded functions with bounded first and second time derivatives.

A.3. Load torque T_L is unknown but constant.

A.4. The parameters of the IM model are known and constant.

Under these assumptions it is necessary to design the vector of control voltages $\mathbf{u}_1=\left(u_{1d}, u_{1q}\right)^T$ in (1), which guarantees that following control objectives are satisfied:

O.1. Global asymptotic tracking of output variables, i.e.

$$\lim_{t\to\infty}\tilde{\omega}=0,\ \lim_{t\to\infty}\tilde{\psi}=0. \tag{5}$$

Together with asymptotic field orientation, given by the condition

$$\lim_{t\to\infty}\psi_{2q}=0. \tag{6}$$

O.2. Asymptotic linearization of the speed control subsystem.

O.3. Asymptotic decoupling of the speed control and the rotor flux control subsystems.

O.4. Asymptotic estimation of the rotor flux vector.

Design of the flux control subsystem

First, we define the family of flux observers for the electrical subsystem of IM in the following form:

$$\begin{aligned}
\dot{\hat{i}}_{1d}&=-\gamma\hat{i}_{1d}+\omega_0\hat{i}_{1q}+\alpha\beta|\hat{\psi}_2|+\frac{1}{\sigma}u_{1d}+v_{1d},\\
\dot{\hat{i}}_{1q}&=-\gamma\hat{i}_{1q}-\omega_0\hat{i}_{1d}-\beta p_n\omega|\hat{\psi}_2|+\frac{1}{\sigma}u_{1q}+v_{1q},\\
|\dot{\hat{\psi}}_2|&=-\alpha|\hat{\psi}_2|+\alpha L_m i_{1d}+v_{2d},\\
\dot{\varepsilon}_0&=\omega_0=p_n\omega+\alpha L_m\frac{i_{1q}}{|\hat{\psi}_2|}+\frac{1}{|\hat{\psi}_2|}v_{2q},\ |\hat{\psi}_2|>0,
\end{aligned} \tag{7}$$

where $\hat{i}_{1d}, \hat{i}_{1q}$ estimates of the stator currents i_{1d}, i_{1q}, $|\hat{\psi}_2|$ estimate of the rotor flux vector magnitude; correction terms $v_{1d}, v_{1q}, v_{2d}, v_{2q}$ will be designed later.

Note that the general form of the observer (7) corresponds to the generalized Verghese observer [14], represented in the (d-q) reference frame with ε_0 in (2) according to the last equation in (7).

Define the observer estimation errors as follows

$$\begin{aligned}
&e_d=i_{1d}-\hat{i}_{1d};\ e_q=i_{1q}-\hat{i}_{1q};\\
&\tilde{\psi}_{2d}=\psi_{2d}-|\hat{\psi}_2|;\ \tilde{\psi}_{2q}=\psi_{2q}.
\end{aligned} \tag{8}$$

Taking into account these definitions, from equations (1) and (7) we obtain the equations for the dynamics of estimation errors

$$\begin{aligned}
\dot{e}_d&=-\gamma e_d+\omega_0 e_q+\alpha\beta\tilde{\psi}_{2d}+\beta p_n\omega\tilde{\psi}_{2q}-v_{1d}\\
\dot{e}_d&=-\gamma e_q-\omega_0 e_d+\alpha\beta\tilde{\psi}_{2q}-\beta p_n\omega\tilde{\psi}_{2d}-v_{1q}\\
\dot{\tilde{\psi}}_{2d}&=-\alpha\tilde{\psi}_{2d}+\omega_2\tilde{\psi}_{2q}-v_{2d}\\
\dot{\tilde{\psi}}_{2q}&=-\alpha\tilde{\psi}_{2q}-\omega_2\tilde{\psi}_{2d}-v_{2q}
\end{aligned} \tag{9}$$

It can be shown that the open-loop observer (7) with $v_{1d}=v_{1q}=v_{2d}=v_{2q}=0$ is globally exponentially asymptotically stable.

Note that the correction signals in (9) can only be the functions of the measured IM variables.

The next step is to design a control algorithm for the estimated rotor flux magnitude. Let define the estimated flux tracking error and excitation current tracking error as

$$\tilde{\tilde{\psi}}=|\hat{\psi}_2|-\psi^*;\ \tilde{i}_{1d}=i_{1d}-i_{1d}^*. \tag{10}$$

According to definition (10), third equation from (7) and the second equation from (1) we obtain the following error dynamics:

$$\begin{aligned}&\dot{\tilde{\psi}}=-\alpha\tilde{\psi}+\alpha L_m\tilde{i}_{1d}+\alpha L_m i_{1d}^*-\dot{\psi}^*-\alpha\psi^*,\\&\dot{\tilde{i}}_{1d}=-\gamma\tilde{i}_{1d}+\omega_0 i_{1q}+\alpha\beta\tilde{\psi}_{2d}+\\&+\beta p_n\omega\tilde{\psi}_{2q}-\gamma i_{1d}^*+\alpha\beta\left|\hat{\psi}_2\right|-\dot{i}_{1d}^*+\frac{1}{\sigma}u_{1d}.\end{aligned}\tag{11}$$

For system (11) we design:

- proportional flux controller

$$i_{1d}^*=\frac{1}{\alpha L_m}\left(\alpha\psi^*+\dot{\psi}^*-k_\psi\tilde{\psi}\right),k_\psi>0;\tag{12}$$

- non-linear current controller

$$u_{1d}=\sigma\begin{pmatrix}\gamma i_{1d}^*-\omega_0 i_{1q}-\alpha\beta\left|\hat{\psi}_2\right|+\\+\dot{i}_{1d}^*-k_{id1}\tilde{i}_{1d}+v_d\end{pmatrix},\tag{13}$$

$$k_{id1}>0.$$

After substituting (12) and (13) into (11), we can write complete equations for the dynamics IM flux subsystem

$$\begin{aligned}&\begin{pmatrix}\dot{e}_d\\\dot{e}_q\\\dot{\tilde{\psi}}_{2d}\\\dot{\tilde{\psi}}_{2q}\\\dot{\tilde{\psi}}\\\dot{\tilde{i}}_{1d}\end{pmatrix}=\begin{pmatrix}\mathbf{A}_{20\_1}&\mathbf{A}_{20\_2}\\\mathbf{A}_{20\_3}&\mathbf{A}_{20\_4}\end{pmatrix}\begin{pmatrix}e_d\\e_q\\\tilde{\psi}_{2d}\\\tilde{\psi}_{2q}\\\tilde{\psi}\\\tilde{i}_{1d}\end{pmatrix}+\\&+\mathbf{Bv}\triangleq\mathbf{A}_{20}\mathbf{x}_2+\mathbf{Bv},\\&\mathbf{y}_2=C\mathbf{x}_2.\end{aligned}\tag{14}$$

where $\mathbf{x}_2=\left(e_d,e_q,\tilde{\psi}_{2d},\tilde{\psi}_{2q},\tilde{\psi},\tilde{i}_{1d}\right)^T$,

$\mathbf{v}=\left(v_{1d},v_{1q},v_{2d},v_{2q}\right)^T,\mathbf{y}_2=\left(e_d,e_q,\tilde{\psi},\tilde{i}_{1d}\right)^T$,

$k_{id}=\gamma+k_{id1}$,

$$\mathbf{A}_{20\_1}=\begin{bmatrix}-\gamma&\omega_0&\alpha\beta&\omega p_n\beta\\-\omega_0&-\gamma&-\omega p_n\beta&\alpha\beta\\0&0&-\alpha&\omega_2\\0&0&-\omega_2&-\alpha\end{bmatrix},$$

$$\mathbf{A}_{20\_2}=\begin{bmatrix}0&0\\0&0\\0&0\\0&0\end{bmatrix},\ \mathbf{A}_{20\_3}=\begin{bmatrix}0&0&0&0\\0&0&\alpha\beta&\omega p_n\beta\end{bmatrix},$$

$$\mathbf{A}_{20\_4}=\begin{bmatrix}-\left(\alpha+k_\psi\right)&\alpha L_m\\0&-k_{id}\end{bmatrix}.$$

Design of correction vector $\mathbf{v}$ in (14) can be carried out in the standard way for linear nonstationary control systems. Consider the Lyapunov function in the form

$$V=\frac{1}{2}\mathbf{x}_2^T\mathbf{P}\mathbf{x}_2,\tag{15}$$

where $\mathbf{P}=\mathbf{P}^T>0$.

The derivative of (15) along trajectories (14) has the form

$$\dot{V}=\frac{1}{2}\mathbf{x}_2^T\left(\mathbf{A}(t)_{20}^T\mathbf{P}+\mathbf{PA}(t)_{20}^T\right)\mathbf{x}_2+\mathbf{x}_2^T\mathbf{PBv}.\tag{16}$$

Designing $\mathbf{v}=\mathbf{\Gamma}(t)C\mathbf{x}_2$, such that

$$\frac{1}{2}\left(\mathbf{A}(t)_{20}^T\mathbf{P}+\mathbf{PA}(t)_{20}^T\right)+\mathbf{PB\Gamma}(t)C=-\mathbf{Q},\quad \mathbf{Q}=\mathbf{Q}^T>0,\tag{17}$$

we get

$$\dot{V}=-\mathbf{x}_2^T\mathbf{Q}\mathbf{x}_2,\tag{18}$$

which guarantees the global exponential stability of the equilibrium point $\mathbf{x}_2=0$.

The structure of the system (14) gives the freedom in design of the correction vector $\mathbf{v}$ in order to obtain not only stability, but also a specification for the estimation and control errors convergence, as well as robustness with respect to parametric disturbances. This is achieved by choosing different matrix configurations $\mathbf{P}$ in the expression for the Lyapunov function (15). For example, when $\mathbf{P}=\mathrm{diag}\left(\frac{1}{\beta},\frac{1}{\beta},1,1,1,\gamma_1\right)$ with vector of correction terms

$$\mathbf{v}=\begin{bmatrix}k_1&0&\alpha\beta&0\\0&k_1&-p_n\omega\beta&0\\\alpha&-p_n\omega&0&\gamma_1\alpha\beta\\-p_n\omega&\alpha&0&\gamma_1 p_n\omega\beta\\0&0&w_{53}&0\end{bmatrix}\begin{pmatrix}e_d\\e_q\\\tilde{\psi}\\\tilde{i}_{1d}\end{pmatrix},\tag{19}$$

where $w_{53}=-\left(\frac{\alpha L_m}{\gamma_1}+\alpha\beta\right)$, we obtain the following solution of the Lyapunov equation

$$\mathbf{Q}=\mathrm{diag}\begin{bmatrix}\beta^{-1}\left(\gamma+k_i\right),\beta^{-1}\left(\gamma+k_i\right),\\\alpha,\alpha,\left(k_\psi+\alpha\right),\gamma_1 k_{id}\end{bmatrix}>0.\tag{20}$$

In this case, the error dynamics is given by the equation

$$\dot{\mathbf{x}}_2 = \mathbf{A}_{21}(t)\mathbf{x}_2, \tag{21}$$

with matrix

$$\mathbf{A}_{21} = \begin{bmatrix} \mathbf{A}_{21\_1} & \mathbf{A}_{21\_2} \\ \mathbf{A}_{21\_3} & \mathbf{A}_{21\_4} \end{bmatrix},$$

$$\mathbf{A}_{21\_1} = \begin{bmatrix} -(\gamma + k_i) & \omega_0 & \alpha\beta \\ -\omega_0 & -(\gamma + k_i) & -p_n\omega\beta \\ -\alpha & p_n\omega & -\alpha \end{bmatrix},$$

$$\mathbf{A}_{21\_2} = \begin{bmatrix} p_n\omega\beta & -\alpha\beta & 0 \\ \alpha\beta & p_n\omega\beta & 0 \\ \omega_2 & 0 & -\gamma_1\alpha\beta \end{bmatrix},$$

$$\mathbf{A}_{21\_3} = \begin{bmatrix} -p_n\omega & -\alpha & -\omega_0 \\ \alpha & -p_n\omega & 0 \\ 0 & 0 & \alpha\beta \end{bmatrix}, \tag{22}$$

$$\mathbf{A}_{21\_4} = \begin{bmatrix} -\alpha & 0 & -\gamma_1 p_n\omega\beta \\ 0 & -(\alpha + k_\psi) & -\gamma_1 w_{53} \\ p_n\omega\beta & w_{53} & -k_{id} \end{bmatrix}.$$

The structure of the closed loop flux subsystem (21) allows to extend the flux and current controllers (12), (13) adding the integral action components x_ψ and xd.

The complete equations of the rotor flux vector control algorithm are defined by (12), (13), (7), (19) in the following form:

– flux controller

$$i_{1d}^* = \frac{1}{\alpha L_m}\left(\alpha\psi^* + \dot{\psi}^* - k_\psi\tilde{\tilde{\psi}} - x_\psi\right),$$

$$\dot{x}_\psi = k_{\psi i}\tilde{\tilde{\psi}}; \tag{23}$$

– d-axis current controller

$$u_{1d} = \sigma\begin{bmatrix} \gamma i_{1d}^* - \omega_0 i_{1q} - \alpha\beta|\hat{\psi}_2| + \dot{i}_{1d}^* - \\ -k_{id1}\tilde{i}_{1d} - \left(\gamma_1^{-1}\alpha L_m + \alpha\beta\right)\tilde{\tilde{\psi}} - xd \end{bmatrix},$$

$$\dot{x}_d = k_{ii}\tilde{i}_{1d},$$

$$\dot{i}_{1d}^* = \frac{1}{\alpha L_m}\begin{Bmatrix} \alpha\dot{\psi}^* + \ddot{\psi}^* - \\ -k_\psi\begin{bmatrix} -(\alpha + k_\psi)\tilde{\psi} + \alpha e_d - \\ -p_n\omega e_q + (\alpha L_m + \gamma_1\alpha\beta)\tilde{i}_{1d} \end{bmatrix} \end{Bmatrix} \tag{24}$$

– flux observer

$$\dot{\hat{i}}_{1d} = -\gamma\hat{i}_{1d} + \omega_0\hat{i}_{1q} + \alpha\beta|\hat{\psi}_2| + \frac{1}{\sigma}u_{1d} + k_1 e_d + \alpha\beta\tilde{\tilde{\psi}},$$

$$\dot{\hat{i}}_{1q} = -\gamma\hat{i}_{1q} - \omega_0\hat{i}_{1d} - \beta p_n\omega|\hat{\psi}_2| +$$

$$+\frac{1}{\sigma}u_{1q} + k_1 e_q - \beta p_n\omega\tilde{\tilde{\psi}},$$

$$|\dot{\hat{\psi}}_2| = -\alpha|\hat{\psi}_2| + \alpha L_m i_{1d} - \alpha e_d - p_n\omega e_q + \gamma_1\alpha\beta\,\tilde{i}_{1d}, \tag{25}$$

$$\dot{\varepsilon}_0 = \omega_0 = p_n\omega + \alpha L_m\frac{i_{1q}}{|\hat{\psi}_2|} +$$

$$+\frac{1}{|\hat{\psi}_2|}\left(\alpha e_q + p_n\omega e_d - \gamma_1\beta p_n\omega\tilde{i}_{1d}\right),$$

where $k_{\psi i} > 0, k_{ii} > 0$ are the coefficients of the integral components of the flux and current controllers [19]. Note that flux and current controllers are extended with components of the integral action according to analysis [19].

Flux controller for robust indirect field orientation

According to [16], the complete equations of the rotor flux vector control algorithm for indirect field orientation can be obtained as:

– flux controller

$$\dot{\varepsilon}_0 = \omega_0 = \omega + \alpha L_m\frac{i_{1q}}{\psi^*} + \frac{\lambda}{\psi^*}\beta\omega\tilde{i}_{1d}, \qquad \lambda > 0,$$

$$i_{1d}^* = \frac{1}{\alpha L_m}\left(\alpha\psi^* + \dot{\psi}^*\right); \tag{26}$$

– flux subsystem current controller

$$u_{1d} = \sigma\left[-k_{id}\tilde{i}_{1d} + \gamma i_{1d}^* - \omega_0 i_{1q} - \alpha\beta\psi^* + \dot{i}_{1d}^*\right],$$

$$\dot{i}_{1d}^* = \frac{1}{\alpha L_m}\left(\alpha\dot{\psi}^* + \ddot{\psi}^*\right), \tag{27}$$

where λ is the flux subsystem tuning parameter.

From comparison of flux controller's equations (23)-(25) for direct control scheme and (26)-(27) for indirect one can conclude, that indirect flux control algorithm is much simpler. The only one flux subsystem correction term $\frac{\lambda}{\psi^*}\beta\omega\tilde{i}_{1d}$ is used in equation (26).

Speed control subsystem

Since the both flux control algorithms for the IM guarantee global asymptotic exponential stability, the result obtained in [16] for the case of indirect vector control of IM can be directly used to control the angular speed also in direct flux orientation scheme.

Speed controller's equations according to [16] are defined as

– non-linear speed controller

$$\begin{aligned} & i_{1q}^* = \frac{1}{\mu\psi^*}\left(-k_\omega\tilde{\omega} + \hat{T}_L + \dot{\omega}^* + \nu\omega^*\right), \\ & \dot{\hat{T}}_L = -k_{\omega i}\tilde{\omega}; \end{aligned} \tag{28}$$

– non-linear q-axis current controller

$$\begin{aligned} & u_q = \sigma\left[\gamma i_{1q}^* + \omega_0 i_{1d} + \beta\omega\psi^* + i_{1q1}^* - k_{iq1}\tilde{i}_{1q} + x_q\right], \\ & x_q = -k_{ii}\tilde{i}_{1q}, \\ & i_{1q1}^* = \frac{1}{\mu\psi^*}\begin{bmatrix} -k_\omega\left(-k_\omega\omega + \mu\psi^*\tilde{i}_{1q}\right) + \\ +\dot{M}_c + \ddot{\omega}^* + \nu\dot{\omega}^* \end{bmatrix} - \frac{\dot{\psi}^*}{\psi^*} i_{1q}^*, \end{aligned} \tag{29}$$

where: $\hat{T}_L$ estimate of $T_L / J = \mathrm{const}$, $\tilde{i}_{1q} = i_{1q} - i_{1q}^*$, $\left(k_\omega, k_{\omega i}\right) > 0$, $\left(k_{iq1}, k_{\eta i}\right) > 0$ are the coefficients of the proportional and integral components of the speed controller and the modified current controller.

Under control algorithm (28), (29) the speed subsystem error dynamics are identical to those defined in [16], when changing the variable $\tilde{\psi}_d$ on the $\left(\tilde{\tilde{\psi}} + \tilde{\psi}_{2d}\right)$. A variable $\tilde{\psi}_d$ in the case of indirect vector control [16], and the variable $\left(\tilde{\tilde{\psi}} + \tilde{\psi}_{2d}\right)$ in the direct vector control, decay exponentially to zero regardless of the IM electromechanical subsystem. Therefore, using the result of [16] we establish that the asymptotic speed reference trajectory tracking is achieved, that is $\lim_{t\to\infty}\tilde{\omega} = 0$, and the control objectives specified by conditions O.2 and O.3 are also achieved.

Block-diagram of direct field-oriented flux-speed vector control algorithm is shown in Fig.1.

Experimental study

A. Experimental set-up

The experiments are carried out using the Rapid Prototyping Station (RPS).

As shown in Fig. 2, the RPS includes: (1) Induction motor with a current controlled loading DC machine; (2) 20 A and 380 V three-phase PWM controlled inverter, operated at 10 kHz switching frequency; (3) Digital Signal Processor (DSP) TMS320F28335 controller which performs data acquisition, implements control algorithms with programmable tracing of selected variables; (4) Personal computer for processing, programming, interactive oscilloscope, data acquisition, etc. A 1000 pulse/revolution optical encoder measures the motor speed; the sampling time is set at 200 μsec. The motor parameters are the following: motor type 4AO80B2, rated power 0.75 kW, rated current 1.7 A, $\cos\phi = 0.86$, rated speed 300 rad/s, $R_1 = 11\,\mathrm{Ohm}$, $R_2 = 5.51\mathrm{Ohm}$, $L_1 = L_2 = 0.95\mathrm{H}$, $L_m = 0.91\mathrm{H}$, $J = 0.003\,\mathrm{kg}\cdot\mathrm{m}^2$.

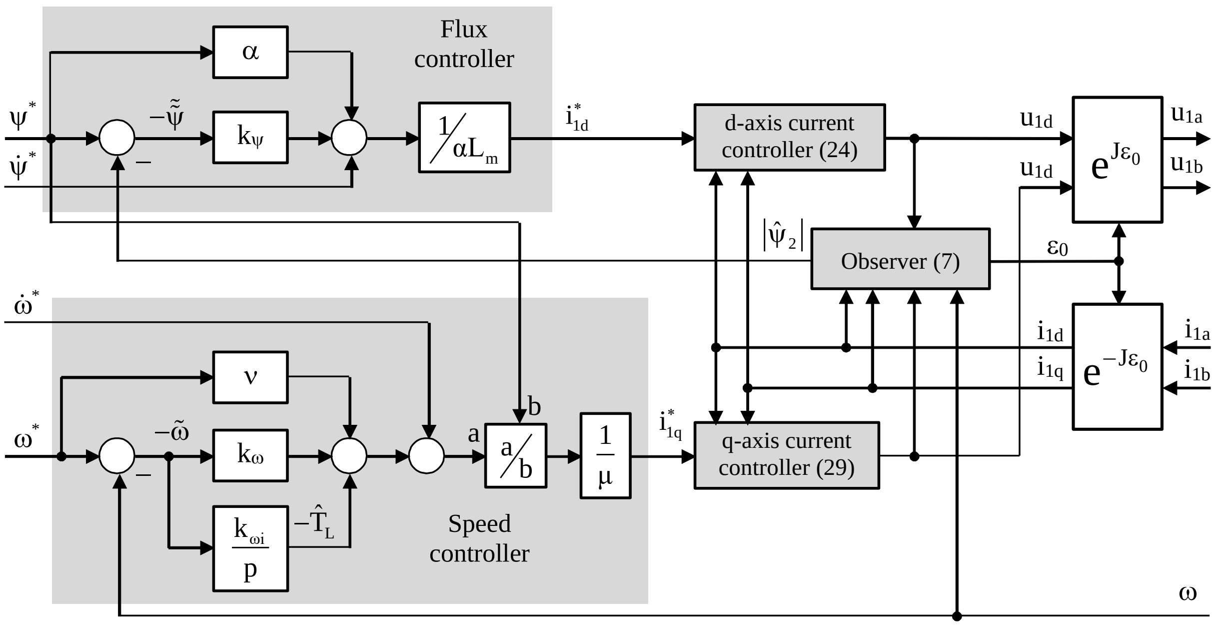

Fig. 1. **Block diagram of the direct vector control algorithm**

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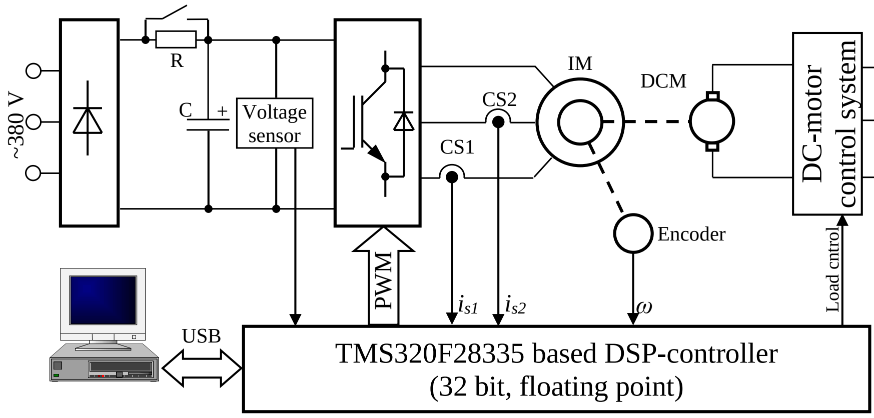

Fig. 2. **Experimental setup**
Source: **compiled by the authors**

During all tests tuning parameters for the controllers are set to: $k_{id1} = k_{iq1} = 700$, $k_{ii} = k_{iq1}^2/4$, $k_{\omega} = 150$, $k_{\omega i} = k_{\omega i}^2/2$, $\lambda = 0.1$, $\gamma_1 = 0.001$, $k_1 = 500$. Note that tuning parameters for current and speed controllers are the same for all control algorism under the test.

B. Operating sequences

The operating sequences, reported in Fig.3, are the following:

– the machine is excited during the initial time interval 0÷0.25s using a flux reference trajectory starting at $\psi^*(0) = 0.02\text{Wb}$ and reaching the motor rated value of 0.9 Wb;

– the unloaded motor is required to track the speed reference trajectory, starting at t = 0.6s from zero initial value and reaching the speed of 50 rad/s;

– at time t = 1.2 s speed reference reversal is applied;

– motor braking to zero speed started from t = 1.8 s;

– during time intervals 0.8-1 s and 1.45-1.65 constant rated torque is applied;

Note that during time interval 1.45-1.65 IM operates in regenerative mode. Tracking of the adopted speed reference trajectory requires a dynamic torque that is equal to the rated value of the IM. Flux and speed reference trajectories are presented in Fig.3 using solid lines; dashed line represents the load torque profile.

In order to study the robustness properties of the different types of commonly used control approaches we test three control schemes: indirect robust controller [16] (Robust Indirect Field Oriented Control – RIFOC) (26),(27), clasical non-robust version of RIFOC – (IFOC) and proposed direct field oriented control (DFOC). From intensive experimental investigation we conclude, that when motor papameters are known, all three control schems provide the same speed-flux control performances, as reported in Fig. 4. From speed tracking error transient in Fig. 4 it follows that asymptotic speed reference trajectory tracking is achieved (speed errors is at zero level when speed reference is changed). Dynamic speed error appears only when constant load torque is applied or removed.

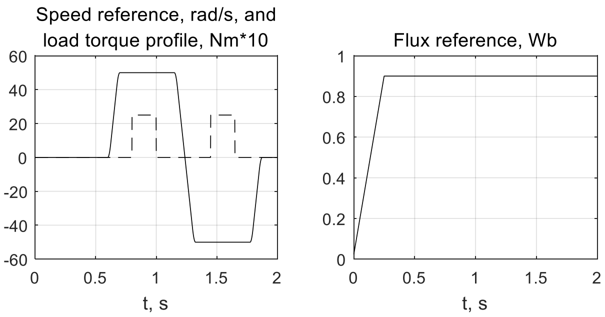

Fig. 3. **Speed-flux reference trajectories and load torque profile**
Source: **compiled by the authors**

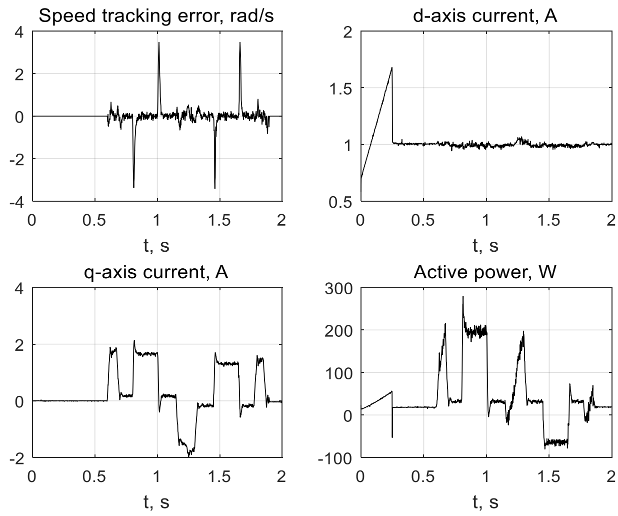

Fig. 4. **Transients of speed, current and active power**
Source: **compiled by the authors**

C. Experimental results

Additional transients for DFOC is reported in Fig.5, they confirm that observer (25) toogether with the flux controller (23) provide asymptotic current-flux estimation and estimated flux vector modulus tracking.

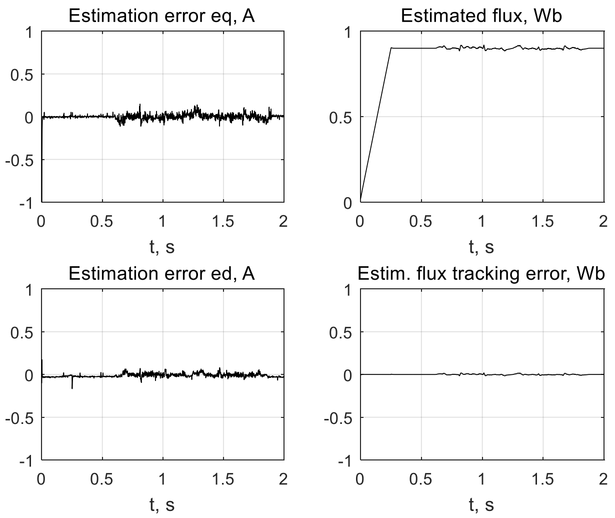

Fig. 5. **Transients of estimated values**
Source: **compiled by the authors**

The next set of experiments are caried out in order to investigate robustness properties of control algorithms, when parameter $\hat{\alpha}$ used in the control algorithm is different from the actual one.

During these tests modified operation sequence was used: after excitation motor track speed reference trajectory from zero speed to value 50 rad/s, as reported in Fig. 3 and speed remain constant;
at t = 0.8 s the rated contant load torque is applied, which remains constant up to the end of the test.

Speed error dynamic behaviour, stedy state values of q-axis current and active power level where analysed. Transients of considered variables in case of known parameters are the same for all controllers as depicted in Fig.6 he transients for all three control schemes, when estimated parameter $\hat{\alpha}=1.7\alpha$ in controller, are shown in Fig.7. According to previous investigations [9], this variation is the most critical for control stability and performances. From Fig.7a it follows, that non-robust version of indirect vector control demonstrates significant control performance and efficiency degradation. In comparison with Fig.6, one can see significant increasing of torque producing current and consumed active power at steady state. Active power level is increased from 200 W to 320 W while mechanical power in all cases remains at the same level of 125 W. RIFOC control scheme provides better dynamic control performances which are similar no undisturbed one. Active power level increased to 240 W, which shows a better efficiency of RIFOC. Proposed DFOC scheme has same good dynamic performance as RIFOC, but level of active power remains equal to 200 W and corresponds to undisturbed case.

By this experimental investigation we proof, that developed DFOC control algorithm provides better control performances and energy efficiency under rotor resistance variations condition. The main reason for such results is robustness properties of the full order closed loop observer.

CONCLUSIONS

A novel configuration of the unified direct vector speed-flux controller for induction motors is presented. The proposed solution allows using any asymptotic observer of the rotor flux vector with exponential stability properties. A Lyapunov based constructive procedure for the design of corrective feedbacks of the observer and the structure of the flux controller is given, which provides the properties of global exponential asymptotic stability of the rotor flux vector control subsystem. Resulting structure of the flux subsystem allows designing the family of flux observers in order to achieve robustness to parametric perturbations of the IM model.

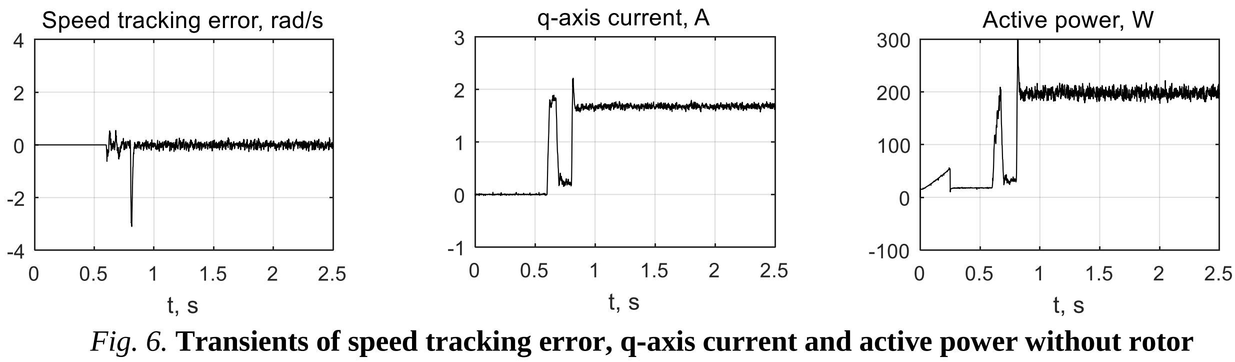

Fig. 6. **Transients of speed tracking error, q-axis current and active power without rotor resistance variation**
Source: **compiled by the authors**

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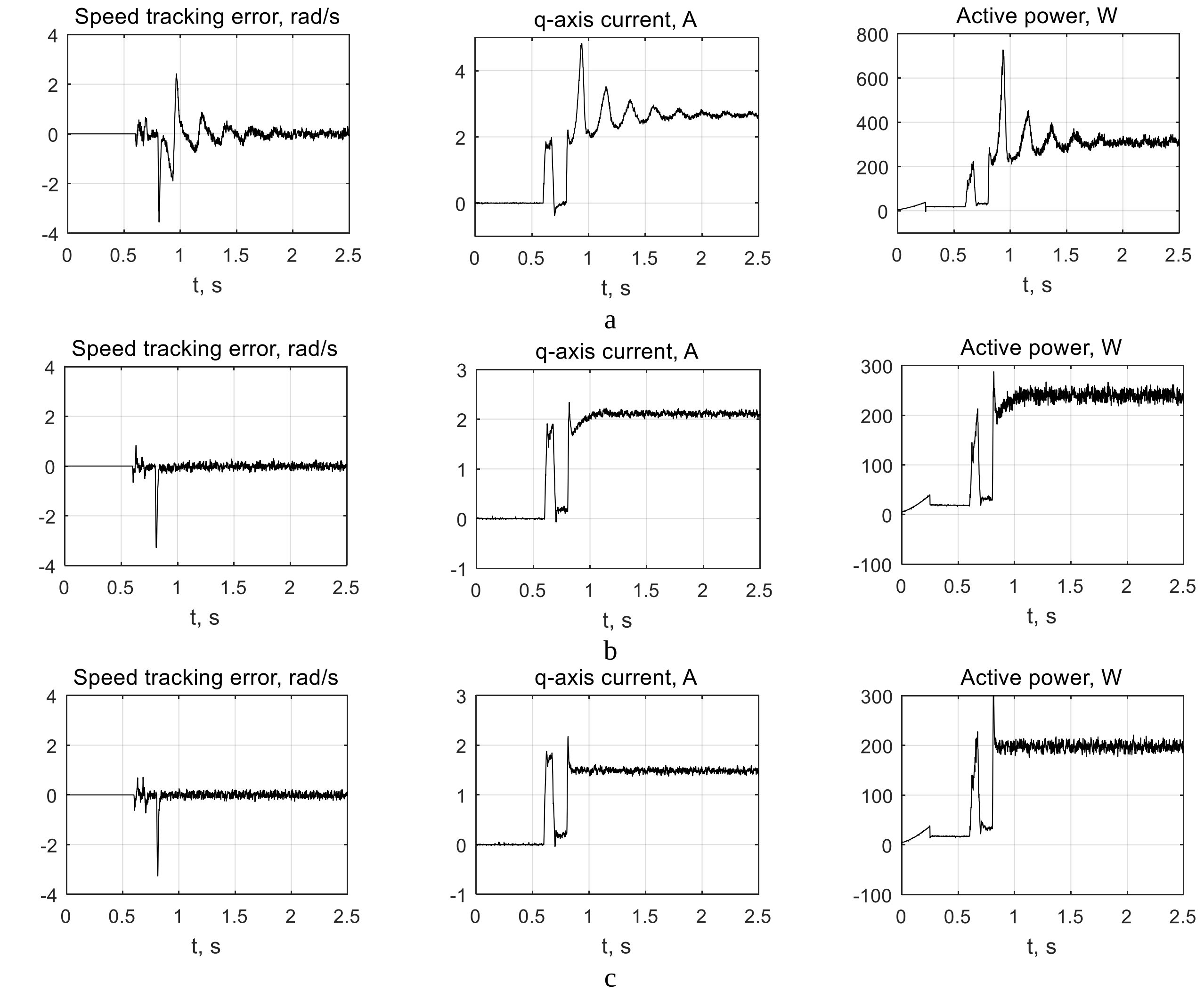

Fig. 7. **Transients of speed tracking error, q-axis current and active power with rotor resistance variation (a – IFOC, b – RIFOC, c – DFOC)**
Source: **compiled by the authors**

This theoretical results lead to decoupling properties of the electromechanical (torque and speed) and electromagnetic (flux) subsystems and allow to use the unified structure of the speed controller. In fact, the modified (more simple) structure of the speed controller without an explicit feedback linearization is used in the paper. Results of the intensive experimental comparative study of the proposed speed-flux controller and standard indirect field-oriented control, together with his robustified version, demonstrate the significantly improved robustness with respect of rotor resistance variation in terms of dynamic performance and energy efficiency.

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Пряме векторне керування швидкістю та потоком асинхронних двигунів: синтез та експериментальне дослідження робастновсті

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АНОТАЦІЯ

У статті представлено загальнотеоретичне рішення проблеми прямого векторного керування швидкістю та потокозчепленням асинхронного двигуна, яке базується на застосуванні другого методу Ляпунова. Структура алгоритму прямого векторного керування передбачає використання будь-якого асимптотичного спостерігача потоку з властивостями експоненціальної стійкості. Такий підхід дозволив підвищити властивості робастності системи векторного керування. Запропоновано конструктивну процедуру синтезу коригуючих зв'язків спостерігача потокозчеплення ротора. Розроблене сімейство спостерігачів потоку гарантує експоненціальну стійкість процесів оцінювання та робастність до параметричних збурень. Показано, що запропоноване рішення гарантує: глобальне експоненціальне відпрацювання заданих траєкторій потокозчеплення та швидкості разом з асимптотичною орієнтацією за полем, асимптотичну експоненціальну оцінку потоку ротора, а також асимптотичну розв'язку керування моментом (швидкістю) та потоком. Порівняльні експериментальні дослідження показують, що розроблений алгоритм керування забезпечує стабілізацію показників якості керування, а також коефіцієнту корисної дії на номінальному рівні при зміні активного опору ротора. Запропоновані структури прямого векторного керування можуть бути використані для розробки енергоефективних високоякісних асинхронних електроприводів для металообробки, пакувального обладнання, сучасного електротранспорту та спеціальної техніки.

Ключові слова: асинхронний двигун; робастне векторне керування; спостерігач потокозчеплення; програмна реалізація алгоритмів керування

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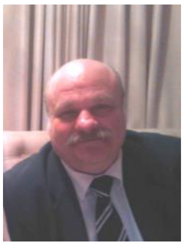

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