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IDENTIFICATION OF STATOR AND ROTOR RESISTANCES OF INDUCTION MOTORS

Abstract. *This article presents a novel identification method to identify the unknown stator and rotor resistances of induction motors. Our method allows one to identify the unknown parameters of nonlinear multivariable dynamic systems in real-time operation. An analysis of parameter convergence is provided, and, the regions of convergence is found by using the Lyapunov stability theory. Analytic, numerical and experimental studies are reported to illustrate and validate the mixed parameter identification and state estimation concept. The motor resistances are identified within practical operating modes. The proposed identification concept with state estimation ensures robustness and reduces shortcomings of identification algorithms. The aforementioned advantages uniquely suit advanced control schemes, such as adaptive and vector controls. Accurate identification of parameters allows one to guarantee optimal and energy-efficient operating envelopes which are of a very significant importance in energy and power systems. The experimental results illustrate that the motor parameters match the real motor parameters. Our findings are applicable in various high-performance energy, power and electromechanical systems.*
Keywords: *induction motor, parameter identification, stator and rotor resistance*

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ИДЕНТИФИКАЦИЯ АКТИВНОГО СОПРОТИВЛЕНИЯ СТАТОРА И РОТОРА АСИНХРОННОГО ДВИГАТЕЛЯ

Аннотация. *Изложен новый метод идентификации сопротивлений статора и ротора асинхронного двигателя, позволяющий производить их асимптотическую оценку в реальном времени. С использованием теории устойчивости по Ляпунову выполнен анализ и определены области устойчивости алгоритма идентификации. Представлены результаты математического моделирования и экспериментальных тестов, свидетельствующие о том, что разработанный алгоритм обеспечивает асимптотическую оценку неизвестных сопротивлений статора и ротора при реальных режимах работы асинхронного двигателя. Разработанный алгоритм идентификации может применяться для построения адаптивных систем векторного управления, которые будут гарантировать высокие показатели энергетической эффективности электромеханических систем. Полученные в работе результаты могут использоваться для проектирования высокоэффективных электромеханических систем.*
Ключевые слова: *асинхронный двигатель, идентификация параметров, активное сопротивлениестатора и ротора*

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ІДЕНТИФІКАЦІЯ АКТИВНОГО ОПОРУ СТАТОРА І РОТОРА АСИНХРОННОГО ДВИГУНА

Анотація. *Викладено новий метод ідентифікації опорів статора та ротора асинхронного двигуна, що дає змогу здійснювати їх асимптотичне оцінювання в реальному часі. З використанням теорії стійкості за Ляпуновим виконано аналіз і визначено області стійкості алгоритму ідентифікації. Наведено результати математичного моделювання та експериментальних тестів, які свідчать про те, що розроблений алгоритм забезпечує асимптотичне оцінювання невідомих активних опорів статора і ротора при реальних режимах роботи асинхронного двигуна. Розроблений алгоритм ідентифікації може використовуватися для побудови адаптивних систем векторного керування асинхронними двигунами, які будуть гарантувати високі показники енергетичної ефективності електромеханічних систем. Отримані в роботі результати можуть застосовуватися для проектування високоефективних електромеханічних систем.*
Ключові слова: *асинхронний двигун, ідентифікація параметрів, активний опір статора і ротора*

I. Introduction

Vector field-oriented control of induction motors (IM) is widely used in various comercial electric drives [1]. One of the major issues, which affect performance and capabilities of closed-loop systems with induction motors, is identification of motor parameters. It is well known that temperature-dependent varying stator and rotor resistances significantly affect the torque-speed characteristics, efficiency, currents and fluxes transients, motor dynamics and other characteristics [1]. Variation of resistances affects effectiveness of initialization of control schemes which are configured during self-commissioning process used in majority commercial electric drives [2]. Real-time identification with consequent self-adaptation and reconfiguration will result in optimal performance and capabilities of application-specific and commercial electric drives.

Starting from pioneering work [3], significant efforts have being performed to develop identification algorithms for real-time identification of the rotor resistance. Different approaches are reported in [4] – [6]. In [5] the estimation algorithm was used to update an observer-based controller for current-fed machines. A control algorithm which uses a full-order induction motor model is designed ensuring adaptation to the rotor resistance [7].

The aforementioned algorithms to identify the rotor resistance require the knowledge of stator resistance which also varies up to 50 % during motor operation. As shown in [5], varying stator resistance causes not accurate estimation of the rotor resistance. Therefore the simultaneous identification of slowly varying stator and rotor resistances is of a particular importance to implement advanced control schemes for squirrel-cage induction motors.

The first effort to the subject of our studies was design of the adaptive flux observer which ensures convergence of both stator and rotor resistance estimates when the electromagnetic torque is constast [8]. The stability analysis is performed by using the Lyapunov method under simplifying assumptions. Globally asymptotically stable adaptive identification and state es timation of the rotor resistance and rotor fluxes, assuming that integrals of stator currents are bounded, proposed in [9]. The algorithm, reported in [10], is similar to [9], however, does not require over-parameterization used in [9]. Simulation and experiments show that algorithms [9, 10] ensure asymptotic identification and estimation of the resistances if the electromagnetic torque is not zero. However, very complex ten- and eleven- order identifiers/observers is difficult to implement in real-time control schemes [9, 10].

A different approach for identification of stator and rotor resistances is given in [11]. The design is based on development of linear parameterization of a mathematical model with respect of resistances assuming that the angular velocity is slow varying. In order to avoid over- parameterization, the authors applied special coordinate transformation which allows estimation of two unknown parameters using a scalar function of stator currents. However, persistency of excitation, required for parameters convergence, is not satisfied in general operating modes.

Our goal is to design new robust identification algorithm to identify varying stator and rotor resistances assuming that the angular velocity varies slowly. The identification algorithm does not use over-parameterization and guaranties local asymptotic estimation of unknown parameters if the persistency of excitations conditions is satisfied. The proposed identification concept is validated by experiments.

II. Identification of Parameters: Problem Formulation

The mathematical model of induction motors in the stationary reference frame is give as [1]

$$\begin{aligned}
\dot{i}_a &= -\left(\frac{R_1}{\sigma} + \frac{R_2}{L_2} L_m \beta\right) i_a - \\
&\quad -\beta\left(-\frac{R_2}{L_2}\psi_a - p_n \omega \psi_b\right) + \frac{1}{\sigma} u_a \\
\dot{i}_b &= -\left(\frac{R_1}{\sigma} + \frac{R_2}{L_2} L_m \beta\right) i_b - \\
&\quad -\beta\left(-\frac{R_2}{L_2}\psi_b + p_n \omega \psi_a\right) + \frac{1}{\sigma} u_b \\
\dot{\psi}_a &= -\frac{R_2}{L_2}\psi_a - p_n \omega \psi_b + \frac{R_2}{L_2} L_m i_a \\
\dot{\psi}_b &= -\frac{R_2}{L_2}\psi_b + p_n \omega \psi_a + \frac{R_2}{L_2} L_m i_b ,
\end{aligned} \qquad (1)$$

where $\boldsymbol{u} = (u_a, u_b)^T$, $\boldsymbol{i} = (i_a, i_b)^T$ and $\boldsymbol{\psi} = (\psi_a, \psi_b)^T$ are the stator voltage, stator current and rotor flux linkage vectors; ω is the angular velocity; p_n is the number of pole pairs, and, we let p_n=1 without lost of generality;

L_1, L_2 and L_m are the stator, rotor and magnetizing inductances; R_1 and R_2 are the stator and rotor resistances.

Define the positive-definite parameter-dependent constants as

$$\beta = \frac{L_m}{\sigma L_2},\ \sigma = L_1 - \frac{L_m^2}{L_2}.$$

Our goal is to design a robust identification algorithm to identify varying stator and rotor resistances R_1 and R_2. These resistances must be identified using measured $(u_a, u_b)^T$, $(i_a, i_b)^T$ and ω.

Induction motors are asymptotically stable systems [6]. Correspondingly, $(u_a, u_b)^T$, $(i_a, i_b)^T$, $(\psi_a, \psi_b)^T$ and ω are bounded for all $t \geq 0$. Assuming the linear magnetic system, the motor inductances are constant. Correspondingly, only R_1 and R_2 are of our interest.

Problem Formulation – Design an identification algorithm for asymptotically stable nonlinear dynamic systems, such as induction motors.

Our identification scheme must guarantee:

1. Asymptotic stability of identification of stator and rotor resistances R_1 and R_2, such that
$$\lim_{t\to\infty}\left(R_1 - \hat{R}_1\right) = 0,\quad \lim_{t\to\infty}\left(R_2 - \hat{R}_2\right) = 0. \quad (2)$$
2. Asymptotic estimation of the currents, such that
$$\lim_{t\to\infty}\left(i_a - \hat{i}_a\right) = 0,\quad \lim_{t\to\infty}\left(i_b - \hat{i}_b\right) = 0, \quad (3)$$

where $\hat{i}_a$, $\hat{i}_b$ are the estimations of $i = (i_a, i_b)^T$.

III. Identification Algorithm

We combine the first and third equations of (1), as well as, combine the second and forth differential equations in (1). One obtains

$$\begin{aligned} \dot{i}_a &= -R_1 i_a/\sigma - \beta\dot{\psi}_a + u_a/\sigma \\ \dot{i}_b &= -R_1 i_b/\sigma - \beta\dot{\psi}_b + u_b/\sigma. \end{aligned} \quad (4)$$

Using (4), and, differentiating the first and second equations of (1), we have

$$\begin{aligned} \ddot{i}_a =& -R_1\dot{i}_a/\sigma - \omega i_b R_1/\sigma - \left(L_m R_2\beta + R_2\right)\dot{i}_a/L_2 + \\ &+ R_2 u_a/\sigma L_2 - R_1 R_2 i_a/\sigma L_2 - \\ &- \omega\dot{i}_b + \omega u_b/\sigma + \dot{u}_a/\sigma + \beta\dot{\omega}\psi_b \\ \ddot{i}_b =& -R_1\dot{i}_b/\sigma + \omega i_a R_1/\sigma - \left(L_m R_2\beta + R_2\right)\dot{i}_b/L_2 + \\ &+ R_2 u_b/\sigma L_2 - R_1 R_2 i_b/\sigma L_2 + \\ &+ \omega\dot{i}_a - \omega u_a/\sigma + \dot{u}_b/\sigma - \beta\dot{\omega}\psi_a. \end{aligned} \quad (5)$$

Following [11] we add to the right and left sides of (5) $c\dot{i}_a$ and $c\dot{i}_b$, respectfully. One has

$$\begin{aligned} \ddot{i}_a + c\dot{i}_a =& c\dot{i}_a - \alpha_1\left(\dot{i}_a + \omega i_b\right) - \\ &-\alpha_2\left[\left(L_m\beta + 1\right)\dot{i}_a - u_a/\sigma\right] - \\ &-\alpha_1\alpha_2 i_a - \omega\dot{i}_b + \omega u_b/\sigma + \dot{u}_a/\sigma + \beta\dot{\omega}\psi_b \\ \ddot{i}_b + c\dot{i}_b =& c\dot{i}_b - \alpha_1\left(\dot{i}_b - \omega i_a\right) - \\ &-\alpha_2\left[\left(L_m\beta + 1\right)\dot{i}_b - u_b/\sigma\right] - \\ &-\alpha_1\alpha_2 i_b + \omega\dot{i}_a - \omega u_a/\sigma + \dot{u}_b/\sigma - \beta\dot{\omega}\psi_a, \end{aligned} \quad (6)$$

where $\alpha_1 = \dfrac{R_1}{\sigma}$ and $\alpha_2 = \dfrac{R_2}{L_2}$; c is the constant, $c>0$.

Using the Laplace operator p, from (6), we obtain

$$\begin{aligned} pi_a\left(p + c\right) =& cpi_a - \alpha_1\left(pi_a + \omega i_b\right) - \\ &-\alpha_2\left[\left(L_m\beta + 1\right)pi_a - u_a/\sigma\right] - \\ &-\alpha_1\alpha_2 i_a - \omega pi_b + \omega u_b/\sigma + \\ &+ pu_a/\sigma + \beta p\omega\psi_b \\ pi_b\left(p + c\right) =& cpi_b - \alpha_1\left(pi_b - \omega i_a\right) - \\ &-\alpha_2\left[\left(L_m\beta + 1\right)pi_b - u_b/\sigma\right] - \\ &-\alpha_1\alpha_2 i_b + \omega pi_a - \omega u_a/\sigma + \\ &+ pu_b/\sigma - \beta p\omega\psi_a. \end{aligned} \quad (7)$$

We multiply the right and left sides of (6) on $w(p) = 1/(p + c)$. Thus,

$$\begin{aligned} pi_a =& cpi_a w(p) - \alpha_1\left(pi_a + \omega i_b\right)w(p) - \\ &-\alpha_2\left[\left(L_m\beta + 1\right)pi_a - u_a/\sigma\right]w(p) - \\ &-\alpha_1\alpha_2 i_a w(p) - \omega pi_b w(p) + \omega u_b w(p)/\sigma + \\ &+ pu_a w(p)/\sigma + \beta p\omega\psi_b w(p) \\ pi_b =& cpi_b w(p) - \alpha_1\left(pi_b - \omega i_a\right)w(p) - \\ &-\alpha_2\left[\left(L_m\beta + 1\right)pi_b - u_b/\sigma\right]w(p) - \\ &-\alpha_1\alpha_2 i_b w(p) + \omega pi_a w(p) - \omega u_a w(p)/\sigma + \\ &+ pu_b w(p)/\sigma - \beta p\omega\psi_a w(p). \end{aligned} \quad (8)$$

One denotes

$$\begin{aligned}&i_{a0}=i_a w(p), i_{b0}=i_b w(p),\\&i_{a1}=pi_a w(p), i_{b1}=pi_b w(p),\\&u_{a0}=u_a w(p), u_{b0}=u_b w(p),\\&u_{a1}=pu_a w(p), u_{b1}=pu_b w(p),\\&\eta_a=\beta p\omega\psi_b w(p),\\&\eta_b=-\beta p\omega\psi_a w(p).\end{aligned} \quad (9)$$

Using (8) and (9), we have

$$\begin{aligned}pi_a &= ci_{a1}-\alpha_1(i_{a1}+\omega i_{b0})-\\&-\alpha_2\left[(L_m\beta+1)i_{a1}-u_{a0}/\sigma\right]-\\&-\alpha_1\alpha_2 i_{a0}-\omega i_{b1}+\omega u_{b0}/\sigma+\\&+u_{a1}/\sigma+\eta_a\\pi_b &= ci_{b1}-\alpha_1(i_{b1}-\omega i_{a0})-\\&-\alpha_2\left[(L_m\beta+1)i_{b1}-u_{b0}/\sigma\right]-\\&-\alpha_1\alpha_2 i_{b0}+\omega i_{a1}-\omega u_{a0}/\sigma+\\&+u_{b1}/\sigma+\eta_b,\end{aligned} \quad (10)$$

where η_a and η_b are the excitations which are proportional to $d\omega/dt$, $\boldsymbol{\eta}=(\eta_a,\eta_b)^T$.

Equations (10) and (9) in the differential form become

$$\begin{aligned}\dot{i}_a &= ci_{a1}-\alpha_1(i_{a1}+\omega i_{b0})-\\&-\alpha_2\left[(L_m\beta+1)i_{a1}-u_{a0}/\sigma\right]-\\&-\alpha_1\alpha_2 i_{a0}-\omega i_{b1}+\omega u_{b0}/\sigma+\\&+u_{a1}/\sigma+\eta_a\\\dot{i}_b &= ci_{b1}-\alpha_1(i_{b1}-\omega i_{a0})-\\&-\alpha_2\left[(L_m\beta+1)i_{b1}-u_{b0}/\sigma\right]-\\&-\alpha_1\alpha_2 i_{b0}+\omega i_{a1}-\omega u_{a0}/\sigma+\\&+u_{b1}/\sigma+\eta_b,\end{aligned} \quad (11)$$

$$\begin{aligned}&\dot{i}_{a0}=i_a-ci_{a0},\\&\dot{i}_{b0}=i_b-ci_{b0},\\&i_{a1}=i_a-ci_{a0}, i_{b1}=i_b-ci_{b0},\\&\dot{u}_{a0}=u_a-cu_{a0},\\&\dot{u}_{b0}=u_b-cu_{b0},\\&u_{a1}=u_a-cu_{a0}, u_{b1}=u_b-cu_{b0},\\&\dot{\eta}_a=-\beta\dot{\omega}\psi_b-c\eta_a\\&\dot{\eta}_b=\beta\dot{\omega}\psi_a-c\eta_b.\end{aligned} \quad (12)$$

Remark 1. From (12) one concludes that $-\dot{\omega}\psi_b$ and $\dot{\omega}\psi_a$ are the inputs of the first-order filters which characterized by the coefficients $\beta(1/c)$ and time constant 1/*c*. The convergence constant *c* affects the overall robustness and convergence of the identification algorithm. For slow-varying or constant angular velocity *ω*, the expressions $-\dot{\omega}\psi_b$ and $\dot{\omega}\psi_a$ are zeros.

We write equations (11) as

$$\begin{aligned}\dot{i}_a &= f_a+\alpha_1 f_{1a}+\alpha_2 f_{2a}-\alpha_1\alpha_2 i_{a0}+\eta_a\\\dot{i}_b &= f_b+\alpha_1 f_{1b}+\alpha_2 f_{2b}-\alpha_1\alpha_2 i_{b0}+\eta_b,\end{aligned} \quad (13)$$

where

$$\begin{aligned}&f_a=ci_{a1}-\omega i_{b1}+\omega u_{b0}/\sigma+u_{a1}/\sigma\\&f_b=ci_{b1}+\omega i_{a1}-\omega u_{a0}/\sigma+u_{b1}/\sigma\\&f_{1a}=-(i_{a1}+\omega i_{b0})\\&f_{1b}=-(i_{b1}-\omega i_{a0})\\&f_{2a}=-\left[(L_m\beta+1)i_{a1}-u_{a0}/\sigma\right]\\&f_{2b}=-\left[(L_m\beta+1)i_{b1}-u_{b0}/\sigma\right].\end{aligned} \quad (14)$$

From (13), one obtains the equation for the currents observer as

$$\begin{aligned}\dot{\hat{i}}_a &= f_a+\hat{\alpha}_1 f_{1a}+\hat{\alpha}_2 f_{2a}-\hat{\alpha}_1\hat{\alpha}_2 i_{a0}+k_i\tilde{i}_a\\\dot{\hat{i}}_b &= f_b+\hat{\alpha}_1 f_{1b}+\hat{\alpha}_2 f_{2b}-\hat{\alpha}_1\hat{\alpha}_2 i_{b0}+k_i\tilde{i}_b,\end{aligned} \quad (15)$$

where $\hat{\boldsymbol{i}}=(\hat{i}_a,\hat{i}_b)^T$ is the vector of estimation for $\boldsymbol{i}=(i_a,i_b)^T$; $\tilde{i}_a=i_a-\hat{i}_a$, $\tilde{i}_b=i_b-\hat{i}_b$ are the current estimation errors; $\hat{\alpha}_1$, $\hat{\alpha}_2$ are the estimations of α_1 and α_2; k_i is the constant, $k_i>0$.

Using (13) and (15), the dynamics of estimation errors are

$$\begin{aligned}\dot{\tilde{i}}_a &= -k_i\tilde{i}_a+\tilde{\alpha}_1 f_{1a}+\tilde{\alpha}_2 f_{2a}-\hat{\alpha}_1\tilde{\alpha}_2 i_{a0}-\\&-\tilde{\alpha}_1\hat{\alpha}_2 i_{a0}-\tilde{\alpha}_1\tilde{\alpha}_2 i_{a0}+\eta_a\\\dot{\tilde{i}}_b &= -k_i\tilde{i}_b+\tilde{\alpha}_1 f_{1b}+\tilde{\alpha}_2 f_{2b}-\hat{\alpha}_1\tilde{\alpha}_2 i_{b0}-\\&-\tilde{\alpha}_1\hat{\alpha}_2 i_{b0}-\tilde{\alpha}_1\tilde{\alpha}_2 i_{b0}+\eta_b,\end{aligned} \quad (16)$$

where $\tilde{\alpha}_1=\alpha_1-\hat{\alpha}_1$ and $\tilde{\alpha}_2=\alpha_2-\hat{\alpha}_2$ are the estimation errors of α_1 and α_2.

The approximation of (16) can be performed neglecting $\tilde{\alpha}_1\tilde{\alpha}_2$ and $\boldsymbol{\eta}$ due to slow-varying or constant ω.

To examine the identification and estimation stability, we use the quadratic positive-definite function

$$V=\frac{1}{2}\left(\tilde{i}_a^2+\tilde{i}_b^2\right)+\frac{1}{2\gamma_1}\tilde{\alpha}_1^{\,2}+\frac{1}{2\gamma_2}\tilde{\alpha}_2^{\,2}, \quad (17)$$

where γ_1 and γ_2 are the constants, $\gamma_1>0$ and $\gamma_2>0$.

Using (16) and defining

$$\begin{aligned} &\dot{\tilde{\alpha}}_1 = -\dot{\hat{\alpha}}_1 = \\ &\quad = -\gamma_1\left[\left(f_{1a}-\hat{\alpha}_2 i_{a0}\right)\tilde{i}_a + \left(f_{1b}-\hat{\alpha}_2 i_{b0}\right)\tilde{i}_b\right] \\ &\dot{\tilde{\alpha}}_2 = -\dot{\hat{\alpha}}_2 = \\ &\quad = -\gamma_2\left[\left(f_{2a}-\hat{\alpha}_1 i_{a0}\right)\tilde{i}_a + \left(f_{2b}-\hat{\alpha}_1 i_{b0}\right)\tilde{i}_b\right], \end{aligned} \quad (18)$$

the first derivative of (17) is given by

$$\dot{V} = -k_i\tilde{i}_a^{\,2} - k_i\tilde{i}_b^{\,2}. \quad (19)$$

From (12), (14), (15) and (18), the identification and estimation equations become

$$\begin{aligned} &\dot{\hat{i}}_a = f_a + \hat{\alpha}_1 f_{1a} + \hat{\alpha}_2 f_{2a} - \hat{\alpha}_1\hat{\alpha}_2 i_{a0} + k_i\tilde{i}_a \\ &\dot{\hat{i}}_b = f_b + \hat{\alpha}_1 f_{1b} + \hat{\alpha}_2 f_{2b} - \hat{\alpha}_1\hat{\alpha}_2 i_{b0} + k_i\tilde{i}_b \\ &\dot{\hat{\alpha}}_1 = \gamma_1\left[\left(f_{1a}-\hat{\alpha}_2 i_{a0}\right)\tilde{i}_a + \left(f_{1b}-\hat{\alpha}_2 i_{b0}\right)\tilde{i}_b\right] \\ &\dot{\hat{\alpha}}_2 = \gamma_2\left[\left(f_{2a}-\hat{\alpha}_1 i_{a0}\right)\tilde{i}_a + \left(f_{2b}-\hat{\alpha}_1 i_{b0}\right)\tilde{i}_b\right] \quad (20) \\ &\dot{i}_{a0} = i_a - c i_{a0} \\ &\dot{i}_{b0} = i_b - c i_{b0} \\ &\dot{u}_{a0} = u_a - c u_{a0} \\ &\dot{u}_{b0} = u_b - c u_{b0}. \end{aligned}$$

Equations (20) contain four parameters c, k_i, γ_1 and γ_2. Here, the inputs are $\boldsymbol{u} = \left(u_a, u_b\right)^T$, $\boldsymbol{i} = \left(i_a, i_b\right)^T$ and ω. The output is $\hat{\boldsymbol{\alpha}} = \left(\hat{\alpha}_1, \hat{\alpha}_2\right)^T$.

The condition for asymptotic stability of (16) and (18) with observer (20) is

$$\lim_{t\to\infty}\left(\tilde{i}_a, \tilde{i}_b, \tilde{\alpha}_1, \tilde{\alpha}_2\right) = 0 .$$

From (17) and (19) it follows that $\tilde{i}_a$, $\tilde{i}_b$ $\tilde{\alpha}_1$ and $\tilde{\alpha}_2$ are bounded. In (20), the vectors $\boldsymbol{u} = \left(u_a, u_b\right)^T$, $\boldsymbol{i} = \left(i_a, i_b\right)^T$ and $\left(-\dot{\omega}\psi_b, \dot{\omega}\psi_a\right)^T$ are also bounded. From (12), we conclude that $\boldsymbol{\eta} = \left(\eta_a, \eta_b\right)^T$ are also bounded. The linear approximation of (16) and (18) gives the evolution of estimation errors as described by equations

$$\begin{aligned} &\dot{\tilde{\boldsymbol{i}}} = \boldsymbol{A}\tilde{\boldsymbol{i}} + \boldsymbol{W}(t)\tilde{\boldsymbol{p}}, \tilde{\boldsymbol{i}} = \left(\tilde{i}_a, \tilde{i}_b\right)^T, \\ &\dot{\tilde{\boldsymbol{p}}} = -\Lambda \boldsymbol{W}^T(t)\tilde{\boldsymbol{i}}, \tilde{\boldsymbol{p}} = \left(\tilde{\alpha}_1, \tilde{\alpha}_2\right)^T, \end{aligned} \quad (21)$$

where $\boldsymbol{A} = \begin{bmatrix} -k_i & 0 \\ 0 & -k_i \end{bmatrix}$;

$$\boldsymbol{W}^T(t) = \begin{bmatrix} f_{1a} - i_{a0}\hat{\alpha}_2 & f_{1b} - i_{b0}\hat{\alpha}_2 \\ f_{2a} - i_{a0}\hat{\alpha}_1 & f_{2b} - i_{b0}\hat{\alpha}_1 \end{bmatrix}; \Lambda = \begin{bmatrix} \gamma_1 & 0 \\ 0 & \gamma_2 \end{bmatrix}.$$

For the bounded $\tilde{\boldsymbol{p}}$, the mapping $\boldsymbol{W}(t)$ is also bounded. From (21), one concludes that $\dot{\tilde{\boldsymbol{i}}}$ is bounded. Using (19), we have

$$\begin{aligned} &\int_0^t \dot{V}(\tau)\,d\tau = -k_i\int_0^t\left(\tilde{i}_a^{\,2}(\tau) + \tilde{i}_b^{\,2}(\tau)\right)d\tau = \\ &\qquad = V(t) - V(0), \end{aligned} \quad (22)$$

and, $\lim_{t\to\infty}\int_0^t\left(\tilde{i}_a^{\,2}(\tau) + \tilde{i}_b^{\,2}(\tau)\right)d\tau < \infty$ for a bounded Lyapunov function $V(t)$. The Barbalat lemma [6] gives

$$\lim_{t\to\infty}\left(\tilde{i}_a, \tilde{i}_b\right) = 0. \quad (23)$$

From (21), we have

$$\lim_{t\to\infty}\boldsymbol{W}(t)\,\tilde{\boldsymbol{p}} = 0. \quad (24)$$

Furthermore, if

$$\int_t^{t+T}\boldsymbol{W}(\tau)\boldsymbol{W}^T(\tau)d\tau \geq c_1\boldsymbol{I} > 0, \quad (25)$$

for $T>0$ and $c_1>0$ for all $t \geq 0$, then, the persistency of excitation conditions for (21) are guaranteed. Furthermore, $\left(\tilde{\boldsymbol{i}}, \tilde{\boldsymbol{p}}\right)$ decay exponentially to zero [6].

IV. Experimental Results

Our goal is simultaneously identify the stator and rotor resistances R_1 and R_2. The developed identification and estimation algorithm (20) is used. We report simulation and experimental results for a 0,75 kW induction motor with the following rated data: torque – 2,5 N-m , speed – 314 rad/sec, current – 2,1 A, voltage – 380 V, R_{1r} =11 Ohm, R_{2r}=5,5 Ohm, L_1=0,95 H, L_2=0,915 H and L_m=0,91 H, total inertia – 0.0036 kg·m^2. In the identification algorithm, let c=20, k_i=700, γ_1=10000 and γ_2=20.

The experimental tests are carried out by using the Rapid Prototyping Station (RPS) at the National Technical University of Ukraine [12]. The RPS includes: 1) 20A and 380 V three-phase PWM controlled inverter, operated at 10 kHz switching frequency; coupled induction motor and current controlled DC loading motor; 2) DSP TMS320C32 based real time controller, connected to the computer. The controller performs data acquisition, implements control algorithms with programmable real-time tracing of selected variables, and, generates two symmetrical three-phase PWM driving signals; 3) personal computer for processing, programming, interactive oscilloscope, data storage, data acquisition, etc.

The motor speed is measured by a 1000 pulse/revolution optical encoder. The sampling time is 200 μ sec. A discrete-time control algorithm is implemented. During identification process, an induction motor operates under the speed-tracking vector control, applying the operation sequence as reported in Fig.1. The motor is excited during the initial time interval [0 0,25] sec using a flux reference trajectory starting at $\psi_0 = 0{,}02$ Wb, reaching the rated flux 0.9 Wb with the rate 3,52 Wb/sec. The unloaded motor is required to track the angular velocity reference trajectory, starting at 0.6 sec from zero to 50 rad/sec with the rates (first and second derivatives) 667 rad/sec^2 and 26667 rad/sec^3. At 1,2 sec, a constant load torque T_L=2,5 N-m is applied.

We numerically and experimentally substantiate the identification algorithm within two phases. First, we examine the evolutions of the estimated states by the observer, and, parameter identification process independently from a control algorithm.

Figure 2 documents experimental evolutions of state estimates and parameter identification for different initial conditions. Asymptotic estimation and identification are achieved, and parameters converge within ~3 sec. The identified resistances are found to be $\hat{R}_1 = 10{,}7$ Ohm and $\hat{R}_2 = 5{,}4$ Ohm. The simulation results are illustrated in Figure 3 for the initial conditions $\hat{R}_1(0) = 2R_{1r}$ and $\hat{R}_2(0) = 2R_{2r}$. The comparison of experimental and simulation results, presented by Figures 2 and 3, show equivalency of dynamics, matching and overall accuracy. The documented findings substantiate our results.

We further verify and demonstrate the effectiveness and capabilities of our concept. Using the derived equations (20), we examine parameter identification when induction motor is controlled using an adaptive version of standard *indirect* vector control algorithm [13], where value of rotor and stator resistances in the field-oriented controller are constantly updated with the estimated ones. Figure 4 reports experimentally measured transient dynamics of angular velocity, torque component of the stator current vector with rated motor parameters. Very good dynamics is guaranteed during speed trajectory tracking and when the rated torque T_L is applied. If the control algorithm is derived under unknown rotor resistance, assuming that $R_2 = 2R_{2r}$=10,8 Ohm, the resulting dynamics is illustrated in Figure 5. The overall performance and capabilities are significantly worsening. The higher value of the resulting current i_q leads to significant reduction of efficiency, overheating, electromagnetic loads, etc.

Figure 6 reports the experimental transient dynamics of the closed-loop induction motor with the adaptive vector control algorithm when R_1 and R_2 are accurately identified using equations (20). The overall performance and motor capabilities are significantly improved. These features are of a great importance in energy and power systems, as well as in electric drives.

V. Conclusions

While a significant progress was made in identification of linear and nonlinear electromechanical systems, the identification of time-varying parameters of induction motors in real-time remains an open problem. Significant improvements of performance and capabilities of energy and power systems can be achieved if control systems are designed using accurately identified parameters. This paper reports a new identification scheme which is centered on a robust identification of rotor and stator resistances.

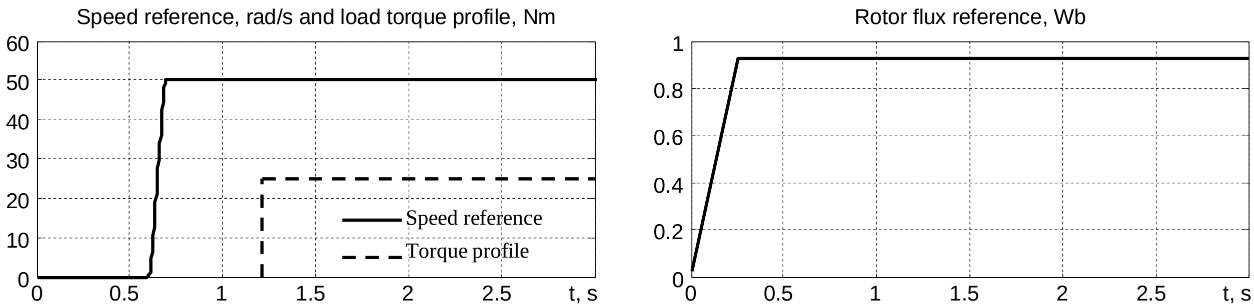

Fig. 1. Speed, flux reference and load torque profiles

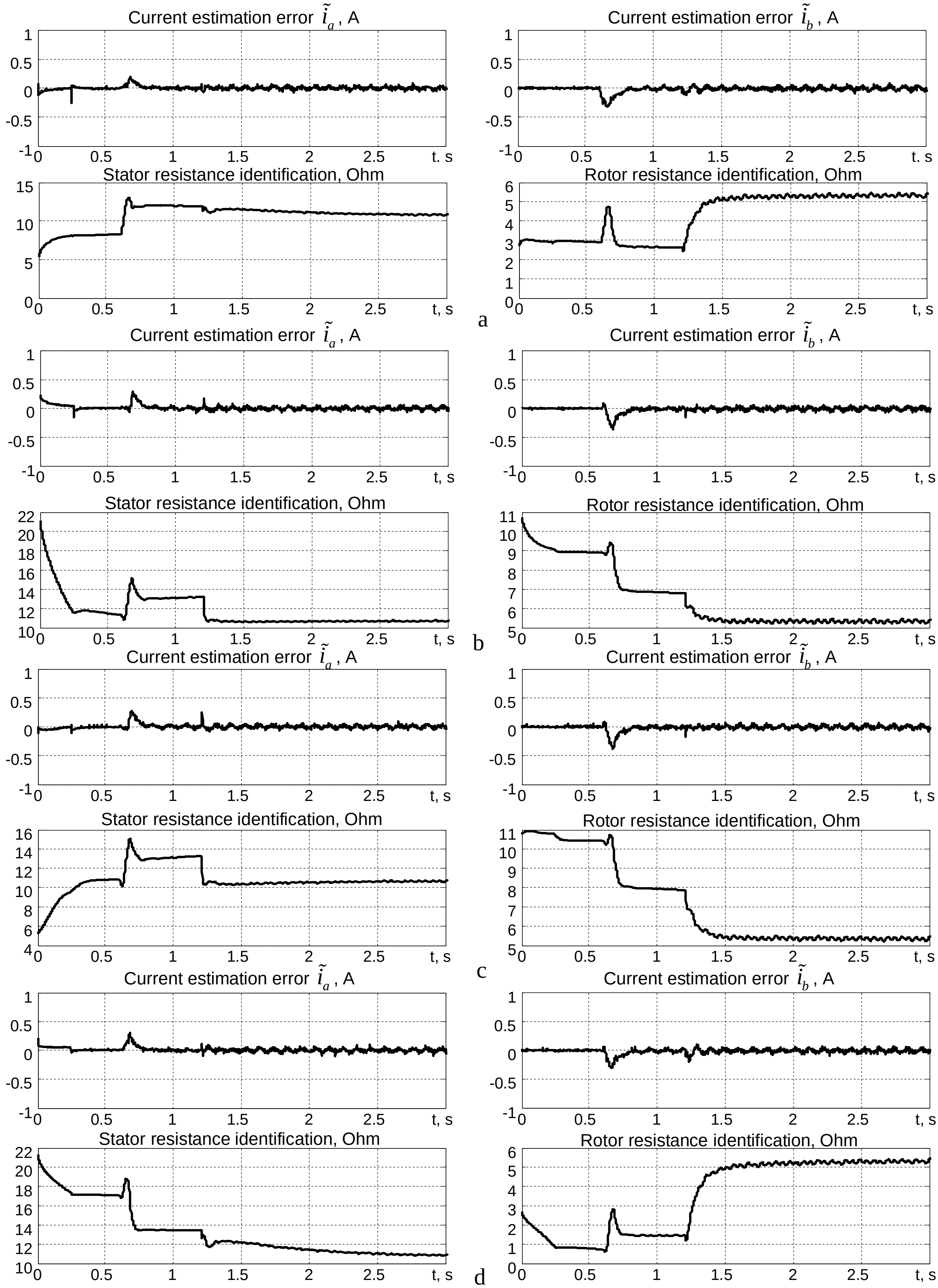

Fig. 2. Transients during identification: a) $\hat{R}_1(0) = 0{,}5R_{1r}$, $\hat{R}_2(0) = 0{,}5R_{2r}$;
b) $\hat{R}_1(0) = 2R_{1r}$, $\hat{R}_2(0) = 2R_{2r}$; c) $\hat{R}_1(0) = 0{,}5R_{1r}$, $\hat{R}_2(0) = 2R_{2r}$; d) $\hat{R}_1(0) = 2R_{1r}$, $\hat{R}_2(0) = 0{,}5R_{2r}$

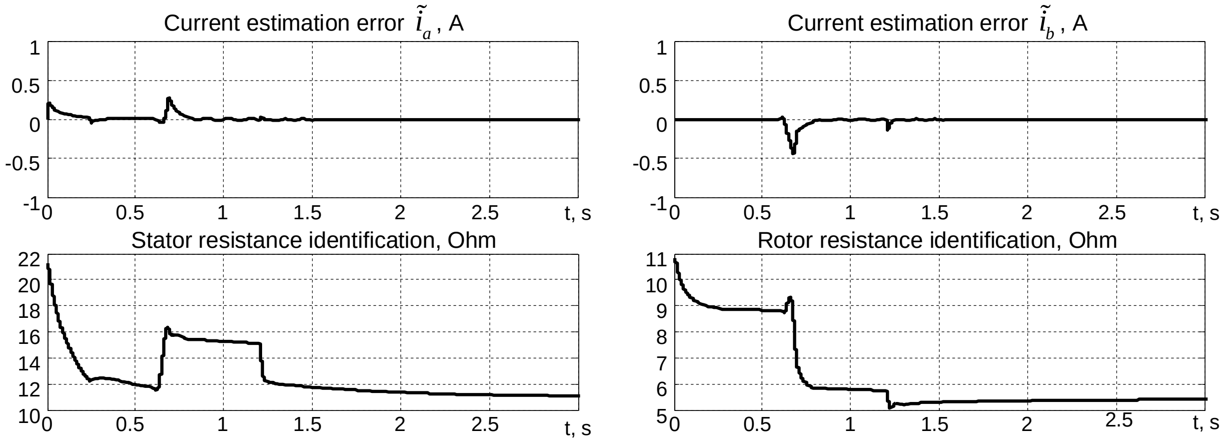

Fig. 3. Simulation results: transients during identification $R_1(0) = 2R_{1r}$, $R_2(0) = 2R_{2r}$

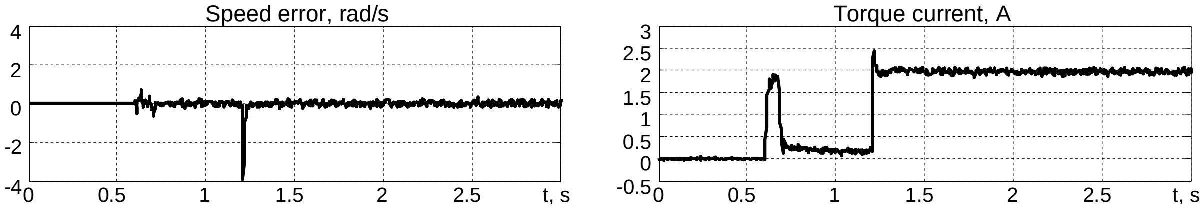

Fig. 4. Transients during speed vector control without adaptation letting R_{1r} and R_{2r}

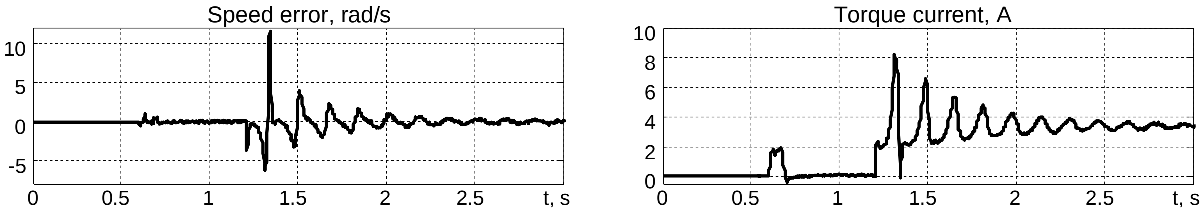

Fig. 5. Transients during speed vector control without adaptation assuming $\hat{R}_2 = 2R_{2r}$

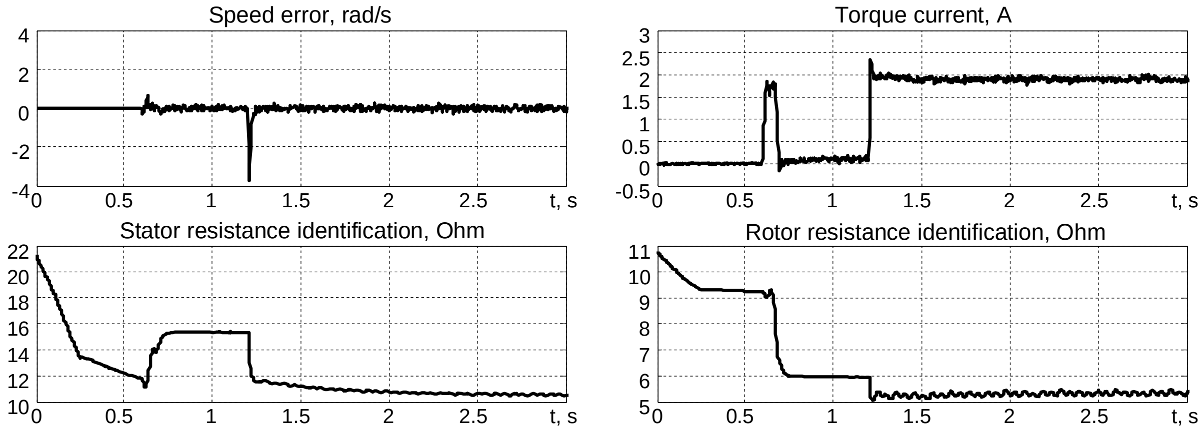

Fig. 6. Transients during adaptive speed control with $R_1(0) = 2R_{1r}$, $R_2(0) = 2R_{2r}$

The Lyapunov stability theory was applied to design a robust and asymptotically stable identification scheme. Robustness, stability and convergence of identified parameters were proven by means of experiments. The documented numeric and experimental results demonstrate a precise match between the actual and identified parameters within an operational motor envelope. The reported identification concept offers a simple, robust, effective and industry-relevant solution which can be used in high-performance commercial and application-specific drives and systems.

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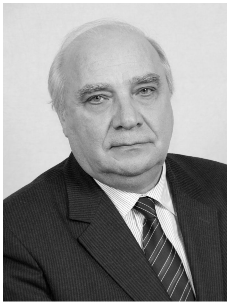

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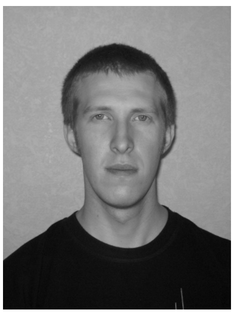

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