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Dynamic Identification of Rotor Magnetic Flux, Torque and Rotor Resistance of Induction Motor

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ABSTRACT In the modern high-performance drive applications, a high-precision and efficient control of the induction motor depends on the accuracy of parameter values. However, the motor parameters may change due to the winding temperature fluctuations, flux saturation and skin effect. Any discrepancy between the values of the motor's actual parameters and the ones used for the design of the controllers may result in degradation of the drive performance. In this work, a new identification method of hardly measurable internal quantities of the induction motor, such as components of the magnetic flux vector and electromagnetic torque, is outlined. Commonly, the measurable quantities of the induction motor like stator currents, stator voltage frequency and mechanical angular speed are used to determine a feedback effect of the rotor flux vector on the vector of the stator currents of induction motor. Based on this feedback, it is also possible to identify the actual value of the rotor resistance, which may alter during the induction motor operation. This has a significant impact on the precision of the identified quantities as well as on the master control of the induction motor. Stability of the identification structure is guaranteed by the position of roots of characteristic equation of its linear transfer function. Simulation and experimental results are given to highlight the quality, effectivity, feasibility, and robustness of the proposed identification method, which is working reliably within the whole range of the motor angular speed.

INDEX TERMS AC motors, magnetic flux, motion control, motor torque, system identification, variable speed drives.

I. INTRODUCTION

The use of the induction motor (IM) in the electric drives introduces indisputable advantages consisting of their simple design, maintenance, operational reliability and efficiency. In the industrial practice, IM is used for both conventional and high-performance applications [1]–[4]. From the control point of view, the control of IM brings difficulties, because it requires more complicated actuators and generally more complex control algorithms, for example the field-oriented control (FOC) [5], direct torque control (DTC) [6], [7], model predictive control (MPC) [8], [9] and model reference adaptive control (MRAC) [10]. Nowadays, frequency converters with advanced control features present excellent actuators for the IM drives. For the high-quality control algorithms, it is necessary to identify respective motor parameters and internal quantities that are changed during the operation of the electric drive system, for example due to the temperature fluctuations or the magnetic flux saturation.

In the literature, there are many approaches to determine the basic stationary parameters of IM, such as winding resistances, main and leakage inductances and moment of inertia on the motor shaft, where their exact determination can significantly affect precision of the IM control. In [11], [12], a finite element method is used to identify the parameters of the IM and its dynamic behavior. The literature [13] compares two methods determining the parameters of IM in a steady state, while the saturation of the magnetic circuit is also considered. In [14], the identification of the steady-state motor parameters are based on integration (at standstill, using integral calculations). An overview of the methods for identification of the stator and rotor resistances in a

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sensorless control of IM together with the design of an observer is outlined in [15]. In [16], there is a review article of the methods of an on-line and off-line identification of IM parameters.

For the high-quality dynamic control of IM, it is necessary to know the motor basic static parameters during its operation. Additionally, it is important to identify the basic internal quantities of the machine, such as the rotor magnetic fluxes and motor internal electrical torque on the rotor shaft. As the direct measurement of these quantities is rather difficult and often impossible, the instantaneous values of these quantities can be obtained indirectly, using various more or less complex identification methods.

The traditional identification methods include methods based on a certain type of IM model (so-called model-based methods). These methods are generally very sensitive to the accuracy of the model parameters and the accuracy of the measurement of machine quantities. Such methods of identifying IM magnetic fluxes include a method based on the so-called reduced voltage model [17], [18]. The advantage of this method is, that it does not need any speed sensing, but on the other hand, it is very sensitive to any offset error of the sensors used (of voltages and/or currents), because a pure integrator is used to observe the stator flux. The disadvantage of this method is, that it can only be used to identify the motor magnetic fluxes at higher IM speeds. This shortcoming is eliminated by the flux identification method based on the reduced current model of IM [19], [20], which is also suitable in range of low speeds of IM. This method is however sensitive to accurate knowledge of the rotor resistance (i.e. its rotor time constant). Combining these two methods has led to the use of the well-known Gopinath-type observer [21], which allows a smooth transition between the current and voltage models, but disturbances caused by sensor noise can lead to significant observation errors.

In order to eliminate the influence of the sensor noise and the inaccurate knowledge of the parameters for identification of IM magnetic fluxes, closed-loop identification methods using observers used. The two most promising approaches are the Luenberger observer [22] and the Kalman filter [23]. On the basis of the measurements of the rotor speed, stator currents, and voltages a nonlinear adaptive on-line observer with exponential convergence properties is designed to identify magnetic flux, as shown in [24]. It utilizes reduction of variations of the three critical parameters (load torque and motor resistances). It results in a complex dynamic nonlinear adaptive observer of the eighth-order. Generally, the problem of using observers in IM flux identification methods is, that the stability analysis of the methods is often difficult and in addition, it requires extended calculation and an estimation of the IM state. This may be negatively influenced by the uncertainty of the parameters that vary during operation (e.g. winding temperature rise, skin effect, and flux saturation [16], [25]).

The methods based on the artificial intelligence in a combination with observers may also be used to improve the identification of the internal IM quantities. An adaptive robust fuzzy observer according to [26] ensures asymptotic convergence of the magnetic flux and the load torque error. The magnetic fluxes and the IM torque are identified in parallel here. Two online methods of the identification of the load torque and the flux of IM combining the adaptive Luenberger observer theory and Takagi-Sugeno fuzzy logic method are outlined in [27]. There are presented sufficient conditions in order to ensure the asymptotic convergence of the flux, which are derived in a complex way according to 2nd Lyapunov's theory.

The identification of the parameters, magnetic fluxes and torque of IM is often applied to a specific type of motor control in order to improve drive dynamics, or to eliminate the angular speed sensor. Improving the identification of IM magnetic fluxes based on a neural network, and thus improving the control at DTC (direct torque control), is described in [28], [29]. The difficulty of the implementation of these methods consists mainly in the computational complexity and the execution time of the neural architecture. Accurate identification of the IM parameters and magnetic fluxes for subsequent control by genetic algorithms using a reduced-order robust observer can be found in the [30]. The method requires a complex state-space model and high computational costs linked to the implementation of a discrete observer. It is suitable for high speeds and it is sensitive to an exact value of machine magnetization inductance.

This article proposes a new high-quality dynamic identification structure for IM based on its mathematical model and allowing to identify rotor magnetic fluxes and IM electromagnetic torque simultaneously with an on-line adaptation of the rotor resistance. There is an identification method based on determining a feedback effect of the rotor flux vector on the vector of stator currents. The method requires only the knowledge of measurable quantities of IM, such as its stator currents and mechanical angular speed. The identification structure for identifying the rotor feedback influence on the stator of IM is simple and linear. Its stability is guaranteed by position of roots of the characteristic equation of its transfer function. It preserves the same properties within the whole range of angular of IM speeds. It also withstands any changes of rotor resistance. Achieved results confirming the efficiency and the quality of the proposed identification method were verified by the simulation and experimental measurements.

The proposed identification procedure can be divided into the following steps:

1) Identification of the influence of the rotor feedback on the stator (more specifically: a feedback of the rotor flux on the stator current)
2) Calculation of the rotor flux components based on the rotor mechanical angular speed and identified values of the rotor feedback influence on the stator
3) Calculation of the motor electromagnetic torque based on the measured stator currents and the identified rotor flux vector components

4) Real-time adaptation of the rotor resistance, using a parameter representing the inverse value of the rotor time constant, by comparing one component of the identified rotor flux using two different estimation methods, which may be reused for all calculations in relations depending on this parameter.

The article is organized as follows: Introduction of the mathematical model of IM is presented in Section 2. Next, Section 3 describes the design of the identification structure of determining of the rotor feedback effects on the stator, which are further used in the Section 4 to identify the rotor fluxes and IM torque. Section 5 describes a method of adaptation of a designed identification structure to changing rotor resistance. The proposed identification method is verified by simulation of various IM operating states and by experimental measurements using a laboratory model in the Section 5 and Section 6. Finally, a comparison of the results with existing methods and conclusions are presented in the Section 7.

II. MATHEMATICAL MODEL OF INDUCTION MOTOR

Many different types of the IM mathematical model are described in the literature, depending on selection of the motor quantities as state variables. In this model, the particular motor quantities (the stator current vector $\mathbf{i_1}$ and the rotor flux vector $\boldsymbol{\psi_2}$) are expressed by their components in the reference frame x-y rotating synchronously with the stator field vector by the angular frequency ω_1.

After selecting the stator current and rotor flux for the model state variables, the induction motor is described by the following set of equations, [20]:

$$\begin{bmatrix} \frac{\mathrm{d}i_{1x}}{\mathrm{d}t} \\ \frac{\mathrm{d}i_{1y}}{\mathrm{d}t} \\ \frac{\mathrm{d}\psi_{2x}}{\mathrm{d}t} \\ \frac{\mathrm{d}\psi_{2y}}{\mathrm{d}t} \end{bmatrix} = \begin{bmatrix} -\omega_0 & \omega_1 & -K_{12}\omega_g & -K_{12}\omega_m n_p \\ -\omega_1 & \omega_0 & K_{12}\omega_m n_p & -K_{12}\omega_g \\ M\omega_g & 0 & -\omega_g & \omega_2 \\ 0 & M\omega_g & -\omega_2 & \omega_g \end{bmatrix} \begin{bmatrix} i_{1x} \\ i_{1y} \\ \psi_{2x} \\ \psi_{2y} \end{bmatrix} + \begin{bmatrix} K_{11} & 0 \\ 0 & K_{11} \\ 0 & 0 \\ 0 & \end{bmatrix} \begin{bmatrix} u_{1x} \\ u_{1y} \end{bmatrix} \tag{1}$$

$$n_p \frac{M}{L_2}\left(\psi_{2x} i_{1y} - \psi_{2y} i_{1x}\right) - T_{load} = J\frac{d\omega_m}{dt} \tag{2}$$

The notation of the IM parameters motor and their values used for simulation and experimentation are given in the Table 1 in the Appendix. The parameters in (1) and (2) can be determined from the motor parameters by simple recalculation according to the following formulas:

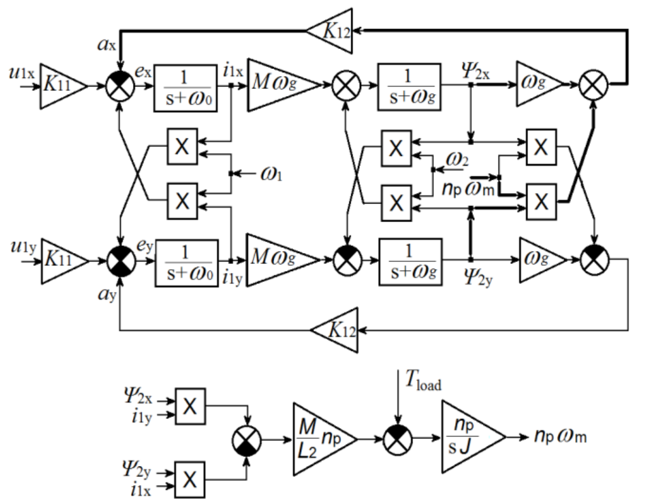

FIGURE 1. Block diagram of IM in the x-y reference frame rotating synchronously with stator field vector.

$$K_{11} = \frac{3}{2}\left(L_{s1} + \frac{L_{s2}L_m}{L_{s2}+L_m}\right)^{-1} \tag{3}$$

$$K_{12} = -\frac{3}{2}\left(L_{s1} + L_{s2} + \frac{L_{s1}L_{s2}}{L_m}\right)^{-1} \tag{4}$$

$$\omega_0 = K_{11}\left[R_1 + \left(\frac{M}{L_2}\right)^2 R_2\right] \tag{5}$$

The block diagram of the IM corresponding to equations (1) and (2) with the parameters adjusted according to the above equations (3) – (8) is shown in Fig. 1.

$$M = \frac{2}{3}L_m \tag{6}$$

$$\omega_g = \frac{R_2}{L_2} \tag{7}$$

$$L_2 = \frac{2}{3}\left(L_{s2} + L_m\right) \tag{8}$$

The dynamics of motor quantities (components of stator currents and rotor fluxes, and angular speed) after motor connection to the stator voltage $U_1 = 40$ V with the angular frequency $\omega_1 = 28.03$ rad/s., obtained by digital simulation in MATLAB/Simulink program is shown in Fig. 2.

The values of the motor parameters used for the simulation are listed in the Table 1 in the Appendix. At verification of the motor properties by numerical simulation it is assumed, that the IM operates under the scalar control, i.e. the ratio $U_1/\omega_1 =$ const. In this case, the x-component of the voltage u_{1x} can be considered identical to the magnitude of the stator voltage U_1, i.e. $U_1 = u_{1x}$.

III. DESIGN OF IDENTIFICATION STRUCTURE OF ROTOR FEEDBACK INFLUENCE ON STATOR CURRENTS

The purpose of the identification of the IM quantities is to determine their exact instantaneous values. The presented

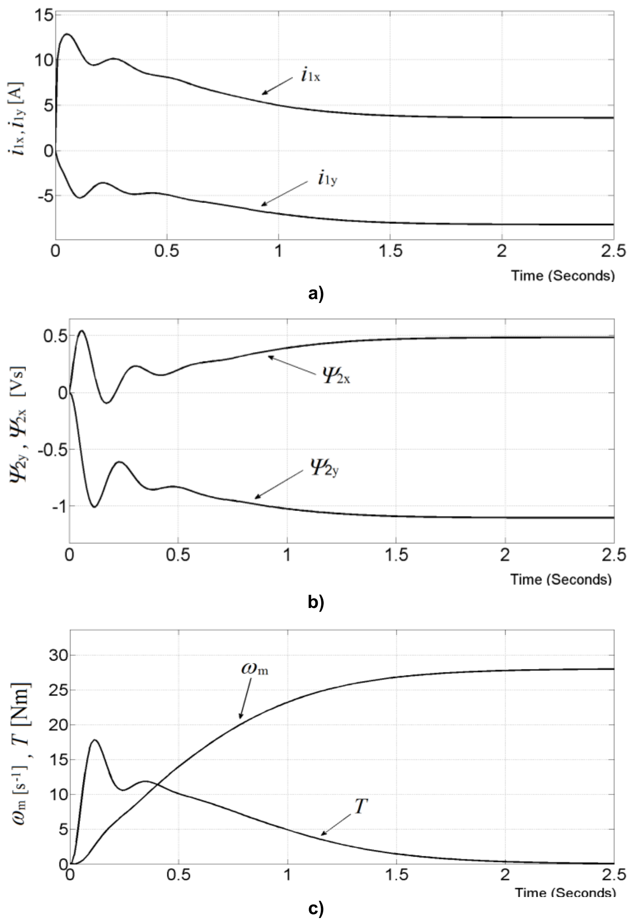

FIGURE 2. Dynamic responses of the induction motor: a) stator currents; b) rotor fluxes; c) mechanical angular speed and torque of the induction motor to step of input voltage and frequency.

idea of the identification of hardly measurable internal quantities of the IM (the rotor flux vector components and electromagnetic torque) is based on the identification of the feedback effect of the rotor flux vector on the stator current vector. The identification of the feedback effect of the rotor flux is commonly completed based on the induced voltage, which is either equal to zero at zero speed or is insignificant at low speeds [17], [18]. However, the considered feedback effect of the rotor flux also reflects itself on the stator current components equally within the entire speed range of IM. This fact was used in the design of the presented identification method. This influence upon the variable a_x is shown in the block diagram (Fig. 1), indicated by a thicker line starting from the quantities ψ_{2x}, ψ_{2y} and ω_m.

A similar path can be found for the variable a_y. For a clarity, the corresponding partial block diagrams are shown separately in Fig. 3.

Let's suppose, that for the induction motor with known instantaneous values of the stator input quantities U_1 a ω_1, the stator current components i_{1x}, i_{1y} and the mechanical angular speed ω_m can be measured. The instantaneous feedback value of the stator current component i_{1x} from the first

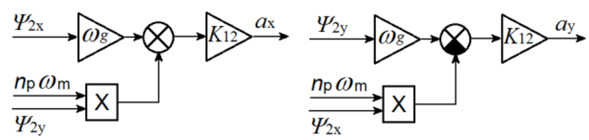

FIGURE 3. Block diagrams for determining variables a_x and a_y.

line of equation (1) is:

$$a_x = K_{12}(\omega_g\psi_{2x} + \omega_m n_p \psi_{2y}) \tag{9}$$

and similarly, for the stator current component i_{1y} it is:

$$a_y = K_{12}(\omega_g\psi_{2y} - \omega_m n_p \psi_{2x}) \tag{10}$$

For a high-quality dynamic identification of the variables a_x and a_y, the following identification structures have been designed (Fig. 4).

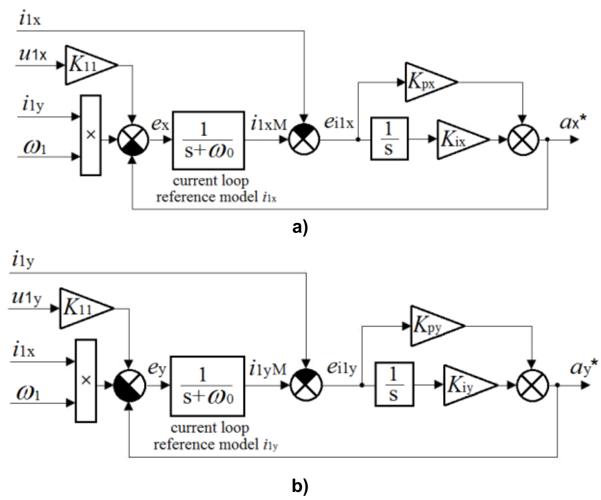

FIGURE 4. The structures of identification of the rotor feedback influence on stator, for the variables: a) a_x^*; b)a_y^*.

Fig. 4a shows the identification structure of identification of the component a_x (where its identified value is denoted by an asterisk: a_x^*). It presents a structure with the reference model for the stator current RL circuit having the first order transfer function. The same input signals e_x and e_y, like the one affecting the stator current component i_{1x} (Fig. 1), are fed into this reference model. The feedback effect of the rotor flux on this stator current component is identified from the deviation of the measured current i_{1x} and its predicted value i_{1xM} obtained from the reference model. This deviation is processed by a PI (proportional-integral) controller, whose output a_x* presents the identified value of a_x. The Laplace transfer function of the linear circuit is:

$$F(s) = \frac{a_x^*(s)}{i_{1x}(s)} = \frac{\frac{K_{px}s+K_{ix}}{s}}{1+\frac{K_{px}s+K_{ix}}{s}\frac{1}{s+\omega_0}} = \frac{K_{px}s^2 + \left(K_{px}\omega_0 + K_{ix}\right)s + K_{ix}\omega_0}{s^2 + \left(K_{px}+\omega_0\right)s + K_{ix}} \tag{11}$$

The circuit stability and its dynamics can be easily adjusted by a suitable choice of the proportional component gain and the integration component gain based on required position of poles of the transfer function (11). The identification structure in Fig. 4a is linear, and therefore its stability is guaranteed by negative real parts of the transfer function poles in the equation (11). This is obviously ensured, because all terms of the characteristic polynomial are positive.

Since the rotor time constant $1/\omega_g$ (for the considered IM it is about 35 ms) has the major influence on the dynamics of the feedback influence, the dynamics of the identification loop with the transfer function according to (11) was chosen. In this case, the two poles of the denominator are negative real and they are equal to -100. The substitute time constant according to the Shanon-Kotelnik theory will be approximately 5 times of the time constants corresponding to the poles, i.e, approx. 10 ms. Based on this, the calculated values of PI controllers in Fig. 4 are: $K_{px} = K_{py} = 10$ and $K_{ix} = K_{iy} = 11870$.

Analogically, according to the similar considerations and results like in the previous case, the identification structure of the identification of the component a_y (i.e., a_y^*) has also been designed (Fig. 4b). The properties of the designed identification structures have been verified by digital simulation (Fig. 5).

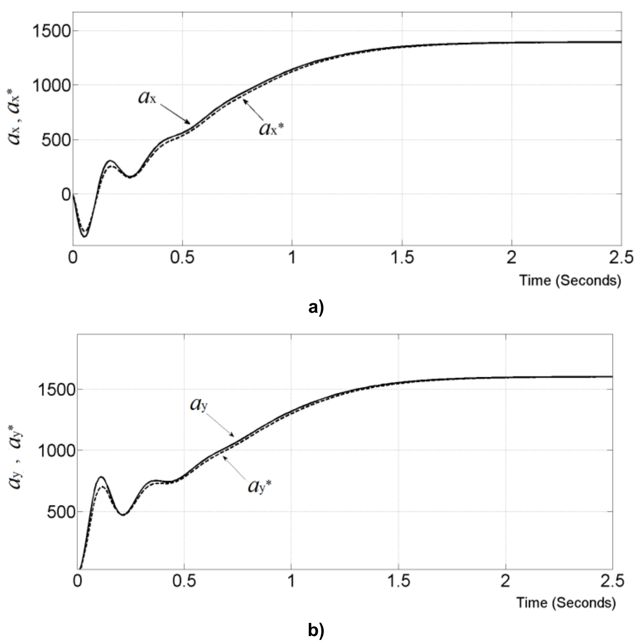

FIGURE 5. Quality of identification of the variables: a) a_x^*; b)a_y^*.

Fig. 5 shows the time-waveforms of the actual internal variables a_x and a_y obtained from the IM model, as well as their identified time courses a_x* and a_y*. From them, it is clear that the proposed identification structures have excellent dynamic properties while being principally stable. Due to accuracy of the sensed currents and voltages, which usually are sized within a range of a few percent, the deviations between the real and the identified values can be considered to be negligible, especially in dynamic states, because in the steady state these deviations ($a_x - a_x^*$ and $a_y - a_y^*$) are zero.

IV. IDENTIFICATION OF ROTOR FLUXES AND TORQUE OF THE INDUCTION MOTOR ON BASIS OF THE VARIABLES a_x^* AND a_y^*

If the signals a_x^* and a_y^* on the output of the identification scheme in Fig. 4 are already known, then the rotor flux components can be estimated using the following equations:

$$\psi_{2x}^* = \frac{1}{K_{12}}\left(\frac{\omega_g a_x^* - \omega_m a_y^*}{\omega_g^2 + \omega_m^2}\right) \tag{12}$$

$$\psi_{2y}^* = \frac{1}{K_{12}}\left(\frac{\omega_g a_y^* + \omega_m a_x^*}{\omega_g^2 + \omega_m^2}\right) \tag{13}$$

The quality of the rotor flux components identification in comparison with their actual values obtained by the simulation follows up from the time courses shown in Fig. 6.

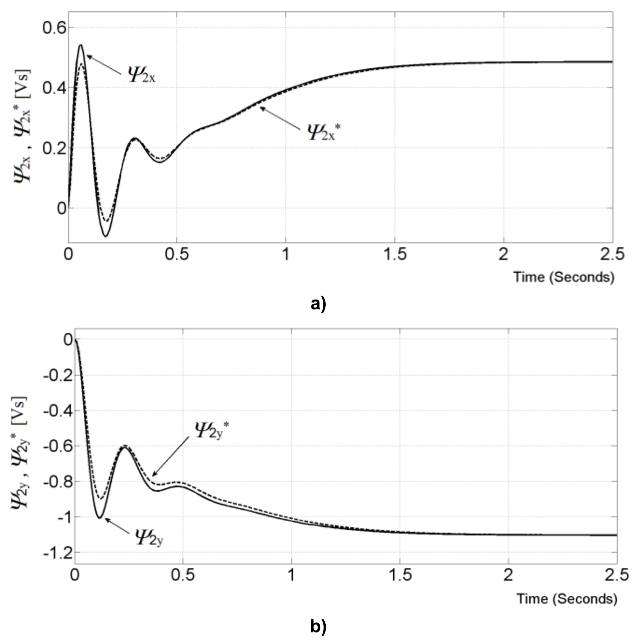

FIGURE 6. Comparison of the identified and the actual components of the IM rotor fluxes *a)*Ψ_{2x}; *b)*Ψ_{2y}.

From the figure, it can be derived, that the courses of the measured rotor flux vector components almost coincide with their identified time courses. The largest deviation in the dynamic state occurs approximately at the time $t = 0.15$ s, and its value is less than 4 %. This precision is fully satisfactory for the use in any practical control application. In the steady state, the deviation between the measured and the identified components of the flux vector is zero, which corresponds to the zero identification deviation of the variables ($a_x - a_x^*$) and ($a_y - a_y^*$) in Fig. 5.

If the individual components of the rotor flux are identified, the motor electromagnetic torque can be determined on the

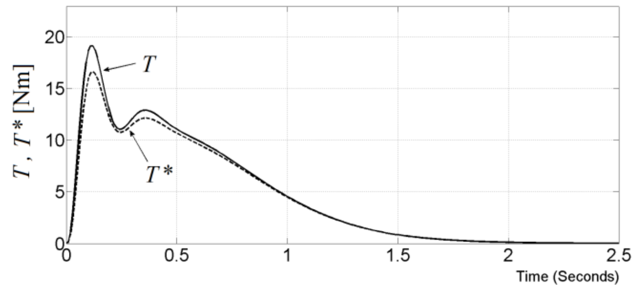

FIGURE 7. Comparison of the identified electromagnetic torque T^* and the actual torqueTof the induction motor.

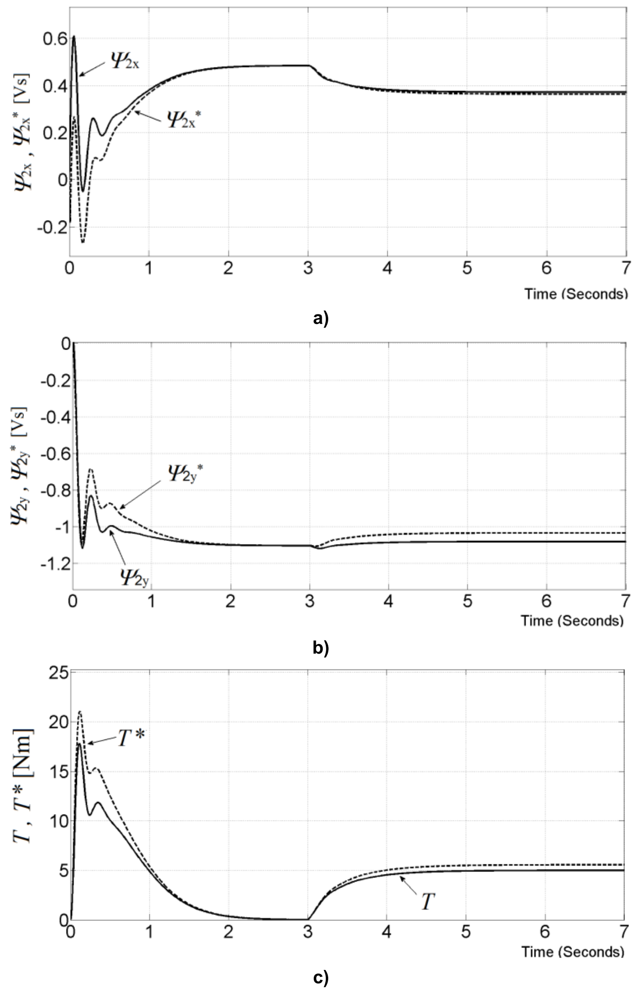

FIGURE 8. Influence of the uncertainty of the parameter ω_g (corresponding to the rotor resistance) on the quality of identification of the rotor flux components: a) Ψ_{2x}; b)Ψ_{2y}; and c) motor torque T.

basis of the equation (2) as follows:

$$T^* = \frac{Mn_p}{L_2}(\psi^*_{2x} i_{1y} - \psi^*_{2y} i_{1x}) \quad (14)$$

The identified electromagnetic torque waveform obtained by the numerical simulation based on the equation (14) is shown in the Fig. 7. The time course confirms the quality of the identification both in steady and dynamic states. In the

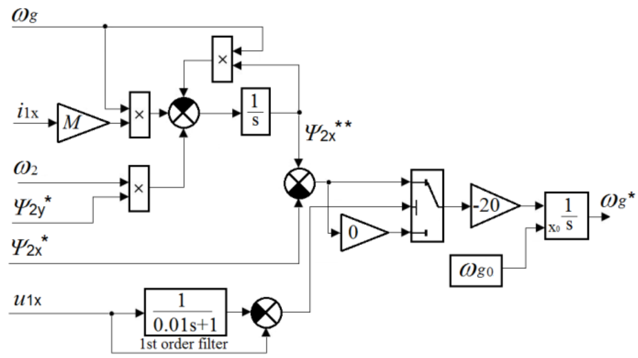

FIGURE 9. Block diagram of a continuous identification of the parameter ω_g.

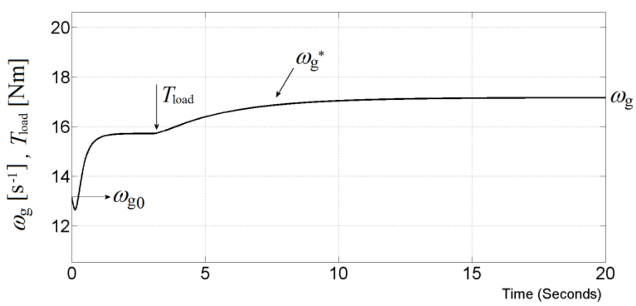

FIGURE 10. Identification of the rotor time constant ω_g.

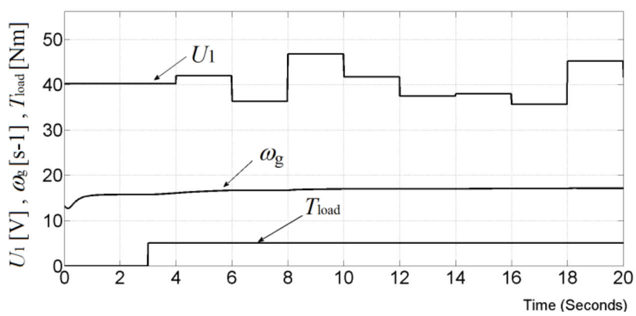

FIGURE 11. Waveforms of the motor input quantities: voltage U_1, and load torque T_{load} together with the course of parameter ω_g adaptation.

steady state, the identification deviation of the torque $(T - T^*)$ is equal to zero again. The largest identification error in the dynamic state occurs at time $t = 0.15$ s and it represents approximately 3 % of the maximum waveform range of the IM electromagnetic torque. The maximum value of the motor torque presents three times of the nominal torque here (in the analyzed case it reaches 60 Nm).

V. ADAPTATION OF IDENTIFICATION STRUCTURE TO CHANGE OF ROTOR RESISTANCE

Quality of the identification according to the proposed identification structure in Chapter 2 (Fig. 3) depends on the quality of the identification of the variables a_x and a_y, representing the feedback effect of the rotor flux on the stator current. Identification of these variables depends substantially on the

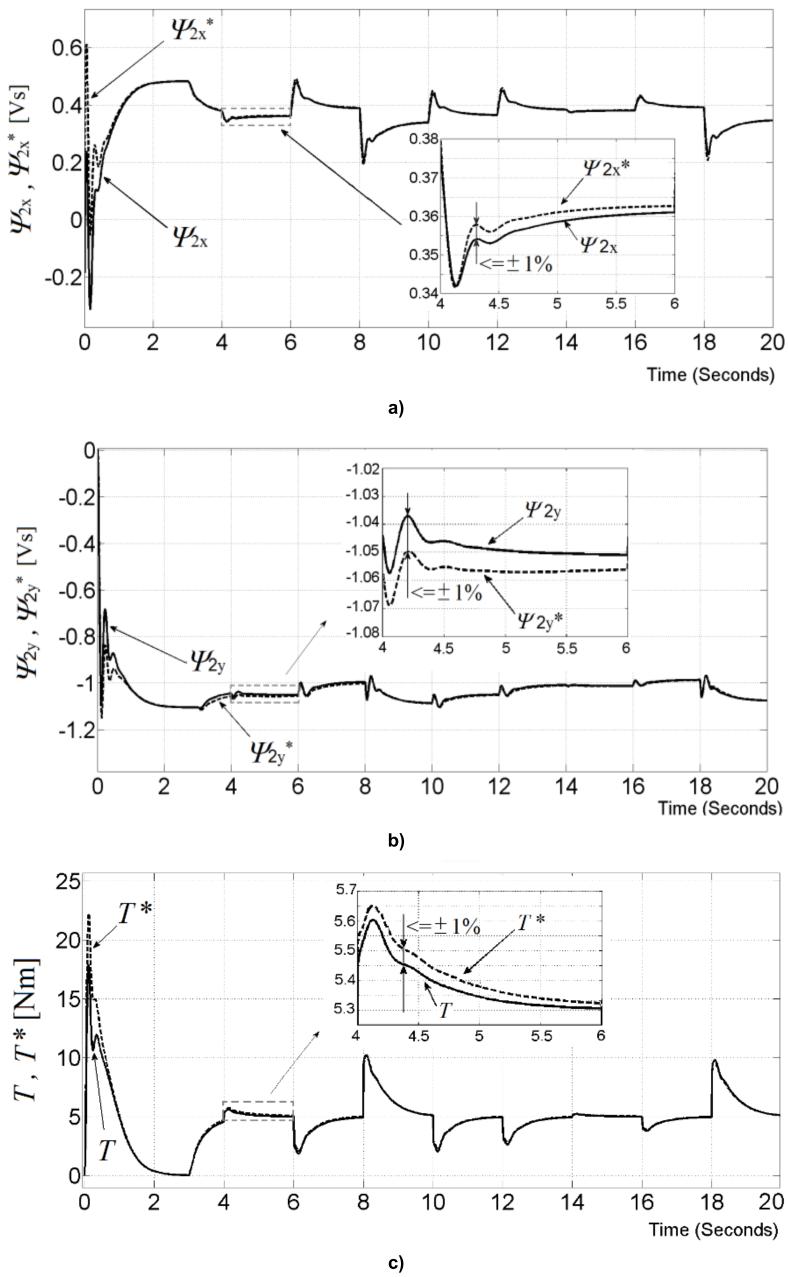

FIGURE 12. Identification of the rotor flux vector components a) Ψ_{2x}; *b)* Ψ_{2y} and c) motor torque *T* taking into account adaptation of the parameter ω_g (presenting the inverse value of the rotor resistance R_2).

exact value of the stator current circuit parameter ω_0:

$$\omega_0 = K_{11}(R_1 + \frac{M^2}{L_2}\omega_g) \tag{15}$$

In addition to the known parameters (Table 1) in equation (15), the parameter ω_g (representing a reciprocal value of the rotor time constant) is directly proportional to a non-measurable value of the rotor resistance R_2 and inductance L_2 according to (7). Furthermore, we assume, that the temperature change of the rotor resistance R_2 in different operating states of the motor has a fundamental influence on the change of the parameter ω_g. On the other side, the influence of the change of the inductance L_2 is not as significant as the change of the resistance R_2. The effect of the change of the parameter ω_g on the quality of the rotor flux identification and IM torque obtained by digital simulation is shown in Fig. 8, where the original value $\omega_g = 17.18$ s^{-1} was reduced by 30 % to $\omega_{g0} = 13.18$ s^{-1}. At the time $t = 3$ s, an additive load torque $T_{load} = 5$ Nm acts on the motor shaft representing a 25 % of the nominal torque.

Comparing the courses in the Fig. 6 and 7 with the Fig. 8, we can see, that the identification of the IM internal quantities is inaccurate, if the exact value of the rotor resistance (i.e. of the ω_g parameter, respectively) is unknown. This

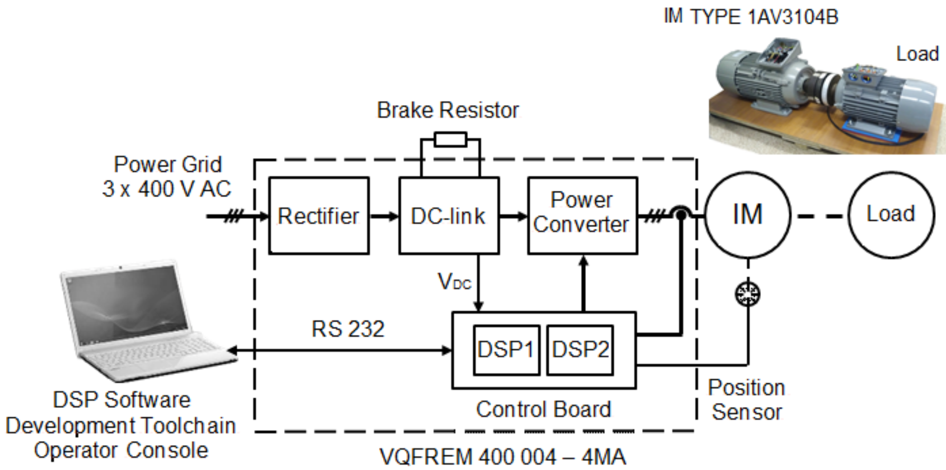

FIGURE 13. **Structure of experimental setup with VQFREM 400 004 – 4MA power converter.**

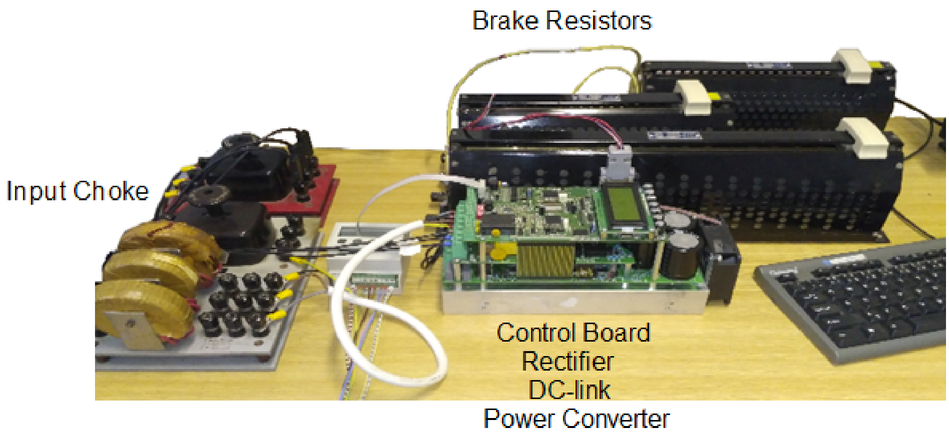

FIGURE 14. **Experimental setup with VQFREM 400 004 – 4MA converter.**

fact can be also derived from (12), (13), and (15), in which the parameter ω_g occurs. In this case, a permanent additive identification deviation occurs for all the identified quantities both in dynamic and steady states.

The exact initial value of the parameter ω_g can be checked by connecting the stator of IM to a dc voltage. In this case, the motor is at standstill, in which the cross-links between its electrical quantities ($\omega_1 = 0$, $\omega_m = 0$, $\omega_2 = 0$) do not occur and from the equations (1), (10) and (15) in steady state, one can derive:

$$0 = K_{11}\left(R_1 + \frac{M^2}{L_2}\omega_g\right) i_{1ust} - a_x^* + K_{11}U_{1x} \quad (16)$$

In this equation, only one parameter ω_g is unknown, whose calculation for particular motor parameters represents a simple task.

During the motor operation, the rotor resistance may change due to rise of the temperature, which results in a change of the parameter ω_g value. To identify the change of the rotor resistance the following considerations are made.

According to (1), the relation for derivation of the identified x-component of the rotor flux ψ_{2x}^* is:

$$\frac{d\psi_{2x}^*}{dt} = M\omega_g i_{1x} - \omega_g\psi_{2x}^* + \omega_2\psi_{2y}^* \quad (17)$$

For the initial value of the parameter $\omega_g = \omega_{g0}$ selected for identification and for the already identified value of the y-component $\psi_{2y}*$ of the rotor flux, the following equation applies:

$$\frac{d\psi_{2x}^{**}}{dt} = M\omega_{g0} i_{1x} - \omega_{g0}\psi_{2x}^* + \omega_2\psi_{2y}^* \quad (18)$$

The value of the parameter ω_g is correct, if the difference between the values of the x-components of the rotor fluxes ψ_{2x}^* and ψ_{2x}^{**} in equations (17) and (18) is equal to zero. For this reason, the block diagram shown in the Fig. 9 allows identification of the parameter ω_g, which enables a continuous setting up of its value in the equation (9) for identification of variable a_x^*, and similarly for a_y^* in the equation (10) for the fluxes ψ_{2x}^* in (12) and ψ_{2y}^* in (13).

The value of difference between the x-components of rotor flux according to the equation:

$$e_{\psi} = \psi_{2x}^{**} - \psi_{2x}^{*} \tag{19}$$

depends on change of the x-component of the stator voltage component u_{1x}. In order to avoid a negative value of the change of the u_{1x}, the block diagram has been completed by a switch causing switching this difference e_{ψ} to zero, if $\psi_{2x}^{**} - \psi_{2x}^{*} < 0$. Assuming the scalar control of the IM by the voltage U_1, for the x-component of voltage u_{1x} it is valid $U_1 = u_{1x}$, as stated in the Section 1. The ideal derivative of the voltage u_{1x} is realized by a derivative block completed by the first order filter, like it is depicted in the Fig. 9.

Fig. 10 shows the successive change of the parameter ω_g from the initial value $\omega_{g0} = 13.18\ s^{-1}$ at the time $t = 0$ s to the value of $\omega_g = 17.18\ s^{-1}$ at the time $t = 20$ s, that corresponds to 30 % increase of the rotor resistance value. Actually, in a real motor, the rotor resistance, never changes so fast within a 20-second interval, and thus the identification is dynamically fast enough.

At the time $t = 3$ s, an additive load torque fault $T_{load} = 5$ Nm acts on the motor shaft causing increase of the deviation in the equation (19), and thus a faster adaptation of the parameter ω_g to the set value ($\omega_g = 17.18\ s^{-1}$).

VI. VERIFICATION OF ROTOR FLUX AND TORQUE IDENTIFICATION OF IM FOR VARIOUS OPERATING STATES

Subsequently, the properties of the rotor flux and torque identification method are verified by the simulation in various operating states.

Let's suppose that IM operating under scalar control is supplied by the voltage U_1 (where $U_1 = u_{1x}$) and is loaded gradually stepwise by the load torque T_{load}. For this reason random step changes of the input voltage U_1 applied to the stator of the motor and a step change of its load torque T_{load} were simulated, as illustrated in Fig. 11.

At the time $t = 3$ s, a torque step of $T_{load} = 5$ Nm is applied to the motor. The stator voltage U_1 and the supply frequency jumps ω_1 are generated randomly in two-second time intervals, under maintaining the relationship $U_1/\omega_1 = const$. This operation can be considered as the worst case for the rotor resistance identification. In the praxis, this step change is not realistic.

In this case the rotor winding time constant ω_g was adapted from the initial value $\omega_{g0} = 13.18\ s^{-1}$ to the value $\omega_g = 17.18\ s^{-1}$ as shown in the Fig. 11.

Fig. 11 shows the gradual improvement of the identification of the quality of the rotor flux components and motor electromagnetic torque identification, when taking into account the adaptation of the parameter ω_g from the initial value $\omega_{g0} = 13.18\ s^{-1}$ to $\omega_g = 17.18\ s^{-1}$.

The induction motor starts its operation from non-excited state and at starting, during the magnetic flux generation, the identification error lies at its maximum. After the motor has been excited, the error lies practically within the range ± 1% of the actual value both in static and dynamic states of the identified quantities, like it follows from the time courses in the Fig. 12.

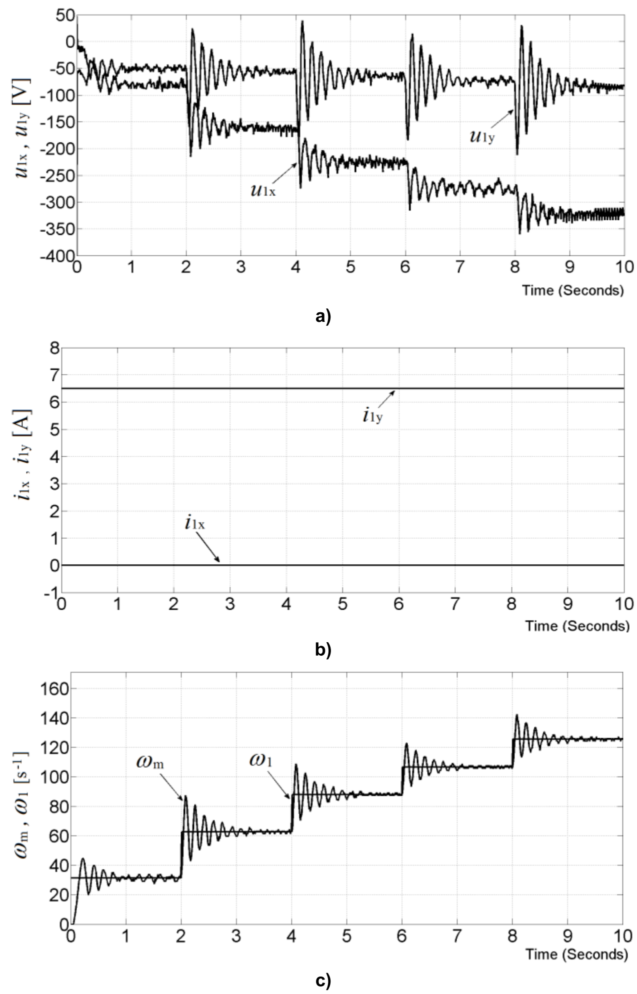

FIGURE 15. Experimentally measured values of the induction motor: a) input voltages, b) currents, c) stator frequency and motor angular speed ω_m.

VII. EXPERIMENTAL VERIFICATION OF IDENTIFICATION OF IM ELECTROMAGNETIC TORQUE OF A LABORATORY MODEL

The laboratory model for experimental verification of the above-described method of the identification of the electromagnetic torque of IM consists of a development system with the SIMOTICS GP motor TYPE 1AV3104B (three-phase squirrel-cage-motor, 400 V, 2.2 kW, 1465 RPM, 14.3 Nm) and the power converter VQFREM 400 004 – 4MA (400 V, 11 A, manufactured by VONSCH Co, Slovakia).

Fig. 13 shows the structure of the experimental setup with VQFREM 400 004 – 4MA power converter.

The power converter is controlled by two processors DSP TMS320F2406. The first one performs the data acquisition from the sensors ADC and IRC sensing stator currents, dc-link voltage and rotor position. It also performs Clarke and Park transformations and controls the stator currents in the d-q co-ordinate system.

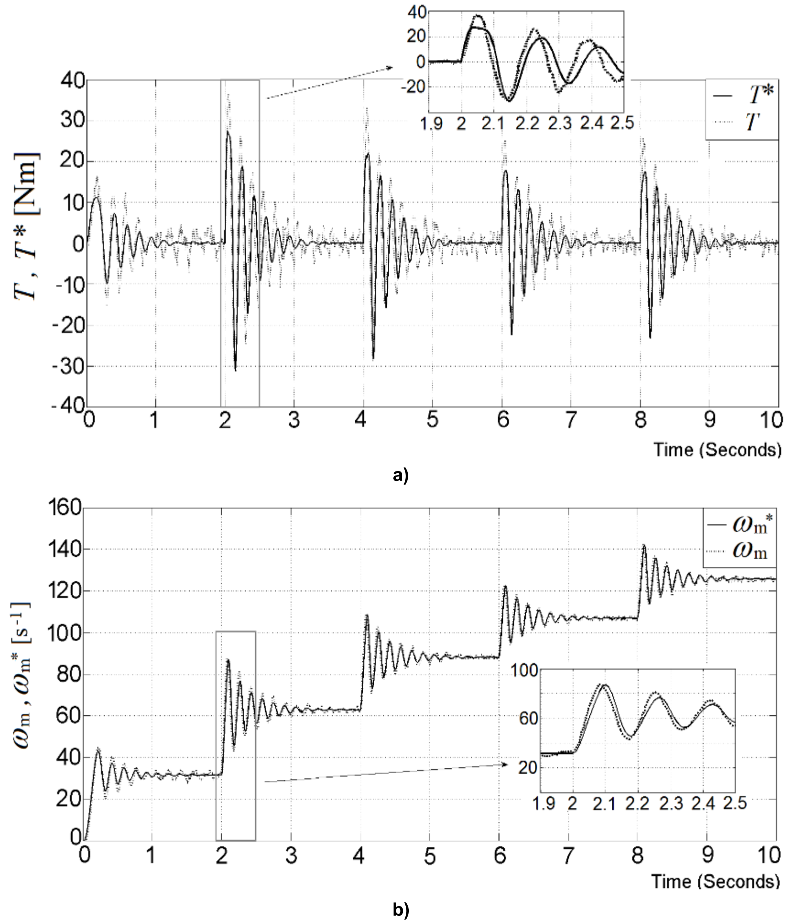

FIGURE 16. Comparison of experimentally measured and identified motor quantities (a) motor torques: T (real) and T* (identified) b) angular speeds: ω_m (real) and ω_m^* (identified).

The second processor contains PI type current controllers to maintain the stator current components i_{1x} and i_{1y} on required values. The sample time for the current control loops is 50μs. The second processor also implements the speed control scheme and communicates with the higher level controller/operator console, which is also a personal computer.

The sample time for calculations in the second processor is 1 ms. Control algorithms are written in the assembler language for the first DSP and in C language for the second processor. For programing DSPs the Texas Instruments software development toolchain is used.

The structure of the converter itself is shown in Fig. 14.

To verify the quality of the IM torque identification, a stepwise increase of the induction motor angular speed ω_1 from zero was chosen while components of the stator current vector were maintained at constant setpoint values for corresponding PI controllers. Measured values of the input voltages and currents, the stator frequency and the angular speed of IM in this mode are depicted in Fig. 15.

Fig. 16 shows comparison of time courses of the measured and identified motor dynamic torque. Considering that stator voltage and current measurements were sensed with 8-bit accuracy and the real value of the rotor resistance was determined from static measurement of the IM at no load and short-circuit had accuracy of about 3%, the error of identification is larger than it was ideally assumed during the digital simulation.

Nevertheless, during dynamic states, it does not exceed 5% of the value of the identified quantities, as shown in the Fig. 16a and Fig. 16b. As showed in the detailed pictures, the slight phase shift between the time courses of the identified and measured quantities is caused by non-zero dynamics of real identification of IM internal variables a_x and a_y, which according to Fig. 3 is realized by the second order dynamic system.

VIII. DISCUSSION AND CONCLUSION

In this article, we have proposed an original method of a high quality identification of rotor fluxes and electromagnetic torque of induction motor (IM).

The results of the published procedures and methods for solving the problem of identification of the IM internal quantities, presented in the Chapter 1, show the following drawbacks:

1) High dependence of the identification on unknown parameter disturbances, especially on the rotor resistance [17]–[20].

2) The use of complex non-linear observers and related problems with demonstration of global stability (the stability within the whole range of parameter changes) of the identification method [16], [21]–[26].
3) Dependence of the use of identification methods on the range of operational speeds of IM. Some methods are suitable for low speeds [19], others for high ones [17], [18], [29].
4) High computational complexity of the identification methods using complex observers in connection with neural networks (demands on the sampling time of computational resources), [27]–[29].

The identification method proposed in this article has practically none of the disadvantages mentioned. It is based on identification of the feedback influence on stator currents with on-line adaptation of the rotor resistance. For its operation, the method utilizes commonly available measurable quantities of IM: stator currents and mechanical angular rotor speed.

The quality and efficiency of the proposed identification method was verified by digital simulation in the MATLAB/ Simulink program and then verified by experimental measurements using laboratory model with IM.

The results from numerical simulation (Fig. 5 – Fig. 7) show, that the proposed method identifies at sufficient accuracy the feedback effect of the rotor flux vector on the stator current vector of the motor in steady and dynamic states.

To determine this feedback effect a reduced model of the IM stator is used. The resulting identification circuit of the feedback effect of the rotor magnetic fluxes on the stator currents presents a simple linear circuit of the 2nd order (Fig. 4). Its stability is guaranteed by the negative real part of the poles of its transfer function (11). It follows that in case of the presented novel method, unlike the said methods, it is not necessary to investigate and prove the stability of the proposed identification method.

Based on the well-identified feedback effect of the rotor fluxes on the stator currents, it is further possible to calculate components of the internal rotor magnetic flux and electromagnetic torque. The accuracy of the identification of these quantities depends on value of the IM rotor resistance expressed by the parameter ω_g according to (7), which is generally not known precisely and varies during operation (e.g. due to the temperature change).

The article also suggests a method of an on-line identification of the rotor resistance during IM operation and the influence of parameter ω_g adaptation on the improvement of the identified IM quantities (Fig. 12). For this reason, the proposed identification structure can be considered as a robust one against changes in rotor resistance.

Experimental measurements performed using laboratory model with IM for verification of the electromagnetic torque identification confirm the quality of the proposed identification structure, its stability and robustness against changes of the rotor resistance (Fig. 15, Fig. 16). The identification method works reliably within the whole range of IM operating speeds, practically from zero (because when the motor is excited, the effect of rotor magnetic fluxes in the feedback to stator currents appears even at zero speed). This fact can be also derived from the equations (9) and (10) for the value of the parameter $\omega_m = 0$.

The implementation of the experimental measurements also confirmed that the proposed method of the identification not only leads to a relatively easy-to-implement system, but also it is applicable at sufficiently fast sampling time of measured stator quantities of IM at a standard accuracy of their measurement (the sampling time of the detection circuit was 1 ms).

Due to the fact that all input and output parameters of the proposed identification structure are sensed directly on the IM, this identification structure is in the principle independent of the superior control structures. For the reason mentioned above, the identified internal quantities of the motor, such as the individual components of its rotor flux and the internal electromagnetic torque of the motor can be used for various commonly used methods of a dynamic drive control like scalar control, different types of a vector control and a direct torque control of the induction motor.

APPENDIX

See Table 1.

TABLE 1. Induction motor parameters and variables.

Symbol	Quantity	Value
P_N	nominal power	3 kW
U_{1N}	nominal voltage	220 V
I_{1N}	nominal current	6.9 A
T_N	nominal torque	20 Nm
n_N	nominal revolution	1430 rev.min^{-1}
n_p	number of poles	2
J	moment of inertia	0.1 kgm^2
L_2	inductance defined by equation (8)	0.14 H
R_1	stator phase resistance	1.8 Ω
R_2	rotor phase resistance	1.85 Ω
L_m	main inductance	0.202 H
M	mutual inductance defined by equation (6)	0.13 H
$L_{s1}=L_{s2}$	leakage inductance	0.0086 H
K_{11}	parameter defined by equation (3)	59.35 H^{-1}
K_{12}	parameter defined by equation (4)	-56.93 H^{-1}
ω_0	parameter defined by equation (5)	207.9 s^{-1}
ω_g	rotor winding time constant – equation (7)	13.18 s^{-1}
ω_1	angular frequency of stator voltage vector	[rad/s]
ω_m	rotor mechanical speed	[rad/s]
ω_2	angular slip frequency (ω_1-ω_m)	[rad/s]
T	motor torque	[Nm]
T_{load}	load torque	[Nm]

REFERENCES

[1] Y. Liu, J. Zhao, R. Wang, and C. Huang, "Performance improvement of induction motor current controllers in field-weakening region for electric vehicles," *IEEE Trans. Power Electron.*, vol. 28, no. 5, pp. 2468–2482, May 2013.

[2] C. Sun, S. Ai, L. Hu, and Y. Chen, "The development of a 20MW PWM driver for advanced fifteen-phase propulsion induction motors," *J. Power Electron.*, vol. 15, no. 1, pp. 146–159, Jan. 2015.

[3] M. H. Holakooie, A. Taheri, and M. B. B. Sharifian, "MRAS based speed estimator for sensorless vector control of a linear induction motor with improved adaptation mechanisms," *J. Power Electron.*, vol. 15, no. 5, pp. 1274–1285, Sep. 2015.

[4] J. Vittek, P. Butko, B. Ftorek, P. Makýš, and L. Gorel, "Energy near-optimal control strategies for industrial and traction drives with A.C. motors," *Math. Problems Eng.*, vol. 2017, Jan. 22017, Art. no. 1857186, doi: 10.1155/2017/1857186.

[5] B. K. Bose, *Modern Power Electronics and AC Drives*. Upper Saddle River, NJ, USA: Prentice-Hall, 2002.

[6] P. Brandstetter, M. Kuchar, H. H. Vo, and C. S. Thien Dong, "Induction motor drive with PWM direct torque control," in *Proc. 18th Int. Sci. Conf. Electr. Power Eng. (EPE)*, Desnou, Czech Republic, May 2017, pp. 1–5, doi: 10.1109/EPE.2017.7967268.

[7] M. P. Kazmierkowski and A. B. Kasprowicz, "Improved direct torque and flux vector control of PWM inverter-fed induction motor drives," *IEEE Trans. Ind. Electron.*, vol. 42, no. 4, pp. 344–350, 1995.

[8] Y. Zhang, Y. Bai, and H. Yang, "A universal Multiple-Vector-Based model predictive control of induction motor drives," *IEEE Trans. Power Electron.*, vol. 33, no. 8, pp. 6957–6969, Aug. 2018.

[9] Y. Zhang, H. Yang, and B. Xia, "Model-predictive control of induction motor drives: Torque control versus flux control," *IEEE Trans. Ind. Appl.*, vol. 52, no. 5, pp. 4050–4060, Sep. 2016.

[10] P. Fedor, D. Perdukova, K. Kyslan, and V. Fedak, "Stable and robust controller for induction motor drive," in *Proc. IEEE 18th Int. Power Electron. Motion Control Conf. (PEMC)*, Budapest, Hungary, Aug. 2018, pp. 764–769, doi: 10.1109/EPEPEMC.2018.8521978.

[11] M. R. Hachicha, N. Ben Hadj, M. Ghariani, and R. Neji, "Finite element method for induction machine parameters identification," in *Proc. 1st Int. Conf. Renew. Energies Veh. Technol.*, Hammamet, Tunisia, Mar. 2012, pp. 490–496.

[12] S. Williamson, L. H. Lim, and A. C. Smith, "Transient analysis of cage-induction motors using finite-elements," *IEEE Trans. Magn.*, vol. 26, no. 2, pp. 941–944, Mar. 1990.

[13] S. Khamehchi, E. Molsa, and M. Hinkkanen, "Comparison of standstill parameter identification methods for induction motors," in *Proc. IEEE 9th Int. Symp. Sensorless Control Electr. Drives (SLED)*, Helsinki, Finland Sep. 2018, pp. 156–161.

[14] S.-H. Lee, A. Yoo, H.-J. Lee, Y.-D. Yoon, and B.-M. Han, "Identification of induction motor parameters at standstill based on integral calculation," *IEEE Trans. Ind. Appl.*, vol. 53, no. 3, pp. 2130–2139, May 2017.

[15] J. Chen and J. Huang, "Stable simultaneous stator and rotor resistances identification for speed sensorless IM drives: Review and new results," *IEEE Trans. Power Electron.*, vol. 33, no. 10, pp. 8695–8709, Oct. 2018.

[16] J. Tang, Y. Yang, F. Blaabjerg, J. Chen, L. Diao, and Z. Liu, "Parameter identification of inverter-fed induction motors: A review," *Energies*, vol. 11, p. 21, Apr. 2018, doi: 10.3390/en11092194.

[17] O. Wallseheid, M. Schenke, and J. Bocker, "Improving torque and speed estimation accuracy by conjoint parameter identification and unscented Kalman filter design for induction machines," in *Proc. 21st Int. Conf. Electr. Mach. Syst. (ICEMS)*, Jeju, South Korea Oct. 2018, pp. 1181–1186, doi: 10.23919/ICEMS.2018.8549514.

[18] H. Becker, "Dynamisch hochwertige Drehzahlregelung einer umrichtergespeisten Asynchronmaschine (in German)," in *Proc. RegelungstechnischePraxis und Prozessrechentechnik*, Oct. 1973, pp. 217–221.

[19] F. Elaschke and K. Böhm, "Verfahren der Felderfassung bei der Regelung stromrichtergespeister Asynchronmaschinen," *IFAC Proc. Volumes*, vol. 7, pp. 635–649, Oct. 1974.

[20] V. Arunachalam, H. Srinivasan, and A. Muthuramalingam, "A novel NN based rotor flux MRAS to overcome low speed problems for rotor resistance estimation in vector controlled IM drives," *Frontiers Energy*, vol. 10, no. 4, pp. 382–392, Dec. 2016, doi: 10.1007/s11708-016-0421-y.

[21] P. L. Jansen and R. D. Lorenz, "A physically insightful approach to the design and accuracy assessment of flux observers for field oriented induction machine drives," *IEEE Trans. Ind. Appl.*, vol. 30, no. 1, pp. 101–110, Dec. 1994, doi: 10.1109/28.273627.

[22] R. Bojoi, G. Griva, and F. Profumo, "Field oriented control of dual three-phase induction motor drives using a luenberger flux observer," in *Proc. IEEE Ind. Appl. Conf. Forty-First IAS Annu. Meeting*, Tampa, FL, USA Oct. 2006, pp. 1253–1260, doi: 10.1109/IAS.2006.256692.

[23] D. J. Atkinson, P. P. Acarnley, and J. W. Finch, "Observers for induction motor state and parameter estimation," *IEEE Trans. Ind. Appl.*, vol. 27, no. 6, pp. 1119–1127, Aug. 1991, doi: 10.1109/28.108463.

[24] C. M. Verrelli, A. Savoia, M. Mengoni, R. Marino, P. Tomei, and L. Zarri, "On-line identification of winding resistances and load torque in induction machines," *IEEE Trans. Control Syst. Technol.*, vol. 22, no. 4, pp. 1629–1637, Jul. 2014, doi: 10.1109/TCST.2013.2285604.

[25] S. Ho Jeon, K. Kyo Oh, and J. Young Choi, "Flux observer with online tuning of stator and rotor resistances for induction motors," *IEEE Trans. Ind. Electron.*, vol. 49, no. 3, pp. 653–664, Jun. 2002, doi: 10.1109/TIE.2002.1005393.

[26] M. Bahloul, L. Chrifi-Alaoui, M. Souissi, S. Drid, and M. Chaabane, "Robust flux and load torque estimation in induction machine," in *Proc. Eur. Control Conf. (ECC)*, Linz, Austria, Jul. 2015, pp. 666–671, doi: 10.1109/ECC.2015.7330618.

[27] M. Bahloul, L. Chrifi-Alaoui, A. N. Vargas, M. Chaabane, and S. Drid, "Online robust estimation of flux and load torque in induction motors," *Int. J. Adv. Manuf. Technol.*, vol. 94, nos. 5–8, pp. 2703–2713, Feb. 2018, doi: 10.1007/s00170-017-1049-8.

[28] A. Venkadesan, S. Himavathi, and A. Muthuramalingam, "Performance comparison of neural architectures for on-line flux estimation in sensor-less vector-controlled IM drives," *Neural Comput. Appl.*, vol. 22, nos. 7–8, pp. 1735–1744, Jun. 2013, doi: 10.1007/s00521-012-1107-y.

[29] S.-W. Gao, "Stator flux linkage observer based on RBF neural network for IM," in *Proc. 3rd Int. Conf. Intell. Netw. Intell. Syst.*, Shenyang, China, Nov. 2010, pp. 152–155, doi: 10.1109/ICINIS.2010.81.

[30] K.-Y. Chen, W.-H. Yang, and R.-F. Fung, "System identification by using RGA with a reduced-order robust observer for an induction motor," *Mechatronics*, vol. 54, pp. 1–15, Oct. 2018, doi: 10.1016/j.mechatronics.2018.07.001.

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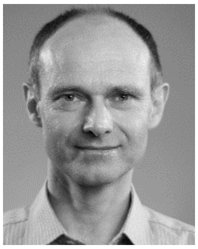

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