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Magnetization Curve Identification of Vector-Controlled Induction Motor at Low-Load Conditions

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In many drives the induction motor torque is transferred to the load through complex mechanical connections which produce an initial load. In such cases, to identify the magnetization curve by the no-load test it is necessary to decouple the motor from the transfer gear what is expensively and time-consuming. In an attempt to avoid the motor mechanical decoupling in the case of hoist crane systems, in this paper a method is proposed for the magnetization curve identification of the indirect rotor flux oriented (IRFO) induction machine at low-load conditions. Two flux controllers with estimated feedback quantities are added to the basic structure of the IRFO control drive. One of these controllers is used to control desired operating point on the magnetization curve and the other for on-line adaption of the reference frame position. The proposed algorithm is implemented in a commercial drive as a part of the self-commissioning module. Afterward, the magnetization curve of the 22 kW induction motor was identified and compared with the one obtained by the no-load test. The method sensitivity to the load torque and the transient inductance has also been considered.

Key words: Induction motors, Magnetization curve, Parameter estimation, Vector control

Određivanje krivulje magnetiziranja vektorski upravljanog asinkronog motora kod malih opterećenja. Brojni su primjeri pogona u kojima se moment asinkronog stroja prenosi na radni mehanizam preko mehaničkih reduktora koji mogu predstavljati značajno opterećenje motoru i u režimu neopterećenog pogona. Da bi se u takvim slučajevima odredila krivulja magnetiziranja iz pokusa praznog hoda nužno je odspojiti motor od reduktora što je skupo i vremenski zahtjevno. U pokušaju da se izbjegne mehaničko odspajanje motora kod dizaličnih sustava, u ovom je radu predložena metoda za određivanje krivulje magnetizranja vektorski upravljanog asinkronog motora kod malih iznosa opterećenja. Dva regulatora ulančenog toka s estimiranim veličinama u granama povratne veze dodani su osnovnoj strukturi vektorski upravljanog pogona. Jedan od regulatora koristi se za upravljanje željenom radnom točkom na krivulji magnetiziranja dok se drugi koristi za podešavanje referentnog položaja vektora toka rotora. Predloženi algoritam implementiran je u komercijalnom uređaju kao jedan od samopodešavajućih modula koji se koriste prilikom puštanja u rad pogona. Nakon toga je opisanim postupkom određena krivulja magnetiziranja 22 kW asinkronog motora i uspoređena s krivuljom dobivenom iz klasičnog pokusa praznog hoda. Također je razmotrena i osjetljivost metode na iznos momenta opterećenja i tranzijentnog induktiviteta.

Ključne riječi: asinkroni motori, krivulja magnetiziranja, estimacija parametara, vektorsko upravljanje

1 INTRODUCTION

Among several different configurations for the field orientation control of induction machines, the indirect rotor flux oriented (IRFO) control is favorable because of its simple structure and excellent decoupling characteristics. When the speed and the current are accurately measured, the performance of the system is sensitive to the rotor time constant only [1]. This constant depends on the rotor temperature and on saturation of the main magnetic circuit of the machine. Changes of saturation are much faster than the heating process and because of that the on-line estimation of the rotor resistance and the magnetizing inductance are commonly performed separately. When the motor operates in the field weakening region, the magnetizing inductance is changed according to the magnetization curve. Although the magnetizing inductance can be estimated on-line by applying a tuning scheme, for instance like that one proposed in [2], the on-line estimation based on the magnetization curve makes the control system more robust. In addition, the magnetization curve must be known if the maximum torque over the whole speed range should be reached [3,4].

Early works dealing with a self-commissioning of PWM-inverter-fed induction motor drives [5,6] introduced a procedure for identifying the motor parameters at standstill. But the magnetizing current and the mutual inductance are identified by the no-load test. A method for the magnetization curve identification at standstill proposed in [7] is based on the motor excitation with different voltage steps and measuring the current responses. The accuracy of the method deteriorates bellow a certain applied voltage, so that this approach is not accurate enough particularly if a large size machine is under the test. In [8] a solution for the magnetization curve identification is proposed which is based on one-test only. The ac current is injected into the machine at standstill. The reference voltage response is used for the magnetization curve identification, and estimation procedure is performed off-line. The main drawback of this method is that it is not suitable for the self-commissioning scheme. The method presented in [9], which deals with the parameter identification of the analytical inverse magnetization curve implemented in the control system, requires that the machine is not coupled to the load and that the voltage measuring equipment is available. A newer paper dealing with the self-commissioning at standstill does not deal with the magnetization curve at all [10].

In this paper a method is proposed for the magnetization curve identification at motor running with a low-load at the shaft. The proposed method is suitable in cases where the motor torque is transferred to the load through complex mechanical connections which produce an initial load, so that the no-load test is not possible. Such cases are, for example, rolling mill plants, paper winding processes, elevators, hoist crane systems, etc. The method is based on applying two flux controllers with estimated feedback quantities, which are added to the basic control structure of the IRFO drive. One of these controllers is used to control desired operating point on the magnetization curve and the other for on-line adaption of the reference frame position.

2 METHOD DESCRIPTION

2.1 Machine Voltage Equations

The induction machine voltage equations in terms of complex space vector quantities, in the arbitrary reference frame, are

$$\bar{u}_s = R_s \bar{i}_s + \frac{d\bar{\psi}_s}{dt} + j\omega_k \bar{\psi}_s, \tag{1}$$

$$0 = R_r \bar{i}_r + \frac{d\bar{\psi}_r}{dt} + j\left(\omega_k - \omega\right) \bar{\psi}_r, \tag{2}$$

where ω and ω_k is the angular speed of the rotor and the reference frame, respectively, and the meaning of the other quantities can be deduced from Figure 1.

The flux linkages are expressed

$$\bar{\psi}_s = L_s \bar{i}_s + L_m \bar{i}_r, \tag{3}$$

$$\bar{\psi}_r = L_m \bar{i}_s + L_r \bar{i}_r, \tag{4}$$

where L_s, L_r and L_m are the stator and rotor inductances and the magnetizing inductance, respectively. In the case of the IRFO control it is convenient to select the stator current and the rotor flux as the state variables. After introducing the space vector of the rotor magnetizing current defined as

$$\bar{i}_{mr} = \frac{L_r}{L_m} \bar{i}_r + \bar{i}_s, \tag{5}$$

relations (3) and (4) become

$$\bar{\psi}_s = L_s' \bar{i}_s + k_r L_m \bar{i}_{mr}, \tag{6}$$

$$\bar{\psi}_r = L_m \bar{i}_{mr}, \tag{7}$$

where $k_r = \frac{L_m}{L_r}$ is the rotor magnetic coupling factor, $L_s' = \sigma L_s$ is the stator transient inductance, and $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ is the total leakage coefficient.

By substituting (6) into (1) and (7) into (2) the voltage equations can be rewritten as follows

$$\bar{u}_s = R\bar{i}_s + L_s' \frac{d\bar{i}_s}{dt} + j\omega_k \left(L_s' \bar{i}_s + k_r L_m \bar{i}_{mr}\right) + k_r L_m \frac{d\bar{i}_{mr}}{dt}, \tag{8}$$

$$0 = k_r^2 R_r \left(\bar{i}_{mr} - \bar{i}_s\right) + j\left(\omega_k - \omega\right) k_r L_m \bar{i}_{mr} + k_r L_m \frac{d\bar{i}_{mr}}{dt}. \tag{9}$$

The above equations suggest the inverse-Γ dynamic equivalent circuit shown in Figure 1, which is convenient for the IRFO control.

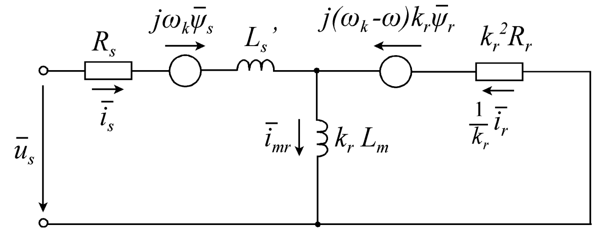

Fig. 1. Inverse-Γ dynamic equivalent circuit of induction machine

In the rotor flux-oriented reference frame, $\omega_k = \omega_{mr}$, the *d*-axis is aligned with the vector $\bar{\psi}_r$. Thus, $\bar{\psi}_r = \psi_{dr} + j0$ and $\bar{i}_{mr} = i_{mr} + j0$. Accordingly, the *d*- and *q*-axis flux linkages are expressed from (6) and (7) as

$$\psi_{ds} = L_s' i_{ds} + k_r L_m i_{mr}, \tag{10}$$

$$\psi_{qs} = L_s' i_{qs}, \tag{11}$$

$$\psi_{dr} = L_m i_{mr}, \tag{12}$$

$$\psi_{qr} = 0. \tag{13}$$

According to (10), at steady-state ($i_{mr} = i_{ds} = I_{ds}$) the d-axis stator flux linkage may be expressed as $\Psi_{ds} = L_s I_{ds}$, and the saturation characteristic $\Psi_{ds}(I_{ds})$ practically does not depend on the load condition [3]. This characteristic can be identified at low-load condition by means of a procedure proposed in this paper. When the characteristic $\Psi_{ds}(I_{ds})$ is known, the magnetization curve $\Psi_m(I_{ds})$ can be obtained using the following relation for the mutual flux linkages

$$\Psi_m = \Psi_{ds} - L_{\sigma s} I_{ds}, \tag{14}$$

where the stator leakage inductance may be approximated as $L_{\sigma s} = \frac{L_s'}{2}$.

2.2 Formulation of Estimation Models

To identify the magnetization curve according to the procedure proposed in this paper, two flux controllers are added to the basic control structure of the IRFO drive (Figure 2). One of them is used for the d-axis stator flux linkage control and the other for the q-axis rotor flux linkage control. The last one is used for a tuning of the reference frame position. According to (13) the reference ψ_{qr}^* is set to zero, and ψ_{ds}^* is changed during the identification process in order to set given operating point on the magnetization curve. The flux estimator computes the feedback quantities $\hat{\psi}_{ds}$ and $\hat{\psi}_{qr}$ which are compared with the corresponding references. Although the control system shown in Figure 2 exhibits similarities to a model reference adaptive control (MRAC) [11], it differs from the MRAC because the reference quantities are predetermined, but not calculated with respect to the machine model.

The output of the d-axis flux controller is the d-axis stator current reference. The rotor magnetizing current is related with the d-axis stator current through the dynamic equation that is derived as the real part of (9) considering $\omega_k = \omega_{mr}$ and $\bar{i}_{mr} = i_{mr} + j0$

$$T_r \frac{di_{mr}}{dt} + i_{mr} = i_{ds}, \tag{15}$$

where $T_r = \frac{L_r}{R_r}$ is the rotor time constant. As one can see in Figure 2, the above equation is implemented in control algorithm as the first order element.

The output of the q-axis rotor flux controller is the estimated value of slip gain ($\hat{K}_s$). According to the expression for the angular velocity of the rotor flux reference frame (16), derived from imaginary part of (9), $\frac{1}{T_r}$ represents the theoretical value of the slip gain.

$$\omega_{mr} = \omega + \frac{1}{T_r}\frac{i_{qs}}{i_{mr}} \tag{16}$$

The error signal of the q-axis rotor flux controller is multiplied by the sign of the q-axis current in order to make possible estimation process independent on the torque and speed direction.

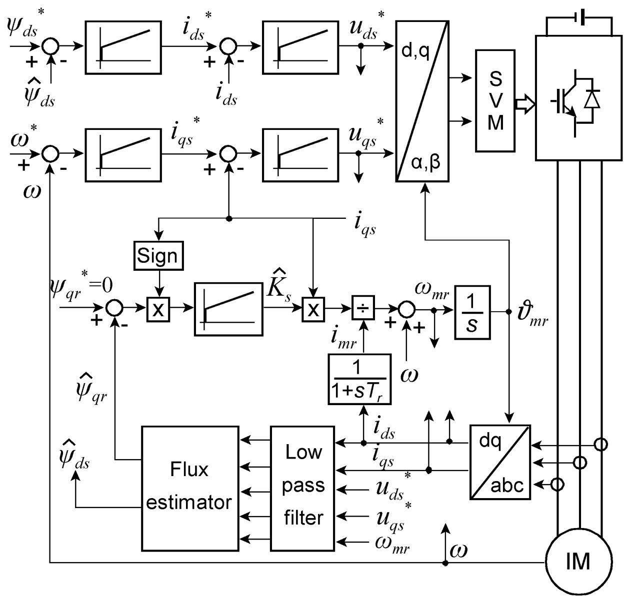

Fig. 2. System block diagram

The estimation models for the feedback quantities are formulated based on the stator steady-state voltage equations in the d- and q-axis and the flux linkage equations (6), (7) and (10)-(13). In the rotor flux reference frame, the d- and q-axis voltage equations derived from (8) at steady-state are

$$U_{ds} = R_s I_{ds} - \omega_{mr} L_s' I_{qs}, \tag{17}$$

$$U_{qs} = R_s I_{qs} + \omega_{mr}\left(L_s' I_{ds} + k_r L_m I_{mr}\right). \tag{18}$$

The d-axis stator flux linkage estimation model is formulated from (18) by taking into account (10). This yields

$$\hat{\psi}_{ds} = \frac{u_{qs}^* - R_s i_{qs}}{\omega_{mr}}. \tag{19}$$

The stator resistance in above equation is supposed to be identified by means of the standstill test.

The q-axis rotor flux linkage estimation model is derived by combining (17) with (6) and (7)

$$\hat{\psi}_{qr} = \frac{1}{k_r}\left(\frac{R_s i_{ds} - u_{ds}^*}{\omega_{mr}} - L_s' i_{qs}\right). \tag{20}$$

Among various estimation models for on-line adaption of the rotor flux reference frame discussed in [11], the selected one has the highest sensitivity to the reference frame detuning at low-load conditions. In (20), as it was the case for estimation model (19), an estimation of the stator resistance and the transient inductance is supposed to be performed by means of the standstill tests. However, it should be pointed out that the flux estimation has low sensitivity to the stator resistance at the two-third of rated frequency at which the identification of the magnetization curve is performed.

2.3 Initialization Module

The control and self-commissioning software is structured in a modular concept, and the initialization module is executed in the very beginning. In this module the stator (R_s) and rotor (R_r) resistance and the transient inductance (L'_s) are first determined by identification at standstill. To identify the rotor resistance and the transient inductance, the single-phase excitation test with injected ac current has been performed and a procedure similar like that discussed in [12] has been applied. For this purpose the PI current controllers are used and their parameters are selected based on the magnitude optimum approach. In accordance with this approach following relation for the controller integral time is obtained

$$T_i = \frac{L'_s}{R_s + k_r^2 R_r}, \tag{21}$$

while the controller proportional gain is given by

$$K_i = \frac{L'_s}{2T_{eq}}, \tag{22}$$

where T_{eq} is the equivalent time constant representing PWM inverter and signal processing time delays. The same parameters are used for both the *d*- and *q*-axis control loops.

An initial value of the magnetizing current may be estimated based on the reactive power balance (see Figure 1) at the nominal operating point as follows

$$I_{mi} = \frac{U_N I_N \sin\varphi_N - \sqrt{3}\cdot 2\pi f_N L'_s I_N^2}{U_N - \sqrt{3}\cdot 2\pi f_N L'_s I_N \sin\varphi_N}. \tag{23}$$

After the initial value of the rotor inductance has been estimated as

$$L_{ri} \approx L_{si} = \frac{1}{\sqrt{3}} \frac{U_N}{2\pi f_N I_{mi}}, \tag{24}$$

the initial value of the rotor time constant may be calculated as follows

$$T_{ri} = \frac{L_{ri}}{R_r}. \tag{25}$$

Based on the initial value of the rotor time constant, the flux controller can be set using the magnitude optimum approach. It should be noted that the same flux controller is used for both the normal operation (standard flux controller) and the identification procedure (*d*-axis flux controller), but the reference and feedback signal are different for these two mode of operation.

Finally, the inertia constant is determined and the speed controller and the *q*-axis flux controller are set. The drive with a low-load torque (T_{L0}) is accelerated from standstill to a given value of the rotor speed (ω_a) by applying constant motor torque. This is performed with the current regulators operating in the field orientation mode. The reference i^*_{mr} is set on the peak value of the initial magnetizing current ($i^*_{mr} = \sqrt{2}I_{mi}$), and i^*_{qs} is set to 60% of the motor rated current. By choosing $i^*_{qs} \geq 0.6$ p.u. it is avoided that an error in estimation of T_{L0} significantly influence on the speed controller setting. If the acceleration torque is constant, the speed rises linearly. Then the inertia constant may be calculated from the acceleration time (t_a) and the speed ω_a as follows

$$H = \frac{1}{2}\frac{\left(k_r L_m i^*_{mr} i^*_{qs} - T_{L0}\right)\ t_a}{\omega_a}. \tag{26}$$

In the above relation L_m and k_r are calculated from the stator inductance estimated by means of (24) and from the identified value of the transient inductance: $L_m = L_s - \frac{L'_s}{2}, k_r = \frac{L_m}{L_r}$.

The speed controller is set using the symmetrical optimum approach. The same approach is used to set the *q*-axis rotor flux controller. The time constant of this controller is bigger than the one of the speed controller due to the low-pass filtering which is applied to feedback signal.

To identify the magnetization curve, first, the drive is accelerated to two-third of the rated speed without an action of the additional flux controllers. The rotor magnetizing current reference is set on the peak value of the initial magnetizing current. When the speed reference is reached, the standard flux controller is turned off and both the *d*- and *q*-flux controllers are turned on. The reference ψ^*_{ds} is set to the nominal value of the stator flux linkage, and ψ^*_{qr} is set to zero. When being adjusted the reference frame, the estimated *d*-axis stator flux is affected. Therefore, both the *d*- and *q*-axis flux controllers are operated simultaneously. The tuning process is finished when the *q*-axis rotor flux is converged to zero, and the *d*-axis stator flux and the rotor magnetizing current are converged to constant values. In that way the nominal operating point of the magnetizing curve is obtained, and by changing ψ^*_{ds} the whole magnetizing curve can be estimated.

3 EXPERIMENTAL RESULTS

To prove the proposed procedure several tests have been performed on 40 kVA vector controlled induction motor drive. It was designed and built for an adjustable speed drive in hoisting applications. The control algorithm is implemented on Texas Instruments TMS320F28335 floating point digital signal controller. The sampling time of the *d*-

Table 1. Rated data and initial motor parameters

Name plate data	Initial parameters
Power, 22 kW	Stator resistance, 0.11 Ω
Speed, 1460 rpm	Rotor resistance, 0.091 Ω
Frequency, 50 Hz	Transient inductance, 4 mH
Voltage, 380 V	Stator inductance, 41.6 mH
Current, 43 A	Rotor time constant, 0.46 s
Power factor, 0.85	Inertia constant, 0.26 s

and q-axis stator current control is 0.5 ms, and it is 10 ms for the speed and flux control. The space vector modulation technique with switching frequency of 2 kHz is used. The name plate data and initial parameters of the motor are presented in Table 1.

3.1 System Dynamics

Figure 3 shows the responses of the machine variables during the tuning process at the nominal flux. Initially, the machine has been excited at standstill by setting the reference i_{mr}^* which is 15% bigger than the nominal magnetizing current ($i_{mr}^* = 0.46$ p.u., $\sqrt{2}I_{mN} = 0.4$ p.u.) and accelerated to 65% of the rated speed with 25% load. The flux estimator has been initiated at 12.5 s, after the speed had reached the reference value. The estimated q-axis rotor flux differs from zero what indicates a detuned condition. At 19.5 s, the standard flux controller with the reference i_{mr}^* is turned off and both the d- and q-axis flux controllers are activated with the references $\psi_{ds}^* = 1$ p.u. and $\psi_{qr}^* = 0$ p.u. The system responds in approximately 2 seconds.

Figure 4 shows the system response to the step of the d-axis flux reference. In Figure 4a) ψ_{ds}^* is set to 1.0 p.u., and when the system responses are settled it is changed to 0.8 p.u. and back to 1.0 p.u. Figure 4b) shows a similar system response which begins from $\psi_{ds}^* = 0.6$ p.u. One can see that the step change of ψ_{ds}^* has a bigger influence over the detuning at unsaturated conditions. The corresponding values of the stator inductance are calculated as ψ_{ds}^* to i_{mr} ratio at steady-state.

Figure 5 shows the test results of the system response to the load torque changes. In this figure the load is reflected by i_{qs} trace. The machine is initially in a detuned condition and i_{qs} is changed from approximately 0.5 p.u. to 0.1 p.u. and back to 0.5 p.u. Both $\hat{\psi}_{ds}$ and $\hat{\psi}_{qr}$ traces change with load, and i_{mr} trace does not change with load. Again, the flux $\hat{\psi}_{qr}$ differs from zero what indicates a detuned condition. At 40.6 s, both the d- and q-axis flux controllers are initiated. The system responds in approximately 2 s. As a result of tuning, the estimated fluxes reach the corresponding reference values, i_{mr} decreases and i_{qs} increases. At 46.5 s, the load is changed in the same manner as before. In steady-state, both $\hat{\psi}_{ds}$ and $\hat{\psi}_{qr}$ do not change with load.

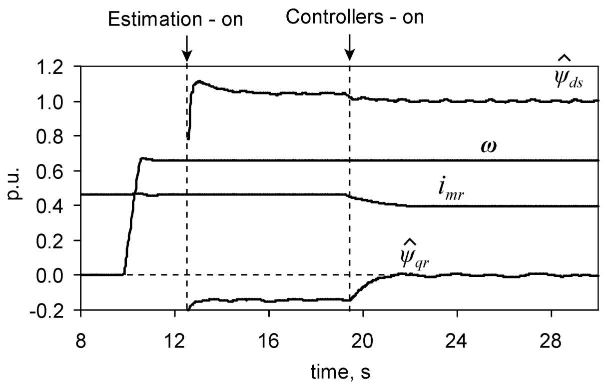

Fig. 3. System dynamic performance during initialization of tuning process

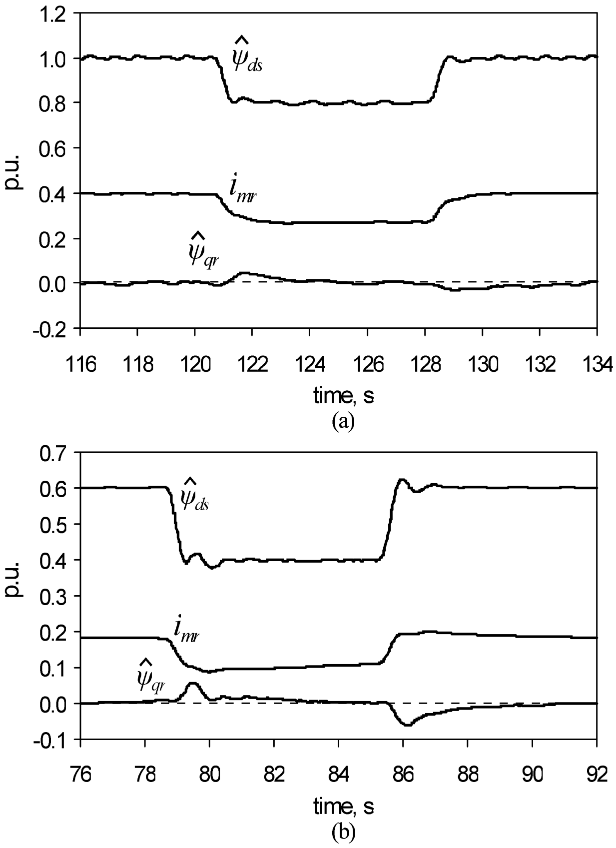

Fig. 4. System response to step change of d-axis stator flux reference for 0.2 p.u.*: a) initial flux reference* 1.0 p.u.*, b) initial flux reference* 0.6 p.u.

3.2 Magnetization Curve

After setting the current and speed controllers, the identification procedure for the saturation characteristic $\Psi_{ds}(I_{ds})$ has been performed as follows: (i) excite the machine by setting $i_{mr}^* = \sqrt{2}I_{mi}$ according to (23), (ii) ac-

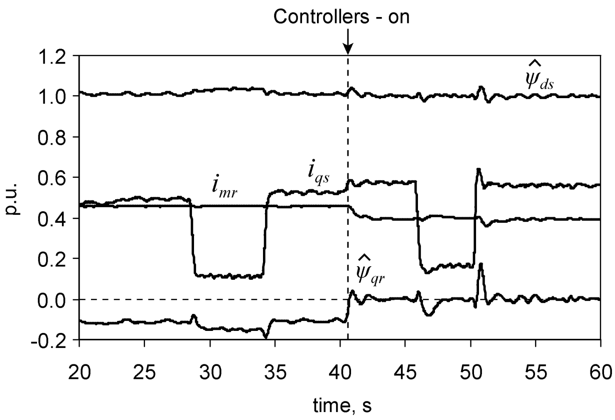

Fig. 5. System response to load torque changes under detuned and tuned conditions

celerate the machine to 65% of the rated speed, (*iii*) initiate the estimation algorithm, (*iv*) activate the flux controllers with the references $\psi^*_{ds} = 1$ p.u. and $\psi^*_{qr} = 0$, (*v*) change the value of the reference ψ^*_{ds} using 10% step from 40% to 120% and measure the corresponding steady-state values of the *d*-axis stator current. After that the magnetization curve can be obtained by using (14). The same test with different levels of the load torque was performed again. For all the load levels, the identified value of the transient inductance was used in the estimation algorithm ($L'_s = 4$ mH). The obtained results are shown in Figures 6 and 7. Figure 6 shows the estimation results obtained when the load torque equals 25% of the rated value in comparison with the characteristic $\Psi_{ds}(I_{ds})$ identified by the no-load test. The estimated results match very well with the reference curve in the whole considered range of the stator *d*-axis flux linkage ($0.4 \leq \psi_{ds} \leq 1.2$ p.u.). In the range $0 \leq \psi_{ds} < 0.4$ p.u. the reference curve in Figure 6 is approximated by a straight line since the no-load test was not performed in this range.

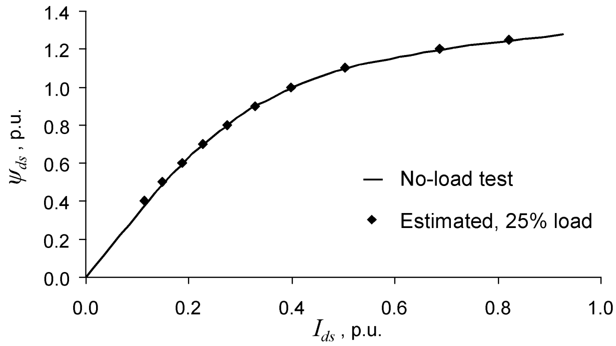

Fig. 6. Experimental results of $\mathbf{\Psi}_{ds}(I_{ds})$ curve estimation at 25% load condition

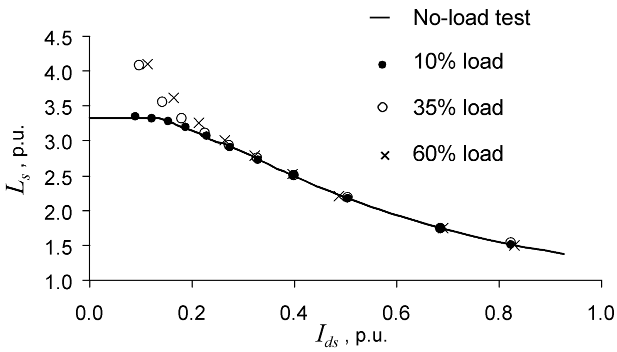

Fig. 7. Experimental results of stator inductance estimation at different load conditions

In Figure 7 the results are shown for three different load torque levels in comparison with the saturation curve $L_s(I_{ds})$ obtained by the no-load test. At a 10% load condition an excellent matching results are obtained. For higher levels of the flux, a very good matching is obtained at a 35% and a 60% load conditions. Significant discrepancies begin at a 70% of the rated flux. These discrepancies are expected since the induction machines are inherently more sensitive to detuning at larger values of i_{qs}/i_{ds} ratio [3].

The *q*-axis rotor flux estimator is sensitive to the transient inductance. Additional tests were performed to assess the error in the stator inductance due to incorrect values of the transient inductance. The results for ±20% error in the transient inductance (in comparison with the identified value $L'_s = 4$ mH) are shown in Figure 8. At the error of -20%, significant discrepancies between the estimates and the reference curve obtained by the no-load test begin at the flux lower than 50% of rated. At the error of +20%, significant discrepancies begin at a higher value of the flux, but it does not exceed 10% at the 50% of rated flux. This suggests that the identified value of the transient inductance could be a little bit bigger than the actual one.

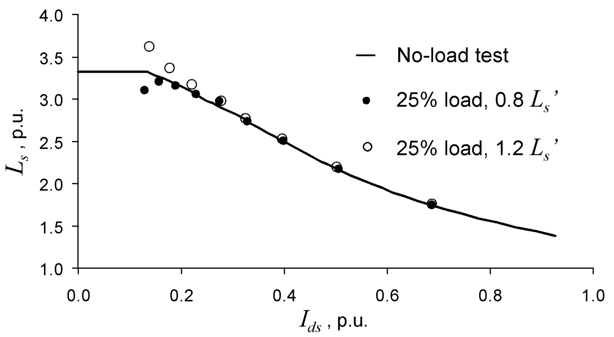

Fig. 8. Experimental results of stator inductance estimation for ±20% error in transient inductance at 25% load condition

4 CONCLUSION

The paper describes a method of the magnetization curve identification of the induction motor in IRFO drive at low-load conditions, suitable for drives where the motor torque is transferred to the load through complex mechanical connections. The estimation algorithm is implemented in a commercial drive as a part of the self-commissioning module. By comparison of the experimental results of the magnetization curve estimation with the no-load test data, the method is verified for the loads lower than twenty-five percentage of the rated torque. A very good accuracy of the magnetization curve estimation has been also obtained at bigger load torques if the flux is not lower than seventy percent of the nominal one. The transient inductance is the only motor parameter on which the presented method is sensitive. Experimental investigations have shown that the estimates for the stator inductance exhibit a very low sensitivity to this parameter if the flux is bigger than fifty percent of the nominal one.

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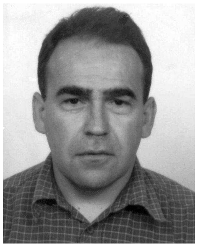

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