On-Line Tuning of the Rotor Time Constant for Vector-Controlled Induction Motor in Position Control Applications

Slobodan N. Vukosavić and Milić R. Stojić

Abstract—**The induction motor drive with an indirect field-oriented controller (IFO) exhibits excellent behavior in a low-speed region. Thus, the motor can be advantageously employed as an actuator in positioning servomechanisms [1]. The substantial sensitivity of IFO controller with respect to changes in the rotor time constant T_r^* calls for an adaptation mechanism, which should allow the tracking of the actual value of rotor time constant through the whole range of operation conditions. This paper proposes the adaptation scheme for the continuous on-line tuning of the parameter T_r^*, suitable for the environment of a position servo. The outlined analytical considerations and experimental results are focused on operation conditions characterized by the zero speed and a light load of the drive. The proposed adaptation scheme is based on the measurement of the terminal voltages that are used as an auxiliary information; then, the scheme is designed in such a way that stator resistance fluctuations and nonlinearities within the analog processing circuitry do not affect the estimated value of the rotor time constant. Experimental results show that the proposed scheme gives good results even in conditions at zero speed and dynamic loads that may be as low as 0.2 p.u.**

I. Introduction

The field-oriented control of the induction motor enables the flux and torque control loops to be decoupled; in addition, the linear and almost instantaneous response of the electrical torque reveal control characteristics of a dc motor. Thus, the induction motor may become suitable also for certain high-performance speed- and position-controlled electrical drives. Positioning servomechanisms require actuators capable of holding the full torque at zero speed. Therefore, in such applications, an IFO controller is preferred rather than a direct field orientation (DFO) controller [2]. Inherently more robust at low speeds, the IFO controller involves the simple control hardware and less complex software than DFO controller. Moreover, the implementation of IFO control calls for the velocity feedback, whereas the slip command is calculated through the slip gain of the feedforward branch. The high-performance servo regularly involves the shaft sensor; thus, the velocity feedback signal is available to the IFO controller. On the other hand, since the feedforward slip calculation implies open-loop decoupling of the induction motor, the IFO controller becomes sensitive to plant parameters. Hence, for the correct field orientation, the parameter T_r^*, used in the feedforward model, must fit the actual rotor time constant T_r of the motor. An inaccurate setting of the parameter T_r^* ($\neq T_r$) results in an undesirable cross-coupling and deterioration of the overall drive performance [2], [3].

Variations of the rotor time constant are caused mainly by the thermal drift of the rotor resistance and by the change of the rotor inductance due to saturation. Although effects of the magnetic saturation can be included into the feedforward flux model [4], [5], the thermal drift cannot be predicted. Hence, for the reliable operation of a drive, an adaptive feature is to be provided within the IFO controller in order to allow the continuous tracking of rotor resistance changes. The adaptation mechanism for the on-line tuning of T_r^* should converge to the actual value of T_r in all operating conditions; the mechanism should not require special test conditions and/or the injection of test signals. Naturally, it is of interest if the implementation of the adaptation scheme does not implicate much penalty in the software and hardware, that is, if the adaptation can be accomplished relying on variables that are already available to the digital controller performing the IFO and (possibly) current control. The usage of dedicated sensors, additional wire connections within the motor-controller assembly, and special configuration of the motor windings are unacceptable because the desirable algorithm should be able to perform the adaptation with standard, off-the-shelf motors without modifications.

Several authors discussed possibilities of adaptation of the parameter T_r^*, and different ideas and solutions were suggested and tested [6]–[14], but solutions were not suited for servo applications. Most of the proposed structures are based on the use of stator (terminal) voltages as an additional information required for the adaptation process. Such an approach is suitable since the stator voltage-related signals are already available within the

Manuscript received March 20, 1992; revised September 9, 1992.
S. N. Vukosavić is with the Electrical Engineering Institute Nikola Tesla, 11000 Belgrade, Yugoslavia.
M. R. Stojić is with the Electrical Engineering Faculty, University of Belgrade, 11000 Belgrade, Yugoslavia.
IEEE Log Number 9204613.

drive controller. Generally, either the stator voltage $\boldsymbol{u}_s$ or the function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$, derived from the stator currents and voltages, may be taken as an input to the adaptation mechanism. The chosen criterion function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ has to be related to the rotor time constant, whereas the reference function F^* is derived from the feedforward model. Hence, the error $\Delta F = F^* - F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ may be used as an indicator of the on-line tuning efficiency. Garces [6] has chosen the function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ related to the motor reactive power. Rowan, Kerkman, and Leggate [7] review possible adaptation schemes and propose a new approach wherein the stator voltage components (u_d and u_q) are adopted as the reference function F^*. Lorenz and Lawson [8] assume the estimated torque as the criterion function and obtain superior results for the stator frequencies $f > 5$ Hz. Chan and Wang [9] propose the method where the parameter T_r is directly calculated from the synchronous watt torque in the steady state. As Schumaher, Heinemann, and Leonhard [10], [11] showed, the adaptation—excluding very low stator frequencies—can be efficiently performed by measuring the phase displacement between the rotor-induced voltage estimated from the terminal quantities and from its counterpart calculated from the reference model.

In cases when some apriori knowledge about the drive is available, the problem of adaptation of the parameter T_r^* can be advantageously resolved. Thus, in the case where the load parameters are known, Ohnishi, Ueda, and Miyachi [12] proposed the efficient adaptation scheme capable of operating at the zero-speed as well. The solution suggested by Zai and Lipo [13] utilizes the commutation noise of the pulsewidth modulation (PWM) inverter as a test signal; in this way, the use of an external test signal is avoided. To this end, the use of the extended Kalman filter was proposed, and the resulting algorithm identifies the parameter T_r^* in cases where the PWM noise has a wide spectrum and when the rotor resistance R_r at the slip frequency corresponds to the value of R_r, within the frequency range of the PWM noise. For motors with star-connected stator windings and with the available potential of the motor neutral, Moreira *et al.* [14] proposed the adaptation structure having the excellent deadbeat dynamic. In the structure, the third harmonic component of the stator-induced voltages is detected from the magnetic saturation phenomena and non-sinusoidal distribution of the stator windings.

If the adaptation at low-input frequencies is not required, the aforementioned adaptation mechanisms can be advantageously applied. Adaptive controllers based on the stator voltage d component U_d^s and the torque referent model [7] offer the superior performance and do not require steady-state conditions in order to tune the drive. However, to tune the IFO controller in the environment of the position servo, the adaptation of the parameter T_r^* must be efficient in both the zero-speed and low-torque conditions. Moreover, such an adaptation mechanism should be active in the steady state and during the transient response; that is, the pulsation character of the torque command $K_T i_q$ should not interfere with the adaptation process. Also, the implementation of the adaptation mechanism should not require dedicated sensors and/or a modification of the motor windings and wiring. Note that up to now these requirements have not been met. Therefore, the problem of adaptation of the parameter T_r^* in the design of positioning servomechanisms with induction motors is discussed in this paper, and the adaptive mechanism suited for the position control requirements is proposed and experimentally tested (Figs. 1–8).

The adaptation mechanism under consideration is driven by the error $\Delta F = F^*(\boldsymbol{\Psi}_r^*, \boldsymbol{i}_s) - F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ detected by subtracting the criterion function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ derived from the stator voltages and currents, from the match function $F^*(\boldsymbol{\Psi}_r^*, \boldsymbol{i}_s)$, related to the model of rotor circuit implemented within the IFO controller. The choice and derivation of the criterion function is discussed in Section II, taking into account the requirement that the function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ must be sensitive to a detuned condition, and that the related hardware must not involve the additional wiring, transducers, and motor modification. Instead of sensing the stator voltages directly, the binary control signals for inverter switches are adopted. These signals are then integrated and correlated with the stator currents. Thus, undesirable effects, such as the lockout time of power switches, do not affect the mean value of the function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$. Section III presents the adaptation mechanism of the parameter T_r^*. The mechanism enables the continuous correction of T_r^* until the error $\Delta F = F^*(\boldsymbol{\Psi}_r^*, \boldsymbol{i}_s) - F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ is reduced to zero. Experimental results are presented in Section IV along with descriptions of the drive prototype and an experimental setup. Concluding remarks are given in Section V.

II. The Choice and Derivation of the Criterion Function

The rotor time constant adaptation within the IFO controller, which decouples the control of induction motor in the position control environments, requires the reliable operation at extremely low stator frequencies. At zero speed, the stator frequency equals the slip frequency, which may be only a fraction of 1 Hz, if the load torque is close to zero. In the absence of flux sensors, the stator voltages remain as a sole additional information that can be used in the process of adaptation. Thus, the major problem appears from the fact that the terminal voltage at zero speed consists predominantly of the voltage drop across the stator resistance R_s. Since the value of R_s varies with temperature and is not generally known, neither the stator voltage itself nor the estimate of the motor torque (derived from the variables $\boldsymbol{u}_s$ and $\boldsymbol{i}_s$) can be used as a criterion function.

The suitable method of detecting the stator voltages can be evaluated using the binary control signals fed to the inverter power switches, especially when the control of stator currents is performed by the digital controller and when the voltage-related signals are already available within the controller memory. In this approach, the problem of the lock-out time of power switches is faced. That is, the lockout time causes a difference between the binary control signals (PWM pattern) and the actualized voltage across the motor. Hence, the proper choice of the criterion function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ must be made in such a way that the knowledge of the stator resistance is not necessary and that the undesirable signal caused by the lockout time does not affect the value of the criterion function. We propose the following criterion function

$$F(u_a, u_b, u_c, i_a, i_b, i_c) = i_a \int u_a\, dt + i_b \int u_b\, dt + i_c \int u_c\, dt \tag{1}$$

or, after transforming the three-phase quantities into the $\alpha-\beta$ stationary coordinate system,

$$F(u_\alpha, u_\beta, i_\alpha, i_\beta) = i_\alpha \int u_\alpha\, dt + i_\beta \int u_\beta\, dt. \tag{2}$$

The influence of stator resistance and lockout time on the detected signal $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ is analyzed and then it is proven that (1) is related to the square of the rotor flux.

Suppose that the stator voltages u_a, u_b, and u_c are reconstructed from the PWM pattern having the lockout time of power switches, PWM period, and dc-bus voltage denoted by τ, T, and E, respectively. Then the detected signals u_a', u_b', and u_c' may be approximated as

$$u_a' = u_a - E\frac{\tau}{T}\operatorname{sgn}(i_a) = R_s i_a + d\Psi_a/dt - E\frac{\tau}{T}\operatorname{sgn}(i_a) \tag{3a}$$

$$u_b' = u_b - E\frac{\tau}{T}\operatorname{sgn}(i_b) = R_s i_b + d\Psi_b/dt - E\frac{\tau}{T}\operatorname{sgn}(i_b) \tag{3b}$$

$$u_c' = u_c - E\frac{\tau}{T}\operatorname{sgn}(i_c) = R_s i_c + d\Psi_c/dt - E\frac{\tau}{T}\operatorname{sgn}(i_c). \tag{3c}$$

It is of interest to investigate the influence of the stator resistance and the lock-out time on the function (1). Denoting the amplitude of stator current and the angular frequency by I and ω, respectively, one obtains

$$F(u_a', u_b', u_c', i_a, i_b, i_c) = \Psi_a i_a + \Psi_b i_b + \Psi_c i_c + \sigma(R_s) + \gamma(\tau) \tag{4}$$

where

$$\sigma(R_s) = R_s I^2 \Big[\cos(\omega t) \int \cos(\omega t)\, dt + \cos(\omega t - 120°) \int \cos(\omega t - 120°)\, dt + \cos(\omega t - 240°) \int \cos(\omega t - 240°)\, dt\Big] = 0 \tag{5}$$

and

$$\gamma(\tau) = E\frac{\tau}{T} I \Big\{\cos(\omega t) \int \operatorname{sgn}[\cos(\omega t)]\, dt + \cos(\omega t - 120°) \int \operatorname{sgn}[\cos(\omega t - 120°)]\, dt + \cos(\omega t - 240°) \int \operatorname{sgn}[\cos(\omega t - 240°)]\, dt\Big\}. \tag{6}$$

After substituting $\operatorname{sgn}[\cos(\omega t)]$ by its Fourier series and performing the integration indicated in (6), $\gamma(\tau)$ becomes

$$\gamma(\tau) = -EI\frac{4\tau}{\pi T} \Big\{\sum_{k=0}^{+\infty} \frac{(-1)^k}{\omega(2k+1)^2} \cdot [\sin[(2k+1)(\omega t)]\cos(\omega t) + \sin[(2k+1)(\omega t - 120°)]\cos(\omega t - 120°) + \sin[(2k+1)(\omega t - 240°)]\cos(\omega t - 240°)]\Big\}. \tag{7}$$

The fundamental component in (7), which corresponds to $2k + 1 = 1$, and all other components in which $2k + 1$ equals some multiple of 3 are identically equal to zero. It can be proven that remaining components in (7) have the zero mean values within the period $[t \in (0.2\pi/\omega)]$ of the fundamental component. Hence, as long as the average value of the function $F(u_a, u_b, u_c, i_a, i_b, i_c)$ is considered, one can conclude that the stator resistance and the inverter lockout time τ do not affect the function value in the steady state.

According to (1), the derivation of $F(u_a, u_b, u_c, i_a, i_b, i_c)$ requires the integration of stator voltages. If the current control loops are realized within the digital controller, this integration can be performed by the software; otherwise the binary control signals are to be processed by an analog integrator. The advantageous solution for the integration of stator voltages has been proposed by Schumaher and Leonhard [10]: the dc bus is fed to the VCO, and the resulting pulses are driving three UP/DOWN counters in accordance with the PWM pattern for each phase. Then, contents of the counters are related to the integrals of the phase voltages.

Whatever the method of integration is, the result may contain a parasitic dc-component (V_{a0}, V_{b0}, and V_{c0}) in each phase. Consequently, the corresponding contribution

$V_{a0}i_a + V_{b0}i_b + V_{c0}i_c$ to the function $F(u_a, u_b, u_c, i_a, i_b, i_c)$ will appear. The contribution has the zero mean value for a periodic excitation as well, but it may introduce pulsations at the supply frequency. Hence, the proposed derivation of the flux-related function $\Psi_a i_a + \Psi_b i_b + \Psi_c i_c$ from the inverter control signals is not affected by the lockout time phenomenon, dc-offset, and voltage drop across the stator resistance. Still, the aforementioned parasitic effects may introduce zero-mean pulsations into the detected signal, as it will be demonstrated in Section IV.

After performing the rotating transformation from the α–β stationary frame into the synchronous d–q frame, the criterion function may be expressed in terms of d–q variables as

$$\begin{aligned} F(\boldsymbol{u}_s, \boldsymbol{i}_s) &= \Psi_{ds} i_{ds} + \Psi_{qs} i_{qs} \\ &= L_\sigma i_s^2 + \frac{M}{L_r}[\Psi_{dr} i_{ds} + \Psi_{qr} i_{qs}]. \end{aligned} \tag{8}$$

where M is the magnetizing inductance, L_r is the rotor inductance, and L_σ is the equivalent leakage inductance of the induction motor.

In (8), $L_\sigma i_s^2$ corresponds to the magnetic energy accumulated in the motor leakage inductances, whereas the term $\Psi_{dr} i_{ds} + \Psi_{qr} i_{qs}$ is related to the rotor flux; hereafter it will be shown that this term is sensitive to a detuned condition of the IFO controller.

Let us consider the induction motor with the rotor time constant T_r. Suppose that the motor is driven by the IFO controller having the parameter T_r^* ($\neq T_r$) and generating the current commands (i_d, i_q) and slip command $\omega_s = (i_q/i_d)/T_r^*$. The components (Ψ_{dr}, Ψ_{qr}) of the rotor flux and the rotor flux amplitude Ψ_r can be derived from the rotor equations

$$\Psi_{dr} = \frac{Mi_d + \omega_s T_r M i_q}{1 + \omega_s^2 T_r^2} \qquad \Psi_{qr} = \frac{Mi_q - \omega_s T_r M i_d}{1 + \omega_s^2 T_r^2} \tag{9}$$

$$|\Psi_r| = \sqrt{\Psi_{dr}^2 + \Psi_{qr}^2} = \frac{Mi_s}{\sqrt{1 + \omega_s^2 T_r^2}} \tag{10}$$

with

$$i_s^2 = i_d^2 + i_q^2.$$

From (8) and (9), one obtains

$$F(\boldsymbol{u}_s, \boldsymbol{i}_s) = L_\sigma i_s^2 + \frac{M}{L_r}\frac{Mi_s^2}{1 + \omega_s^2 T_r^2} = L_\sigma i_s^2 + \frac{|\Psi_r|^2}{L_r}. \tag{11}$$

To investigate the sensitivity of $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ with respect to a detuned condition, let us assume that $T_r = T_r^* + \Delta T_r$. Then, the sensitivity function can be conveniently introduced as the partial derivative

$$\frac{\partial F(\boldsymbol{u}_s, \boldsymbol{i}_s)}{\partial(\Delta T_r)} = -\frac{2T_r M^2 i_s^2}{L_r(1 + \omega_s^2 T_r^2)^2}\cdot \omega_s^2 \tag{12}$$

calculated readily from (10) and (11).

In virtue of (12), positive and negative changes in T_r cause respectively a decrease and an increase of the criterion function, in all operating conditions. Thus the corrective action of the adaptation mechanism does not require the extraction of the criterion function error ΔF coherent with the feedforward command. On the other hand, the sensitivity function (12) is mainly influenced by the slip. Thus, if the slip is zero (i.e., the output torque is zero), the motor is fed by the magnetizing current i_d alone, and then the deviation ΔT_r has no effect on the function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$. Hence, the drive tuning at the no-load condition is not feasible. Note, this problem is common to all adaptation schemes that do not utilize additional test signals.

III. The Mechanism for On-Line Tuning of the Parameter T_r^*

As it was shown in the previous section, the derivation of $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ from the motor terminal variables enables the extraction of the term $\Psi_{dr} i_{ds} + \Psi_{qr} i_{qs}$, which reflects changes in the actual rotor time constant T_r. After calculating the reference function $\Psi_{dr}^* i_{ds} + \Psi_{qr}^* i_{qs}$ related to the rotor circuit model, the on-line tuning of the model parameter T_r^* can be performed through the PI- or I-actions, which drive the difference ΔF between the criterion function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ and its reference F^* to zero.

In the case where $T_r^* = T_r$, the estimate $\Psi_{dr} i_{ds} + \Psi_{qr} i_{qs}$ corresponds to the reference function $\Psi_{dr}^* i_{ds} + \Psi_{qr}^* i_{qs}$ in both the steady state and transient conditions. Consequently, the suggested adaptation mechanism will not require steady-state conditions in order to tune the drive. Still, undesirable effects, discussed in Section II, will inject a zero-mean pulsation into the function $F(\boldsymbol{u}_s, \boldsymbol{i}_s)$ at the supply frequency. Since the position control environment includes the frequency range below 1 Hz, the corrective action must reveal a low-pass nature to prevent possible pulsations in the value of estimated parameter T_r^*. Therefore, it is desirable to process the error signal ΔF through the integrator with the integration time constant T_i; the output of integrator represents the estimate T_r^*. A large value of T_i and resulting slow adaptation dynamic will attenuate the noise, decouple the adaptation loop from the drive dynamic, and make proportional corrective action unnecessary.

The slow adaptation process implies that changes in the rotor time constant, produced by variations of the flux level and saturation phenomena, cannot be accurately tracked since the time of response of the servo with induction motor from the basic speed region up to the

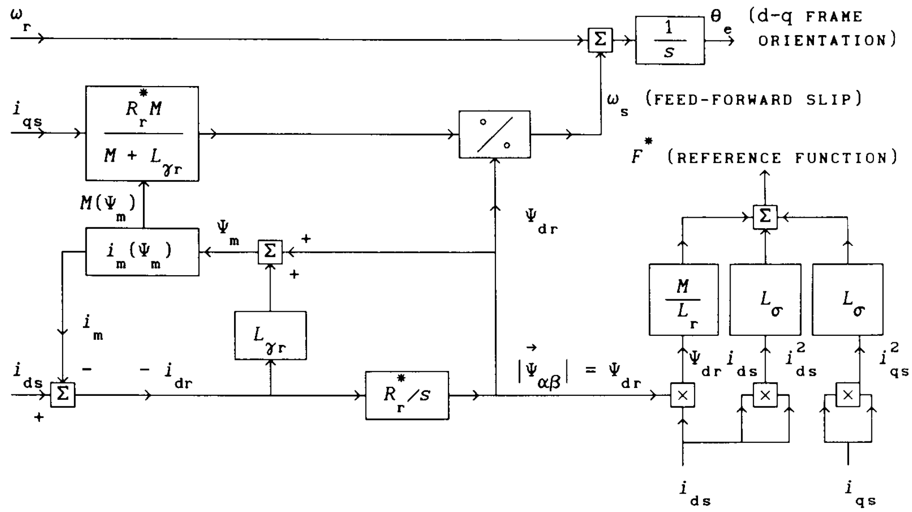


Fig. 1. Modification of the model of rotor circuit and the derivation of the reference function $F^*(\Psi_r^*, i_s)$; L_σ-equivalent leakage inductance, L_{γ_r} = rotor leakage inductance, L_r = rotor inductance, and ω_r = shaft speed.

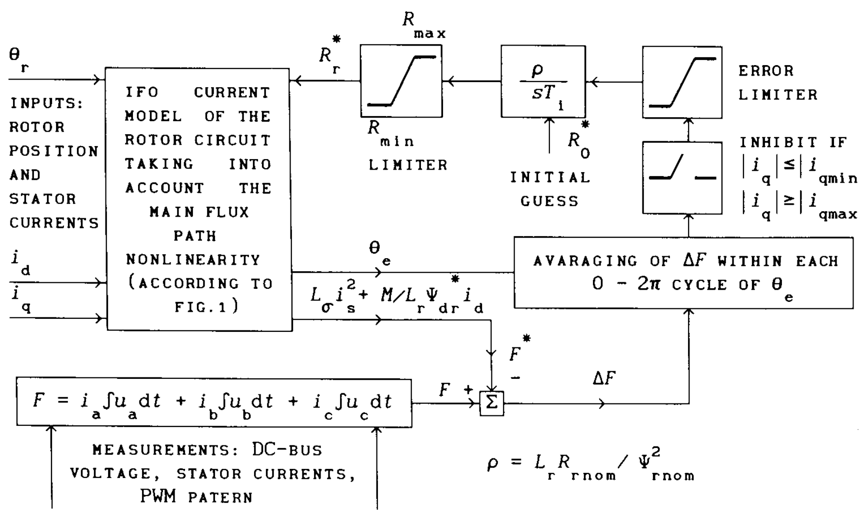


Fig. 2. Proposed adaptation mechanism for the correction of thermal drift in R_r.

field weakening and back may be below 1 s. Therefore, in the following discussion and during the experimental verification, the adaptation mechanism is used to track the thermal drift of rotor resistance R_r, whereas the influence of the magnetic saturation is compensated in the modified model [4], [5] of the rotor circuit (Fig. 1). When the air-gap magnetizing characteristic $i_m(\Psi_m)$ is available, the current model of the rotor circuit, contained in the IFO controller, is to be rearranged, as shown in Fig. 1.

In an overload operation regime of the drive, where the stator current, output torque, and slip exceed their rated values several times, the phenomenon of cross-saturation [4], [5], saturation of leakage inductances, and frequency dependence of the rotor parameters must be taken into account as well. Since such a kind of working regime may not last very long, the adaptation process may be inhibited while the drive is at a heavy overload mode. Under the normal rated conditions, the phenomena mentioned above have a negligible effect, and the model illustrated in Fig. 1 can be advantageously applied.

Proposed adaptation mechanism is shown in Fig. 2. In the figure, the parameter ρ is introduced for scaling purposes. The proposed adaptation process is inhibited under no-load conditions since the sensitivity function (12), in this mode of operation, drops to zero. The integration time constant T_i determines the velocity of adapta-

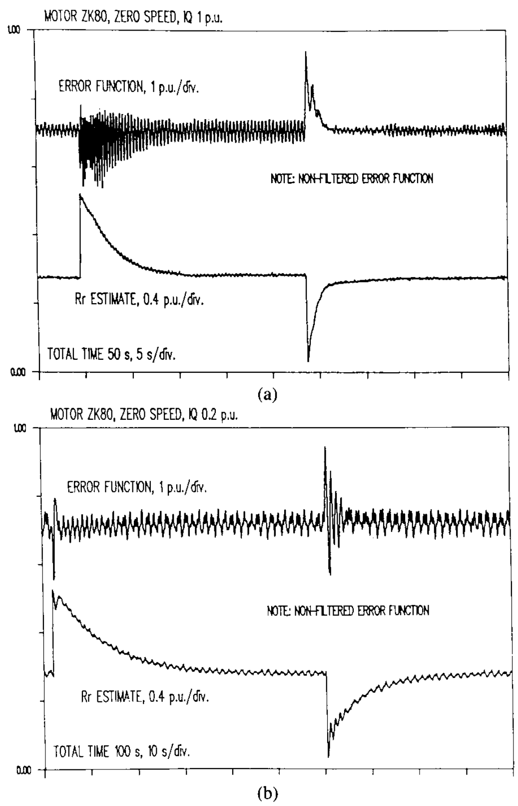


Fig. 3. Motor ZK80 with P_n = 750 W. Identification traces of R_r^* in the steady state with $\omega_r = 0$ and disabled filtering of ΔF. (a) T_e = const. (= 1 p.u.). (b) T_e = const. (= 0.2 p.u.). Upper traces: (a) and (b) ΔF with 1 p.u./div. Lower traces: (a) and (b) R_r^* with 0.4 p.u./div. Total times: (a) 50 s, (b) 100 s.

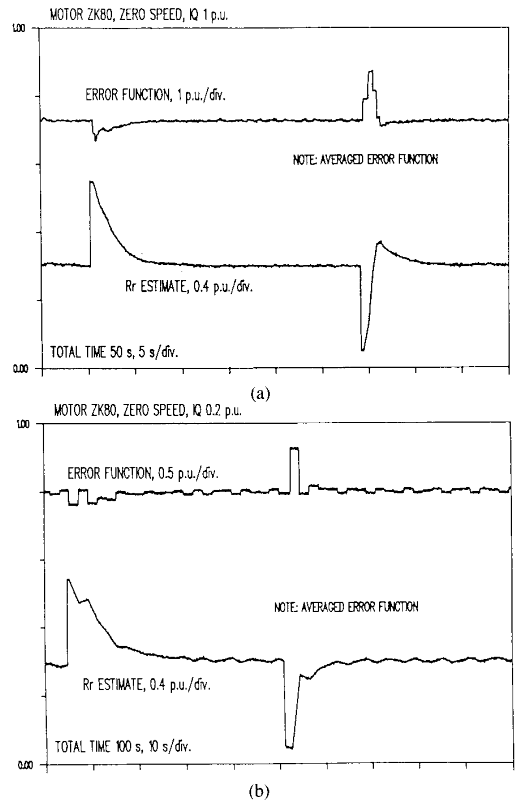


Fig. 4. Motor ZK80 with P_n = 750 W. Identification traces of R_r^* in the steady state with $\omega_r = 0$ and enabled filtering of ΔF. (a) T_e = const. (= 1 p.u.). (b) T_e = const. (= 0.2 p.u.). Upper traces: (a) ΔF with 1 p.u./div. (b) ΔF with 0.5 p.u./div. Lower traces: (a) and (b) R_r^* with 0.4 p.u./div. Total times: (a) 50 s, (b) 100 s.

tion loop response, that is, the speed of convergence. The error signal is filtered in the following manner: whenever the rotor flux (d-axis angle θ_e) comes into the alignment with the phase a winding ($\theta_e = 0$), a dedicated register starts accumulating samples of ΔF. At the next alignment, the average value of ΔF is calculated by dividing the sum of samples of ΔF by the expired time, and the new cycle is initialized. In this way, the low-frequency pulsations, discussed in Section II, are greatly attenuated. The benefit of such filtering is proven by the experimental results that follow.

IV. Experimental Results

The efficiency of the proposed adaptation scheme is verified on the prototype of the drive with induction motor, employing two standard induction motors having different rated powers. The motor data are given in Table I.

The IFO controller and the adaptation mechanism were implemented on the microcontroller system based in the INTEL 8088/8087, along with the interactive/monitoring software. Since all measurements were performed at the zero speed, neither speed nor position controller were implemented, but the motor shaft was mechanically blocked, and the encoder feedback was overridden by $\theta_r \equiv 0$. The torque command $K_T i_q$ to the drive was assigned by the user, or generated in the software as a pulse sequence. The motor was additionally cooled, and all tests were undertaken only after a steady state temperature has been achieved.

The $(d\text{-}q) \rightarrow (abc)$ transformation is performed each 1 ms, and references for the motor phase currents are fed from the 8-b D/A converters to the current controller implemented by the analog circuitry. The hardware was essentially the same for both motors, except for the shunt resistors that were scaling the range of the Hall-effect current sensors. The average switching frequency and the lock-out time of the IGBT-transistor bridge were 8 kHz and 4 μs, respectively.

The stator voltages were reconstructed from the binary signals controlling the IGBT inverter and the dc-bus voltage. Such obtained signals were fed to operational amplifiers configured as integrators with the clamping back-to-back zener diodes. The result of integration was fed to the 10-b A/D converter, and further processed by the digital microcontroller.

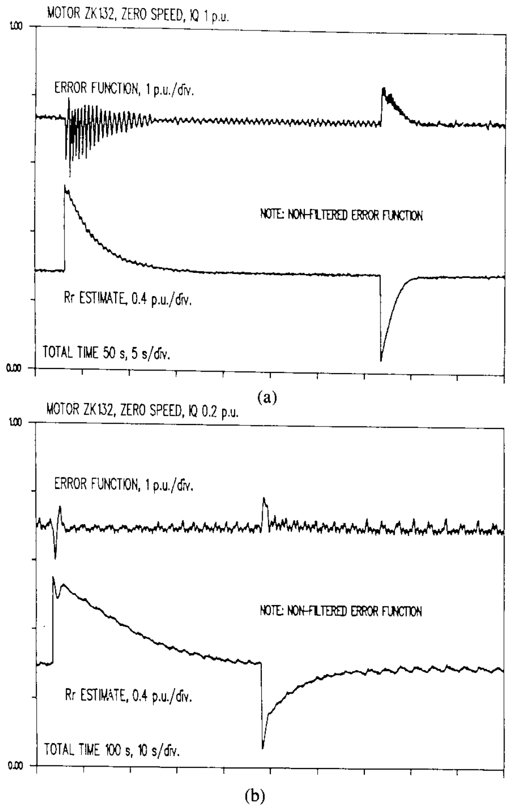


Fig. 5. Motor ZK132 with P_n = 7.5 kW. Identification traces of R_r^* in the steady state with ω_r = 0 and disabled filtering of ΔF. (a) T_e = const. (= 1 p.u.). (b) T_e = const. (= 0.2 p.u.). Upper traces: (a) and (b) ΔF with 1 p.u./div. Lower traces: (a) and (b) R_r^* with 0.4 p.u./div. Total times: (a) 50 s, (b) 100 s.

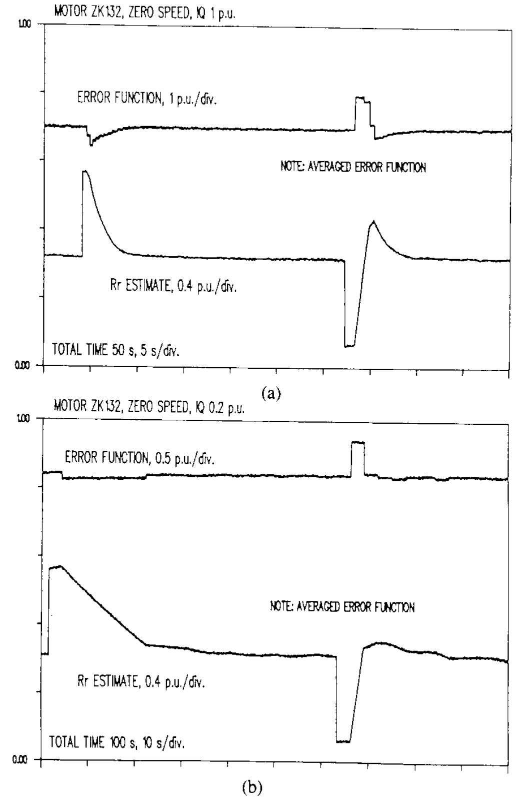


Fig. 6. Motor ZK132 with P_n = 7.5 kW. Identification traces of R_r^* in the steady state with ω_r = 0 and enabled filtering of ΔF. (a) T_e = const. (= 1 p.u.). (b) T_e = const. (= 0.2 p.u.). Upper traces: (a) ΔF with 1 p.u./div. (b) ΔF with 0.5 p.u./div. Lower traces: (a) and (b) R_r^* with 0.4 p.u./div. Total times: (a) 50 s, (b) 100 s.

TABLE I
PARAMETERS OF THE MOTORS UNDER THE TEST

Motor type	ZK80	ZK132
Rated power	0.75 kW	7.5 kW
Number of poles	4	4
Rated voltage	3 × 380 V	3 × 380 V
Rated current	2.1 A	16 A
Rated frequency	50 Hz	50 Hz
Connection	Y	D
Stator resistance (p.u.)	0.095	0.038
Rotor resistance (p.u.)	0.06	0.043
Stator inductance (p.u.)	1.38[a]	1.993[a]
Rotor inductance (p.u.)	1.38[a]	1.993[a]
Magnetizing inductance (p.u.)	1.26[a]	1.91[a]
Equivalent leakage inductance (p.u.)	0.23	0.164

[a] Under rated conditions.

The traces obtained in an experimental setup (Figs. 3–8) present the waveforms of the error signal $\Delta F = F(\boldsymbol{u}_s, \boldsymbol{i}_s) - F^*$ and the time changes of the parameter R_r^*, obtained through the proposed adaptation mechanism. As an excitation of the adaptation loop, the parameter R_r^* step changes of $\pm 50\%$ were introduced by the software.

To visualize the importance of filtering of the error signal ΔF, the same set of experiments was repeated with and without the averaging filter indicated in Fig. 2. The process of adaptation corresponding to the regime of constant torque or to steady-state conditions is illustrated by Figs. 3–6. The experimental traces of the rotor resistance adaptation for the motor ZK80 are shown in Fig. 3 corresponding to the load torques of 1.0 p.u. and 0.2 p.u.; in the experiment, the filter of error ΔF (Fig. 2) was not included in the adaptation mechanism. Fig. 4 is related also to the motor ZK80 with the same test conditions, except for the averaging filter of the error ΔF, which was involved into the adaptation loop. Figs. 5 and 6 show the experimental results obtained from the same set of tests repeated for the motor ZK132.

In order to verify the ability of the proposed adaptation mechanism to operate in transient conditions of the drive as well, additional tests were performed. In the tests, the torque, or the torque command $K_T i_q$, was pulsating within 0 and 1 p.u. at the frequency of 1 Hz; the duty cycles of 1 p.u. pulses was set to 20%, and hence the average value of the torque was 0.2 p.u. Under these conditions, the efficiency of the proposed adaptation scheme was analyzed

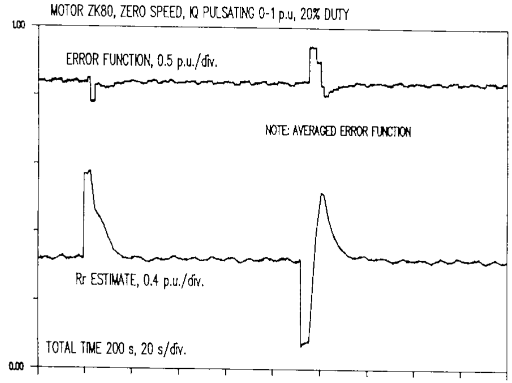


Fig. 7. Motor ZK80 with $P_n = 750$ W. Identification traces of R_r^* in transient conditions: T_e consists of a series of pulses having the frequency of 1 Hz, amplitude 1 p.u., duty cycle 20%, and average value 0.2 p.u. Shaft speed is $\omega_r = 0$, and the filtering of ΔF is enabled. Upper trace: ΔF with 0.5 p.u./div. Lower trace: R_r^* with 0.4 p.u./div. Total time 200 s.

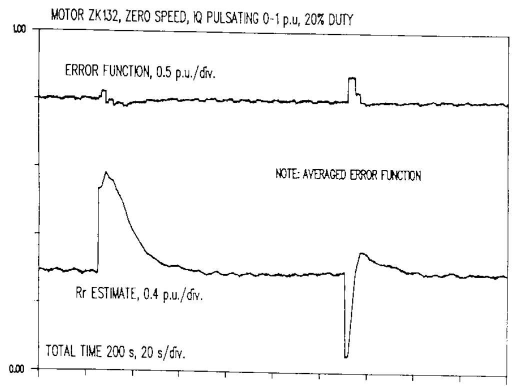


Fig. 8. Motor ZK132 with $P_n = 7.5$ kW. Identification traces of R_r^* in transient conditions: T_e consists of a series of pulses having the frequency of 1 Hz, amplitude 1 p.u., duty cycle 20%, and average value 0.2 p.u. The shaft speed is $\omega_r = 0$, and the filtering of ΔF is enabled. Upper trace: ΔF with 0.5 p.u./div. Lower trace: R_r^* with 0.4 p.u./div. Total time 200 s.

experimentally for both motors, and the test results are shown in Figs. 7 and 8 for the motors ZK80 and ZK132, respectively. In the experimental tests, the averaging filter was enabled.

The presented experimental results are in agreement with the analytical analysis and design guidance presented in Sections II and III. It is of particular interest to note that the proposed adaptation scheme is capable of tracking rotor resistance changes at the zero speed as well, and even under low-torque conditions. Still, as the supply frequency decreases, the noise introduced by parasitic effects becomes more emphasized. Therefore, instead of processing the error signal ΔF, the better way is to take the mean value of this signal within each rotation of the d–q reference frame for 2π electrical degrees. Notice from the experimental results that the behavior of the adaptation loop in transient conditions of the drive is very much the same as in the steady state.

V. Conclusion

The novel adaptation mechanism for on-line tuning of the IFO controller of the induction motor is proposed and tested. The mechanism is suitable for application in the design of positioning servomechanism with induction motor, where the adaptation is required at the zero speed as well. As it was proven experimentally, the proposed adaptation mechanism operates even in the regime of low torques. Moreover, the drive tuning is feasible in both the steady-state and transient operating conditions. Note, an implementation of the proposed adaptation mechanism does not require additional sensors, extra wiring, nor modifications of standard motors. In drives with the digital current control, the developed adaptation technique can be entirely implemented in the software, without hardware modifications of the drive.

References

[1] M. R. Stojić and S. N. Vukosavić, "Design of microprocessor-based system for positioning servomechanism with induction motor," *IEEE Trans. Ind. Electron.*, vol. 38, pp. 369–378, Oct. 1991.

[2] D. W. Novotny and R. D. Lorenz, "Introduction to field orientation and high performance AC drives," in *Publication Industrial Drives Committee IEEE Ind. Applications Soc.*, 1985.

[3] R. Krishnan and F. C. Doran, 'Study of parameter sensitivity in high performance inverter-fed induction motor drive systems," *IEEE Trans. Industry Applications*, vol. IA-23, pp. 623–635, July/Aug. 1987.

[4] E. Levi and V. Vučković, "Field oriented control of induction machine in the presence of magnetic saturation," *Electrical Machines Power Syst.*, vol. 16, no. 2, pp. 133–147, 1989.

[5] E. Levi, S. Vukosavić and V. Vučković, "Study of main flux saturation effects in field oriented induction motor drives," in *Conf. Rec. IEEE Ind. Electron. Soc. Ann. Meet., IECON'89*, pp. 219–224.

[6] L. Garces, "Parameter adaptation for the speed-controlled static ac-drive with squirrel-cage induction motor operated with variable frequency power supply," *IEEE Trans. Industry Applications*, vol. IA-16, pp. 173–178, Mar./Apr. 1980.

[7] T. Rowan, R. Kerkman, and D. Leggate, "A simple on-line adaptation for indirect field orientation of an induction machine," *IEEE Trans. Industry Applications*, vol. 37, pp. 720–727, July/Aug. 1991.

[8] R. D. Lorenz and D. B. Lawson, "A simplified approach to continuous on-line tuning of field oriented induction machine drives," in *Conf. Rec. IEEE Industry Applications Soc. Ann. Meet.*, 1988, pp. 444–449.

[9] C. C. Chan and H. Wang, "An effective method for rotor resistance identification for high-performance induction motor vector control," *IEEE Trans. Ind. Electron.*, vol. 37, pp. 477–482, Dec. 1990.

[10] W. Schumaher and W. Leonhard, "Transistor-fed AC-servo drive with microprocessor control," in *Proc. 1983 IPEC*, Tokyo, pp. 1466–1476.

[11] G. Heinemann and W. Leonhard, "Self-tuning field orientated control of an induction motor drive," in *Proc. 1990 IPEC*, Tokyo, pp. 465–472.

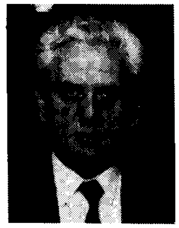
[12] K. Ohnishi, Y. Ueda, and K. Miyachi, "Model reference adaptive system against rotor resistance variations in induction motor drives," *IEEE Trans. Ind. Electron.* vol. IE-33, pp. 217–223, Aug. 1986.

[13] L. C. Zai and T. A. Lipo, "An extended Kalman filter approach to rotor time constant measurement in PWM induction motor drives," in *Conf. Rec. IEEE Industry Applications Soc. Ann. Meet.*, 1987, pp. 177–183.

[14] T. C. Moreira, K. T. Hung, T. A. Lipo, and R. D. Lorenz, "A simple and robust adaptive controller for detuning correction in field oriented induction machines," in *IEEE Industry Applications Soc. Ann. Meet.*, 1991, pp. 397–403.

Slobodan N. Vukosavić was born in Sarajevo, Bosnia and Hercegovina, Yugoslavia, on January 27, 1962. He received the B.C., M.S., and Ph.D. degrees from the Electrical Engineering Faculty, University of Belgrade, in 1985, 1987, and 1989, respectively.

Since 1986, he has been with the Nikola Tesla Institute, Belgrade, Yugoslavia, where he conducts research in the areas of static power converters and microcomputer-based control of electrical drives. He is also engaged part time at the Electrical Engineering Faculty, University of Belgrade, where he teaches graduate and postgraduate courses in power electronics and control of electrical drives. In 1988, he spent six months with the ESCD Laboratory of Emerson Electric, St. Louis, MO, in the cooperative research program. Since October 1991, he has been with the Vickers Co., Milano, Italy, under the one-year research program in the design and control of electrical drives for robots. His scientific interests are in the areas of the system modeling and identification, microcomputer-based real-time control, power electronics, and rotating electrical machines design and control.



Milić R. Stojić was born in Užice, Serbia, Yugoslavia, on February 27, 1940. He received the B.S., M.S., and Ph.D. degrees from the Electrical Engineering Faculty, University of Belgrade, in 1963, 1965, and 1967, respectively.

Since 1980, he has been a full Professor at the Electrical Engineering Faculty, University of Belgrade, teaching graduate and postgraduate courses in automatic control. In 1980, 1984, and 1988, he spent six months at the Physics Department, University of Birmingham, England, under a cooperative research program. His scientific interests are in the sensitivity analysis of dynamic systems, system simulation, and microcomputer-based real-time control of electrical drives and industrial processes. He has published over 100 papers and six books in English, Russian, and Serbocroation. He is the author of *Continuous Control Systems* and *Digital Control Systems*, which are used as standard textbooks on automatic control at universities in Yugoslavia. He also developed two analog and digital student laboratories.