FINITE SET MODEL PREDICTIVE SPEED CONTROL OF INDUCTION MOTOR WITH LONG HORIZON*

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Abstract: This paper presents predictive speed control of induction motor. The structure with finite control set and long prediction horizon is proposed. In methods with finite set of solutions there is assumed the use of signals that can be obtained in inverter only. Long prediction horizons allow better performance of predictive control structures to be achieved. In this case, long prediction horizon refers to systems with more than one prediction step. The weighting factors determine the properties of control structure via an optimization method using genetic algorithms. The results of simulation studies has been presented.

Keywords: *finite control set, predictive control, induction motor, speed control*

1. INTRODUCTION

In recent years requirements for driving systems are steadily growing. System responses should be quick and exact. Also predetermined boundaries of signals should be taken into account. A model predictive control (MPC) has been continuously developed for a few decades. It is considered as one of the most important advances in process control [9]. MPC ensures fast and precise responses of the system due to its complex algorithm. The effect of future control actions on the process state is predicted based on a current state. The explicit process model is usually used for predicting the behavior of the system. Predicted system states are compared to a set of objectives, then available control actions are adjusted to achieve the objectives while respecting constraints (Fig. 1c). One disadvantage of MPC methods is the high computational complexity. However, the modern control platforms have possibility to compute the MPC algorithm on-line even for high frequency sampling system such as electrical drives.

Predictive algorithms used in induction drive control can be classified as: Continuous Control Set and Finite Control Set MPC (FS-MPC) [2], [7]. In the first one, it is

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assumed that any value of the control signals can be obtained (Fig. 1a). This algorithm requires the use of a modulator which, through selection of appropriate switching sequences, makes it possible to get control signal generated by the control circuit. The second one involves the use of signals that can only be obtained in inverter (possible vectors for the two-level voltage source inverter are presented in Fig. 1b).
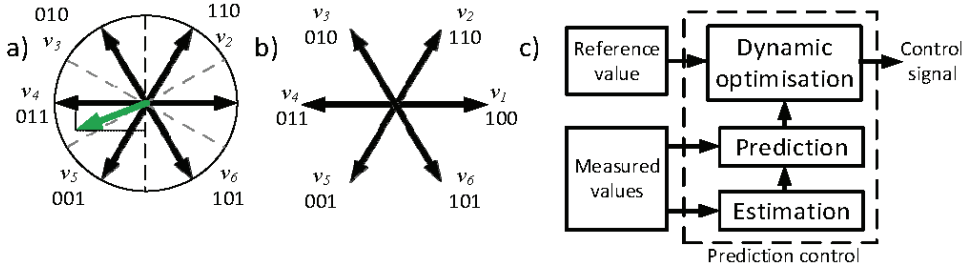

Fig. 1. Control signal available in CC-MPC (a) and FC-MPC (b). Idea of prediction control (c)

Predictive algorithms can also be classified into two main categories according to the length of the prediction horizon. In the case of FS-MPC, it is assumed that short prediction horizon refers to systems with one prediction step ($N_p = 1$) and long prediction horizon refers to systems with more than one prediction step ($N_p \geq 2$). Long prediction horizons allow better steady-state performance, reduction of current distortion and switching to be achieved. But elongation of prediction horizon significantly increases computational complexity (in the worst case exponentially) [4]–[7].

This paper presents a finite set model predictive speed control of induction motor with long prediction horizon. The weighting factors of the cost function are calculated via an optimization method using genetic algorithms. There are presented simulations results.

Similar studies have been carried out with promising results for PMSM drive systems [8]. The moment of inertia of the PMSM used in those studies was significantly smaller compared to the induction motor considered. This may cause some problems related to the different dynamics of currents and speed. The studies are designed to test the properties of the proposed system.

2. CONTROL STRUCTURE

Control structure presented in Fig. 2 consists of a finite set model predictive speed regulator, an estimator of unavailable state variables and a model of induction motor (Fig. 2). Mathematical model of induction motor is represented by the following equations

$$\mathbf{V}_s = R_s \mathbf{I}_s + \frac{d}{dt} \mathbf{\Psi}_s + j\Omega_k \mathbf{\Psi}_s, \tag{1}$$

$$\mathbf{V}_r = R_r \mathbf{I}_r + \frac{d}{dt}\mathbf{\Psi}_r + j(\Omega_k - \Omega)\mathbf{\Psi}_r , \tag{2}$$

$$\mathbf{\Psi}_s = L_s \mathbf{I}_s + L_m \mathbf{I}_r , \tag{3}$$

$$\mathbf{\Psi}_r = L_r \mathbf{I}_r + L_m \mathbf{I}_s , \tag{4}$$

$$M_e = \frac{3}{2} p \, \text{Im}\{\overline{\mathbf{\Psi}}_s \mathbf{I}_s\} , \tag{5}$$

where $\mathbf{\Psi}_s$, $\mathbf{\Psi}_r$ – vectors of stator and rotor flux, $\mathbf{I}_s$, $\mathbf{I}_r$ – vectors of stator and rotor current, $\mathbf{V}_s$, $\mathbf{V}_r$ – vectors of stator and rotor voltage, R_s, R_r – resistance of stator and rotor, L_s, L_r, L_m – inductance of stator, rotor and magnetizing, Ω – speed, Ω_k – rotation speed of the coordinate system, M_e – electromagnetic torque, p – the number of pole pairs.
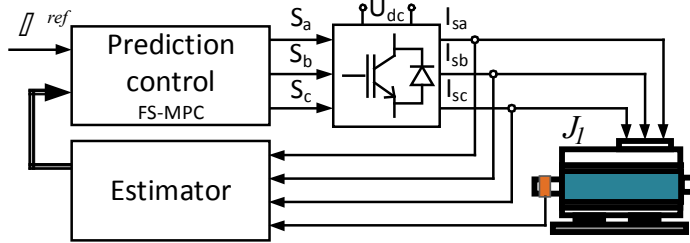

Fig. 2. Scheme of control structure

The proposed controller uses a model of the motor in *α-β* coordinates. The sequence of generation of the control signal consist of [10]: estimation of rotor flux (6), prediction of stator flux (7), stator current (8) electromagnetic torque (9) and speed (10).

$$\mathbf{\Psi}_r(k) = \frac{L_r}{L_m}\mathbf{\Psi}_s(k) + \mathbf{I}_s(k)\left(L_m - \frac{L_r L_s}{L_m}\right) , \tag{6}$$

$$\mathbf{\Psi}_s(k+1) = \mathbf{\Psi}_s(k) + T_s \mathbf{V}_s(k) - T_s R_s \mathbf{I}_s(k) , \tag{7}$$

$$\mathbf{I}_s(k+1) = \left(1 + \frac{T_s}{\tau_\sigma}\right)\mathbf{I}_s(k) + \frac{T_s}{\tau_\sigma + T_s}\left\{\frac{1}{R_\sigma}\left[\left(\frac{k_r}{\tau_r} - k_r j\Omega\right)\mathbf{\Psi}_r(k) + \mathbf{V}_s(k+1)\right]\right\} , \tag{8}$$

$$M_e(k+1) = \frac{3}{2} p \, \text{Im}\{\overline{\mathbf{\Psi}}_s(k+1)\mathbf{I}_s(k+1)\} , \tag{9}$$

$$\Omega(k+1) = \frac{T_s}{J_1}(M_e(k+1) - M_L) + \Omega(k) , \tag{10}$$

where: T_s – sample time, $\tau_\sigma = \sigma L_s/R_\sigma$, $R_\sigma = R_s + k_r^2 R_r$, $k_r = L_m/L_r$, $\tau_r = L_r/R_r$, $\sigma = 1 - L_m^2/(L_r L_s)$, k – sampling instant, J_1 – moment of inertia, M_L – load torque.

This sequence is repeated for each of the assumed steps of prediction. The last step is to estimate the value of the objective function, which in the general case takes the form (11). The optimal control vector is selected based on an assessment of the value of the objective function.

$$g = \sum_{n=1}^{N} \mu_n \left| \Omega^{\mathrm{ref}} - \Omega^p(k+n) \right| + \sum_{n=1}^{N} \lambda_n \left\| \Psi_s^{\mathrm{ref}} \right| - \left| \Psi_s^p(k+n) \right\| + \sum_{n=1}^{N} (\alpha_n f_n + \beta_n h_n + \gamma_n d_n), \quad (11)$$

where: Ω^{ref}, Ω^p – reference and predictive speed, Ψ_s^{ref}, Ψ_s^p – reference and predicted stator flux, h_n, f_n – components of penalties for exceeding the limits and switching the keys of converter, $\mu, \lambda, \alpha, \beta, \gamma$ – weighting factors, N – prediction horizon, d_n – component integrating an error as

$$d_n = [\Omega^{\mathrm{ref}} - \Omega^p(k+n)] + [\Omega^{\mathrm{ref}} - \Omega(k)], \quad (12)$$

where: Ω^{ref}, Ω^p, Ω – reference, predictive and measured speed.

In the reporting system, the control process is divided into two parts: coarse adjustment and regulation in the area of steady state. In the first case, when the speed error is greater than 3 [rad/s], the objective function is limited to stabilize the speed and flux deviations, and maintain the predetermined limits. When the speed error is less than 3 [rad/s] form (11) of objective function is used.

Components of penalties for switching the keys of converter and integrating an error have been introduced in order to reduce pulsation velocity and flux near the steady state. Smaller number of switching should result in a reduction in switching losses [7]. A long prediction horizon ($N = 2$) has also been introduced in order to smooth out the waveforms obtained.

The key issue in MPC operation is the selection of the weighting factors in the cost function [9]. Weighting factors determine the influence of particular deviations on the value of the objective function. Thus they have a key impact on dynamic parameters of the system [3]. There are no formal methods that allow the optimal weighting factors to be obtained. In this work, all weighting factors (for each step of prediction and for each control zone) have been calculated by the genetic algorithm. The objective function used in the optimization process has the form (13)

$$y = k_1 k_2 + k_1 + k_2, \quad (13)$$

$$k_1 = \frac{1}{L} \sum_{s=1}^{L} \left| \Omega - \Omega^{\mathrm{ref}} \right|, \quad (14)$$

$$k_2 = \frac{1}{L} \sum_{s=1}^{L} \left| \Psi_s - \Psi_s^{\mathrm{ref}} \right|. \quad (15)$$

The load torque and the moment of inertia can be estimated in a real system [11]. In the present work, the load torque was taken directly from the adjuster of this value. In the future works appropriate estimators will be designed. The moment of inertia is assumed to be constant.

3. SIMULATION RESULTS

During simulation studies model of induction motor represented by equations (1)–(5) has been used. It determined the current limit as $3\sqrt{2}I_n$. The specification of the induction motor is shown in Table 1. Simulations were carried out as follows:

1. At time $t = 0$ s a nominal flux value was specified.
2. At time $t = 0.1$ s some positive value of speed was specified.
3. At time $t = 0.15$ s the nominal load torque was provided.
4. At time $t = 0.2$ s the load torque was reduced to zero.
5. At time $t = 0.25$ s some negative value of speed was specified.
6. At time $t = 0.325$ s the nominal load torque was provided.
7. At time $t = 0.375$ s the load torque was reduced to zero.

Simulation results obtained for nominal speed (144 rad/s) are presented in Figs. 3–6. In Fig. 3, waveforms of speed are shown. Speed obtained in the model of induction motor (Ω) follows the trajectory determined by speed predicted in control structure (Ω_p). The speed is stabilized on the reference value (Ω_{ref}). In Fig. 4, waveforms of fluxes are presented. The fluxes obtained in model of induction motor (Ψ_s, Ψ_r – stator and rotor flux, respectively) correspond to values of estimated rotor flux (Ψ_r^p, calculated using equation (6)) and predicted stator flux (Ψ_s^p, calculated using equation (7)). The changes in stator waveform (0.07 s) are the result of selection of the weighting factors of the objective function. It means that it is more profitable not to change the control signal (set zero vector) than to stabilize the flux to the reference value. Hereby is achieved the elimination of the oscillation in waveforms of torque and speed (Fig. 5, 0.07–0.1 s). As tested, even in the absence of changes in reference values of other variables (M_L, Ω_{ref}) after a short time a stator flux returns to the reference value. The value of the flux may fall even to 0.35 [Wb]. However, after the step change in the speed or torque, the stator flux is immediately reconstructed.

Table 1. Induction machine specifications

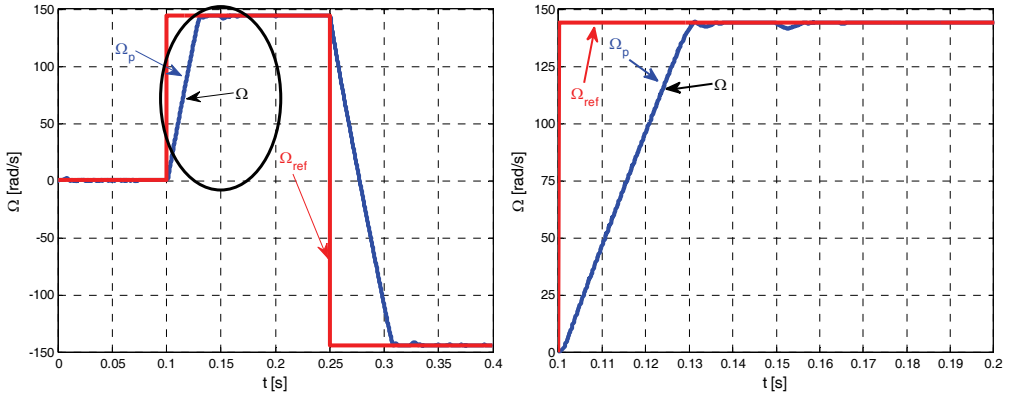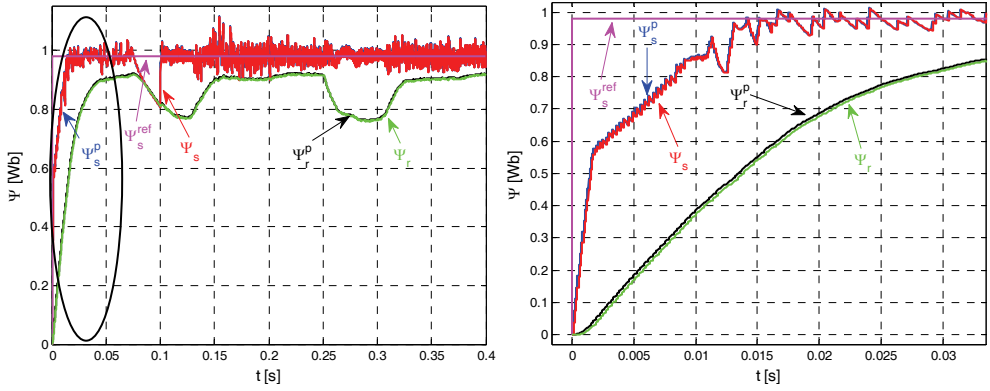| $P_n = 1100$ [kW]	$p = 2$	$L_s = 417.3$ [mH]
$U_n = 230$ [V]	$M_{en} = 7.6118$ [Nm]	$L_r = 417.3$ [mH]
$I_n = 2.9$ [A]	$R_s = 5.9$ [Ω]	$L_m = 392.5$ [mH]
$\Omega_n = 1380$ [rpm]	$R_r = 4.559$ [Ω]	$f_{sn} = 50$ [Hz]

Fig. 3. Waveforms of speeds ($\Omega = \Omega_N$)
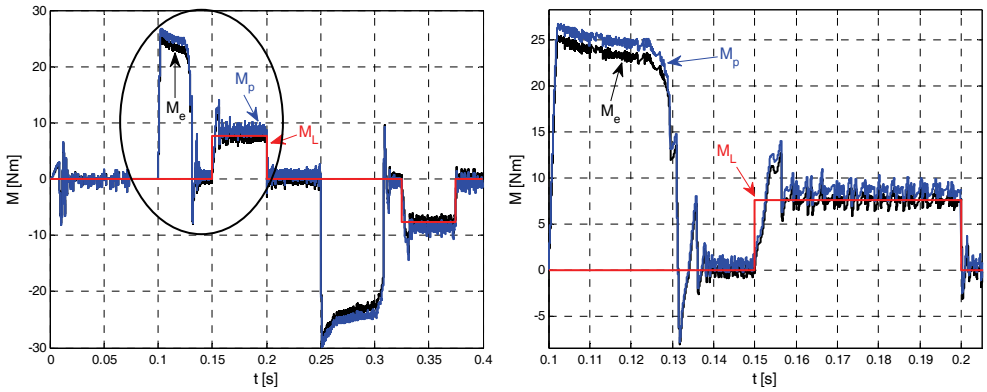

Fig. 4. Waveforms of fluxes ($\Omega = \Omega_N$)
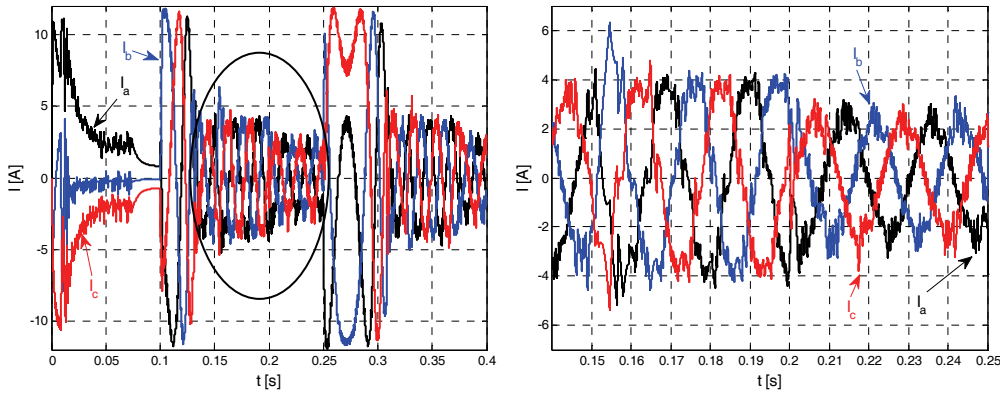

Fig. 5. Waveforms of torques ($\Omega = \Omega_N$)
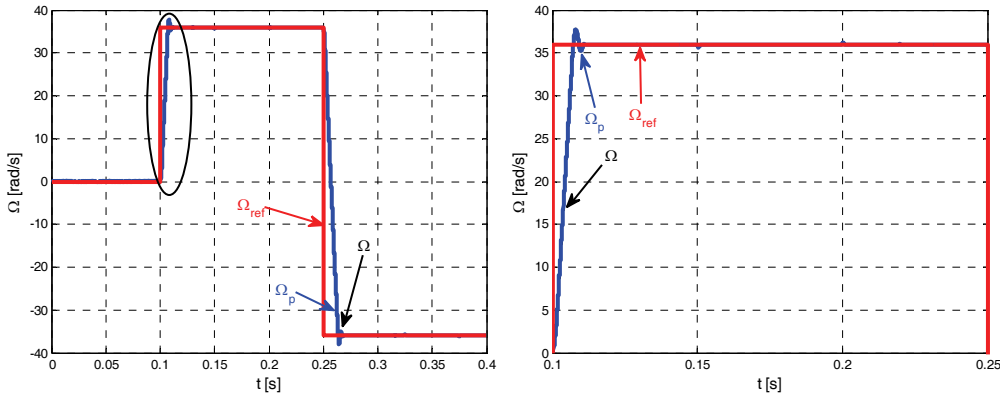
Fig. 6. Waveforms of currents ($\Omega = \Omega_N$)

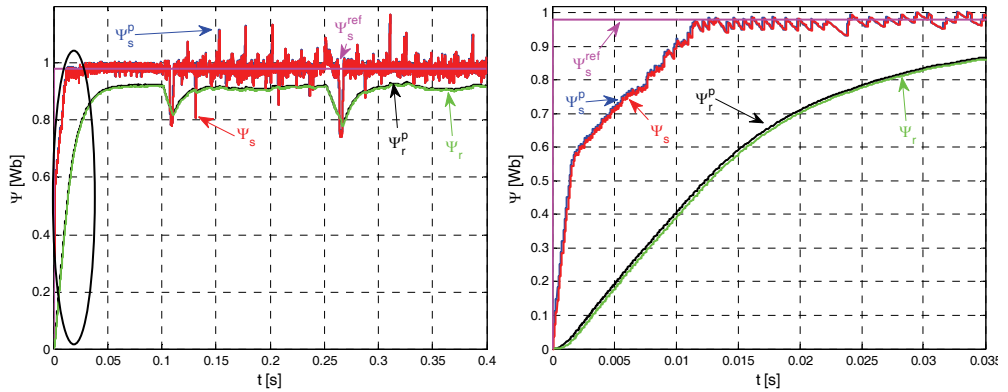

Fig. 7. Waveforms of speeds ($\Omega = 0.25\Omega_N$)

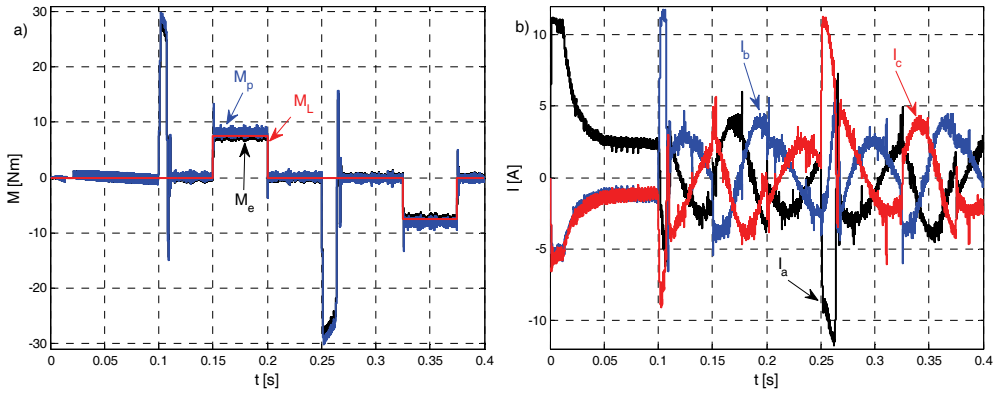

Fig. 8. Waveforms of fluxes ($\Omega = 0.25\Omega_N$)

Fig. 9. Waveforms of torques (a) and currents (b) ($\Omega = 0.25\Omega_N$)
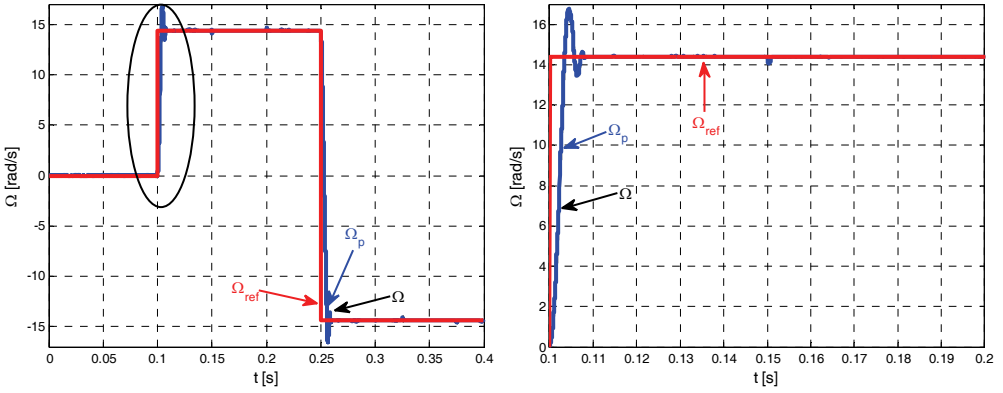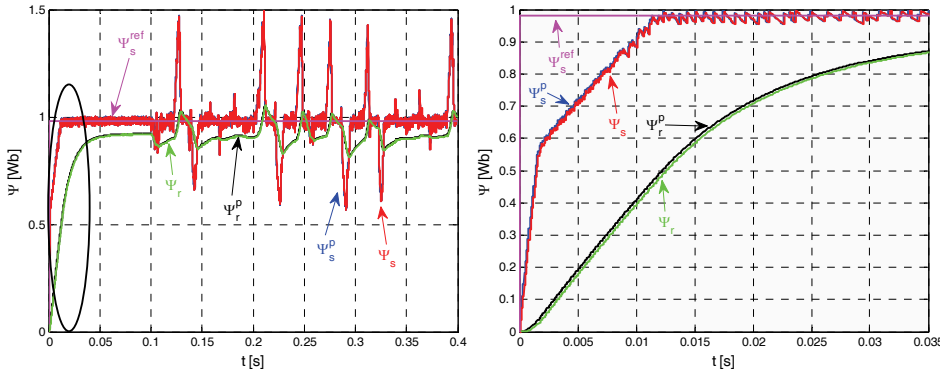

Fig. 10. Waveforms of speeds ($\Omega = 0.1\Omega_N$)
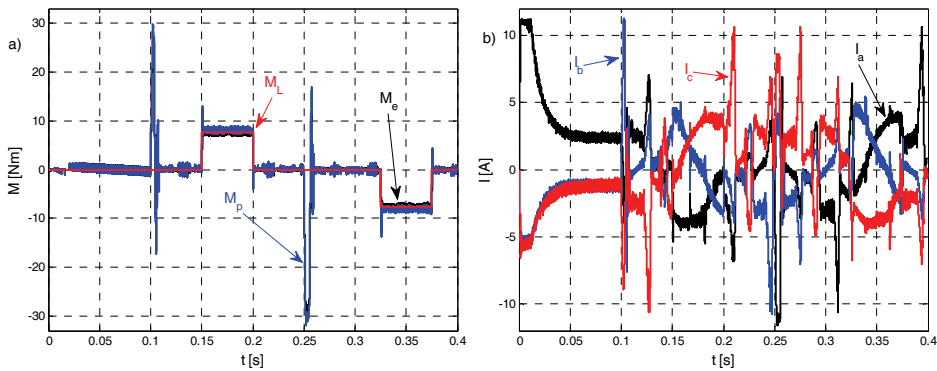

Fig. 11. Waveforms of fluxes ($\Omega = 0.1\Omega_N$)

Fig. 12. Waveforms of torques (a) and currents (b) ($\Omega = 0.1\Omega_N$)

In Fig. 6 transients of system torques are demonstrated. Electromagnetic torque (M_e) responds quickly to changes of reference values and load torque (M_L). Some inaccuracies between torque values obtained in the model of the drive and predicted in the structure of control are visible. However, they do not have significant effect on the operation of the system. Some small torque oscillations are evident in the torque transients. Oscillations are also visible in the current waveforms (Fig. 6). They result from discrete character of the control signals. Set current limits are maintained. In Figs. 7–9 and Figs. 10–12 results obtained for 25% and 10% of nominal speed are presented. For these speeds the system also works correctly and has similar property.

4. CONCLUSION

In the paper, the results of application of modified MPC algorithm for induction motor drive are presented. Contrary to commonly used approach the integral element, cost function and components of penalties are proposed in the work. Using those factors it is possible to reduce frequency of torque oscillations and thereby to reduce vibrations of speed, which results in more smooth transients. The remaining oscillations are the result of the discrete character of control signals because the converter switching states can be changed only at discrete time instant. The choice of the particular vector is a compromise between the stabilization of the flux and stabilization of the speed. The proposed algorithm gives a non-zero steady-state error [1]. The system has good dynamics, resulting from the direct character of this control strategy. In future works will be checked how the control system works in the case of motor parameter uncertainty or in the case of generating inaccurate voltage.

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