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PERFORMANCE IMPROVEMENT OF A CLASS OF POSITION DRIVES

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Abstract: *The performance of a precise DC motor position drive is discussed in this paper. A new approach to control of such drives is introduced. Using a discrete vector-matrix description of the controlled object, an optimal modal state observer has been synthesized, as well as the respective optimal modal controller. Detailed study of position drives has been carried out by means of computer simulation for the dynamic and static regimes at various loading and work conditions. Some results are presented which show that the applied method of control can provide the desired performance.*

Key words: *position control, DC drive system, optimal modal control, state observer, state controller.*

1. INTRODUCTION

A number of precise industrial applications, such as manipulators, robots, machine tools, etc., require high quality position control, without overshoot at maximum rate.

Such performance can be provided by a cascade control system, including a non-linear position controller with shifting structure [1]. Similar controller provides for maximum deceleration pace, but approaching the reference position, its gain should be limited in accordance with the condition of lack of overshoot. This, on the other hand, leads to some deterioration of the control system dynamics.

A new approach to solving this problem has been suggested in this paper, applying optimal modal control [2, 3, 5].

The procedure utilizes a combination between both - setting the closed-loop system poles (modal control) and optimal control through the quadratic quality criterion minimization, i.e. in this case a complex criterion for optimization has been introduced.

This paper discusses main problems concerning optimal modal control of precise DC motor position drives. Detailed study carried out by means of modeling and computer simulation shows that this type of control can provide the desire performance.

2. MODEL OF THE CONTROLLED OBJECT

The controlled object is an electromechanical system which consists of a four-quadrant transistor chopper and a permanent magnet DC motor. This configuration is shown in Fig. 1, where the following notations have been used: UR – uncontrollable rectifier; PWM – pulse width modulator; C – filter capacitor; T1 $\div$ T4 – electronic switches; D1 $\div$ D4 – freewheeling diodes; M – motor; PE – position encoder; L – Load; V_{dc} – DC link voltage; θ – angular position; T – motor torque; T_l – load torque.

The basic parameters of the controlled object are as follows:

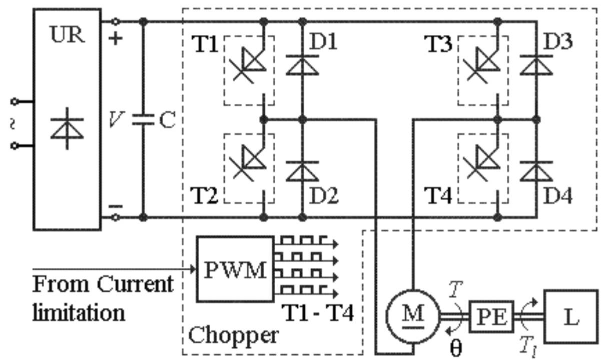


Fig. 1. The controlled object under consideration.

- armature circuit resistance $R_a = 0.61\,\Omega$;
- armature inductance $L_a = 0.003\,\text{H}$;
- back EMF coefficient $K_e = 0.191\,\text{V.s/rad}$;
- torque coefficient $K_t = 0.191\ \text{Nm/A}$;
- total inertia $J = 0.0043\ \text{kg.m}^2$;
- amplifier gain of the chopper $K_c = 3.16$;
- armature circuit time-constant $\tau_a = 0.005\ \text{s}$;
- electromechanical time-constant $\tau_m = 0.072\ \text{s}$;
- position encoder gain: $6\,000\ \text{imp/rev}$.

The rated data of the used permanent magnet DC motor are:

$$V_{\text{rat}} = 30\,\text{V},\ I_{\text{rat}} = 15.7\,\text{A},\ \omega_{\text{rat}} = 115.19\,\text{rad/s}.$$

The state-space model of the controlled object is as follows:

$$\begin{bmatrix} \dfrac{d\theta}{dt} \\ \dfrac{d\omega}{dt} \\ \dfrac{di}{dt} \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & \dfrac{k_t}{J} \\ 0 & -\dfrac{k_e}{L_a} & -\dfrac{R_a}{L_a} \end{bmatrix} \begin{bmatrix} \theta \\ \omega \\ i \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ \dfrac{k_c}{L_a} \end{bmatrix} u + \begin{bmatrix} 0 \\ -\dfrac{1}{J} \\ 0 \end{bmatrix} i_l, \quad (1)$$

where: ω is motor speed; i – armature current of the motor; u – chopper controlling code; i_l – armature current, which is determined by the respective load torque.

The following notations of state variables have been adopted: $x_1 = \theta$, $x_2 = \omega$, $x_3 = i$. Measurable coordinate in this case is the angular position θ, i.e.

$$\mathbf{y}(t) = \mathbf{C}\mathbf{x}(t),$$

where: $\mathbf{C} = \begin{bmatrix} 1 & 0 & 0 \end{bmatrix}$, $\mathbf{x}^{\mathrm{T}} = \begin{bmatrix} x_1 & x_2 & x_3 \end{bmatrix}$.

The discrete state-space model of the controlled object can be represented as follows:

$$\begin{bmatrix} x_1(k+1) \\ x_2(k+1) \\ x_3(k+1) \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \end{bmatrix} \begin{bmatrix} x_1(k) \\ x_2(k) \\ x_3(k) \end{bmatrix} + \begin{bmatrix} b_1 \\ b_2 \\ b_3 \end{bmatrix} u(k) + \begin{bmatrix} l_1 \\ l_2 \\ l_3 \end{bmatrix} i_l . \quad (2)$$

In order to use the quadratic quality criterion in the process of synthesis, the error of $e(k) = \theta_r(k) - \theta(k)$, should be formulated, where $\theta_r(k)$ is the reference position.

It is assumed that both the reference and disturbance inputs are constant, i.e. $\theta_r(k) = \mathrm{const}$ and $i_l = \mathrm{const}$. The following equation concerns the error and state variables, which are not outputs [2]:

$$\begin{bmatrix} x_{1e}(k+1) \\ x_{2e}(k+1) \\ x_{3e}(k+1) \\ x_{4e}(k+1) \end{bmatrix} = \begin{bmatrix} 1 & 1 & 0 & 0 \\ 0 & a_{11} & -a_{12} & -a_{13} \\ 0 & -a_{21} & a_{22} & a_{23} \\ 0 & -a_{31} & a_{32} & a_{33} \end{bmatrix} \begin{bmatrix} x_{1e}(k) \\ x_{2e}(k) \\ x_{3e}(k) \\ x_{4e}(k) \end{bmatrix} + \begin{bmatrix} 0 \\ -b_1 \\ b_2 \\ b_3 \end{bmatrix} u_e(k) \quad (3)$$

or

$$\mathbf{x}_e = (k+1) = \mathbf{A}_e \mathbf{x}_e(k) + \mathbf{b}_e \mathbf{u}_e(k),\ \mathrm{x}_e(0) = \mathbf{x}_{e0},\ k = 0, 1, 2, ...;$$

$$\mathbf{y}(k) = \mathbf{C}_e \mathbf{x}_e(k).$$

where:

$$\begin{aligned} x_{1e}(k) &= e(k-1) = \theta_r(k) - \theta(k-1); \\ x_{2e}(k) &= e(k) - e(k-1) = -[\theta(k) - \theta(k-1)]; \\ x_{3e}(k) &= \omega(k) - \omega(k-1); \\ x_{4e}(k) &= i(k) - i(k-1); \\ u_e(k) &= u(k) - u(k-1); \\ \mathbf{C}_e &= \begin{bmatrix} 1 & 0 & 0 & 0 \end{bmatrix}. \end{aligned} \quad (4)$$

Eq. (3) has been used for the synthesis of both an optimal modal digital observer and the respective state controller.

Based on this equation the model of the controlled object has been developed, shown in Fig. 2.

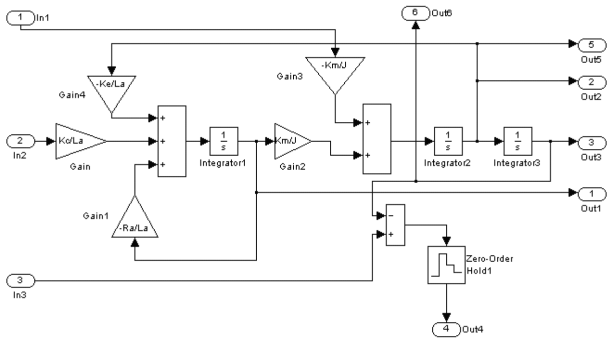


Fig. 2. Model of the controlled object.

3. SYNTHESIS OF THE CONTROL SYSTEM

3.1. State observer

Synthesis of the digital observer has been realized by an algorithm presented in [3]. This procedure utilizes the transpositioned additional object [4]:

$$\boldsymbol{\alpha}(k+1) = \mathbf{A}_e^{\mathrm{T}} \boldsymbol{\alpha}(k) + \mathbf{C}_e^{T} \beta(k) \quad (5)$$

or

$$\begin{bmatrix} \alpha_1(k+1) \\ \alpha_2(k+1) \\ \alpha_3(k+1) \\ \alpha_4(k+1) \end{bmatrix} = \begin{bmatrix} 1 & 1 & 0 & 0 \\ 0 & a_{11} & -a_{21} & -a_{31} \\ 0 & -a_{12} & a_{22} & a_{32} \\ 0 & -a_{13} & a_{23} & a_{33} \end{bmatrix} \begin{bmatrix} \alpha_1(k) \\ \alpha_2(k) \\ \alpha_3(k) \\ \alpha_4(k) \end{bmatrix} + \begin{bmatrix} 1 \\ 0 \\ 0 \\ 0 \end{bmatrix} \beta(k). \quad (6)$$

The $\mathbf{A}_e^{\mathrm{T}}$ matrix eigenvalues are determined solving the following equation:

$$\det \left\{ \begin{bmatrix} 1 & 0 & 0 & 0 \\ 1 & a_{11} & -a_{21} & -a_{31} \\ 0 & -a_{12} & a_{22} & a_{32} \\ 0 & -a_{13} & a_{23} & a_{33} \end{bmatrix} - \begin{bmatrix} \chi & 0 & 0 & 0 \\ 0 & \chi & 0 & 0 \\ 0 & 0 & \chi & 0 \\ 0 & 0 & 0 & \chi \end{bmatrix} \right\} = 0. \quad (7)$$

At quantization period of $T = 0.001$ the following eigenvalues are obtained:

$$\chi_1 = 1; \quad \chi_2 = 1; \quad \chi_3 = 0.9851; \quad \chi_4 = 0.8284.$$

In this case there are two undesired roots of the open-loop system ($\chi_1 = 1$ and $\chi_2 = 1$), which must be displaced.

Locations for the closed-loop system roots $\mu_1 = 0.1$ and $\mu_2 = 0.2$ are defined, where χ_1 and χ_2 should be placed. The locations of μ_3 and μ_4 are the same as in the open-loop system, i.e. $\mu_3 = \chi_3$ and $\mu_4 = \chi_4$.

In order to define the observer $\mathbf{H}$ matrix, it is necessary to find the elements of $\mathbf{q}_1$ and $\mathbf{q}_2$ eigenvectors corresponding to χ_1 and χ_2, respectively.

The $\mathbf{q}_1$ eigenvector is obtained solving this system of homogenous algebraic equations:

$$(\mathbf{A}_e - \mathbf{I}\chi_i)\mathbf{q}_i = 0, \text{ for } i = 1. \quad (8)$$

For the elements of both eigenvector $\mathbf{q}_1$ and weight matrix $\mathbf{Q}_1$ the following is obtained:

$$\mathbf{q}_1 = \begin{bmatrix} 1 \\ 0 \\ 0 \\ 0 \end{bmatrix},$$

$$\mathbf{Q}_1 = \mathbf{q}_1\mathbf{q}_2^{\mathrm{T}} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix}.$$

These products are computed:

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{q}_1\mathbf{q}_1^{\mathrm{T}} = \begin{bmatrix} 1 & 0 & 0 & 0 \end{bmatrix}$$

and

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{q}_1\mathbf{q}_1^{\mathrm{T}}\mathbf{b}_e = 1 .$$

Weight coefficient $r_1 = 0.1235$ and the $\lambda_1 = 1.1111$ coefficient are calculated.

After the first iteration, for the optimal modal feedback gain the following is obtained:

$$\gamma_1^{\mathrm{T}} = \begin{bmatrix} -0.9 \\ 0 \\ 0 \\ 0 \end{bmatrix}.$$

In order to displace χ_2 to location μ_2, the new system with a state matrix should be optimized:

$$\mathbf{A}_e^c = \mathbf{A}_e + \mathbf{b}_e\gamma_1 .$$

The $\mathbf{q}_2$ eigenvector is derived after solving the following system of homogeneous algebraic equations:

$$(\mathbf{A}_e^c - \mathbf{I}\,\chi_i)\mathbf{q}_i = 0, \text{ for } i = 2 . \tag{9}$$

For the elements of eigenvector $\mathbf{q}_2$ and weight matrix $\mathbf{Q}_2$ respectively, the following is obtained:

$$\mathbf{q}_2 = \begin{bmatrix} 0.7433 \\ 0.6690 \\ 0 \\ 0 \end{bmatrix},$$

$$\mathbf{Q}_2 = \mathbf{q}_1\mathbf{q}_1^{T} = \begin{bmatrix} 0.5525 & 0.4972 & 0 & 0 \\ 0.4972 & 0.4475 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix}.$$

The following products are computed:

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{q}_1\mathbf{q}_1^{\mathrm{T}} = \begin{bmatrix} 0.5525 & 0.4972 & 0 & 0 \end{bmatrix}$$

and

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{q}_1\mathbf{q}_1^{\mathrm{T}}\mathbf{b}_e = 0.5525 .$$

The respective weight coefficient $r_2 = 0.1727$ and the $\lambda_2 = 1.25$ coefficient are calculated.

After the first iteration, for the optimal modal feedback gain the following is obtained:

$$\gamma_2^{\mathrm{T}} = \begin{bmatrix} -0.8 \\ -0.72 \\ 0 \\ 0 \end{bmatrix}.$$

Since in this case there are two undesired values ($\chi_1 = 1$ and $\chi_2 = 1$), the optimal modal feedback gain becomes:

$$\gamma = \gamma_1 + \gamma_2 = \begin{bmatrix} -1.70 & -0.72 & 0 & 0 \end{bmatrix}.$$

The observer feedback vector is formulated:

$$\mathbf{H} = \begin{bmatrix} h_1 \\ h_2 \\ h_3 \\ h_4 \end{bmatrix} = \begin{bmatrix} -1.7 \\ -0.72 \\ 0 \\ 0_4 \end{bmatrix}.$$

The observer equation is as follows [4]:

$$\hat{\mathbf{x}}_e(k+1) = \mathbf{A}_e\hat{\mathbf{x}}_e(k) + \mathbf{b}_e\mathbf{u}_e(k) + \mathbf{H}\Delta\mathbf{e}(k) = $$
$$= \mathbf{A}_e\hat{\mathbf{x}}_e(k) + \mathbf{b}_e\mathbf{u}_e(k) + \mathbf{H}[y(k) - \mathbf{C}\hat{\mathbf{x}}(k)]$$

or

$$\begin{bmatrix} \hat{x}_{1e}(k+1) \\ \hat{x}_{2e}(k+1) \\ \hat{x}_{3e}(k+1) \\ \hat{x}_{4e}(k+1) \end{bmatrix} = \begin{bmatrix} 1 & 1 & 0 & 0 \\ 0 & a_{11} & -a_{12} & -a_{13} \\ 0 & -a_{21} & a_{22} & a_{23} \\ 0 & -a_{31} & a_{32} & a_{33} \end{bmatrix} \mathrm{x}$$
$$\mathrm{x} \begin{bmatrix} \hat{x}_{1e}(k) \\ \hat{x}_{2e}(k) \\ \hat{x}_{3e}(k) \\ \hat{x}_{4e}(k) \end{bmatrix} \begin{bmatrix} 0 \\ -b_1 \\ b_2 \\ b_3 \end{bmatrix} u_e(k) + \begin{bmatrix} h_1 \\ h_2 \\ h_3 \\ h_4 \end{bmatrix} \Delta e(k) \quad , \tag{10}$$

where $\Delta e(k) = \mathbf{y}(k) - \mathbf{C}\hat{\mathbf{x}}(k)$.

These equations produce the state variables valuation. Based on them the optimal modal observer has been developed. Block diagram of the observer model is shown in Fig.3.

3.2. State controller

Synthesis of the optimal modal controller has been realized by an algorithm described in [2]. In this case synthesis is carried out based on Eq. (3).

At quantization period of $T = 0.001$ for the matrix $\mathbf{A}_e$ eigenvalues, the following is obtained:

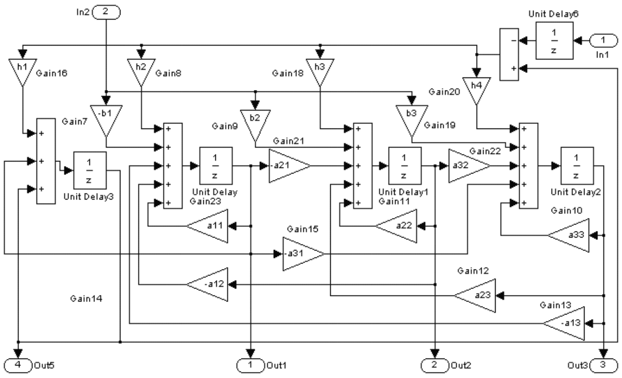


Fig. 3. Model of the optimal modal observer.

$$\chi_1 = 0.8284; \quad \chi_2 = 0.9851; \quad \chi_3 = 1; \quad \chi_4 = 1.$$

Among these values two undesired roots exist ($\chi_3 = 1$ and $\chi_4 = 1$), which should be displaced.

Locations for the closed-loop system roots $\mu_3 = 0.98$ and $\mu_4 = 0.1$ are defined, where χ_3 and χ_4 should be placed. The locations of μ_1 and μ_2 are the same as in the open-loop system, i.e. $\mu_1 = \chi_1$ and $\mu_2 = \chi_2$.

In order to determine the optimal modal controller matrix $\mathbf{K}$, it is necessary to find the elements of the eigenvector $\mathbf{q}_4$, corresponding to χ_4, as well as the eigenvector $\mathbf{q}_3$, corresponding to χ_3.

The $\mathbf{q}_4$ eigenvector is obtained after solving the following system of homogeneous algebraic equations:

$$\left(\mathbf{A}_e^{\mathrm{T}} - \mathbf{I}\chi_i\right)\mathbf{q}_i = 0, \text{ for } i = 4. \quad (11)$$

The elements of eigenvector $\mathbf{q}_4$ and weight matrix $\mathbf{Q}_4$ are obtained as follows:

$$\mathbf{q}_4 = \begin{bmatrix} 0.0000 \\ -0.9973 \\ 0.0717 \\ 0.0157 \end{bmatrix},$$

$$\mathbf{Q_4} = \begin{bmatrix} 0.0000 & 0.0000 & 0.0000 & 0.0000 \\ 0.0000 & 0.9946 & -0.0715 & -0.0156 \\ 0.0000 & -0.0715 & 0.0051 & 0.0011 \\ 0.0000 & -0.0156 & 0.0011 & 0.0002 \end{bmatrix}.$$

Products are calculated:

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{Q}_4 = [0.0000 \;\; -0.0165 \;\; 0.0012 \;\; 0.0003]$$

and

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{Q}_4\mathbf{b}_e = 2.7225 \times 10^{-4}.$$

For these coefficients the following values are obtained: $r_4 = 3.3611 \times 10^{-5}$; $\lambda_4 = 1.1111$.

The optimal modal feedback gain is determined:

$$\boldsymbol{\gamma}_1^{\mathrm{T}} = \begin{bmatrix} 0.0000 \\ 54.3987 \\ -3.9113 \\ -0.8544 \end{bmatrix}. \quad (12)$$

In order to displace χ_3 to location μ_3, the new system with a state matrix should be optimized:

$$\mathbf{A}_e^c = \mathbf{A}_e + \mathbf{b}_e\gamma_1.$$

The $\mathbf{q}_3$ eigenvector is obtained after solving the following system of homogeneous algebraic equations:

$$(\mathbf{A}_e^{c\,\mathrm{T}} - \mathbf{I}\chi_{\mathrm{i}})\mathbf{q}_i = 0, \text{for } i = 3. \quad (13)$$

For the elements of eigenvector $\mathbf{q}_3$ and weight matrix $\mathbf{Q}_3$ respectively, the following is obtained:

$$\mathbf{q}_3 = \begin{bmatrix} -0.0137 \\ -0.9999 \\ 0.0054 \\ 0.0001 \end{bmatrix},$$

$$\mathbf{Q}_3 = \mathbf{q}_3\mathbf{q}_3^{\mathrm{T}} = \begin{bmatrix} 0.0002 & 0.0137 & -0.0001 & 0.0000 \\ 0.0137 & 0.9998 & -0.0054 & -0.0001 \\ -0.0001 & -0.0054 & 0.0000 & 0.0000 \\ 0.0000 & -0.0001 & 0.0000 & 0.0000 \end{bmatrix}.$$

These products are defined:

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{q}_1\mathbf{q}_1^{\mathrm{T}} = [-0.0034 \;\; -0.2517 \;\; 0.0014 \;\; 0]$$

and

$$\mathbf{b}_e^{\mathrm{T}}\mathbf{q}_1\mathbf{q}_1^{\mathrm{T}}\mathbf{b}_e = 6.3379 \times 10^{-8}.$$

For these coefficients the following values are obtained: $r_3 = 1.5528 \times 10^{-4}$; $\lambda_3 = 50$.

At the second iteration, the optimal modal feedback gain becomes:

$$\boldsymbol{\gamma}_2^{\mathrm{T}} = \begin{bmatrix} 1.0880 \\ 79.4347 \\ -0.4324 \\ -0.0104 \end{bmatrix}.$$

Since there are two undesired values ($\chi_3 = 1$ and $\chi_4 = 1$), the optimal modal feedback gain is:

$$\boldsymbol{\gamma} = \gamma_1 + \gamma_2 = [1.0880 \;\; 133.8335 \;\; -4.3437 \;\; -0.8649].$$

The feedback vector obtains this form:

$$\mathbf{K} = \begin{bmatrix} k_1 \\ k_2 \\ k_3 \\ k_4 \end{bmatrix} = \begin{bmatrix} 1.0880 \\ 133.8335 \\ -4.3437 \\ -0.8649 \end{bmatrix}$$

and control of the following type is formulated:

$$u_e(k) = \mathbf{K}\mathbf{x}_e(k) = k_1 x_{1e} + k_2 x_{2e} + k_3 x_{3e} + k_4 x_{4e} \,. \quad (14)$$

After substitution of $u_e(k)$ in Eq. (4), for the optimal modal controller this expression is obtained:

$$u(k) = u(k-1) + k_1 x_{1e} + k_2 x_{2e} + k_3 x_{3e} + k_4 x_{4e} \,. \quad (15)$$

Analyzing Eq. (15) it can be seen, that the optimal modal controller includes an integral component into its structure. This means that when the driven mechanism is far from the reference position, the integral component would increase at each controlling cycle. This would quickly bring to saturation of the control loop. As a result, the motor will be supplied with maximum voltage. When the controlled mechanism approaches the reference position, the integral component will continue to increase and will become the dominant part of the control signal, forcing the drive to exceed the set position.

To solve this problem it is necessary to provide the following condition: when the mechanism enters some preliminary set range $(\Delta\theta_s = \theta_r - \theta)$, control signal is established to the maximum admissible value of $u_{\max}$, after which this error of $\Delta\theta_s$ is processed.

Based on these considerations, as well as on Eq. (15), the model of an optimal modal controller has been constructed. It is represented in Fig. 4.

Overtaking current limitation has been applied. The respective function is as follows:

$$u_{cl}(k) = u_n + k_m \omega(k) \,, \quad (16)$$

where: u_n is the current limitation initial code; k_m – scale coefficient.

Hence, the control condition in the presence of current limitation will be:

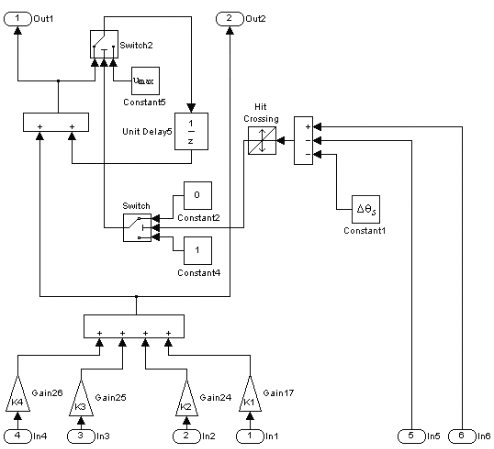


Fig. 4. Model of the optimal modal controller.

$$u_c(k) = \begin{cases} u(k) & \text{at} \quad u(k) \le u_{cl}(k); \\ u_{cl}(k) & \text{at} \quad u(k) > u_c(k). \end{cases} \quad (17)$$

In real systems the limitation set on the control signal should also be taken into account:

$$u_{cr}(k) = \begin{cases} u_c(k) & \text{for} \quad u_c(k) \le u_{\max} \\ u_{\max} & \text{for} \quad u_c(k) > u_{\max}, \end{cases} \quad (18)$$

where $u_{\max}$ is the maximum value of the control signal.

The controlling code, which should be applied to the chopper control scheme, is determined by conditions (17) and (18).

In accordance with these equations the current limitation model is composed, and it is shown in Fig. 5.

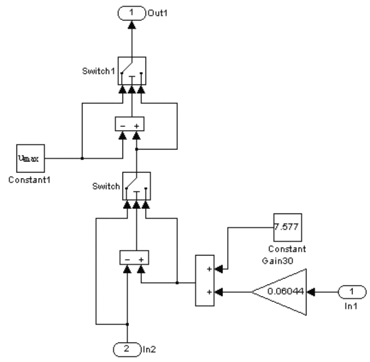


Fig. 5. Model of the current limitation.

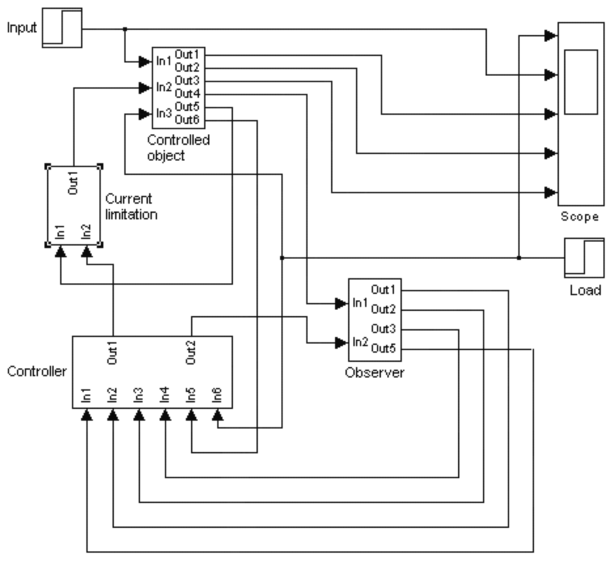
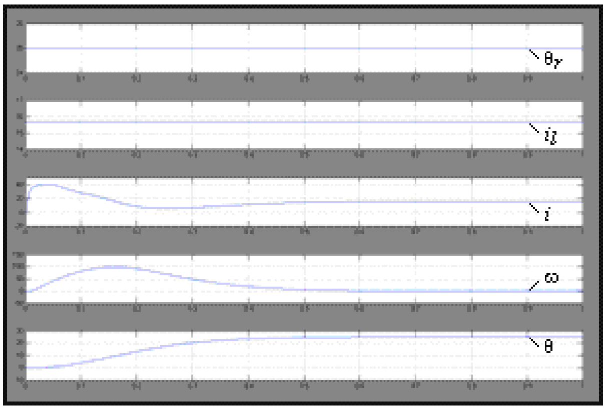


Fig. 6. Model of the position drive system under consideration.

Practically, the optimal modal control is achieved through consequent realization of Eqs. (14), (15), (16), (17) and (18).

4. DRIVE SYSTEM PERFORMANCE ANALYSIS

To prove the offered control algorithm functionality a computer simulation model has been developed, using the MATLAB/SIMULINK software package. The block diagram of the drive model is represented in Fig. 6.



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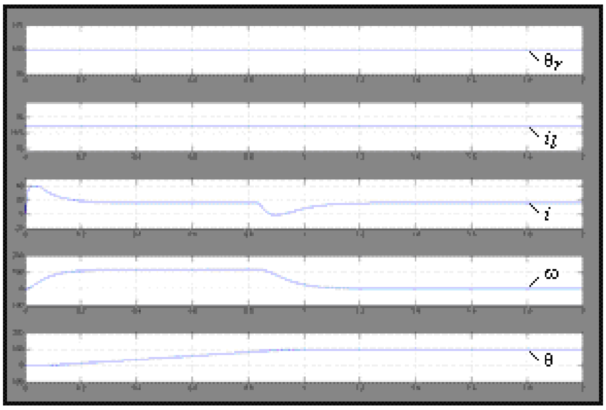

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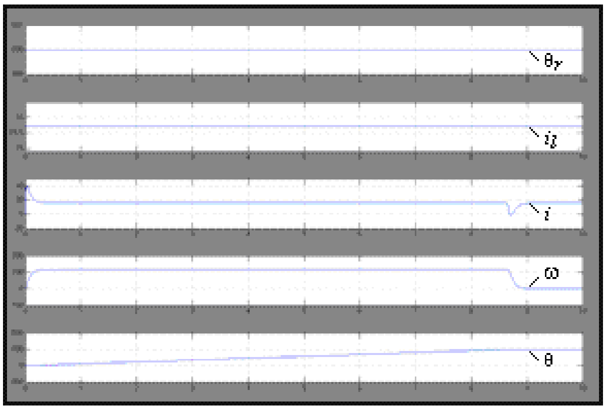

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Fig. 7. Time-diagrams illustrating the drive performance.

Detailed study of the positioning system under consideration has been carried out for the dynamic and static regimes at various loading and work conditions.

Fig. 7 shows some results illustrating the performance of the drive system. The reference angular positions are as follows: $\theta_r = 25$ rad (Fig. 7a); $\theta_r = 100$ rad (Fig. 7b); $\theta_r = 1000$ rad (Fig. 7c).

The reference static current is equal to the rated value of $I_{\text{rat}} = 15.7\ \text{A}$, and the motor speed is limited to the rated value of $\omega_{\text{rat}} = 115.19\ \text{rad/s}$.

During the respective transient regimes the armature current is limited to the maximum admissible value of $I_{a\max} = 39.25\ \text{A}$, which provides good dynamics of the drive system.

5. CONCLUSIONS

An approach to control of precise DC motor position drives is introduced in this paper.

The main features of such drive with optimal modal control have been described and discussed. The synthesis of the respective control system implements a combination between both - poles setting of the closed-loop system (modal control) and quadratic quality criterion minimization (optimal control).

Detailed study has been carried out by means of modeling and computer simulation for the respective transient and steady state regimes of operation.

The analysis shows that the represented control method provides good performance, which makes it suitable for a variety of industrial applications.

The developed simulation models as well as the results obtained can be used in the design of precise position drives. They can also be successfully applied in the process of teaching about such types of positioning systems.

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