ЕЛЕКТРОМЕХАНІЧНЕ ПЕРЕТВОРЕННЯ ЕНЕРГІЇ

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SENSORLESS SPEED CONTROL OF THE DIRECT CURRENT MOTORS

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In this paper, a new speed control algorithm for a permanent magnet DC motor which does not require implementation of the angular speed sensor is presented. Three steps are performed to develop the control system: design of speed tracking control algorithm assuming the speed measurement; design of speed observer; design of sensorless speed control algorithm based on the principle of separation. Information about speed is taken from the speed observer using the motor current value. The stability of the composite system dynamics consisting of three subsystems (the speed regulation loop, current regulation loop, and speed observer) is analyzed. The feedback gains tuning procedure for decoupling of three subsystems is given. The simulation results show that the dynamic performance of the designed system is similar to the performance of the system with angular speed measurement. The resulting closed-loop system has structural robustness properties with respect to parametric and coordinate disturbances. References 12, figures 2.
Keywords: DC motor, speed observer, estimation error dynamics, separation principle.

Introduction. Sensorless control of the electrical motors remains a relevant task of the modern control theory. As a result, a lot of papers have been written to solve the problem of sensorless control of different kinds of electric motors.

Most of the existing solutions for sensorless DC drives, as well as for other types of motors, are based on the estimation of electromotive force. The disadvantages of such systems are well known, the most significant of them is system degradation at low speeds [1].

From the viewpoint of the control theory, this problem relates to the adaptive control of linear plants [2]. The general task is to design a speed observer-based controller on the base of measurable signals: armature current and voltage. The observer in [3] uses a simplified model of DC motor, neglecting the armature inductance and current dynamics. An optimal observer [4] provides high-performance tracking only in specific operating conditions. The sliding mode observer [5], which provides local stability of speed estimation, has a considerable level of noise in the current regulation loop. The estimation algorithm [6] implements a Kalman filter. The estimation algorithm based on torque disturbance estimation [7] uses overparametization. A hybrid fuzzy-PI observer [8] does not provide an optimal solution in all operation modes of the motor. The controller [9] provides speed estimation based only on the duration of the voltage spikes in pulse-width modulation.

The solutions [4], [6], [9] do not provide asymptotic speed tracking. The papers [3] – [5] provide the stability analysis of the estimation convergence only, but none of them demonstrate proof of the close-loop system stability. The solutions [4] – [9] have complex structures of the observer and controller. The configuration is strongly dependent on the DC motor parameters. The robustness properties of the algorithms [6], [8], [9] have not been studied and established.

The paper aims to design a sensorless speed tracking control algorithm of the DC motor that provides the following properties:

– dynamic and static performance close to similar systems with speed measurement;
– cascaded structure of the control system;
– the simplicity of controller tuning.

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The purpose of the paper is to design a speed tracking sensorless control algorithm for permanent magnet DC motors that has robustness properties to parametric and coordinate disturbances. Some preliminary results are reported in [10] (see [11] as well).

Control problem formulation. The model of the DC motor with permanent magnet excitation is given by

$$\begin{aligned} \dot{\omega} &= \mu i - \frac{T_L}{J}, \\ \dot{i} &= -\frac{R}{L} i - \frac{c}{L}\omega + \frac{1}{L}u, \end{aligned} \tag{1}$$

where ω is the motor speed, i is armature current, u is control voltage, T_L is the load torque, R and L denote armature resistance and inductance respectively, c is torque constant, and $\mu = c/J$.

Considering the DC motor model (1), the following assumptions are taken into consideration:

A1. The speed reference trajectory ω^* is a smooth, bounded function together with its first $\dot{\omega}^*$ and second-time derivatives.

A2. The load torque T_L is unknown, constant, or changing slowly and limited.

A3. All motor parameters are known and constant.

A4. Current i is available for measurement; speed ω is unmeasured.

The controller for the system (1) must ensure asymptotic speed tracking, i.e. $\lim\limits_{t\to\infty}\tilde{\omega} = 0$, where $\tilde{\omega} = \omega - \omega^*$ – speed tracking error.

The algorithm is developed in the following stages: a) design of a speed control algorithm with robustness properties considering that speed is measured; b) design of the speed observer; c) development of the integrated sensorless control algorithm which consists of controller and observer; d) stability analysis of the developed closed-loop system.

Design of control algorithm. According to the back-stepping design procedure [12], the speed controller is derived first.

1. Controller with the speed measurement.

From (1) the speed error dynamics can be written as

$$\dot{\tilde{\omega}} = \mu i - \hat{T}_L - \tilde{T}_L - \dot{\omega}^*, \tag{2}$$

where $\hat{T}_L$ is the estimate of the load torque component T_L/J, and $\tilde{T}_L = T_L/J - \hat{T}_L$ is the load torque estimation error.

In case of the current fed condition, a speed controller for the system (2) is

$$\begin{aligned} i &= \frac{1}{\mu}\left(-k_\omega\tilde{\omega} + \hat{T}_L + \dot{\omega}^*\right), \\ \dot{\hat{T}}_L &= -\dot{\tilde{T}}_L = -k_{\omega i}\tilde{\omega}, \end{aligned} \tag{3}$$

where $(k_\omega, k_{\omega i}) > 0$ are speed controller proportional and integral gains.

From (2) and (3) the speed loop error dynamics is given by

$$\begin{aligned} \dot{\tilde{T}}_L &= k_{\omega i}\tilde{\omega}, \\ \dot{\tilde{\omega}} &= -\tilde{T}_L - k_\omega\tilde{\omega}. \end{aligned} \tag{4}$$

The system (4) is asymptotically stable $\forall (k_\omega, k_{\omega i}) > 0$, i.e. $\lim\limits_{t\to\infty}\left(\tilde{T}_L, \tilde{\omega}\right) = 0$.

The armature current is not the real control action in (2), so (3) can be considered as the reference i^* for the current i. Defining current tracking error as

$$\tilde{i} = i - i^*, \tag{5}$$

system (4) becomes

$$\dot{\tilde{T}}_L = k_{\omega i}\tilde{\omega},$$
$$\dot{\tilde{\omega}} = -\tilde{T}_L - k_\omega \tilde{\omega} + \mu \tilde{i}. \tag{6}$$

The control voltage in (1), formed by the current controller, should guarantee the current error tracking, i.e. $\lim\limits_{t\to\infty} \tilde{i} = 0$. The equation (1) in error form is presented as

$$\dot{\tilde{i}} = -\frac{R}{L}\tilde{i} + \frac{1}{L}u - \frac{R}{L}i^* - \frac{c}{L}\omega - \dot{i}_1^* - \dot{i}_2^*, \tag{7}$$

where the reference current derivative is divided into the known function $\dot{i}_1^*$ and the unknown term $\dot{i}_2^*$ is defined from (3) as

$$\dot{i}_1^* = \frac{1}{\mu}\left(-k_\omega\left(-k_\omega\tilde{\omega} + \mu\tilde{i}\right) + \dot{\hat{T}}_L + \ddot{\omega}^*\right),$$
$$\dot{i}_2^* = \frac{k_\omega}{\mu}\tilde{T}_L. \tag{8}$$

The current controller is constructed from (7) and (8) as

$$u = L\left(\frac{R}{L}i^* + \frac{c}{L}\omega + \dot{i}_1^* - k_{i1}\tilde{i} - y\right),$$
$$\dot{y} = k_{ii}\tilde{i}, \tag{9}$$

where y is the current controller integral term, $\left(k_{i1}, k_{ii}\right) > 0$ are the proportional and integral gains of the current controller.

From (6), (7) – (9) the resulting closed-loop error dynamics is given by

$$\begin{pmatrix} \dot{\tilde{T}}_L \\ \dot{\tilde{\omega}} \\ \dot{y} \\ \dot{\tilde{i}} \end{pmatrix} = \left[\begin{array}{cc|cc} 0 & k_{\omega i} & 0 & 0 \\ -1 & -k_\omega & 0 & \mu \\ \hline 0 & 0 & 0 & k_{ii} \\ -k_\omega/\mu & 0 & -1 & -k_i \end{array}\right]\begin{pmatrix} \tilde{T}_L \\ \tilde{\omega} \\ y \\ \tilde{i} \end{pmatrix}, \tag{10}$$

where $k_i = R/L + k_{i1}$.

The linear system (10) is asymptotically stable with the suitable tuning of the speed and current controllers gains $\left(k_\omega, k_{\omega i}\right)$, $\left(k_{i1}, k_{ii}\right)$. It is known from the theory of cascaded systems that the dynamics of the current closed-loop, given by the two last equations in (10), should be at least two times faster than speed control loop dynamics.

Time-scale separation between speed and current dynamics may be obtained using the standard frequency-domain approach on the base of the characteristic equation:

$$p^2 + k_p p + k_{in} = 0, \tag{11}$$

where $\left(k_p, k_{in}\right)$ are proportional and integral gains of controllers.

The tuning of each second-order subsystem is $k_{in} = k_p^2/4$ for damping factor $\xi = 1$, and $k_{in} = k_p^2/2$ for $\xi = 0.707$. The resulting relationship between the natural frequency of undamped oscillations $\omega_0^2 = k_{in}$ becomes $\omega_{0i} = \left(2 \div 4\right)\omega_{0s}$, where ω_{0i}, ω_{0s} stand for current and speed loops respectively.

The resulting closed-loop error dynamics (10) has structural robustness properties with respect to parametric and coordinate disturbances. This is due to the cascaded connection of second-order systems with the two time-scale separations. If initial conditions are set to zero, the system (10) tracks speed references without errors.

It follows from the analysis that asymptotic speed tracking is guaranteed if assumptions A1, A2, A3 are satisfied and speed is available for measurement.

2. Speed observer.

Let us define the current and speed estimation errors as

$$e_i = i - \hat{i}, \ e_\omega = \omega - \hat{\omega}, \tag{12}$$

and speed observer in the following form:

$$\begin{aligned} \dot{\hat{\omega}} &= \mu i - \hat{T}_L - k_1 e_i, \\ \dot{\hat{i}} &= \frac{1}{L}\left(u - R\hat{i} - c\hat{\omega}\right) + k_2 e_i, \end{aligned} \tag{13}$$

where $\left(k_1, k_2\right) > 0$.

From (1) and (13), the estimation error dynamics can be written as

$$\begin{aligned} \dot{e}_\omega &= k_1 e_i - \tilde{T}_L, \\ \dot{e}_i &= -k_{0i} e_i - \frac{c}{L} e_\omega, \end{aligned} \tag{14}$$

where $k_{0i} = R/L + k_2$.

Under the condition $\tilde{T}_L = 0$, the globally stable solution of (14) is $e_i = e_\omega = 0$. At the same time, the observer (14) is not asymptotic due to the presence of perturbation $\tilde{T}_L$.

To prove the robustness properties of the system (14) with respect of $\tilde{T}_L$, the following coordinate transformation is considered

$$\begin{aligned} z &= -k_{0i}\eta - e_\omega, \\ \eta &= \frac{L}{c} e_i. \end{aligned} \tag{15}$$

The error dynamics (14) in new coordinates (15) is given by

$$\begin{aligned} \dot{e}_\omega &= -\frac{k_{02}}{k_{0i}} e_\omega - \frac{k_{02}}{k_{0i}} z - \tilde{T}_L, \\ \dot{z} &= -\left(k_{0i} - \frac{k_{02}}{k_{0i}}\right) z + \frac{k_{02}}{k_{0i}} e_\omega + \tilde{T}_L, \end{aligned} \tag{16}$$

where $k_{02} = c k_1 / L$.

Defining $k_{02} = k_{0i}^2/2$, equations (16) become

$$\begin{aligned} \dot{e}_\omega &= -\frac{k_{0i}}{2} e_\omega - \frac{k_{0i}}{2} z - \tilde{T}_L, \\ \dot{z} &= -\frac{k_{0i}}{2} z + \frac{k_{0i}}{2} e_\omega + \tilde{T}_L. \end{aligned} \tag{17}$$

From (17) it can be concluded that $\left\|x(t)\right\| = \left\|x(0)\right\| e^{-(k_{0i}/2)t}$ under conditions of $\tilde{T}_L = 0$, where $x = \left(e_\omega, z\right)^T$. Consequently, the load disturbance can be arbitrarily attenuated by increasing k_{0i}. It should be noted that the damping factor for (17) is $\xi = 0.707$.

3. Composite electromechanical system on the base of separation principle.

Substituting $\hat{\omega}$ instead of ω in (3) and (9), the speed estimation error is defined as

$$\tilde{\tilde{\omega}} = \hat{\omega} - \omega^*. \tag{18}$$

The sensorless speed controller (3) can be presented in the following form:

$$\begin{aligned} i^* &= \frac{1}{\mu}\left(-k_\omega \tilde{\tilde{\omega}} + \hat{T}_L + \dot{\omega}^*\right), \\ \dot{\hat{T}}_L &= -\dot{\tilde{T}}_L = -k_{\omega i}\tilde{\tilde{\omega}}. \end{aligned} \tag{19}$$

Taking into account that $\tilde{\tilde{\omega}} = \tilde{\omega} - e_\omega$, the system (6) becomes

$$
\begin{aligned}
&\dot{\tilde{T}}_L = k_{\omega i}\tilde{\omega} - k_{\omega i}e_\omega,\\
&\dot{\tilde{\omega}} = -\tilde{T}_L - k_\omega\tilde{\omega} + \mu\tilde{i} + k_\omega e_\omega.
\end{aligned} \tag{20}
$$

The current controller is formed similarly to (9) using estimated speed $\hat{\omega}$ as

$$
\begin{aligned}
&u = L\left(\frac{R}{L}i^* + \frac{c}{L}\hat{\omega} + \dot{i}^* - k_{i1}\tilde{i} - y\right),\\
&\dot{y} = k_{ii}\tilde{i}.
\end{aligned} \tag{21}
$$

The current derivative $\dot{i}^*$ is known from the solution of the equation (19)

$$
\begin{aligned}
&\dot{i}^* = \frac{1}{\mu}\left(-k_\omega\dot{\tilde{\tilde{\omega}}} + \dot{\hat{T}}_L + \ddot{\omega}^*\right),\\
&\dot{\tilde{\tilde{\omega}}} = -k_\omega\tilde{\tilde{\omega}} - k_1 e_i + \mu\tilde{i}.
\end{aligned} \tag{22}
$$

Substituting the current control algorithm (21), (22) to (7), and taking into account (17) and (20), the resulting tracking and estimation error dynamics becomes

$$
\begin{pmatrix} \dot{\tilde{T}}_L \\ \dot{\tilde{\omega}} \\ \dot{y} \\ \dot{\tilde{i}} \\ \dot{e}_\omega \\ \dot{z} \end{pmatrix}
=
\left[\begin{array}{cc:cc:cc}
0 & k_{\omega i} & 0 & 0 & -k_{\omega i} & 0 \\
-1 & -k_\omega & 0 & \mu & k_\omega & 0 \\
\hdashline
0 & 0 & 0 & k_{ii} & 0 & 0 \\
0 & 0 & -1 & -k_i & -c/L & 0 \\
\hdashline
-1 & 0 & 0 & 0 & -k_{0i}/2 & -k_{0i}/2 \\
1 & 0 & 0 & 0 & k_{0i}/2 & -k_{0i}/2
\end{array}\right]
\begin{pmatrix} \tilde{T}_L \\ \tilde{\omega} \\ y \\ \tilde{i} \\ e_\omega \\ z \end{pmatrix}. \tag{23}
$$

The composite system (23) represents a cascaded connection of three subsystems: speed regulation loop, current regulation loop, and speed observer. If the observer dynamics is at least 3–4 times faster than the dynamics of the current loop then the dynamics of the composite system is similar to the system with the speed measurement, i.e. when $e_\omega = 0$. This corresponds to the

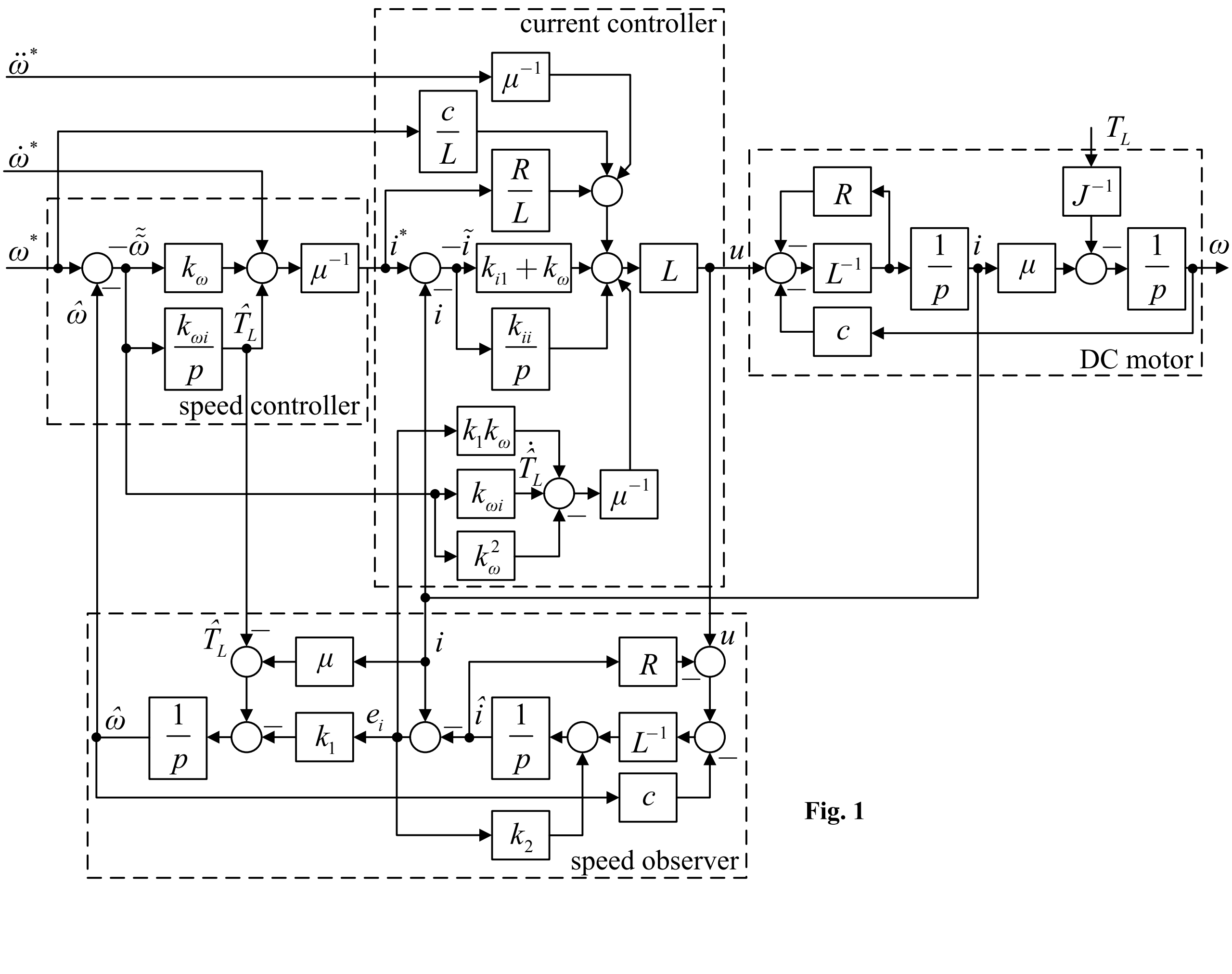

Fig. 1

following natural frequency of undamped oscillations: $\omega_{0o} = (3 \div 4)\omega_{0i}$, $\omega_{0o}^2 = k_{0i}^2/2$, where ω_{0o} related to observer dynamics. In (23) it was assumed that $c/L \ll k_i$, which usually takes place in practice.

The block diagram of the designed system is presented in Fig. 1. It consists of the speed controller (19), current controller (21), speed observer (13) , and DC motor (1).

Simulation results. The designed control algorithm was applied for DC motor, whose rated data are $P_N = 500$ W, $\omega_N = 100$ rad/s, $R = 1$ Ohm, $L = 5$ mH, $J = 0.01$ kg m^2, $c = 1$ Nm/A.

The controllers' parameters were set at: for speed controller (19): $k_\omega = 200$, $k_{\omega i} = k_\omega^2/2$; for current controller (21): $k_{i1} = 1000$, $k_{ii} = k_i^2/2$; tuning parameters of the speed observer (13) are $k_2 = 2000$, $k_1 = k_2^2/2$.

The operating sequences have been configurated as following: at the initial time the unloaded motor is required to track the speed reference trajectory, starting from zero initial value and reaching a rated value of 100 rad/s at time $t = 0.15$ s. Dynamic torque during speed transient equals double of rated value. At time $t = 0.3$ s, a rated load torque of 5 Nm is applied; at time $t = 0.5$ s load torque is set to zero.

Transients of speed trajectory tracking are depicted in Fig. 2. From Fig. 2 it follows that the speed tracking error only occurs at the moments of time when applying the load. Transient $\tilde{\omega}(t)$ is almost the same as for the algorithm (3), (8) with speed sensor because the speed estimation error is negligibly small.

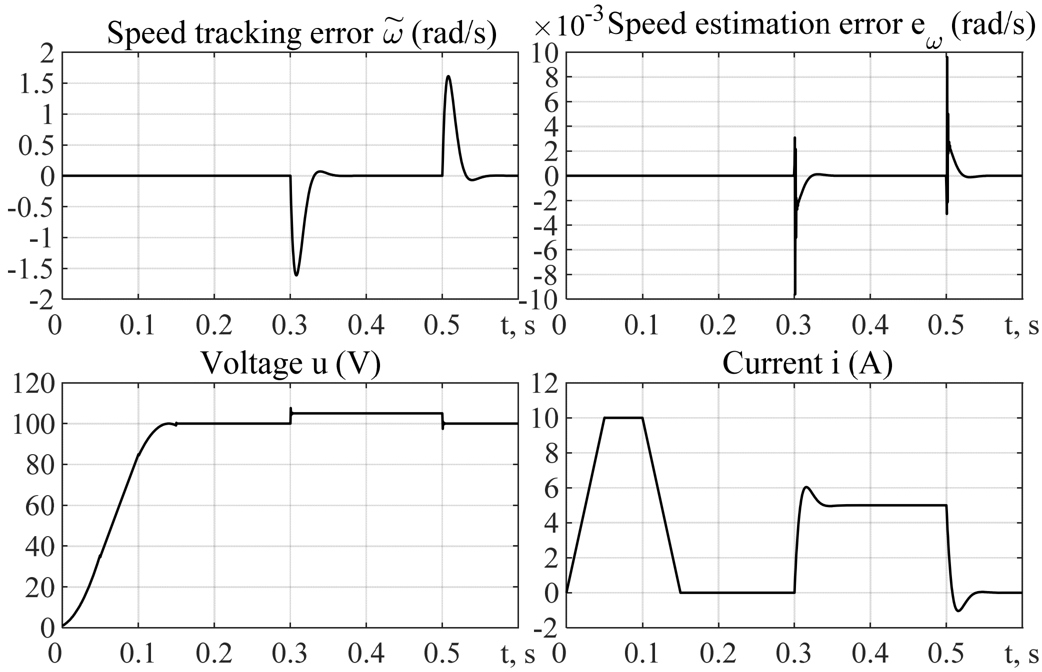

Fig. 2

Conclusions. A novel sensorless speed tracking control algorithm for a permanent magnet DC motor has been designed. Development of the algorithm performed in three steps: design of the speed tracking control algorithm considering using speed measurements; design of the robust speed observer; design of sensorless speed control algorithm on the base of separation principle. The dynamic performance of the designed system is close to the performance of the system with speed measurement. The proposed algorithm ensures system robustness to parametric disturbances as well.

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БЕЗДАВАЧЕВЕ КЕРУВАННЯ ДВИГУНОМ ПОСТІЙНОГО СТРУМУ

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У роботі розроблено алгоритм керування швидкістю ДПС з незалежним збудженням, який не вимагає використання давача швидкості. Розробка алгоритму відбувається в три етапи: синтез алгоритму керування кутовою швидкістю за умови вимірюваності швидкості; синтез спостерігача швидкості; синтез алгоритму бездавачевого керування швидкості на основі принципу розділення. Оцінювання швидкості забезпечується завдяки запропонованому замкненому спостерігачу на основі інформації про струм двигуна. Проаналізовано стійкість композитної системи (контур регулювання кутової швидкості, контур регулювання струму і спостерігач кутової швидкості). Запропоновано процедуру конфігурації коефіцієнтів зворотних зв'язків алгоритму для досягнення розділення в часі процесів у трьох підсистемах. Результати моделювання свідчать, що запропонована система керування без використання давача швидкості забезпечує динамічні показники, які є близькими до отриманих у системах на основі вимірювання кутової швидкості. Результуюча замкнена система має властивості робастності до параметричних та координатних збурень. Бібл. 12, рис. 2.
Ключові слова: двигун постійного струму, спостерігач кутової швидкості, динаміка похибок оцінювання, принцип розділення.

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