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SPEED VECTOR CONTROL SYSTEM OF SWITCHED INDUCTOR-TYPE ELECTRICAL DRIVE

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Purpose. To propose the method of development of control laws for the electrical drive with switched inductor motor with independent electromagnetic excitation, and to study dynamic behaviour of the developed electromechanical system. **Methodology.** Identification of control laws is based on an idea of the reversibility of the Lyapunov direct method for the stability analysis, and using the instantaneous value of energy as the predetermined Lyapunov function. Modelling was used to study the control system's performance. **Results.** The control laws have been identified and respective current and speed controllers have been developed, allowing a lesser sensitivity to variations of the motor's parameters, as well as the simplicity of realization of control system. Modelling of the proposed electrical drive confirmed good control performance and the workability of the system. **Originality.** Although the known elements of the control theory and vector control are used, this is the first time when these tools are applied to the development of the considered type electrical drive. **Practical value.** This study will contribute to further development of electrical drive based on switched inductor motor with independent electromagnetic excitation, and consequently support its practical application in various industries, as a good alternative to the most widely used induction motor drive. References 11, figures 9.

Key words: electrical drive, switched inductor motor, control laws, speed vector control, sensitivity to parameters' variation.

СИСТЕМА ВЕКТОРНОГО КЕРУВАННЯ ШВИДКІСТЮ ВЕНТИЛЬНО-ІНДУКТОРНОГО ЕЛЕКТРОПРИВОДА

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Електричний привод споживає близько 65 % загального обсягу виробленої у світі електроенергії. Асинхронний електропривод є найбільш поширеним у різних галузях промисловості, проте володіє відомими недоліками, зокрема невисоким коефіцієнтом потужності. Електричний привод на основі вентильно-індукторного двигуна з незалежним електромагнітним збудженням може бути альтернативою для асинхронного приводу завдяки низці переваг, пов'язаних з енергетичною ефективністю. На електропривод діють координатні та параметричні збурення, які можуть вплинути на ефективність перетворення енергії та якість керування в різних технологічних процесах. Запропоновано метод розробки законів керування на основі ідеї зворотності прямого методу Ляпунова для аналізу стійкості й мінімізації миттєвого значення енергії як заданої функції Ляпунова. Це забезпечує ефективну роботу з меншою чутливістю до зміни параметрів двигуна, а також простоту реалізації системи керування. Моделювання запропонованого електричного приводу підтвердило високу якість керування й технологічність системи. Результати даного дослідження будуть сприяти подальшому впровадженню вентильно-індукторного електропривода в промислових технологіях.

Ключові слова: електропривод, вентильно-індукторний двигун, закони керування.

PROBLEM STATEMENT. In many industries, the promising alternative to the most widely used induction electrical drive is one based on a switched inductor motor. Main advantages of this type of motors [1–6] are: high efficiency factor within a wide speed range; power factor is about 100 %; a simple design and low production costs; high manufacturability and reliability; a wider speed control range in a zone of reduced magnetic flux; an easier heat removal.

There are several types of inductor motors. This study is dedicated to switched inductor motor with independent electromagnetic excitation (SIM IE). It has a passive rotor with tooth structure and a stator with a classic distributed "star" 3-phase winding. Additionally, there is an excitation winding which is supplied from a direct current source. Electromagnetic flow of this motor, in its nature, is active.

Miscalculations during identification of the parameters of the equivalent circuit of the SIM IE can be caused by assumptions used in an applied methodology, as well as by the lack of basic information. During the motor operation, resistance of windings may be changed because of heating, and inertia moment may be deviated through changes of the kinematics. These parametric deviations resulted in differences between estimated and actual parameters of the electrical drive, which, in turn, leads to worsening of control performance.

Naturally, the SIM IE, as well as other types of alternating current motors, is an interrelated controlled object, substantially dependent on influence of inducted eddy currents. In this case, electrical drive control requires compensation of negative influence of these coordinate disturbances.

Solution of the above mentioned problems by the classic methods of the automatic control theory, under the under conditions of uncertainties in a mathematical model, is rather complicated because requires additional algorithms of identification, adaptation or compensation.

Analysis of methods for control law optimization showed [7] that solutions can be found based on a concept of reverse task of dynamics in combination with minimization of local functionals of instantaneous values of energies [8–11].

The reverse task of dynamics is to identify the control law which would ensure a given quality of control with desired static and dynamic performance of the system. The proposed method is based on an idea of the reversibility of the Lyapunov direct method for the stability analysis. This allows defining control laws which ensure that a closed loop has the predetermined Lyapunov function in form of the instantaneous value of energy. In this case, the specificity of optimization is not obtaining the absolute minimum of the quality functional, as usually used in traditional systems, but rather getting a certain minimal value which would assure a technically allowable dynamic error of the system.

This paper is aimed at the identification of respective control laws which would allow a lesser sensitivity to variations of the motor's parameters, as well as the simplicity of realization of the control system, and consequently ensure good control performance of electrical drive, required for most of industrial technologies.

EXPERIMENTAL PART AND RESULTS OBTAINED. A mathematical model of SIM IE in the coordinate system (d-q), oriented by the rotor magnetic axis, can be described by known non-liner equation system (1).

$$\begin{cases} \dfrac{di_d}{dt} = \dfrac{1}{L_s}\left(u_d - R_s i_d + L_m \dfrac{di_f}{dt} - \omega\psi_q\right); \\ \dfrac{di_q}{dt} = \dfrac{1}{L_s}\left(u_q - R_s i_q + \omega\psi_d\right); \\ \dfrac{di_f}{dt} = \dfrac{1}{L_f}\left(u_f - R_f i_f + L_m \dfrac{di_d}{dt}\right); \\ \dfrac{d\omega_r}{dt} = \dfrac{1}{J}\left(M - M_c\right); \\ \psi_d = L_s i_d + L_m i_f; \\ \psi_q = L_s i_q; \\ \psi_f = L_f i_f + L_m i_d; \\ M = \sqrt{3} Z_p \left[\psi_d i_q - \psi_q i_d\right], \end{cases} \tag{1}$$

where i_d, i_q and u_d, u_q are d-axis and q-axis stator currents and voltages respectively; i_f and u_f are excitation current and voltage; $\omega = Z_p\omega_r$ and ω_r are electrical and angular rotor speed; Z_p is pole couple number; J is inertia moment; M, M_c are electromagnetic motor torque and load torque; ψ_d, ψ_q, ψ_f are d- and q-axis, as well as excitation winding fluxes; L_s, L_f, L_m are stator, excitation winding and mutual inductance; R_s, R_f are stator and excitation winding resistances.

It is shown from (1) that motor's coordinates are interrelated because of the existing nonlinearity caused by the operation of multiplication and coordinative disturbances. In classic control systems, compensation of the negative influence of coordinative disturbances is to be realized through setting specific feedbacks, the effectiveness of which depends on the accuracy of motor's parameters. It is also possible to identify control laws based on the static decomposition of the controlled object (1) resulting in complication of the control system. In this study, solution is being found through the dynamic decomposition [11], using optimization method proposed in [8]. In this case, the initial system (1) should be transformed into system (2).

During the control system design, coordinate deviations $F_1 = L_m \dfrac{di_f}{dt} - \omega\psi_q$, $F_2 = \omega\psi_d$, and $F_3 = L_m \dfrac{di_d}{dt}$ are usually interpreted as indeterminate, but value limited $F_1 \le F_{1\max}$, $F_2 \le F_{2\max}$, $F_3 \le F_{3\max}$, while values of control signals u_d, u_q, u_f are sufficient for their compensation. In this case, a problem to control the interrelated controlled object (1) comes to finding solution of local tasks to control four liner subsystems (2):

$$\begin{cases} L_s \dfrac{di_d}{dt} + R_s i_d = u_d + F_1; \\ L_s \dfrac{di_q}{dt} + R_s i_q = u_q + F_2; \\ L_f \dfrac{di_f}{dt} + R_f i_f = u_f + F_3; \\ J\dfrac{d\omega_r}{dt} = M - M_c; \\ \psi_d = L_s i_d + L_m i_f; \\ \psi_q = L_s i_q; \\ \psi_f = L_f i_f + L_m i_d; \\ M = \sqrt{3} Z_p \left[\psi_d i_q - \psi_q i_d\right]. \end{cases} \tag{2}$$

From the control point of view, the SIM IE is similar to a classic synchronous motor, and its constructive features allow applying the direct vector control system with rotor position orientation.

The vector control system, according to first four differential equations of the system (2) consist of four control loops: for stator d-axis current i_d, q-axis current i_q, excitation current i_f, and motor speed ω_r. The speed loop is external to the internal loop of current i_q. This current defines a value of the electromagnetic torque of a motor. The excitation current i_f can be easily controlled within the range 1:8. This allows increasing a range of speed control with a constant power, in comparison with induction motor.

An object of the local control loop for the stator current i_d according to the 1st equation of the system (2)

$$L_s \frac{di_d}{dt} + R_s i_d = u_d + F_1 \tag{3}$$

can be described by the first order linear differential equation with control signal u_d and disturbance F_1. A desired equation of the closed current loop, which defines expected control performance, can also be described by the first order differential equation [4, 5]

$$\dot z + \alpha_{0i_d} z = \alpha_{0i_d} i_d^*, \tag{4}$$

where i_d^* is referenced current. The equation (4) enables a type 1 astatic system for control variable, as well as smooth (with no overcontrol) current transients. Required transient time $t_n \approx 3/\alpha_{0i_d}$, is defined only by the coefficient α_{0i_d}.

The extent to which the real current control process is close to desirable one can be estimated through the functional, which depends on inductance-normalized instantaneous energy of the magnetic field by the 1st derivation of the current

$$G(u_d) = \frac{1}{2}\left[\dot z(t) - \dot i_d(t)\right]^2. \tag{5}$$

To minimize the functional, the gradient law of the 1st order can be used

$$\frac{du_d(t)}{dt} = -\lambda_{i_d} \frac{dG(u_d)}{du_d}, \tag{6}$$

where λ_{i_d} is a constant.

Substituting (3) and (5) into (6), the control law for the current i_d can be obtained i_d

$$\dot u_d(t) = k_{i_d}(\dot z - \dot i_d), \tag{7}$$

where $k_{i_d} = \lambda_{i_d}/L_s$ is the gain coefficient of the controller.

Necessary condition for a convergence of the functional minimization process with $t \to \infty$:

$$\frac{dG(u_d)}{dt} < 0; \quad G(u_d) \to 0 \tag{8}$$

is met according to a mark rule

$$sign(k_{i_d}) = sign(1/L_s). \tag{9}$$

A variable $\dot z$ in the control law (7) plays a role of a necessary derivative on the current, which can be found in real time from the equation (4) through closing feedback on the current component $z = i_d$

$$\dot z = \alpha_{0i_d}(i_d^* - i_d). \tag{10}$$

Integrating both parts of the equation (7) and taking into account (10), the control law for the current i_d can be finally obtained

$$u_d(t) = k_{i_d}(z - i_d); \quad z = \alpha_{0i_d}\int_0^t (i_d^* - i_d)dt. \tag{11}$$

The block diagram of the current i_d controller, based on the equation (11) is presented in Fig. 1.

Contrary to classic controllers, the designed one does not contain parameters of the controlled object (1), and has only the parameter α_{0i_d} which defines the desired equation of the closed-loop system performance (4).

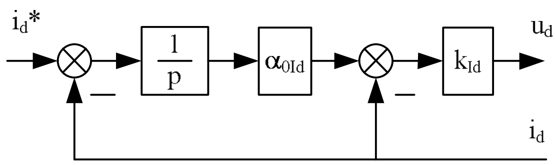

Figure 1 – Block-diagram of current controller i_d

The differential equation of closed control loop of the current i_d can be derived through substituting the control law (11) to (3):

$$\ddot i_d + (R_s/L_s + k_{i_d}/L_s)\dot i_d + (k_{i_d}\alpha_{0i_d}/L_s)i_d = = (k_{i_d}\alpha_{0i_d}/L_s)i_d^*. \tag{12}$$

It demonstrates that control process is asymptotically stable. According to the Hurwitz criterion, coefficients of the equation (12) are positive $(k_{i_d}\alpha_{0i_d}/L_s) > 0$,

$$(R_s/L_s + k_{i_d}/L_s) > 0.$$

It is important that the stability of the control loop is maintained under unlimited increasing of the controller gain coefficient $k_{i_d} \to \infty$; and real (12) and designed (4) control processes are fully coincident. This is made obvious, if to divide all elements of th e equation (12) by the coefficient k_{i_d}/L_s under condition $k_{i_d} \to \infty$

$$\frac{L_s}{k_{i_d}}\ddot i_d + \left(\frac{R_s}{k_{i_d}} + 1\right)\dot i_d + \alpha_{0i_d} i_d = \alpha_{0i_d} i_d^*. \tag{13}$$

This specificity provides the dynamic decomposition of the system (1) and the robustness to parametric disturbances. During operation, the interrelated system is broke down into relatively independent local control loops, with their transients which run in accordance with the desired performance equation (4). Clearly, if the gain coefficient of the controller is technically limited, there is a dynamic error which is set through technical requirements to the quality of control.

The transfer function of the open loop of the current i_d, which was obtained based on (12),

$$W_r(p) = \frac{i_d(p)}{i_d^*(p)} = \frac{k_{id}\alpha_{0id}/L_s}{p[p + (R_s/L_s + k_{id}/L_s)]} \tag{14}$$

shows that real system is of the type 1 astatic for control variable without a static error.

According to (14), if the controller gain is increasing, the actual speed merit factor of the current control system

$$D_\omega = \frac{k_{i_d}\alpha_{0i_d}/L_s}{R_s/L_s + k_{i_d}/L_s} = \frac{\alpha_{0i_d}}{\frac{R_s}{k_{i_d}}+1} \quad (15)$$

approximates to the predetermined value set in accordance with (4) $D_\omega^z = \alpha_{0i_d}$. This ensures the maximal permissible dynamic error of the current control.

During the development of the current control law, a small uncompensated time constant of the power frequency convertor T_μ, which is in the closed loop, was not taken into consideration. Assessment of its influence in form of the 1st order aperiodic unit can be carried out through the 3rd order differential equation of the closed loop system which is derived similarly to (12)

$$T_\mu \dddot{i}_d + (1+T_\mu R_s/L_s)\ddot{i}_d + (R_s/L_s + k_{i_d}/L_s)\dot{i}_d + +(k_{i_d}\alpha_{0i_d}/L_s)i_d = (k_{i_d}\alpha_{0i_d}/L_s)i_d^*. \quad (16)$$

According to the Hurwitz criterion, the current loop stability can be achieved under the following condition

$$(1+T_\mu R_s/L_s)(R_s/L_s + k_{i_d}/L_s) > T_\mu k_{i_d}\alpha_{0i_d}/L_s. \quad (17)$$

Assuming that $k_{i_d} \to \infty$, the stability condition can be finally presented as follows

$$\alpha_{0i_d} < 1/T_\mu + R_s/L_s \quad (18)$$

the system, which is set by the coefficient a_{0d}. Similarly, following the 2nd equation of the system (2)

$$L_s\frac{di_q}{dt} + R_s i_q = u_q + F_2 \quad (19)$$

and the desired equation (4) the control law for the current i_q, which is proportional to a motor's torque, can be obtained

$$u_q(t) = k_{i_q}(z - i_q); \qquad z = \alpha_{0i_q}\int_0^t (i_q^* - i_q)dt. \quad (20)$$

As seen from (20), the control law for stator current i_q also does not have parameters of the object (1) and its performance is similar to one for current i_d (11).

The block diagram of the current controller i_q, which is developed based on the equation (20), is presented in Fig. 2.

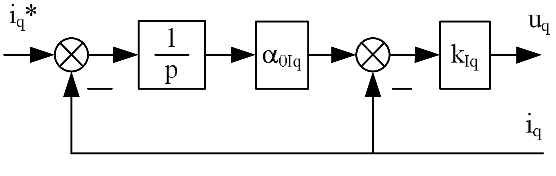

Figure 2 – Block-diagram of the current controller i_q

Following the 3rd equation of the system (1)

$$L_s\frac{di_f}{dt} + R_f i_f = u_f + F_3 \quad (21)$$

the control law of the excitation current i_f can be derived

$$u_f(t) = k_{i_f}(z - i_f); \qquad z = \alpha_{0i_f}\int_0^t (i_f^* - i_f)dt. \quad (22)$$

Block diagram of the excitation current controller i_f is shown in Fig. 3.

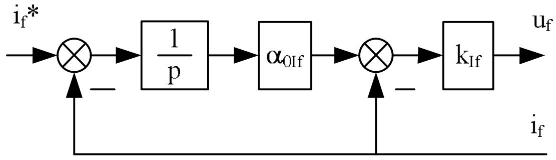

Figure 3 – Block-diagram of excitation current controller i_f

A speed control loop consists of the optimized internal loop of the current i_q and the local controlled object which is described by the 4th equation of the system (1)

$$J\frac{d\omega_r}{dt} = M - M_c. \quad (23)$$

Developing the speed control law and respective controller, inertia of the optimized loop of the current is not to be taken into consideration. The object, described by (23), is the integrating unit. Therefore, in order to achieve the type 1 astatism, the desired equation of the closed speed loop is of the 1st order as well

$$\dot{z} + \alpha_{0\omega}z = \alpha_{0\omega}\omega^*. \quad (24)$$

To reduce the influence of the dynamics of the control current loop i_q on speed control performance, a coefficient of the equation (24) should be identified under condition $\alpha_{0i_q} > (3-5)\alpha_{0\omega}$.

It is necessary to find a control function of a speed controller in a way that actual speed control performance would approximate to desired one set by the equation (24). The extent of such approximation can be estimated by means of the functional

$$G(i_{1q}^*) = \frac{1}{2}\left[\dot{z}(t) - \dot{\omega}(t)\right]^2. \quad (25)$$

To minimize this functional, similarly to the current loop, the gradient law of the 1st order can be used

$$\frac{di_q^*(t)}{dt} = -\lambda_\omega\frac{dG(i_q^*)}{di_q^*}, \quad (26)$$

where $\lambda_\omega > 0$ is a constant.

Substituting (25) to (26) the speed control law can be obtained

$$i_q^*(t) = k_\omega (z - \omega) \; ; \quad z = \alpha_{0\omega} \int (\omega^* - \omega) dt, \tag{27}$$

where k_ω is the gain coefficient of a speed controller.

Based on the equation (27), the speed controller of the 1st order can be designed. Its black diagram is presented in Fig. 4.

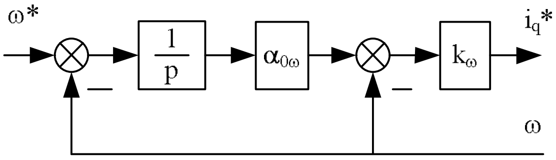

Figure 4 – Block-diagram of the 1st order speed controller

The speed controller, similarly to current controllers, contains only parameters of the desired control law and does not have parameters of the controlled object (1). Increasing the gain coefficient of the speed controller will result in approximation of dynamic processes in the loop to the desired ones, set by the equation (24). The system, according to the Hurwitz criterion, is stable even if the gain coefficient of the speed controller is unlimitedly increased $k_\omega \to \infty$.

The developed speed control law (27) provides a type 1 astatic system for control variable. The technological conditions may require ensuring the type 2 astatism. In this case, the control law has to be designed based on the desired equation (24) with an order of one unit higher than an order of the equation of the local object (23)

$$\ddot{z} + \alpha_{1\omega} \dot{z} + \alpha_{0\omega} z = \alpha_{1\omega} \dot{\omega}^* + \alpha_{0\omega} \omega^*. \tag{28}$$

Variation of two coefficients $\alpha_{0\omega}$, $\alpha_{1\omega}$ of the equation (28) can be used, based on well-known methods, to set desired performance indicators of speed control, such as particularly a response time and a value of overcontrol.

Based on the above presented methodology, the optimized speed control law will be obtained as follows

$$\begin{aligned} & i_q^*(t) = k_\omega \left[z - \omega \right]; \\ & z = \int f_0 dt; \\ & f_0 = \alpha_{0\omega} \int (\omega^* - \omega) dt + \alpha_{1\omega} (\omega^* - \omega). \end{aligned} \tag{29}$$

Block-diagram of 2nd order speed controller, designed based on the equation (29), is presented in Fig. 5.

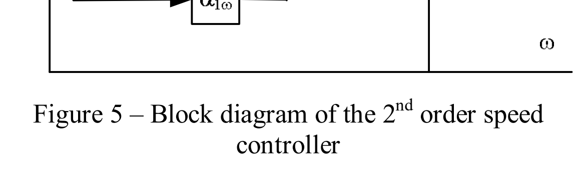

Figure 5 – Block diagram of the 2nd order speed controller

This controller also has only parameters of the desired differential equation of the speed closed loop (28).

The developed vector control system was investigated through modelling with the following parameters of the SIM IE: rated power $P_н$=208 кW; rated motor torque $M_н$=663 Nm; rated current $I_н$=458 A; rated speed $n_н$=3000 rpm. Controllers had parameters as follows: current controller i_d: α_{0id} =500, k_{id} =250; current controller i_q: α_{0iq} =500, k_{iq} =260; current controller i_f: α_{0if} =50, k_{if} =250; speed controller: $\alpha_{0\omega}$ =150, k_ω =50.

Fig. 6 presents transients of referenced speed ω* during the electrical drive start period. Transients of motor torque during start and applying the constant load torque (at time t=5 s), equal to rated motor torque, are shown in Fig. 7. In the steady state the motor torque is equal to load torque M=663 Nm.

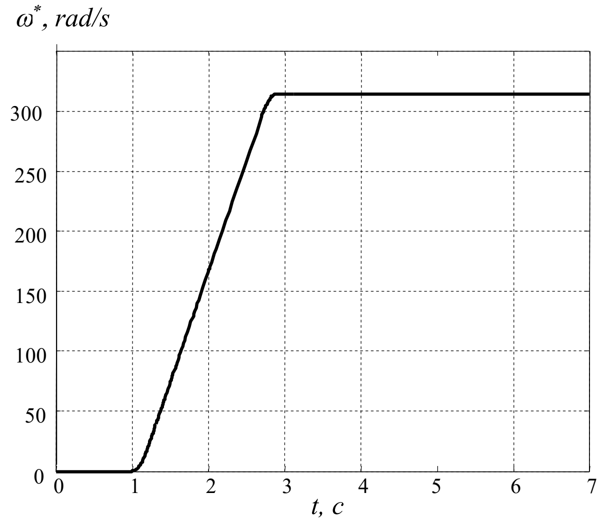

Figure 6 – Transients of referenced speed

Fig. 8 presents speed tracking error under variation of the stator resistance R_s: R_s = 0.0029 Ω (rated value), R_s = 0.00145 Ω (0.5 rated value) and R_s = 0.0058 Ω (2.0 rated value). As seen, this parametric disturbance does not affect the dynamic performance of the proposed system: three transients are identical, no recognizable differences. The maximal dynamic speed error during start is not over 4 rad/s, and during the applying the load torque – 3.3 rad/s.

Fig. 9 presents transients of speed error under variation of the inertia moment J of electrical drive: J= 3.6 kgm^2 (rated value); J= 7.2 kgm^2 (2 rated value) and J= 10.8 kgm^2 (3 rated value). It is seen that variation of the inertia moment does not influence the system dynamic performance. The maximal speed error is 4.8 rad/s during the start, and 2.5 rad/s under the load torque applied.

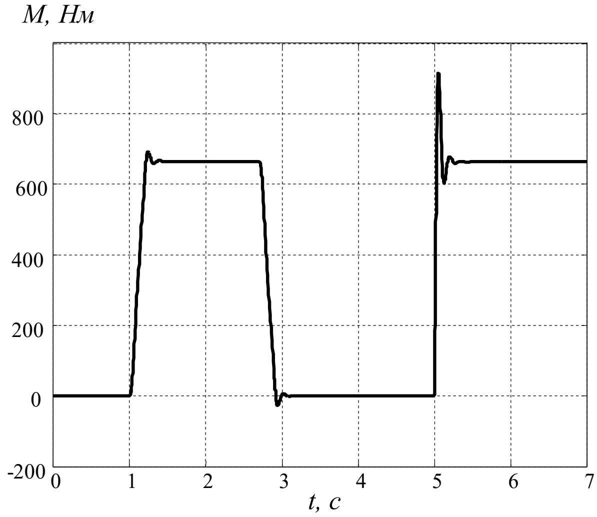

Figure 7 – Transients of motor torque

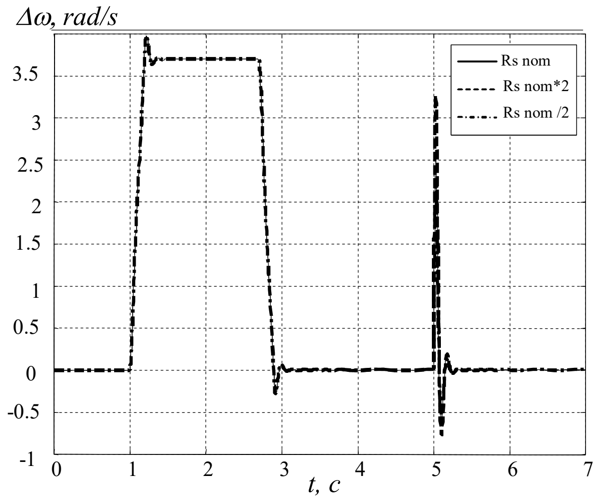

Figure 8 – Transients of speed error under stator winding resistance variation

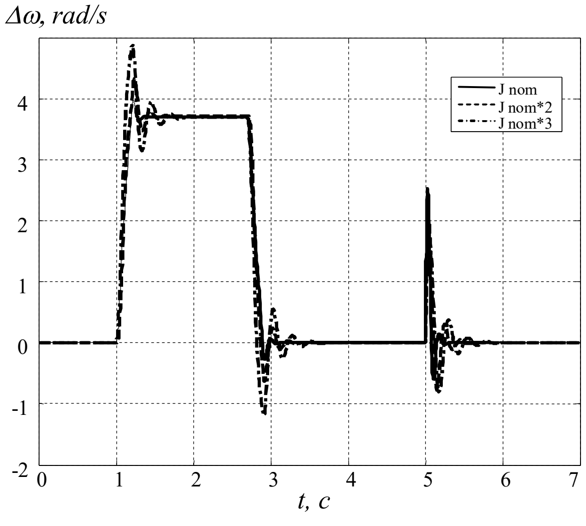

Figure 9 – Transients of speed error under inertia moment variation

The results of study presented above, clearly demonstrate that the electrical drive with SIM IE, designed based on the proposed methodology, has good control performance, is simple for development, and allows required operation under the parametric disturbances

CONCLUSIONS. Proposed electrical drive based on the switched inductor motor with independent electromagnetic excitation (SIM IE) can be designed based on the relatively simple methodology, applying a concept of reverse task of dynamics in combination with minimization of local functionals of instantaneous values of energies. This approach allows practical development of the controllers of the electro-mechanical system which would ensure a given quality of control and adequately simple practical realization under conditions of variation of the parameters of the controlled object and the uncertainties in a mathematical model. As a result, this type of electrical drive can be recommended for further development and promotion, to be used in technological processes and installations of various industries.

REFERENCES

1. Song, S., Ge, L., Ma, S. et al. (2015), “Accurate measurement and detailed evaluation of static electromagnetic characteristics of switched reluctance machines”, *IEEE Transactions on Instrumentation and Measurement*, Vol. 64, no. 3, pp. 704–714.

2. Xu, Y., Chen, H. and Gu, J. (2015), “Power loss analysis for switched reluctance motor converter by using electrothermal model”, *IET Power Electronics*, Vol. 8, no. 1, pp. 130–141.

3. Kozachenko, V., Rusakov, A., Sorokin, A. and Kochanov, A. (2011), "Switched reluctance drive – a promising direction of development of modern controlled electric drive", *Novosti teplosnabzheniya,* Vol. 11, pp. 24–26. (in Russian)

4. Rafiq, M., Rehman, S.U., Butt, Q.R. and Bhatti, A.I. (2009), “Power efficient sliding mode control of SR motor for speed control applications”, *in Proceedings of the 13th International Multitopic Conference (INMIC '09)*, pp. 1–6.

5. Pupadubsin, R., Chayopitak, N., Nulek, N. et al. (2010), “An improved adaptive sliding mode position control of a linear variable reluctance motor”, *in Proceedings of the International Conference on Electrical Machines and Systems (ICEMS '10),* pp. 1583–1588.

6. Baik, W.-S., Kim, M.-H., Kim, D.-H. et al. (2006), “A high performance motion control system of switched reluctance motor”, *in Proceedings of the 37th IEEE Power Electronics Specialists Conference (PESC '06)*, pp. 1–5.

7. Potapenko, E.M. and Kazurova, A.E. (2007), "Precision control uncertain objects. Comparison of methods of control", *Sb. nauchn. trud. Dneprodzerzhinskogo gos. tehn. univer-ta*, pp. 353–356. (in Russian)

8. Krutko, P.D. (2005), "Robustly stable structures of control systems of high dynamic precision. Algorithms and dynamics of control of model objects", *Izves-*

tiya RAN. TiSU, Vol. 2, pp. 120–140. (in Russian)

9. Ostroverkhov, N. and Buryk, N. (2011), "Control of coordinates electric drives based on the concept of inverse dynamics problems for minimization local functionals momentary values of energy", *Elektrotehnika i elektroenergetika,* Vol. 1, pp. 41–49. (in Ukrainian)

10. Ostroverkhov, N.J. (2008), "Method for the synthesis of regulators of electromechanical systems based on of the concept of inverse problems of dynamics in combination with the minimization of local functionals of the instantaneous motion energy values", *Vestnik NTUU "KPI",* Vol. 30, pp. 105–100. (in Ukrainian)

11. Chernousko, F.L., Ananyevskiy, I.M. and Reshmin, S.A. (2006), *Metody upravleniia nelineynymi mekhanicheskimi sistemami* [Methods of control nonlinear mechanical systems], Fizmatlit, Moscow. (in Russian)

СИСТЕМА ВЕКТОРНОГО УПРАВЛЕНИЯ СКОРОСТЬЮ ВЕНТИЛЬНО-ИНДУКТОРНОГО ЭЛЕКТРОПРИВОДА

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Электрический привод потребляет около 65 % общего объема вырабатываемой в мире электроэнергии. Асинхронный электропривод является наиболее распространенным в разных отраслях промышленности, однако имеет рад известных недостатков, в частности, невысокий коэффициент мощности. Электрический привод на основе вентильно-индукторного двигателя с независимым электромагнитным возбуждением может быть альтернативой для асинхронного привода благодаря ряду преимуществ, связанных с энергетической эффективностью. На электропривод действуют координатные и параметрические возмущения, которые могут повлиять на эффективность преобразования энергии и качество управления в различных технологических процессах. В исследовании предложен метод разработки законов управления на основе идеи обратимости прямого метода Ляпунова для анализа устойчивости и минимизации мгновенного значения энергии в качестве заданной функции Ляпунова. Это обеспечивает эффективную работу с меньшей чувствительностью к изменению параметров двигателя, а также простоту реализации системы управления. Моделирование предложенного электрического привода подтвердило высокое качество управления и технологичность системы. Результаты данного исследования будут способствовать дальнейшему внедрению вентильно-индукторного электропривода в промышленных технологиях.

Ключевые слова: электропривод, вентильно-индукторный двигатель, законы управления.

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