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Review Article

TECHNIQUE OF SETTING UP A PIPELINE VALVE ELECTRIC ACTUATORS CONTROL SYSTEM USING THE EQUIVALENT CIRCUIT PARAMETERS, ESTIMATED BY FALLING CURRENT CURVE

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Abstract

To support modern requirements for an induction electric actuators of pipeline valves, the most promising is the organization of a filed oriented control system, the setting of which is extremely difficult without the parameters of the equivalent circuit of an induction motor. This article provides a technique of setting a field-oriented control system for induction electric drive of pipeline valves based on equivalent circuit parameters, estimated by falling phase current curve. **The main aim** of the research is to test the functioning of an induction filed oriented electric drive using a load test bench, the setting of which is made on the basis of parameters previously estimated by falling phase current curve. **Methods.** To achieve the goal of the research, theoretical and experimental research methods were used. Theoretical research methods include the theory of electric drive, the theory of automatic control systems, the theory of electrical machines. Experimental research were carried out using a load test bench that provides the required level of load on the shaft of the tested induction motor. **Results.** Suggested the technique of setting a field-oriented control system for induction electric drive of pipeline valves based on equivalent circuit parameters, estimated by falling phase current curve. Relative values of deviations of current, speed and torque are obtained at the nominal level of load on the shaft of an induction motor. The applicability of the proposed technique using a loading test bench was confirmed.

Keywords: pipeline valves; electric actuators; induction motor; field-oriented control; off-line estimation; subordinate regulation; equivalent circuit; falling current curve.

1. Introduction

In the economy of the Russian Federation, the fuel and energy complex occupies a significant place and plays the role of basic infrastructure, the basis for generating revenues for the budget system of the Russian Federation and the largest customer for other industries [1]. The most effective way to transport hydrocarbons a long land routes in the Russian Federation is pipeline transport [2]. The indications for the appointment of electric actuators of pipeline valves with an electronic control unit in accordance with GOST R 55511-2013 to ensure general requirements include [3], [4]:

- nominal and maximum torque (or force) on the output shaft;
- maximum torque (or force) developed by the electric drive in case of failure of the disconnecting devices;
- electric current parameters (AC or DC, frequency, voltage, number of phases, current strength, current parameters of discrete control signals and analog output for information about the position of the output link, etc.);
- output shaft speed limit or output shaft rated travel;

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- the time of making the limit number of revolutions of the output shaft or the rated travel;
- output shaft speed (this indicator can be indicated instead of the indicator «time to complete the maximum number of revolutions»);
- motor power (W or VA);
- duty cycle;
- climatic version;
- resistance to external influences (seismic loads, vibration, fire resistance, moisture, dust, harmful substances in the environment);
- explosion protection.

To ensure the specified quality indicators [5], as well as the above-mentioned special requirements for the electric drive of pipeline valves [6]–[8], the most important of which is the limitation of the maximum torque (or force) on the output shaft, the most acceptable is the use of a field-oriented (vector) control system, implemented on the basis of mathematical and software-algorithmic support of the electronic control unit.

2. Field-oriented (vector) control system for pipeline valves electric drive

The first vector control system based on an induction motor (IM) was designed by Felix Blaschke at Siemens in 1971 [9]–[11]. The regulation of the stator current in it was organized in a rotating coordinate system oriented along the measured rotor flux ψ_r with the help of sensors mounted in the motor on the Hall elements. In the future, the designers abandoned the flux sensor and began to estimate the position and size ψ_r by indirect methods according to the machine model.

A simpler approach is to form a vector control system with indirect orientation along the rotor field [12], [13], which was implemented in the software of the ESD-VCX control unit for the electric drive of pipeline valves (EleSy company, Tomsk) to ensure high accuracy of speed control, as well as limiting the moment of the working body (fig. 1). This system does not contain a flux stabilization circuit ψ_r. A similar approach is applicable when there is no need to implement flux linkage regulation ψ_r, for example, to organize a second control zone, which is not required for the electric drive of pipeline valves. The advantage of this approach when building a control system is a good balance between accuracy and ease of regulation, which is determined by the absence of the need to build a state observer ψ_r.

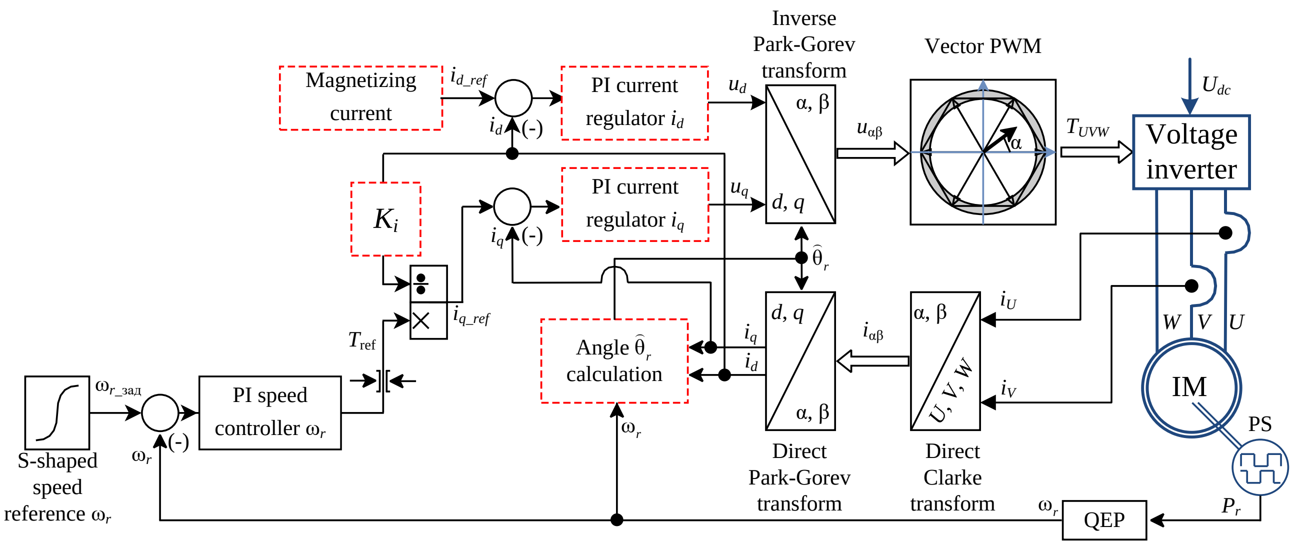

Fig. 1. Components in the structure of a vector control system with indirect orientation along the rotor field, the setting of which requires estimates of the parameters of the IM equivalent circuit

The electric drive control system is made in a rotating coordinate system according to the principles of subordinate regulation [14], [15]. Proportional-integral (PI) controllers of control loops ensure the maintenance of the required state variables, as well as the quality of transients [5]. So the speed control loop, with the control loop subordinate to it, the component of the stator current vector i_q, responsible for the formation and limitation of the motor torque T_{ref} and motor shaft speed control ω_r. The formation of a smooth setting of the speed loop is provided by an S-shaped speed reference, and the feedback signal ω_r is calculated on the basis of position change signals P_r coming from the position sensor (encoder). The control loop of the component of the stator current vector i_d, the statutory for which is the magnetizing current, is responsible for indirectly maintaining the given constant level of rotor flux linkage $\hat{\psi}_r$. The angle $\hat{\theta}_r$ is used to orient the rotating coordinate system in the d, q axes relative to the rotor flux linkage, and its calculation requires instantaneous values of the rotor speed ω_r, as well as currents i_d, i_q. The control of stator currents in the motor phases is provided by at least two measurement channels. The transition from a real, three-phase unrotating coordinate system in the U, V, W axes to an orthogonal fixed two-phase coordinate system in the α, β axes and vice versa is performed using the direct and inverse Clarke transformations. The transition from an orthogonal unrotating two-phase coordinate system in the α, β axes to a rotating orthogonal coordinate system in the d, q axes and vice versa is performed using the direct and inverse Park-Gorev transformation. To effectively use the DC link voltage U_{dc}, a vector pulse-width modulation (PWM) block is used, which generates T_{UVW} signals that go directly to the power switches of an autonomous voltage inverter.

On the structure of the vector system (fig. 1), components are highlighted, the correct setting of which requires estimates of the parameters of the IM equivalent circuit. These blocks include current controllers, while the adjustable coefficients of the speed controller in this case do not depend on the electromagnetic parameters of the IM, but depend on the parameters of the mechanical subsystem of the electric drive. To ensure the nominal level of the stator current IM at the nominal level of the load on the shaft, it is required to ensure the correct ratio between the active and reactive power entering the machine [7], [16]. This is determined by the value of the magnetizing current, as well as the time constant of the rotor used to orient the coordinate system (angle calculation block $\hat{\theta}_r$), the calculation of which also requires finding estimates of the parameters of the equivalent circuit. To organize the limitation of the torque at the output of the speed controller, proportional to the current level of the magnetizing current, it is required to apply the coefficient K_i, the calculation of which is also carried out according to the estimates of the parameters of the IM equivalent circuit.

3. IM equivalent circuit parameters estimating algorithm

To find estimates of the parameters of the IM equivalent circuit, an algorithm for preliminary identification by the falling curve of phase currents was applied [17], [18]. This procedure for preliminary identification of IM parameters according to the phase falling current curve by means of a digital device is carried out in two stages. At the first stage of pumping by means of a transistor switch, the IM windings are connected to the power source, which causes the flow of a direct current of a given value. At the first stage of pumping by means of a transistor switch, the IM windings are connected to the power source, which causes the flow of a direct current of a given value. The pumping stage continues until the complete and guaranteed completion of the phase current transient process, the time of which is determined by the properties of IM, after which the steady value of the current $i(0+)$ is stored by a digital device and used in further calculations as a non-zero initial condition. At the end of the first stage of pumping, a transition is made to the second stage of identification – the formation of a falling current curve, which is ensured by disconnecting the windings of the tested IM from the power source and shorting them together. During current falling instantaneous values are measured and stored in the same winding in which the steady-state current value was previously measured. Based on the mathematical expression describing the current falling process, as well as the experimental falling current curve, an objective function is formed, the extremum of which corresponds to the desired estimates of the parameters of the IM equivalent circuit. To minimize the values of the objective function, the Levenberg-Marquardt method, which has a high convergence rate, was applied. The main assumptions made when compiling a customizable regression mathematical model of IM with a short-circuit rotor, which is subsequently used to build a preliminary identification algorithm, include the linearity of the magnetic system, the absence of losses in the magnetic circuit, as well as the equality of the leakage inductance of the stator winding and the leakage inductance of the rotor winding reduced to the stator $L_{1\sigma} = L'_{2\sigma} = L_\sigma$ [19]. As a result, the parameters of the IM equivalent circuit, identified as a result of the calculation of the preliminary identification procedure, are the leakage inductance $\hat{L}_\sigma$, the inductance of the main magnetization circuit $\hat{L}_m$, as well as the active resistance of the rotor winding reduced to the stator circuit $\hat{R}_2'$. The active resistance of the stator winding R_1 is assumed to be a priori known and predetermined.

4. Technique of setting a vector control system based on estimated parameters

Since the vector control system (fig. 1) is a subordinate regulation control system, the most critical for the operation of the electic drive as a whole is the correct setting of the fastest current loop, subordinate to the slower speed loop. The optimization of the current loop [14], [15], taking into account the inertia of the feedback with the PI controller, is close to tuning to the modular optimum of the 2nd order system [20]–[22]. The contour is an astatic system of the 1st order in terms of control. According to the contour optimization criterion, the coefficients of two current controllers i_d, i_q are determined as follows. The block diagram of an optimized current loop with inertial feedback and full compensation of internal negative feedback on the electromotive force (EMF) IM is shown in fig. 2. The contours of the currents i_d, i_q are identical.

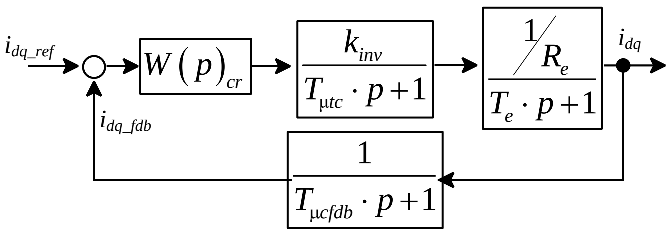

Fig. 2 Structural diagram of the generalized current loop i_{dq} with a PI controller

Transfer function of the current PI controller (1)

$$W(p)_{cr} = k_{cr}\cdot\frac{T_{cr}\cdot p+1}{T_{cr}\cdot p}. \tag{1}$$

Current regulator gain (2)

$$k_{cr} = \frac{\hat{T}_e\cdot\hat{R}_e}{a_c\cdot k_{inv}\cdot\left(T_{\mu tc}+T_{\mu cfdb}\right)}, \tag{2}$$

where $\hat{T}_e$ – estimate of the electromagnetic time constant IM, s; $\hat{R}_e = R_1 + \frac{\hat{R}_2'\cdot\hat{L}_m^{\,2}}{L_2^{\,2}}$ – estimate of the equivalent active resistance of stator circuits IM, Ohm; $a_c = 2$ – current loop optimization factor, o.u.; $k_{inv} = 311$ – inverter gain when the inverter is powered by 380 V and using vector PWM modulation, V; $T_{\mu tc}$ – small time constant in the forward channel, s; $T_{\mu cfdb}$ – small time constant in the feedback circuit, s.

Integral component of the current regulator $T_{cr} = \hat{T}_e$.

The estimate of the electromagnetic time constant $\hat{T}_e$, characterizing the dynamics of current changes in the IM stator windings during the transient process is defined as (3)

$$\hat{T}_e = \frac{\hat{\sigma}\cdot\hat{L}_1}{\hat{R}_e}, \tag{3}$$

where $\hat{\sigma} = 1 - \frac{\hat{L}_m^{\,2}}{\hat{L}_1\cdot\hat{L}_2}$ – leakage coefficient estimate, o.u.

Taking into account the assumptions made when compiling the mathematical model:

- $\hat{L}_1 = \hat{L}_{1\sigma} + \hat{L}_m = \hat{L}_\sigma + \hat{L}_m$ – estimation of the equivalent inductance of the stator winding, H;
- $\hat{L}_2 = \hat{L}_{2\sigma}' + \hat{L}_m = \hat{L}_\sigma + \hat{L}_m$ – estimation of the equivalent inductance of the rotor winding, H.

The constant $T_{\mu tc}$ characterizes the time during which, with the help of switching the autonomous voltage inverter keys, a control action is formed on the windings in the form of an average voltage value over the PWM period. The constant $T_{\mu cfdb}$ characterizes the time required to measure the average current value over the PWM period. Thus $T_{\mu tc} = T_{\mu cfdb} = 1/F_{pwm}$, where $F_{pwm} = 10\,\text{kHz}$ is the PWM modulation frequency generated by the hardware timer of the ESD-VCX control unit. A block diagram of the speed loop with inertial feedback, PI controller, as well as the possibility of limiting the electromagnetic torque of the motor under the condition of a constant magnetizing current i_d and full compensation of internal negative EMF feedback, is shown in fig. 3.

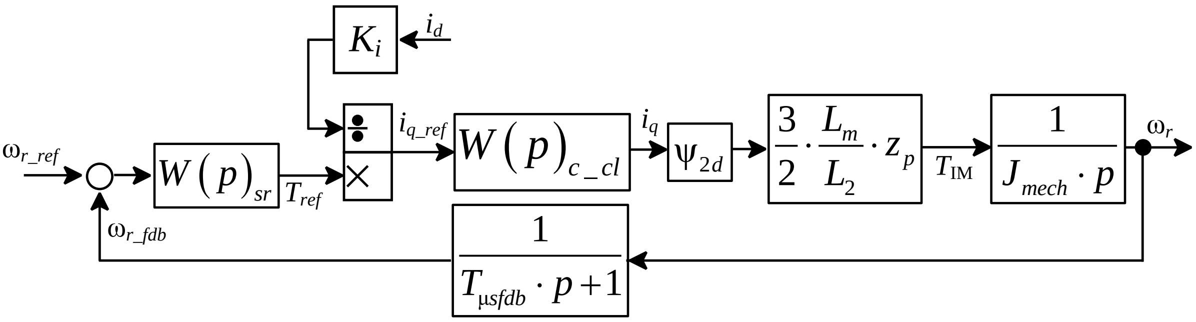

Fig. 3. Structural diagram of a speed loop with a PI controller

In speed loop optimization, the internal optimized current loop i_q is represented by a truncated 1st order transfer function (4)

$$W(p)_{c\_cl} \approx \frac{1}{T_c \cdot p+1}, \tag{4}$$

where $T_c = a_c \cdot \left(T_{\mu tc} + T_{\mu sfdb}\right)$ – equivalent time constant of the optimized current loop, s.

Transfer function of the speed PI controller (5)

$$W(p)_{sr} = k_{sr} \cdot \frac{T_{sr} \cdot p+1}{T_{sr} \cdot p}. \tag{5}$$

The gain (6) of the speed loop, at the output of which the task of the required electromagnetic torque is formed T_{ref}

$$k_{sr} = \frac{J_{mech}}{a_c \cdot T_{\mu se}}, \tag{6}$$

where J_{mech} – moment of inertia of a mechanism driven by an electric drive, kg·m^2; a_c – optimization coefficients from the speed loop; $T_{\mu se} = T_{\mu sc} + T_{\mu sfdb}$ – equivalent short time constant of the optimized speed loop, s; $T_{\mu sc} = T_c$ – small time constant in the forward channel of the speed loop, s; $T_{\mu sfdb}$ – small time constant in the feedback loop of the speed loop, determined by the frequency of calculation of the speed signal ω_{r_fdb} and the frequency of calculation of the current regulator in the microcontroller, s.

The integral component of the current regulator (7) is defined as,

$$T_{sr} = a_s \cdot b_s \cdot \left(T_{\mu sc} + T_{\mu sfdb}\right), \tag{7}$$

where b_s – speed loop optimization factor.

Thus, setting the parameters of the speed regulator according to the structural diagram of the speed loop (fig. 3), which provides the possibility of limiting the electromagnetic torque of the IM, does not require the values of the estimates of the parameters of the IM equivalent circuit and depends only on the parameters of the mechanical subsystem of the electric drive.

The presented vector control system with indirect orientation along the rotor field (fig. 1), in which there is no rotor flux linkage control loop, does not have a link that calculates the components of the rotor flux vector ψ_{2d}, ψ_{2q}. Thus, under the condition of a constant magnetizing current i_d, which is responsible for the formation of flux linkage in the rotor, to organize the torque limitation at the output of the speed regulator, a torque coefficient (8) is introduced, the value of which depends on the parameters of the equivalent circuit estimates according to the expression,

$$\hat{K}_i = \frac{3}{2} \cdot \frac{\hat{L}_m^2}{\hat{L}_2} \cdot z_p, \tag{8}$$

where z_p – number of pole pairs IM.

In an induction electric drive, the angular velocity of the rotor, by definition, is not equal to the angular velocity of the rotor flux vector. This means that the required position of the rotor magnetic flux vector cannot be detected directly by the position sensor mounted on the IM shaft (fig. 1). In an induction electric drive, the angular speed of the rotor, by definition, is not equal to the angular speed of the rotor flux vector. This means that the required position of the rotor magnetic flux vector cannot be detected directly by the position sensor mounted on the IM shaft (fig. 1). An iterative calculation of the relative estimate of the rotor field angle $\hat{\theta}_r$ for the orientation of the rotating coordinate system in the axes d, q is based on estimates of the parameters of the IM equivalent circuit according to the system of equations (9) [12], [13], [23].

$$\begin{cases} \hat{i}_{mR_{k+1}} = \hat{i}_{mR_k} + \dfrac{T_{samp}}{T_r} \cdot \left(i_{d_k} - \hat{i}_{mR_k} \right) \\ \hat{f}_{S_{k+1}} = \omega_{e_{k+1}} + \dfrac{1}{\hat{T}_r \cdot \omega_b} \cdot \dfrac{i_{q_k}}{\hat{i}_{mR_{k+1}}} \\ \hat{\theta}_{r_{k+1}} = \hat{\theta}_{r_k} + K \cdot \hat{f}_{S_{k+1}} \end{cases}, \tag{9}$$

where $\hat{i}_{mR_k}$ – estimation of the rotor magnetization current at the corresponding calculation step, A; $T_{samp} = 1/F_{pwm}$ – sampling period, s; $\hat{T}_r = \dfrac{\hat{L}_2}{\hat{R}_2'}$ – rotor time constant estimation, s; i_{d_k}, i_{q_k} – currents in the rotating coordinate system *d*, *q*, oriented relative to the rotor field IM, obtained at the corresponding calculation step using the measuring instruments of the frequency converter, A; $\hat{f}_{S_{k+1}}$ – angular speed of rotation of the rotor field, rad/s; $\omega_e = \dfrac{\omega_r}{z_p}$ – electrical angular speed of the rotor at the corresponding calculation step, calculated on the basis of pulses from the position sensor, taking into account the number of IM pole pairs z_p, rad/s; $\omega_b = 2 \cdot \pi \cdot f_b$ – nominal angular speed of rotation of the field in the IM magnetic gap, rad/s; $f_b = 50$ – nominal electric frequency of the field in the magnetic gap IM, Hz; $K = T_{samp} \cdot f_b$ – nominal electric frequency of the field in the magnetic gap IM; *k* – current model sample time.

To calculate the magnetization current $\hat{i}_{flux}$ (10), it is required to determine the EMF of the magnetization branch $\hat{E}_{mr}$. induced by the air gap flux in the stator winding in the nominal operating mode according to the expression [[24]

$$\hat{E}_{mr} = \sqrt{\begin{matrix} \left(U_{1r} \cdot \cos\varphi_{1r} - R_1 \cdot I_{1r} \right)^2 + \\ + \left(U_{1r} \cdot \sqrt{1 - \cos^2 \varphi_{1r}} - \hat{X}_{1\sigma} \cdot I_{1r} \right)^2 \end{matrix}}, \tag{10}$$

where U_{1r} – rated phase voltage, V; I_{1r} – rated stator current, A; $\cos\varphi_{1r}$ – rated cosine; $\hat{X}_{1\sigma} = 2 \cdot \pi \cdot f_b \cdot \hat{L}_\sigma$ – estimation of leakage reactance, Ohm.

According to the estimates of the parameters of the IM equivalent circuit, the magnetization current (11) is defined as

$$\hat{i}_{flux} = \frac{\hat{E}_{mr} \cdot \omega_b}{\hat{L}_m}, \text{ A.} \tag{11}$$

5. Description of the test bench

One of the objective criteria to evaluate the applicability in practice and the correctness of the parameters of the IM equivalent circuit obtained as a result of the preliminary identification procedure is the analysis of the behavior of the vector control system configured on their basis in various modes, under various external influences, including variable torque loads. The research of the vector control system, tuned on the basis of the estimated equivalent circuit parameters and organized on the basis of the ESD-VCX electric drive software (manufactured by EleSy company, Tomsk) was carried out using a load test bench (fig. 4), ensuring the formation of the required level of load moment on the test shaft IM. IM with a squirrel-cage rotor of the ELAS series were used as test machines used as part of an electric drive for pipeline valves.

The formation of the load moment on the shaft of the tested IM as part of the ESD-VCX electric drive is provided by a loading machine, which is a DC machine of independent excitation (fig. 4, 5). The frequency converter FC1 provides stabilization of the current of the field coil (FC) of the DC machine, forming the required level of flux linkage corresponding to the synchronous speed loaded by the IM. The two-loop control system, organized in the software of the frequency converter FC2, closed according to the signals of the armature circuit (AC) current and the speed ω_r, calculated on the basis of quadrature encoder pulses (QEP) from the test bench position sensor (PS0), ensures stable maintenance of the specified level of torque T_{DC} on the DC machine shaft, opposite to the direction movement of the common shaft and torque of T_{IM} on the IM shaft, regardless of speed. The test bench is provided with a torque sensor (TS), which is polled by FC2 via the RS-485 interface, providing monitoring of the torque determined by the difference in the moments of the DC machine and IM, as well as emergency removal of voltage from the DC machine windings in case of emergency situations. Since the DC machine, when forming the torque T_{dc}, operates in the regenerative braking mode, FC2 organizes a decrease in the excess voltage of the DC link using a braking resistor BR and converts it into heat. FC1 and FC2 are powered from a 220V mains, and the voltage of the DC machine windings is limited using PWM modulation. The frequency converter FC3, which is part of the ESD-VCX electric drive, ensures the operation of a vector control system, configured on the basis of the estimated equivalent circuit parameters, using a signal from the built-in position sensor PS1. The ESD-VCX electric drive was powered from the 380 V mains. The appearance of the load test bench, corresponding to the functional design (fig. 4), is shown in fig. 5.

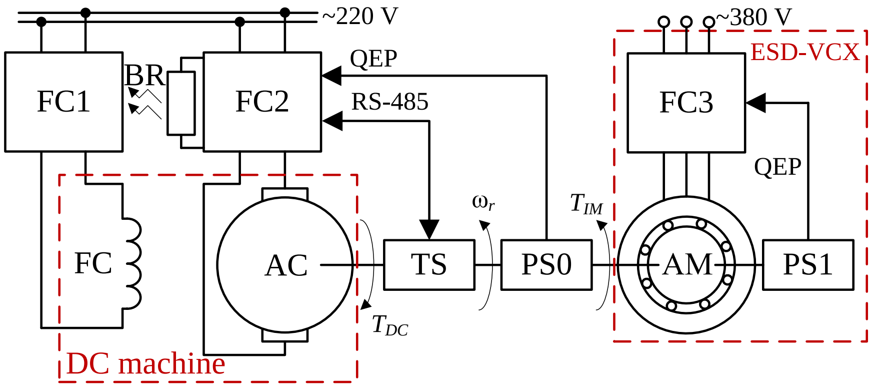

Fig. 4. Functional diagram of the load test bench for the electric drive of pipeline valves ESD-VCX

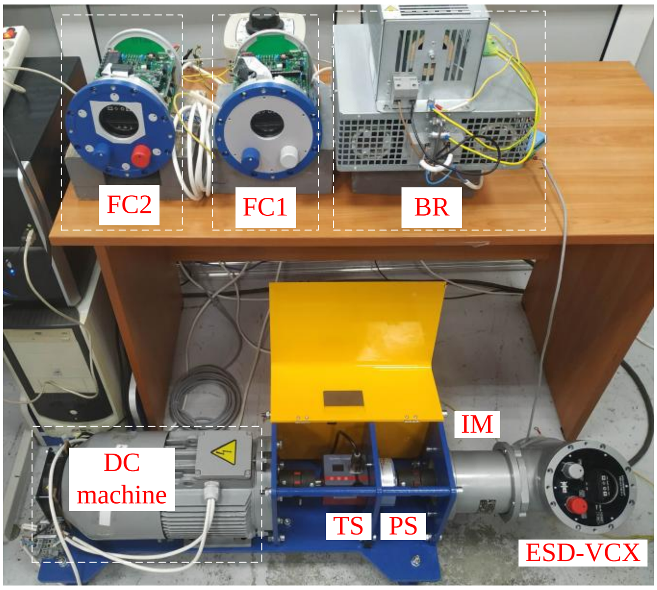

Fig. 5. Appearance of the test bench

6. Research of a tuned induction electric drive of pipeline valves using a test bench

The values of experimental and estimated equivalent circuit IM parameters of the ELAS series, obtained from the falling curves of phase currents [17], [18] are presented in table 1.

In addition to the coefficients of current and speed controllers, the key parameters of the vector control system that directly affect the characteristics and performance of the electric drive as a whole are the rotor time constant $\hat{T}_r$, the torque coefficient $\hat{K}_i$ and the magnetization current $\hat{i}_{flux}$, the values of which are determined based on the estimated parameters of the IM equivalent circuit. The settings of the vector control system calculated on the basis of estimates of the parameters of the IM equivalent circuit of the ELAS series are presented in table 2.

Table 1. The values of the experimental and estimated parameters of the equivalent circuit IM with a short-circuit rotor of the ELAS series

IM Model	R_1, Ohm	$\hat{R}_2'$, Ohm	$\hat{L}_m$, H	$\hat{L}_\sigma$, H
ELAS 120	72,95	36,76	1,419	0,17
ELAS 180	43,10	21,96	1,042	0,12
ELAS 370	21,35	11,04	0,638	0,06
ELAS 550	6,27	6,27	0,653	0,03

Table 2. Vector control system settings based on estimated equivalent circuit parameters, as well as discrepancies from nominal parameters

IM Model	$\hat{T}_r = \frac{\hat{L}_2}{\hat{R}_2'}$, s	$\hat{K}_i = \frac{3}{2}\cdot\frac{\hat{L}_m^2}{\hat{L}_2}$, o.u.	$\hat{i}_{flux}$, A	ΔT_{rated}, %	ΔI_{rated}, %	$\Delta\omega_{rated}$, %
ELAS 120	0,043	3,79	0,3	22,38	10	1,5
ELAS 180	0,052	2,8	0,46	24,6	7,5	0,3
ELAS 370	0,063	1,74	0,81	28,7	2,4	2,7
ELAS 550	0,11	0,92	1,05	12,28	8,5	1,6

As an example, the timing diagrams of a vector control electric drive configured on the basis of the identified parameters are given, providing speed stabilization during a load surge up to the nominal value (fig. 6).

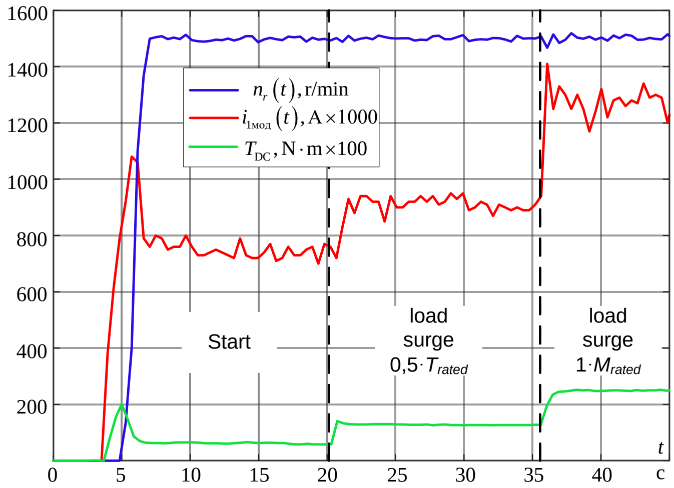

Fig. 6. Operation of a vector control system tuned using the estimated equivalent circuit parameters of the IM ELAS 370

The vector control system of the electric drive, tuned on the basis of the estimated IM parameters, is operational (fig. 6) and provides speed stabilization when the load torque on the IM shaft changes, which indicates the correct setting of the current and speed circuit controllers. The maximum value of the relative torque error (table 1) ΔT_{rated}, determined at the output of the speed controller of the vector control system relative to the signal from the torque sensor at the nominal load level on the IM shaft, does not exceed 29%. The achieved result is acceptable for classical systems of electric drive subordinate regulation with constant regulators coefficients. Nevertheless, the relative deviations of the stator current ΔI_{rated} and rotor speed $\Delta\omega_{\text{rated}}$ at the nominal level of the load on the IM shaft relative to the nominal values do not exceed 10 and 2.7 %, respectively, which indicates the correctness of the control system settings, reflecting the real processes occurring in IM. The procedure for preliminary identification by the phase current falling curves can be applied to setting the vector control system of pipeline valves electric drive.

7. Conclusion

An engineering method of setting vector control system for an induction electric drive of pipeline valves with an indirect orientation along the rotor field is proposed based on estimates of the equivalent circuit parameters obtained using the preliminary identification procedure from the phase falling current curves. Based on the estimated parameters of the equivalent circuit IM, namely the magnetizing main circuit inductance, leakage inductance, as well as the active resistance of the rotor winding reduced to the stator, the settings of the vector control system, namely the rotor time constant $\widehat{T}_r$, the torque coefficient $\widehat{K}_i$ and the magnetization current $\widehat{i}_{flux}$, are obtained, the values of which are key for ensuring the performance of the electric drive.

Tests of a vector control electric drive of pipeline valves, tuned using the proposed methodology, as part of a load test bench, made it possible to estimate the magnitude of the relative deviations of the torque ΔT_{rated} determined by the control system, the current in the stator windings ΔI_{rated}, as well as the speed dip $\Delta\omega_{\text{rated}}$ when working out the nominal load level on the shaft of the tested IM in a static mode of operation. The deviations of ΔI_{rated} and $\Delta\omega_{\text{rated}}$ do not exceed 10 %, but deviations ΔT_{rated} do not exceed 29 %, which allows us to recommend the proposed method for setting up the vector control electric drive of pipeline valves.

Further development of the proposed methodology can be aimed at finding the quality indicators of current and speed loops in dynamic operating modes, as well as at studying the effect of fluctuations in the parameters of the IM equivalent circuit, caused by temperature changes, on the static and dynamic characteristics of the vector control electric drive of pipeline valves.

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