Nonlinear Control of Synchronous Servo Drive
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Abstract—The complete control design for a permanent magnet synchronous (PMAC) motor, derived from the input-output linearization, is presented in the paper. The motor model, written in the rotor's d–q coordinates, is nonlinear with respect to the state variables and linear in the control. The input-output linearization makes it possible to write the motor's model in Brunovski decoupled canonical form which makes the synthesis of linear controllers possible. The proposed control structure allows perfect tracking of smooth references in the case of nominal parameters. In relation to the bounded parameter perturbations, the robustness of the feedback system is improved by introducing an additional compensation signal which assures the stability of the perturbed system in Lyapunov's sense. The influence of parameter variations that prevents exact compensation of the control plant nonlinearities was analyzed for the PMAC motor. A description of the laboratory setup is given, and the experimental results of the proposed PMAC servo drive control are presented.

I. INTRODUCTION

The control of electrical motors used in high-performance servo drives and robots demands control concepts that allow one to achieve high dynamics and the prescribed accuracy for all operating conditions. The class of reference input signals, the class of external disturbances, constraints of system states and inputs, parameter and structure perturbations, and the power source characteristic define the operating conditions of the servo drive. The feedback system is additionally expected to provide a linear dependence of the system's inputs and the controlled outputs within the defined domain.

At present, permanent magnet synchronous (PMAC) motors cover the largest part of servo applications in the power range from 1–10 kW which is predominantly due to the high power/weight ratio based on the use of high quality rare-earth magnetic materials. It is evident from the analysis of compared model structures and characteristic electrical and mechanical time constants of dc and ac motors that PMAC motors provide the best possible controllability and high dynamics of drives.

Despite the above-mentioned advantages of PMAC motors and their wide applicability, very few experimentally evaluated system-based control concepts are available. Generally, the control design is based on linear models and on the assumption that the mechanical and the electrical time constants differ at least for one order of magnitude which results in the well-known series control structures. Apart from the fact that its formal treatment is inconsistent, series control also does not allow a methodical design of the control structure and the determination of the optimal control parameters. The influence of nonlinearities, external disturbances (mechanical torque), and parameter and structural perturbations (inertia, friction) is usually considered only by the demanded “adequate robustness” of linear controllers. Consequently, the series control structure cannot assure a satisfactory dynamic behavior in the entire operating range.

Thus a strong tendency to use advanced, predominantly nonlinear control design methods for high performance ac servo drives has prevailed in the last decade. These numerous approaches can be divided into three basic groups, considering the mathematical tools used by different authors. The first approach is based on the extended linear control law with an adaptation mechanism which the authors use to consider the variations of relevant parameters (inertia, rotor resistance in induction motor (IM), etc.) and the influence of unmodeled dynamics. To prove the convergence of the adaptive law and the stability of feedback system, the motor models are linearized [3], [8].

The second group of authors solves the problem of ac drives on the basis of discontinuous control which assures sliding motion of the closed-loop system according to the prescribed manifold in the state space. Invariance to parameter and structure perturbations inside the bounded domain is proven for motion in the sliding mode [4]. Modifications of the initial approach in the direction of the continuous sliding mode eliminate the chattering of the controlled output signal and allow a relatively simple implementation [5].
The third approach, where our contribution belongs, is based on differential geometric methods [9]-[11]. This approach yields exact linearized models in some open neighborhood of the initial state, and in case of multi-input multi-output systems, the (local) decoupling of the input-output behavior can be achieved under certain conditions. The differential geometric methods make it possible to introduce the definitions of nonlinear system structure properties (reachability, observability, relative degree) which are similar to those established for linear systems. The main advantage of this approach for control design is the fact that a number of linear methods that allow the selection of the most appropriate control structure and a systematic determination of its parameters can be used. Unlike series control structures, the exact linearization implies a parallel structure, the measurement of all state variables, and a good knowledge of the model (structure and parameters). In the field of servo drives, the mentioned formalism was used by Taylor [7] to control a permanent magnet brushless dc motor. In [6], Georgiu and Le Piouffle have shown the basic control structure for PMAC motors, thus exposing the problem of robustness. Marino [11] has extended the approach toward adaptation, where he introduced the estimated variations of the inertia and rotor resistance, following the example of IM. It should be noted, however, that electromechanical systems represent plants that allow a successful application of exact linearization methods because nonlinearities can be modeled with sufficient accuracy.

In our contribution, a complete control design of a voltage fed PMAC motor is proposed. The mathematical model is derived on the basis of Park’s transformation with which the model in natural coordinates is written in orthogonal rotor (d, q) coordinates. The obtained electrical model and the equations of mechanical balance can be grouped into a special class of nonlinear systems that is characterized by nonlinearity in states and by linearity in the control (voltage) vector (affine input). It is shown that the relative degree of the linearized and decoupled PMAC motor model is equal to the degree of the original (the internal dynamics is not present), so the procedure is equivalent to the exact input-state linearization. Thus the PMAC motor model represents a reachable and observable nonlinear system for chosen inputs and outputs, and moreover, both conditions are derived globally (unlike the IM model where the two conditions are not assured).

In the proposed linearization procedure, the load torque is expressed as

$$u = Ri + \frac{d}{dt} \Psi(i, \Theta)$$

where $u = [u_a, u_b, u_c]^T$ and $i = [i_a, i_b, i_c]^T$ are vectors of phase voltages and currents, $\Psi = [\Psi_a, \Psi_b, \Psi_c]^T$ is the vector of magnetic flux linkages, the angle $\Theta$ is defined by the axis of the reference stator winding $a$ and the rotor’s permanent magnet field axis, and $R$ is the diagonal matrix of ohmic resistances. The magnetic flux linkages $\Psi(i, \Theta)$ can be expressed as

$$\Psi(i, \Theta) = L_m i + \Psi_m(\Theta)$$

where the components of vector $\Psi_m(\Theta)$ are a sin projection of permanent magnet constant flux linkage $\psi_m$ on the stator’s windings. Due to the symmetric structure of the rotor and star connection of the stator windings, the total stator inductance can be expressed as $L_s = \text{diag}(L_\alpha + \frac{2}{3}L_m)$. The motor’s electrical torque is obtained from the expression for electric power $p_e = \dot{\mathbf{t}}^T \dot{\mathbf{u}}$, where only torque building components are considered. From the above equations we can write the PMAC motor model as a system of nonlinear differential equations of the fifth order

$$\frac{d}{dt} i = L_s^{-1} \left[ -Ri - \frac{d}{dt} \Psi_m(\Theta) + u \right]$$

$$\frac{d}{dt} \omega = J^{-1} \left[ \omega^{-1} T_L - \frac{d}{dt} \Psi_m(\Theta) - T_L \right]$$

$$\frac{d}{dt} \Theta = \omega$$

where $T_L$ is the load torque. Due to the star connection of stator windings, (3) represents an unreachable nonlinear system with the dimension of the unreachable subspace equal to one. This structure property is physically conditioned by the linear dependence of stator currents. The usual Park’s transformation applied to (2) means, in terms of differential geometry, a special example of a nonsingular diffeomorphism of $x_2 = T(x)$ type with which the original model is decoupled into the reachable part with four dimensions and the unreachable part with one dimension (triangular decoupling). The reachable part of the system is given by the transformed state vector $x_1 = [i_d, i_q, \omega, \Theta]^T$ and the unreachable part
by the state \( i_0 \) (due to the star connection it always holds that \( i_0 = 0 \)). In the geometric sense, Park's transformation applied to the currents, voltages, and magnetic flux linkages can be seen as the transition into the new base \((\alpha, \beta, 0)\) and the rotation of this base for the angle \( \Theta \). Using the standard procedure [14], the PMAC motor model can be expressed in the rotor’s base

\[
\frac{d}{dt} i_d = -\frac{R}{L} i_d + p\omega i_q + \frac{1}{L} u_d \\
\frac{d}{dt} i_q = -\frac{R}{L} i_q - p\omega i_d + \frac{p\psi_m}{L} \omega + \frac{1}{L} u_q \\
\frac{d}{dt} \omega = -\frac{3}{2} \frac{p\psi_m}{J} i_q - T_L J \\
\frac{d}{dt} \Theta = \omega
\]

where \( d \) refers to the axis of the rotor’s permanent magnet field, index \( q \) to its orthogonal axis, \( p \) means the number of pole pairs, \( J \) the rotor’s inertia, and \( L = L_a + \frac{3}{2} L_m \). If the coordinate \( i_d \) and the mechanical rotor position \( \Theta \) are defined as the system’s outputs, the above equation can be given in the following form:

\[
\frac{d}{dt} x = f(x) + \sum_{i=1}^{2} g_i(x) u_i + d; \quad x \in X \subset \mathbb{R}^4, \quad u \in \mathbb{R}^2, \quad f(0) = 0, \\
y_1 = h_1(x); \quad i = 1, 2; \quad h_1(0) = 0
\]

where \( f(.) \), \( g_1(.) \), and \( g_2(.) \) are smooth vector fields on the open set \( U \) in \( \mathbb{R}^4 \), \( h_1(.) \) is a smooth function mapping \( X \) into \( \mathbb{R} \), and \( d \) is a disturbance vector including only the load torque component. The new system’s inputs are the voltage coordinates in the direct \((u_1 = u_d)\) and in the quadrature axes \((u_2 = u_q)\). The system of (5) neglecting the disturbance vector represents the starting point for the input–output linearization and decoupling.

### III. INPUT–OUTPUT DECOUPLING AND LINEARIZATION

The input–output relation of (5) is derived by the successive derivation of outputs \( y_i \) so that at least one input \( u_i \) occurs in the expression for \( y_{i}^{(k)} \) \((k = 1, 2, \ldots)\). Simultaneously the notation \( L \phi h \equiv \nabla h f; \mathbb{R}^n \rightarrow \mathbb{R} \) will be adopted for Lie’s directional derivative of \( h \) along the direction of the vector field \( f \). Thus the following is obtained for the system’s output \( y_1 \):

\[
y_1 = i_d \\
y_1^{(1)} = L f h_1 + \sum_{i=1}^{2} L g_i L f h_1 u_i = f_1(x) + \frac{R}{L} u_d = -\frac{R}{L} i_d + p i_q \omega + \frac{1}{L} u_d.
\]

Using successive derivations, the following is obtained for the second system’s output \((y_2)\), chosen as the rotor position \( \Theta \)

\[
y_2 = \Theta \\
y_2^{(1)} = L f h_2 + \sum_{i=1}^{2} L g_i L f h_2 u_i = f_2(x) = \omega \\
y_2^{(2)} = L f h_2 + \sum_{i=1}^{2} L g_i L f h_2 u_i = f_3(x) = \frac{3p\psi_m}{2J} i_q \\
y_2^{(3)} = L f h_2 + \sum_{i=1}^{2} L g_i L f h_2 u_i = \frac{3p\psi_m}{2J} f_2(x) + \frac{3p\psi_m}{2J} u_d + \frac{3p\psi_m}{2J} u_q.
\]

Since it is obvious that \( L g_1 L f h_1 \neq 0 \) and \( L g_2 L f h_2 \neq 0 \), the total relative degree is \( r = 1 + 3 \) which indicates that the input–output PMAC motor model linearization is equivalent to the input-state linearization, or putting it in other words, the PMAC motor model in \( d-q \) coordinates represents a completely reachable and observable nonlinear system, without the so-called internal dynamics for \( dz \in X \).

Hence it follows:

\[
\begin{bmatrix}
y_1^{(1)} \\
y_2^{(1)} \\
y_2^{(2)} \\
y_2^{(3)}
\end{bmatrix} =
\begin{bmatrix}
L f h_1(x) \\
L f h_2(x) \\
L f h_2(x) \\
L f h_2(x)
\end{bmatrix} + D(x)
\begin{bmatrix}
u_d \\
u_q
\end{bmatrix}
\]

\[
\begin{bmatrix}
y_1^{(1)} \\
y_2^{(1)} \\
y_2^{(2)} \\
y_2^{(3)}
\end{bmatrix} = \begin{bmatrix}
\frac{R}{L} i_d + p i_q \omega \\
-\frac{R}{L} i_d + p i_q \omega \\
\frac{3p\psi_m}{2J} R i_q - p i_q \omega - \frac{p\psi_m}{L} \omega + \frac{3p\psi_m}{2J} u_q
\end{bmatrix}
\]

\[
+ \begin{bmatrix}
\frac{1}{L} \\
0 \\
3p\psi_m \\
2J L
\end{bmatrix}
\begin{bmatrix}
u_d \\
u_q
\end{bmatrix}.
\]

The decoupling matrix \( D(x) \) is obviously nonsingular since it is independent of the system states, so the control vector \( u \) can be generated as

\[
\begin{bmatrix}
u_d \\
u_q
\end{bmatrix} =
D^{-1}(x)
\begin{bmatrix}
u_d \\
u_q
\end{bmatrix}
\]

\[
\begin{bmatrix}
u_1^{(1)} \\
u_2^{(1)} \\
u_2^{(2)} \\
u_2^{(3)}
\end{bmatrix} =
\begin{bmatrix}
u_d \\
u_q
\end{bmatrix}.
\]
The above equation describes the decoupled linear relationship between the transformed model states \( y \) and the new inputs \( v \) in the well-known Brunovski canonical form

\[
\begin{align*}
\frac{d}{dt} y &= Ay + Bv \\
&= \begin{bmatrix} 0 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} y + \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} v_d + \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix} v_q 
\end{align*}
\]

(11)

where the vector of transformed states is defined as \( y = [i_d, \Theta, \omega, \dot{\omega}]^T \). Equation (11) allows one to determine the linear control structure and its optimal parameters.

IV. ROBUST TRACKING CONTROLLER DESIGN

It is expected of servo drives that the rotor position should follow a reference trajectory which is assumed to be an element of \( C_\infty \). In the synthesis of controllers, the method of the prescribed characteristic polynomial of the closed-loop system is applied. To improve the steady-state accuracy, the state controller is extended with the integral control action on both inputs. The problem of tracking is solved by defining the error vector as

\[
e = \begin{bmatrix} e_d \\ e_{q1} \\ e_{q2} \\ e_{q3} \end{bmatrix} = \Phi R - y
\]

(12)

where \( \Phi R \) is the vector of reference trajectories and notation \( \Phi^{(j)}_R \) stands for the \( j \)th derivative of reference position trajectory. The estimated state \( \hat{y}^{(2)}(z) \) (acceleration) which is used in error vector calculation is obtained from the load observer [3]. If we take into account that maximum efficiency operation demands that \( \Phi_{1x}(t) = 0 \) for all \( t \), then the following can be introduced for the new control vector \( u \):

\[
\begin{align*}
\begin{bmatrix} v_d \\ v_q \end{bmatrix} &= \begin{bmatrix} -k_d i_d - k_i \int i_d dt \\ k_i \int e_q dt + \sum_{j=1}^{3} k_q e_{qj} + \Phi^{(2)}_{2x} \end{bmatrix} 
\end{align*}
\]

(13)

Considering (9), the control voltage vector \( u \) is

\[
\begin{align*}
\begin{bmatrix} u_d \\ u_q \end{bmatrix} &= -D^{-1}(x) \begin{bmatrix} f_1(x) \\ 3p_{bm} \omega_f(t) \end{bmatrix} + D^{-1}(x) \\
&= -k_d i_d - k_i \int i_d dt \\
&\quad + k_i \int e_q dt + \sum_{j=1}^{3} k_q e_{qj} + \Phi^{(2)}_{2x} 
\end{align*}
\]

(14)

In the case of nominal parameters and smooth reference trajectories, (14) generally assures asymptotic tracking and perfect tracking if \( \Phi_R(0) = y(0) \). The robustness of the feedback system depends only on the robustness of the linear control law. In case of significant parameter perturbations, (14) no longer assures the stability of the feedback system because the nonlinearities are insufficiently compensated. Stability can be assured for the class of parameter perturbations that do not affect the relative degree by introducing additional compensation signals. In the case of a perturbed plant, the error vector dynamics can be expressed as

\[
\begin{align*}
\frac{d}{dt} e &= \Phi e + B[v + \nu(x, v)] 
\end{align*}
\]

(15)

where \( \nu(x, v) \) is a generally unknown nonlinear function whose upper bound is assumed to be known. For (15), the stability in Lyapunov’s sense can be assured with the introduction of an additional control signal \( \Delta v \), so that \( v = v_n + \Delta v \), where \( v_n \) refers to (14). Thus the feedback system is

\[
\begin{align*}
\frac{d}{dt} e &= \Phi e + B[\Delta v + \nu(x, v)] 
\end{align*}
\]

(16)

where \( \Phi \) is the Hurwitz’s matrix of the controlled nominal system, \( B \) is the input matrix of the form \( B = [b_1, b_2] \), and \( \nu(\cdot) \) the nonlinear function driven by the control signal \( v \) and the tracking error \( e \). If Lyapunov’s function is introduced for (16)

\[
V = e^T P e; \quad \begin{align*}
V &> 0, \quad e \neq 0, \\
V &= 0, \quad e = 0, \\
\lambda^T P + PA_c &= -Q, \\
Q &= \text{diag}(q_i), \quad q_i > 0
\end{align*}
\]

(17)

its derivative is

\[
\begin{align*}
\frac{d}{dt} V &= e^T P e + e^T P \frac{d}{dt} e \\
&= e^T (A^T P + PA_c)e + 2e^T PB[\Delta v + \nu(x, v)] \\
&= e^T Qe + 2e^T PB[\Delta v + \nu(x, v)].
\end{align*}
\]

(18)

The first term in (18) is certainly negative definite. The negative definiteness of the second term is obtained by choosing the following for \( \Delta v \) [12]:

\[
\Delta v = \begin{cases} 
-\rho(e) \frac{B^T P e}{||B^T P e||} & ||B^T P e|| \neq 0 \\
0 & ||B^T P e|| = 0
\end{cases}
\]

(19)

Equation (16) will be stable in Lyapunov’s sense if the following condition is fulfilled:

\[
||\nu(x, v)|| \leq \rho(e).
\]

(20)

For the PMAC motor, the influence of uncompensated nonlinearities on the feedback system is expressed analytically by taking into account (6), (7), (9), and (16). Nonlinear perturbations functions \( \nu_i(\cdot) \) are calculated in the form

\[
\begin{align*}
\nu_d(x, v_d) &= \frac{L}{L - 1} (v_d + \Delta v_d) \\
+ \frac{R}{L - 1} \left( I_d + \frac{1 - L}{L} \right) i_d \omega \\
\nu_q(x, v_q)
\end{align*}
\]
With the above equations we can determine the function $\rho(e)$ in case of known or assumed parameter perturbations. In the case of the PMAC servo drives, it is primarily reasonable to consider the variations of inertia. If inertia variations in the range from 0.5-1.5 $J/J$ are assumed, then the perturbation parameter is $|\alpha| \leq |(J/J) - 1| = 0.5$. It follows from (21):

$$\nu_q(x, v_q) = \alpha(v_{q_n} + \Delta v_q).$$

(22)

Having fulfilled (20) and after choosing $\|\Delta v_q\| = \rho_q(e)$, it follows:

$$\|\nu_q(x, v_q)\| = |\alpha|\|v_{q_n}\| + |\Delta v_q|$$

$$\rho_q(e) \geq |\alpha|\|v_{q_n}\| + \rho_q(e)$$

$$\rho_q(e) \geq \frac{|\alpha|}{1 - |\alpha|} \|v_{q_n}\| = \|v_{q_n}\|.$$

(23)

Fig. 3. Activities inside one sampling period.

The structure of the proposed control algorithm is shown in Fig. 1.
V. ELEMENTS OF THE SYSTEM SOFTWARE AND HARDWARE

The proposed control concept was tested on a 3 kW PMAC servomotor with a current controlled dc motor as the active load. The basic hardware consists of a microprocessor controlled voltage source inverter, an A/D converter with simultaneous sampling of all measurement channels, and a digital signal processor in the PC environment (Fig. 2).

A modified industrial inverter INES (Industrial Electronics Sevnica-Slovenia) was used for the voltage supply of the servo drive. It can operate stand alone, or it allows on-line receiving of the reference voltage vector \( u = [u_q, u_d]^T \) from the digital signal processor. The frequency of the inverter can be varied in the range from 4–18 kHz. The data transmission was serial at a transmission speed of 2.5 Mbit/s. The total transmission time per one reference voltage vector was approximately 100 \( \mu s \) which is due to synchronization. Current sensors, a resolver and a tachogenerator, were used for data acquisition. Simultaneous sampling of the five channels (two current signals, two resolver signals, and a speed signal) was synchronized by a zero-state voltage vector generated by the inverter. Two four-channel A/D converters were directly connected to the signal processor by the DSPLINK bus. The effective time of the five channel-12 bit A/D conversion was 50 \( \mu s \). The entire control algorithm was written in the programming language C for a PC extension card SMIS 32C (with an AT&T DSP 32C digital signal processor) with a computation power of 25 MFLOPS.

The activities of the signal processor, the peripheral unit, and the inverter microcontroller within one sampling period are shown in Fig. 3.

At the beginning of each sampling interval the voltage inverter generates the trigger signal for the S/H circuits on A/D converters. After converting the first channel, the A/D’s controller sends the interrupt to the signal processor which starts the successive software controlled A/D conversions of the remaining four channels. This is done within the interrupt routine. Having calculated the new control voltage vector, the signal processor sends the interrupt request to the inverter’s microcontroller. At the beginning of the next sampling period, the inverter forms the actual control voltage vector and again starts S/H circuits. Due to time constraints caused by A/D conversion, the communication between DSP and inverter microcontroller and the computation of the control algorithm a sampling period of 400 \( \mu s \) at the selected converter frequency of 10 kHz was attained. In this way, the voltage inverter repeated the same control voltage vector four times within sampling interval.

VI. EXPERIMENTAL RESULTS

The proposed control algorithm was implemented in four program modules. The first module acted as the generator of smooth reference trajectories so that the rotor position is a three times continuously derivable function.

The reference trajectories were so planned that the state variables (currents, acceleration, and speed) did not exceed the prescribed bounds. In the case of presented experiments,
the parameters $\Omega_{\text{max}} = 209.4$ rad/s and $\omega_0 = 10.471$ rad/s were selected, so the motor reached the final reference position value $\Theta_R = 10\pi$ rad in $t = 0.3$ s. The second program module contained the transformation of phase currents into the rotor's coordinates on the basis of sinusoidal signals from the resolver. The new voltage vector in $dq$ coordinates was calculated in accordance with (14) or completed with the compensation signal $\Delta u_q$ (19) for a perturbed system. The parameters of the controller in the $q$ axis for (14) were selected so that the poles of the feedback system had the following values: $s_1 = -100$, $s_2 = -110$, $s_3 = -120$, and $s_4 = -130$. The voltage control vector in the last program module was transformed into $\alpha\beta$ coordinates and transmitted to the voltage source inverter.

Fig. 4 shows the tracking control of the motor at no load for nominal parameters. Due to the noise in measured signals, numerical errors, and most of all because of the delays in the processing of the new control voltage vector, the torque control is not ideally solved at a high speed.

The tracking control of a servo drive with load is given in Fig. 5 for similar reference trajectories as in the previous case. A constant load torque of 7.5 Nm was generated with the current controlled DC motor at the time $t_L = 0.02$ s. The influence of the load is clearly seen in the current $i_q$.

Fig. 6 refers to the case of the perturbed inertia. For the prescribed poles of the feedback system and for the selected weighting matrix $Q = \text{diag}(1, 0.1, 0.01, 0.001)$, Lyapunov's matrix equation $A_x^T P + PA_x$ is solved so that the generalized error of the state vector from the prescribed reference trajectories vector is

$$b^T Pe_q = 10^{-3}[0.05, -0.3, -0.01, 0.06]e_q.$$  (24)
If an inertia variation in the range from 0.5–1.5 \( \tilde{J}/J \) is assumed, the obtained compensation signal \( \Delta v_q \) is

\[
\Delta v_q = \begin{cases} 
-|v_q| \frac{b^T P e_q}{|b^T P e_q|}; & b^T P e_q \neq 0 \\
0; & b^T P e_q = 0.
\end{cases}
\]  

(25)

In Fig. 6 the graphs show the tracking control of the perturbed system without and with the compensation signal. In the first case, a substantial worsening of the dynamic behavior is observed, whereas in the second case the compensation signal improves the dynamics, and, most importantly, it assures the stability of the perturbed system within the given parameter bounds.

VII. CONCLUSION

In the paper, nonlinear tracking control structure for PMAC servo drives derived from the exact linearization and decoupling of the model in rotor coordinates is presented. With the described procedure, a globally linearized and decoupled PMAC motor model can be written in the reachable and observable Brunovski canonical form that allows a methodical determination of the control structure and an optimal parametrization of tracking controllers for the class of smooth references. Compared to classical (serial) solutions, the proposed control approach improves the dynamics performances and the accuracy of the servo drive. The influence of the load is considered implicitly by introducing a disturbance observer. Also, a detailed analysis of the influence of parameter perturbations that cause an incomplete compensation of the plant’s nonlinearities is included. In these cases the stability of the feedback system is assured by introducing an additional compensation signal which is calculated from Lyapun’s function of the perturbed plant. The advantages of the proposed approach are confirmed through laboratory experiments. Further work will be focused on the analysis of voltage inverter delays, sensor noise and the influence of unmodeled dynamics.

REFERENCES