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Microcomputer based sliding regime discrete control for synchronous AC motors

Control discreto por microordenador basado en regímenes deslizantes para motores de corriente alterna síncronos

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Abstract

The basic ideas to design discrete algorithms used to calculate unmeasured mechanical or electrical variables, in order to control electrical motors in mechanical systems are presented. The main point of the paper is to simplify the algorithms by using a linear discrete time model without any variable limitations and to reduce the computing capacity requirements of the controller. Limited references, based on the discrete sliding mode for the exception of the influences of variable limitations, are proposed and designed. The main advantage of the controller presented in this paper is that a large control error does not bestow any problems. In such a case, the system always works in an area without the limitations. Original observation algorithms of both position and rotational velocity for exterior permanent magnet synchronous motors are designed. A computer simulation was performed and the results are presented, showing high dynamic accuracy.

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Key words: sliding mode, observer based control, simulation, limitation algorithms.

Resumen

Se presentan las ideas básicas para diseñar algoritmos discretos para calcular las variables mecánicas o eléctricas, no medidas sino estimadas, para controlar motores eléctricos en sistemas mecánicos. El objetivo principal de este artículo es simplificar los algoritmos usando un modelo de tiempo discreto lineal sin ninguna limitación en las variables y reducir los requisitos de capacidad informática del controlador. Las referencias limitadas, basadas en el modo deslizante discreto para la excepción de las influencias de limitaciones de las variables, se proponen y diseñan. La ventaja principal del controlador presentada en este artículo es que un error de control grande no causa ningún problema. En dicho caso, el sistema trabaja siempre en un área sin limitaciones. Se diseñan los algoritmos de la observación originales de la posición y de la velocidad rotatoria para el imán permanente exterior de motores síncronos. Se llevó a cabo una simulación en computadora y se presentan los resultados, que muestran una exactitud dinámica alta.

Palabras claves: modos deslizantes, control basado en observador, simulación, algoritmos de limitación.

1 Introduction

An innovative, new, efficient discrete control system for complex electromechanical position and speed control systems is a real necessity these days [1, 2, 3, 4], because of the increasingly demanding requirements for the control and the need to lower costs. The evolution of microprocessors has lowered system prices and improved their efficiency and speed, opening new doors to the design of discrete control systems. Exterior permanent magnet synchronous motors (EPMSM) have found wide applications due to their high power density and ease of control in relation to other AC motors. Luenberger observers are the foundation of the state estimation theory. He proposes AC motor control –with or without– observers for mechanical variable sensors.

Drives with a wide range of rotation speed, such as one to hundred or one to thousand and even higher, have a wide range of practical applications in real life, such as machine–tool construction and robotics. Several schemes have been proposed to identify the rotor velocity using stator windings voltage and current measurements [1, 2, 4, 5, 6] in past decades. However, the problem is very complex and a full solution for such a problem is yet to be found and improved. Most of such electromechanical systems are highly nonlinear. Their dynamic behaviour depends on the operating point of the system. Discrete control should solve two important tasks: to calculate (or observe) unmeasured control variables such as rotation speed Ω, angular position Θ, load torque, T_L, and so on, and to control the servomechanism. A mathematical model for state variable calculation and control is required. Such a model would be very complicated, mainly because there are several nonlinearities in the system, which can be divided in two main categories, namely, those which are inherent to the system and therefore are always present, and those dependent on the control error values.

The first class nonlinearities are the physical nonlinearities of AC motors and those of the voltage source inverter (VSI). The second class nonlinearities are gear backlash, dry friction and the different limits of the system variables, such as voltages and currents, mechanical angular velocity, system output dynamics and the rate of the control calculations being carried out by the computer. Nonlinearities make control design very difficult.

The first class nonlinearities are always in the system and must be taken into account when designing the unmeasured variable estimator and the control. The second ones would be taken into account only when the control error is very low or very high. In the first case, only such nonlinearities as the gear backlash –and if an error sign is changed– dry friction must be considered in the mathematical model. In the second case only the limits of the variables must be taken into account. We will only research the limitations case, and later on, understand the situation under which the small control error is such, that it does not reach those limits. A large control error creates another error. It is very important to consider the case of a large control error, even if it does not occur normally under typical operation. These errors occur when operation starts or changes quickly or suddenly. The incorrect estimation of the variables that are used for control purposes could make the system unstable. The estimation of unmeasured variables in such high order systems, with nonlinearities and limits, is a very difficult task and needs the fault mathematical model of the system with limits. A “fault mathematical model” must be used in case we need a very complicated high order mathematical model. This model again is very nonlinear, and therefore it is very difficult to analyse it. It considers extreme cases to achieve recovery in cases where the system fails to operate correctly. Such model for variable estimation would be a very complicated observation algorithm. The same is true for the control model too. However, these complicated algorithms would be useful only for the short term, because there are no limitations in the system in typical operation, so more simple estimation algorithms could be used.

One of the possible solutions is to use only simple algorithms, such as formatting the "smooth" reference for the control system. The main goal is that in case a large step reference cannot be realised by the system, because of limitations, the reference limiter produces a different reference, which can be realised by system. In this case, the system by the large control error works on the board of the limitation in the area of the system nonlinearities only. The designed closed loop system uses the estimated variables to produce a small control error. The system has a good dynamic behaviour and makes good use of the system resources. The primary aim of this paper is to explain basic ideas related to the design of a combination of a limiter and an observer for the electromechanical system with a synchronous motor with external permanent magnets (to be referred later on as "motor") and VSI in the digital form. The paper can be outlined as follows. Section 2 deals with a brief introduction to the digital model of the motor and assumptions used. The observation algorithms for the unmeasured control variables and load is given in section 3. In section 4 the proposed reference limitation algorithm, on the base of the digital sliding mode, is introduced. The control algorithm is outlined in section 5. The results of the numerical simulation, that illustrate the properties of the complete control, are presented as graphs as a function of time in section 6 followed by conclusions.

2 Motor discrete–time model

Exterior permanent magnet synchronous motors have found wide applications due to their high power density and ease of control in relation to other AC motors. However, designing the control for such VSI fed motor is a difficult undertaking, due to complicated nonlinearities in the motor and VSI. As the controller works in discrete time, it is necessary to have a discrete model for the control system. Park's equations for the motor, written down in rotor–oriented rotating reference frame (d,q), and in stationary (stator) one (α,β), can be used as a base. Axis Od is assumed to be in the direction of the rotor flux, and axis $O\alpha$ is to be in the direction of axis R (one of the axis of three–phase reference frame).

The transformation of motor phase currents (i_A, i_B, i_C) and voltages (u_A, u_B, u_C) to the one of the vector components in the reference frame $(\alpha,\beta)(i_\alpha, i_\beta)$ and (u_α, u_β) is done using (1) (i.e. for windings currents).

$$\begin{bmatrix} 1_\alpha \\ 1_\beta \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix}. \tag{1}$$

The transformation of the variables from the reference frame (α,β) to one (d,q) (i_d, i_q) and (u_d, u_q) is performed using Park's transformation, where $\theta(\theta = p\Theta$, p are pole pairs) is the electrical angle between d–axes of rotor rotating frame (d,q) and α–axes of stationary (stator) one (α,β) (i.e. for current)

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix}. \tag{2}$$

The discrete–time model of the motor works under the following assumptions:

- Motor behaviour is analysed and variables are recalculated on a constant sampling period Δ.
- The mechanical time constant (typically 10–100 ms) of the system, and the switching time constant (typically 10–100 μs) of the controller, differ significantly. The switching time constant is much smaller than the mechanical time constant.

The angular velocity ω and the angle θ are quasi–static parameters in the electrical equations of the motor, and the motor is described by the following discrete–time equations:

$$i_d^{n+1`} = i_d^{n`} + \frac{\Delta}{L}(-r i_d^{n`} + L\omega_{eq}^{n} i_q^{n`} + u_d^{n`}) \tag{3}$$

$$i_q^{n+1} = i_q^n + \frac{\Delta}{L}(-ri_q^n - L\omega_{eq}^n i_d^n - \Psi_f \omega_{eq}^n + u_q^n) \tag{4}$$

$$i_\alpha^{n+1} = i_\alpha^n + \frac{\Delta}{L}(-ri_\alpha^n + \Psi_f \omega_{eq}^n \sin\theta_{eq}^n + i_\alpha^n) \tag{5}$$

$$i_\beta^{n+1} = i_\beta^n + \frac{\Delta}{L}(-ri_\beta^n - \Psi_f \omega_{eq}^n \cos\theta_{eq}^n + u_\beta^n) \tag{6}$$

$$\theta^{n+1} = \theta^n + \Delta\omega^n + \frac{\Delta^2}{2J}(T_{eq}^n - T_L^n) \tag{7}$$

$$\omega^{n+1} = \omega^n + \frac{\Delta}{J}(T_{eq}^n - T_L^n) \tag{8}$$

$$T_L^{n+1} = T_L^n \tag{9}$$

where r is stator winding resistance, L is stator winding inductance, ω is electrical rotational velocity ($\omega = p\Omega$), Ψ_f is the no–load magnet flux linkage, T is the electrical drive created torque, J is the moment of inertia, n is the sampling period number (variable upper index), $\omega_{eq}^n = \omega^n + (\Delta/2)(1/J)(T_{eq}^n - T_L^n)$ is the average value of the rotation velocity on a sampling period Δ, θ_{eq}^n is the average value of the angle on a sampling period Δ, $T_e^n = \Psi_f(i_q^{n+1} + i_q^n)/2$ is the average value of the torque on a sampling period Δ, and it is estimated as the arithmetic average of the mean values found above the component of the current at the beginning and at the end of a sampling period Δ.

3 Discrete–time observation algorithms

The measured variables are the currents and voltages in the stator windings of the VSI fed motor. First of all, the PWM component in both measurements must be excluded. As an exception of a switching component in current measurements, it is necessary to measure the currents, not in any, but only in certain moments of time within the PWM period. As a value for the average voltage without the PWM component, the reference value of the voltage from the digital controller can be used. Due to limitations of the computing capacity of the processor, it is necessary to calculate estimations of the control variable only once for every sampling period. The following features caused by the discrete–time character of calculations in a controller are inherent in the decision of the observation task and then the control.

First of all the calculation has to be synchronized with the PWM period Δ, which is constant (the PWM frequency is fixed). Second, the regulation task cannot be carried out faster than for two sampling periods. Namely, the first period $[n, n+1]$ is used for calculations of the value of the unmeasured variables and load and then of the desired value of VSI voltage by using of the initial information about variables, received as a result of measurements and available in the processor. The second period $[n+1, n+2]$ is used to implement the designed control.

Discrete–time equations enable simultaneous identification of all mechanical variables. However, they are too difficult, both for the analysis, and for observer design. Therefore, a step–by–step design procedure is used. There are current measurements, i_α^{n+1}, i_β^{n+1} (made at the beginning of the period) and past ones i_α^n, i_β^n, during the $(n+1)$–th period. Also, the values of voltages, which were applied to stator windings u_α^n, u_β^n, are known. First estimation of the angle θ_{eq}^n can be solved by using equations (5) and (6)

$$\theta_{eq}^n = -\arctan\frac{i_\alpha^{n+1} - i_\alpha^n - \frac{\Delta}{L}(u_\alpha^n - ri_\alpha^n)}{i_\beta^{n+1} - i_\beta^n - \frac{\Delta}{L}(u_\beta^n - ri_\beta^n)}\,. \tag{10}$$

Observers are useful if there are errors that are introduced by sensors and/or system information processing (in particular due to noise and digital behaviour of measurements) or if there are not enough control variables. For example, suppose the value of the load torque is not available. In that case it is possible to use a state observer [3]. The observer design is carried out based on the mechanical variable equations (7), (8) and (9) (the observer variables have upper indices *):

$$\theta^{(n+1)^*} = \theta^{n^*1} + \Delta\omega^{n^*}\frac{\Delta^2}{2J}(T_{eq}^n - T_L^{n^*}) + l_1(\theta^{n^*} - \theta_{eq}^n) \tag{11}$$

$$\omega^{(n+1)^*} = \omega^{n^*1} + \frac{\Delta}{J}(T_{eq}^n - T_L^{n^*}) + l_2(\theta^{n^*} - \theta_{eq}^n) \tag{12}$$

$$T_L^{(n+1)^*} = T_L^{n^*} + l_3(\theta^{n^*} - \theta_{eq}^n)\,, \tag{13}$$

where l_i are observer coefficients parameters, based on the difference between estimated and calculated values of angle. The calculation of coefficients l_i were carried out in depend of the convergence rate of an estimation error to zero. In our case the same three real modes $\lambda_i = \lambda_1 = \lambda_2 = \lambda_3$ in the characteristic equations have been used:

$$-\lambda^3 + \lambda^2(\lambda_1 + \lambda_2 + \lambda_3) - \lambda(\lambda_1\lambda_2 + \lambda_1\lambda_3 + \lambda_2\lambda_3) + \lambda_1\lambda_2\lambda_3 = 0 \tag{14}$$

$$-\lambda^3 + \lambda^2(3 + l_1) - \lambda(3 + 2l_1 + \frac{l_3\Delta^2}{2J} - l_2\Delta) + (1 + l_1 - l_2\Delta - \frac{l_3\Delta^2}{2J}) = 0\,. \tag{15}$$

Equations (11) and (12) are the observers but they are also digital filters. These are simply two ways to look upon them or to designate them. The structure of a full–order observer and a Kalman filter are essentially the same. Equation (13) may be an observer, but it is also a model–based or model–free observer.

4 Sliding mode based references limitation algorithm

There are limitations in mechanical motions, due to current limitations of the power converter, the maximal value of the motor torque $T_{\max}$ and then the acceleration E is bounded by the minimal and maximal values

$$-\frac{T_{\max} + T_L}{J} = E_{\min} < E < E_{\max} = \frac{T_{\max} - T_L}{J}\,. \tag{16}$$

Due to voltage limitations of power supply and mechanical durability, the output velocity is bounded by

$$|\Omega| < \Omega_{\max}\,. \tag{17}$$

We have to consider the following points for the control system:

- Ability to acquire the reference under dynamic limitations in case the reference cannot be obtained.
- Absence of a dynamic error by the submitting the accrued reference to the control.
- Staying–in–stability linear zone on deviations of control variables.

The reference rate limiter represents dynamic system with limitations on variable and on rate of their change. The input and output signals of the limiter have such indices: upper one "n" is the period number, lower one "lim" is the output signal that is the reference signal for the control and "z" is the drive reference signal that is the input signal. The high level regulator (e.g., digital control system) forms a reference value of position Θ_z^n. The discrete time equations of the rate limiter can be written as follows:

$$(\delta\Theta)_z^{n+1} = (\delta\Theta)_z^n + (\delta\Omega)_z^n\Delta + \frac{(\delta E)_z^n\Delta^2}{2} + \frac{v^n\Delta^3}{6} \tag{18}$$

$$(\delta\Omega)_z^{n+1} = (\delta\Omega)_z^n + (\delta E)_z^n\Delta + \frac{v^n\Delta^2}{2} \tag{19}$$

$$(\delta E)_z^{n+1} = (\delta E)_z^n + v^n\Delta\,, \tag{20}$$

where $(\delta\Theta)_z^n = \Theta_{\lim}^n - \Theta_z^n$, $(\delta\Omega)_z^n = \Omega_{\lim}^n - \Omega_z^n$, $(\delta E)_z^n = E_{\lim}^n - E_z^n$, v is the rate limiter control.

It must ensure convergence to zero from deviations between the limiter input and output signals, on conditions that acceleration and rate of limiter output signals do not surpass their minimal and maximal values. In this case there are no additional requirements to reference signals, in particular, the position reference can change in steps, which is characteristic, for example, when positioning. This problem can be solved with the help of a sliding mode organization on the surfaces [6]:

$$S_1^{n+1} = E_{\lim}^{n+1} - E_{\max} = 0 \tag{21}$$

$$S_2^{n+1} = E_{\lim}^{n+1} - E_{\min} = 0 \tag{22}$$

$$S_3^{n+1} = c(\Omega_{\lim}^{n+1} - \Omega_{\max}) + E_{\lim} = 0 \tag{23}$$

$$S_4^{n+1} = c(\Omega_{\lim}^{n+1} + \Omega_{\max}) + E_{\lim} = 0 \tag{24}$$

$$S_5^{n+1} = (\delta E)_z^{n+1} + b_1(\delta\Omega)_z^{n+1} + b_2(\delta\Theta)_z^{n+1} = 0\,. \tag{25}$$

The acceleration and velocity limitations are organized by means of sliding modes along surfaces (21)–(24). The sliding mode on a surface (25) is used for the ensure of the convergence to zero of deviations between the limiter input and output signals by conditions that acceleration and rate of limiter output signals do not surpass their minimal and maximal values.

The characteristics of moving an error to zero after occurrence of a sliding mode on a surface $S_j = 0$, $(j = 1, \ldots, 5)$ is defined by a choice of the coefficients b_1, b_2, c.

Additionally, the limiter control v^n can be bounded within minimal and maximal values

$$-v^n_{\max} < v^n < v^n_{\max} \,. \tag{26}$$

The values of v^n that are necessary for performance, for conditions (21)–(25) on the $(n+1)$–th period, have been calculated (table 1).

Table 1: values of the limiter control

$\mathbf{j}$	$\mathbf{v_j^n}$
1	$-\frac{E^n_{\lim}-E_{\max}}{T}$
2	$-\frac{E^n_{\lim}-E_{\min}}{T}$
3	$-\left(\frac{c(\Omega^n_{\lim}+E^n_{\lim}T-\Omega_{\max})}{\Delta+\frac{b_1\Delta^2}{2}}+\delta E^n_{\lim}\right)$
4	$-\left(\frac{c(\Omega^n_{\lim}+E^n_{\lim}T+\Omega_{\max})}{\Delta+\frac{b_1\Delta^2}{2}}+\delta E^n_{\lim}\right)$
5	$-\frac{\delta E^n_z\left(1+b_1\Delta+\frac{b_2\Delta^2}{2}\right)+\delta\Omega^n_z(b_1+b_2\Delta)+\delta\Theta^n_z b_2}{\Delta+\frac{b_1\Delta^2}{2}+\frac{b_2\Delta^3}{6}}$

Then the absolute values of these controls are compared with each other, and the control with the least absolute value used for the calculation of the limiter output signals (18)–(20), i.e. control the reference ones

$$v^n = \gamma \min\{|v^n_1|, |v^n_2|, |v^n_3|, |v^n_4|, |v^n_5|, |v^n_{\max}|\}$$

where γ is a sign of limiter control with the least absolute value.

5 Discrete–time control algorithm

The control of a VSI fed motor is depicted with a block diagram. It accomplishes direct discrete vector control of the motor, using a discrete model of the process, independent variable estimation, and reference rate limitation. The closed loop control system presented consists on a reference, a squirrel cage motor with permanent magnets and a servomechanism, its electronic interface and drive and a computer executing control algorithms on a periodical basis (constant PWD period), to update the values of the variables and the control signals. The control output signals are the control signals for the VSI power keys. The control system has two blocks, namely, a calculator of the reference value of the torque and a current control block. The structure of a drive with control is shown in figure 1.

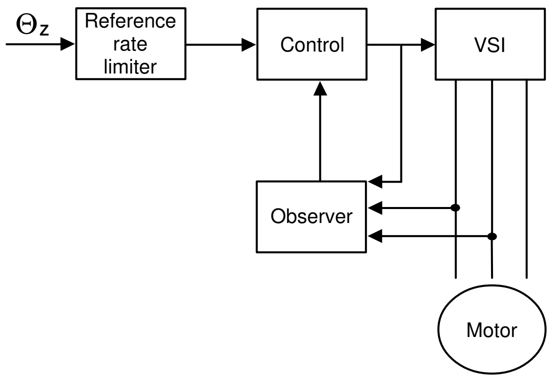


Figure 1: Structure of a digital motor control

The proposed controller is used for as speed control. Position estimation is not necessary, since the angle can be calculated by using the information on the current and voltage, to estimate the speed and torque, which are used in turn for control. The input of measurement and task signals is carried out once on each calculation period of the vector control, which is equal to multiple PWM periods. It is assumed that at the beginning of the operation of the system, all values for the variables $[n, n+1]$ are known, i.e.:

- Actual values of current components i_d^{n+1}, i_q^{n+1}, which are calculated using the phase currents (i_A, i_B, i_C) and the estimated rotor angle θ_{eq}^n or θ^{n^*}.

- Actual values of the voltage components u_d^{n+1}, u_q^{n+1}, which were calculated on the previous calculation period and are used in this one.

- Estimated value of the rotation speed ω_{eq}^n.
- Limiter output signals, which are the reference signals for control.

The reference value of the torque is calculated by using of the limiter output signals

$$T_{eq\,\hat{u}}^n = T_L^n + J\left[E_{\lim}^n + a_0\left(E_{\lim}^n - \frac{T^{n^*}}{J}\right) + a_1(\omega_{\lim}^n - \omega^*) + a_2(\theta_{\lim}^n - \theta^*)\right], \quad (27)$$

where the value for the coefficients a_0, a_1, a_2 are chosen from the needed value of the mode of the characteristic equation

$$\det\begin{vmatrix} 1+\frac{\Delta^2}{2}a^2-\lambda & \Delta+\frac{\Delta^2}{2}a_1 & \frac{\Delta^2}{2}a_0 \\ \Delta a_2 & 1+\Delta a_1-\lambda & \Delta a_0 \\ a_2 & a_1 & a_0-\lambda \end{vmatrix} = 0\,. \quad (28)$$

The job of the current control block consists in forming the reference value of the voltage u_d, u_q, which ensures solving the control task. The control goal is to keep the actual and reference values of the control variable equal (for example, angular velocity or position). If the control law can be written as the following function of the control error of the angular velocity $\delta\Omega$ (or error of rotating angle $\delta\theta$)

$$S = \delta\Omega + \alpha(d\delta\Omega/dt)\,. \quad (29)$$

There is asymptotic convergence of a control error $\delta\Omega$ when $S = 0$, with the time constant $\tau = 1/\alpha$. By $\alpha = 0$ there is finite step procedure of regulation (digital sliding mode [6]) when the equality of actual value of a rotation velocity to the reference one is provided to the time moment $(n+2)$. It is possible to calculate the needed average value of controlled variables (the torque T_{eq}^{n+1} and the value of the rotation velocity on the $(n+1)$ period end). The condition $\delta\Omega^{n+2} = \alpha^2\delta\Omega^n$ can be used to find the needed value of the torque T_{eq}^{n+1} and then the needed value of the voltage components u_d^{n+1} and u_q^{n+1}:

$$u_d^{n+1} = -(L/\Delta)(i_d^{n+2} - i_d^{n+`1}) + ri_d^{n+`1} - L\omega_{eq}^n i_q^{n+`1} \quad (30)$$

$$u_q^{n+1} = -(L/\Delta)(i_q^{n+2} - i_q^{n+`1}) + ri_q^{n+`1} + L\omega_{eq}^n i_d^{n+`1} + \Psi_f\omega_{eq}^n\,, \quad (31)$$

where the current component i_d^{n+2}, i_q^{n+2} on the calculation period $[k+2, k+3]$ are $i_d^{n+2} = 0$, $i_q^{n+2} = T_{eq}^{n+2}/\Psi_f$.

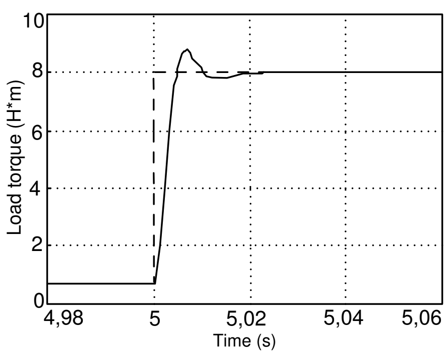
Components u_d, u_q would be used for the next step rotation speed and position estimation in the observer. Output signals of a regulator are the reference signals u_α, u_β in stationary coordinate system for PWM control of VSI. They could be found by using an estimated value for the rotor position θ.

6 Simulation results

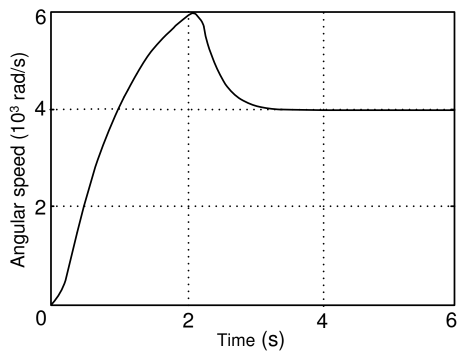
The above mentioned algorithm was simulated using a program in MatLab 6.5.0 and Simulink 5.0. The motor used has the following parameters: $R = 0,038\Omega$, $L = 0,24\text{mH}$, $X = 1,5\Omega$, $U_\Phi = 220\text{V}$, $f = 50\text{Hz}$, $J = 0,01\text{kg}\cdot\text{m}^2$, $p = 2$, $\Psi_f = 0,06\text{N}\cdot\text{m/A}$, $P = 50\text{kW}$. The system used for simulation includes a motor model, a model for VSI, reflecting PWM algorithm, an observer algorithm, a digital controller algorithm, a reference rate limiter, some means of formation of the given velocity, an algorithm for processes indication, and another one for motor parameters input.

The models of VSI, observer and control work in discrete time. The opportunity of input and change motor parameters is stipulated during simulation. The motor and VSI simulation was carried out in real physical variable by using primary SI units (V, A, N·m, radians, sec). The analysis of processes was carried out under the transitive characteristics of drive. They are the reaction of the observer and the drive on a step input of the speed reference and on one of the load torque. The reference value of the rotation speed is 4.000 electrical rad/sec. The step input of the torque is from $0,7$Nm to 8Nm on the fifth second. The modes of the observer are identical and equal to 0,9. The modes of the control are identical too and equal $0,98$. The control coefficients are $a_0 = 0,9412$; $a_1 = -17,73$ and $a_2 = -1782$.

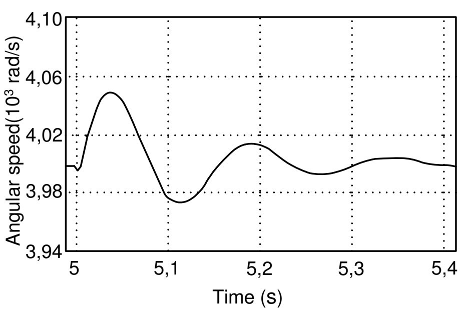
On the given figures, the horizontal scale is time in seconds, the vertical scale is the control variable or the estimated one. In figures 2(a)–2(f) the observer reaction are depicted. Here the soft line corresponds to the actual value of the variable, while the bright one is the estimated value. The estimation of the step input of the load torque is shown in figure 2(a). The estimation of the rotation speed is depicted in figures 2(b)–2(d). The estimation value of the rotation speed and its actual value are practically equal. The error can be observed only in the first moment under the torque load step input in figures 2(c)–2(d) and is very small. The estimation of the rotor angle is depicted in figures 2(e)–2(f). There is no error under the torque load step input. The step behaviour of the estimation value of the angle depends on the digital character of its calculation. The lower bright step line is calculated by using equation (10). The higher one is a result of the angle estimation by using of the observer (11)–(13). The observer has practically no error.



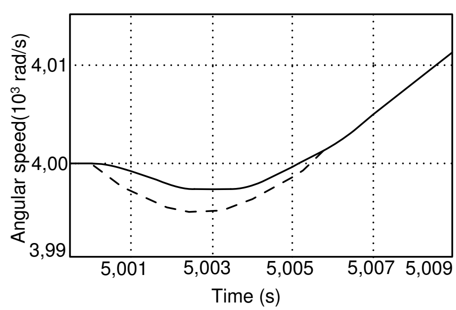
(a) Estimate step input of load torque



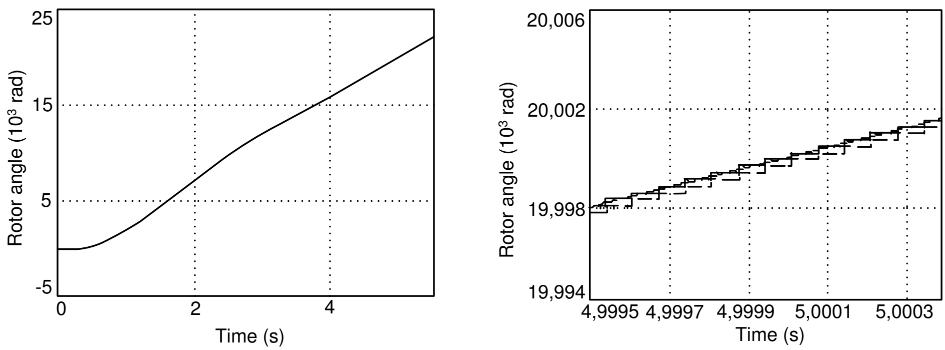
(b) Estimated angular speed



(c) Estimated angular speed under load torque



(d) First moment estimated angular speed under torque load



(e) Estimated rotor angle

(f) Estimated rotor angle under torque load

Figure 2:

7 Conclusions

Algorithms for the discrete–time observer and reference limiter and the observer–based electromechanical system for the synchronous motor and VSI were developed. They offer high qualitative characteristics for variable estimation. The procedure of observation design is based on decomposition of an initial task on two independent tasks: control variable observation in the system without variable limitations, and formation of the control plant references, which is independent of variable limitations influences. The simulations performed for the control plant have confirmed fitness and appropriateness for use and adequate quality of the discussed algorithms.

We demonstrated experimentally that the system is stable. The graphs presented here with the transitive characteristics confirmed stability. We verified the stability using transitive characteristics. We could find out if the system is stable mathematically, but the analysis is particularly complex because of the nonlinearities.

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