Digital sliding mode based references limitation law for sensorless control of an electromechanical system

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Abstract. This paper is intended to explain the basic ideas related to designing digital algorithms using unmeasured mechanical variables (but observed without sensors), to control electromechanical systems digitally. The main idea is to simplify algorithms by using a linear discrete-time model without any variable limitations. Original references limiter based on the digital sliding mode for the exception of the variable limitations influences is proposed and designed. The simulation confirmed high dynamic accuracy, simplification of the digital control algorithm, and reduction of the computing capacity requirements of the controller.

1. Introduction
Designing an effective digital control system for the complex electromechanical system without sensors for the mechanical coordinates is a rather urgent problem these days [1, 2, 5] and the two main reasons are the increasing requirements for control and the need to lower costs. Drives with a wide range of rotation speed (1:100, 1:1000 and higher) could be used in machine-tool construction, robotics, and other applications. During the last decades, several schemes have been proposed to identify the rotor velocity by stator windings voltages and currents measurements [1, 2, 4, 5]. Nevertheless, the complete solution of such a problem is very complicated. Most of such electromechanical systems are not only highly nonlinear but exhibit a large number of different nonlinearities. Their dynamic behaviour depends on the operating point of the system. Digital control should solve two important tasks: to observe unmeasured control variables (e.g. rotation velocity $\Omega$ and angular position $\Theta$, unmeasured load torque $\tau_l^L$, etc) and to control variables. A mathematical model for state variable observation and control is needed. Such a model would be very complicated, because there are many nonlinearities in the system, related to the value of the control error and can be divided in two large classes, namely, those which are always present, and those dependent on the control error value.

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 limitations of the system variables, such as VSI voltages and currents, mechanical angular velocity, system output dynamics and the rate of the control calculations being carried out.

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The first class nonlinearities are always in the system and must be taken into account when designing the unmeasured variable estimator and the control as well. 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 limitations must be taken into account. In our case, we will investigate only the limitations case and later on understand the situation under which the small control error is such, that it does not raise limitations. A large control error raises another error. It is very important to consider the case of the large control error, even if it does not occur under typical operation. These are when operation starts or changes quickly. The incorrect estimation of the variables that are used for the control design could make the system unstable. The estimation of the unmeasured variables in such high order systems, with nonlinearities and a number of limitations, is a very difficult task and needs the fault mathematical model of the system with limitations. 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, as in the typical operation there are no limitations in the system and 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. 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 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 Oα 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 (α,β) (i_α, i_β) and (u_α, u_β) is done using the following equation (i.e. for windings currents):

\[
\begin{bmatrix}
    i_α \\
    i_β \\
\end{bmatrix} = \sqrt{2/3} \begin{bmatrix}
    1 & -1/2 & -1/2 \\
    0 & \sqrt{3}/2 & -\sqrt{3}/2
\end{bmatrix} \begin{bmatrix}
    i_A \\
    i_B \\
    i_C \\
\end{bmatrix}
\]  

(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 θ (θ = pθ, 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):
The discrete-time equations of the motor is carried out under the following assumptions:

- Motor behaviour is analysed on a sampling period $\Delta$.
- The mechanical time constant (typically 10 - 100 ms) and switching time constant (typically 10 - 100 µs), of the processes in PMSM, are essentially different.

In this case velocity $\omega$ and angle $\theta$ are quasi-static parameters in the electrical equations of the motor, and the motor is described by the following discrete-time equations:

$$
\begin{align*}
[i_d]_{n+1} &= [i_d]_n + \frac{\Delta}{L}(-ri_d^n + L\omega^n_{eq,i}i_d^n + u_d^n) \\
[i_q]_{n+1} &= [i_q]_n + \frac{\Delta}{L}(-ri_q^n - L\omega^n_{eq,i}i_q^n - \Psi_f \omega^n_{eq} + u_q^n) \\
[i_a]_{n+1} &= [i_a]_n + \frac{\Delta}{L}(-ri_a^n + \Psi_f \omega^n_{eq} \sin \theta_{eq} + i_a^n) \\
[i_b]_{n+1} &= [i_b]_n + \frac{\Delta}{L}(-ri_b^n - \Psi_f \omega^n_{eq} \cos \theta_{eq} + u_b^n) \\
\theta^n_{eq} &= \theta^n + \Delta \omega^n + \frac{\Delta^2}{2J} (T_{eq}^n - T_L^n) \\
\omega^n &\equiv \omega^n + \frac{\Delta}{J} (T_{eq}^n - T_L^n) \\
T_L^n &\equiv T_L^n
\end{align*}
$$

where $r$ is stator winding resistance, $L$ is stator winding inductance, $\omega$ is electrical rotational velocity ($\omega = p\Omega$), $\Psi_f$ 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), $\omega^n_{eq} = \omega^n + (\Delta/2)(1/J)(T_{eq}^n - T_L^n)$ is the average value of the rotation velocity on a sampling period $\Delta$, $\theta^n_{eq}$ is the average value of the angle on a sampling period $\Delta$, 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 $\Delta$.

### 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 a sampling period $\Delta$. 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 is to be synchronized with the PWM period $\Delta$, 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, \( i_{n+1}^\alpha, i_{n+1}^\beta \) (made at the beginning of the period) and past ones \( i_n^\alpha, i_n^\beta \), during the \((n+1)\)-th period. Also, the values of voltages, which were applied to stator windings \( u_n^\alpha, u_n^\beta \), are known. First estimation of the angle \( \vartheta_{eq}^n \) can be solved by using equations (5) and (6):

\[
\vartheta_{eq}^n = -\arctan \frac{i_{n+1}^\alpha - \frac{\Delta}{L} (u_n^\alpha - ri_n^\alpha)}{i_{n+1}^\beta - \frac{\Delta}{L} (u_n^\beta - ri_n^\beta)}
\]

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) - (9) (the observer variables have upper indices *):

\[
\theta^{n+1*} = \theta'^n + \Delta \theta^n + \frac{\Delta}{2J} (T_{eq}^n - T_L^n) + l_i (\theta'^n - \theta_{eq}^n)
\]

(11)

\[
\omega^{n+1*} = \omega'^n + \frac{\Delta}{J} (T_{eq}^n - T_L^n) + l_i (\omega'^n - \omega_{eq}^n)
\]

(12)

\[
T_L^{n+1*} = T_L^n + l_i (\theta'^n - \theta_{eq}^n)
\]

(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 \( \lambda_1 = \lambda_2 = \lambda_3 \) in the characteristic equations have been used:

\[
-\lambda + \lambda^2 (\lambda_1 + \lambda_2 + \lambda_3) - \lambda (\lambda_2 \lambda_3 + \lambda_1 \lambda_3 + \lambda_1 \lambda_2) + \lambda_1 \lambda_2 \lambda_3 = 0
\]

(13)

\[
-\lambda^2 + \lambda^3 (3 + l_i) - \lambda (3 + 2l_i + \frac{l_i \Delta^2}{2J} - l_2 \Delta) + (1 + l_i - l_2 \Delta - \frac{l_i \Delta^2}{2J}) = 0
\]

(14)

4. Digital 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:

\[
-\frac{T_{max} + T_L}{J} = E_{min} < E < E_{max} = \frac{T_{max} - T_L}{J}
\]

(16)

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

\[
\left| \Omega \right| < \Omega_{max}
\]

(17)

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

Ability to realise the reference under dynamic limitations in case the reference cannot be realised.
Absence of a dynamic error by the submitting the realised 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 in- 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 $\Theta^n_z$. The digital equations of the rate limiter can be written:

$$ (\mathcal{D})_{\zeta}^{n+1} = (\mathcal{D})_{\zeta}^n + (\mathcal{D})_{\zeta}^n \Delta + \frac{(\mathcal{E})_{\zeta}^n \Delta^2}{2} + \frac{v^n \Delta^3}{6} \quad (18) $$

$$ (\mathcal{D})_{\zeta}^{n+1} = (\mathcal{D})_{\zeta}^n + (\mathcal{D})_{\zeta}^n \Delta + \frac{v^n \Delta^2}{2} \quad (19) $$

$$ (\mathcal{E})_{\zeta}^{n+1} = (\mathcal{E})_{\zeta}^n + v^n \Delta \quad (20) $$

where $(\mathcal{D})_{\zeta}^n = \Theta^n_{lim} - \Theta^n_{\zeta}$, $(\mathcal{D})_{\zeta}^n = \Omega^n_{lim} - \Omega^n_{\zeta}$, $(\mathcal{E})_{\zeta}^n = E^n_{lim} - E^n_{\zeta}$, $v$ is the rate limiter control.

It must ensure 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. 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]:

$$ S^n_{1+1} = E^n_{lim} - E^n_{max} = 0 \quad (21) $$

$$ S^n_{2+1} = E^n_{lim} - E^n_{min} = 0 \quad (22) $$

$$ S^n_{3+1} = c(\Omega^n_{lim} - \Omega^n_{max}) + E^n_{lim} = 0 \quad (23) $$

$$ S^n_{4+1} = c(\Omega^n_{lim} + \Omega^n_{max}) + E^n_{lim} = 0 \quad (24) $$

$$ S^n_{5+1} = (\mathcal{E})_{\zeta}^{n+1} + b_1 (\mathcal{D})_{\zeta}^{n+1} + b_2 (\mathcal{D})_{\zeta}^{n+1} = 0 \quad (25) $$

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 $S_j = 0, (j = 1, ..., 5)$ is defined by a choice of coefficients $b_1, b_2, c$. Additionally, the limiter control $v^n$ can be bounded by the minimal and maximal values:

$$ -v^n_{max} < v^n < v^n_{max} \quad (26) $$

The values of $v^n$ that are necessary for performance of a condition, (21) - (25) on the $(n+1)-th$ period have been calculated (Table I).
Table 1. Values of the limiter control.

| j | \( v^n_j \) |
|---|---|
| 1 | \( \frac{E_{\text{lim}}^n - E_{\max}}{T} \) |
| 2 | \( \frac{E_{\text{lim}}^n - E_{\min}}{T} \) |
| 3 | \( -\frac{c(\Omega_{\text{lim}}^n + E_{\text{lim}}^n T - \Omega_{\max})}{\Delta + \frac{b_1 \Delta^2}{2}} + \delta E_{\text{lim}}^n \) |
| 4 | \( -\frac{c(\Omega_{\text{lim}}^n + E_{\text{lim}}^n T + \Omega_{\max})}{\Delta + \frac{b_1 \Delta^2}{2}} + \delta E_{\text{lim}}^n \) |
| 5 | \( \frac{\delta E^n_e (1 + b_1 \Delta + \frac{b_2 \Delta^2}{2}) + \delta \Omega^n_e (b_1 + b_2 \Delta) + \delta \theta^n b_2}{\Delta + \frac{b_1 \Delta^2}{2} + \frac{b_2 \Delta^3}{6}} \) |

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:

\[
v^n = \gamma \min \left\{ |v^n_1|, |v^n_2|, |v^n_3|, |v^n_4|, |v^n_5|, |v_{\max}^n| \right\}
\]

where \( \gamma \) is a sign of limiter control with the least absolute value.

5. Discrete-time control algorithm

The control of a VSI fed motor is designed by a block principle and realizes direct digital vector control of motor with use of discrete model of processes, independent variable estimation, and reference rate limitation. The structure of a drive with control is shown in figure 1.

The input of measurement and task signals is carried out once on each calculation period of vector control, which is equal to multiple PWM periods. At the beginning of the calculation, periods \([n, n+1]\) are known, i.e.:

- Actual values of current components \( i_{d}^{n+1} \) and \( i_{q}^{n+1} \), which are calculated using the phase currents (iA, iB, iC) and the estimated rotor angle \( \theta_{eq}^n \) or \( \theta^{n*} \).
- Actual values of the voltage components \( u_{d}^{n+1} \) and \( u_{q}^{n+1} \), which were calculated on the previous calculation period and are used in this one.
- Estimated value of the rotation speed \( \theta_{eq}^{n*} \).
- Limiter output signals, which are the reference signals for control.

Control output signals are the control signals for the VSI power keys on the PWM period.
Figure 1. Structure of a digital EPMSM control. The control himself has two blocks: the calculator of the reference value of the torque and the current control block.

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

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

(27)

where the value of the coefficients \(a_0, a_1, a_2\) are chosen from the needed value of the mode of the characteristic equation:

\[
\det \begin{vmatrix} 1 + \frac{\Delta^2}{2}a_2 - \lambda & \frac{\Delta^2}{2}a_i & \frac{\Delta^2}{2}a_0 \\ \frac{\Delta a_2}{\Delta a_i} & 1 + \Delta a_i - \lambda & \frac{\Delta a_0}{\Delta a_i} \\ \frac{\Delta a_2}{a_i} & \frac{\Delta a_0}{a_i} & 1 / \lambda \end{vmatrix} = 0
\]  

(28)

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

\[
S = \delta \Omega + \alpha (d \delta \Omega / dt)
\]  

(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 \(u_q^{n+1}\).
\[ u_{d}^{n+1} = -(L / \Delta)(i_{d}^{n+2} - i_{d}^{n+1}) + r_{d}^{n+1} - L \omega_{d}^{n+1} \frac{q}{q} \]
\[ u_{q}^{n+1} = -(L / \Delta)(i_{q}^{n+2} - i_{q}^{n+1}) + r_{q}^{n+1} + L \omega_{q}^{n+1} i_{d}^{n+1} + \Psi_{d}^{n+1} \]

(30)

(31)

where the current component \( i_{d}^{n+2}, i_{q}^{n+2} \) on the calculation period \([k + 2, k + 3]\) are \( i_{d}^{n+2} = 0, i_{q}^{n+2} = T_{eq}^{n+2} \frac{q}{j} \).

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 \( u_{ax}, u_{ay} \) in stationary coordinate system for PWM control of VSI. They could be found by using estimated value of the rotor position \( \theta \).

6. Simulation results

The above mentioned algorithm was simulated using MatLab 6.5.0 and Simulink 5.0. The motor used has the following parameters: \( R = 0.038 \Omega, L = 0.24 mH, X = 1.5 \Omega, U_{\phi} = 220 V, f = 50 Hz, \)
\( J = 0.01 kg \cdot m^2, p=2, \Psi_{f} = 0.06 N \cdot m \cdot A, P = 50 kW \). 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 one, 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, rad, 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 8 Nm 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 \( a_{0} = 0.9412, a_{1} = -17.73, a_{2} = -1782 \).

On given below figures horizontal scale is time one (sec), vertical scale is the control variable or estimated one. In figures 2 - 7 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. The estimation of the rotation speed is depicted in figures 3 - 5. 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 4 - 5 and is very small. The estimation of the rotor angle is depicted in figures 6 and 7. 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 equestrian (10). The higher one is a result of the angle estimation by using of the observer (11) - (13). The observer has no error in practice.
7. Conclusions
The algorithms of the discrete-time observer and reference limiter, and the sensorless electromechanical system based on the synchronous motor and VSI were developed. They ensure 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 simulation carried out for the control plant has confirmed serviceability of the algorithms mentioned.
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