Speed Sensorless Sliding Mode Torque Control of Induction Motor

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Abstract

Novel induction motor control optimizing both torque response and efficiency is proposed in the paper. The main contribution of the paper is a new structure of rotor flux observer aimed for the speed-sensorless operation of induction machine servo-drive at both, low and high speed, where rapid speed changes can occur. The control differs from the conventional field-oriented control. Stator and rotor flux in stator fixed coordinates are controlled instead of the stator current components in rotor field coordinates $i_{sd}$ and $i_{sq}$. In principle the proposed method is based on driving the stator flux towards the reference stator flux vector defined by the input command, which are the reference torque and the reference rotor flux. The magnitude and orientation angle of the rotor flux of the induction motor are determined by the output of the closed-loop rotor flux observer based on sliding mode control and Lyapunov theory. Simulations and experimental tests are provided to evaluate the consistency and performance of the proposed control technique.

Keywords

Induction motors, Motor drives, Torque control, Observers

I. Introduction

Due to its rigidness, maintenance free operation and many other advantages, the squirrel cage induction motor is the most widely used electrical motor in industrial applications. However, due to its highly coupled nonlinear structure it has certain limitations in sensorless applications in servodrives. This is due to the fact that in currently used control schemes like field oriented control [1], [2] or direct torque control [3], errors occur due to the parameters variation, integral drift and noise.

The traditional Field Oriented Control (FOC) technique ([1], [2], [4]) represents the attempt to reproduce, for an induction motor (IM), a dynamical behaviour similar to that of the dc machine, characterized by the fact that developed torque is proportional to the modulus of the stator current: to reach this objective, it is necessary to keep the rotor flux value constantly equal to the nominal value, so that, contemporarily, the optimal magnetic circuits exploitation guarantees the maximum power efficiency. Even though FOC is largely adopted, the asymptotical decoupling between speed and flux control presents a problem in the currently employed techniques, since when a field weakening is required to exceed the speed nominal value, flux transients have effects on the speed control, too. Another aspect that makes the control of IM quite complex is the rotor resistance variation during motor operation, due to the heating of rotor. This parameter may strongly change, assuming values up to twice the nominal one. But even with control algorithms, which guarantee good robustness properties against rotor resistance variations, flux observers are, in general, very sensitive to imperfect system parameters knowledge. As a consequence, to allow a correct flux estimation, and consequently a proper operation of the control, it is required to estimate the rotor resistance value.

Fifteen years after the appearance of FOC, another technique to control the torque in IM was developed and presented: the Direct Torque Control (DTC) [3]. With the use of DTC it is possible to obtain good dynamic control of the torque without any mechanical transducers on the machine shaft. Thus, DTC can be considered a "sensorless type" control technique. The name DTC is derived from the fact that, on the basis of the errors between the reference and the estimated values of the torque
and flux, it is possible to directly control the inverter states in order to reduce the torque and flux errors within the prefixed band limits.

Unlike FOC, DTC does not require any current regulator, coordinate transformation and PWM signals generators. In spite of its simplicity, DTC allows good torque control in steady-state and transient operating conditions. In addition, this controller is less sensible to the parameter detuning in comparison with FOC. On the other hand, it is well known that DTC presents some disadvantages that can be summarized as follows: difficulty to control torque and flux at very low speed, high current and torque ripple, variable switching frequency behaviour, high noise level at low speed and lack of direct current control.

The aim of this paper is to show use of the advantages found in both FOC and DTC so as to develop a new nonlinear robust speed sensorless control scheme for IM. The idea was to apply the continuous sliding mode control design technique ([5], [6]) to a voltage-fed IM control scheme, where the stator flux is assumed to be the control input. A unified control scheme for the stator flux and torque control based on the discrete time chattering free continuous torque control from an IM drive was developed. In this way the chattering of control input was eliminated and the excitation of the power drive without high frequency oscillation was achieved. The continuous sliding mode design possessing the robustness to the matched and unmatched uncertainties will be employed in the stator flux controller as in the rotor flux observer.

Controlled IM drives without speed sensor extract information on the mechanical shaft speed from measured stator voltages and currents at the motor terminals. The majority of speed identification methods rely on the approximated fundamental model of the machine. The use of the stator equation, particularly the integration of the stator voltage vector, is common for all methods. Its solution is fairly accurate when the switched stator voltage waveform is measured at high bandwidth, and when the parameters that determine the contributions of the resistive and the leakage voltage components are well known. The influence of these parameters dominates the estimation at lower speed, the steady-state accuracy of speed sensorless operation tends to be poor in the lower speed range. The dynamic performance depends on the accuracy of the field angle estimation, which is also parameter dependent [7].

Using the findings in [8], [9], [10] a closed-loop observer topology is presented with a simple design methodology that allows precise control of estimated rotor and stator flux accuracy attributes. The observer is built of the fundamental stator voltage equation of IM. The continuous sliding mode rotor flux compensator together with the stator voltage equation constitutes the structure of closed-loop observer. The sliding mode rotor flux controller exhibits two structures, i.e. I-PD and PD as feedback action of the observer. The integral part of the controller is active via a non-linear gain function whose gain value depends strictly on stator frequency. The I-PD structure is active at zero and low speed and will change its structure to a PD structure at moderate and high speed. The I-channel of the controller will replace the information of stator voltage at zero speed. Furthermore it will allow the motor to drive over the zero stator frequency, where the steady-state accuracy of speed sensorless operation tends to be poor.

The observed values of rotor and stator flux converge towards the actual values. Convergence is influenced by the stator resistance variations resulting in the need for stator resistance adaptation algorithm. Both, convergence proof and stator resistance algorithm are provided.

The paper is organized as follows: in Section II the IM model, sliding mode theory and design of the sliding mode stator flux controller are presented. Section III is devoted to the rotor flux observer. Section IV presents the behaviour of IM in a flux weakening region. Finally, in Section V simulation and experimental results are presented. Section VI gives some conclusion remarks on the new speed sensorless control of IM.
II. Machine Dynamics and Proposed Sliding Mode Control

A. Induction Motor Model

The main purpose of this paper is to apply the nonlinear continuous sliding-mode control combined with Lyapunov design approach to the torque and flux control of IM so that the following is achieved:

- the generated torque becomes a continuous output with respect of control states;
- regulating the rotor flux amplitude can increase the power efficiency and make it possible to operate in the flux-weakening region;
- employing the nonlinear continuous sliding-mode control on the torque and flux possesses the robustness to the matched and mismatched uncertainties.

Under the assumptions of linearity and symmetry of electric and magnetic circuits and neglecting iron losses, the dynamic model of the IM in the fixed stator reference frame will be described with state variables stator flux $\psi_s^s$ and rotor flux $\psi_r^s$ as

$$\frac{d\psi_s^s}{dt} = -\frac{R_s}{\sigma L_s} \psi_s^s + \frac{R_s L_m}{\sigma L_s L_r} \psi_r^s + u_s^s,$$

(1)

$$\frac{d\psi_r^s}{dt} = -\frac{R_s}{\sigma L_s} \psi_r^s + \frac{R_r L_m}{\sigma L_s L_r} \psi_s^s - j p \omega_r \psi_r^s,$$

(2)

where $\psi_s^s = [\psi_{sa}^s, \psi_{sb}^s]$, $\psi_r^s = [\psi_{ra}^s, \psi_{rb}^s]$, $u_s^s = [u_{sa}^s, u_{sb}^s]$ denote stator flux, rotor flux linkage and stator terminal voltage, respectively. $R_s$ and $R_r$ stand for stator and rotor resistance, $\omega_r$ denotes the rotor electrical speed, $L_m$ is the mutual inductance, and $\sigma L_s$ is the redefined leakage inductance ($\sigma L_s = L_s - \frac{L_r^2}{L_r}$). Stator current $i_s^s = [i_{sa}^s, i_{sb}^s]$ of IM is expressed by (3):

$$i_s^s = \frac{1}{\sigma L_s} \left( \psi_s^s - \frac{L_m}{L_r} \psi_r^s \right).$$

The generated torque, which is the control output of the control plant (IM), can be expressed in terms of stator and rotor flux linkage as

$$T_e = \frac{2}{3} p \frac{L_m}{\sigma L_s L_r} (\psi_r^s \times \psi_s^s) = \frac{2}{3} p \frac{L_m}{\sigma L_s L_r} |\psi_r^s||\psi_s^s| \sin \delta,$$

(4)

where $p$ is the number of pole pairs.

The mechanical dynamic equation is given by

$$J \frac{d\omega_r}{dt} = T_e - T_L,$$

(5)

where $J$ denotes the moment of inertia of the motor and $T_L$ a load torque.

The signal flow diagram (Fig.1) presents the graphic interpretation of the dynamic model of IM. This graph exhibits two fundamental structures, one on the upper portion on the left hand side which represents the stator winding, and one on the lower portion representing the rotor winding. Each winding is characterized by a first order delay element with corresponding time constant $\sigma L/R$. In the diagram two nonlinearities appear. The first one describes the cross-coupling between orthogonal space vector components stator current and rotor flux of the active torque, and the second one shows the cross-coupling between the rotor flux and speed of IM in the internal feedback path of the rotor winding. The mechanical equation of the drive is represented by an integrator characterized with drive inertia $J$. 

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B. Sliding Mode Theory

In the last decade, a wide range of non-linear methods for feedback control, state estimation and parameter identification have emerged. Among them the sliding mode control gained wide acceptance due to the use of straightforward fixed non-linear feedback control functions, which operate effectively over a specified magnitude range of system parameter variations and disturbances. The essential property of sliding mode control is that the discontinuous feedback control switches on one or more manifolds in the state space. Ideally, the switching of control occurs at infinitely high frequency to eliminate the deviations from sliding manifolds. In the practice, the switching frequency is not infinitely high due to the finite switching time and with effects of unmodeled dynamics causes undesired chattering of the control. This is a well-known problem and is widely covered in literature. The design of the control system will be demonstrated for a nonlinear system presented in the canonical form [5], [6]

\[
\frac{dx}{dt} = f(x, t) + B(x, t)u(x, t) + d(x, t),
\]
\[
x \in \mathbb{R}^n, \quad u \in \mathbb{R}^m, \quad \text{rank}(B(x, t)) = m.
\]

In sliding mode control, the goal is to keep the system motion on the manifold \( S \), which is defined as:

\[
S = \{ x : \sigma(x, t) = 0 \},
\]

\[
\sigma = x^d - x.
\]

Control input \( u \) has to grant that the motion of the system described in (6) is restricted to belong to the manifold \( S \) in the state space.

The "chattering free" sliding mode control should be chosen such that the candidate Lyapunov function satisfies the Lyapunov stability criteria:

\[
V = \frac{1}{2} \sigma^T \sigma,
\]

\[
\dot{V} = \sigma^T \dot{\sigma}.
\]

This can be assured for:

\[
\dot{V} = -\sigma^T D \sigma,
\]

where \( D \) is the positive definite matrix, then the system (6) will be stable.

Therefore (9), (11) satisfy the Lyapunov conditions. From (10) and (11) we get the reaching condition \( \dot{\sigma} = -D \sigma \). With the selected Lyapunov function the stability of the whole control system in the case of the initial conditions and parameter mismatch is guaranteed. The control function will satisfy reaching conditions in the following form [5],[6]:

\[
u(t^+) = u(t^-) + (GB)^{-1} (D \sigma + \dot{\sigma}); \quad t = t^- + \Delta, \quad \Delta \to 0.
\]

The next step in controller design is discretization of (12) into the state equation on the assumption that the input vectors \( u(k) \) and \( u(k-1) \) are constant within small \( k \)-th and \( (k+1) \)-th sampling intervals.

In the discretized form, the continuous first derivative of the control error can be replaced by its first back step approximation. Thus the output of the controller is calculated as:

\[
u(k) = u(k-1) + (GBT_S)^{-1} ((I + TS)D)\sigma(k) + \sigma(k-1)),
\]

where \( I \) is the identity matrix and \( G \) is the feedback gain.
C. Continuous Discrete Time Stator Flux Controller

The motor stator flux controller is designed to achieve a real sliding mode motion on the manifold \( \sigma_s = 0 \) and at the same time provides a first order transient towards sliding surface.

\[
u_s^d(k) = u_s^d(k - 1) + \frac{K_{rs}}{T_S} ((I + T_S D) \sigma_s(k) + \sigma_s(k - 1)),
\]

where:

\[
\sigma_s = \psi_s^d - \hat{\psi}_s^s.
\]

The stator flux components in a rotor flux reference frame \( \psi_{sd} \) and \( \psi_{sq} \) are calculated from the torque \( T^d \) and rotor flux \( \psi_r^d \) commands:

\[
\psi_{sd} = \frac{L_s}{L_m} \left( \psi_r^d + \sigma \frac{L_r}{R_r} \frac{d\psi_r^d}{dt} \right),
\]

\[
\psi_{sq} = \frac{3}{2} \frac{\sigma L_s L_r T^d}{p L_m} \psi_r^s,
\]

and afterwards transformed into a stator reference frame. The transformation angle \( \theta^s_r \) is the phase angle of the rotor flux vector in a stator reference frame. Elements of the transformation matrix can be calculated directly from the estimated rotor flux values:

\[
\sin \hat{\theta}^s_r = \frac{\hat{\psi}^s_{rb}}{\left| \psi_r^s \right|} = \frac{\hat{\psi}^s_{rb}}{\sqrt{(\hat{\psi}^s_{ra})^2 + (\hat{\psi}^s_{rb})^2}},
\]

\[
\cos \hat{\theta}^s_r = \frac{\hat{\psi}^s_{ra}}{\left| \psi_r^s \right|} = \frac{\hat{\psi}^s_{ra}}{\sqrt{(\hat{\psi}^s_{ra})^2 + (\hat{\psi}^s_{rb})^2}}.
\]

Calculated reference voltage is continuous and regular PWM techniques may be used for switching converter control. For the relatively small sampling time \( T_S \) the magnitude of supply voltage can be assumed to remain constant, so the new value of the voltage vector can be predicted from the previous value by rotation in the plane in the following way:

\[
u_s^d(k + 1) = C(\Delta \theta^s_r) u_s^d(k) + \frac{K_{rs}}{T_S} ( (I + T_S D) C(\Delta \theta^s_r) \sigma_s(k) - C(2 \Delta \theta^s_r) \sigma_s(k - 1) ),
\]

The output of the controller is then calculated with:

\[
u_s^d(k + 1) = C(\Delta \theta^s_r) u_s^d(k) + \frac{K_{rs}}{T_S} ( (I + T_S D) C(\Delta \theta^s_r) \sigma_s(k) - C(2 \Delta \theta^s_r) \sigma_s(k - 1) ) ,
\]

The elements of the transformation matrix \( C(\Delta \theta^s_r) \) can be calculated from successive values of \( \sin(\Delta \theta^s_r) \) and \( \cos(\Delta \theta^s_r) \). The system is asymptotically stable and theoretically will reach the sliding mode manifold \( S \) in infinite time, but \( \varepsilon = o(\Delta) \)-vicinity of the manifold is reached in finite time. The matrices \( C(\Delta \theta^s_r) \), \( C(2 \Delta \theta^s_r) \) are used in algorithm for compensation of computational delay, namely the stator flux errors \( \sigma_s(k), \sigma_s(k - 1) \) are measured and computed from the observed and reference stator flux in previous sampling time one step \( (\Delta \theta^s_r) \) and two steps \( (2 \Delta \theta^s_r) \) backwards. The proposed control scheme is presented in Fig. 2.
The ac phase voltages of the PWM inverter are uniquely determined from $u_s^d(k+1)$ using the simple coordinate transformation:

$$u_{1,2,3} = K_{ab}^{1,2,3} u_s^d, \quad K_{ab}^{1,2,3} = \begin{bmatrix} 1 & 0 \\ -1/2 & -\sqrt{3}/2 \\ -1/2 & \sqrt{3}/2 \end{bmatrix}.$$  

(22)

From sinusoidal PWM, which constitutes a suitable sequence of active and zero inverter output vectors $u_k$:

$$u_k = \begin{cases} 2/3 U_{de} e^{j(k-1)\pi/3}, & \text{for } k = 1, \ldots, 6 \\ 0, & \text{for } k = 0, 7 \end{cases}.$$  

(23)

The stator flux moves along a track resembling a circle. The rotor flux, however, rotates continuously with actual synchronous speed along a near-circular path, because its components are sinusoidal.

III. NON-LINEAR FLUX OBSERVER

A. Closed Loop Flux Observer

Speed and flux observers for sensorless IM drives are a huge research area. There are many algorithms, but very few comparative studies ([7], [8], [9], [10]). In the excellent review paper [14] a summary of claimed performance is presented, the paper does not attempt to verify these claims. In [7] the authors carry out an independent review of EKF, Luenberger observer and MRAS approaches and the tests on them. Further more, since used stator flux controllers are constructed on the rotor flux frame, the characteristics of the controllers are dominated by the error in the rotor flux estimation. Therefore, to incorporate the observer into the controller, it is of importance to evaluate the error of the observer caused by the disturbance.

Speed sensorless observers are based on the fundamental equation of IM, which is widely known as a voltage model of the rotor flux [7], [11]:

$$\frac{d\hat{\psi}_s^r}{dt} = \frac{\hat{L}_r}{\hat{L}_m} \left( u_s^s - \hat{R}_s \hat{u}_s^s - \hat{\sigma} \hat{L}_s \frac{di_s^s}{dt} \right),$$  

(24)

where `\` denotes estimated values.

However, this scheme has some disadvantages. It is not robust to the parameter’s variations and initial conditions of the system. This problem is especially important at low speeds, and since at startup the motor has to operate at low speed, problems occur. Therefore improved schemes have to be used.

In our work a scheme using non-linear feedback compensator $C_r(s)$ and stator resistance adaptation algorithm is used (Fig. 2).

The idea of the proposed scheme is to use the classical voltage model (24) and then improve its performance with the compensation of influences caused by parameter uncertainties and variations. To achieve a low pass filtering action, a rotor flux controller (compensator) is used.

If the loop is closed around the rotor flux calculation, the rotor and stator fluxes will converge to a reference value even at zero speed. The estimated value used for zero and low speed convergence is obtained from the second order closed-loop rotor flux observer. This is similar to the topology in [6] but the PI controller is replaced with the sliding mode rotor flux controller and a speed dependent gain function $K(\omega_s)$, which makes the flux observer non-linear from stator frequency.

The output value of the rotor flux sliding mode controller is expressed with:

$$v_r(k) = K(\hat{\omega}_s) v_r(k - 1) + K_{rr} (D_r \sigma_r(k) + \dot{\sigma}_r(k)), \quad (25)$$

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which in discrete time form can be written as:

\[
v_r(k) = K(\dot{\omega}_s)C(\Delta \theta^*_r)\nu_r(k - 1) + \frac{K_{rr}}{T_s} \left( (I + T_s D_r)C(\Delta \theta^*_r)\sigma_r(k) - C(2\Delta \theta^*_r)\sigma_r(k - 1) \right), \tag{26}
\]

where

\[
\sigma_r(k) = \psi^*_d(k) - \dot{\psi}^*_r(k). \tag{27}
\]

The compensator structure is changed during the operation depending on the observed value of stator frequency \(\dot{\omega}_s\). At moderate and zero stator frequencies I-PD structure is used, whereas at medium and high frequencies it transfers to PD structure by deactivating the integral part. Thus the compensator algorithm passes over from recursive to non-recursive structure when stator frequency increases from low to higher values. The second order structure of the observer passes to the first order structure.

The transfer from I-PD to PD structure is performed with the use of non-linear gain \(K(\omega_s)\). The choice of the type of function that will represent \(K(\omega_s)\) is strictly dependent on when and how the transition will take place. During low speed \(K(\omega_s)\) should have a positive value and for higher rotor speeds \(K(\omega_s)\) is to roll off to zero. This means that for high speed the integral action of the compensator will be deactivated. The polynomial of the third order is used as the transfer function in order to enable smooth transfer.

Speed dependent gain function \(K(\omega_s)\) described with \(28\) dynamically changes observer bandwidth which will allow complete control and give a crisp function from low pass filter of second order at zero and low speed to the low pass filtering action of first order at moderate and high speed of IM (Fig. 3).

\[
K(\omega_s) = \begin{cases} 
1; & \omega_s \leq \omega_{p1} \\
 a_3\omega^3_s + a_2\omega^2_s + a_1\omega_s + a_0; & \omega_{p1} < \omega_s < \omega_{p2} \\
 0; & \omega_s \geq \omega_{p2} 
\end{cases} \tag{28}
\]

Fig. 3 shows the transfer function of the closed loop rotor flux observer. Resulting frequency response consists of two parts: influence of command rotor flux \((\dot{\psi}^*_r/\dot{\psi}^*_d)\) and influence of actual rotor flux of IM \((\dot{\psi}^*_r/\psi^*_r)\). For low frequencies including zero \((\omega_s < \omega_{p1})\) the observer suppresses the influence of actual rotor flux part as a high-pass filter. This is required, because the information obtained from the stator voltage and current are incorrect. The output \(\dot{\psi}^*_r\) is mainly determined by command input \(\dot{\psi}^*_d\). For moderate frequencies \((\omega_{p1} < \omega_s < \omega_{p2})\) the transition takes place with the change of nonlinear gain \(K(\omega_s)\) and both inputs are active. For medium and high frequencies \((\omega_{p2} < \omega_s < \omega_{max})\) the influence of command rotor flux \(\dot{\psi}^*_d\) will be suppressed like if the low-pass filter would be used and estimated value of rotor flux \(\dot{\psi}^*_r\) will be determined mainly by rotor flux of IM \(\psi^*_r\). Thus for all the working frequencies \((0 \leq \omega_s < \omega_{max})\) the gain and phase of the closed-loop observer transfer function remain constant.

B. Speed Estimation

An obvious way to obtain the value of stator frequency would be to use the \(\text{atan2}\) (four quadrant arc tangens) function for the calculation of stator flux angle and then derivate the result:

\[
\dot{\omega}_s = \frac{d}{dt}(\dot{\theta}^*_s) = \frac{d}{dt}\left(\text{atan2}(\dot{\psi}^*_r/\psi^*_r)\right). \tag{29}
\]

If derivation is performed analytically, we get the following result:

\[
\dot{\omega}_s = \frac{\dot{\psi}^*_r \dot{\psi}^*_r - \dot{\psi}^*_r \dot{\psi}^*_r}{(\psi^*_r)^2 + (\psi^*_r)^2}, \tag{30}
\]
which using the equations (3) and (24) giving a well known basic equation:

\[
\dot{\psi}_s = \frac{d\dot{\psi}_s}{dt} = u_s - \hat{R}_s i_s
\]  \hspace{1cm} (31)

can be rewritten as:

\[
\dot{\omega}_s = \frac{\left( u_{sa} - \hat{R}_s i_{sa} \right) \dot{\psi}_{sb} - \left( u_{sb} - \hat{R}_s i_{sb} \right) \dot{\psi}_{sa}}{\left( \dot{\psi}_{sa} \right)^2 + \left( \dot{\psi}_{sb} \right)^2}.
\]  \hspace{1cm} (32)

The estimated value of rotor speed can be calculated from the equation below:

\[
\hat{\omega}_s = \frac{\dot{\omega}_s}{p} - \hat{\omega}_{sl},
\]  \hspace{1cm} (33)

where slip frequency estimation is directly obtained from reference values:

\[
\hat{\omega}_{sl} = \frac{3 R_r T^d}{2 p} |\psi_r^d|^2.
\]  \hspace{1cm} (34)

C. Convergence Proof and Stator Resistance Adaptation

Further more, since the used stator flux controllers are constructed on the rotor flux frame, the characteristics of the controllers are dominated by error in the flux estimator. Therefore, to incorporate the rotor flux observer into the controller, it is of great importance to evaluate the error of the observer caused by the load of IM.

The error differential equation between actual stator flux \(\psi_s\) in the IM and estimated stator flux \(\hat{\psi}_s\) used in the scheme is expressed with:

\[
\frac{d}{dt} \left( \psi_s - \hat{\psi}_s \right) = -(R_s - \hat{R}_s) i_s - C_r(s) (\psi^s_r - \hat{\psi}_r).
\]  \hspace{1cm} (35)

The convergence of (35) depends mainly on the difference between the estimated value of stator resistance \(\hat{R}_s\) and the actual stator resistance \(R_s\) can be influenced by the error between reference \(\psi^d_r\) and estimated \(\hat{\psi}^d_r\) rotor flux. The stator resistance change very slowly with the load and temperature of the machine, so we can use (35) to adopt the stator resistance \(\hat{R}_s\). By equating the left hand side of (35) to zero we get:

\[
\Delta R_s = R_s - \hat{R}_s = -C_r(s) (\psi^s_r - \hat{\psi}_r) (i_s)^{-1}.
\]  \hspace{1cm} (36)

The estimated value of stator resistance \(\hat{R}_s\) can be tuned with a combination of (36) and the output of (26):

\[
\hat{R}_s(k) \approx \hat{R}_s(k - 1) - k_{Rs} \frac{v_{rb}(k)i_{sa}(k) - v_{ra}(k)i_{sb}(k)}{|i_s|^2}.
\]  \hspace{1cm} (37)

IV. Flux Control for Efficiency Improvement

Torque control is skillfully obtained by controlling the instantaneous slip frequency, whereas improvement of the efficiency can be achieved by controlling the amplitude of \(\psi^s_r\) like a field-weakening operation. In the case of frequent changes of the torque command, it is necessary to keep the flux amplitude at maximum. Under the field-weakening state, the sufficient torque or quick torque response would not be expected because of the slow response of the flux increase. On the other hand, in the steady-state operation, especially at light loads, the maximum efficiency can be obtained at a low flux level. Therefore, in order to obtain the maximum efficiency the flux level is adjusted in accordance with the torque command. The maximal allowed rotor flux will depend on actual stator frequency and dc-link voltage with:

\[
\psi^d_{r,\text{max}} = \frac{U_{dc} \sqrt{3} - |i_s| \hat{R}_s}{\omega_s},
\]  \hspace{1cm} (38)
and torque dependent efficiency optimal rotor flux will be [13]:

$$
\psi_{r \text{ opt}}^d = \frac{T_d}{p} \sqrt{\frac{R_s + R_r}{R_s + \frac{L_d^2}{R_{gl}s + R_r \omega_s^2}}}.
$$

(39)

$R_{gl}s$ is the resistance of iron losses and should be determined by experimental identification. The actual rotor flux command will be used:

$$
\psi_{r \text{ opt}}^d \leq \psi_{r \text{ max}}^d.
$$

(40)

A. Simulation Results

The control system for induction motor was simulated as described below, modeled in stator reference frame. The motor is voltage fed. Ideal voltage and current are assumed. Estimated stator resistance deviates from the actual value ($\hat{R}_s = 0.8R_s$) and adaptation algorithm was used.

Fig. 4 features reference, actual and load torque, Fig. 5 shows reference, actual and estimated rotor flux, and Fig. 6 shows actual and estimated mechanical speed together with estimated stator frequency. As shown in the presented figures, the applied algorithm is robust to the parameter variations. Rotor and stator flux will track their reference values very closely even at fast changes of reference values and load. Fig. 5 shows that the value of rotor flux can be changed without causing problems to the control and observer schemes. Thus the reference value of rotor flux can be chosen with respect to the available stator voltage and improvement of efficiency. Rotor and stator flux will track their reference values very closely even at fast changes of command values (torque and rotor flux) and load.

B. Experimental Results

Experimental results were acquired with the use of Texas Instruments’ DSP TMS320C31 on IM motor as described below. The PWM inverter was built with an IGBT module produced by Semikron. The algorithm was calculated every 160$\mu$s.

Brushless AC servomotor of type 142UMD400CACAA produced by UNIMOTOR driven by 7.5kW Digital AC Drive UNI2402 was used as the Load Machine. The control of both, speed and applied torque is possible, thus by physical linking of Drive and Load Machine’s axes Hardware-in-the-loop operation can be performed. By changing the applied torque of the Load Machine various types of loads can be emulated. Since the torque is set by the same program that is running the Drive Machine, graphical results can be obtained from the PC.

For the operation at low and zero speed the reference value of the Load Machine speed is set to zero and speed controller parameters are not optimized allowing coupled axes to change speed in a reasonably small vicinity of the zero speed. The applied torque of the Load Machine is measured and thus the value of the torque applied by the Drive Machine is obtained.

Experiments were performed at both, low and high speed thereby describing the behavior of the control system at low and medium stator frequencies. The reference and observed value of torque are featured in Fig. 7 for locked rotor and in Fig. 9 for rotating rotor. In Fig. 7 the actual torque measured from the applied current of the load machine is presented, whereas in Fig. 9 transient of the load torque applied by the load machine is included. Fig. 8 shows actual rotor speed and the estimated stator frequency for locked rotor and Fig. 10 shows those values for rotating rotor.

VI. Conclusion

Sensorless torque and flux control of IM is an emerging new technology, though in the early state of development. An algorithm for torque and flux tracking control is presented in the paper. The control algorithm utilizes the stator flux as control input and the rotor flux is selected in accordance
with torque demand of IM to achieve the efficiency optimized drive performance. The proposed scheme allows a smooth transition into the field weakening region and the full utilization of the inverter voltage and current capability. Continuous sliding mode controller without chattering of control input are used in the stator flux controller and in the new rotor flux observer. Non-linear stator frequency dependent gain was used in the rotor flux observer which enables the operation at zero, low and high speed. The adaptation algorithm for stator resistance is included. The performance of the proposed algorithm was verified by simulation and experiments.

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