Induction Motor DTC Strategy Using Discrete-Time Sliding Mode Control

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Abstract— **A new algorithm for discrete-time sliding mode torque and flux control is proposed. The induction motor equations are written in a stator flux reference frame. The torque and flux controllers are shown to be robust to errors in the machine parameters. The controllers provide d and q reference voltages to be applied to the induction motor and no current controllers are present. However, differently from most sliding mode techniques, the calculated voltage vector serves as a reference for a vector PWM scheme and a fixed switching frequency is used. Simulation and experimental results are presented to show the strategy effectiveness.**

I. Introduction

Stator flux oriented induction motor drives [1], [2] have been thoroughly studied recently. In most schemes, torque and flux current commands can be calculated without rotor speed or position measurement.

In the first schemes proposed [1], the torque and flux current commands are obtained from the torque and flux errors and current regulators are used to determine the firing signals to the inverter.

Some authors have proposed methods to generate a dq voltage reference based directly on the torque and flux errors and PI controllers [2]. The current controllers are then avoided, but the use of linear controllers still limit the transient performance of these techniques.

Alternative strategies to avoid using current controllers, based on the so called Field Acceleration Method (FAM) have been proposed [3], [4]. Torque and flux errors are used as inputs to hysteresis controllers which determine the space voltage vector necessary to reduce both torque and flux errors. These techniques are usually called DTC (Direct Torque Control). They are robust and generally lead to very good transient performance. However, a variable switching frequency must be used to guarantee the errors to remain inside the hysteresis bands. If a fixed switching frequency is used, one voltage vector is chosen and applied during one entire switching interval. For the usual switching frequencies, the torque oscillations are then relatively high. Further, the strategies as originally proposed have problems for low speed operation.

Some deadbeat DTC strategies have been presented [5], [6]. These methods allow the excellent transient response to be maintained. The problem of using a variable switching frequency is solved and high torque and flux oscillations are avoided. However, the robustness is reduced, once the determination of the stator voltage vector to be applied for the deadbeat control requires the knowledge of many machine parameters.

A deadbeat stator flux oriented DTC strategy has been recently presented [7]. Writing the equations in a stator flux reference frame makes the torque and flux control algorithms very easy to implement, reducing the computational effort for this technique as compared to the previous deadbeat DTC strategies. However the robustness problem persist.

In this paper, a new stator flux oriented DTC strategy using discrete-time sliding mode controllers for both torque and flux control is proposed. The controllers are designed to be robust to errors in the machine parameters. As in any DTC strategy, the flux and torque controllers produce voltage commands and no current controllers are necessary. However, instead of applying one voltage vector during each sampling period, d and q voltage references are generated and a PWM algorithm is applied to produce the commanded voltage vectors. Thus, a fixed switching frequency is used and the torque oscillations are reduced.

II. Control Strategy

In a stator flux reference frame, the induction motor model is described by the equations below:

$$\vec{v}_s = R_s \vec{i}_s + \dot{\vec{\lambda}}_s + j\omega_e \vec{\lambda}_s \tag{1}$$

$$0 = R_r \vec{i}_r + \dot{\vec{\lambda}}_r + j\left(\omega_e - \omega_r\right)\vec{\lambda}_r \tag{2}$$

$$\vec{\lambda}_s = L_s \vec{i}_s + L_m \vec{i}_r \tag{3}$$

$$\vec{\lambda}_r = L_m \vec{i}_s + L_r \vec{i}_r \tag{4}$$

where $\vec{v}_s$ is the stator voltage vector, $\vec{i}_s$ and $\vec{i}_r$ are the stator and rotor current vectors, $\vec{\lambda}_s$ and $\vec{\lambda}_r$ are the stator and rotor flux vectors, L_s, L_r, and L_m are the stator, rotor and mutual inductances, R_s, R_r are the stator and rotor resistances, ω_e is the stator flux vector speed and ω_r is the rotor speed.

A. Flux Control

Since the equations are written in a stator flux reference frame, the q axis component of the stator flux is zero and

the flux dynamics can be imposed only by the d axis stator voltage component:

$$v_{sd} = R_s i_{sd} + \dot{\lambda}_{sd} \tag{5}$$

Defining a sliding surface based only on the error would not allow the imposition of the dynamics for the error correction. As an alternative, some authors define the sliding surface using the error and its integral.

$$S = x_2 + c_1 \bar{x}_1 \tag{6}$$

where x_2 is the error and $\bar{x}_1$ is the integral of the error.

With this choice, the sliding mode operation makes the system to have a first order dynamics:

$$S = \dot{S} = 0 \tag{7}$$

$$\Longrightarrow \dot{x}_2 + c_1 x_2 = 0 \tag{8}$$

However, the switching surface is reached only when the error is equal to $-c_1\bar{x}_1$. Choosing c_1 large enough to ensure a fast dynamic response would generally lead to a big overshoot.

In order to reduce the overshoot, a nonlinear sliding surface will be written in terms of two state variables defined as:

$$x_2 = \lambda_{sd}^* - \lambda_{sd} \tag{9}$$

$$x_1 = \int f(x_2)\, x_2 dt = \int e^{-p x_2^2} x_2 dt \tag{10}$$

It should be noted that $\lim_{|x_2| \longrightarrow \infty} f(x_2) = 0$. Thus, when the flux error is high the effect of the integral term is greatly reduced and high overshoots are not expected to occur. On the other hand, when the flux error is becoming small, the system tend to have a first order dynamics with a time constant imposed by the sliding surface parameter c_1, since

$$\lim_{x_2 \longrightarrow 0} f(x_2) = 1 \Longrightarrow \lim_{x_2 \longrightarrow 0} S(x) = x_2 + c_1 \int x_2 dt \tag{11}$$

For the sliding mode flux controller design, the d axis stator equation must be written in terms of the state variables x_1 and x_2 above. From (5), (9) and (10):

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{bmatrix} = \begin{bmatrix} 0 & f(x_2) \\ 0 & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} 0 \\ -1 \end{bmatrix} v_{sd} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} R_s i_{sd} \tag{12}$$

Assuming that teh sampling interval T is small, as compared to the flux time constant, Eqn. (12) can be written in a discrete-time form as:

$$\begin{bmatrix} x_{1,k+1} \\ x_{2,k+1} \end{bmatrix} = \begin{bmatrix} 1 & Tf(x_{2,k}) \\ 0 & 1 \end{bmatrix} \begin{bmatrix} x_{1,k} \\ x_{2,k} \end{bmatrix} + \begin{bmatrix} -\frac{T^2}{2} f(x_{2,k}) \\ -T \end{bmatrix} v_{sd,k} - \begin{bmatrix} -\frac{T^2}{2} f(x_{2,k}) \\ -T \end{bmatrix} R_s i_{sd,k} \tag{13}$$

or

$$x_{k+1} = F_k x_k + G_k v_{sd,k} - G_k R_s i_{sd,k} \tag{14}$$

The sliding surface is:

$$S_k = \begin{bmatrix} c_1 & 1 \end{bmatrix} \begin{bmatrix} x_{1,k} \\ x_{2,k} \end{bmatrix} = \tag{15}$$

The following control law is proposed:

$$u_k = v_{sd,k}^* = \frac{1}{\frac{c_1 T^2}{2} f(x_{2,k}) + T} \beta S_k \tag{16}$$

where

$$c_1 T f(x_{2,k}) < \beta < c_1 T f(x_{2,k}) + 2 \tag{17}$$

It can be easily proved that if control law (16) is applied, then the system representative point will move monotonically to a sliding surface neighborhood defined by the inequalities:

$$-c_1 x_{1,k} < x_{2,k} < -\frac{\beta c_1 x_{1,k} - c^t G_k R_s i_{sd,k}}{\beta - c_1 T f(x_{2,k})} \tag{18}$$

$$-c_1 x_{1,k} < x_{2,k} < -\frac{(2-\beta) c_1 x_{1,k} - c^t G_k R_s i_{sd,k}}{2 - \beta + c_1 T f(x_{2,k})} \tag{19}$$

$$-\frac{\beta c_1 x_{1,k} - c^t G_k R_s i_{sd,k}}{\beta - c_1 T f(x_{2,k})} < x_{2,k} < -c_1 x_{1,k} \tag{20}$$

$$-\frac{(2-\beta) c_1 x_{1,k} - c^t G_k R_s i_{sd,k}}{2 - \beta + c_1 T f(x_{2,k})} < x_{2,k} < -c_1 x_{1,k} \tag{21}$$

The system representative point will tend to remain in sliding movement on one border of the region defined by inequalities (18)-(21). Considering that a small sampling time is used, this region is narrow and a dynamic behavior very similar to the imposed by the sliding surface can be assumed to occur.

B. Torque Control

In a stator flux reference frame

$$T_e = \frac{3}{2}\frac{P}{2} \lambda_{sd} i_{sq} \tag{22}$$

and

$$v_{sq} = R_s i_{sq} + \omega_e \lambda_{sd} \tag{23}$$

Considering that the flux is conveniently controlled, the torque depends only on the q axis current. This current is strongly related to the stator flux vector speed as it can be

seen from the quadrature component of Eqn. (2), rewritten in terms of stator flux and current components:

$$\omega_e = \omega_r + \frac{\frac{L_s}{\tau_r} i_{sq} + \sigma L_s \dot{i}_{sq}}{\lambda_{sd} - \sigma L_s i_{sd}} \tag{24}$$

where τ_r is the rotor time constant and σ is the leakage coefficient.

Using equations (23) and (24), the following state equation can be written:

$$\dot{i}_{sq} = -\left(\frac{1}{\sigma\tau_r} + \frac{1}{\sigma\tau_s}\frac{\lambda_{sd} - \sigma L_s i_{sd}}{\lambda_{sd}}\right) i_{sq}$$

$$+\frac{\lambda_{sd} - \sigma L_s i_{sd}}{\lambda_{sd}}\frac{1}{\sigma L_s} v_{sq} - \omega_r \frac{\lambda_{sd} - \sigma L_s i_{sd}}{\sigma L_s} \tag{25}$$

If the stator flux has been previously established and is well controlled, it can be assumed that

$$\dot{\lambda}_{sd} \cong 0 \tag{26}$$

and

$$\dot{T}_e \cong \frac{3}{2}\frac{P}{2}\lambda_{sd}\dot{i}_{sq} \tag{27}$$

From (22), (25), (27) and considering that the sampling interval is small as compared to the time constants for i_{sd}, λ_{sd} and ω_r, the above state equation can be written in terms of the electromagnetic torque in a discrete-time form as:

$$T_{e,k+1} = A_k T_{e,k} - B_k v_{sq,k} + B_k \lambda_{sd,k}\omega_{r,k} \tag{28}$$

where

$$A_k = 1 - \left(\frac{1}{\sigma\tau_r} + \frac{1}{\sigma\tau_s}\frac{\lambda_{sd,k} - \sigma L_s i_{sd,k}}{\lambda_{sd,k}}\right) T \tag{29}$$

and

$$B_k = -\frac{3}{2}\frac{P}{2}\frac{\lambda_{sd,k} - \sigma L_s i_{sd,k}}{\sigma L_s} T \tag{30}$$

A_k and B_k must be calculated each sampling interval, since they depend on λ_{sd} and i_{sd}.

As in the sliding mode flux controller, it would be advantageous to write the sliding surface not only in terms of the error, but using also some nonlinear function of the error. By doing so, overshoots would be avoided and a first order dynamics for the correction of the error could be ensured. However, the system state equations should be written in terms of the variables used to define the sliding surface. The system state matrix would then contain terms dependent on the torque error. If the sampling interval were small enough, the torque could be considered constant during each sampling interval and a discrete-time linear model could be obtained each sampling interval. Unfortunately, this is not the usual case. With the aim of reducing the computational effort and to make the sliding mode design as simple as possible, a sliding surface defined only be the torque error will be used.

$$S = x = 0 \tag{31}$$

where

$$x = T_e^* - T_e \tag{32}$$

Considering $\dot{T}_e^* = 0$:

$$x_{k+1} = A_k x_k + B_k v_{sq,k} + (1 - A_k) T_e^* - B_k \lambda_{sd,k}\omega_{r,k} \tag{33}$$

Imposing $x_{k+1} = 0$, a deadbeat control law can be easily obtained [7]:

$$v_{sq,k} = -B_k^{-1} A_k x_k - B_k^{-1}(1 - A_k) T_e^* + \lambda_{sd,k}\omega_{r,k} \tag{34}$$

However, this control law is dependent on the exact knowledge of the many parameters involved and on the stator flux and rotor speed.

In order to improve the robustness of the torque controller, a discrete-time sliding mode technique according to Gao et al [9] will be used.

The discrete-time state equation for the torque error (33) can be rewritten to explicit the uncertainties:

$$x_{k+1} = \hat{A}_k x_k + \hat{B}_k v_{sq,k} + \left(1 - \hat{A}_k\right) T_e^*$$

$$-\hat{B}_k \lambda_{sd,k}\omega_{r,k} - \Delta A_k T_{e,k} + F_k \tag{35}$$

where $-\Delta A_k T_{e,k}$ represents the uncertainties present in the first and third terms of equation (35) and F_k represents the uncertainties present in the second and fourth terms.

The following convergence condition must be satisfied for the discrete-time sliding mode according to Gao et al. [9] to occur.

$$S_{k+1} = (1 - qT) S_k - \varepsilon T sgn(S_k) - \Delta A_k T_{e,k} + F_k$$

$$-(S_1 + F_1) - (S_2 + F_2) sgn(S_k) \tag{36}$$

where

$$S_1 = \frac{\sup(-\Delta A T_e) + \inf(-\Delta A T_e)}{2} \tag{37}$$

$$S_2 = \frac{\sup(-\Delta A T_e) - \inf(-\Delta A T_e)}{2} \tag{38}$$

$$F_1 = \frac{\sup(F) + \inf(F)}{2} \tag{39}$$

$$F_2 = \frac{\sup(F) - \inf(F)}{2} \tag{40}$$

It can be proved [9] that if the defined convergence condition is satisfied, then the sliding surface is reached in a

finite time interval and after that it will be crossed every sampling interval. Further, if

$$S_2 + F_2 < \frac{qT\varepsilon T}{2\left(1-qT\right)} \tag{41}$$

the distance from the system representative point to the sliding surface, after the sliding surface to be crossed for the first time, is limited to the quasi sliding mode band below:

$$|S| < \frac{\varepsilon T}{1-qT} \tag{42}$$

The only problem now is to choose an input such that the convergence condition is imposed. From (31), (35) and (36) the q axis stator voltage can be obtained by

$$v_{sq,k} = B_k^{-1}[-A_k x_k - \left(1-A_k\right)T_e^* + B_k\lambda_{sd,k}\omega_{r,k}$$

$$+\left(1-qT\right)x_k - \varepsilon T sgn\left(x_k\right) - \left(S_1+F_1\right)$$

$$-\left(S_2+F_2\right)sgn\left(x_k\right)] \tag{43}$$

The first three terms correspond to the deadbeat control law (34) and the additional terms are necessary to improve the robustness of the controller.

In fact, if the control law (43) is applied to Eqn. (35):

$$x_{k+1} = \left(1-qT\right)x_k - \varepsilon T sgn\left(x_k\right)$$

$$-\left(S_1+F_1\right) - \left(S_2+F_2\right)sgn\left(x_k\right) - \Delta A_k T_{e,k} + F_k \tag{44}$$

Then, if $x_k > 0$,

$$x_{k+1} = \left(1-qT\right)x_k - \varepsilon T sgn\left(x_k\right)$$

$$-\left[\sup\left(-\Delta A_k T_{e,k}\right) - \left(-\Delta A_k T_{e,k}\right)\right] - \left[\sup\left(F_k\right) - \left(F_k\right)\right] \tag{45}$$

$$\Longrightarrow x_{k+1} \leq \left(1-qT\right)x_k - \varepsilon T sgn\left(x_k\right) \tag{46}$$

Since $0 < \left(1-qT\right) < 1$, the error will decrease by at least $-\varepsilon T$. Furthermore, if $0 < x_k < \frac{\varepsilon T}{1-qT}$, then the error will become negative.

Similarly, if $x_k < 0$, the error will increase by at least εT and if $-\frac{\varepsilon T}{1-qT} < x_k < 0$, then it will become positive.

B.1 Designing a Robust Torque Controller

The torque controller design can be easily made in order to ensure robustness to parameter deviations and measuring errors.

The first step is to compute the maximum deviations that are expected and, based on them, to calculate S_1 , S_2, F_1 and F_2. Then, control law (43) is obtained. However, it should be noted that condition (41) must be satisfied. If large parameter deviations are expected, in order to impose condition (41) ε and q will be chosen so that the ratio $\varepsilon T/\left(1-qT\right)$ is big. In this case, the torque oscillations will be confined to a large quasi sliding mode band.

Physically, the voltage vectors to be applied need to be increased for the robustness to be improved. In practice, the ability to guarantee robustness will be limited, since the magnitude of the voltage vectors are limited by the inverter. The maximum robustness that can be obtained will then correspond to a mode of operation very similar to the conventional DTC strategies.

As an example of the torque controller robustness to errors in the system parameters, the torque controller will be designed considering that the rotor time constant can vary from 2/3 to 2 times its previously measured value.

Rotor time constant is present only in the parameter ΔA in Eqn. (35). Thus, the disturbances term F will be neglected in the present example.

From (29):

$$1 - \left(\frac{1.5}{\sigma\hat{\tau}_r} + \frac{1}{\sigma\tau_s}\frac{\lambda_{sd,k}-\sigma L_s i_{sd,k}}{\lambda_{sd,k}}\right)T < A_k = \hat{A}_k + \Delta A_k$$

$$< 1 - \left(\frac{0.5}{\sigma\hat{\tau}_r} + \frac{1}{\sigma\tau_s}\frac{\lambda_{sd,k}-\sigma L_s i_{sd,k}}{\lambda_{sd,k}}\right)T \tag{47}$$

$$\Longrightarrow -\frac{0.5}{\sigma\hat{\tau}_r}T < \Delta A_k < \frac{0.5}{\sigma\hat{\tau}_r}T \tag{48}$$

Considering the torque to be limited to some saturation values $\pm T_{e_{sat}}$, S_U and S_L are determined:

$$S_U = \frac{0.5}{\sigma\hat{\tau}_r}TT_{e_{sat}} \tag{49}$$

$$S_L = -\frac{0.5}{\sigma\hat{\tau}_r}TT_{e_{sat}} \tag{50}$$

Thus, $S_1 = 0$ and $S_2 = S_U$. The control law is then

$$v_{sq,k} = B_k^{-1}[-A_k x_k - \left(1-A_k\right)T_e^* + B_k\lambda_{sd,k}\omega_{r,k}$$

$$+\left(1-qT\right)x_k - \varepsilon T sgn\left(x_k\right) - \frac{0.5}{\sigma\hat{\tau}_r}TT_{e_{sat}}sgn\left(x_k\right)] \tag{51}$$

Figs. 1 and 2 show the torque response to a 5N.m. step command with the measured rotor time constant equal to 0.5 and 1.5 times the actual value respectively. The sampling interval is 0.25ms. The quasi sliding mode band is reached in two sampling intervals. A faster dynamic response is not possible due to a limit imposed in the maximum q axis voltage the inverter can generate.

Sometimes a voltage command saturation may be necessary to avoid overcurrents. As expected, after crossing the sliding surface for the first time, the system representative point crosses it again each sampling interval and is confined to the quasi sliding mode band. Any combination of parameter variations is allowed, since S and F are

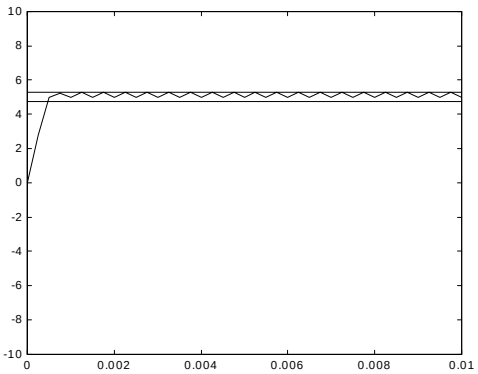

Fig. 1. Torque response to a step command considering $\hat{\tau}_r = 0.5\tau_r$.

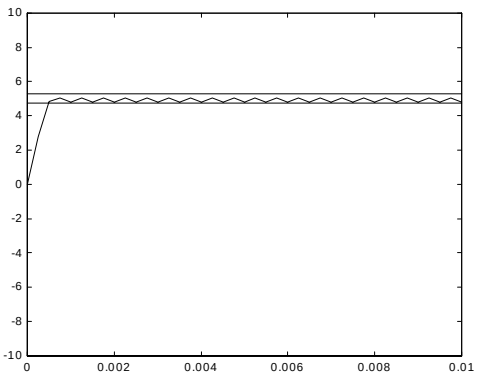

Fig. 2. Torque response to a step command considering $\hat{\tau}_r = 1.5\tau_r$.

maintained between their lower and upper limits. The bigger the S and F expected variations, the bigger the quasi sliding mode band necessary to ensure the robustness. The designer must seek a solution with the minimum torque oscillations (or minimum quasi sliding mode band) that can be obtained to ensure robustness to acceptable parameter variations and measuring errors in the system.

III. Flux Estimation

For flux estimation, a reduced order Gopinath stator flux observer was implemented [7]. In a stator fixed dq reference frame, the stator flux can be calculated from the induction motor stator voltage equation. The observed flux is obtained by adding a correction term based on the stator current derivative error:

$$\dot{\vec{\lambda}}_s = \vec{v}_s - R_s \vec{i}_s \tag{52}$$

$$\dot{\widehat{\vec{\lambda}}}_s = \vec{v}_s - R_s \vec{i}_s - L\left(\dot{\vec{i}}_s - \dot{\widehat{\vec{i}}}_s\right) \tag{53}$$

The stator current derivatives in Eqn. (53) can be calculated from stator voltage, stator current and stator flux, using the induction motor state model:

$$\dot{\vec{i}}_s = \frac{1}{\sigma L_s}\left[\vec{v}_s - R'_s \vec{i}_s + \left(\frac{R_r}{L_r} - j\omega_r\right)\vec{\lambda}_s\right] - \left(\frac{R_r}{L_r} - j\omega_r\right)\sigma L_s \vec{i}_s \tag{54}$$

$$\dot{\widehat{\vec{i}}}_s = \frac{1}{\sigma L_s}\left[\vec{v}_s - R'_s \vec{i}_s + \left(\frac{R_r}{L_r} - j\omega_r\right)\widehat{\vec{\lambda}}_s\right] - \left(\frac{R_r}{L_r} - j\omega_r\right)\sigma L_s \vec{i}_s \tag{55}$$

where $R'_s = R_s + R_r \frac{L_m^2}{L_r^2}$.

The stator flux error is then governed by

$$\dot{\vec{\varepsilon}} = \dot{\vec{\lambda}}_s - \dot{\widehat{\vec{\lambda}}}_s = \frac{1}{\sigma L_s} L \left(\frac{1}{\tau_r} - j\omega_r\right)\vec{\varepsilon} \tag{56}$$

or, considering $L = \begin{bmatrix} l_1 & -l_2 \\ l_2 & l_1 \end{bmatrix}$:

$$\dot{\vec{\varepsilon}} = \frac{1}{\sigma L_s}\begin{bmatrix} \frac{l_1}{\tau_r} - l_2\omega_r & -\left(\frac{l_2}{\tau_r} - l_1\omega_r\right) \\ \frac{l_2}{\tau_r} - l_1\omega_r & \frac{l_1}{\tau_r} - l_2\omega_r \end{bmatrix}\vec{\varepsilon} = A_{obs}\vec{\varepsilon} \tag{57}$$

Any observer dynamics can be easily imposed by an adequate choice of l_1 and l_2. However, in order to make the real part of the eigenvalues to be independent from the speed, l_2 will be chosen equal to zero. The observer poles are then

$$s_{1,2} = \frac{1}{\sigma L_s}\left(\frac{l_1}{\tau_r} \pm j\omega_r l_1\right) \tag{58}$$

As it can be seen, ω_r influences only the imaginary part of the eigenvalues. However, l_1 cannot be arbitrarily chosen for the discrete time flux computation requires a limited sampling rate.

Stator current differentiation in (53) is avoided by defining a new variable:

$$\widehat{\vec{z}} = \widehat{\vec{\lambda}}_s + L\vec{i}_s \tag{59}$$

From (53), (54) and (59), a state equation to obtain $\widehat{\vec{z}}$ is:

$$\dot{\widehat{\vec{z}}} = A_{obs}\widehat{\vec{z}} + \left[-A_{obs}L - R_s I - \frac{R'_s}{\sigma L_s}L - \left(\frac{1}{\tau_r}I - \omega_r J\right)\right]\vec{i}_s + \left(I + \frac{1}{\sigma L_s}L\right)\vec{v}_s \tag{60}$$

where $I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$ and $J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$.

Eqn. (60) is solved each sampling period and the observed stator flux is obtained from (59).

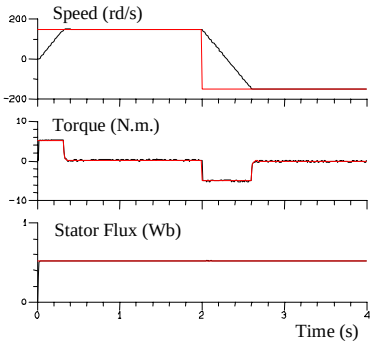

Fig. 3. Speed, Torque and flux responses to steps in the commanded speed. Simulation Results.

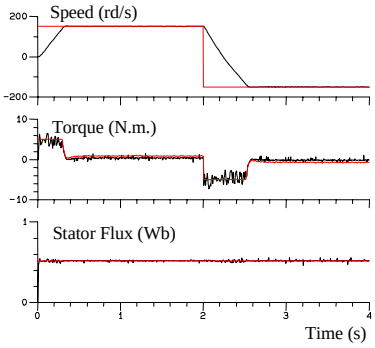

Fig. 4. Speed, Torque and flux responses to steps in the commanded speed. Experimental Results.

IV. Simulation and Experimental Results

The prototype implemented to obtain the experimental results contains an AC/DC/AC converter, a 2hp induction motor coupled to a DC generator that serves as a load torque, a signal conditioning circuit and a 133MHz Pentium microcomputer equipped with an acquisition board and two PWM interfaces.

The induction motor data are:

- Power: $2hp$
- Rated voltage: $220/380V$
- 4 Poles
- Inertia: $0.018kg.m^2$
- $R_s = 3.85\Omega$, $R_r = 3.77\Omega$
- $L_m = 257mH$, $L_s = 266mH$, $L_m = 269mH$

Some simulation results are presented. The system data used for the simulations are equal to the measured ones of the prototype used to obtain experimental results.

Fig. 3 shows speed, torque and flux simulation responses to steps in the speed command, using the proposed strategy. Torque and flux follow their command values without significant ripples. A speed command of 150 electrical rad/s is applied and a speed reversal is commanded at 2.0s. The experimental results to the same speed commands are presented in Fig. 4. The experimental results corroborate with the simulation and with the theoretically predicted results. Simulation and experimental responses to load torque steps are presented in Figs 5 and 6 respectively.

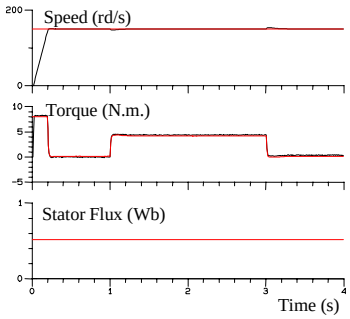

Fig. 5. Speed, Torque and flux responses to steps in the load torque. Simulation Results.

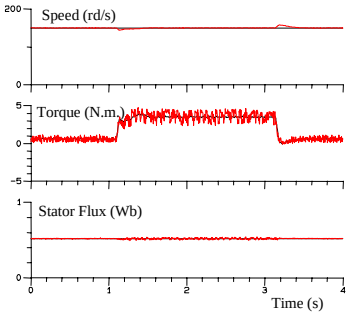

Fig. 6. Speed, Torque and flux responses to steps in the load torque. Experimantal Results.

V. Conclusion

The proposed control strategy provides a solution for conventional DTC schemes, in which a variable switching frequency must be used to ensure the torque and flux errors to remain inside the hysteresis bands. Further, differently from deadbeat DTC strategies, the discrete-time sliding mode torque and flux controllers are shown to be robust to parameter variations. Instead of applying one voltage vector during all switching period, a vector PWM algorithm is used and the torque oscillations are reduced. However, some torque oscillations are expected, since some terms must be added to the deadbeat control law to guarantee the system robustness. The simulation and experimental results confirm the theoretical analysis presented.

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References

[1] Xu, X., DeDoncker, R. and Novotny, D., "A stator flux oriented induction motor drive", 1988 Power Electronics Specialists Conference, pp. 870-876.

[2] Xue, X., Xu, X. Habetler, T. and Divan, D., "A low cost stator flux oriented voltage source variable speed drive", IEEE IAS Annual Meeting, 1990, pp. 410-415.

[3] Takahashi, I. and Noguchi, T., "A new quick-response and high-efficiency control strategy of an induction motor", IEEE Transactions on Industry Applications, Vol. IA-22, Sept./Oct./1986.

[4] Depenbrock, M., "Direct self-control (DSC) of inverter-fed induc-

tion machine", IEEE Transactions on Power Electronics, Vol. 3, No. 4, Oct./1988.

[5] Habetler, T., Profumo, F., Pastorelli, M. and Tolbert, L., "Direct torque control of induction machines using space vector modulation", IEEE Transactions on Industry Applications, Vol. 28, No. 5, pp. 1045-1053, Sept./Oct./1992.

[6] Griva, G., Habetler, T., Profumo, F. and Pasorelli, M., "Performance evaluation of a direct torque controlled drive in the continuous PWM-square wave transition region", IEEE Transactions on Power Electronics, Vol. 10, No. 4, pp. 464-471, July/1995.

[7] Neves, F. , Menezes, B. and Silva, S., "A deadbeat torque and flux field oriented controller for induction motor drives", Proceedings of the Brazilian Automatic Conference CBA'98, Vol. 6, pp. 2243-2247, 1998.

[8] Neves, F. , Menezes, B. and Silva, S., "Discrete-time Sliding mode control: a new design strategy applied to induction motor drives", Proceedings of the American Control Conference ACC'98, Vol. 1, pp. 125-130, Philadelphia-USA, June, 1998.

[9] Gao, W., Wang, Y. and Homaifa, A., "Discrete-time variable structure control systems", Transactions on Industrial Electronics, Vol. 42, No. 2, pp. 117-122, April/1995.