ROTOR FLUX ORIENTED TORQUE-CONTROL OF INDUCTION MACHINES BASED ON STATOR FLUX VECTOR CONTROL

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Abstract. This paper describes a speed sensorless rotor flux oriented torque-control of induction machines which utilises stator flux components as control variables. Combining the advantages of rotor flux orientation and stator flux space vector control leads to high performance drives using a simple control scheme. This scheme is affected by total leakage inductance variations which determine detuned operation. A method is presented that provides the possibility to estimate the total leakage inductance besides the identification of stator self inductance. The effectiveness of the proposed system is verified by computer simulation and experimental tests.

Keywords. AC drives, field-oriented control, direct-torque control, parameter adaptation.

INTRODUCTION

For high performance drives based on induction machines, the principle of field oriented control is usually employed. Field oriented control can be exercised on stator, air-gap or rotor flux, requiring the knowledge of the instantaneous magnitude and position of the flux vector. Considering rotor flux orientation, a complete decoupling of flux and torque control variables is possible, and the highest pull out torque can be obtained. A torque command at constant rotor flux is followed instantaneously without transient rotor and stator currents. However, in all rotor flux field oriented control schemes a correct estimation of the machine parameters is essential to obtain an accurate control. In stator and air-gap flux orientation sophisticated parameter estimation methods are not required to produce robust torque control. On the other hand the pull-out torque is not so high as in rotor flux orientation and the simplicity of rotor flux decoupling equation is lost [1].

Recently a direct torque-control method was introduced which directly control the stator flux and torque on the basis of the instantaneous errors in flux and torque [2]. According to the switching capabilities of voltage source inverters, the selection of the switching configuration is made in order to maintain the stator flux and the torque within the limits of two hysteresis bands. This method, however, results in a variable switching frequency due to the hysteresis control technique. In [3] a method, based on the Space Vector Modulation (SVM) technique, is presented which allows constant switching frequency operation.

In this paper a control scheme based on rotor flux orientation which utilises the stator flux components as control variables is considered [4]. Combining the advantages of rotor flux orientation and stator flux space vector control leads to a high performance drive using a simple control scheme without the requirement for speed or position feedback. The input commands of the control system are the torque and rotor flux. The determination of the stator flux reference requires the knowledge of the total leakage inductance and the ratios of stator and rotor self inductance to mutual inductance. An identification method of the total leakage inductance which is based on the analysis of the irregularities in the estimated rotor flux vector locus is proposed. Furthermore, utilising the relationship between stator flux and current, a method for accurate estimation of the stator self inductance is analysed.

Simulation results showing the good performance of the parameter adaptation algorithm are given. For the experimental tests a DSP based high-speed A/D conversion board and an IGBT inverter have been utilised. The experimental results are in good agreement with the theoretical results.

FIELD ORIENTATION MODEL

The stator and rotor flux of an induction machine, in the rotor reference frame can be written in terms of spatial vectors as

$$\bar{\Phi}_s^r = L_s \bar{i}_s^r + M \bar{i}_r^r \tag{1}$$

$$\bar{\Phi}_r^r = L_r \bar{i}_r^r + M \bar{i}_s^r \tag{2}$$

By eliminating $\bar{i}_s^r$ from the preceding equations we obtain

$$\bar{\Phi}_s^r = \frac{L_s}{M}\left(\bar{\Phi}_r^r - \sigma L_r \bar{i}_r^r\right) \tag{3}$$

where $\sigma = 1 - \dfrac{M^2}{L_s L_r}$.

From the rotor voltage equation the rotor current is

$$\bar{i}_r^r = -\frac{1}{R_r} p\bar{\Phi}_r^r \tag{4}$$

where $p = d/dt$.

Substituting (4) in (3) yields

$$\bar{\Phi}_s^r = \frac{L_s}{M}\left(\Phi_r + \sigma\tau_r p\Phi_r + j\,\omega_r\sigma\tau_r\Phi_r\right) e^{j\vartheta_r^r} \tag{5}$$

where $\tau_r = L_r/R_r$, $\omega_r = p\vartheta_r^r$ being ϑ_r^r the phase angle of $\bar{\Phi}_r^r$ in the rotor reference frame.

For a two-pole machine, the electromagnetic torque T can be expressed as

$$T = \omega_r \frac{\Phi_r^2}{R_r} \tag{6}$$

By eliminating ω_r between (5) and (6) we obtain

$$\bar{\Phi}_s^r = \frac{L_s}{M}\left(\Phi_r + \sigma\tau_r p\Phi_r + j\,\sigma L_r \frac{T}{\Phi_r}\right) e^{j\vartheta_r^r} \tag{7}$$

The field orientation concept is accomplished choosing a synchronously rotating reference frame and locking the phase of the reference system such that the rotor flux is entirely on the d-axis (Fig. 1).

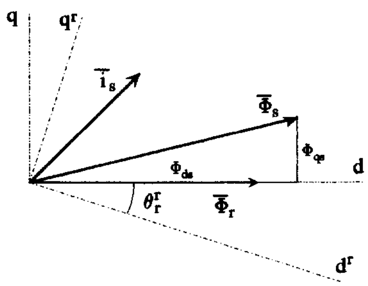


Fig. 1 - Space-phasor diagram with rotor and synchronous reference frame.

Whit this choice, the d-q stator flux components derived from Eq. 7 result

$$\Phi_{ds} = \frac{L_s}{M}\left(1 + p\,\sigma\,\tau_r\right)\Phi_r \tag{8}$$

$$\Phi_{qs} = \frac{L_r}{M}\sigma L_s \frac{T}{\Phi_r} \tag{9}$$

Eqs. 8 and 9 represent the basic equations of the rotor field orientation where the control variables are the stator flux components. The control of torque and rotor flux is obtained in terms of stator flux components Φ_{qs} and Φ_{ds} respectively.

From Eq. 8 it appears that a change in the flux command Φ_{ds} is followed with a time constant $\sigma\tau_r$ which is much lower than the rotor time constant τ_r involved in traditional stator current control. With constant rotor flux command, a change in Φ_{qs} is followed instantaneously by a corresponding change in torque. The relationship between Φ_{qs} and T requires the quantities L_r/M, which is moderately affected by saturation, and σL_s which is slightly dependent on the current due to saturation effects.

TORQUE CONTROL IMPLEMENTATION

Using stator flux components as control variables, the open-loop torque control can be achieved by employing standard, sensorless induction motors. A basic implementation scheme is represented in Fig. 2.

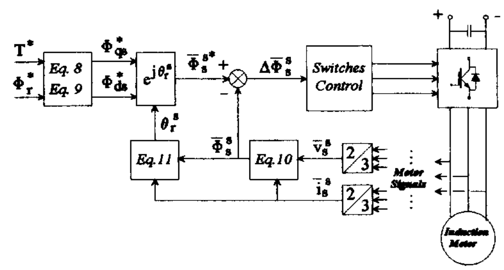


Fig. 2 - Block diagram of field orientation scheme.

In this scheme the d-q components of the stator flux reference are obtained from the torque and rotor flux commands using Eqs. 8 and 9. The terminal voltages and the phase currents are employed to estimate the stator flux vector in a stationary reference frame by integrating the stator voltage equation as follows

$$\bar{\Phi}_s^s = \int\left(\bar{v}_s^s - R_s\bar{i}_s^s\right)dt \tag{10}$$

The rotor flux vector is obtained combining Eqs. 1 and 2, rewritten in a stator reference frame (Fig. 3), leading to

$$\bar{\Phi}_r^s = \frac{L_r}{M}\left(\bar{\Phi}_s^s - \sigma L_s\bar{i}_s^s\right) \tag{11}$$

The phase angle ϑ_r^s of the rotor flux in a stator reference frame can now be calculated from the values of the d-q components of $\bar{\Phi}_r^s$ obtained from Eq. 11.

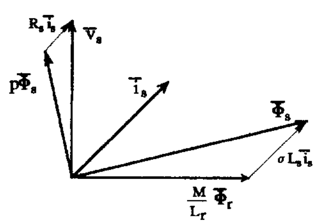


Fig. 3 - Space-phasor diagram for stator and rotor quantities

The reference value of the stator flux vector $\bar{\Phi}_s^{s*}$ is then calculated by applying the coordinate transformation from synchronous to stator reference frame. By comparing the estimated and the reference value of the stator flux vector, the flux error $\Delta\bar{\Phi}_s^s$ is readily obtained. The knowledge of this error allows the determination of the proper voltage vector which has to be applied to the motor during the next sampling period t_s. For this purpose traditional PWM or SVM control technique can be used according to the following equation

$$\vec{v}_s^{s*} = \frac{\Delta\vec{\Phi}_s^s}{t_s} + R_s\vec{i}_s^s \qquad (12)$$

A further solution can be obtained applying the 3-phase flux errors to three independent hysteresis controller working under "bang-bang" control strategy [5].

The control scheme shown in Fig. 2 requires the determination of the instantaneous position of rotor flux. However, the torque control can be achieved utilising a simplified and efficient scheme which is based only on the determination of stator flux magnitude. A block diagram of this control scheme is shown in Fig. 4.

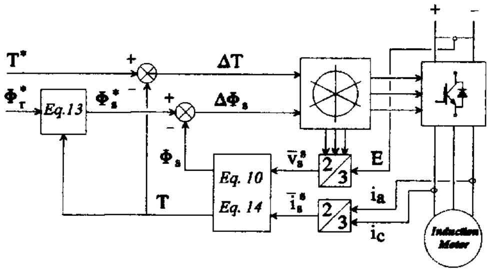


Fig. 4 - Block diagram of direct torque-control scheme.

This scheme allows torque control without transient conditions when the rotor flux magnitude is kept constant. The input commands are the torque T^* and the rotor flux magnitude Φ_r^*. The torque and the stator flux are directly controlled utilising the space vector modulation technique on the basis of two hysteresis band [2]. The stator flux reference is calculated from Eq. 8, written in steady-state, and Eq. 9, giving

$$\Phi_s^{*2} = \Phi_{ds}^{*2} + \Phi_{qs}^{*2} = \left(\frac{L_s}{M}\Phi_r^*\right)^2 + (\sigma L_s)^2\left(\frac{L_r}{M}\frac{T}{\Phi_r^*}\right)^2 \qquad (13)$$

In Eq. 13 the estimated value of torque is used rather than reference value because this choice improves the dynamic performance to step torque commands. As it is shown in Fig. 4, the actual values of $\vec{v}_s^s$ and $\vec{i}_s^s$ are calculated by using the DC link voltage E, the inverter switching states and the motor line currents i_a and i_c.

The actual value of $\vec{\Phi}_s^s$ is evaluated utilising Eq. 10, while the torque estimation is carried out according to the following equation

$$T = \vec{i}_s^s \cdot j\,\vec{\Phi}_s^s \qquad (14)$$

MOTOR PARAMETER ADAPTATION

In sensorless field oriented control schemes, the calculation of the actual value of stator flux requires the knowledge of the stator resistance (Eq. 10). An accurate value of this resistance is necessary particularly at low excitation frequency. Therefore, the speed range is to be limited unless some form of resistance estimator or tuning is incorporated. Some methods for the determination of the stator resistance are described in [5], [6].

The proposed control scheme requires also the evaluation of the stator flux reference Φ_s^* according to Eq. 13. This calculation is based on the knowledge of the total leakage inductance σL_s. An error in σL_s has the same effects at all excitation frequencies where it is zero when the torque command is zero and becomes worse under load. Thus it is important to minimise the error in σL_s. In the following a method of σL_s tuning is described. For this purpose it is useful to determine the relationship between the stator and the rotor flux vectors. Substituting (4) in (3) and solving for $\vec{\Phi}_r^r$, yields

$$\vec{\Phi}_r^r = \frac{M/L_s}{1+p\sigma\tau_r}\vec{\Phi}_s^r \qquad (15)$$

This equation clearly shows the nature of the rotor flux dynamic response for changes in stator flux.

The space vector modulation leads to a stator flux vector locus characterised by a ripple due to the switching processes, while the rotor flux vector locus is practically a circle owing to the first order, low-pass filtering action expressed by (15). In fact, the cut-off frequency of this filter is usually much lower than the control frequency.

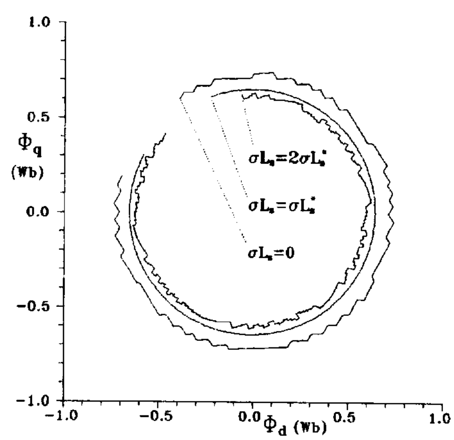


Fig. 5 - Vector locus of estimated rotor flux in tuned and detuned operation.

This analysis leads to the conclusion that the estimation of Φ_r, achieved by Eq. 11, is expected to be a constant value when the flux command Φ_r^* is held constant. Hence, the variations at the control frequency of the two terms in Eq. 11 counterbalance each other only

when σL_s assumes the correct value.
When σL_s is incorrect, detuned operation determines a ripple in the estimated value of Φ_r. Fig. 5 gives some vector loci of the estimated rotor flux to illustrate the effects caused by detuning. Note that an incorrect value of σL_s causes also a phase displacement of rotor flux vector determining the loss of field orientation in control scheme based on the determination of the rotor flux instantaneous position.
The analysis of detuned operation suggests that it is possible to implement an algorithm in order to find the value of σL_s which minimises a suitable error function. This algorithm should be simple and efficient in order to reduce the computational time. Good results have been obtained utilising an error function calculated by adding, at each sampling period, the absolute value of the difference between the actual and the previous estimated value of Φ_r

$$\varepsilon_N = \sum_{k=1}^{N} \left| \Phi_r(k) - \Phi_r(k-1) \right| \tag{16}$$

Each value of ε_N is calculated for a prefixed value of σL_s on the basis of N samples. Applying the same procedure for different values of σL_s and interpolating it is possible to obtain ε_N as function of σL_s. The correct value of the parameter σL_s is readily obtained finding the minimum of this function.
Once the correct value of σL_s is known, an accurate estimation of the stator self-inductance L_s can be obtained. For this purpose, writing Eqs. 1 and 4 in a stator reference frame and manipulating leads to

$$L_s\, \bar{i}_s^s \cdot e^{j\vartheta_r^s} - \bar{\Phi}_s^s \cdot e^{j\vartheta_r^s} = \frac{M}{R_r}\, p\Phi_r \tag{17}$$

For constant rotor flux operation, Eq. 17 gives

$$L_s = \frac{\bar{\Phi}_s^s \cdot e^{j\vartheta_r^s}}{\bar{i}_s^s \cdot e^{j\vartheta_r^s}} = \frac{\Phi_{ds}}{i_{ds}} \tag{18}$$

where the rotor flux angular position ϑ_r^s is obtained from Eq. 11.
In application of the torque control scheme represented in Fig. 4 the calculation of Φ_s^* is based on Eq. 13, where the knowledge of L_s/M, L_r/M and σL_s is required. The mutual inductance coefficient M can be easily evaluated assuming the total leakage inductance σL_s equally distributed between stator and rotor windings. This assumption leads to small errors in the parameter L_s/M and L_r/M. Hence, in Eq. 13 the only parameter having a significant influence on Φ_s^* calculation is σL_s, which can be accurately estimated as above explained.
Utilising the described procedure it is then possible to determine with few calculations the parameters required by the control scheme of Fig. 4. The proposed control scheme can then be applied to any standard induction motor allowing the system to be correctly tuned.

SIMULATION OF THE CONTROL SCHEME

To analyse the performance of the control scheme and the phenomena related to detuned operation it is useful to simulate numerically the whole system [4]. The numerical results can be utilised to find a suitable algorithm for parameter adaptation and to anticipate an appropriate hardware structure. The motor under test is a standard squirrel-cage induction motor. The nameplate data are as follows

3-phase, 2-pole, rated voltage 220 V, 50 Hz
Output power 4 kW at 2890 rpm

The machine parameters are

$R_s = 0.402\ \Omega$ $\quad R_r = 0.307\ \Omega$
$L_s = 87.9$ mH $\quad L_r = 89.2$ mH $\quad M = 84.8$ mH

The DC link voltage is $E = 310$ V.

With reference to this machine it has been analysed how the error function ε_N, defined in Eq. 16, varies as function of the error in σL_s. As above mentioned, an error in the value of σL_s will affect the vector locus of the estimated rotor flux leading to a rugged circle.
Fig. 6 illustrates the calculated values of ε_N in steady-state operation at rated torque and for a speed of 150 rad/s. Each value of ε_N is calculated considering N=500 samples and is plotted as function of the ratio between the value of the total leakage inductance σL_s, used to calculate ε_N, and the correct value σL_s^*.

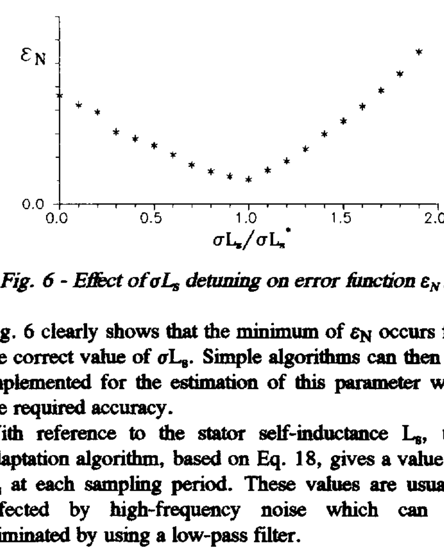


Fig. 6 - Effect of σL_s detuning on error function ε_N.

Fig. 6 clearly shows that the minimum of ε_N occurs for the correct value of σL_s. Simple algorithms can then be implemented for the estimation of this parameter with the required accuracy.
With reference to the stator self-inductance L_s, the adaptation algorithm, based on Eq. 18, gives a value of L_s at each sampling period. These values are usually affected by high-frequency noise which can be eliminated by using a low-pass filter.
The convergence from an initial zero estimate to the correct final value is illustrated in Fig. 7.
Finally, it will be shown that in the control scheme proposed in Fig. 4 the decoupling between the stator flux d and q components is achieved as in the control scheme of Fig. 2. For this purpose a torque reference

alternating between 50% and 200% of rated torque is applied to the system at steady state running speed of 150 rad/s.

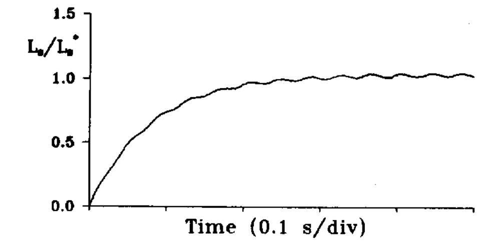


Fig. 7 - Simulated response of L_s identification.

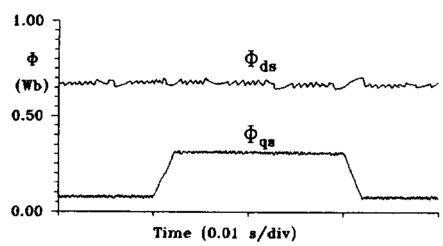


Fig. 8 - Stator flux components in a synchronous reference frame for a torque pulse demand.

Fig. 8 gives the corresponding d and q stator flux components obtained by numerical simulations. It can be noticed that Φ_{ds} is practically constant as it could be expected according to Eq. 8 for constant rotor flux command, while Φ_{qs} varies in response to the torque command. As specified in Eq. 9 the developed torque follows exactly Φ_{qs} .

PERFORMANCE OF THE EXPERIMENTAL SYSTEM.

The direct torque control scheme has been implemented on an interrupt-based program for DSP. The signal processor chosen is the TMS320E15 with a clock rate of 20 MHz. The DSP is interfaced to a 1 MHz, 8-channel, 12-bit A/D converter and to a 2-channel, 16-bit D/A converter as represented in Fig. 9.
The feedback signals to the DSP-based controller are the measured motor line currents and the DC link voltage. These quantities are measured by Hall-effect sensors. The reference rotor flux Φ_r^* is a fixed constant for constant flux operation. The execution of the stator flux and torque control loops are carried out in a straightforward manner. These include the computation of the reference stator flux and the actual values of stator flux and torque. The outputs of the controller are pulse signals that are sent to the driver circuits of the inverter switches on the basis of the instantaneous errors in torque ΔT and flux $\Delta\Phi_s$. According to the combination of the switching modes, the voltage vectors are specified for eight kinds of vectors [4], [2].

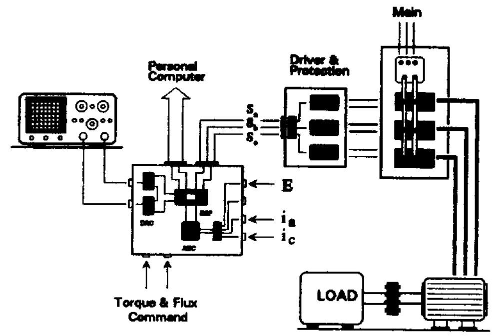


Fig. 9 - Hardware configuration of the experimental set-up.

The selection of the proper voltage vector is made in order to maintain the torque and the stator flux within the limits of two hysteresis bands.
The experimental set-up includes an IGBT inverter and a 4 kW standard induction motor. The main switches are rated at 1000 V, 50 A.
The total time required to read in the data, execute the software and output the switching state to the inverter is about 30 μs. The flowchart of the algorithm and the approximate time consumption of each block are shown in Fig. 10.

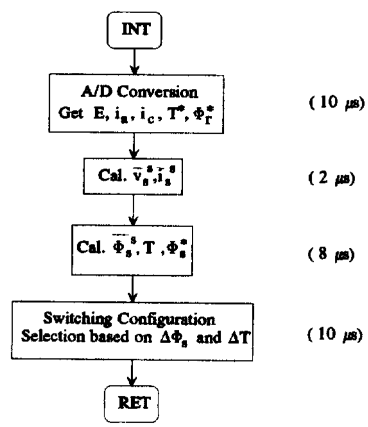
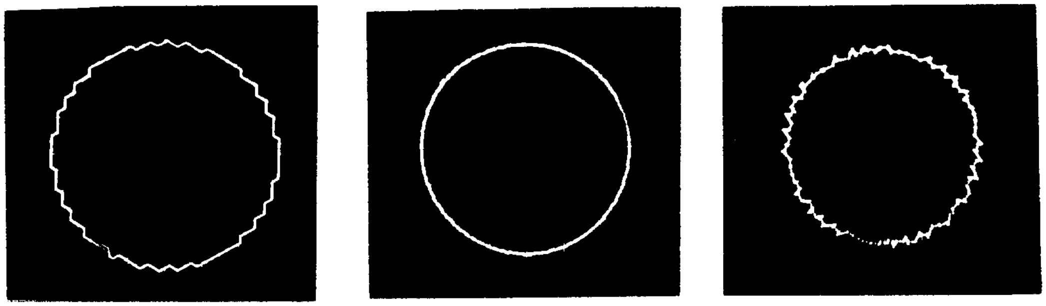


Fig. 10 - *Flow-chart of the control loop.*

The required computational time allows a 20 kHz sampling frequency leaving enough time for the implementation of the parameter adaptation algorithm.
The experimental results are shown in Figs. 11 and 12. The oscillograms have been obtained by converting the digital estimated values to analog form using D/A converters. Fig. 11 shows the vector loci of the estimated rotor flux in tuned and detuned operation.
It should be noted that in Fig. 11a, being $\sigma L_s=0$, the



a) $\sigma L_s/\sigma L_s^* = 0$ b) $\sigma L_s/\sigma L_s^* = 1$ c) $\sigma L_s/\sigma L_s^* = 2$

Fig. 11 - Experimental vector loci of estimated rotor flux.

trace represents the stator flux vector locus scaled by the factor L_r/M, as results from Eq. 11. Fig. 11b shows the circle representing the vector locus in tuned operation. Fig. 11c illustrates the typical rugged circle which appears when the parameter σL_s is over-estimated.

Fig. 12 illustrates the d-q components of stator flux, estimated during the response to a torque reference alternating between 50% and 200% of rated torque as previously considered in Fig. 8. These experimental results show that the decoupling between stator flux components is achieved controlling directly the stator flux magnitude as previously pointed out. However, the decoupling is lost under σL_s detuned operation.

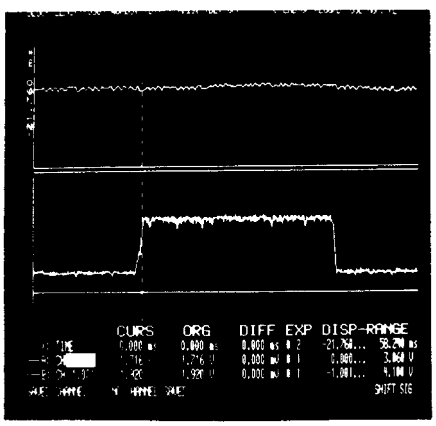


Fig. 12 - Experimental stator flux behaviour for a torque pulse command.

CONCLUSIONS

A direct torque-control method for induction machines which uses the rotor flux and the torque as references has been presented. The control technique is based on stator flux vector control and leads to a simple and robust torque control scheme which does not require speed feedback. To predict the influence of certain physical phenomena related to parameter variations, a simulation study of the overall system has been carried out. The analysis has shown that the irregularities in rotor flux vector locus are related to incorrect values of the total leakage inductance. Based on this consideration a method for the estimation of σL_s has been described. In addition a method for the estimation of the stator self inductance has been presented. The experimental tests validate the simulation results and emphasise the good performance of the control system realised.

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