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APPLICATION SPECIFIC SPEED IDENTIFICATION FOR INDUCTION MOTORS

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Abstract

In this paper we describe and analyze an application specific speed and flux estimation method of induction motors. The technique utilizes the mechanical equation of the drive, assuming a certain load torque characteristic. The goal is to extend the stable operating range towards the low speeds. The stability and the steady state performance of the method is investigated numerically and by simulation. The dynamic performance is verified by experiments and simulation.

Keywords: asynchronous drive, sensorless control, speed estimation, flux estimation, field oriented control.

1. Introduction

1.1. General Description

Today, cancelling the speed sensor became almost a base requirement for the vector controlled induction drives. The obvious merit is the reduced costs regarding the speed sensor and the cables. The maintaining and repairing costs are also smaller. It effects the reliability and robustness of the drive, so the fields where it can be applied is also widening.

On the other hand to realize field oriented control it is necessary to identify the rotor flux of the machine, which is quite problematic without speed sensor. At high speeds the well known identification methods [1], [2] work well, but in the range of low velocities their performance declines.

1.2. Limits of the Known Identification Structures

Let us assume, that we used a simplified modell for the iduction machine, where the parameters of machine are transformed in such a manner, that the transformed rotor part of the leakage inductance is zero. We used the transformed parameters and quantities in the following without any notice.

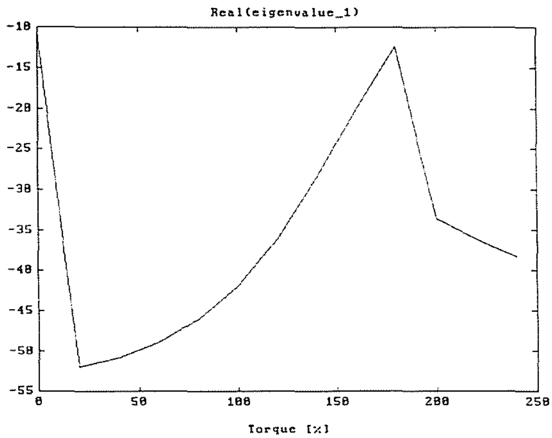

Fig. 1. The real part of the first eigenvalue

A possible way to determine the rotor flux vector in the stationary coordinate system is simply to integrate the induced voltage computed from the stator side:

$$\psi_r = \int \left(\mathbf{u}_s - \mathbf{i}_s \cdot R_s - L_s \cdot \frac{d\mathbf{i}_s}{dt} \right) dt \,. \quad (1)$$

Because of the open loop integration, any offset in the measured quantities causes a cumulative flux error. We can have a low pass element instead of the integrator, but then the estimation would be inaccurate in the low frequencies. The low frequency operating range is critical from other aspects too. When the frequency is low, the voltage amplitude is also small. In this case the real and calculated voltage drops can differ considerably because of the uncertain knowledge of motor parameters. Moreover in most of the applications it is assumed that the voltage of the motor is unambiguously determined by the pulse width. So this reference voltage signal is used instead of the measured phase values. This also causes an error because of the dead time, the voltage drop of the switching devices and the ripple of the DC voltage. Such an estimator can estimate the speed by the rotor voltage equation.

The other possibility is to calculate the flux from the rotor side [3] through a first order element, where the input is the stator current:

$$\psi_r = \int \left(-\frac{\psi_r}{L_r} \cdot R_s + j \cdot \omega \cdot \psi_r + \mathbf{i}_s \cdot R_r \right) dt \,, \quad (2)$$

where L_r is the rotor inductance and we neglect the effect of saturation on its change. Because the rotor equation involves the rotation, we must use the stator side quantities to estimate the speed, which is then used again in the calculation of the flux. So the problems raised by the stator equation also exist here.

At the end the conventional estimators can do nothing at standstill when there is no load torque. At this point the supply voltage has zero frequency. The only measured quantity, the current is determined simply by the Ohm's law: $\mathbf{i}_s = \frac{\mathbf{u}_s}{R_s}$, and the stator and rotor totally disjoint. For that reason the machine's speed and flux are not observable.

2. Estimation of the Rotor Flux and the Speed in the Proposed System

The estimator is working in the rotating, rotor flux-oriented coordinate system.

In this model we use the rotor voltage equation to calculate the magnitude of the flux on the basis of the d current and the magnetizing current:

$$\frac{d\hat{\psi}_r}{dt} = (\hat{i}_d - \hat{i}_m) \cdot R_r ; \tag{3}$$

The magnetizing curve is approximated by the following expression:

$$\hat{i}_m = i_{mn} \cdot \left(p_1 \cdot \frac{\hat{\psi}_r}{\psi_n} + (1 - p_1) \cdot \left(\frac{\hat{\psi}_r}{\psi_n} \right)^{p_2} \right) . \tag{4}$$

We get the angular velocity by utilizing the mechanical equation:

$$\frac{d\hat{\omega}}{dt} = \frac{1}{J} \cdot (\hat{m} - \hat{m}_{load}) , \tag{5}$$

where the torque is:

$$\hat{m} = 1.5 \cdot P \cdot \hat{\psi}_r \cdot \hat{i}_q . \tag{6}$$

The angle position of the flux is necessary for the transformation of the currents from standing to field coordinates and its derivative is calculated by the sum of the mechanical speed and the slip angular frequency:

$$\frac{d\rho}{dt} = \hat{\omega}_s = \hat{\omega} + \frac{\hat{i}_q \cdot R_r}{\hat{\psi}_r} . \tag{7}$$

The transformation of the current vector is the following: $\hat{i}_{d,q} = i_{x,y} \cdot e^{-j \cdot \hat{\rho}}$.

What this method requires -differently form the other known concepts- is the knowledge of the inertia and the load torque. The inertia can be measured, but the load torque is usually treated as a disturbance and unknown signal. However, in several applications such as machine tools, the load torque is well determined at low stator frequency or at standstill. In this range the torque-speed characteristics of these machines are very close to zero, hence neglecting the load-torque they provide a good approximation.

It is obvious that the knowledge of the stator parameters and the stator voltage is unnecessary, so the errors that originate from the stator voltage equation are eliminated.

3. Stability of the Identification

The small sign stability can be studied from the state-space model of small deviations.

Let us introduce the following small deviations:

$$\Delta\psi_r = \psi_r - \hat{\psi}_r, \quad \Delta\omega = \omega - \hat{\omega}, \quad \Delta\rho = \rho - \hat{\rho}_m, \quad \Delta m_{load} = m_{load} - \hat{m}_{load}. \quad (8)$$

We intend to find a model:

$$\frac{d}{dt}\begin{bmatrix} \Delta\psi_r \\ \Delta\omega \\ \Delta\rho \end{bmatrix} = \mathbf{A} \cdot \begin{bmatrix} \Delta\psi_r \\ \Delta\omega \\ \Delta\rho \end{bmatrix} + \mathbf{b} \cdot \Delta m_{load}. \quad (9)$$

If the real part of the eigenvalues of the matrix **A** is negative than the system is stable for small changes.

From the (3), (5), (7) differential equations we can derive the model for the small sign performance.

3.1. Small Deviations of the Flux Amplitude

Rewriting (3) with the substitution of the small deviations and comparing with (3) one can get:

$$\frac{d\psi_r - \Delta\psi_r}{dt} = i_d - \Delta i_d - (i_m - \Delta i_m)) \cdot R_r, \quad (10)$$

$$\frac{d\Delta\psi_r}{dt} = (\Delta i_d - \Delta i_m) \cdot R_r. \quad (11)$$

We must express the deviation of the magnetizing current in the term of the flux deviation. Taking the (2) expression:

$$i_m - \Delta i_m = i_{mn} \cdot \left(p_1 \cdot \frac{\psi_r - \Delta\psi_r}{\psi_n} + (1 - p_1) \cdot \left(\frac{\psi_r - \Delta\psi_r}{\psi_n} \right)^{p_2} \right). \quad (12)$$

If we approximate the second term with the first two elements of the Taylor series: $f(x - \Delta x) \approx f(x) - \Delta x \cdot \frac{df(x)}{dx}$, we get

$$\left(\frac{\psi_r - \Delta\psi_r}{\psi_n}\right)^{p_2} \approx \left(\frac{\psi_r}{\psi_n}\right)^{p_2} - \frac{\Delta\psi_r}{\psi_n} \cdot p_2 \cdot \left(\frac{\psi_r}{\psi_n}\right)^{p_2-1} . \tag{13}$$

Substituting this into (12)

$$i_m - \Delta i_m = i_{mn} \cdot \left(p_1 \cdot \frac{\psi_r}{\psi_n} + (1 - p_1) \cdot \left(\frac{\psi_r}{\psi_n}\right)^{p_2} - \right.$$
$$\left. -p_1 \cdot \frac{\Delta\psi_r}{\psi_n} - (1 - p_1) \cdot \frac{\Delta\psi_r}{\psi_n} \cdot p_2 \cdot \left(\frac{\psi_r}{\psi_n}\right)^{p_2-1}\right) \tag{14}$$

and comparing this with (4) it is seen that

$$\Delta i_m = i_{mn} \cdot \left(p_1 \cdot \frac{\Delta\psi_r}{\psi_n} + (1 - p_1) \cdot \frac{\Delta\psi_r}{\psi_n} \cdot p_2 \cdot \left(\frac{\psi_r}{\psi_n}\right)^{p_2-1}\right) . \tag{15}$$

The deviations of the d current component exist on the right hand side of the (11) equation. Therefore we should find an approximation for this, which must be a function only of the small changes.

As we previously stated: $i_{x,y} \cdot e^{-j\cdot\hat{\rho}} = \hat{i}_{d,q}$. Forming the real and imaginary part of the expression with the similar substitution as above yields:

$$(i_x + j \cdot i_y) \cdot (\cos(\rho - \Delta\rho) - j \cdot \sin(\rho - \Delta\rho)) = i_d - \Delta i_d + j \cdot (i_q - \Delta i_q) . \tag{16}$$

If the deviations are really small then $\cos(\Delta\rho) \approx 1$, $\sin(\Delta\rho) \approx \Delta\rho$, the above expression becomes the following:

$$i_x \cdot \cos(\rho) + i_y \cdot \sin(\rho) + (i_x \cdot \sin(\rho) - i_y \cdot \cos(\rho)) \cdot \Delta\rho +$$
$$+j(i_y \cdot \cos(\rho) - i_x \cdot \sin(\rho) + (i_y \cdot \sin(\rho) + i_x \cdot \cos(\rho)) \cdot \Delta\rho) = \tag{17}$$
$$= i_d - i_q \cdot \Delta\rho + j \cdot (i_q + i_d \cdot \Delta\rho)$$

Combining (16) and (17):

$$\Delta i_d = i_q \cdot \Delta\rho, \quad \Delta i_q = -i_d \cdot \Delta\rho . \tag{18}$$

So the change of the current components can be expressed by the change of the flux angle, and the change of the flux amplitude is then:

$$\frac{d\Delta\psi_r}{dt} = \left(i_q \cdot \Delta\rho - \frac{i_{mn}}{\psi_n} \cdot \Delta\psi_r \cdot \left(p_1 + (1 - p_1) \cdot p_2 \cdot \left(\frac{\psi_r}{\psi_n}\right)^{p_2-1}\right)\right) \cdot R_r . \tag{19}$$

3.2. Small Deviations of the Velocity

Using (5) and (6):

$$\frac{d(\omega-\Delta\omega)}{dt}=\frac{1}{J}\cdot(1.5\cdot P\cdot(\psi_r-\Delta\psi_r)(i_q-\Delta i_q)-(m_{load}-\Delta m_{load}))\approx$$

$$\approx\frac{1}{J}\cdot(1.5\cdot P\cdot(\psi_r\cdot i_q-\Delta\psi_r\cdot i_q-\psi_r\cdot\Delta i_q)-(m_{load}-\Delta m_{load})) \qquad (20)$$

neglecting the product of the deviations.
From this and (5):

$$\frac{d\omega}{dt}=\frac{1}{J}\cdot(1.5\cdot P\cdot\psi\cdot i_q-m_{load}). \qquad (21)$$

Taking (18):

$$\frac{d\omega}{dt}=\frac{1}{J}\cdot(1.5\cdot P\cdot\psi\cdot i_q-m_{load}). \qquad (22)$$

3.3. Small Deviations of the Flux Angle

The same way as before:

$$\frac{d(\rho-\Delta\rho)}{dt}=\omega-\Delta\omega+\frac{i_q-\Delta i_q}{\psi_r-\Delta\psi_r}\cdot R_r. \qquad (23)$$

Because $\frac{1}{\psi_r-\Delta\psi_r}=\frac{\psi_r+\Delta\psi_r}{\psi_r^2-\Delta\psi_r^2}\approx\frac{1}{\psi_r}+\frac{\Delta\psi_r}{\psi_r^2}$, the result

$$\frac{d\rho}{dt}=\Delta\omega+\Delta i_q\frac{R_r}{\psi_r}-\Delta\psi\frac{i_q\cdot R_r}{\psi_r^2}, \qquad (24)$$

that is with (18)

$$\frac{d\rho}{dt}=\Delta\omega-\Delta\rho\omega\frac{i_d\cdot R_r}{\psi_r}-\Delta\psi_r\frac{i_q\cdot R_r}{\psi_r^2}. \qquad (25)$$

Now we can form the state-space model of the system for small deviations from (19), (22) and (25):

$$\frac{d}{dt}\begin{bmatrix}\Delta\psi_r\\ \Delta\omega\\ \Delta\rho\end{bmatrix}=$$

$$=\begin{bmatrix}-\frac{i_{mn}}{\psi_n}\cdot\left(p_1+(1-p_1)\cdot p_2\cdot\left(\frac{\psi_r}{\psi_n}\right)^{p_2-1}\right)\cdot R_r & 0 & i_q\cdot R_r\\ \frac{1.5\cdot P\cdot i_q}{J} & 0 & -\frac{1.5\cdot P\cdot\psi_r\cdot i_d}{J}\\ -\frac{i_q\cdot R_r}{\psi_r^2} & 1 & -\frac{i_d\cdot R_r}{\psi_r}\end{bmatrix}\times \qquad (26)$$

$$\times \begin{bmatrix} \Delta\psi_r \\ \Delta\omega \\ \Delta\rho \end{bmatrix} + \begin{bmatrix} 0 \\ -\frac{1}{J} \\ 0 \end{bmatrix} \cdot \Delta m_{load} \, .$$

Notice that the A matrix does not contain the speed, so the stability of the system does not depend on it.

In a given operating (steady state) point, which is given by ψ, m, ω the values in the matrix (from (3), (4) and (6)):

$$i_d = i_m = i_{mn} \cdot \left(p_1 \cdot \frac{\psi_r}{\psi_n} + (1 - p_1) \cdot \left(\frac{\psi_r}{\psi_n} \right)^{p_2} \right) , \quad i_q = \frac{m}{1.5 \cdot P \cdot \psi_r} , \quad (27)$$

hence the A matrix is available.

In the system matrix there are two parameters: the flux and the torque. We presume $\psi = \psi_{ref}$, which is set by the flux regulator below the field weakening range. Then stability can be studied by varying the torque. In *Fig. 1, 2* and *3*, the real parts of the eigenvalues of A in the function of the torque can be seen. The figures and the calculations were performed by a MATLAB program. It can be stated that over a certain torque the identification becomes unstable while the real component of the third eigenvalue is positive. With our motor this torque limit was 225% of the nominal torque.

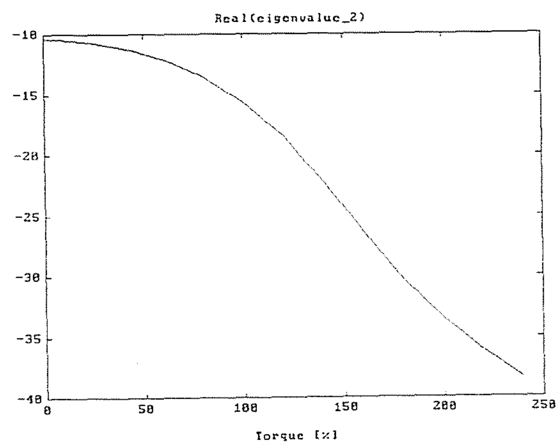

Fig. 2. The real part of the second eigenvalue

4. Stationer Condition

Now let us investigate the steady state errors in stationer condition in case of a load torque estimation error, so $\hat{m}_{load} \neq m_{load}$. If we assume that

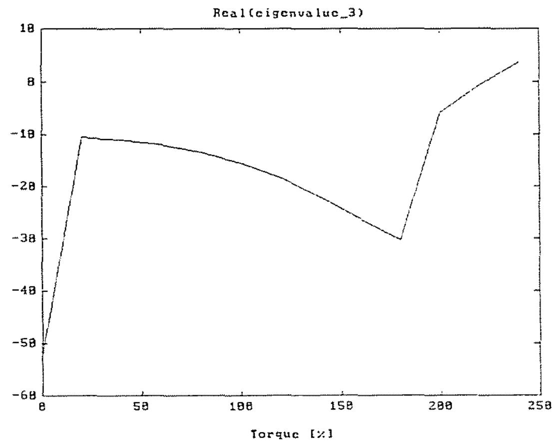

Fig. 3. The real part of the third eigenvalue

stationer condition exists, then $\frac{d}{dt}\begin{bmatrix}\Delta\psi_r\\ \Delta\omega\\ \Delta\rho\end{bmatrix} = 0.$

From (9) for small deviations of m_{load}:

$$\begin{bmatrix}\Delta\psi_r\\ \Delta\omega\\ \Delta\rho\end{bmatrix} = 0 = -\mathbf{A}^{-1}\cdot\mathbf{b}\cdot\Delta m_{load}\,, \tag{28}$$

We normalize the different errors: $\frac{\Delta\psi_r}{\Delta m}\cdot\frac{m_n}{\psi_n}$, $\frac{\Delta\omega}{\Delta m}\cdot\frac{m_n}{\omega_n}$, $\frac{\Delta\rho}{\Delta m}\cdot m_n\cdot\frac{180}{\pi}$ these deviations are seen in *Fig. 4, 5, 6*.

As we can see at *Fig. 4*, the amplitude of the estimated rotor flux is not highly affected by the deviation of load-torque when the load is small. For example at a load of 20% of the rated torque and in case of 1% torque deviation is of the rated torque, the error of the estimated amplitude of the rotor flux will be as small as 0.04%.

As we can see at *Fig. 5*, the deviation of load-torque has not significant effect on the estimated speed when the load is small. For example at a load of 20% of the rated torque and the torque deviation is 1% of the rated torque, the error of the estimated speed is 0.048%.

As we can see at *Fig. 6*, the phase of the estimated rotor flux is not highly affected by the deviation of load-torque when the load is small. For example at a load of 20% of the rated torque the torque deviation is 1% of the rated torque, the error of the rotor flux phase will be 0.41°.

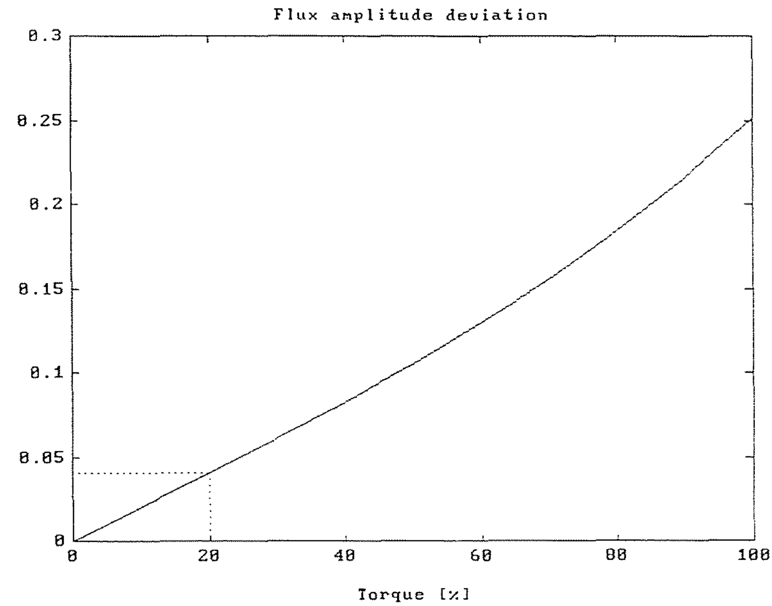

Fig. 4. The steady state error of the flux amplitude

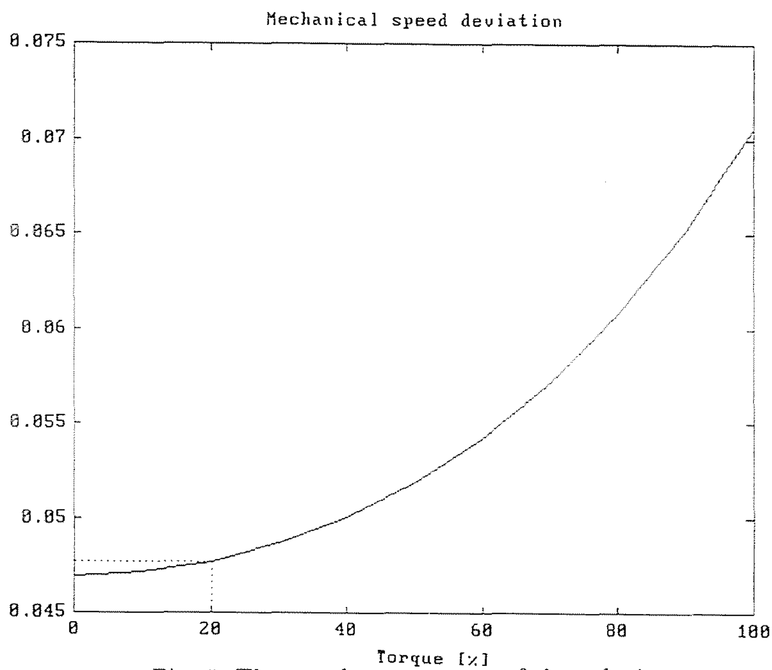

Fig. 5. The steady state error of the velocity

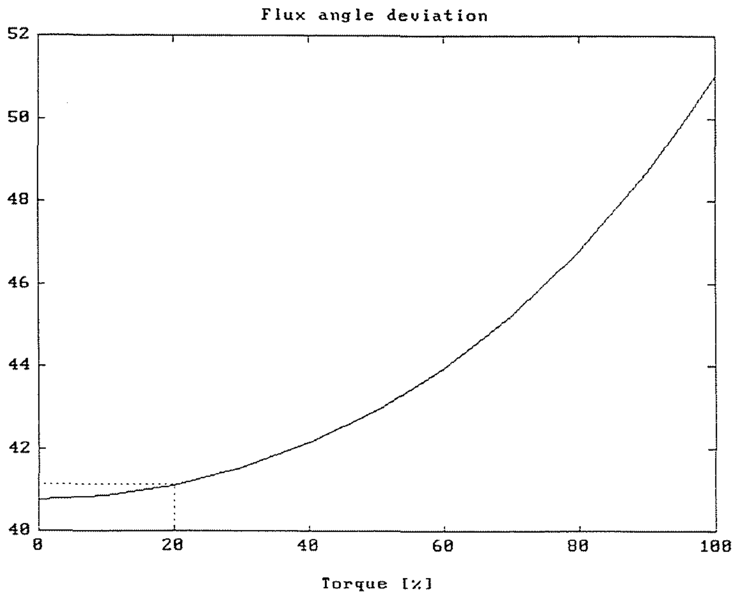

Fig. 6. The steady state error of the flux angle

5. Transient Behaviour

5.1. Estimations for Small Deviations

It is still not enough to investigate the proposed identification from the point of view of small signal stability, since in the most of applications there are large steps in the speed reference. When there is high acceleration or deceleration, two new problems of the identification will arise.

As it was shown in Chapter 3, the stability depends on the motor torque at the working point. As the torque is higher so the stability is worse, and at acceleration the static and dynamic load will be added, the stability will not be as good as it was estimated in Chapter 3.

Theoretically the identification will be unstable, when sum of the static and dynamic torque

$$m = m_{load} + J * \frac{d\omega}{dt} \tag{29}$$

will reach a value, where the real component of one of the eigenvalues of A is positive (See *Fig. 3*). In practice for small positive eigenvalue (comparing to 1/the duration of acceleration) the deviations of the identified state-variables at the end of the acceleration can roughly be estimated as

$$\Delta x \approx \Delta x_0 \cdot \lambda \cdot T_{acc} \,, \tag{30}$$

where Δx is the amplitude of the deviation in the direction of $\mathbf{x}$ eigenvector, Δx_0 is its initial value at the start of the acceleration, λ is the eigenvalue belonging to $\mathbf{x}$, T_{acc} is the duration of the acceleration.

An other problem arises at high acceleration or deceleration because of the uncertainty of the J inertia. In this case the deviation of the motor-torque is

$$\Delta m = \frac{\omega_{acc}}{T_{acc}} \cdot \Delta J , \tag{31}$$

where ω_{acc} is the change of the mechanical speed during T_{acc}. From this the resulted speed and flux deviations can be estimated according to Chapter 4.

5.2. Transient Simulation

We have developed a transient simulator to check the results of the analytical estimations of the chapters above, and investigate the dynamic behaviour of the identification for large steps and large deviations. The transient simulator was written in the C language and utilise the method of 4th order Runge-Kutta. Its output is a data file, which is compatible with the evaluation version of PSPICE, allowing detailed investigation of the resulted time-functions.

First we checked the analytical results of the stability analysis in Chapter 2. For this we used a model of an induction motor powered by unregulated symmetrical sinusoidal 3 phase voltage. The flow-chart of the identification can be seen at *Fig. 7*.

Finally we checked the dynamic behaviour of a complete control system, including the motor, the flux and speed identification and an indirect field-oriented controller. The flow-chart of the field-oriented current controller can be seen at *Fig. 9*.

As it can be seen in the flow-chart, we used only proportional controller. In spite of the lack of the integrator the current deviations a kept small, since we used feed-forward compensation from the estimated stator voltages valid for a working point:

$$\begin{aligned} uhd &= \frac{d\Psi}{dt} + Rs * id\_ref - \omega_s * ls * iq\_ref \\ uhq &= \omega_s * \Psi + Rs * iq\_ref + w_s * ls * id\_ref \end{aligned} \tag{32}$$

Since the proposed method is going to be used especially in the low speed range, we did not deal with field-weakening, so the reference of the flux is constant.

For the sake of the easy implementation we used a proportional flux controller, but to avoid large flux deviation we used also current feed-forward

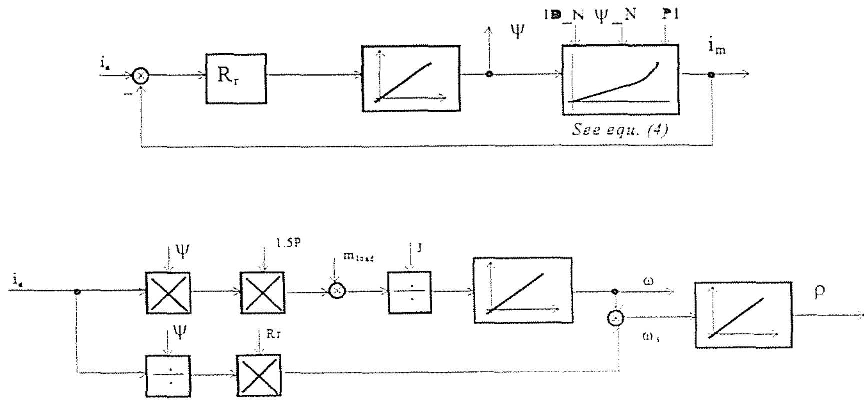

Fig. 7. Identification of rotor-flux. After the damping of the initial transients the identification was started with a small deviation in ρ. We found the stability border at a load equal 225% of the rated torque at rated rotor flux.. A little increase in torque caused instability as can be seen at *Fig. 8*, where the load was 230% (the plots are the flux amplitude error in [%], the flux angle error in degrees, the speed reference, the real and the estimated speed [rad/sec]).

from the magnetizing current (*Fig. 10*). With this method at steady state the output of the flux-controller is near zero.

The main purpose of the transient simulation was to check the dynamic behaviour of the identification at large steps in the speed reference. We also checked the dynamic behaviour for some uncertainty of the J inertia.

At *Fig. 11* there was no inertia deviation. The time-functions of the state variables are very similar as if the controller were working with speed-sensor. The result is unrealistically good. In a real system there are always disturbances at the current sensors and deviations of the motor parameters. The system has a pretty large positive feedback during acceleration, so a little deviation of the identified variables would be multiplied for the end of the acceleration.

We have also checked the transient behaviour of the controller, when the inertia used in the identification was different from the inertia used in the motor model.

In both cases the control system remained stable, although the estimated torque is roughly different from the real one.

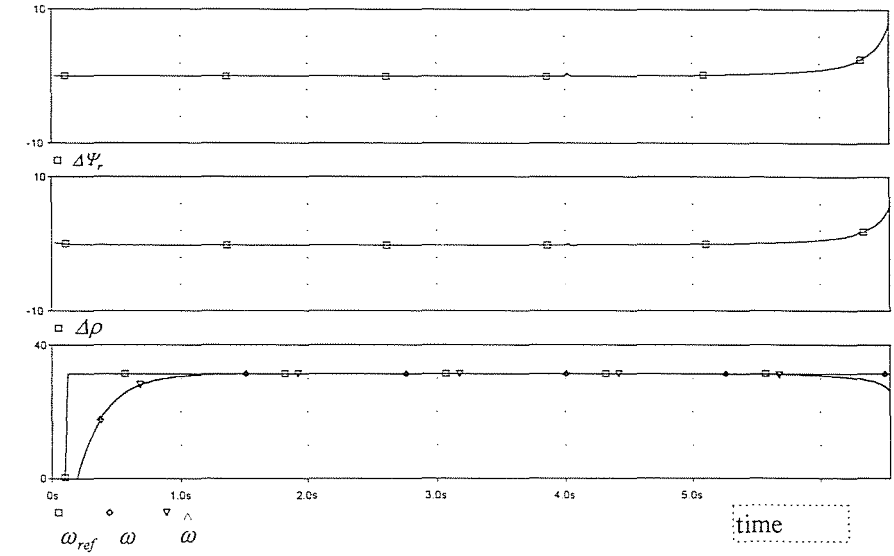

Fig. 8. Instability above the mentioned torque limit

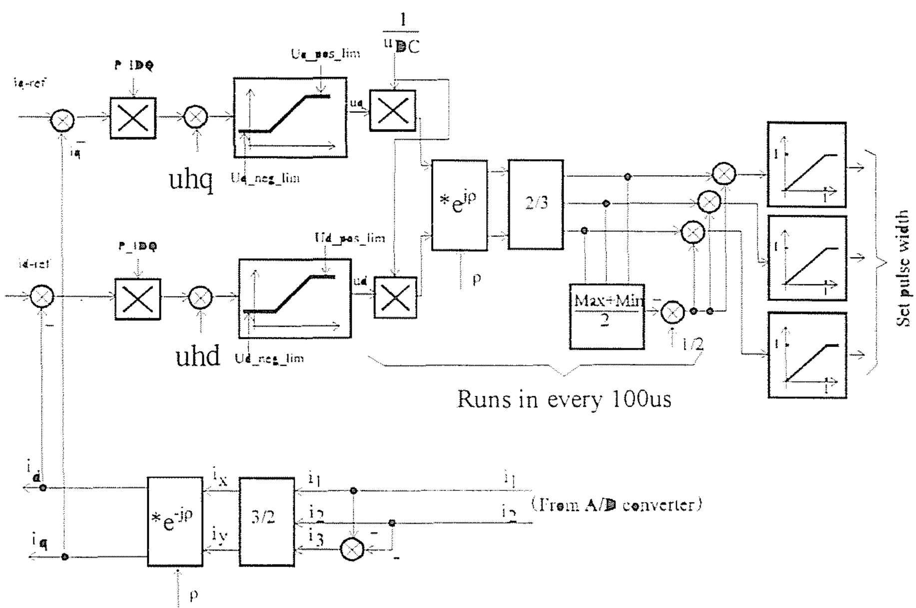

Fig. 9. Current control with voltage feed-forward and voltage-limitation

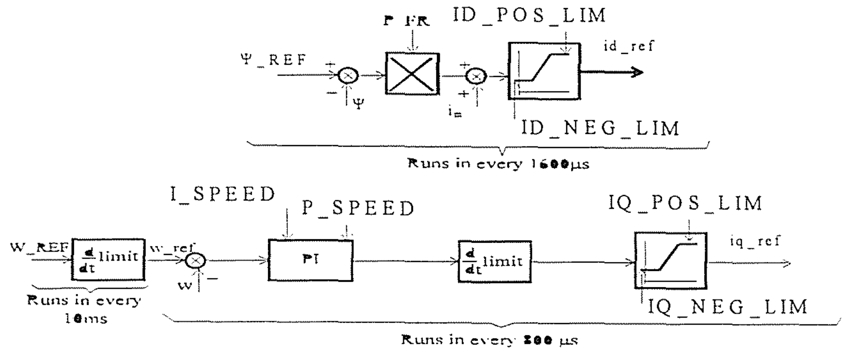

Fig. 10. Flux and speed control

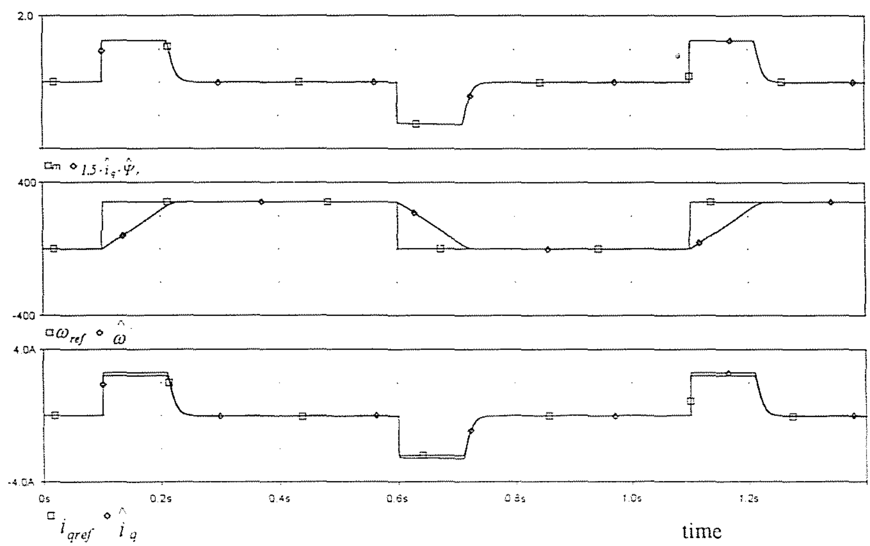

Fig. 11. Transient behaviour of the identification using 100% inertia

6. Experimental System

The control system and identification was implemented on a 16 bit micro-controller.

We used the uPD78P368 controller from NEC, which was designed directly to control three-phase inverters. The main features of the chosen controller are:

- built in three phase pulse-width modulator with dead-time generation,
- a 10 bit A/D converter with 8/1 multiplexer,

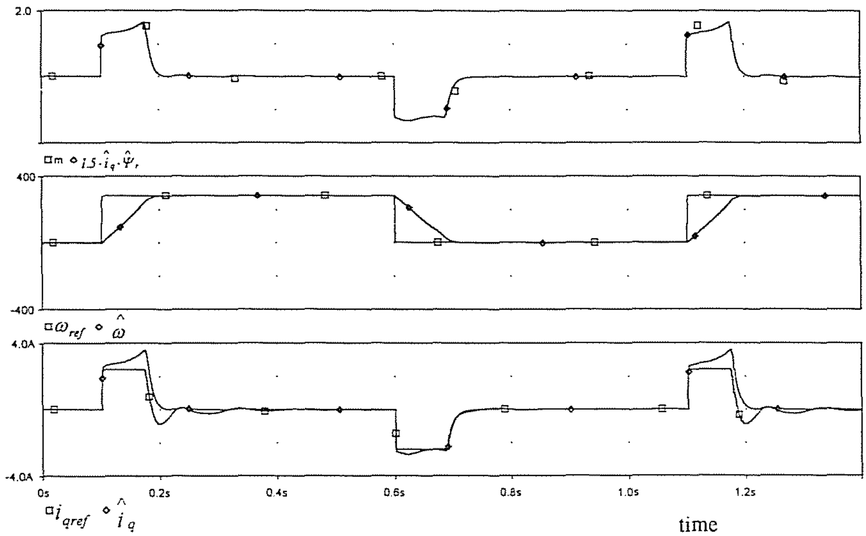

Fig. 12. Transient behaviour of the identification using 80% inertia

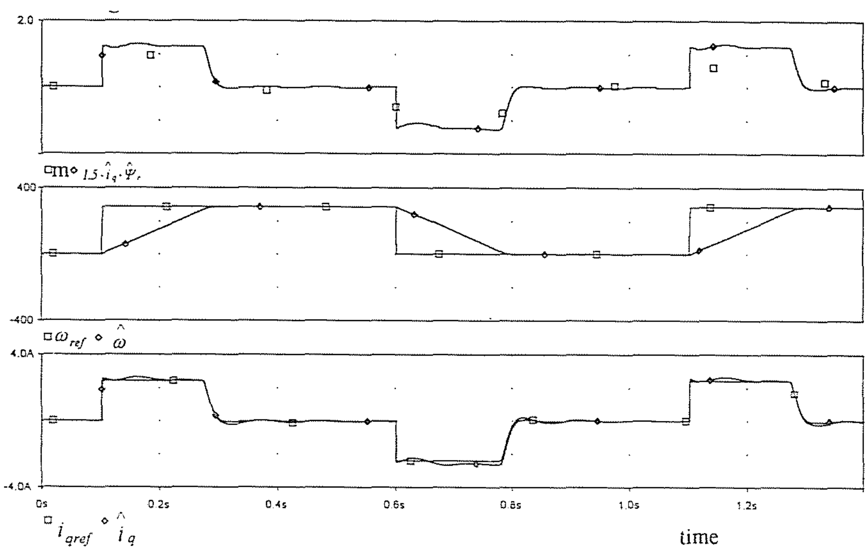

Fig. 13. Transient behaviour of the identification using 150% inertia

- a quick core with pipe-line structure (minimal instruction execution time is 188ns)
- 2 kbyte internal RAM can be used for time-critical programs,
- asynchronous and synchronous serial lines,
- macro facilities for data transfer without processor overhead.

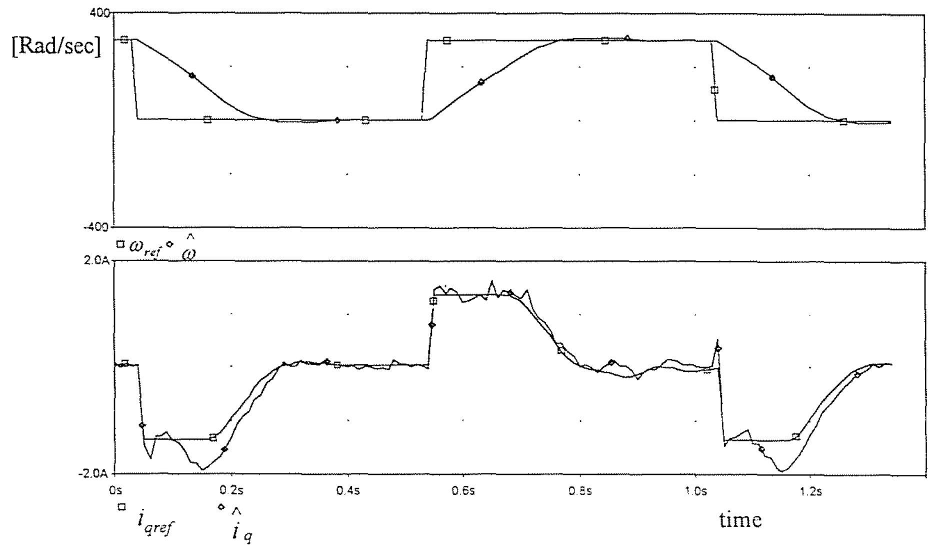

Fig. 14. Transient behaviour of the identification at the experimental system

(For other details see [5] and [6])

The time critical part of the software (including the identification , the flux, speed and current controllers) was programmed using assembly language.

Almost every part runs in every $400\mu s$, except for the speed and flux controller, which runs in every $800\mu s$ and $1600\mu s$ respectively and the PWM system which runs in every $100\mu s$. The flow-chart of the implemented system is equal to the system was used for the transient simulation.

The main circuit of the 1 kW inverter was made by D-Tech GmbH, Germany. The 1 kW motor was made by Hanning Elektowerke, Germany. The motor parameters can be found in the Appendix.

The inverter needed a relatively large, $2\mu s$ dead-time, which roughly disturbed the output voltage. An other difficulty raised because of the method of current sensing. The current was measured by 100 mV shunt-resistors with a common mode voltage jumping up and down with about 300V/100ns. In spite of the effort that was made to filter out noises from the current signals it still contains about 10% white noise.

In this environment we still could reach a stable, safe control system with a relatively quick response for speed reference steps.

The signals in *Fig. 14* were measured or estimated by the controller of the experimental system and was send through an RS232 serial line to an IBM PC, where the data were converted into a data file compatible with the evaluation version of SPICE.

7. Conclusions

We investigated the proposed method of identifying flux and speed both from the viewpoint of stability and the remaining errors. We found that the method is useful in low speed domain, when there is no large static load torque. The stability could be improved, when the motor is somewhat overexcited.

We suggest using the proposed method in the low frequency range, where currently this is the only stable system in noisy environment. At higher frequency an other observer like method (such as suggested in [4]) should be used. The border of the method is approximately 10-30% of the nominal speed, depending on the circumstances as noises in current and voltage signals, uncertainty of the motor parameters used in the identification.

8. Acknowledgement

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9. Appendix

Motor Parameters

ω_n = $2 * pi * 50.0$[rad/s], (Rated motor frequency)
P = 1.0[1],
ψ_{ref} = 0.326[Vs]
R_s = 1.236[Ω]
R_r = 1.417[Ω],
L_s = 8.777e-3[H],
ψ_n = 0.267[Vs], (parameter of flux model)
i_{mn} = 3.0874[A], (parameter of flux model)
$p1$ = 0.7832[1], (parameter of flux model)
$p2$ = 5.0[1], (parameter of flux model)
J = 0.001075[kgm^2],
m_n = 3.2[Nm] (nominal torque)

Nomenclature

u_s	is the stator voltage vector
i_s	is the stator current vector
y_r	is the rotor flux vector
i_x	is the real component of the stator current vector
i_y	is the imaginary component of the stator current vector
i_d	is the d component of the stator current vector
i_q	is the q component of the stator current vector
i_m	is the magnetising current
ψ_r	is the magnitude of the rotor flux
ρ	is the angle of the rotor flux
ω	is the angular rotor velocity
ω_s	is the angular stator velocity
m	is the electromagnetic torque
m_{load}	is the load torque
P	is the number of pole pairs
R_r	is the reduced rotor resistance (so that the rotor leakage inductance is zero)
R_s	is the stator resistance
L_s	is the reduced stator leakage inductance
J	is the moment of inertia
p_1, p_2	are parameters of the magnetising curve
^{n}	denotes to the nominal values
$\wedge$	denotes to the estimated values

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[5] uPD 78365 HW Description

[6] uPD 78365 Software Manual