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APPLICATION OF DIFFERENT METHODS FOR FLUX ESTIMATION IN INDUCTION MOTOR DRIVES

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Abstract: *Different methods for flux estimation in induction motor drives are presented in the paper. The aim of the paper is to present various solutions for flux estimation and to emphasize their advantages and disadvantages for particular application. A special attention is given for drives with flux control in close-loop (both stator and rotor), like direct torque control drives. All shown algorithms are analyzed using simulation models on computer and several of key methods are verified by experiments on a laboratory prototype.*

Key words: ***Estimation/Flux/Induction motor/Drives***

1. INTRODUCTION

Direct flux measurement in electrical drives is possible, but it request embedding special coils in machine or adequate Hall sensors. To avoid application of such a coils, a voltages induced in a parts of machine coils could be selected and summed in a such a manner that in a total there is no voltage drop on the stator resistance. Results are equivalent to those with the embedding of special coils, but it should be more taps created on the machine coils. All these are unacceptable for industrial applications, since it requires use of special motors and additionaly increases the costs of drive production. For that reasons, it is necessary to estimate the flux, what is now days is solved using appropriate software implemented in the control hardware.

2. FLUX ESTIMATION

Flux estimation is an important task in implementing high-performance motor drives [1]-[3]. There are, in general, two methods for flux estimation: one is based on measured motor currents, and the other is based on measured voltages [3]-[5].

In the current-based method, the motor air-gap flux is identified by solving a set of equations in which motor parameters are required as well as measured motor currents, speed, or position [3]. One of the problems associated with this method is that the parameters change with motor operating conditions, e.g., variations in rotor temperature and magnetic saturation level. In order to overcome this problem, an on-line motor parameter identification scheme should be implemented, which increases the complexity of the drive system. Furthermore, the motor speed or position has to be detected, which is undesirable practice in most industrial applications, since the use of tachometer will deteriorate the reliability of the drive.

In the voltage-based method, the motor flux can be obtained by integrating its back electromotive force (emf). The only motor parameter required is the stator winding resistance, which can be easily obtained. Taking into account the fact that the motor speed signal is not required, this method is much preferred [4],[5] and is given by the following relation:

$$\Psi_s = \int (u_s - R_s \cdot i_s)dt \qquad (1)$$

However, implementation of an integrator for motor flux estimation is not an easy task. A pure integrator has dc drift and initial value problems. A dc component in measured motor back emf is inevitable in practice. This dc component, no matter how small it is, can finally drive the pure integrator into saturation. For that reasons, a different methods of compensation are applied in order to resolve all drawbacks of pure integrator and to achieve better accuracy [6], without increase of the implementation complexity or to violence overall drive dynamics. In the paper a detail analysis of several solutions are given, with the indication of advantages and drawbacks [7]-[16].

The initial value problem associated with the pure integrator can be explained as follows. When a sine signal is applied to the integrator, a cosine wave is expected at its output. This is true only when the input sine wave is applied at its positive or negative peak. Otherwise, a constant dc offset will appear at the output. A common solution to these problems is to replace the pure integrator with a first-order low-pass (LP) filter. Obviously, the LP filter will produce errors in magnitude and phase angle, especially when the motor runs at a frequency lower than the filter cutoff frequency. Results of initial value influence on pure integrator and LP filter, obtained by simulation, are shown in Fig. 1.

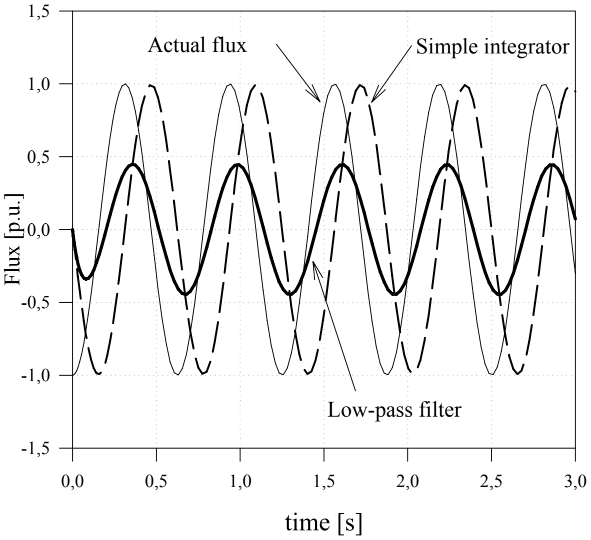

Fig. 1. *Initial value problem in the flux estimation*

The response of the pure integrator to the sine input with a dc drift obtained by simulation is shown in Fig. 2.

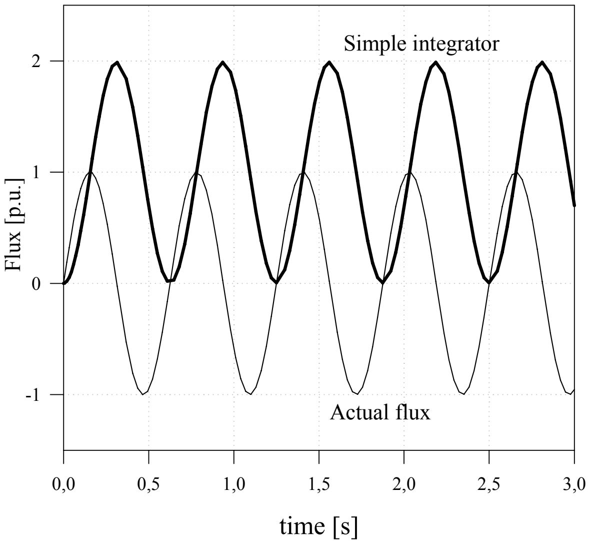

Fig. 2. *Initial value problem in the flux estimation*

The response of the pure integrator to the step change of flux magnitude is shown in Fig. 3.

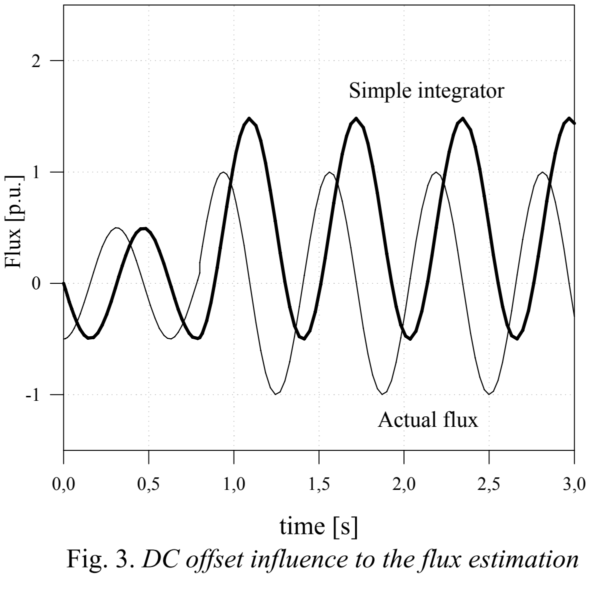

Fig. 3. *DC offset influence to the flux estimation*

3. DIFFERENT ESTIMATION APPROACHES

As explained in previous section, simple integrator is not useful for flux estimator due to the initial value and dc offset problems. Solutions such as modified integrators proposed in [5] are good for improving estimation, but shows moderate accuracy in closed-loop flux drives, such as DTC drives. Algorithm that uses amplitude limiter in the compensation feedback is not useful for DTC drives, since limit could not be keep at the same level in order to obtain fast flux and torque response. This could be seen in Fig. 4, where flux response is shown during speed change from 150 rpm to 600 rpm. The response is rather slow, but also in new steady-state estimated flux has dc offset.

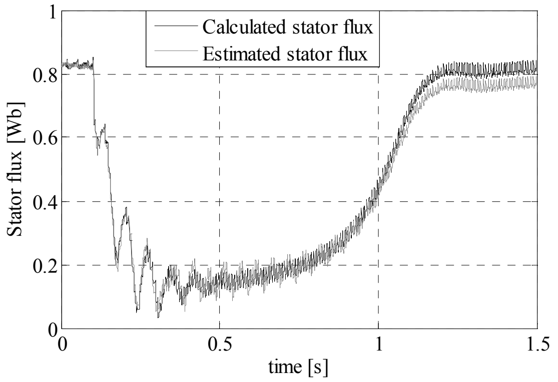

Fig. 4. *Flux response on speed change*

The next algorithm, proposed in [5] as modified integrator with adaptive compensation, is designed for the drives with variable flux operation. Although this would be the correct solution in principle, the magnitude of the induced voltage becomes extremely small at very low speed, which makes dc offset and other disturbances the dominant signals. Another major drawback of this method is the dynamic delay of the closed-loop control used for error correction. This delay generates dynamic errors at transient conditions. Possibly for this reason have the authors applied their methods only for flux monitoring, but not for field-oriented control in a closed loop. The rotor flux response of the CSI drive with this algorithm used for flux estimation is presented in Fig. 5.

Result taken from the real drive presented in [16] are compared to the calculated at low speed about 100 rpm. Flux oscillations are influenced by the values of compensation parameters and it could not be easily tuned if drive runs at low speeds.

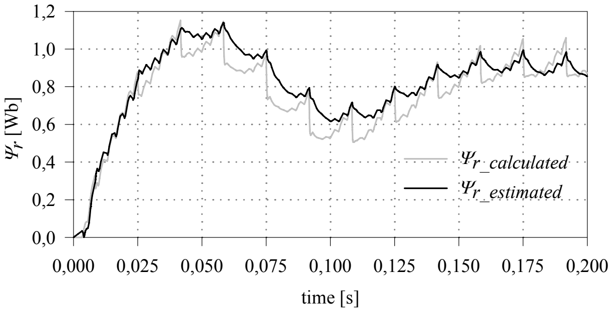

Fig. 5. *Rotor flux response of the modified integrator with adaptive compensation*

Kubota et al. propose to estimate the dc offset using a full-order observer [8]. They exploit the fact that oscillations in speed and rotor flux magnitude occur in the presence of dc offset. The approach is highly computational. It requires computing the average values within a fundamental period of the estimated rotor flux components in stationary coordinates. These values are subsequently multiplied by coefficients derived from the system matrix, which in turn depend on the estimated speed. The results are summed up to yield increments of the estimated offset voltage components.

Also, Rodic et al. [9] use the deviations of the estimated rotor flux magnitude from its reference value to build a nonlinear flux observer. This observer is constructed as a secondorder low-pass filter at low stator frequency; it converts to a first-order low-pass filter at higher stator frequency. The experimental results obtained with this method demonstrate moderate performance.

In order to overcome these drawbacks of several estimation techniques, Holtz in [13] proposes estimation of the stator voltage vector with a precise dc offset estimator. This estimator is based on the fact that the trajectory of flux vector is not circular in the presence of dc offset. Since its undisturbed radius equals reference stator flux, the offset components tend to drive the entire trajectory toward one of the reference flux boundaries.

A contribution to the EMF offset vector can be estimated from the displacement of the flux trajectory [13], as:

$$EMF_{\alpha\beta}^{off} = \frac{1}{\Delta t} \cdot \left(\Psi_{\alpha\beta_{\max}} + \Psi_{\alpha\beta_{\min}}\right) \qquad (2)$$

where the maximum and minimum values in (2) are those of the respective components $\Psi_{s\alpha}$ and $\Psi_{s\beta}$, and Δt is the time difference that defines a fundamental period. The signal EMF^{off} is fed back to the input of the integrator so as to cancel the offset component in EMF. The input of the integrator then tends toward zero in a quasi-steady state, which makes the estimated offset voltage vector equal the existing offset. The trajectory of Ψ_s is now exactly circular, which ensures a precise tracking of the EMF offset vector. Since offset drift is mainly a thermal effect that changes the dc offset very slowly, the response time of the offset estimator is not at all critical. It is important to note that the dynamics of stator flux estimation do not depend on the response of the offset estimator.

This algorithm could be easily modified and applied to CSI drive with direct torque control [16]. Modification shows some more advantages, due to the fact that CSI drive has lower switching frequency and there is no need for current zero detection as in VSI application shown in [13]. Rotor flux response to the flux reference equal to nominal flux is shown in Fig. 6.

The better observation of this flux estimator could be seen in Fig. 7 and Fig. 8. Fig. 7 shows rotor flux trajectory after motor start from standstill, while Fig. 8 presents the influence of the offset compensator to the determination of rotor flux position.

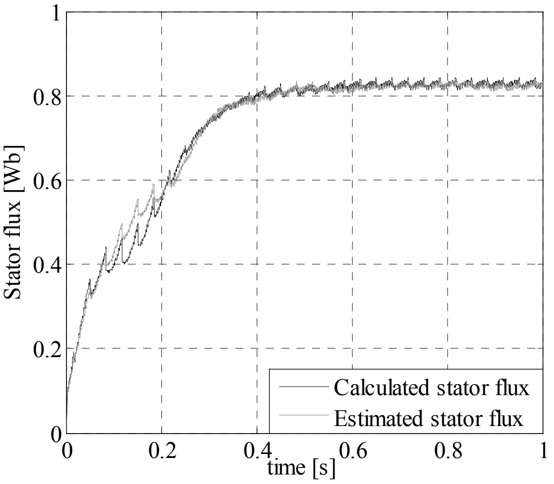

Fig. 6. *Rotor flux response of the modified Holtz algorithm*

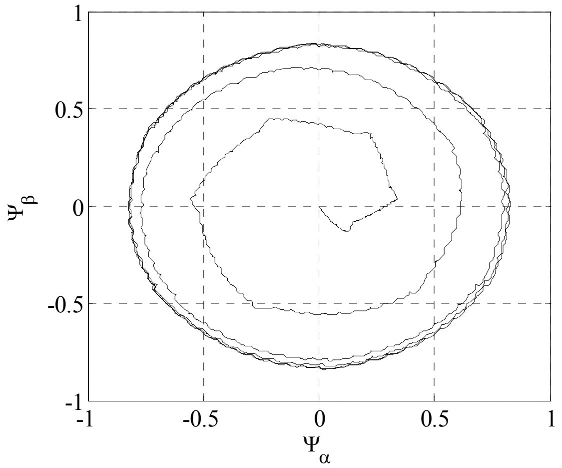

Fig. 7. *Rotor flux trajectory during motor start*

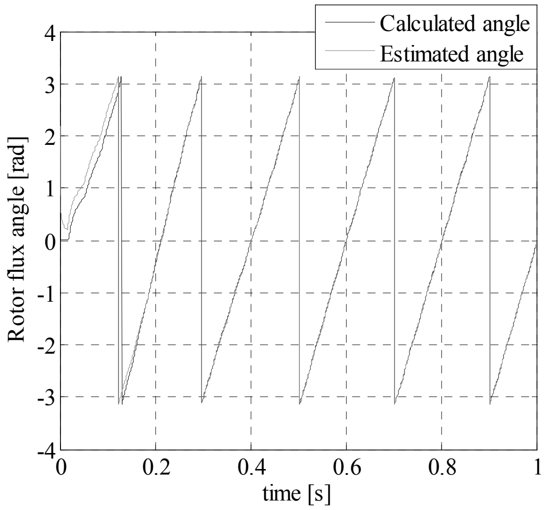

Fig. 8. *Offset compensation of the estimated rotor flux angle*

Some problems could be also arised if the signal on the offset estimator output is not filtered. This is due to fact that in CSI drive current waveform is trapezoidal and there are significant voltage spikes at time instants during thyristor switching.

The same flux response presented in Fig. 6 but without mentioned filter is shown in Fig. 9. It could be also observed that there is a greater flux disturbance at 0.1 s than in the previous case in Fig. 6.

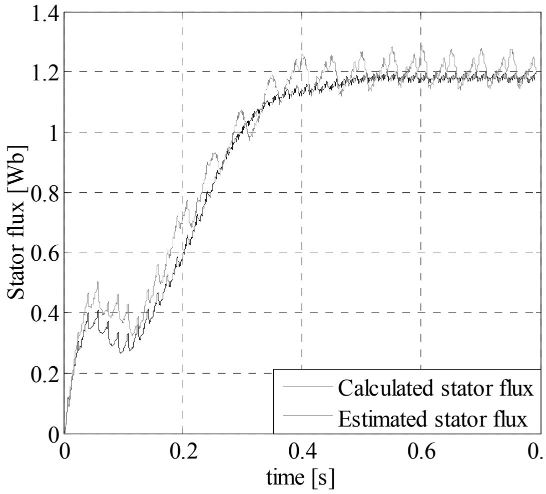

Fig. 9. *Rotor flux response of the modified Holtz algorithm without estimated offset filtering*

Observers are also very popular for flux estimation [10]-[12],[14],[15]. The flux observer is aimed at determining the actual stator and rotor flux vectors by using the stator current and voltage measurements. A closed-loop estimator is based on the principle that feeding back the difference between the measured output of the observed system and the estimated output, and continuously correcting the model by the error signal, the estimate error should be minimized.

In the case of a flux estimator, the motor flux cannot be directly measured, but the idea of realizing a closed-loop system is still applicable if the difference between the reference rotor flux vector and the estimated rotor flux vector is used as feedback signal. Such a solution is proposed in [10], where the behavior of the flux observer can be described by the following equations:

$$\frac{d\overline{\Psi}_s}{dt} = \overline{u}_s - R_s \cdot \overline{i}_s + G \cdot \left(\Psi_r^* \cdot e^{j\theta} - \overline{\Psi}_r\right) \qquad (3)$$

$$\overline{\Psi}_r = \frac{L_r}{L_m} \cdot \overline{\Psi}_s - \frac{\sigma \cdot L_s \cdot L_r}{L_m} \cdot \overline{i}_s \qquad (4)$$

where θ is the rotor flux phase angle, *G* is the gain of the flux observer and Ψ_r^* is the reference rotor flux. This observer is much more robust than an open loop estimator, and shows a performance that is comparable to that of much more time-consuming observers, such as full-order observers and Kalman filters.

Similar solution are proposed by Lascu et al. in [12], but with observer developed as a combination of voltage and current model. Output of the current model operated in open loop is stator flux:

$$\overline{\Psi}_s^i = \frac{L_m}{L_r} \cdot \overline{\Psi}_r^i + \frac{L_s \cdot L_r - L_m^2}{L_r} \cdot \overline{i}_s \qquad (5)$$

Voltage model uses measured stator voltage and current, so vector of the stator flux is:

$$\overline{\Psi}_s = \int \left(\overline{u}_s - R_s \cdot \overline{i}_s - \overline{U}_{comp}\right) \qquad (6)$$

where U_{comp} is PI based compensation signal:

$$\overline{U}_{comp} = \left(K_P + K_I \cdot \frac{1}{s}\right) \cdot \left(\overline{\Psi}_s - \overline{\Psi}_s^i\right) \qquad (7)$$

Compensator coefficients K_P and K_I are calculated in such a manner that at zero frequency only current model is active, while at higher frequencies only voltage model is operating:

$$K_P = \omega_1 + \omega_2,\ K_I = \omega_1 \cdot \omega_2 \qquad (8)$$

Poles ω_1 and ω_2 are choosen usualy in practice so there are no abrupt changes between two models (ω_1 = 2 - 5 rad/s, ω_2 = 20 - 30 rad/s).

Fig. 10 shows a block diagram of the flux observer based on (3) and (4).

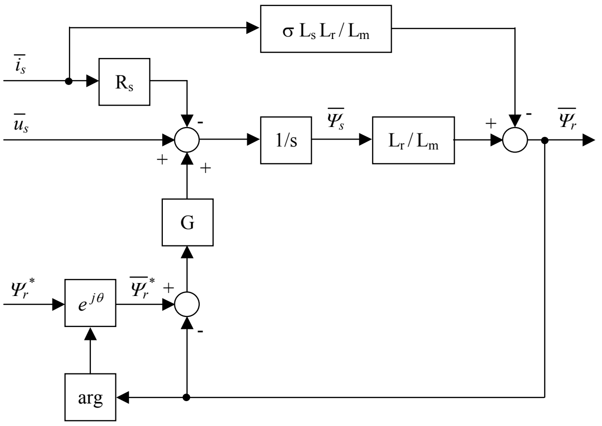

Fig. 10. *Observer based flux estimator*

Although these observers shows a excellent performance and parameter sensitivity [11],[12], they are not a good solution for a low output inverter frequency drive like CSI, what is presented in Fig. 11 using standard deviation of the estimated flux.

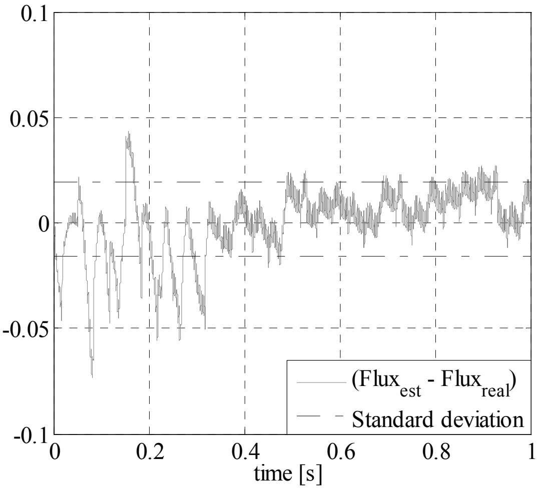

Fig. 11. *Deviation of the estimated flux amplitude*

Most flux estimators are working without sensor on the motor shaft, but in [15] authors proposed DTC algorithm with flux estimator based on rotor model that uses measured stator current and motor speed obtained from the incremental encoder. Although authors presented good experimental results, possible problems could be at low speed when influence of the temperature changes on rotor time constant are significant. The other errors could be expected at higher speeds when slip is small, so small error in a measured speed would result in a large error in magnetizing current (both amplitude and angle).

Finally, now days a powerful simulation software packages like Matlab/Simulink are used for algorithm development, what could yield to some errors in analysis. Most authors uses simple integrator as a flux estimator in their control algorithms like FOC and DTC with preset sample time of 2μs. Although results are correct, in a practical realization it could not be expected to have such a low value of sample time. In the only commercial available industrial DTC drive, sample time of the control algorithm is 25μs [17]. It could be shown in Fig. 12 that even in simulation flux estimation could have significant disturbances after setting torque reference different from zero, due to the higher values of integrator sample time than 20μs.

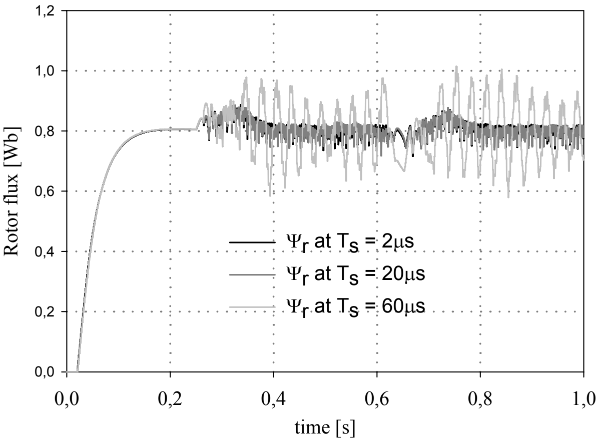

Fig. 12. *Sample time influence on the flux estimation*

4. CONCLUSION

Several mostly used flux estimator techniques are proposed in the paper. A detail analysis is given, with the indication of advantages and drawbacks using simulation and experimental results. Also, it is noted that some algorithms, although with a superior performance capabilities, are not good for application in every drive.

The final aim of the paper is to emphasize importance of analysis when and where to apply particular flux estimation algorithm.

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