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# SPEED ESTIMATION METHOD FOR AC DRIVES

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Abstract: Two structures of speed estimator for AC drive system are developed in this paper. The mathematical models and the simulation results, via numerical simulation, are presented. The utility of the estimator is important for the speed control or for the advanced control synthesis such as optimal vector control of induction motor drives. Besides of speed estimation objective, the proposed method increases the robustness of estimators to measurement signal noises.

Keywords: speed estimation, first order estimator, second order estimator, rotor field oriented control, mathematical model, AC drives.

## 1. INTRODUCTION

The modern control of AC drives requires more and more information about system states and perturbations. The most important perturbation of a drive is the load torque. The load torque can be measured by using adequate equipment (torque sensor) or can be estimated from a torque estimator. The problem of speed estimation appears at the electric drives with the different speed operation than the synchronous one (Holtz, 1996). Compared to measuring speed, the estimation speed brings some important advantages such as increased reliability and reducing cost. For advanced electric drives (vector control) accurate speed estimation is necessary for a wider speed range of operation than the scalar controlled PWM inverter drives. Various speed estimation methods have been developed for sensorless control of induction motor drives (Mostafa, *et al.*, 2009; Jezernik, *et al.*, 2003; Hossein, 2009; Haron and Idris, 2006; Kim *et al*, 1994; Proca, and Keyhani, 2007). There are two basic strategies for speed estimation: the first is to use the mathematical model (Holtz,1996; Jezernik, *et al.*, 2003; Bose, et al., 1995) and the second relies on saliency within induction machine (based on the space harmonics) (Xepapas, *et al.*, 2003; Goran, *et al.*, 2009; Jansen and Lorenz, 1995). The former method depends on the model accuracy having low performances at low rotor frequencies. The later method had obtained very good performance at low speeds. In order to implement this method a complicated hardware is needed for measuring and filtering the stator currents. In order to estimate the speed advanced control techniques were taken into account: model reference adaptive systems (Hossein, 2009; Haron and Idris, 2006; Kalman filtering methods (Kim, et al., 1994); sliding mode (Proca, and Keyhani, 2007; Mezouar, et al., 2008; Goran, et al., 2009) or artificial intelligent methods (Abbou, and Mahmoudi, 2009). Speed estimation plays a key role in sensorless drive. This paper presents simple speed estimation method for vector control drive in rotor field coordinates. For constant flux operation of the induction motor the active stator current is proportional with rotor frequency. Therefore, the torque component of the stator current, $i_{sq}$, can be used as one of the speed estimator input. The

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magnitude of the stator currents depends on the load conditions. Hence, the second input of the speed estimator is the applied load torque. The direct measurement of the angular speed is possible by using a speed sensor of from an encoder. By using an encoder, the accurate position can be obtained but expensive equipment is necessary. The second possibility consists of the integration of the mathematical model (Rosu, and Gaiceanu, 1998). Mathematical models of AC drives of higher order were developed including the nonlinearities. Some authors have been used different linearization methods in order to use these models for speed estimation (Barada, *et al*, 2002). The major disadvantage is that the models integration needs a strong hardware as a digital signal processor.

The proposed method is based on the mathematical model of the induction motor in rotor field coordinates. By using an adequate control ($i_{sd}$=ct), the nonlinear model becomes linear one which is more adequate for speed estimation purposes. Estimators of 1[st] and 2[nd] order are used, which can be implemented easier, because they do not need so complex calculus. The proposed method is distinguished from the others by performance and simplicity. Compared to other estimation methods (Mostafa, *et al.*, 2009; Barada, *et al*, 2002) the proposed estimators assure high accuracy of the estimated speed over a large range, in stationary and dynamic regimes, especially for low rotor frequency operation.

## 2. THE MODEL OF THE DRIVE SYSTEM

The well-known model of the AC drive using an induction machine in rotor field coordinates supplied

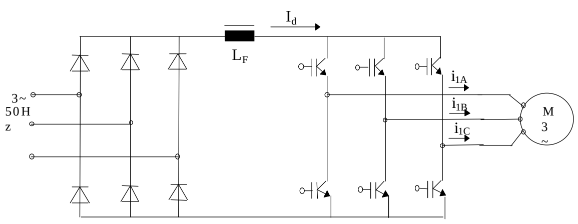


Fig.1. Three-phase squirrel cage induction motor supplied

from the current inverter (Fig. 1), is of the form (Murphy, *et al.*, 1998), (Rosu, *et al.*, 1998):

$$(1)\quad \begin{cases} \tau_R \dfrac{di_{mR}}{dt} + i_{mR} = i_{sd} \\ m_e = \dfrac{2}{3} p \dfrac{M}{1+\sigma_R} i_{mR} \cdot i_{sq} \\ J \dfrac{d\omega_m}{dt} = m_e - m_S \\ \dfrac{dq}{dt} = \omega_m + \dfrac{i_{sq}}{\tau_R i_{mR}} \end{cases}$$

where:

$i_{sd}$ the flux component current;
$i_{sq}$ the torque component current;
$i_{mR}$ the rotor magnetizing current;
$\omega_m$ the instantaneous angular velocity of the motor;
$m_e$ the electromagnetic torque;
$m_s$ the load torque;
$J$ the combined inertia of the motor and load;
$M$ the mutual inductance between the stator and the rotor d, q equivalent windings;
$\tau_R$ the rotor time constant;
$\sigma_R$ the rotor leakage factor;
$p$ the number of the pole pairs;
$q$ the rotor field angle.

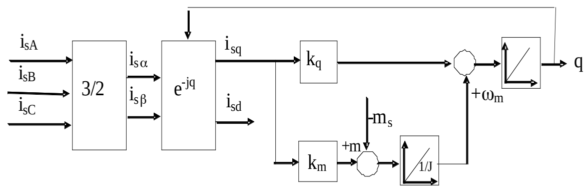


Fig.2. The model of the induction machine

There are two different regions for the control of the angular speed: the constant flux and the variable one. In the first case, the constant flux, $i_{sd}$, the flux current component, has a constant value corresponding to the rating flux. This means that the magnetizing current $i_{mR}$ has also a constant value and the system (1) gets the form:

$$(2)\quad \begin{cases} \dfrac{dq}{dt} = \omega_m + k_q \cdot i_{sq} \\ J \dfrac{d\omega_m}{dt} = k_m \cdot i_{sq} - m_S \end{cases}$$

where

$$(3)\quad k_m = \frac{2}{3} p \frac{M}{1+\sigma_R} i_{mR}$$

and

$$(4)\quad k_q = \frac{1}{\tau_R i_{mR}}$$

The block diagram of the model (2) is presented in (fig.2) where:

$$(5)\quad \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \end{bmatrix} = \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \cdot \begin{bmatrix} i_{sA} \\ i_{sB} \\ i_{sC} \end{bmatrix}$$

and

$$(6)\quad \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} = \begin{bmatrix} \cos q(t) & \sin q(t) \\ -\sin q(t) & \cos q(t) \end{bmatrix} \cdot \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \end{bmatrix}$$

Obviously, provided by adequate control of the current inverter, the condition $i_{sd}=ct$ can be accomplished.

## 3. THE SPEED ESTIMATOR OF THE SECOND ORDER

The block diagram of the second degree estimator is presented in (Fig. 3), where the input needs the measure of the load torque $M_s(s)$ and the calculus of the torque current component, $i_{sq}$., from the measured stator current. The output of the estimator is the estimated angular speed $\hat{\Omega}(s)$. Therefore, having the transversal current component of the stator current and load torque signals from the adequate transducers the angular speed can be estimated properly. The parameters of the estimator to be determined are $c$ and $\tau$.

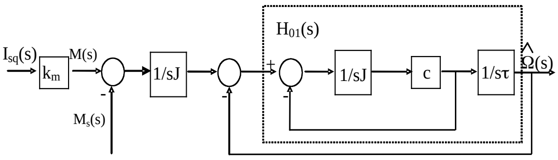


Fig.3. Speed estimator of the second order

The estimator (Fig. 3), after some manipulations, gets the form presented in (Fig. 4).

Using Laplace transform the second equation from (2) gets the form

$$(7)\quad sJ\Omega_m(s) = M(s) - M_s(s)$$

or in another form

$$(8)\quad \Omega_m(s) = \frac{1}{sJ}\left[M(s) - M_s(s)\right]$$

The closed loop transfer function, $H_{01}(s)$, shown in the dashed rectangle from Fig. 3, after some manipulations become:

$$(9)\quad H_{01}(s) = \frac{c}{sJ + c}$$

This means that the block diagram from (Fig. 3) can be redrawing as in (Fig. 4).

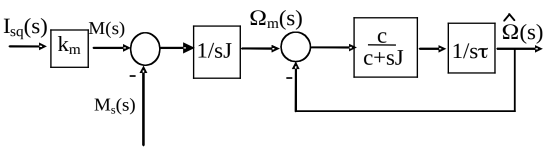


Fig.4. Final form of the angular speed estimator

The output of the second order angular speed estimator is obtained from:

$$(10)\quad \hat{\Omega}(s) = \frac{1}{sJ}\left[M(s) - \hat{M}_s(s)\right]$$

or from

$$(11)\quad \hat{\Omega}(s) = \Omega_m(s)\frac{1}{s^2\frac{\tau J}{c} + s\tau + 1}.$$

## 4. CALCULUS OF THE PARAMETERS

The problem consists of the calculation of the parameters $c$ and $\tau$ such that the error between the estimated angular speed $\hat{\Omega}(s)$ and the actual angular speed $\Omega_m(s)$ to be insignificant. The transfer function of the estimator, (Fig.4), is given by:

$$(12)\quad G(s) = \frac{\hat{\Omega}(s)}{\Omega_m(s)} = \frac{1}{s^2\frac{\tau J}{c} + s\tau + 1}$$

Considering a step variation for the $\Omega_m(s)$, by setting:

$$(13)\quad \Omega_m(s) = \frac{\Omega_m}{s}$$

the estimated angular speed get the form

$$(14)\quad \hat{\Omega}(s) = \Omega_m \frac{1}{s(s^2\frac{\tau J}{c} + s\tau + 1)}$$

The usual form of the equation (11) is given by

$$(15)\quad \hat{\Omega}(s) = \Omega_m \frac{1}{s(T_0^2 s^2 + 2\xi T_0 s + 1)}$$

where the damping factor is:

$$(16)\quad \xi = \frac{\tau\cdot\omega_0}{2}$$

and the pulsation factor

$$(17)\quad \omega_0^2 = \frac{1}{T_0^2} = \frac{c}{\tau\cdot J}$$

The parameters $c$ and $\tau$ are chosen such as that the response $\hat{\Omega}(s)$ to have an acceptable overshoot

$$(18)\quad \sigma = e^{-\frac{\pi\xi}{\sqrt{1-\xi^2}}},$$

a small step time response

$$(19)\quad t_a = \frac{\ln(0.05\sqrt{1-\xi^2})}{-\xi\cdot\omega_0^2},$$

and a minimum output noise (Calin S., *et al.*, 1980), (Goodwin, *et al.*, 2001).

## 5. THE SPEED ESTIMATOR OF THE FIRST ORDER

The second order estimator has a good behavior regarding to measurement noises, but it implies a relative complex calculus.

The order of the estimator can be reduced as in Fig. 5.

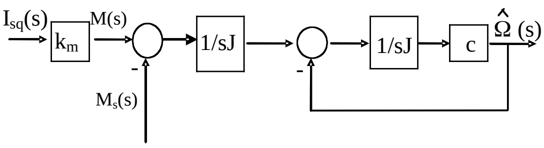


Fig.5. The speed estimator of the first order

After some manipulations, like for the 2nd degree estimator, the estimator gets the form shown in (Fig.6).

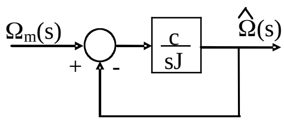


Fig.6. The simplified block diagram

The response to a step input $\Omega_m(s)$, Fig.7, is given by

$$(20)\quad \hat{\Omega}_s(s) = \Omega_m \frac{1}{s(sT_E + 1)}$$

where

$$(21)\quad T_E = \frac{J}{c}$$

is the time constant of the estimator. By adequate choosing of the c parameter, respectively time constant $T_E$, an acceptable step response $\hat{\Omega}(s)$ can be obtained.

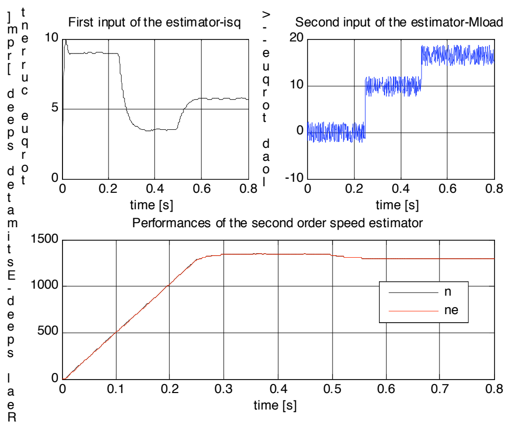


Fig.7 The input signals ($i_{sq}$ and $m_s$), and the comparative results of the real speed (n) with the estimated one (ne) of the second order speed estimator.

Using Z-transform and zero order hold the estimated speed can be calculated. The 1st degree estimator is simpler, but the filtering of the noise is not so good.

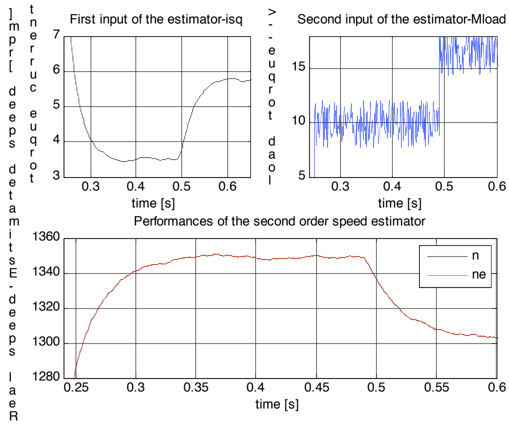


Fig.8 More details of the inputs ($i_{sq}$ and $m_s$), and the output of the second order speed estimator.

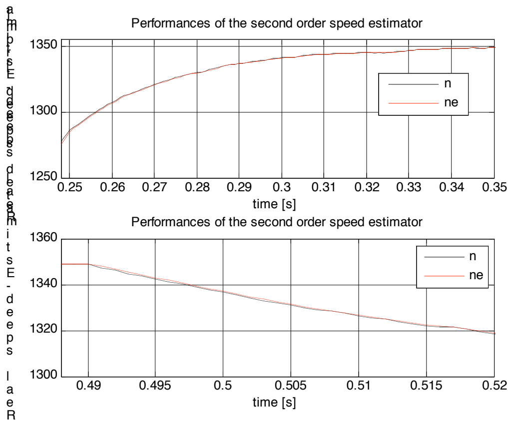


Fig. 9. The actual speed (n) and the output response (ne) of the 2nd order speed estimator

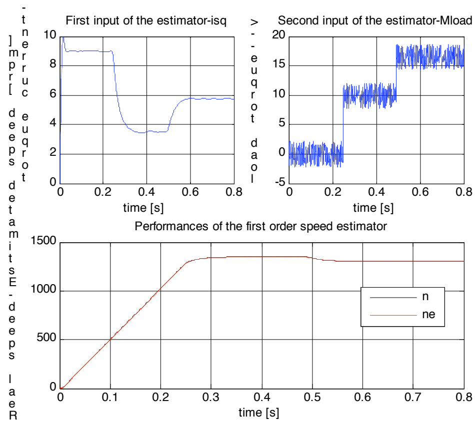


Fig.10 The input signals ($i_{sq}$ and $m_s$), and the comparative results of the real speed (n) with the estimated one (ne) of the first order speed estimator

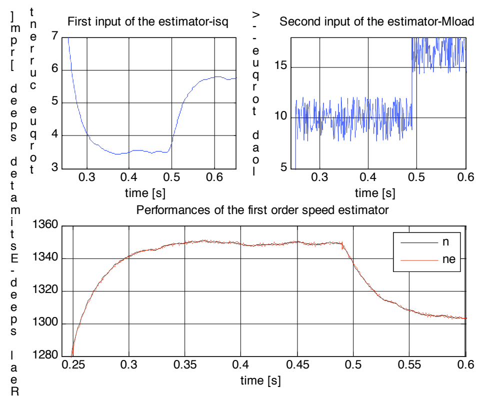


Fig.11 More details of the inputs ($i_{sq}$ and $m_s$), and the output of the first order speed estimator.

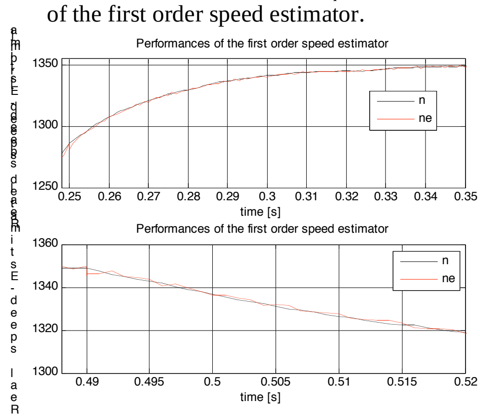


Fig. 12. The actual speed (n) and the output response (ne) of the 1st order speed estimator

## 6. SIMULATION RESULTS

The 1st degree and 2nd degree estimators were numerically simulated for an induction motor drive system K100L-4 FRAME, IEC TYPE, 2.2 KW, 1420 RPM, 3PHASE, 4.8AMPS, 14.81 Nm rated load torque, J=0.1kgms^2, VALIADIS MANUFACTURER.

The numerical simulations were done for a starting with no-load, under normal torque operation (10Nm, load torque being applied at 0.25s) and for overload (18 Nm at 0.49s) conditions for both estimator types: second (Fig.7-9) and first orders (Fig.10-11). It is well-known that the load torque signal is rich in noises (Fig.8). The parameters of the second order estimator were chosen taking into consideration 0.707 damping factor, and 0.1 ms sampling time.

In these conditions the following estimator parameter values were resulted: $\tau = 1{,}4 \cdot 10^{-4}$ and c=148.

Fig. 7 shows the input signals (the active current, $i_{sq}$, and the load torque, $m_s$) of the second order speed estimator and the comparative results of the real speed (n) with the estimated one (ne).

A detailed figure of the above mentioned signals are shown in Fig.8. In Fig. 9 the actual speed (n) and the output response (ne) of the 2nd order speed estimator are shown.

The output of the 1st degree estimator and to an angular speed step is shown in Fig. 8. The 34.88 of the *c* parameter value of the first order estimator has been taken for $T_E$= 0,3ms time constant. The same simulations with the second order speed estimator were done under the same load torque conditions for the first order speed estimator. In Figs.10-12 the simulation results of the first order speed estimator are presented. The error of speed estimation is obtained from:

$$(22)\quad \varepsilon(k) = \Omega_m(k) - \hat{\Omega}(k)$$

Analyzing the results obtained from both estimator types (Fig.10 and Fig 12) the lower estimation error is obtained for the second degree speed estimator. The 1st degree estimator is simpler, but in the first moments at applying a step load torque (at 0,25s and 0,49s) the speed error estimation is higher than in the second order speed estimator case. The increased robustness to load torque sensor noises is obtained.

## 7. CONCLUSIONS

The numerical simulation results confirm the realizing of a good estimation of the angular speed, without using additional equipment in all load conditions: no-load, normal operation conditions and overload for both types of the speed estimators: the first or the second order. The speed estimators presented in this paper can reject the load torque signal noises from the load torque sensor, increasing the robustness of the speed estimated signal. Moreover, compared to existing estimation methods (Mostafa, *et al.*, 2009; Barada, et al, 2002), the proposed one is available with high accuracy over a large speed range (including low rotor frequency), in stationary and dynamic regimes. The rotor field oriented control is the spread method used in drive applications with induction motors. The utility of the estimator is important for the speed control or for the advanced control synthesis (Rosu and Gaiceanu, 1998), with high performances such as optimal vector control of induction motor drives (Rosu, et al., 1998).

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