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견인용 유도전동기의 센서리스 제어

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SENSORLESS CONTROL FOR INDUCTION MOTOR USED IN TRACTION APPLICATION

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Abstract - The paper describes a new and rigorous mathematical model for the rotor field oriented system with induction motor which uses the estimated speed and rotor flux based on a Model Reference Adaptive System, as well as the real-time approach. The estimated speed and rotor flux is used for the speed and flux feedback control. The stability and the convergence of the estimator are improved on the basis of hyperstability theory for non-linear systems. The real-time controller and estimator are implemented with a sampling period of 926 μs using a dual TMS320C44 floating-point digital signal processor. The validity of the proposed method is verified by simulation, and also, the sensorless control was tested on the propulsion system simulator, used for the development of Korean High-Speed Railway Train (KHSRT) [5].

Keywords. Adjustable-speed drives, sensorless control, rotor field oriented induction motor.

1. INTRODUCTION

The adjustable speed ac drives and especially the vector control drives are widely used due to the development of power electronics and modern control technology. The vector control represents a high performance control of the induction motor. In general it needs a speed sensor for obtaining the speed information (tacho-generator, pulse-encoder) and sensing coils (or Hall sensors) for obtaining the flux information. The utilization of speed and flux sensors is expensive, sensitive at noise and reduces the inherent robustness of the induction motor. That is why, the research in the domain of **sensorless-control** represent an active preoccupation of many researchers for eliminate these two kind of transducers.

Speed and rotor flux estimation is performed from the information about motor parameters, voltage and current. For the speed sensorless vector system there are proposed several methods, such as:

- by estimating the slip frequency using the harmonic voltage due to the rotor slot in the motor phase voltage [1],
- by estimating the slip frequency from voltage, current and motor parameters [2],
- by calculating the rotor speed realizing the control system so that the motor main flux axis may be coincided with the calculated one in the control system [3],
- using the third harmonic component of the air gap flux [4].

The proposed method for estimate the speed and rotor flux uses the error between ***voltage model*** and ***current model*** of the rotor field oriented induction machine. This method introduces the least errors in comparison with other methods because the control is an adaptive one.

2. THE MODEL FOR SENSORLESS CONTROL

The control of the induction machine is a vectorial one and that is why it is necessary to use the space phasor model. So, the well-known p.u. state equations of the induction machine in the general reference frame *d-q* rotating with synchronous speed ω_S, are:

$$\begin{aligned}
&\underline{u}_s = r_s \underline{i}_s + \frac{1}{\omega_b}\frac{d\underline{\psi}_s}{dt} + j\nu_s \underline{\psi}_s \\
&-\underline{u}_r = 0 = r_r \underline{i}_r + \frac{1}{\omega_b}\frac{d\underline{\psi}_r}{dt} + j(\nu_s - \nu)\underline{\psi}_r \\
&\underline{\psi}_s = x_s \underline{i}_s + x_m \underline{i}_r \;\; ; \;\; \underline{\psi}_r = x_m \underline{i}_s + x_r \underline{i}_r \\
&t_e = \Im\{\underline{\psi}_s^* \cdot \underline{i}_s\} = \Im\{\underline{\psi}_r \cdot \underline{i}_r^*\} = x_m \Im\{\underline{i}_s \cdot \underline{i}_r^*\} \\
&\frac{d\nu}{dt} = \frac{t_e - t_r}{T_m} \;\; ; \;\; T_m = J\frac{\Omega_b}{M_n}
\end{aligned} \tag{1}$$

Because the measured quantities are the voltage and the current in the stationary frame α–β, the state equations of the induction machine have to be written in this reference frame and then one can obtain the *voltage model* and the *current model*. Having in view that the induction machine is rotor field oriented, in both models must be present as states, the orthogonal components of rotor flux space phasor.
The rotor flux operating equations for the *voltage model* are:

$$\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix}\psi_{r\alpha}\\ \psi_{r\beta}\end{bmatrix} = \frac{x_r}{x_m}\left[\begin{bmatrix}u_{s\alpha}\\ u_{s\beta}\end{bmatrix} - \begin{bmatrix}r_s & 0\\ 0 & r_s\end{bmatrix}\cdot\begin{bmatrix}i_{s\alpha}\\ i_{s\beta}\end{bmatrix} - \begin{bmatrix}\sigma x_s & 0\\ 0 & \sigma x_s\end{bmatrix}\cdot\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix}i_{s\alpha}\\ i_{s\beta}\end{bmatrix}\right] \tag{2}$$

Where: $\sigma = 1 - x_m^2 / x_m x_r$ is the total leakage coefficient of the induction machine,
and the equations for the *current model* are:

$$\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix}\psi_{r\alpha}\\ \psi_{r\beta}\end{bmatrix} = \begin{bmatrix}-\frac{r_r}{x_r} & -\nu\\ \nu & -\frac{r_r}{x_r}\end{bmatrix}\cdot\begin{bmatrix}\psi_{r\alpha}\\ \psi_{r\beta}\end{bmatrix} + \frac{r_r x_m}{x_r}\begin{bmatrix}i_{s\alpha}\\ i_{s\beta}\end{bmatrix} \tag{3}$$

Taking into account that the *voltage model* (2) does not contain the rotor speed ν, this observer may be regarded as a reference model of the induction machine, and the *current model* (3) which contains the rotor speed may regarded as an adjustable one. If it will be constituted an adaptive law so that, the output of the *voltage model* coincides with the output of the *current model*, one can estimate the rotor speed and the orthogonal components of the rotor flux space phasor. Considering the *voltage model* as a reference one, this can be written under the following form:

$$\frac{1}{\omega_S}\frac{d}{dt}\begin{bmatrix}\psi_{r\alpha}^*\\ \psi_{r\beta}^*\end{bmatrix} = \frac{x_r}{x_m}\left[\begin{bmatrix}u_{s\alpha}\\ u_{s\beta}\end{bmatrix} - \begin{bmatrix}r_s & 0\\ 0 & r_s\end{bmatrix}\cdot\begin{bmatrix}i_{s\alpha}\\ i_{s\beta}\end{bmatrix} - \begin{bmatrix}\sigma x_s & 0\\ 0 & \sigma x_s\end{bmatrix}\cdot\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix}i_{s\alpha}\\ i_{s\beta}\end{bmatrix}\right] \tag{4}$$

Where: $u_{s\alpha}, u_{s\beta}, i_{s\alpha}, i_{s\beta}$ are the measured quantities.

If exists error between these two models of the induction motor, it appears due to the rotor speed ν. If the error does not exists, the *current model* should be identical with the *voltage model* (reference model), and it can be written as follows:

$$\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix}\psi_{r\alpha}^*\\ \psi_{r\beta}^*\end{bmatrix} = \begin{bmatrix}-\frac{r_r}{x_r} & -\nu^*\\ \nu^* & -\frac{r_r}{x_r}\end{bmatrix}\cdot\begin{bmatrix}\psi_{r\alpha}^*\\ \psi_{r\beta}^*\end{bmatrix} + \frac{r_r x_m}{x_r}\begin{bmatrix}i_{s\alpha}\\ i_{s\beta}\end{bmatrix} \tag{5}$$

Subtracting the *current model* – the adjustable model (5) from the *voltage model* – the reference model, (4), the state error equations are:

$$\frac{1}{\omega_b}\frac{\mathrm{d}}{\mathrm{d}t}\begin{bmatrix}\psi^*_{r\alpha}-\psi^e_{r\alpha}\\ \psi^*_{r\beta}-\psi^e_{r\beta}\end{bmatrix}=\begin{bmatrix}-\frac{r_r}{x_r} & -\nu^*\\ \nu^* & -\frac{r_r}{x_r}\end{bmatrix}\cdot\begin{bmatrix}\psi^*_{r\alpha}-\psi^e_{r\alpha}\\ \psi^*_{r\beta}-\psi^e_{r\beta}\end{bmatrix}+\begin{bmatrix}-\psi^e_{r\beta}\\ \psi^e_{r\alpha}\end{bmatrix}\cdot\left[\nu^*-\nu^e\right] \qquad (6)$$

Where: $\psi^*_{r\alpha},\psi^*_{r\beta}$ are the outputs of the *voltage model* - reference model; $\psi^e_{r\alpha},\psi^e_{r\beta}$ are the outputs of the *current model* - adjustable model; ν^e is the estimated rotor speed.

The state equations are written under the following compacted form:

$$\frac{1}{\omega_b}\frac{\mathrm{d}}{\mathrm{d}t}x=A\cdot x+B\cdot y \qquad (7)$$

Where: $x=\begin{bmatrix}\psi^*_{r\alpha}-\psi^e_{r\alpha}\\ \psi^*_{r\beta}-\psi^e_{r\beta}\end{bmatrix}$; $A=\begin{bmatrix}-\frac{r_r}{x_r} & -\nu^*\\ \nu^* & -\frac{r_r}{x_r}\end{bmatrix}$; $B=\begin{bmatrix}-\psi^e_{r\beta}\\ \psi^e_{r\alpha}\end{bmatrix}$; $y=\nu^*-\nu^e$.

The defined system (6)-(7) is non-linear because the vector ***B*** depends on the state x ($\psi^e_{r\alpha},\psi^e_{r\beta}$). For determining the rotor speed estimator it should be used the theory of the ***non-linear systems***. Using the well-known notations it is necessary to make references to the base structure of a non-linear system as is presented in Figure 1.

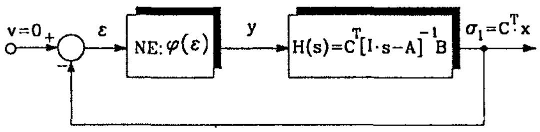


Figure 1. The base structure of a non-linear system.

x - state vector;
v - perturbation vector;
ε - command vector (input vector);
A,B,C - real vectors;
$\varphi(\sigma_1)$=y - real function depending on the real scalar variable $\sigma_1=C^Tx=\varepsilon$;
NE - non-linear element.

If ***B*** is a real variable vector, it will be introduced in the framework of the non-linear element *NE*. Indifferent of the non-linear characteristic $\varphi(\varepsilon)$, the equation $y=\varphi(\sigma_1)$ represents the non-linear operator in the absence of the perturbation (v=0).

$$\frac{\sigma_1}{y}(s)=\frac{-\varepsilon}{y}(s)=C^T\left[I\cdot s-A\right]^{-1}\cdot B=\Delta H(s) \qquad (8)$$

For the synthetization of a non-linear system with a given structure and unknown parameters it is used the ***Popov hyperstability criterion***.

The ***hyperstability*** represents the propriety of an entity composed by a dynamic system and a particular integral with the following form:

$$\mathrm{Int}=\int_0^T\varepsilon^T(t)\cdot y(t)\mathrm{d}t\geq 0 \quad \forall T>0 \qquad (9)$$

Where:.

$\varepsilon(t)$ is the input of the non-linear system;
$y(t)=\varphi(\sigma_1)$ is the output of the non-linear system.

For a non-linear system $y=A_1\cdot\varepsilon$ with continuous functions *y* and *ε* for $t>t_0$, the hyperstability can be defined, if it is associated the integral index:

$$\eta(t_0,T)=\int_{t_0}^T\varepsilon^T(t)\cdot y(t)\mathrm{d}t \qquad (10)$$

Taking into account the above explanations, the non-linear system ***y*** is ***hyperstable*** if, for a pair of functions **ε(t), y(t)**, exists a constant, so that:

$$\eta(t_0,T)\geq-\gamma^2 \quad \forall T\geq t_0 \qquad (11)$$

Using this criterion, it will be determined the structure of the non-linear system for the rotor speed estimation, imposing a hyperstable structure, as is presented in Figure 2.

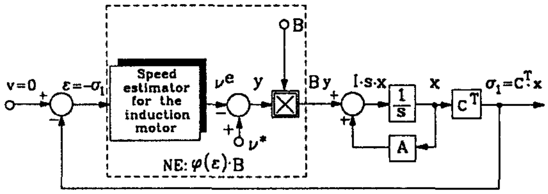


Figure 2. The equivalent non-linear system for rotor speed estimation.

Applying to this system the hyperstability criterion, the integral index $\eta(0,T)$ has the following form:

$$\eta(0,T)=\int_0^T\varepsilon^T(t)\cdot(B\cdot y)(t)\mathrm{d}t\geq-\gamma^2 \quad \forall T\geq 0 \qquad (12)$$

Where:

$$\begin{cases}(B\cdot y)(t)=\begin{bmatrix}-\psi^e_{r\beta}\\ \psi^e_{r\alpha}\end{bmatrix}(\nu^*-\nu^e)\\ \varepsilon^T(t)=-\sigma_1^T(t)=-[C^T\cdot x]^T=-x^T\cdot C=\\ \quad=-[\psi^*_{r\alpha}-\psi^e_{r\alpha} \quad \psi^*_{r\beta}-\psi^e_{r\beta}]\cdot C\end{cases} \qquad (13)$$

The integral index becomes:

$$\eta(0,T)=\int_0^T C\left(\psi^e_{r\alpha}\cdot\psi^*_{r\beta}-\psi^e_{r\beta}\cdot\psi^*_{r\alpha}\right)\cdot\left(\nu^e-\nu^*\right)\mathrm{d}t\geq-\gamma^2 \,\forall T\geq 0 \qquad (14)$$

It may be written as:

$$\eta(0,T)=\int_0^T\varepsilon^{\mathrm{T}}(t)\cdot y'(t)\mathrm{d}t\geq-\gamma^2 \quad \forall T\geq 0 \qquad (15)$$

Where: $\varepsilon'^T=\varepsilon'$

For solving the integral equation (15), one can use the following integral:

$$\int_0^T K\frac{\mathrm{d}f(t)}{\mathrm{d}t}f(t)\mathrm{d}t=\int_0^T\frac{K}{2}\frac{\mathrm{d}[f(t)]^2}{\mathrm{d}t}\mathrm{d}t=\frac{K}{2}f^2(t)\Big|_0^T= \\ =\frac{K}{2}f^2(T)-\frac{K}{2}f^2(0) \qquad \forall K>0\ ;\ T>0 \qquad (16)$$

It results that:

$$\int_0^T K\frac{\mathrm{d}f(t)}{\mathrm{d}t}f(t)\mathrm{d}t\geq-\frac{K}{2}f^2(0) \qquad \forall K>0\,;T>0 \qquad (17)$$

Equalising the integral (16) with the integral (17):

$$\int_0^T C\varepsilon'(\nu^e-\nu^*)\mathrm{d}t=\int_0^T K\frac{\mathrm{d}f(t)}{\mathrm{d}t}f(t)\mathrm{d}t \qquad (18)$$

results the constant ***K*** and the function ***f***(t):

$$f(t)=\nu^e-\nu^*;\ K\frac{\mathrm{d}f(t)}{\mathrm{d}t}=C\varepsilon' \ \Rightarrow\ Kf(t)=C\cdot\int_0^T\varepsilon'\mathrm{d}t \qquad (19)$$

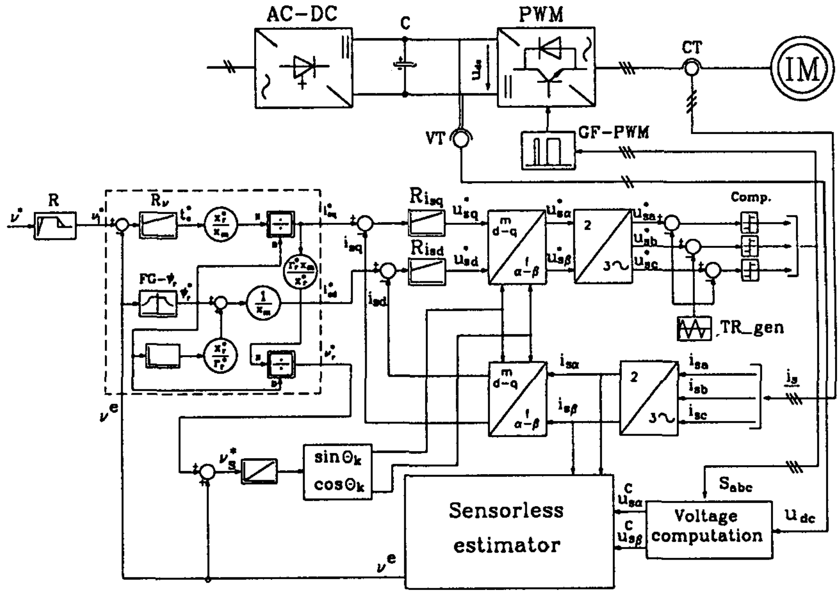


Figure 3. The indirect-rotor field control of the induction machine with sensorless estimator.

For $t=0$ $\nu^e(0)=\nu^*(t)$ (the estimated speed must be equal with the reference model speed) and, taking into account the relations (19), the solution of the integral equation (15) is the following:

$$\nu^e(t)=K_I\int_0^T \varepsilon'(t)dt+\nu^e(0) \qquad \forall K_I>0\ ;\ T>0 \tag{20}$$

This solution represents the adaptive law of the speed and rotor flux estimator. Also, on can be used a *PI* controller instead of an integrator.

The observer estimate the rotor speed so that the error ε' become zero and there are used like inputs, the induction motor voltage and current. For the numerical real-time estimator, the equations of the both models ((2) and (3)) must be numerical handled.

The rotor field oriented control can be made in two manners: the direct-rotor field control, and the indirect-rotor field control. For the traction aplication is used the indirect-rotor field control, as is presented in Figure 3.

3. THE MODEL FOR SENSORLESS CONTROL WITHOUT ROTOR FLUX DERIVATION

Because the rotor flux derivation can lead sometimes to instability of the vector control system, the above model is improved by using the counter-EMF vector, written in p.u. system. The general system of equations (1) can be rearranged in such a manner, to put in evidence the counter-EMF vector, avoiding the rotor flux integration:

$$\begin{cases} \underline{u_s}=r_s\underline{i_s}+\underline{e_m}+\dfrac{1}{\omega_b}\sigma x_s\dfrac{d\underline{i_s}}{dt}+j\,\nu_s\left(x'_m\underline{i'_m}+\sigma x_s\underline{i_s}\right)\\ -\underline{u_r}=0=\dfrac{r_r}{x_r}\left(\underline{i'_m}-\underline{i_s}\right)+\dfrac{1}{\omega_b}\dfrac{d\underline{i'_m}}{dt}+j(\nu_s-\nu)\underline{i'_m}\\ \underline{\psi_r}=x_m\underline{i'_m}\quad ;\quad \underline{i'_m}=\underline{i_s}+\dfrac{x_r}{x_m}\underline{i_r}\\ \underline{e_m}=\dfrac{1}{\omega_b}\dfrac{x_m}{x_r}\dfrac{d\underline{\psi_r}}{dt}=\dfrac{1}{\omega_b}\dfrac{x_m^2}{x_r}\dfrac{d\underline{i'_m}}{dt}=\dfrac{1}{\omega_b}x'_m\dfrac{d\underline{i'_m}}{dt} \end{cases} \tag{21}$$

Where:

$\underline{i'_m}$ is the magnetizing current corresponding to the rotor flux.

Expressing the system of equations (21) in the stator reference frame α–β, one can obtain the modified *voltage model - reference model* and the modified *current model - adjustable model*.

- the *voltage model* - the reference model:

$$\begin{bmatrix} e_{m\alpha}\\ e_{m\beta}\end{bmatrix}=\begin{bmatrix} u_{s\alpha}\\ u_{s\beta}\end{bmatrix}-\begin{bmatrix} r_s & 0\\ 0 & r_s\end{bmatrix}\cdot\begin{bmatrix} i_{s\alpha}\\ i_{s\beta}\end{bmatrix}-\begin{bmatrix} \sigma x_s & 0\\ 0 & \sigma x_s\end{bmatrix}\cdot\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix} i_{s\alpha}\\ i_{s\beta}\end{bmatrix} \tag{22}$$

- the *current model* - the adjustable model:

$$\begin{bmatrix} e_{m\alpha}\\ e_{m\beta}\end{bmatrix}=\frac{1}{\omega_b}x'_m\frac{d}{dt}\begin{bmatrix} i'_{m\alpha}\\ i'_{m\beta}\end{bmatrix} \tag{23}$$

Where:

$$\frac{1}{\omega_b}\frac{d}{dt}\begin{bmatrix} i'_{m\alpha}\\ i'_{m\beta}\end{bmatrix}=\begin{bmatrix} -\dfrac{r_r}{x_r} & -\nu\\ \nu & -\dfrac{r_r}{x_r}\end{bmatrix}\cdot\begin{bmatrix} i'_{m\alpha}\\ i'_{m\beta}\end{bmatrix}+\frac{r_r}{x_r}\begin{bmatrix} i_{s\alpha}\\ i_{s\beta}\end{bmatrix} \tag{24}$$

In this case, according with the hyperstability criterion, the error that has to be minimised (to be maintained near zero) is the following:

$$\varepsilon'=e^e_{m\alpha}\cdot e^*_{m\beta}-e^e_{m\beta}\cdot e^*_{m\alpha} \tag{25}$$

Because the reference model does not require integration, this system can achieve good performance at high speed as well as at low speed, even the stator resistance varies with the temperature or the value of rotor circuit time constant is quite wrong.

The schematic representation of the MRAS (*Model Reference Adaptive System*) is presented in Figure 4, and it is based on the equations of the voltage and current models of the induction machine.

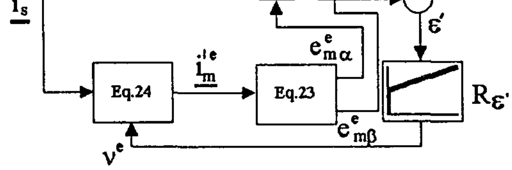


Figure 4. The schematic representation of the MRAS.

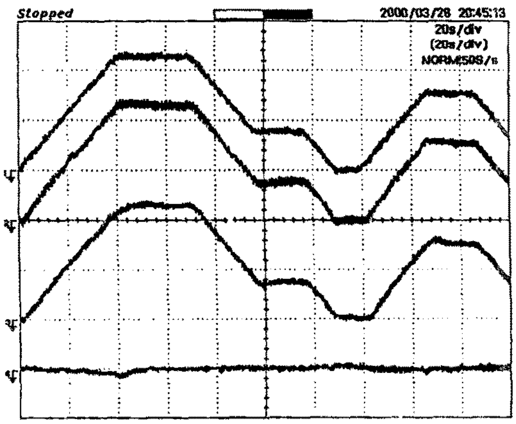
4. CONCLUSIONS & EXPERIMENTAL RESULTS

The MRAS based on the counter-EMF vector which consists of the observer equation – reference model (22) and the adaptive law for speed estimating – adjustable model (23), (24) must be implemented in real-time. If the sampling period is less then 200[μs], witch is considered the normal case, the equations of the observer can be solved using the Euler method, reducing the computation time. It is well-known that in the case of using Euler method, the truncation errors increase with the increasing of sampling period. That is why, especially for the adjustable model which is more sensitive of truncation errors then the reference one, it is needed to use an improved numerical method to reduce the truncation errors, even the sampling period is increased, as in the railway traction application. So, to avoid the numerical instability, for the adjustable model it was used the Runge-Kutta-Gill method. With this method it is solved only the equation (24), the equations (22) and (23) being solved by Euler method.

The rotor speed and orthogonal counter-EMF components are estimated using an adaptive control, and it is introduced as a controller into a rotor field oriented system with induction machine. This estimator can also be used in other direct or indirect vector control schemes as well as in scalar schemes.

In Figure 5. is presented the system behaviour of the indirect rotor filed control using MRAS. The figure shows the plots of ramp-reference speed v_l^*, the estimated motor speed v^e, the real motor speed v, the difference v^e-v.

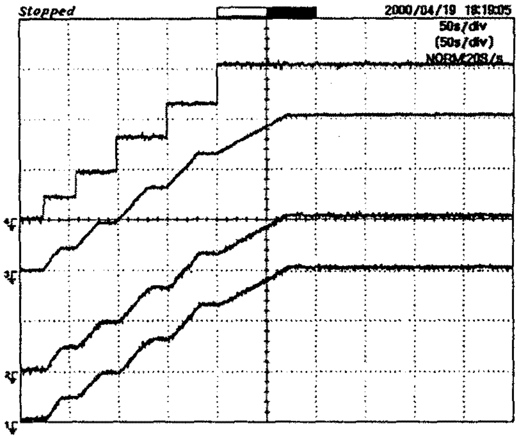
The stability and the least errors are maintained all over the range of operating concerning the rotor speed and electromagnetic torque.
This estimator has the main advantage, the higher convergence and precision. This approach is completely robust to the stator and rotor resistance thermal variations.



Experimental - Indirect Rotor-field control.
Figure 5. The system behaviour: 1 - ramp-reference speed v_l^* ;
2 - estimated speed v^e ; 3 - real speed v ;
4 - error between estimated speed and real speed v^e-v ;
4 motors at full-load; (*652rpm/div., 20s/div.*)

In this figure is presented also the steady state operation at very low and at high speed, to prove the performances in speed estimation of the proposed MRAS. It is important to notice, that it can operate also at low speed (for instance – 50 rpm; it means 2.9% from the rated speed). In the same time one can be seen the operation at zero frequency of the converter (with the powered converter), and the estimation of zero speed is also quite good.

The MRAS estimator, which uses the counter-EMF vector, can operate in good conditions at very low speed because it doesn't require the integration in the reference model. Because of this, the estimator can operate without errors and instabilities also at very high speed (the maximum motor speed of KHSRT - 4200[rpm]) as is presented in Figure 6. This figure shows from bottom to top the plots of real motor speed v, the estimated motor speed v^e, ramp-reference speed v_l^* and the step-reference speed v^*.



Experimental - Indirect Rotor-field control.
Figure 6. The system traction operation from zero to the maximum train speed:
1 - measured speed v ; 2 - estimated speed v^e ; 3 - ramp-reference speed v_l^* ;
4 - step-reference speed v^* ;
4 motors at full-load; (*1355rpm/div., 50s/div.* ; n_{max}=*4200rpm.*)

One can be seen a very good dynamic of the sensorless system and also a very good agreement between the measured speed and the estimated one all over the range of speed and load torque.

Taking into account the above conclusions, the proposed MRAS speed estimator is suitable for traction application. Using this MRAS there are avoided the speed pulse-encoders and also the phase voltage transducers. There are used only line current transducers and one DC voltage transducer, that represents the minimum number of transducers used in closed loop speed control scheme.

This approach was presented in a way that will contribute to a better understanding of the MRAS to motion control, especially for traction applications.

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