COMPARISON ON SENSORLESS CONTROL OF SYNCHRONOUS MOTORS

Radu GIUCLEA, **Cristian BOGUS**, **Liviu KREINDLER**

University "Politehnica" Bucharest – Electrotechnics Faculty
Splaiul Independenței 313, Bucharest – România

Abstract: The paper compares two different methods for speed and position estimation in AC permanent magnet synchronous motors vector control applications. The first method implies two observer blocks — one for the speed, and the other for the electrical position, using the voltage equations in the (d,q) reference frames. The second method estimates the same variables starting from the calculation of instantaneous reactive power. The tests have proved excellent behaviour in steady state (method 1) as well as in transient state (method 2). The implementation has been made on the 16 bits fixed-point DSP - TMS320F240 from Texas Instruments.

Keywords: sensorless control, DSP controller

1. INTRODUCTION

AC motors vector control schemes require the knowledge of the rotor flux position at each sampling. Based on this position and on the torque demand, the control system can compute the required current distribution in the stator phases in order to obtain maximum efficiency from the motor.

Various sensorless solutions have been proposed as the number of applications requiring such approach constantly increases. In the followings, the authors present and compare two sensorless methods that were implemented for permanent magnet synchronous motors (PMSM).

The work was validated by comprehensive tests made in industrial applications.

2. SPEED AND POSITION OBSERVED DIRECTLY FROM THE VOLTAGE EQUATIONS

This method is based on two observer blocks: one for the speed, and the other for the position of the rotor flux.

The principle used in both cases is that of a reference model applied to the voltage equations in the (d,q) reference frame (Equation 1).

$$\begin{aligned} v_q &= Ri_q + L\frac{di_q}{dt} + Li_d\omega + \varphi_0\omega \\ v_d &= Ri_d + L\frac{di_d}{dt} - \omega L i_q \end{aligned} \tag{1}$$

where R and L are equivalent-phase resistance and inductance; i_d and i_q are the motor currents in the (d,q) frame; ω is the motor speed; φ_0 is the flux of the permanent magnets; and v_q is the voltage on the *q*-axis.

The speed observer

The *q*-axis voltage equation is used as a reference model for the speed observer block. The ideal $\hat{v}_q$ voltage computed from the equation is compared to the applied voltage, v_q, calculated from the DC bus voltage and the inverter dead time values.

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The estimated speed $\hat{\omega}$ is obtained by integrating the voltage error considering a speed estimation gain k_{spd}

$$(2)\ \hat{\omega} = k_{spd} \int \left(v_q - Ri_q - L\frac{di_q}{dt} - \varphi_0 \cdot \hat{\omega} \right) dt$$

which translated to the discrete domain with a sampling time Ts gives the equations (2) presented under the mode in which they were implemented on the DSP:

$$(3)\ \begin{aligned} I_i &= I_{i-1} + k_{spd}\left(v_q - R \cdot i_q - \varphi_0 \cdot \hat{\omega}\right)T_s \\ \hat{\omega} &= I_i - k_{spd} \cdot L \cdot i_q \end{aligned}$$

The position observer

The *d*-axis voltage equation is used to minimise the angle error between a hypothetical (δ, γ) reference frame and the actual (d, q) reference frame.

$$(4)\ v_\delta = Ri_\delta + L\frac{di_\delta}{dt} - \hat{\omega} \cdot L \cdot i_\gamma$$

$$(5)\ e_\delta \overset{\Delta}{=} v_d - v_\delta = \omega \cdot \varphi_0 \cdot \sin\delta$$

Integrating the *d*-axis voltage difference minimises the error angle δ. With $J = \frac{1}{2}e_\delta^2$ as minimising criteria, k_{pos} the gain of the position observer and $\frac{d\delta}{dt} = -k_{pos} \cdot \frac{dJ}{d\delta}$, one obtains

$$(6)\ \delta = -k_{pos} \int \left(v_d - Ri_d - L\frac{di_d}{dt} + \hat{\omega} \cdot L \cdot i_q \right) \cdot \omega \cdot \varphi_0 \cdot \cos\delta \cdot dt$$

Considering small variations near the operation point, when $\delta \longrightarrow 0$ one has $\cos\delta \approx 1$, $i_\delta \longrightarrow i_d = 0$ and ω constant during a current sampling period. Then, the rotor flux position estimated angle can be computed as

$$(7)\ \hat{\theta} = \int \hat{\omega} \cdot dt + \delta$$

Thus, the implementation on the DSP follows the equation (8):

$$(8)\ \hat{\theta}_i = \hat{\theta}_{i-1} + \hat{\omega}\left[1 - k_{pos}\varphi_0 \cdot \left(u_d + \hat{\omega} \cdot L \cdot i_q\right)\right]T_s$$

Finally, a correction is applied in steady state regime to compensate for the uncertainty and variation of the motor parameters. The error between the estimated position angle increment ($theta\_inc$) and a target value ($theta\_inc*$) computed from the speed reference is used to adjust the motor model parameters on-line and correct the speed estimation. The magnet flux coefficient (noted $k\varphi_0$) was chosen to be the on-line adjustable model parameter:

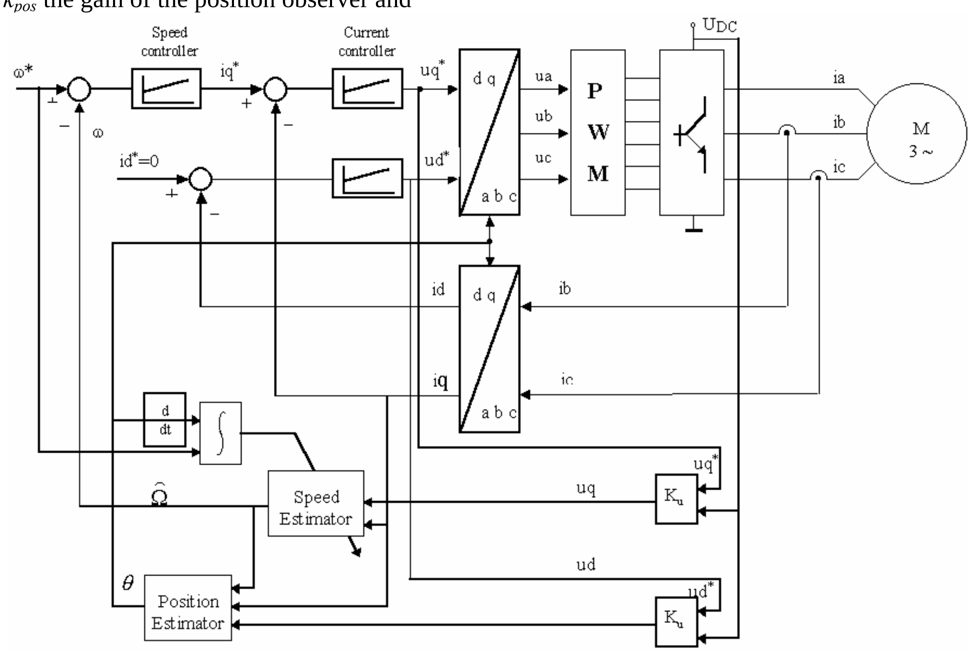

Fig. 1. Sensorless control implementation of method 1 – block diagram

$$k\varphi_0 = k\varphi_0 + gain\_k\varphi_0(theta\_inc * - theta\_inc) \tag{9}$$

where $gain\_k\varphi_0$ represents the integrator gain that cancels the error between desired synchronous speed and estimated speed for constant reference.

Figure 1 presents the block diagram of the DSP sensorless control implementation scheme.

Tests of this method have been made on different types of permanent magnet brushless motors in different applications. Figure 2 presents the start-up ramp for a 60 W brushless motor. An incremental encoder was used to compare the estimated speed and position with the actual ones (*omg_ref* - speed reference, *omg1* - actual speed, *omg* - estimated speed, *theta1* - actual electrical position, *theta* - estimated electrical position).

Figure 3 presents the statistical data gathered on a wide speed range for steady state application with a 1 kW motor.

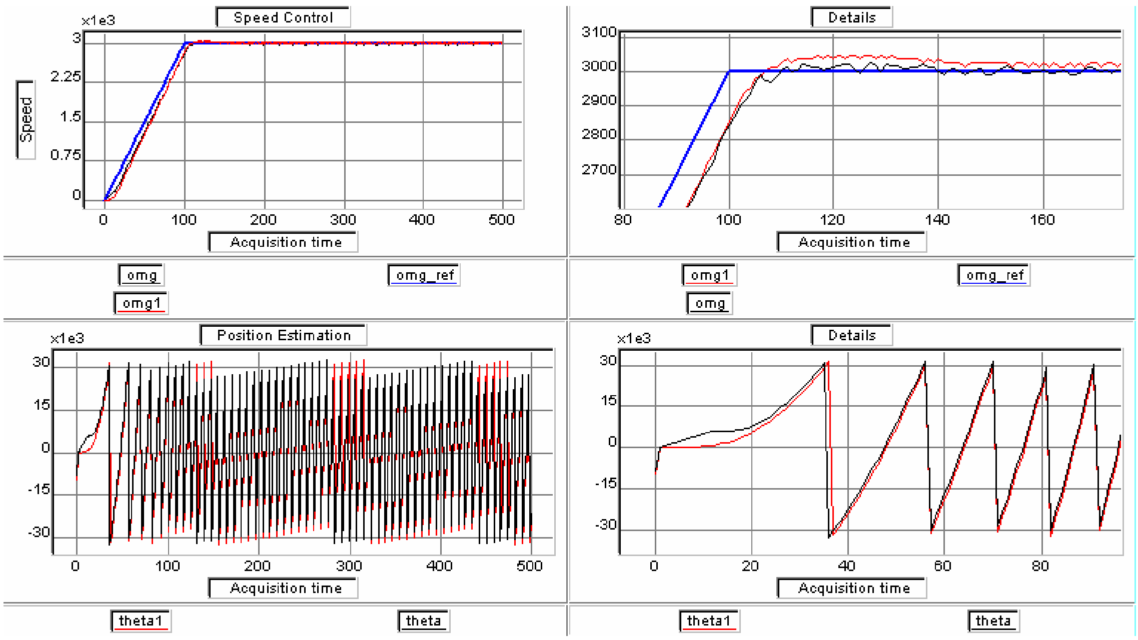

Fig. 2. Sensorless control of 60 W brushless motor - experimental results, method 1

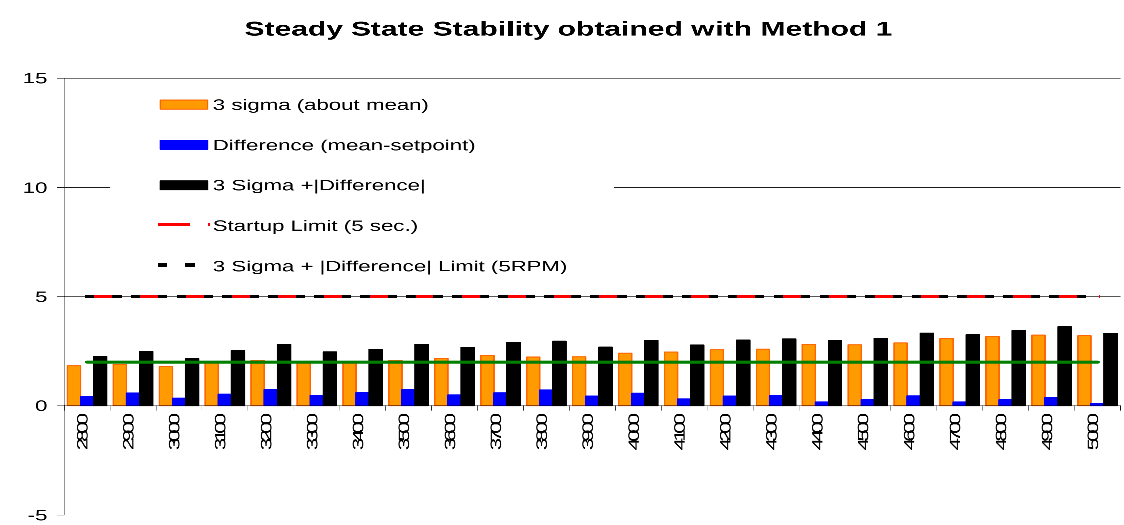

Fig. 3. Method 1 experimental results on 1 kW motor –steady state stability in a wide speed range

3. SPEED ESTIMATION FROM THE INSTANTANEOUS REACTIVE POWER

The principle used in this method is that of a reference model applied to the instantaneous reactive power computed from the voltage equation in the (d, q) reference frame (1).

The instantaneous reactive power, q, is by definition:

$$(10)\; q = u_q i_d - u_d i_d$$

and also, starting from (1), can be expressed as:

$$(11)\; q = \omega L_d i_d^2 + \omega L_q i_q^2 + \omega \varphi i_d + L_q i_d \frac{di_q}{dt} - L_d i_q \frac{di_d}{dt}$$

The reference model

Equation (10) is considered the reference model for the reactive power:

$$(12)\; q_{ref} = u_q i_d - u_d i_d$$

In order to eliminate the perturbations caused by the variations of the DC bus voltage level, a DC bus level compensation is included in q_{ref} computation.

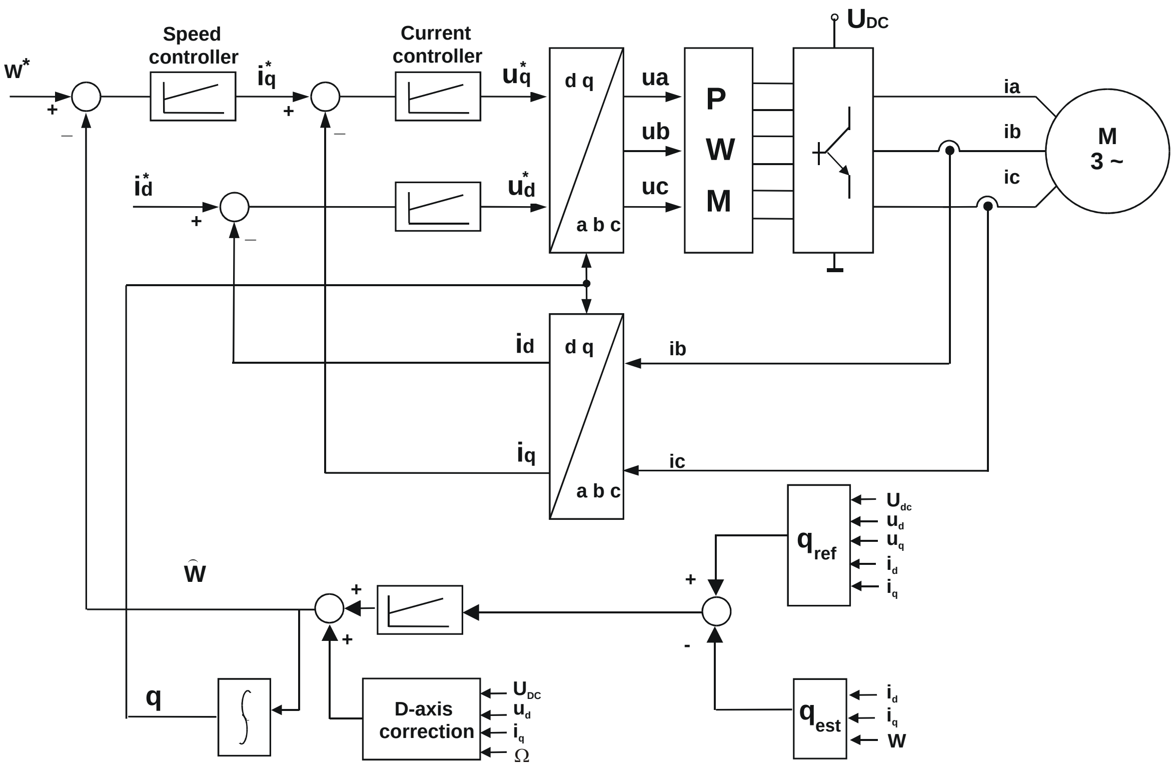

Fig. 4. Sensorless control implementation of method 2 – block diagram

The adaptive model

The adaptive model construction is based on the equation (11). Assuming:

$$\frac{di_q}{dt} = 0 \; ; \; \frac{di_d}{dt} = 0 \text{ and } L_d = L_q = L_s$$

equation (11) can be further written as:

$$(13)\; q_{est} = \omega L_s \left(i_d^2 + i_q^2\right)$$

The term $\omega \varphi i_d$ has been intentionally left out.

A proportional integral controller is used to adjust ω so that $q_{est} = q_{ref}$.

If $q_{est} = q_{ref}$ then $\omega \varphi i_d = 0 \Rightarrow i_d = 0$ so the d axis of the control is aligned with the d axis of the rotor flux.

In order to enhance the functionality of the control in dynamic regimes such as the braking mode, a correction is added to the synchronous speed computed by the PI controller. The correction is based on the d axis voltage model. Starting from the d axis voltage equation:

$$u_d = R i_d + L_s \frac{di_d}{dt} - \omega L_s i_q$$

and assuming $i_d = 0$,

the ideal d axis voltage can be computed as:

$$(14)\ \hat{u}_d = -\omega L_s i_q$$

The ideal d axis voltage is compared to the actual applied voltage, computed from the d axis current controller output and the DC bus voltage. A second proportional controller minimizes the error between the two voltages (ideal and actual). The controller output is added as a correction to the estimated (electrical) speed.

In order to obtain the position of the rotor, the synchronous speed is properly scaled and integrated.

The block diagram of the DSP sensorless control implementation scheme is presented in Figure 4.

Sensorless control experimental results obtained with this method on a 200W brushless motor are presented in the last two figures. Figure 5 captured the start-up and steady state operation in case of constant torque reference. Figure 6 presents closed loop speed control performances. An incremental encoder was also used to present the accuracy of the speed estimation.

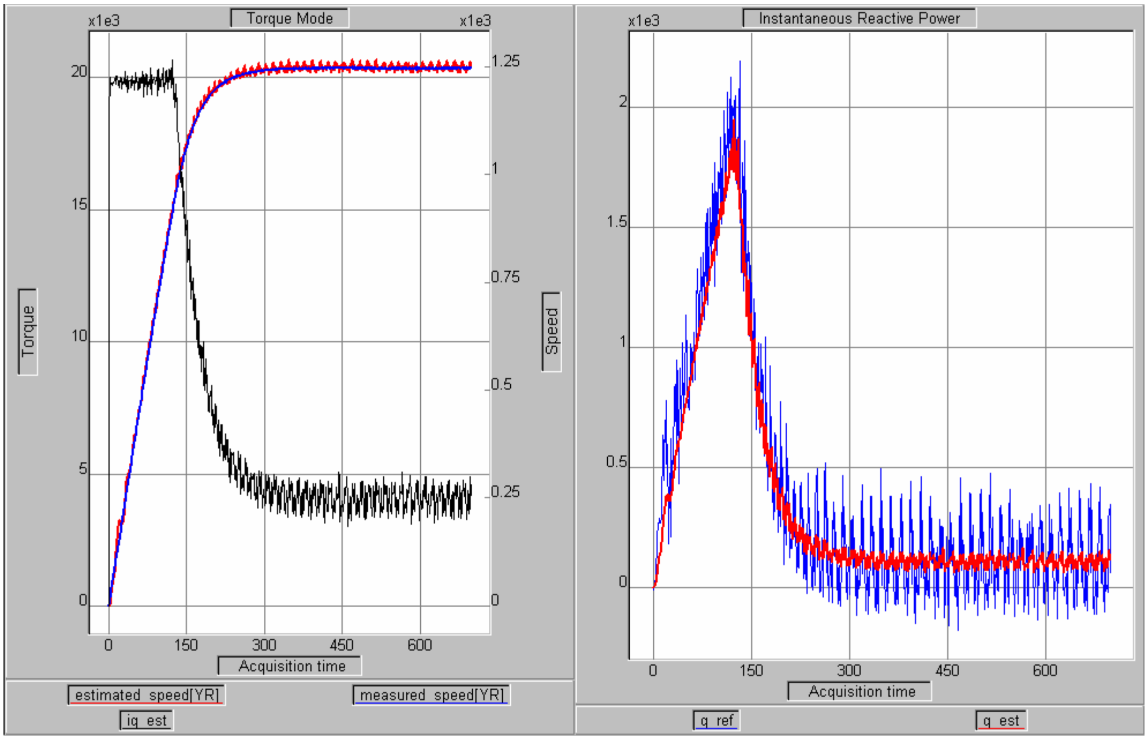

Fig. 5. Sensorless torque control of 200W brushless motor - experimental results, method 2

4. CONCLUSIONS

Two sensorless methods for estimating speed and position in PMSM vector control applications have been presented and compared.

Method 1 estimates the speed and the electrical position directly from the voltage equations in the (d,q) reference frames. Method 2 estimates the same variables by applying a reference and adaptive model to the instantaneous reactive power. Extensive tests have been made with both methods validating them in industrial applications. Method 1 demands an exhausting tuning procedure due to the great number of parameters. Having fewer parameters, method 2 can be much easily adapted to a new motor.

The tests have proved excellent behaviour in steady state regime for the first method. When fine-tuning was accomplished, this method proved very good stability on a wide speed range. Even more, the speed control error was found less than 5 rpm in a whole range from 2800 to 5000 rpm that is remarkable for a sensorless solution (see Figure 3).

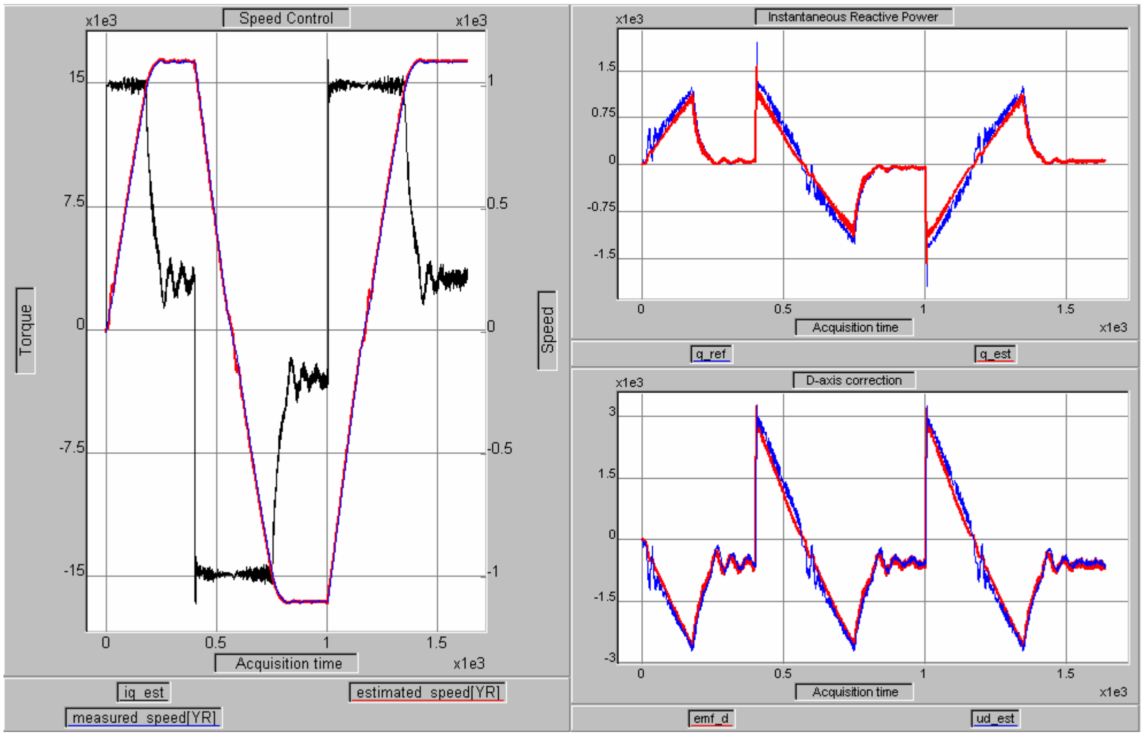

Fig. 6. Sensorless speed control of 200W brushless motor - experimental results, method 2

Tests recommend the second method for its robustness in dynamic regimes. Specific application requirements characterised by random power interruptions were satisfied by this implementation. Method 2 proved to have the possibility to gain control of a running motor without having to wait for a complete stop before restarting the motion.

A DSP solution was used for implementation - the 16 bit fixed-point DSP TMS320F240. A high bandwidth was attained with the H/W and S/W solution adopted: 1 ms speed loop, 100 microsec current loop, 20 kHz symmetric PWM modulation frequency.

5. REFERENCES

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