

Back-EMF Sensorless Control Algorithm for High Dynamics Performances PMSM

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Abstract—In the paper a low time consuming and cost sensorless control algorithm for high dynamics performances Permanent Magnet Synchronous Motors (PMSM) both surface or internal permanent magnet mounted for position and speed estimation is introduced discussed and experimentally validated. This control algorithm is based on the estimation of rotor speed and angular position starting from the back electromotive force space vector determination without voltage sensors by using the reference voltages given by the current controllers instead of the actual ones. This choice obviously introduces some errors that must be vanished by means of a compensating function.

The novelties of the proposed estimation algorithm are the position estimation equation and the process of compensation of the inverter phase lag that also suggests the final mathematical form of the estimation.

The mathematical structure of the estimation guarantees a high degree of robustness against parameters variation as shown by the sensitivity analysis reported in the paper.

Experimental verifications of the proposed sensorless control system have been made with the aid of a flexible test bench for Brushless Motor Electrical Drives. The test results presented in the paper show the validity of the proposed low cost sensorless control algorithm and, above all, underline the high dynamic performances of the sensorless control system also with a reduced equipment.

Index Terms— Brushless machines, Synchronous motor drives, Transducers, Control equipments

I. INTRODUCTION

IN FIELD ORIENTED control for Brushless Machines the exact knowledge of rotor angular position is needed. The rotor angular position and speed of an electrical machine can be obtained in real time with specific sensors (encoders, resolvers), but their presence may introduce several disadvantages. As a matter of fact, the choice of angular position and speed sensors and the consequent choice of the related signal conditioning circuit is not much influenced by electrical drive rated power. Then the influence of the above mentioned sensor's cost grows up in percentage with respect to the electrical drive comprehensive cost, when the rated power of electrical machine is small or fractional. Finally the signal transmission between sensor and control system can

be submitted to Electro-Magnetic Interference (EMI) coming from external sources, producing an error in measurement that may be significant for feedback control. Sensorless drives are a viable solution for all the mentioned situations in which the presence of a speed/position sensor is an element of weakness in the drive.

Various sensorless control techniques have been developed for Brushless DC and Brushless AC machines and many are still in development. All the solutions proposed during the last years have advantages and drawbacks and cannot be considered resolute for the widest range of applications and for the different types of motors. References [1]-[26] give an idea of the manifolds of the proposed solutions and of the development during long time of the state of the art issued to sensorless control of Brushless drives.

Fundamentally two big families of sensorless control for IPMSM are issued. The first develop the difference of saliency between direct and quadrature axis and the rotor position information is derived from the signal conditioning of the currents output due to the injection of high frequency voltage on the stator winding (see e.g. [5],[6],[7]). In this way the machine is used itself as a resolver device. Signal injection on the stator is effective and allows the control of the IPMSM also at standstill, but it is quite complicated and require great attention in the design of the control system and of the signal conditioning system to extract rotor position. The latter is based on back Electro Motive Force (EMF) calculated by integration of the total flux linkage on the stator phase circuits (see e.g. [1],[8],[11],[16]). This system is simpler but cannot assure control at standstill or at very low speed and suffer also for flux integrator's drift problem, especially in the analog realizations. Furthermore it is very sensitive to the variations of the stator resistance during operation. Actually due to fast enhancement in microcomputer technology the application of classical state observers of modern control theory such as Luemberger's, optimal state observers, extended Kalman filter and so on, are gaining a renewed attention (e.g. [11],[25],[23]).

In this paper, a novel low time consuming and low cost sensorless control algorithm for permanent magnet synchronous motor (PMSM) drives, both surface or internal permanent magnet mounted, based on the determination of the back electromotive force (back-EMF) without the aid of voltage probes, is presented, discussed and experimentally verified. Estimation of the back EMF is made via the reference voltages given by the current controller. Sensorless control with only current sensors is not new (see e.g. reference [19], [20], [21], and [22]), in particular in the reference [21] the reference voltages given by the current controllers are used for estimation with

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identification of the machine parameter. This identification concurs to make the estimation system quite still complex and computationally expensive. In a previous paper (see [22]) an early version of the proposed control algorithm have been discussed and experimentally tested. In the same paper the main proposed innovation was essentially the way in which the error introduced by reference voltage usage is corrected as well. This error is due to the time lag introduced by the converter operation. The correction was there determined in real time, by implementing a compensating function.

In this paper an enhanced version of the estimation is proposed. The enhancements consists basically in the introduction of a control loop in the estimation that causes the vanishing of the error in position estimation due to currents transients effects. The presence of the control loop allows to increase the dynamic performances of the electrical drive and the precision of the compensating functions introduced in [22].

It is clear that this correction method and estimation process is more suitable for high dynamic industrial electrical drives working near the rated speed and not for motion control systems for the lack of the behavior at low speed, in particular for the oscillation that the estimation exhibit at low speed.

As a matter of fact, the absence of voltage probes reduces the cost of the system and improves its reliability and electromagnetic susceptibility. The rotor position used for the field oriented control with null direct component of the stator current (i_d) is determined on the basis of the mathematical model of the motor. The motor model allows to calculate the argument of the complex vector of the stator back EMF. From this one, the real instantaneous rotor position is hereafter determined introducing some offset correction terms depending on the actual stator currents and on rotor speed.

The paper is summarized as follows: In section II after a brief recall of the PMSM mathematical model, rotor position and speed determination are discussed. Estimation method is then the object of a further discussion in which the Authors explain what are the electrical quantities and the machine parameters on which position and speed offset terms depend and in which way these must be taken into account.

Experimental results, carried out with the aid of a PMSM drive flexible test bench, giving the validation of the proposed technique and showing the high performances reached, are presented in Section III.

In section IV the degree of robustness of the estimation method against stator resistance and quadrature inductance variations is analytically investigated.

In section V conclusions are finally summarized.

II. CONTROL ALGORITHM DESCRIPTION.

A. PMSM mathematical model

The proposed control system is based on the motor model expressed in the dqo reference frame [5], [6], [7].

The equations of the motor model are:

$$\begin{aligned} v_d &= R i_d + L_d \frac{di_d}{dt} - L_q P_p \omega i_q \\ v_q &= R i_q + L_q \frac{di_q}{dt} + L_d P_p \omega i_d + P_p \omega \lambda_{PM} \\ T_m &= \frac{3}{2} P_p [\lambda_{PM} i_q + (L_d - L_q) i_d i_q] \\ \frac{d\omega}{dt} &= \frac{1}{J} (T_m - T_L - F\omega) \\ \frac{d\theta}{dt} &= \omega \end{aligned} \quad (1)$$

in which:

$v_d v_q$ are the stator voltages in the rotor reference frame;

$i_d i_q$ are the stator currents in the rotor reference frame;

R is the stator phase resistance;

$L_d L_q$ are the machines inductances, respectively along direct quadrature axis;

P_p is the number of motor pole pairs;

λ_{PM} is the flux generated by permanent magnets;

T_m is the electromagnetic torque;

T_L is the load torque;

F is the viscous friction coefficient

J is the moment of inertia of all rotating masses;

ω is the instantaneous angular speed;

θ is the instantaneous angular position.

In industrial field oriented control for the sake of simplicity and for reduction of costs the condition $i_d = 0$ is used. In particular when the saliency ratio is not so high (<3) the contribution of the reluctance torque is hard to justify if compared to the major complexity of the control algorithm implemented to follow the condition of a maximum Torque - current ratio.

The $i_d = 0$ condition is imposed also in the control algorithm presented in this paper because of the industrial target of the possible applications.

In this case the torque expression in eq. (1) set becomes:

$$T_m = \frac{3}{2} P_p \lambda_{PM} i_q \quad (2)$$

Because of the constant permanent magnet flux, the torque depends only on the quadrature component of the stator current.

B. Description of the estimator.

As already established in the introduction, a characteristic of the proposed sensorless control technique is the determination of the back e.m.f space vector without voltage measurements by using the reference voltages instead of the actual ones, so that the system is a low cost and reliable one. In the control system, whose block scheme is shown in Fig. 1, motor supply voltage values that constitute some inputs of the estimator, are replaced with reference voltage signals (v_a^* , v_b^* , and v_c^*) generated by the current regulator. Considering the inverter reference voltage signals instead of the real supply voltages implies, obviously, some approximations in the back e.m.f determination. In fact, by not considering the inverter operations follows to neglect the voltage harmonics and the

Now consider that, neglecting all harmonics, v_α and v_β are respectively cosine and sine functions of ωt . The same can be said for v_α^* and v_β^* . In particular it is:

$$\begin{aligned} v_\alpha^* &= V \cos(\omega(t+T)) \\ v_\beta^* &= V \sin(\omega(t+T)) \\ v_\alpha &= V \cos(\omega t) \\ v_\beta &= V \sin(\omega t) \end{aligned} \quad (9)$$

where V is the RMS of the stator voltage. It is now clear that, being T very small compared to $2\pi/\omega$:

$$\begin{aligned} v_\alpha^* &= v_\alpha + \delta v_\alpha = v_\alpha - v_\beta \omega T \\ v_\beta^* &= v_\beta + \delta v_\beta = v_\beta + v_\alpha \omega T \end{aligned} \quad (10)$$

In this way our previous expansion of (8) reduces to:

$$\frac{-(i_\beta v_\beta + i_\alpha v_\alpha)R + v_\beta^2 + v_\alpha^2}{R^2(i_\alpha^2 + i_\beta^2) - 2R(i_\alpha v_\alpha + i_\beta v_\beta) + v_\alpha^2 + v_\beta^2} \omega T \quad (11)$$

that in a more compact form becomes:

$$\frac{V^2 - VRI \cos(\varphi)}{V^2 + R^2 I^2 - 2VRI \cos(\varphi)} \omega T \quad (12)$$

being $\cos(\varphi)$ the power factor in motoring operation and V , I and E , respectively, the RMS values of the stator voltages currents and back EMF.

Equation (12) may be written also in complex number form:

$$\frac{\vec{V} \cdot (\vec{V} - R\vec{I})^*}{|\vec{E}|^2} \omega T \quad (13)$$

where (*) denotes the complex conjugate:

eq (12) express the correction term to be used to estimate the argument of the back EMF when the reference voltages are used. Furthermore, as with $i_d = 0$ the motor drive operates with near unity power factor, eq. (12) can be further simplified as follows:

$$\frac{V^2 - VRI \cos(\varphi)}{V^2 + R^2 I^2 - 2VRI \cos(\varphi)} \omega T \simeq \frac{VE}{E^2} \omega T = \frac{V}{E} \omega T \quad (14)$$

However the complex products on space vector eq. (13) are not difficult to implement.

It is clear that E depends on the motor speed but also on the current. So the term $(V/E)\omega T$ in (14) depends simultaneously on speed and on stator current (by means of E). This term is referred as the ‘‘speed offset’’ because ω appears explicitly, even if E depends on stator current.

Introducing these speed and current offset corrections in the (4), one finally get a suggestion for a first expression of ‘‘estimated’’ position:

$$\theta_{est} = \arctan\left(\frac{v_\beta^* - Ri_\beta}{v_\alpha^* - Ri_\alpha}\right) - \frac{V}{E} \omega T - \arctan\left(\frac{\lambda_{PM}}{L_q i_q}\right) \quad (15)$$

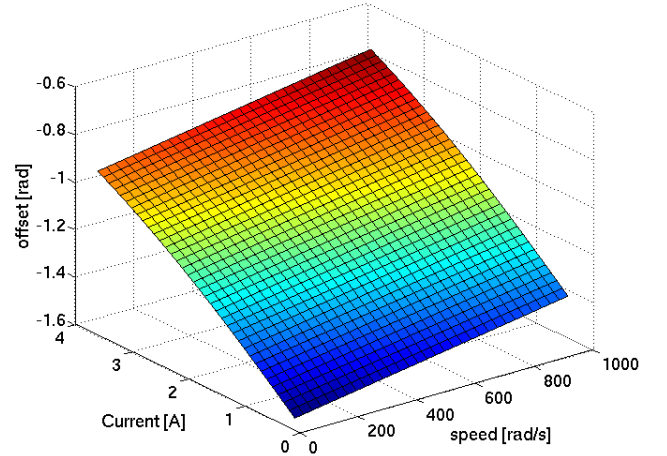


Fig. 2. Position offset at steady state as a function of speed and current.

Fig. 2 shows the sum of ‘‘speed’’ and ‘‘current’’ offsets $\frac{V}{E} \omega T + \arctan\left(\frac{\lambda_{PM}}{L_q i_q}\right)$ as a function of speed and current i_q at steady state with reference to the IPMSM used for experiments in this paper (see Table I above for nameplate ratings).

Speed and current offset may be easily calculated off line, or better determined via measurements from which the offset correction terms may be determined via polynomial interpolation with data fitting procedures. Interpolated data can be stored in discrete form in two-dimensional look up table (a single ROM chip) from which they are recalled depending on the particular motor operation state. This reduces the mathematical burden of estimation and involve to bring an increased step of economy and simplicity in the electrical drive.

In the next sections, the on line realization of the estimator is considered.

C. Estimator realization.

Eq (15) may be rewritten in differential form by substituting ω with the derivative of θ_{est} :

$$\theta_{est} = \arctan\left(\frac{v_\beta^* - Ri_\beta}{v_\alpha^* - Ri_\alpha}\right) - T \frac{V}{E} \frac{d\theta_{est}}{dt} - \arctan\left(\frac{\lambda_{PM}}{L_q i_q}\right) \quad (16)$$

It is important for the sake of a correct estimation to avoid the occurrence of inverter overmodulation by oversizing the DC Link voltage.

Equation 16 describes the estimation dynamic that converge to the value of the real rotor position with $i_d = 0$ field oriented control, but in practical realization it suffers for steady state error due to variation of i_d during transients. In fact, even if the PI current regulators can theoretically vanish the cross coupling between the current channels via feed-forwarding, the presence of the delay introduced by the inverter make this decoupling not perfect so that i_d exhibit some significant variations during speed change or load torque insertion transients.

The steady state error on θ_{est} may be vanished with a PI controller creating an estimation control loop.

To establish this control loop we consider that the estimated direct current $\hat{i}_d = i_\alpha \cos(\theta_{est}) + i_\beta \sin(\theta_{est})$ is forced to zero at steady state only when $\theta_{est} = \theta$. The PI regulator so introduces a further correction term on θ_{est} vanishing the steady state position error. Taking the presence of the PI into account, the ultimate estimator equation form is:

$$\theta_{est} = \arctan\left(\frac{v_\beta^* - Ri_\beta}{v_\alpha^* - Ri_\alpha}\right) - T \frac{V}{E} \frac{d\theta_{est}}{dt} + \arctan\left(\frac{\lambda_{PM}}{L_q i_q}\right) - k_p \hat{i}_d - k_i \int \hat{i}_d dt \quad (17)$$

where k_p and k_i are the proportional and integral PI parameters.

Before the PI tuning it is mandatory to add a pole in the origin of the complex plane with the aim to obtain a system of the second type. This ensures that the position estimation error $\theta - \theta_{est}$ vanishes. In order to ensure stability to the control loop the Routh Criterion is applied to the determinant of the approximated closed loop transfer function. The Routh polynomial is a function of the k_p and k_i parameters. In this way it is possible to establish a relational bound between these parameters whose numerical determination is after made with the traditional methods of the control system theory.

The estimator block diagram is shown in Fig. 3.

III. EXPERIMENTAL RESULTS

A. Description of the test bench.

In order to validate the effectiveness of the proposed sensorless control system a flexible test bench has been built and set up.

The electrical drive test bench is constituted by [17], [18]:

- an internal permanent magnet synchronous motor;
- a controlled hysteresis brake;
- a dSPACE (digital Signal Processing And Control Engineering) board;
- a resolver (used only for comparison purpose).

In particular, the dSPACE, based on floating point microprocessors allows the fast implementation, verification and real-time simulation of algorithms; the control system here used is implemented on the DS1103 board which is equipped with two processors: a master Power PC 604E and a Texas Instruments slave DSP of the type TMS320F240, characterized by cycle frequencies respectively of 400 kHz and 80 MHz.

The controlled hysteresis brake is a Magtrol - INC. model HD-705-8 (6 Nm) and the resolver is a MAGNETIC BLQ-41 used only to measure the real rotor position to be compared with the estimated one to verify the effectiveness of the estimation. The test bench view is shown in the picture of Fig. 4.

The basic drive module under test, composed of a tangential flux IPMSM and an insulated-gate-bipolar-transistor (IGBT)-based converter (the DPS 30) controlled by the dSPACE that manage the converter according to the $i_d = 0$ field-oriented control strategy.

The sensorless control algorithm, including the speed and the current loops, has been fully implemented on the dSPACE

TABLE I
IPMSM PARAMETERS AND RATED VALUE.

maximum speed [rpm]	4000
rated current [A]	3.6
phase rated voltage [V]	77
rated torque [Nm]	1.8
Number of poles	6
Stator phase resistance [Ω]	2.21
Direct axis inductance [mH]	9.77
Quadrature axis inductance [mH]	17.94
permanent magnet flux [Wb]	0.084
Friction Torque [Nm]	0.04

TABLE II
CHARACTERISTICS OF THE DPS 30 POWER ELECTRONIC INVERTER.

Input phase rated voltage (at 50 Hz) [V]	230
DC Link voltage [V]	310
Output max power [kW]	6.5
Output max current (peak) [A]	30
Output max current (RMS) [A]	21
PWM carrier frequency [kHz]	20
Maximum braking power [W]	3200

board with a sampling frequency of $f_s = 10kHz$ while the down sample factor was chosen as $n = 10$.

The output signals of the PWM generator that come out from the dSPACE board are then directly fed to the drivers of the IGBT switches.

In Table I and II the nameplate and per phase parameters of the IPMSM and the electrical characteristics of the IGBT power converter are respectively reported.

B. Results and discussion.

Various tests have been made in order to validate the proposed estimation algorithm. In particular the results of the following tests are herein after reported:

- 1) step change in motor speed from 400 rpm up to 4000 rpm (nominal speed) and back again to 400 rpm;
- 2) sudden application of a 1.8 Nm load torque while the motor run at 4000 rpm speed;

In order to improve the strictness of the tests, the used PI parameters in currents and speed control loops have been locked to the values tuned with the sensed version of the electrical drive. It is clear that a PI re-tuning, taking the presence of estimator into account, could reach a further improvement of the performance in sensorless control.

In the next figures dashed lines are used for the estimated rotor position and speed while solid line are used for the measured ones.

Fig. 5 shows a comparison between the estimated speed and the measured one by the resolver of the test bench for verification purpose. Estimated and real motor speed are in good accordance. Furthermore, the electrical drive shows a very good performance in settling and rise time. The maximum estimation error did never exceeded 35 rpm during transient and become lower (a small oscillation with about 2 rpm peak value) at steady state. Fig. 6 shows instead real and estimated rotor position. It is evident that at steady state there is no estimation error and during the transient a small error appears

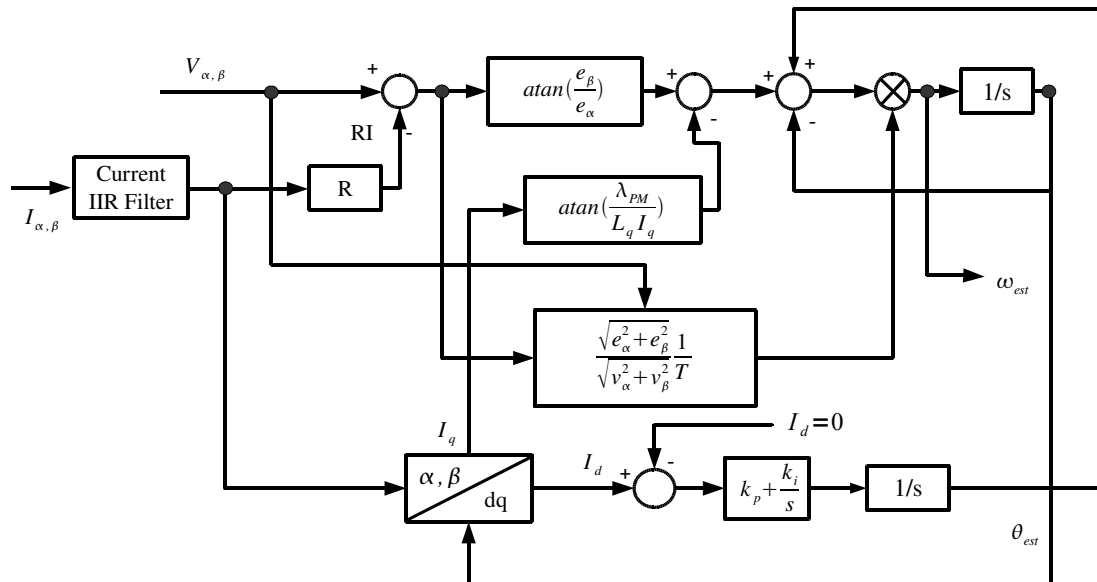


Fig. 3. Estimator block diagram

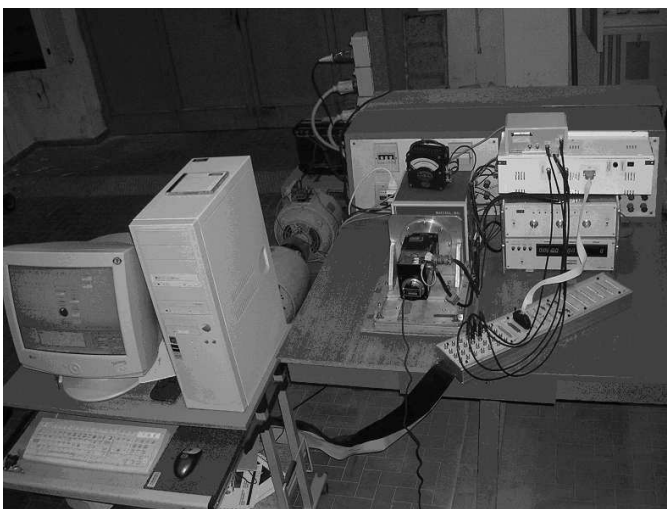


Fig. 4. The Test bench for IPMSM electrical drive machine.

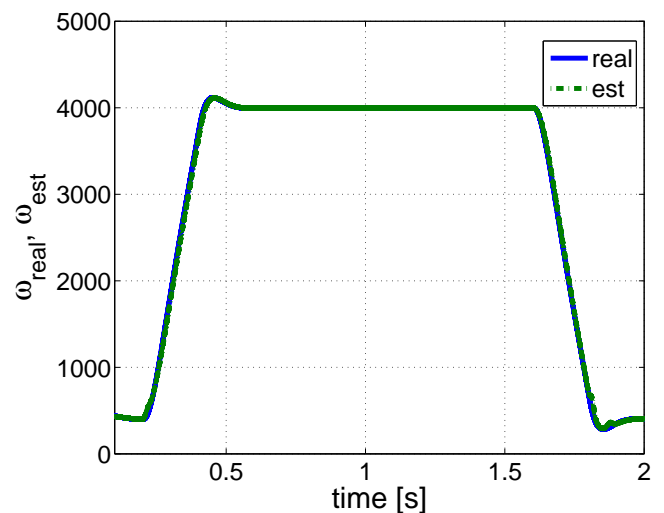


Fig. 5. Comparison between real (solid line) and estimated (dashed line) speed during the execution of test n. 1.

as it can be seen by the little shifting between the two curves. The estimation error vanishes within two cycles.

Test 2 is really significant because the motor works under load condition and it can be observed during a transient operation.

In Fig. 7 estimated and real angular position measured with resolver are shown when an abrupt (step) load torque is applied while the motor rotates at 4000 rpm. Real rotor position is still measured with a resolver on the motor shaft. After the load insertion the position recovery is really fast and only a high level zoom make the rotor position error visible. Real and estimated speed are shown in Fig. 8. It can be observed that suddenly, after torque application, the estimation error increases, but, as Fig. 9 shows, this error tends to vanish and only residual small oscillation and noise due to the presence of the converter remain.

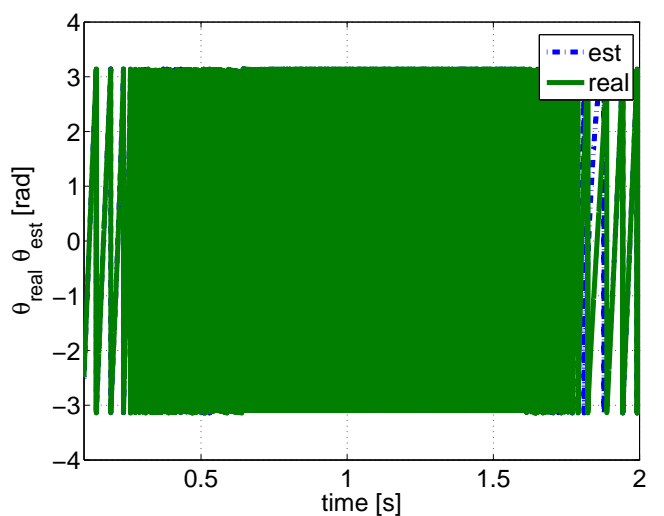


Fig. 6. Real (solid line) and estimated (dashed line) position during the execution of test n. 1.

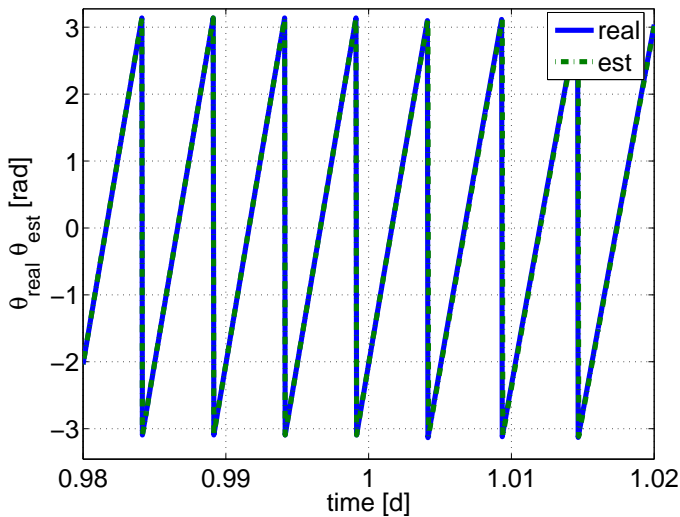


Fig. 7. Comparison between real (solid line) and estimated (dashed line) rotor position during the execution of test n. 2.

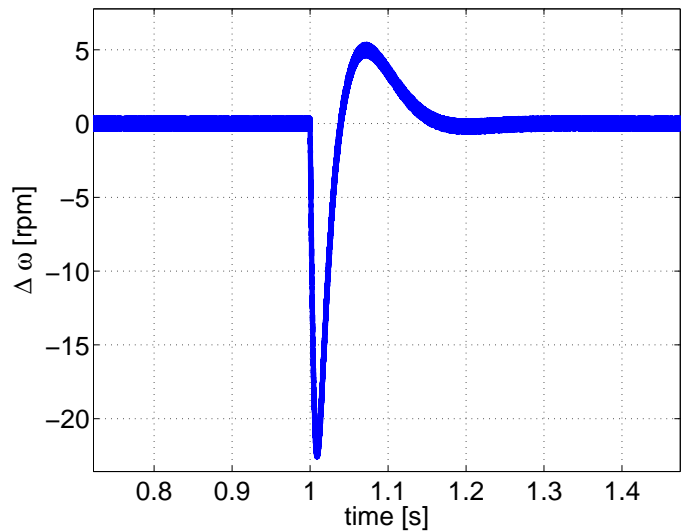


Fig. 9. Estimation error for rotor speed during the execution of test n. 2.

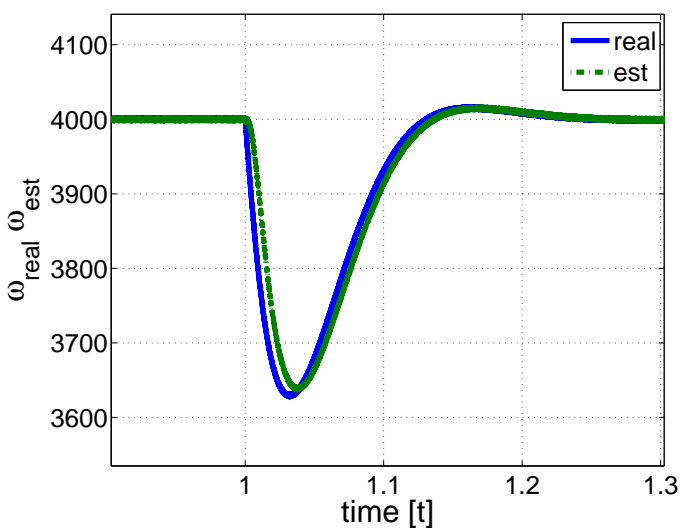


Fig. 8. Comparison between real (solid line) and estimated (dashed line) rotor speed during the execution of test n. 2.

This is a confirmation of the effectiveness and of the proper action of the introduced offset terms.

IV. SENSITIVITY TO PARAMETER VARIATIONS.

As eq. 16 shows the estimation process depends on the values of the stator resistance R and of quadrature inductance L_q . In order to investigate the robustness of the estimation vs the parameter variations the relative sensitivity functions have been calculated. It is clear that in the same equation the term $T_R = \arctan(e_\beta/e_\alpha)$ depends only on the resistance while the current offset term $T_L = \arctan(\lambda_{PM}/L_q i_q)$ depends only on the quadrature inductance.

The sensitivity function related to resistance variation is defined as:

$$S_R = R \frac{dT_R}{dR} \quad (18)$$

Developing this formula the following form is obtained:

$$\begin{aligned} S_R &= R \frac{i_\alpha v_\beta - v_\alpha i_\beta}{R^2 (i_\alpha^2 + i_\beta^2) - 2R (i_\alpha v_\alpha + i_\beta v_\beta) + v_\beta^2 + v_\alpha^2} = \\ &= R \frac{VI \sin(\varphi)}{V^2 - 2RVI \cos(\varphi) + R^2 I^2} \end{aligned} \quad (19)$$

that can be further simplified in:

$$S_R = \frac{R S \sin(\varphi)}{3 |E|^2} \quad (20)$$

where the symbols used are the same of eq. 14 and where $S = 3VI$ is the machine apparent power.

Eq 20 shows that the sensitivity function S_R is very small at high speed when the back EMF reaches its highest values and in each case results to be small at near unity power factor ($\sin(\varphi) \simeq 0$) that is assured by the $i_d = 0$ control.

A similar calculation lead to the sensitivity function on quadrature inductance S_L :

$$S_L = L_q \frac{dT_L}{dL_q} = - \frac{\lambda_{PM} L_q i_q}{(L_q^2 i_q^2 + \lambda_{PM}^2)} \quad (21)$$

from which it can be deduced that sensitivity to variation vs quadrature inductance is smaller at high mechanical loads.

An estimation at the rated values for the motor under test shows that for a parameters variation of 100% the estimation error amount is about 1%. The proposed estimation process appears to be very robust against parameters variation.

V. CONCLUSIONS

In this paper a low time consuming and low cost sensorless control algorithm for PMSM without voltage probes for position and speed estimation was introduced, discussed and experimentally verified.

The sensorless control system is based on the back electromotive force space vector estimation. The use of the back EMF space vector is advantageous respect to any other system using flux estimation because of the integrator elimination avoiding

the problem of integration drift that requires opportune devices or sub systems for its compensation. This simplification also include that the control system will be less susceptible against the EMI external sources. The proposed system in general is more reliable and cheaper than a complicated one without loss of performance respect to other control systems proposed in literature or in industrial applications and also it does not require any voltage probe.

In fact in the proposed control systems the reference voltages, instead of the actual voltages, are used for the back EMF estimation so eliminating the presence of voltage probes. Also the absence of voltage probes imply a significant smaller cost for the proposed system and increased reliability. It is obvious that the choice of this structure, imply errors in position estimation that must be compensated, but this compensation requires only the knowledge of the time delay introduced by the Voltage Source Inverter and the use of Field Oriented Control for the brushless motor with $i_d = 0$. This compensation may also be made by means of pre - collected experimental data for the motor used in the drive. This could allow to avoid on-line tasks if these reveal to be an obstacle in some types of plant.

The comparison with traditional sensorless control methods also based on back-EMF determination shows the high dynamic performance of the proposed algorithm and the effectiveness of the correction terms introduced in the estimation process.

Furthermore a sensitivity analysis with respect to the stator resistance thermal variations have been carried out and the results have shown that the algorithm is robust against this variations. For the typical range of parameter variation the error on estimation is quite not appreciable.

Clearly the presented correction method is intended above all to make the electrical drive cheaper and suitable for industrial drives both surface or internal mounted permanent magnet working within the nominal speed range, as for example for spindle drives, while the field wakening is not taken into account.

Drive starting is made with open loop operation. Speed loop is closed when speed has exceeded almost 10% of the synchronous speed.

In conclusion the proposed algorithm may be considered a very good alternative in terms of economy and precision without lack of performances and, furthermore exhibits an increase in reliability.

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