Self-Commissioning for sensorless Field Oriented Control of PM motors

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Self-commissioning for sensorless field oriented control of PM motors

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***Abstract*— Methods for estimation of the parameters in the inverter and the PM motor, based on special designed experiments, are given. Input to the system is the reference values for the stator voltages given as duty cycles for the Pulse With Modulated power device. The system output is the measured stator currents. Inverter parameters, stator resistance and inductance are measured at stand-still before the initial start-up of the drive. The strength of the permanent magnet is measured during a initial manual start-up of the drive. The measured parameters are used in a field oriented control system and the performance is experimentally verified.**

NOMENCLATURE

a	complex spatial operator $e^{j2\pi/3}$
$i_{sA,B,C}$	stator phase currents A,B and C
$u_{sA,B,C}$	stator phase voltages A,B and C
i_s	stator current complex space vector
u_s	stator voltages complex space vector
ψ_s	stator flux
ψ_r	rotor flux
R_s	stator resistances
L_s	stator inductance
ω	synchronous angular frequency
ω_{mech}	rotor speed
p	time derivative operatord/dt
Z_p	number of pole pair
θ	rotor position
θ_f	filtered rotor position

I. INTRODUCTION

Because standard inverters must be able to operate on different PM motors, the parameters of the PM motor have to be known before starting the drive. Even if the operator were familiar with system identification, it would be too expensive to let him carry out a system identification procedure every time installing a PM motor drive. This calls for an automated process, where the required electrical parameters (inverter parameters for dead time compensation and stator resistance, inductance and the strength of the permanent magnet for field angle estimation) are obtained from on-line measurements. For Induction motors two main groups dealing with this are separated. One is identification during the first power-up, based on special designed test cycles on the motor (Schierling [11], Summer and Asher [12], Vas [13], Rasmussen et. all [7] and [8]), and the other group is on-line identification of parameters in normal operation of the drive (Garces [1], Nilsen and Kazmierkowski [5], Marino et.all [4], Ortega and Espinosa [6], Hurst et. all [2] and Leonhard [3]).

The first group, usually called self-commissioning, is the scope of this paper for field oriented control of PM motors The steps taken in this paper is shortly described by the following.

1. For standard power converters using IGBT's, the turn on and turn off time of the device change significantly with current level, resulting in a noticeable change in output mean-voltage. Particularly at low voltages, this inverter nonlinearity has to be considered when using voltage references instead of measured voltages. This experiment also gives the stator resistance.
2. The current control system is adjusted in advance by an auto-tuning method for PI controllers for the d-axis and q-axis currents [14].
3. The stator inductance is calculated from the relay experiment data or by measuring the open loop current response for a step in the stator voltage.
4. The strength of the permanent magnet is measured in a closed loop current system. Variation of the stator current vector leads to a variation of the angle of the permanent magnet and the strength is computed based on the voltage in the stator.

In the first tree experiments the motor is given a single phase excitation, i.e., two of the phases are impressed the same voltage. No net torque is acting on the rotor and the motor is at stand-still. The stand-still experiments are performed before the initial start-up of the motor. The last item is performed as the first part of the command for running the motor in the direction specified.

II. PM MOTOR MODEL AND ROTOR FLUX OBSERVER

Figure 1 shows the rotor field oriented control system described in details in [10] and [9]. The observer for the angle of the permanent magnet is shown in fig. 2. It is based on complex space phasors for voltage and currents

$$\begin{aligned} i_s &= \tfrac{2}{3}(i_{sA} + a i_{sB} + a^2 i_{sC}) \\ u_s &= \tfrac{2}{3}(u_{sA} + a u_{sB} + a^2 u_{sC}) \end{aligned} \tag{1}$$

The dynamics for the motor is then given by the following stator voltage equation and flux linkage equations

$$\begin{aligned} \frac{d\psi_s}{dt} &= u_s - R_s i_s \\ \psi_s &= L_s i_s + \psi_r \\ \psi_r &= \psi_M e^{j\theta} \end{aligned} \tag{2}$$

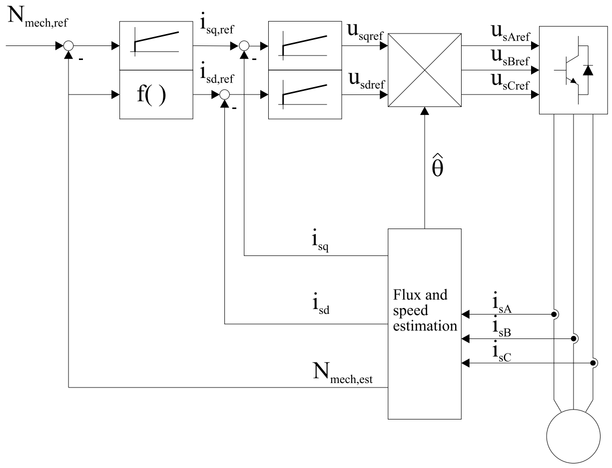

Fig. 1. Rotor Field Oriented Control

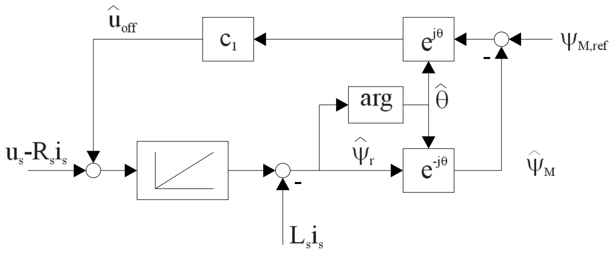

Fig. 2. Signal flow graph of the rotor angle observer

where θ is the rotor angle. The electrical rotor speed is expressed from the mechanical rotor speed as

$$\frac{d\theta}{dt} = \omega = Z_p \omega_{mech} \tag{3}$$

The estimation of ψ_r given in (2) requires an open integration of the voltage equation. The unavoidable offsets contained in the inputs make the output drift. If the offsets are modelled as $\hat{u}_{off}$ the estimator for the rotor flux $\hat{\psi}_r$ is

$$\begin{aligned} \frac{d\hat{\psi}_s}{dt} = &\quad = u_s - R_s i_s + \hat{u}_{off} \\ \hat{\psi}_r = &\quad \hat{\psi}_M\, e^{j\hat{\theta}} \end{aligned} \tag{4}$$

where $\hat{u}_{off}$ has to be designed in a way leading to a flux estimate with constant amplitude $|\hat{\psi}_r|$.

Figure 2 shows how this is obtained by a feedback of the estimated error $\psi_{M,ref} - \hat{\psi}_M$. For known magnitude of the permanent magnet $\psi_{M,ref} = \psi_M$ is used and the observer may then be seen as a rotor field angle estimator.

At zero and very low speed BEMF gives no information of the field angle. A new principle is introduced in this situation by impressing a current in the direction of the default estimated angle. This current then forces the permanent magnet to align to this angle and in this way the estimated angle becomes the correct one only with the error caused by friction and load.

Figure 3 shows the signal flow graph for the rotor speed estimator. The $arg(e^{je})$ block solves the modulus problem.

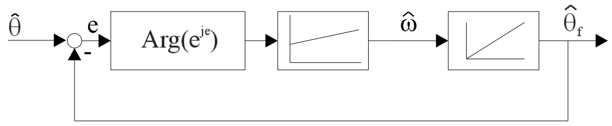

Fig. 3. Speed estimator

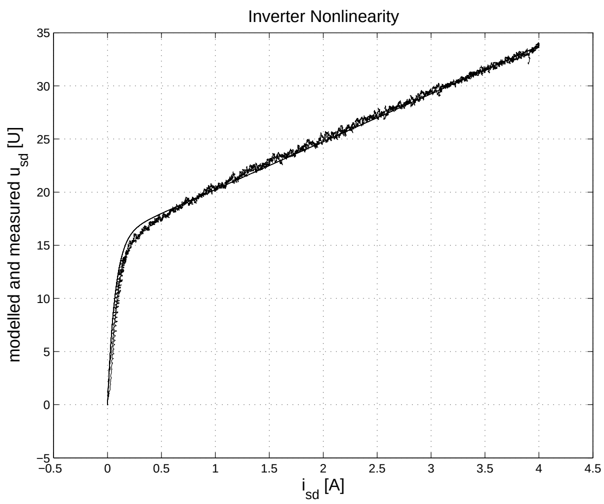

Fig. 4. Inverter nonlinearity giving $R_s = 4.5$, $U_{th} = 11$ and $I_{th} = 0.07$

The algorithm for the estimator then becomes

$$\begin{aligned} \frac{d\hat{\psi}_s}{dt} &= u_s - R_s i_s + c_1(\psi_{M,ref} - \hat{\psi}_M)e^{j\hat{\theta}} \\ \hat{\psi}_r &= \hat{\psi}_s - L_s i_s \\ \hat{\theta} &= \arg(\hat{\psi}_r) \\ \hat{\omega} &= K_p(1 - \tfrac{1}{T_i p}) \arg e^{j(\hat{\theta} - \theta_f)} \\ \frac{d\hat{\theta}_f}{dt} &= \hat{\omega} \end{aligned} \tag{5}$$

III. STATOR RESISTANCE AND INVERTER PARAMETERS

Figure 4 shows the modelled and measured stator reference voltage $u_{s,ref}$ as a function of a slow sweep in the stator current i_s. The stator resistance and the parameters in the inverter model are based on a static measurement of the reference stator voltage $u_{s,ref}$ as a function of the stator current i_s.

The correction of phase voltages are given by

$$\begin{aligned} u_{sA,ref} = &\quad u_{sA} + U_{inv}(i_{sA}) \\ u_{sB,ref} = &\quad u_{sB} + U_{inv}(i_{sB}) \\ u_{sC,ref} = &\quad u_{sC} + U_{inv}(i_{sC}) \end{aligned}$$

with

$$U_{inv}(i) = \begin{cases} +U_{th}(1 - e^{-i/I_{th}}) & \text{for } i > 0 \\ -U_{th}(1 - e^{+i/I_{th}}) & \text{else} \end{cases}$$

In the experiment we have $i_{sB} = i_{sB} = 0.5 i_{sA}$ giving

$$f_{inv}(i_{sd}) = \mathrm{real}\{\frac{2}{3}(U_{inv}(i_{sd}) + aU_{inv}(\frac{-i_{sd}}{2}) + a^2 U_{inv}(\frac{-i_{sd}}{2})\}$$

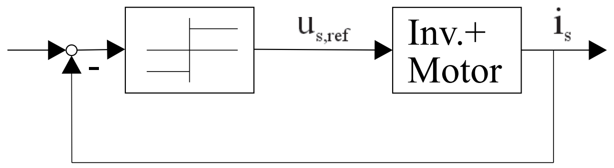

Fig. 5. Setup for relay experiment

A least square fit of $R_s i_{sd} + f_{inv}(i_{sd})$ to the measured value for u_{sd} gives as shown in fig. 4 the parameters R_s, U_{th} and I_{th}.

IV. TUNING OF CURRENT CONTROLLERS

Fig. 5 shows the relay experiment setup for tuning of the current controllers. The tuning of the PI-controller is based on Ziegler-Nichols ultimate period method. An automatic and robust method giving this performance is developed by Åström and Hägglund [14] based on a relay experiment with the controller replaced by a relay. When steady state is achieved, the ultimate period T_u is approximated by the period of the limit cycle.With the notation Δ_u and Δ_i for the change of u_s and i_s, the relay experiment gives, as shown by Åström [14], an approximation of the ultimate gain K_u :

$$K_u \approx \frac{4\Delta_u}{\pi\Delta_i}$$

With K_u and T_u determined from the relay experiment in fig. 6, the parameters for a PI-controller may be computed as $K = 0.45K_u$ and $T_i = T_u/1.2$ if Ziegler-Nichols rules are used. In this paper the tuning parameters are chosen as $K = 0.3K_u$ and $T_i = T_u$. The step response for a current controller tuned with these parameters is shown in fig. 7
With all the motor parameters estimated, a controller, giving a faster response, may be designed and used in the field oriented control system. Lots of laboratory tests have shown that not much is gained in the total performance by doing this.

V. STATOR INDUCTANCE

The stator inductance L_s may be calculated from an experiment where $u_{s,ref}$ is switched between $u^0_{s,ref}$ and $u^0_{s,ref}+d$ as shown in fig. 8. $u^0_{s,ref}$ is the steady state value for $i_{s,ref} = i_{s,0}$ From the initial slope of the curve α_0 the stator inductance is calculated as

$$L_s = \frac{\Delta u_d}{\alpha_0} \tag{6}$$

It is also possible to obtain L_s from the relay experiment using the steepest slope as α_0 in (6). For the motor used the difference between the two methods was less than 10%. The step response method is considered as the most reliable and is used in the following.

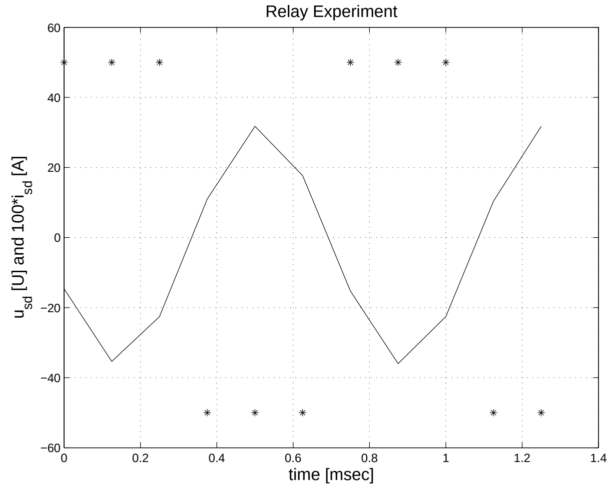

Fig. 6. Relay experiment giving $K_u = 169$ and $T_u = 0.000125$

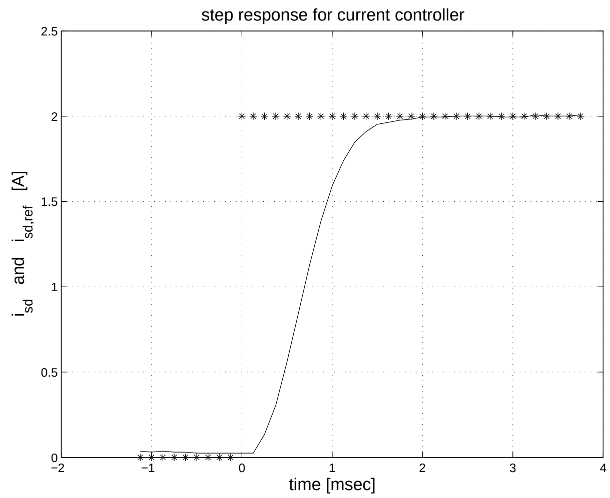

Fig. 7. Step response for current controller

VI. THE PERMANENT MAGNET

The strength of the permanent magnet is measured in a closed loop current system. The stator current vector is rotated with a constant velocity ω_0 giving the relation

$$\psi_M = \frac{1}{\omega_0}(u_{sq} - R_s i_{sq}) - L_s i_{sd}$$

Averaging over a cycle gives the strength of the permanent magnet. Fig. 9 shows that it is possible to obtain a better approximation by looking at the angle dependency.

VII. MEASURED MACHINE PARAMETERS

For comparison the following table shows values for the motor in the test setup and factory measured values at $20°C$

	R_s	L_s	ψ_M	Z_p	N_{mech}
Fact. meass	4.0	0.013	0.30	3	1500
Self comm.	4.5	0.016	0.32		

(7)

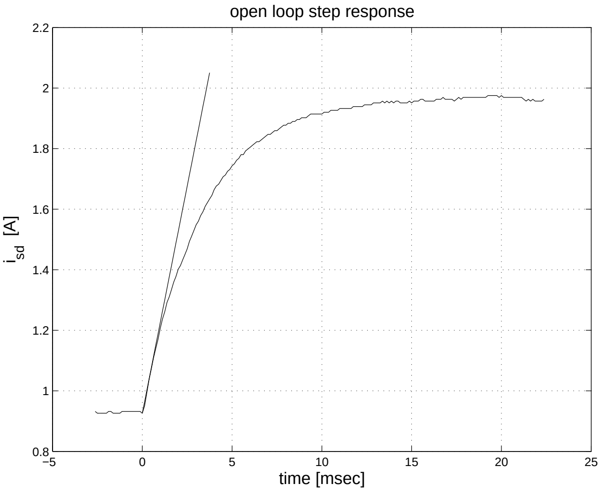

Fig. 8. open loop step response giving $L_s = 0.016$

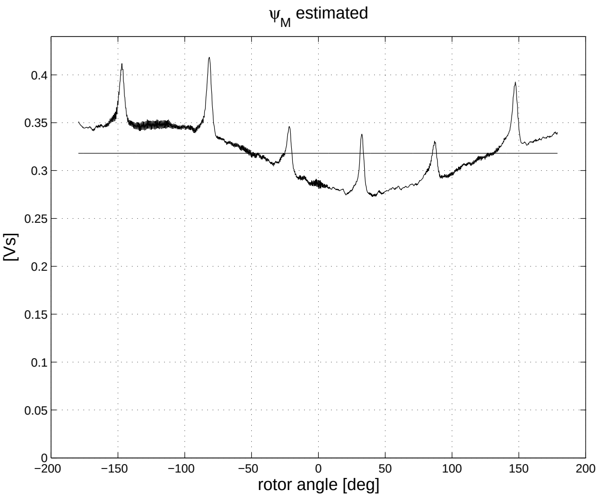

Fig. 9. ψ_M estimated with forced speed during one cycle. Averaging gives $\psi_M = 0.318$

Comments on the result:

- R_s: Because reference voltages to the inverter are used in the estimate of the stator resistance internal output resistance of the inverter is included in the result. This fact as well as the temperature variation may explain the difference between the estimated and the factory measured resistance, and correction would be possible, but because reference voltages are used later in the controller, the measured resistance is more appropriate.
- L_s: The difference between the factory measured and the estimated value may be explained by saturation effects. What value is most correct is difficult to say, and for the overall control performance it is not important at all.
- ψ_M: The difference between the factory measured and the estimated values may be explained by the angle dependency, which is not considered in the factory measured data. The peaks on fig. 9 is due to uncompensated inverter nonlinearities.

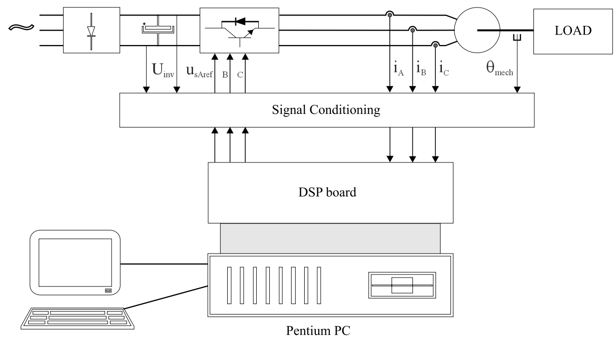

Fig. 10. Laboratory setup

VIII. EXPERIMENTS

Fig. 1 shows the rotor field oriented control system. Having the parameters for dead time compensation, the motor parameters for flux and speed estimation and the current controllers only parameters for the speed controller have to be determined. A relay method as used for the current controllers may be used for on line tuning of PI or PID controller for the speed. At zero and low speed the reference value for i_{sd} is given a value different from zero. If the rotor angle is estimated correctly a value for $i_{sd} \neq 0$ gives no torque. If the rotor angle is estimated with an error the rotor is forced in the direction of $\hat{\theta}$ used in the controller. This new principle means that no manual mode at zero and low speed is necessary, closed loop control is obtained for all values of the speed reference. The function for $i_{sd,ref}$ is

$$i_{sd,ref} = i_{ref0} e^{-|\hat{\omega}_r|/\omega_0} \tag{8}$$

The laboratory setup shown in Fig. 10 is based on Real Time Workshop, Simulink from MathWorks and a DSP based control board from dSPACE. The drive system is via a signal conditioner connected to a DSP board in the computer. The control software is Simulink blocks written in C. Figure 11 shows a step response for the speed and fig. 12 shows the response of the $i_{sd,ref}$-function. A high speed step response is shown in fig. 13 and low speed step response is shown in fig. 14.

IX. CONCLUSIONS

Controllers for the d- and q-axis stator currents in a field oriented coordinate system is automatically tuned based on Ziegler-Nichols ultimate gain method. By using a relay-feedback, approximate values for the ultimate gain and period are found and used for tuning the PI-controller. By impressing different functions for the current reference and measure values for $(i_s, u_{s,ref})$, estimates of the inverter and motor parameters are found. The basic idea of the methods are the same as in the referenced self-commissioning liter-

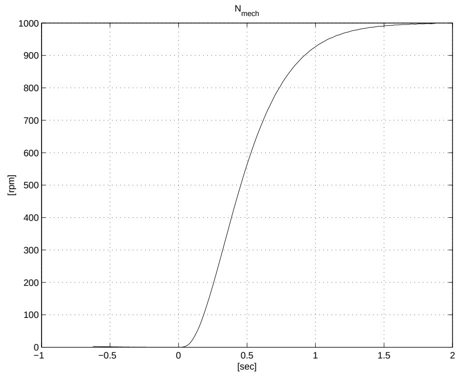

Fig. 11. Normal Rotor speed step response $0 - 1000rpm$

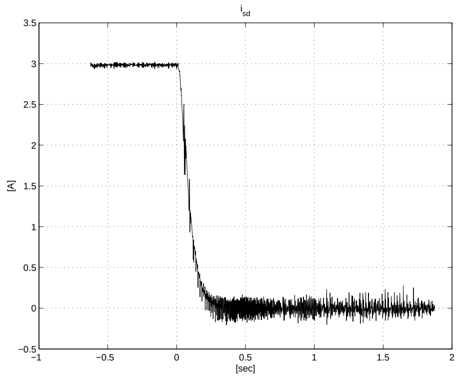

Fig. 12. i_{sd} response for $i_{sd,ref}$ function enabled

ature, but the implementation deviates in many ways from this literature, making it more feasible for a fully automated process. Results for a $800W$-motor are shown and compared to factory measurements. The auto-tuned field oriented control system is tested and the performance of the speed sensor-less drive is excellent.

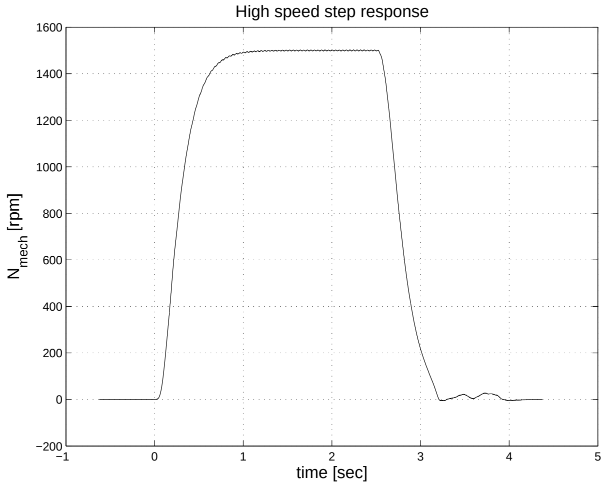

Fig. 13. High speed step response

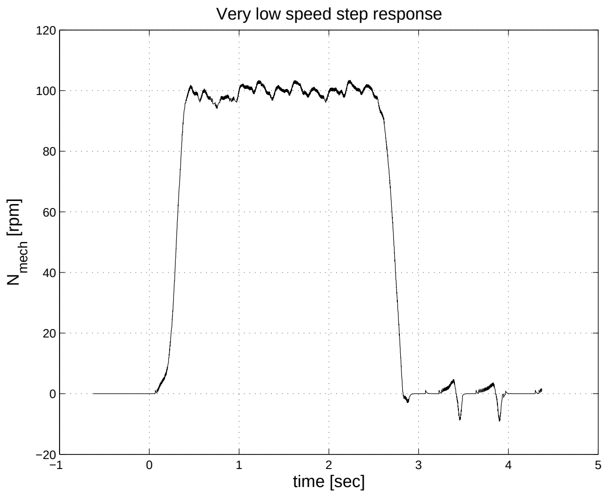

Fig. 14. Low speed step response

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