3. SHAFT SENSORLESS FORCED DYNAMICS CONTROL OF SYNCHRONOUS MOTOR DRIVES

3.1. SYNCHRONOUS MOTOR DRIVE WITH FORCED DYNAMICS CONTROL

Abstract: A new control method for electric drives employing synchronous motors with forced closed-loop dynamics is presented. The system operates without shaft sensors, only the stator currents being measured, the applied stator voltages being determined by the computed inverter switching algorithm with a knowledge of the dc link voltage. The prescribed response to the reference speed demand can be chosen as direct acceleration control, linear first order and second order speed response. The control system, as developed to date, would be suited very well to applications requiring control to a moderate accuracy. Experimental results obtained indicate good agreement with the theoretical predictions.

3.1.1 Introduction

The new FDC based control method for electric drives employing synchronous motors will first be described and then some experimental results presented.

The system operates without shaft sensors, only the stator currents being measured, the applied stator voltages being determined by the computed inverter switching algorithm with knowledge of the dc link voltage. Various prescribed dynamic responses to speed demands are possible, according to *dynamic modes* such as those listed and described in Chapter 1.

The drive control system contains a set of two observers for estimation of the rotor speed and the load torque given the magnetic flux from a flux estimation algorithm. The control system, as developed to date, would be suited very well to applications requiring control to a moderate accuracy. Experimental results obtained indicate good agreement with the theoretical predictions.

It will be recalled that the FDC method was formulated in the α, β frame, when applied to induction motor (IM) drives, yielding oscillatory three-phase stator currents automatically producing a continuously rotating magnetic field in the motor due to an inner oscillatory mode of the closed-loop system, no assumptions being made about the stator currents and fluxes being oscillatory and sinusoidal. In the case of the synchronous motor, however, the method is formulated in the d, q frame to avoid complications due to rotor saliency, in cases where this is present. In contrast with the IM drives, the rotating field is produced instead through the time-varying d, q / α, β transformation, as in other vector control schemes.

The drive control system has a nested loop structure, shown in Fig. 3.1.1 comprising an inner current control loop and an outer speed control loop realising the closed-loop dynamic behaviour of the selected dynamic mode. The inner control loop forces the three-phase stator currents to follow their demands with negligible dynamic lag by setting the switching state of the three-phase inverter to oppose the errors between the demanded and measured stator currents, upon every iteration of the digital processor.

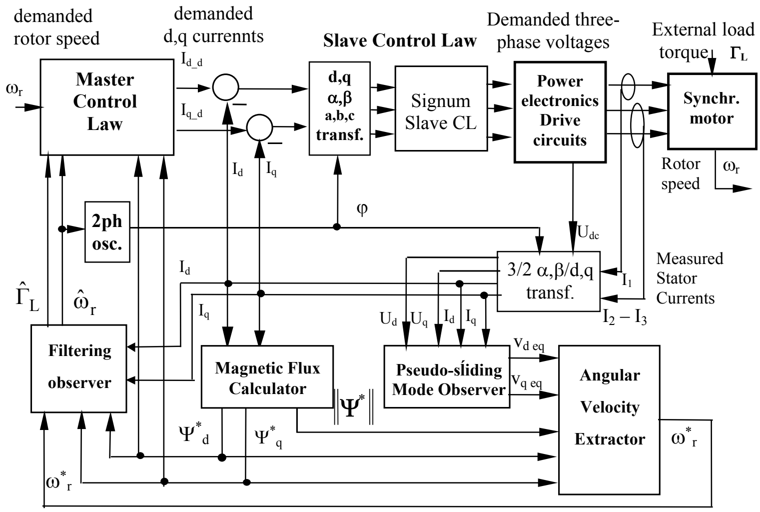


Fig. 3.1.1 Overall control system block diagram

In common with the FDC based approach already created for induction motor drives, *the synchronous motor is treated as a nonlinear multivariable plant in which the control variables are the two stator current vector components and the controlled variable is the rotor speed.* Since there are two control variables and only one controlled variable, there is one degree of freedom available to optimise the performance of the whole system. In this case, the control variables are chosen to maintain the stator current vector and the magnetic flux vector at right angles, as in conventional vector control.

Since the only measurement variables are the stator currents, as for the previously described FDC based IM drive, a rotor speed estimator is employed which requires just these measurements together with the known stator voltages and estimated magnetic flux components from a magnetic flux calculator. An observer whose real time model is based on the motor mechanical equation produces a load torque estimate required by the outer loop control law. This observer is in common with the previously described IM drive control system as was already presented in section 2.1.3. It requires the output of the speed estimator, the measured stator current components and calculated magnetic flux components as inputs.

3.1.2 The Control Law Development

In the interest of simplification, again in common with the IM based FDC system, the control system is arranged in a hierarchical structure [1] in which the stator current demands are generated as primary control variables by a master control law, to be followed closely by a slave control law using the true control variables, i.e., stator voltages, as shown in Fig, 3.1.2.

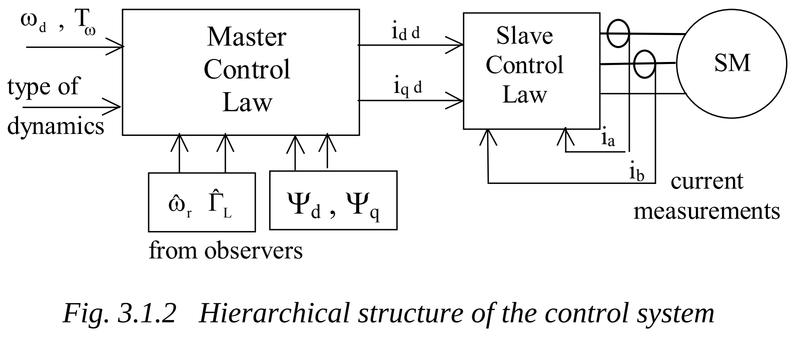


Fig. 3.1.2 Hierarchical structure of the control system

2a) Model of Motor and Load

The permanent magnet SM is described in the synchronously rotating d, q co-ordinate system:

$$\frac{d}{dt}\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \frac{-R_s}{L_d} & p\omega_r\frac{L_q}{L_d} \\ -p\omega_r\frac{L_d}{L_q} & \frac{-R_s}{L_q} \end{bmatrix} \cdot \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \frac{p\omega_r}{L_q}\begin{bmatrix} 0 \\ \Psi_{PM} \end{bmatrix} + \begin{bmatrix} \frac{1}{L_d} & 0 \\ 0 & \frac{1}{L_q} \end{bmatrix} \cdot \begin{bmatrix} u_d \\ u_q \end{bmatrix} \tag{3.1.1, 3.1.2}$$

$$\frac{d\omega_r}{dt} = \frac{1}{J}\left\{c_5\left[\Psi_{PM} i_q + \left(L_d - L_q\right) i_d i_q\right] - \Gamma_L\right\} = \frac{1}{J}\left[\Gamma_{el} - \Gamma_L\right], \tag{3.1.3}$$

where i_d, i_q and u_d, u_q are, respectively, the stator current and voltage components, ω_r is the rotor angular velocity and Γ_L is the total load torque defined in Chapter 1. Ψ_{PM} is the linkage permanent magnet flux, R_s is the stator resistance, L_d and L_q are the direct and quadrature axis inductances, and p is the number of pole pairs.

2b) Master control law

The master control law is the FDC based speed control law and, in common with the FDC based IM drive control system, is derived using the general approach presented in Chapter 1.

First, the master control law is based on the *linearising equation* [2] that yields the first order linear dynamic mode, obeying the following differential equation:

$$\frac{d\omega_r}{dt} = \frac{1}{T_\omega}\left(\omega_d - \omega_r\right). \tag{3.1.4}$$

The motor modelled by equation (3.1.3) for the rotor velocity is forced to follow the desired dynamics of (3.1.4) by equating their right hand sides:

$$\frac{1}{J}\left\{c_5\left[\psi_d i_q - \psi_q i_d\right] - \Gamma_L\right\} = \frac{1}{T_\omega}\left(\omega_d - \omega_r\right) \tag{3.1.5}$$

This is the required linearising function.

The control and estimation algorithms given below are expressed with the state variables (x) replaced by their estimates $(\hat{x})$ and the motor parameters (p) replaced by their estimates $(\tilde{p})$ as they cannot be known with infinite precision in practice.

The rotor magnetic flux calculator estimates the individual rotor flux components, Ψ_d and Ψ_q simply using the well known relations:

$$\begin{aligned} \psi_d^* &= \tilde{L}_d i_d + \tilde{\psi}_{PM} \\ \psi_q^* &= \tilde{L}_q i_q \end{aligned} \tag{3.1.6}$$

The second part of the control law is formulated on the basis of the permanent magnet synchronous motor (PMSM) construction, which has the magnets mounted on the rotor surface. The maximum torque sensitivity [3] is achieved with:

$$i_d = 0 \tag{3.1.7}$$

An equation yielding the required value of i_q is then obtained from equations (3.1.7), (3.1.6) and (3.1.5). On the assumption that the current control loop causes i_d and i_q to follow the demanded currents, i_{d_d} and i_{q_d}, with negligible errors, equation (3.1.7) and the equation yielding i_q together constitute the FDC control law, when i_d and i_q are replaced, respectively by i_{d_d} and i_{q_d}:

$$\begin{aligned} i_{d\_d} &= 0 \\ i_{q\_d} &= \frac{1}{c_5 \tilde{\psi}_{PM}} \left[\hat{\Gamma}_L + \frac{\tilde{J}}{T_\omega} \left(\omega_d - \hat{\omega}_r \right) \right] \end{aligned} \tag{3.1.8}$$

This is the FDC master control law that was first derived for PMSM drives. This master control law may simply be generalised to encompass other dynamic modes by noting in control law (3.1.8) that the term, $\frac{\tilde{J}}{T_\omega}(\omega_d - \hat{\omega}_r)$ is the *inertial torque* (sometimes called the dynamic torque), Γ_I, introduced in the section 1.2.2 and shown in Figure 1.3.1. Then, $\Gamma_I = \tilde{J} acc_d$ where a_d is the *demanded rotor angular acceleration* defined in the section 1.3 for various dynamic modes. The generalised form of control law (3.1.8) is therefore:

$$\begin{aligned} i_{d\_d} &= 0 \\ i_{q\_d} &= \frac{1}{c_5 \tilde{\psi}_{PM}} \left[\hat{\Gamma}_L + \tilde{J} a_d \right]. \end{aligned} \tag{3.1.9}$$

2c) Current Control Law (Slave Control Law)

The slave control law of the inner loop controls the stator voltages so that the stator current components, i_d and i_q, follow the demanded values, i_{d_d} and i_{q_d}. The power electronics and bang-bang inner loop control law are identical to those used in the IM drive and the reader is referred to section 2.1.2c for details. The only difference in the implementation is that the 2/3 transformation indicated in Figure 3.1.1 is time varying and a time varying α, β to d, q urrent measurement transformation is required, due to the control system formulation in the d, q frame.

2d) The discrete time two-phase oscillator

The discrete time two-phase oscillator [6] of Figure 3.1.1 is an unconventional approach to generating the matrix elements, $S = \sin(p\hat{\omega}_r t)$ and $C = \cos(p\hat{\omega}_r t)$ of the rotational transformation needed for the three transformation blocks shown in Figure 3.1.1, *without the need to evaluate the trigonometric functions directly*. They are instead produced directly as transformed state variables C and S of the following discrete time dynamical system:

$$\begin{aligned}
x_1(k+1) &= x_1(k) - (p\hat{\omega}_r h)x_2(k) \\
x_2(k+1) &= x_2(k) + (p\hat{\omega}_r h)x_1(k+1) \\
S(k+1) &= x_2(k+1) \\
C(k+1) &= x_1(k+1) - \left(\tfrac{1}{2}p\hat{\omega}_r h\right)x_2(k+1)
\end{aligned} \tag{3.1.10}$$

where $S=\sin(p\hat{\omega}_r t)$ and $C=\cos(p\hat{\omega}_r t)$.

It should be noted that this scheme would also be applicable in conventional vector control methods.

3.1.3 State Estimation and Filtering

3a) The Pseudo Sliding Mode Observer and Angular Velocity Extractor

This speed estimator uses the measured stator currents and known stator voltages together with the PMSM model to generate a rotor speed estimate and is based on the same principles as the one presented in section 2.1.3b for the IM based drive. The reader is referred to this section for a detailed discussion but the basic derivation is given here in view of the different motor model. The stator current vector pseudo sliding-mode observer used here is based on equations (3.1.1) and (3.1.2) as a real time model but *purposely using only the terms without the rotor speed,* ω_r. Thus:

$$\frac{d}{dt}\begin{bmatrix} i_d^* \\ i_q^* \end{bmatrix} = \begin{bmatrix} \frac{1}{\tilde{L}_d} & 0 \\ 0 & \frac{1}{\tilde{L}_q} \end{bmatrix} \cdot \begin{bmatrix} u_d \\ u_q \end{bmatrix} + \begin{bmatrix} v_{eq\,d} \\ v_{eq\,q} \end{bmatrix}, \qquad (3.1.11)$$

where v_d and v_q are the model corrections, i^*_d and i^*_q are estimates of i_d and i_q as in a conventional observer, but the useful observer outputs here are v_d and v_q. The classical sliding mode observer uses the following bang-bang correction loop actuation function:

$$\begin{bmatrix} v_{eq\,d} \\ v_{eq\,q} \end{bmatrix} = V_{max}\,\mathrm{sgn}\begin{bmatrix} i_d - i_d^* \\ i_q - i_q^* \end{bmatrix}. \qquad (3.1.12)$$

In the sliding mode, v_d and v_q would both switch rapidly to maintain $i_d^* \cong i_d$ and $i_q^* \cong i_q$. Under these circumstances, the short-term average values of $v_d(t)$ and $v_q(t)$, which are denoted the *equivalent values* $v_{d\,eq}(t)$ and, $v_{q\,eq}(t)$ *replace exactly the rotor speed dependent terms* in expressions (3.1.1) and (3.1.2) that are purposely omitted from the real-time model of the observer. Equation (3.1.12) cannot directly generate $v_{d\,eq}$ and, $v_{q\,eq}(t)$. To solve this problem, a *pseudo-sliding-mode* observer [4], as shown in Fig. 3.1.3 (a) may be formed by replacing equation (3.1.12) with (3.1.13):

$$\begin{bmatrix} v_{eqd} \\ v_{eqq} \end{bmatrix} = K_I \begin{bmatrix} i_d - i_d^* \\ i_q - i_q^* \end{bmatrix}. \tag{3.1.13}$$

Here, the gain, K_I, is made as high as possible within the stability limit, since as, $K_I \to \infty$ then $v_d \to v_{d\ eq}$ and $v_q \to v_{q\ eq}$. For large K_I, the errors between real motor currents and fictitious observer currents are driven almost to zero, yielding:

$$\begin{bmatrix} v_{eqd} \\ v_{eqq} \end{bmatrix} = \begin{bmatrix} \dfrac{-\tilde{R}_s}{\tilde{L}_d} & p\omega^*{}_r \dfrac{\tilde{L}_q}{\tilde{L}_d} \\ -p\omega^*{}_r \dfrac{\tilde{L}_d}{\tilde{L}_q} & \dfrac{-\tilde{R}_s}{\tilde{L}_q} \end{bmatrix} \cdot \begin{bmatrix} i_d \\ i_q \end{bmatrix} - \frac{p\omega^*{}_r}{\tilde{L}_q} \begin{bmatrix} 0 \\ \tilde{\Psi}_{PM} \end{bmatrix}. \tag{3.1.14}$$

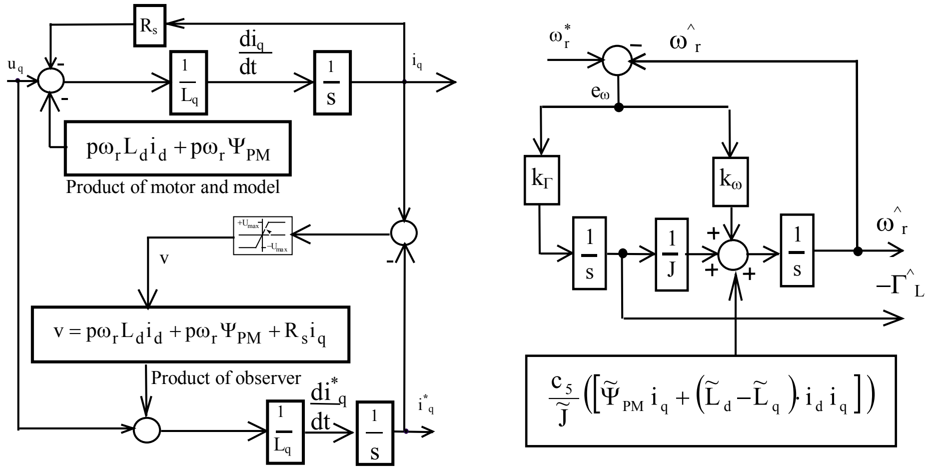
Based on equation (3.1.14) an unfiltered angular rotor speed estimate, ω_r^*, can be extracted as (3.1.15):

$$\omega_r^* = \frac{-\tilde{L}_q v_{eqq} - R_s i_q}{p \cdot \left(\tilde{L}_d i_d + \tilde{\Psi}_{PM}\right)}. \tag{3.1.15}$$

3b) Observer for Load Torque Estimation and Rotor Speed Estimate Filtering

This observer, shown in Fig. 3.1.3 (b) is of the same form regardless of the type of motor and is fully presented in section 2.1.3. In the case of the PMSM drive, the computed electrical torque is given by $\Gamma_{el} = c_5 \left[\Psi_d i_q - \Psi_q i_d\right]$:

$$\begin{aligned} e_\omega &= \omega_r^* - \hat{\omega}_r \\ \dot{\hat{\omega}}_r &= \frac{1}{\tilde{J}} \left(\tilde{c}_5 \left[\tilde{\Psi}_{PM} i_q + \left(\tilde{L}_d - \tilde{L}_q\right) i_d i_q \right] - \hat{\Gamma}_L \right) + k_\omega e_\omega \\ \dot{\hat{\Gamma}}_L &= k_\Gamma e_\omega \end{aligned} \tag{3.1.16}$$



a) Pseudo-sliding mode observer *b) Filtering observer*

Fig. 3.1.3 Block diagrams of pseudo-sliding mode and filtering observer

3.1.4 Experimental Results

The parameters of the permanent magnet SM and ancillary devices used for the experiments are listed in the Appendix. The control law was implemented via a Pentium PC166, the stator currents being measured through LEM transformers and evaluated using a PC Lab Card PCL818 built into the PC. An IGBT transistor module 6MBI10L-060 was used as a three-phase inverter, with a dc bus voltage of U_{dc}=90 V. Data logging of the experimental variables was carried out for a 0,85 s time interval and synchronous motor was idle running.

Experimental results for the FDC based PMSM drive in the constant acceleration control mode are shown in Fig. 3.1.4. The range of rotor speeds achieved is ω_d=20-80 rad/s with prescribed acceleration times T_{acc}=0,05-0,2 s.

Preliminary experimental results for an idle running PMSM drive in the first order linear dynamic mode are shown in Fig. 3.1.5. The achieved control range of the shaft angular velocity is ω_d=20-80 rad/s with a prescribed time constant of T_ω=0,15 s.

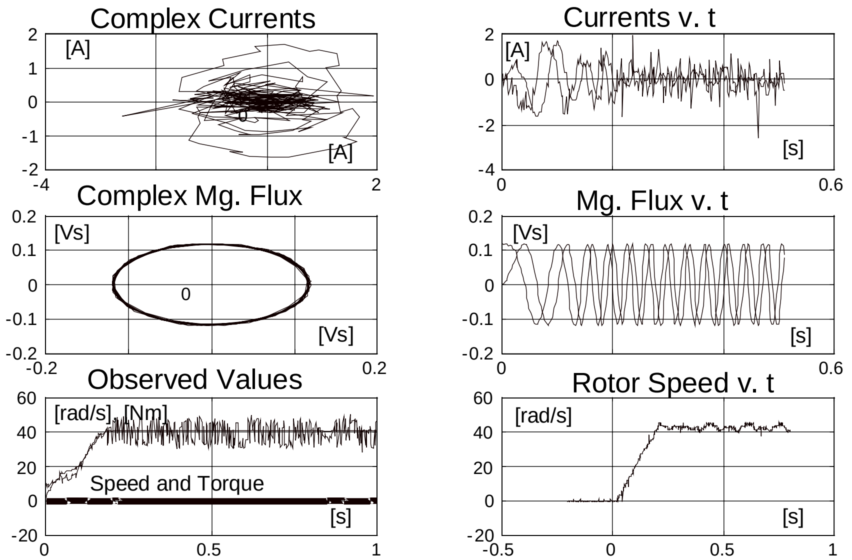


Fig. 3.1.4 Experimental results for synchronous motor in constant acceleration mode

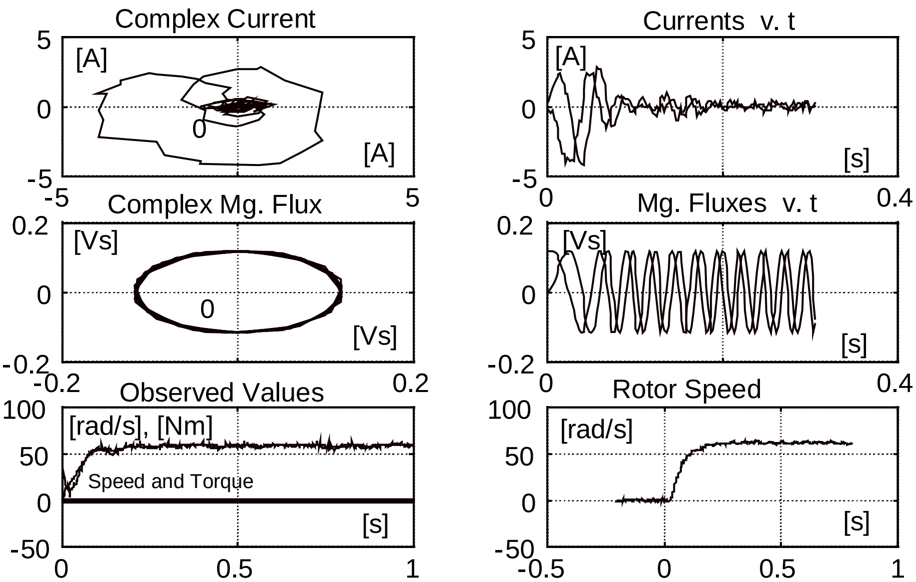


Fig. 3.1.5 Experimental results for synchronous motor with first order dynamic

Preliminary experimental results for the PMSM drive in the second order linear dynamic mode are shown in Fig. 3.2 6. The achieved control range of shaft angular velocity is ω_d=20-80 rad/s and a prescribed settling time of T_s = 0,2 s.

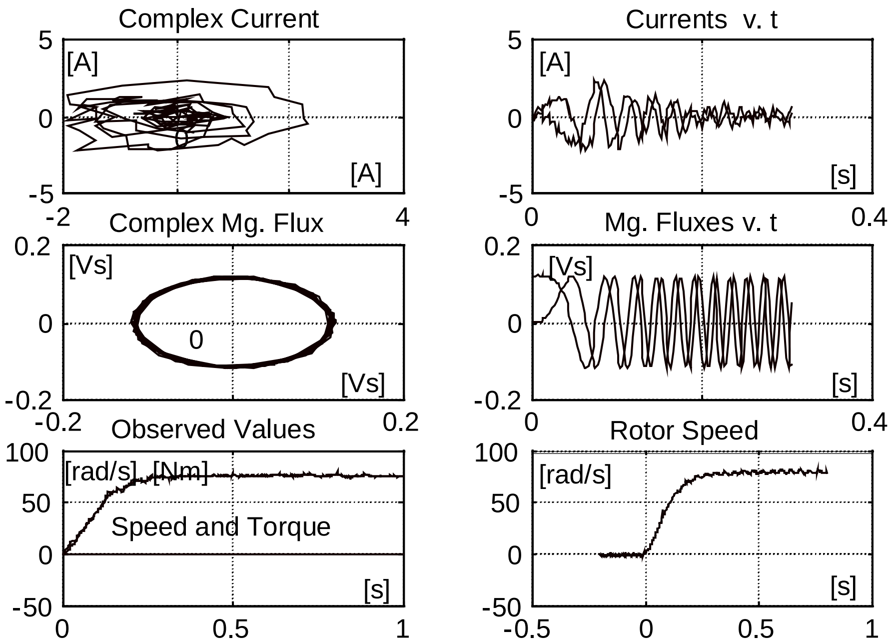


Fig. 3.1.6 *Experimental results for synchronous motor with second order dynamics*

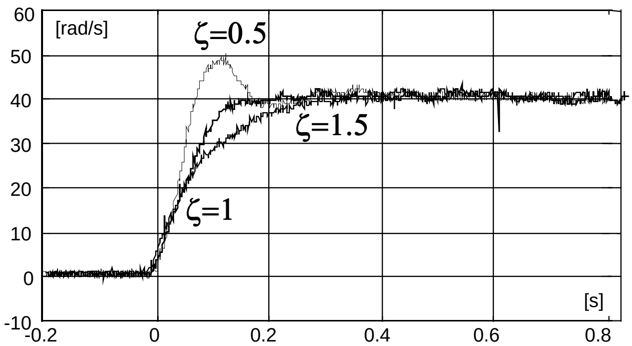


Fig. 3.1.7 *Experimental results with second order dynamics*, ω_d = 40 rad/s, T_ω = 0,1 s *and three various damping factors,* ζ = 0,5, ζ = 1 and ζ = 1,5

Finally Fig. 3.1.7 shows experimental results for the PMSM drive in the second order linear dynamic mode for ξ=0,5 (under-damped), ξ=1 (critically damped), and ξ=1,5 (over-damped).

3.1.5 Conclusions and Recommendations

The preliminary investigations of the proposed new FDC based control method for electrical drives employing permanent magnet synchronous motors show good agreement with the theoretical predictions. This can be clearly observed in figures 3.1.4, 3.1.5 and 3.1.6. It should be noted that the motor was not subject to an external load torque. The significant, though not very large, departure from the ideal performance is due mainly to the non-zero iteration interval, '*h*', and time delay in the load torque estimation as well as due to errors in the motor and load parameter estimation.

The control system, as developed to date, would be suitable for applications requiring sensorless speed control to moderate accuracy ($\approx 5\%$). Further research is required to investigate automatic shaft alignment for the start-up conditions.

3.1.6 References

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Appendix

Three-phase permanent magnet synchronous motor, EVAX EVPU Nova Dubnica are as follows:

Permanent Magnet SM parameters		***Parameters for equivalent circuit***	
Rated power	P_n=720 W	Direct inductance	L_d = 6.06 mH
Rated speed	n_n=3000 rpm	Quadrature inductance	L_q = 5.73 mH;
Rated current	I_n= 3 A	Permanent mg. flux	Ψ_{PM} = 0.119 Vs
Number of poles	2p=8	Stator resistance	R_s = 2.2 Ω
Terminal voltage	U_t=90 V	Moment of inertia	J =3.5.e-4 kgm^2

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