Discrete Multi-Layer Estimator Implementation for Sensorless Control of Elastic Drive Systems – An Industrial Case Study

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Abstract – **Sensorless AC drives do not require speed sensor and have the advantage of high reliability and low cost. The sensorless control techniques require the estimation of internal state variables and the accuracy of this estimation influences the overall control performances. Moreover, implementation of a state observer which computes speed signal and the use of digital filter introduce intrinsic delays which affect electrical drive performances. The result is a deterioration of control performances even if control parameters are correctly tuned. The control of a dual-inertia elastic system is a critical application because of the undesired mechanical oscillations and its sensitivity to motor speed feedback estimation. This paper presents a multi-layer estimator to control a dual inertia elastic system**

I. INTRODUCTION

The performances and the response of any control architecture depend mainly on the quality of the feedback signal, which can be measured or estimated A better quality of them implies better performances of the overall control strategies. The study of sensorless electrical drives, which allow to substitute speed and position measures sensor with their estimation, has allowed a relevant reduction of drive costs especially in small and medium power applications without affecting its reliability. However, a higher robustness cannot be always reached due to implementation issues. The control platform and the estimator unit are usually implemented in the same digital environment such as Digital Signal Processor or dedicated microcontroller [6]. The use of the same sampling time to implement both system estimator and control algorithm could affect the overall system performances because of the presence of different dynamics in the system under analysis. Moreover, the use of digital filters necessary to improve signal quality introduces other delays that can be unacceptable.

This paper proposes a specific industrial application where the above mentioned problems are really critical in terms of control system response: the control of an elastic drive. The presence of a mechanical transmission characterised by an elastic behaviour is a common problem in industrial drives application. The easiest model for an elastic drive system is a dual-inertia mechanical model where the first inertia is related to the electrical motor and the second one is the mechanical load inertia. This kind of system is affected by torsional resonance excitation at low speed [3]. This phenomenon is produced by the pulsating electromechanical torque and by the closed loop interaction that modifies the overall damping of the process [6]. The system is controlled though a nested loop control strategy [4][5][8] able to adjust the speed variation of the motor according to the load process.

Hence, the actual speed feedback estimation is really critical as it has a relevant impact on the speed controller performances. When a resonance frequency is excited it produces effect both on speed and torque: a safety limit for the oscillations amplitude can be evidenced [2]. Speed oscillations are due to different causes: firstly, the control system and the estimator algorithm refer to signals characterised by different dynamics but they are implemented on the same digital environment and discretised with the same sample step. Secondly the presence of digital filters on measured signals causes a delay which contribute to produce undesired oscillations. This paper presents a *discrete multi-layer estimator* that considers the different internal dynamics to solve the above mentioned problems. Special attention is paid to the concept of a *good output quality* and especially to how much a good signal behaviour and time-delays influence it.

II. INDUSTRIAL CASE-STUDY

Nowadays, in the industrial field (i.e., turbo-machinery, liquefied natural gas, rolling mills, metals) the manufacturing plants driven by Variable-Speed-Drives (VSD) are equipped with a mechanical transmission that presents an elastic behaviour. A not-stiff mechanical shaft lead to electromechanical interactions between the VSD and the shaft-line.

A. Torsional Issues on Elastic Drive Model

One of the most critical issue of the aforementioned interactions are the torsional vibrations coming from the

torsional resonance frequencies (TNF) excitations. Typical problems of uncontrolled torsional vibrations are coupling failures, broken shafts, worn gears as reported in [3]. An induction machine driven by a voltage source inverter with a PWM modulator is considered in the paper. The mechanical load has a higher inertia compared with the motor one and the load torque is an available input. The data are reported in Table I.

Table I. Mechanical load characteristics

J_{mot} [pu]	J_{load} [pu]	K_{el} [pu]	C_{el} [pu]
200	1120	78	0.25

The electromechanical interactions that excite the TNF's of the shaft are coming from:
- torque pulsations produced by the VSD (spectrum).
- closed-loop control topology that influences the overall damping of the system and consequently the torsional effects amplification.

First of all, we need to define an operating region for the process under control, then according to the mechanical proprieties of the system components a multi-inertia model can be defined. The number of TNFs is equal to the number of inertias minus one. Usually, a dual-inertia mechanical model is enough to cover most of the common operating region adopted in the industries and the related critical issues. Besides a specific frequency, each TNF is characterized by a damping factor. The full drive topology i.e. multilevel structure, NPC (neutral point clamped), active/passive rectifier unit. With such topology, the number of the inverter levels defines the harmonic spectrum which affects electromagnetic of torque. Usually the inverter unit mainly defines the torque harmonics with higher magnitude. However, these harmonics depends on motor frequency. Consequently, certain torque pulsations can match the TNFs. Furthermore, the mechanical damping can be not the real damping coefficient related to a TNF due to the closed-loop influences. In fact, there is also the electrical damping thus it is better to consider a process damping as function of the previous two. In the industrial field the most used control architecture is based on a *cascade solution* as in Fig.1. Conventional PI controllers are used for current/torque and speed loop. Instead of an explicit flux control loop, a steady reference value for the flux is used. There are different dynamics for the drive and load phenomena. Usually they are split into three groups of variables that have a comparable time constant:

- fastest dynamics: i.e. current, torque
- medium dynamics: i.e. flux
- slow dynamics: i.e. speed, angle

The cascade architecture tries to fulfil this hierarchy by using the slow dynamic variables only in the outer loop and the fastest one only in the inner loop. A different set-up for the controller is done for each cascading loop.
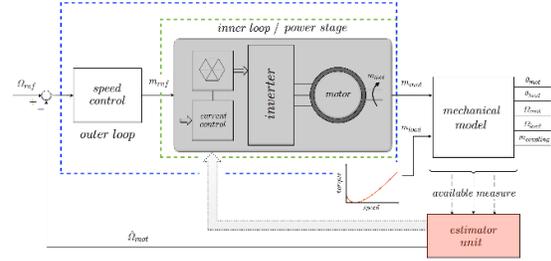

Fig. 1. cascade control

Some detailed explanation of the vibrations generation and control architectures are reported in [4][5][7]. The aim of this paper is to use this industrial example and these particular issues as a case-study to see the different response effects related to different estimator structure.

III. CONTROL AND ESTIMATION ASPECTS

A. Sensitivity to the Estimation

The feedback needed for closing the loop are coming from the estimator unit as in Fig.1. According to the aforementioned TNF excitations oscillation effects are produced both on torque and load speed. The nature of the oscillations is related to a slow dynamic thus for the previous explanation the speed loop. If the controller tuning is poor. the produced resonance effect might be critical (i.e. crank). However, the same results can be reached if the controller is well-tuned but the speed estimation contains errors because in this case it does not give the right reference output. We can summarize the estimation errors in:
- *wrong estimation*, due to the parameters variations in the estimator structure.
- *good estimation but at the wrong time instance*, due to time-delays affecting the estimator structure.

B. Digital Implementation of an Estimator

Let us consider one of the aforementioned cases where the controller is well-tuned but there are problems in the estimator chain. Usually this is one of the most critical cases in the industry because it is related to the latency affecting the digital environment that is sometimes not taken into account. Time-delays can be related to different aspects such as: delays in the signal communication, computational delays, cumulative delays given by filters and/or anti-aliasing filters. Latency aspect becomes even more critical when strictly related to a particular hardware implementation procedure. As explained in [2] a control platform and also an estimator structure can be implemented on a Digital Signal Processor (DSP) or on a dedicated microcontroller. Thus, they are affected by the sampling time of the signals and the processing time itself of the discrete environment. This environment has

computational and actuation performances limited by the microprocessor. Further, the sampling time is not always enough to have a good discrete signal behaviour. Hence, it is necessary to filter the most sensitive signal not to compromise the overall performances. However, a too large use of digital filters is a perfect example of a good estimation procedure at the wrong time instance that implies a bad estimation due to an increasing latency given by the filter itself.

IV. DISCRETE MULTI-LAYER ESTIMATOR

The paper presents a complete discrete-time multi-layer estimator. Based on it, the latency impact on control performances and, for this particular case-study, on the speed estimation is presented in the paper. A suitable tuning of both the layers and of the digital filters can be done keeping a good trade-off between output signal quality and time-delays addiction. Furthermore, the multi-layer structure is based on an open-loop form in order to be reliable as a possible industrial solution.

A. Estimator Structure

The full structure is based on simple state estimators as VI-estimator, $I\omega$-estimator and electrical machine equations. The stator current and the DC link voltage are measured. The measure disturbances e.g. noise are negligible. No high bandwidth is required for the measurement of the DC link voltage while the current measurement should have a high bandwidth. The applied stator voltage vector can be reconstructed from the DC link voltage and the known switching patterns of the inverter though the space vector formula [1]. A simple representation is reported in Fig.2.
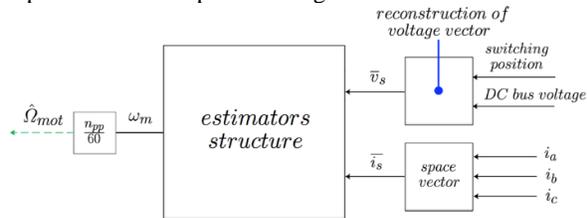

Fig.2. estimator unit representation

Commutation delays and dead times of the inverter are neglected according to the paper aim. Let us consider the stator voltage and current as the inputs and the mechanical speed as the main output. The scheme is reported in Fig.3. Due to the sinusoidal nature of all the variables, electrical machine dynamic model is referred to Park fixed reference frame (α,β). However a rotor flux oriented reference frame (d,q) is adopted for the slip frequency computation.
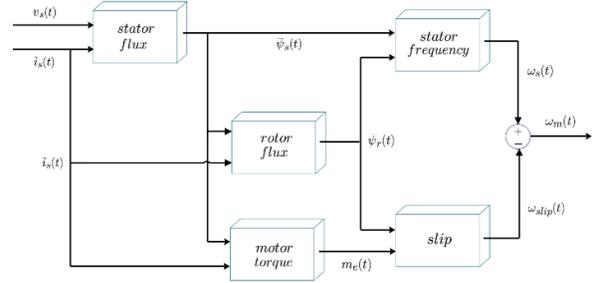

Fig.3. full estimator unit scheme

C. Stator Flux and Torque estimators (α,β)

The stator flux can be integrated in the stationary coordinate system from the voltage space vector like a *VI*-estimator. Hence, from the stator equation of the model (1) are computed the equations in (4).

$$\begin{cases} \psi_{s\alpha} = \int v_{s\alpha} - R_s i_{s\alpha} \; dt \\ \psi_{s\beta} = \int v_{s\beta} - R_s i_{s\beta} \; dt \end{cases} \tag{1}$$

Sensorless becomes critical at zero stator frequency as at low speed the model becomes not globally observable. One suitable solution to this problem is to substitute the pure integral with a low-pass filter. The torque can be calculated from the stator flux and the stator current (2).

$$m_e = n_{pp} Im(\bar{i}_s \underline{\psi}_s) = n_{pp}(\psi_{s\alpha} i_{s\beta} - \psi_{s\beta} i_{s\alpha}) \tag{2}$$

D. Stator frequency estimator (α,β)

The stator frequency computation can be done through the angle θ_s. The information of the angle will be useful for the *dq*-components transformation necessary into the other parts of the control structure.

$$\theta_s = atan\left(\frac{\psi_{r\beta}}{\psi_{r\alpha}}\right) \rightarrow \omega_s = \frac{d}{dt}\theta_s \tag{3}$$

We will pay a special attention to the angle limitation in the digital implementation because we need to limit the variable to $-2\pi \div 2\pi$. If there are many oscillations in between another digital filter in addition to the standard one can be added.

E. Rotor Flux estimator (α,β)

Rotor flux can be based easily estimated using the reduced order model of the induction machine (4).

$$\bar{\psi}_s = L_{ks}\bar{i}_s + \bar{\psi}_r \rightarrow \bar{\psi}_r = \bar{\psi}_s - L_{ks}\bar{i}_s \tag{4}$$

where L_{ks} is defined in (1.5) where L_s L_r L_m are the stator, rotor and mutual inductance respectively.

$$L_{ks} = L_s - L_m^2/L_r \tag{5}$$

F. Slip estimator (d,q)

The slip estimator is based on the $I\omega$ -estimator on the

rotor flux oriented reference frame. We already have the information of the angle θ_s that is a requirement for reference transformation. The equation on *q-axis* of the rotor flux can be rewritten like in (6).

$$\omega_{slip} = \omega_s - \omega_m = \frac{R_r L_m}{L_r}\left(\frac{i_{sq}}{\psi_{rd}}\right) \qquad (6)$$

Where the rotor flux is coming from the previous rotor flux estimation. Now, from (1.6) knowing the stator frequency and the slip frequency we can compute the mechanical speed.

$$\omega_m = \omega_s - \omega_{slip} \quad \rightarrow \quad \hat{\Omega}_{mot} = \frac{\omega_m}{n_{pp}}60 \qquad (7)$$

G. Layers Hierarchy

The choice of the sampling time is crucial for the discretization process in order to be consistent with the industrial equipment e.g. the microcontroller. It is possible, but not useful, to implement the estimator unit in a single time layer where all the computation is executed at the same sampling time. This procedure is shown in Fig.4. As previously reported, the variables do not have the same dynamic and this aspect should be taken into account in the control architecture. If the sample time is based on the fastest dynamic, then an unnecessary high precision of the slower dynamic variable is obtained and an unnecessary computational effort for the microcontroller is produced. On the other hand, if the sample time is decided on the basis of the slowest dynamic the fastest dynamic variables are not computed in the correct way, and thus they are not enough sensitive to variations. Hence, it is quite intuitive that an estimator structure on a multi-layer base with different sampling time could be the right choice. For this application, the paper suggests the following structure and parameters.

$$y(t) = \frac{1}{1+sT_F}x(t) \quad \rightarrow \quad \dot{y}(t) = \frac{1}{T_F}\big(x(t)-y(t)\big)$$
$$\frac{y(k)-y(k-1)}{T_{si}} = \frac{1}{T_F}\big(x(k)-y(k)\big) \quad i=1..3 \qquad (8)$$
$$y(k) = \underbrace{y(k-1)}_{y(k-1)} + \underbrace{\frac{T_{si}}{T_F}}_{G_F}(x(k)-y(k)) \quad i=1..3$$

The faster variables are more sensitive to the time-domain variations, thus the smallest sampling time available named from now $T_{S_1} = 25\mu s$ is chosen. For the medium dynamics is chosen an intermediate sampling time $T_{S_2} = 1ms$ and for the slowest one, such as the mechanical speed and the stator frequency, a sampling step $T_{S_3} = 2ms$ is selected. A representation of the hierarchy is reported in Fig.6. Each block of the scheme is equipped with a digital filter related to the layer sampling time. Starting from a

continuous-time domain a backward Euler rule is selected for the discretization. In this way the poles are constrained in a narrow circle. This allows to avoid unstable behavior but we should also check if are not present distortion in the frequency response.

A discrete representation of a digital filter is given in (8). If the aim is to decrease the time delays, these filters time constants can be decreased, however, this choice implies the presence of a noise signal. Hence, we can consider this as a sensitivity trade-off.

Fig.4. single-layer estimator unit structure

V. PERFOMANCES COMPARISON

The control and the estimator unit are implemented in Simulink in a discrete form according to the previous sampling times. The drive power stage and the mechanical model are implemented in continuous time. The simulation runs with an ODE3 integration step. According to the previous control topology, it is considered a constant nominal input for the reference speed. At the time 3s an input step of the load torque is applied to evaluate the load torque rejection performances.

Fig.5. digital filter implementation scheme

For example, if we assume that the 1[th]TNF is located at low frequency around 17Hz, the content of the step will excite it. Hence, with these setups we are not interested in evaluating the command tracking performances. Although the estimated speed signal is a critical aspect also the coupling torque effects should be evaluated because it is representative of the torque applied to the shaft and it is not directly controlled. High torque oscillation amplitude can produce crack. In fact the amplitude of these oscillations must be limited to $\pm 10\%$ of the rated torque.

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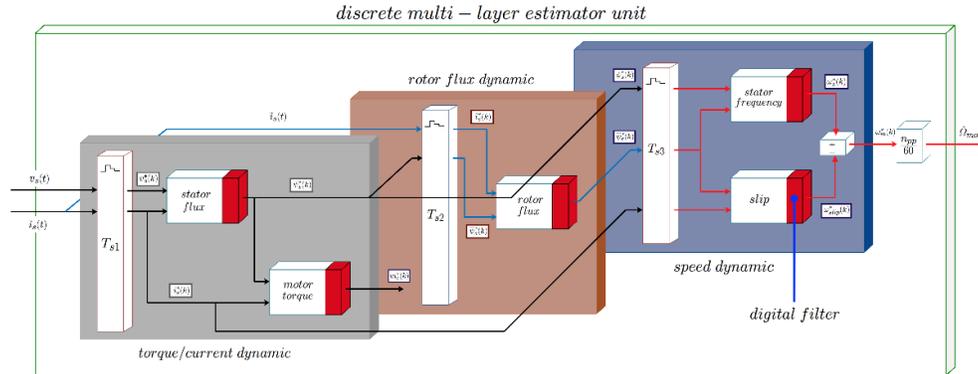Milan, Italy, June 27-28, 2016

Fig.6. full multi-layer estimator unit structure

A. First Setup – Hard Filtering

A first possible setup is a very hard filtering in order to avoid all the effects above the resonance frequency and theoretically limiting the mechanical oscillation effects due to smoother signals. The time constant settings are given in (9), where G_{F1} refers to torque layer, G_{F2} to rotor flux and G_{F3} to speed one.

$$G_{F1} = \frac{25\,\mu s}{0.25ms} = 0.1 \quad \rightarrow \quad f_c = 4kHz$$

$$G_{F2} = \frac{1ms}{10ms} = 0.1 \quad \rightarrow \quad f_c = 100Hz \qquad (9)$$

$$G_{F3} = \frac{2ms}{20ms} = 0.1 \quad \rightarrow \quad f_c = 50Hz$$
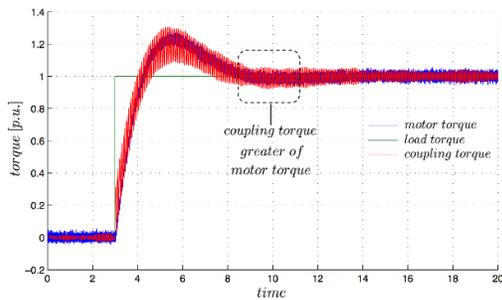
Fig.7. torque step response

Fig.7 shows that the mechanical oscillations in the coupling torque is coincident with the 1[th]TNF. The same effects can be seen in Fig.8 on the motor and load speed. The effects on the load speed are smaller because filtered by a higher inertia instead of the motor one.

It is useful to analyze the torque and speed harmonic content in the time window $12s \div 20s$. Fig.9 shows the magnitude of the 1[th]TNF which is the highest one and it is strictly related to the amplitude of the mechanical oscillations. If the speed controller is correctly tuned it is able to damp the mechanical oscillations. However, the digital filter settings and the inherent delay are responsible

for the oscillations amplitude. With a strong filtering, thus a relevant delay, oscillating behavior is not reduced because the application is much more sensitive to the delays.
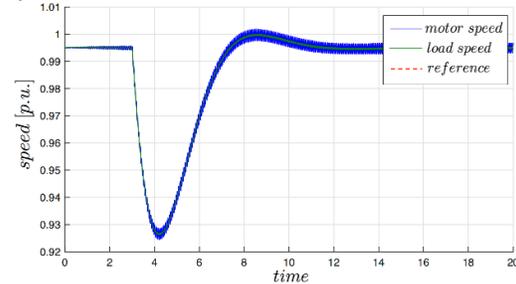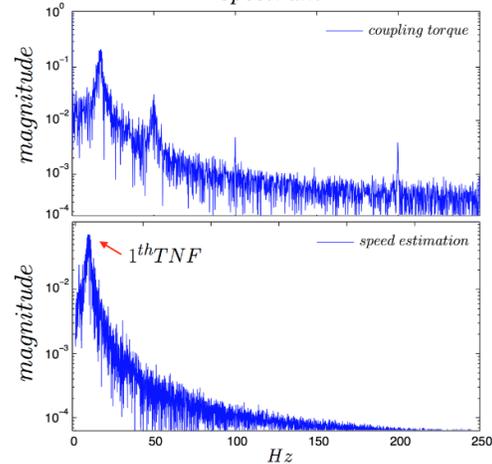

Fig.8. speed step response

Fig.9. torque and speed harmonic contents

B. Second Setup – Reduced Latency

The filter setting for the torque is maintained to filter the higher frequencies due to the intrinsic switching behavior of the drive. However, decreasing the filtering action on the medium and slower level implies to decrease the inherent amount of delays.

$$G_{F1} = \frac{25\mu s}{0.25ms} = 0.1 \quad \rightarrow \quad f_c = 4kHz$$

$$G_{F2} = \frac{1ms}{2ms} = 0.1 \quad \rightarrow \quad f_c = 500Hz \qquad (10)$$

$$G_{F3} = \frac{2ms}{4ms} = 0.1 \quad \rightarrow \quad f_c = 250Hz$$

Tuning the filters as in (10) we obtain better performances in terms of coupling torque oscillations as shown in Fig.10.

Fig.10. torque step response
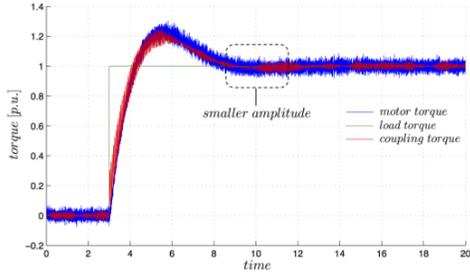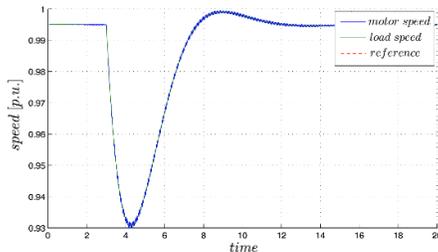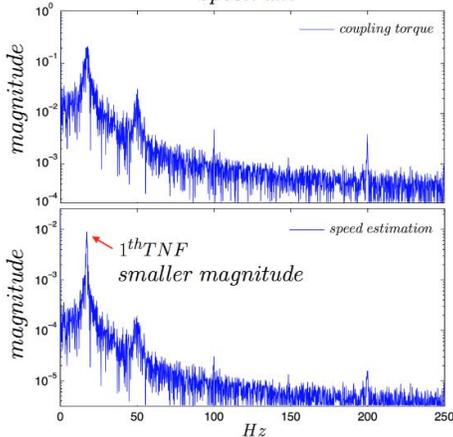

Fig.11. speed step response

Fig.12. torque and speed harmonic contents

In fact, from Fig.12 the harmonic content of the 1^{th}TNF has less magnitude compared to Fig.9. The second setup has better performances with the drawback of higher magnitude at the higher harmonic component in the coupling torque. This result demonstrates that a stronger filtering action does not always increase the overall control system performances. The choice of having a very clean signal behaviour could be the wrong one as the delay in the signal become unacceptable. As in this case-study, this is a very critical aspect in the control architecture that has to deal with the control of different dynamics.

VI. CONCLUSIONS

Digital filters inside a discrete estimator need to be tuned according to the application sensitivity to the delays. In this case-study a wrong estimation due to a bad setting of the filters produces a delayed speed that leads to an underperformed control. Reducing the filtering action meanwhile keeping a good trade-off on the quality output leads to better performances. In this paper the controller is assumed to be able to damp the mechanical oscillations. However, it must be checked if the coupling torque oscillations are under the safety limit. After this, only the digital filters should be managed evaluating the sensitivity to the delays of the specific process. If a reduction of the filtering action improves the control performance, as in this case-study, then the application belongs to the aforementioned category. The paper achieves this conclusion by a complete discrete-time multi-layer structure able to manage both the different dynamics and the side-effects of the digital filters settings in terms of latency.

REFERENCES

[1] M.i Li, J. Chiasson, M. Bodson, and L. Tolbert. "Observability of speed in an induction motor from stator currents and voltages", Proceedings of IEEE conference on decision and control and the European control conference, 2005

[2] API 684, API Standard Paragraphs "Rotor dynamic Tutorial: Lateral Critical Speeds, Unbalance Response, Stability", Train Torsionals and Rotor Balancing, 2005

[3] M. Mauri, M. Rossi and M. Bruha, "Generation of Torsional Excitation in a Variable-Speed-Drive System", IEEE 23nd International Symposium on Power Electronics, Electrical Drives, Automation and Motion, 2016

[4] L. Harnefors, S. Saarakkala, M. Hinkkanen, "Speed Control of Electrical Drives Using Classical Control Methods", IEEE Transactions in Industrial Electronics ,2013

[5] M. Hinkkanen, K. Zenger, S. Saarakkala, "Speed Control of Two-Mass Mechanical Loads in Electrical Drives", IEEE Transactions in Industrial Electronics, 2012

[6] Tajima H., G. Guidi, and H. Umida, "Consideration about problems and solutions of speed estimation method and parameter tuning for speed sensorless vector control of induction motor drives", IEEE Transactions on Industry Applications, 2002.

[7] M. Rossi, M. Mauri, M. S. Carmeli, and F. Castelli-Dezza, "Latency Effect in a Variable Speed Control on Torsional Response of Elastic Drive Systems", International Conference on Electrical Machines, in press, 2016

[8] R. Marino, P. Tornei, and C. M. Verrelli, "A global tracking control for speed sensorless induction

motors", Automatica, 2004