Preprints of the 19th World Congress
The International Federation of Automatic Control
Cape Town, South Africa. August 24-29, 2014

Experimental evaluation of a DO-FPID controller with different filtering properties

M.Huba* ** I. Belai*

* *Institute of Automotive Mechatronics, Faculty of Electrical Engineering and IT, Slovak University of Technology, Ilkovičova 3, 812 19 Bratislava, Slovakia (e-mail: mikulas.huba,igor.belai@stuba.sk)*
** *FernUniversität in Hagen, FMI, PRT, Hagen, Germany*

Abstract: The paper brings experimental results and evaluation to modular disturbance observer based filtered PD and PID controller design based on theoretical results of papers Huba (2013b,c). The approach is focused on achieving the possibly best filtering properties by keeping nearly constant dynamics of the setpoint responses. The developed controller applied to a positional DC motor control is evaluated for different values of the tuning parameter T_D and different filter orders n by using time and shape related performance measures.

Keywords: noise attenuation, filter design, disturbance observer, PD control, position control

1. INTRODUCTION

Besides the traditional robustness-versus-speed trade off of different control approaches one is frequently faced with another trade off related to noise-versus-speed of transients. Is is important especially in situations with a dominant role of noise attenuation, e.g. in servo systems based on incremental position sensors. It is well known that a measurement, or quantization noise have a negative influence on the overall control performance. They contribute to an increased equipment wear, heat dissipation, production costs increase, reduced control precision, undesirable acoustic noise, etc. Thus, one may find many papers dealing with appropriate noise filtration (Åström and Hägglund, 2006; Hägglund, 2012; Larsson and Hägglund, 2012).

A higher attention to a filter design is paid in the disturbance observer based control (Ohishi et al., 1987; Schrijver and van Dijk, 2002; Radke and Gao, 2006; She et al., 2011). However, in all up to now known solutions, this holds just for the disturbance observer loop reconstructing an equivalent input disturbance that is then used for its compensation by a counteracting signal applied to the controller output. In all this publications, the filter design focusses on the possibly best closed loop robustness and does not cover filtration properties of the basic controller used to stabilize the plant. Despite the fact that the effects of noise exposure can be very similar to impact of a lower closed loop robustness.

For tasks with the dominant first-order dynamics this problem has been recently reformulated in papers Huba (2012a,b, 2013a). The corresponding experimental results achieved by an application to a DC motor speed control (Huba and Bélai, 2014b) confirmed a huge performance improvement with respect to traditional two-degree-of-freedom (2DOF) PI control and gave a motivation to deal with this problem also in a more general setup.

⋆ This work has been partially supported by the grants VEGA 1/0937/14 Advanced methods for nonlinear modeling and control of mechatronic systems and APVV-0343-12 Computer aided robust nonlinear control design

An extension of the underpinning ideas to a modular design of a disturbance observer based filtered PD and PID control (FPD and DO-FPID) devoted to plants with a second order dominant dynamics has been treated in Huba (2013c,b). Whereas they have been based on analytical and simulation calculations, this paper reports experimental verification of the proposed approaches in their application to a DC motor positional control.

The paper is structured as follows. Section 2 deals briefly with a pole assignment control of second order plants. An expected performance and deviations from ideal shapes of transients at the plant input and output are discussed in Section 3. Section 4 summarizes basic features of the filtered PD control (FPD) and disturbance observer based filtered PID control (DO-FPID) and its optimal tuning for control loops extended by an unmodelled and filter dynamics. Section 5 describes application of the proposed DO-FPID control to a DC motor positional control. The achieved results are discussed in Section 6 and summarized summarized in Section 7.

2. A SECOND ORDER PLANT AND ITS CONTROL

Firstly, let us is consider design of FPD and DO-FPID controllers for a dominant 2nd-order plant dynamics with the input disturbance d_i, with $\mathbf{x} = (y, \dot{y})'$, $\dot{y} = dy/dt$ and y being the plant state, or output

$$\ddot{y} = K_s(u_r + d_i) - a_1\dot{y} - a_0 y \tag{1}$$

Such a plant with parameters K_s, a_1 and a_0 may be described by a "pole-zero form" transfer function

$$F(s) = \left[\frac{Y(s)}{U_r(s)}\right]_{d_i=0} = \frac{K_s}{s^2 + a_1 s + a_0} \tag{2}$$

For stepwise constant setpoint values r, under pole assignment control of the plant (2) one can require the

Copyright © 2014 IFAC

setpoint-to-output relation characterized by the closed loop poles α_1, α_2, or by the corresponding time constants $T_{ri} = -1/\alpha_i$

$$F_r(s) = \frac{Y(s)}{R(s)} = \frac{1}{(T_{r1}s+1)(T_{r2}s+1)} \tag{3}$$

When considering plant (1) and solving the given problem for u_r one gets a PD controller

$$u_r = K_P(r-y) - K_D\dot{y} + u_c\,;\; u_c = a_0 r/K_s - d_i$$
$$K_P = \frac{\alpha_1\alpha_2 - a_0}{K_s};\; K_D = -\frac{\alpha_1+\alpha_2+a_1}{K_s} \tag{4}$$

Thereby, the feedforward control u_c is necessary for keeping the output in a steady state at the required reference value r under influence of a constant disturbance d_i.

A closed loop with the extended PD-controller (4) remains stable, when its poles remain negative, i.e.

$$K_P K_s + a_0 > 0;\; K_D K_s + a_1 > 0 \tag{5}$$

For stable and marginally stable plants ($a_i \geq 0$) this holds for any $K_P K_s > 0$, $K_D K_s > 0$ and stability will be satisfied for any $0 < T_{ri} < \infty$. For unstable plants with $a_0 < 0$ and $T_{r1} = T_{r2} = T_r$ it must hold

$$T_r < \sqrt{-1/a_0} = T_p \tag{6}$$

i.e. the controller gain K_P cannot be arbitrarily decreased (the closed loop time constant T_r cannot be arbitrarily increased), just to a value fulfilling (6).

The paper deals with the problem that in tuning controllers for real plants, due to a nonmodelled loop dynamics, intentionally used filters, as well as due to the always present measurement and quantization noise, a control designer has to look for some optimal controller gains K_P, K_D, or the task may be expressed as looking for "optimal" closed loop poles α_1, α_2.

3. EXPECTED CONTROL PERFORMANCE

For the plant output changes, the optimal closed loop performance is frequently specified by nearly monotonic (MO) transients corresponding to setpoint step changes (Huba, 2012c, 2013d,a). According to the plant dynamics inversion, the corresponding input has to yield "two-pulses" and will be denoted as a 2P input. At the plant output, ideal disturbance responses may be characterized by "one-pule" shapes.

Measures for evaluating deviations of a signal $u(t)$ from these ideal shapes may be proposed by modifying the Total Variance (TV) measure (Skogestad, 2003)

$$TV = \int_0^\infty \left|\frac{du}{dt}\right| dt \approx \sum_i |u_{i+1} - u_i| \tag{7}$$

For evaluating deviations from strictly MO plant output setpoint response $y_s(t)$ with the initial value $y_{s,0}$ and the final value $y_{s,\infty}$ it is the $TV_0(y_s)$ criterion

$$TV_0(y_s) = \sum_i |y_{s,i+1} - y_{s,i}| - |y_{s,\infty} - y_{s,0}| \tag{8}$$

$TV_0(y_s) = 0$ just for strictly MO response, else $TV_0(y_s) > 0$.

For disturbance responses $y_d(t)$ with a 1P output shape having one extreme point $y_{d,m} \notin (y_{d,0}, y_{d,\infty})$ the TV_1 criterion may be defined as

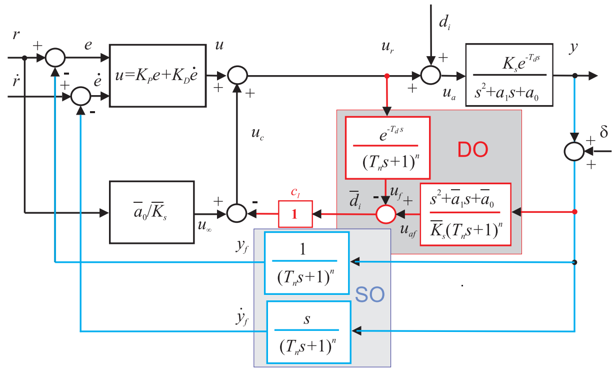

Fig. 1. FPD (with $c_I = 0$) and DO-FPID control ($c_I = 1$) using equal filtration of all channels of the state observer (SO) and of the disturbance observer (DO); δ-quantization noise

$$TV_1(y_d) = \sum_{d,i} |y_{d,i+1} - y_{d,i}| - |2y_{d,m} - y_{d,\infty} - y_{d,0})| \tag{9}$$

Again, $TV_1(y_d) = 0$ just for strictly 1P response, else for output signals with superimposed higher harmonics $TV_1(y_d) > 0$.

Integral deviations from an ideal 2P input shape with two extreme points u_{m1}, u_{m2} may be characterized by

$$TV_2(u) = \sum_i |u_{i+1} - u_i| - |2u_{m1} - 2u_{m2} + u_\infty - u_0| \tag{10}$$

Again, for ideal 2P control functions $u(t)$, $TV_2(u) = 0$.

The speed of the transients at the plant output is usually quantified by the IAE (Integral of Absolute Error)

$$IAE = \int_0^\infty |e(t)|\, dt \tag{11}$$

4. DO-FPD AND DO-FPID TUNING

4.1 *Extended loop dynamics*

To deal with the implementation and filtration problem, the 2nd order plant approximation (1) will now be extended by filters $F_n(s)$ used in all controller channels (Fig. 1) to

$$S_n(s) = \frac{K_s e^{-T_d s}}{(s^2 + a_1 s + a_0)(T_n s + 1)^n}\,;\; F_n(s) = \frac{1}{(T_n s + 1)^n}$$
$$0 < T_n << T_p = \sqrt{1/|a_0|};\; n = 1,\, 2,\, ... \tag{12}$$

In a limit for $n \to \infty$ it was shown to be equivalent to a dead time T_D

$$S_D(s) = \frac{K_s e^{-T_D s}}{s^2 + a_1 s + a_0}\,;\; 0 < T_D << T_p = \sqrt{1/|a_0|} \tag{13}$$

what allows to simplify the overall treatment. Critical and optimal tuning of such a configuration with $a_0 = 0$ have been analyzed in Huba (2013c,b, 2014).

Let us firstly consider a plant delay $T_d \to 0$. Then, for both $F_n(s)$ and T_D located in the feedback one gets the input-to-output transfer functions

$$F_{rn}(s) = \frac{Y(s)}{R(s)} = \\ = \frac{(K_P K_s + a_0)(T_n s + 1)^n}{(s^2 + a_1 s + a_0)(T_n s + 1)^n + K_s(K_P + K_D s)} \\ F_{in}(s) = \frac{Y(s)}{D_i(s)} = \\ = \frac{K_s((T_n s + 1)^n - 1)}{(s^2 + a_1 s + a_0)(T_n s + 1)^n + K_s(K_P + K_D s)} \tag{14}$$

$$F_{rD}(s) = \frac{Y(s)}{R(s)} = \frac{(K_P K_s + a_0) e^{T_D s}}{(s^2 + a_1 s + a_0) e^{T_D s} + K_s(K_P + K_D s)} \\ F_{iD}(s) = \frac{Y(s)}{D_i(s)} = \frac{K_s(e^{T_D s} - 1)}{(s^2 + a_1 s + a_0) e^{T_D s} + K_s(K_P + K_D s)} \tag{15}$$

It is important to note (Huba, 2013b) that both setpoint-to-output responses $F_{rn}(s)$ and $F_{rD}(s)$ and thus also the corresponding characteristic polynomials are the same for the FPD and DO-FPID control.

4.2 Normalized loop parameters

By introducing parameters

$$\overline{K}_{Pn} = K_P K_s T_n^2 \,;\; \overline{K}_{Dn} = K_D K_s T_n \\ A_{0n} = a_0 T_n^2 \,;\; A_{1n} = a_1 T_n \,;\; p = T_n s \\ \overline{K}_{PD} = K_P K_s T_D^2 \,;\; \overline{K}_{DD} = K_D K_s T_D \\ A_{0d} = a_0 T_d^2 \,;\; A_{1d} = a_1 T_d \,;\; p = T_d s \tag{16}$$

they may be normalized to

$$A(p) = (p^2 + A_{1n} p + A_{0n})(p + 1)^n + \overline{K}_{Dn} p + \overline{K}_{Pn} \\ A(p) = (p^2 + A_{1D} p + A_{0D}) e^p + \overline{K}_{DD} p + \overline{K}_{PD} \tag{17}$$

4.3 Optimal loop tuning

Since, in the nominal case, the FPID and DO-FPID characteristic polynomials are equal, an optimal controller tuning based on the closed loop poles may be for both the FPD and DO-FPID control the same. The triple real dominant pole (TRDP) method represents one of the first methods used for an analytical controller tuning (see e.g. Oldenbourg and Sartorius (1944, 1951)). This method requires to fulfill identities $A(p_o) = 0$, $\dot{A}(p_o) = 0$ and $\ddot{A}(p_o) = 0$. Although a corresponding solution exist for any a_0, a_1, for the sake of simplicity it was shown just for the double integrator ($a_0 = a_1 = 0$), when

$$p_{on} = -\frac{2 - \sqrt{2n/(n+1)}}{(n+2)};\; N = n(n+1) \\ \overline{K}_{oPn} = 2\frac{n + (5n+2)\sqrt{2N} - 7N}{(n+2)^2(N + \sqrt{2N})}\left(\frac{N + \sqrt{2N}}{(n+1)(n+2)}\right)^n \\ \overline{K}_{oDn} = \frac{\sqrt{2N} - n}{N + \sqrt{2N}}\left(\frac{N + \sqrt{2N}}{(n+1)(n+2)}\right)^n \tag{18}$$

For the limit case with dead time one gets

$$p_{oD} = -2 + \sqrt{2} \\ \overline{K}_{oPD} = 2e^{-2+\sqrt{2}}(5\sqrt{2} - 7) \\ \overline{K}_{oDD} = 2e^{-2+\sqrt{2}}(\sqrt{2} - 1) \tag{19}$$

As shown in Huba (2014), this tuning approximates well also the dynamics of considered positional motor control.

4.4 IAE closed loop values

In the case of MO setpoint and 1P disturbance responses, when the control error does not change its sign, one may derive the IAE values corresponding to unit input steps of particular inputs by Laplace transform as

$$IAE_{sn} = \frac{a_1 + K_s K_{Dn} - K_{Pn} K_s n T_n}{K_{Pn} K_s + a_0} \\ IAE_{in} = \frac{K_s n T_n}{K_s K_{Pn} + a_0} \\ IAE_{sD} = \frac{a_1 + K_s K_{DD} - K_{PD} K_s T_D}{K_{PD} K_s + a_0} \\ IAE_{iD} = \frac{T_D K_s}{K_{PD} K_s + a_0} \tag{20}$$

4.5 s_o - Equivalence of loop delays

Similarly as in tuning controllers for the 1st order plants (Huba, 2013a), in tuning the FPD and DO-FPID controllers one may get a closed loop performance nearly invariant against the filter order n by requiring a fixed dominant closed loop pole position in the complex s plane, which may be expressed by means of (18) and (19) as

$$s_{on} = p_{on}/T_n = s_{oD} = p_{oD}/T_D \tag{21}$$

what corresponds to a closed loop equivalence among the time constants T_n and T_D

$$T_n = \frac{2(n+1) - R}{(2 - \sqrt{2})(n+1)(n+2)} T_D \,;\; R = \sqrt{2n(n+1)} \tag{22}$$

Thereby T_D may be used as a tuning parameter influencing primarily the speed of transients and the filter parameters (the order n and the time constant T_n) for modifying the noise attenuation.

4.6 Control of loops with mixed delays

The above treated situations considered just the limit situations with either $T_D = 0, T_n \neq 0$, or $T_D \neq 0, T_n = 0$. Thereby, T_D has been considered as an equivalent limit case corresponding to T_n for $n \to \infty$. In practice, one has usually to treat mixed situations, when a loop contains not only a filter dynamics $F_n(s)$, but also a dead time T_d representing e.g. an nonmodelled plant dynamics. Analysis carried out in Huba (2013a) showed that for $T_d << T_D$ the controller tuning may be simplified by using the values K_{oPD}, K_{oDD} (19), whereas the filter time constants are calculated according to (22) with $\overline{T}_D$ substituded for T_D

$$\overline{T}_D = T_D - T_d \tag{23}$$

5. ILLUSTRATIVE EXAMPLE

Since a DO-FPID control has been shown in Huba (2013c,b) to be much more noise sensitive than a FPD control, this comparative experiment deals just with the worse situation with a DO-FPID positional control of HSM 150 DC motor. The identified plant parameters are:

$J = 0.00012$ [kgm^2]; moment of inertia

$B = 0.00016$ Nm.s.rad^{-1}; viscous friction

$T_{GM} = 0.00025$ [s]; torque generator time constant

$\Delta\,\phi = 6.283/10000$; [rad]; position resolution

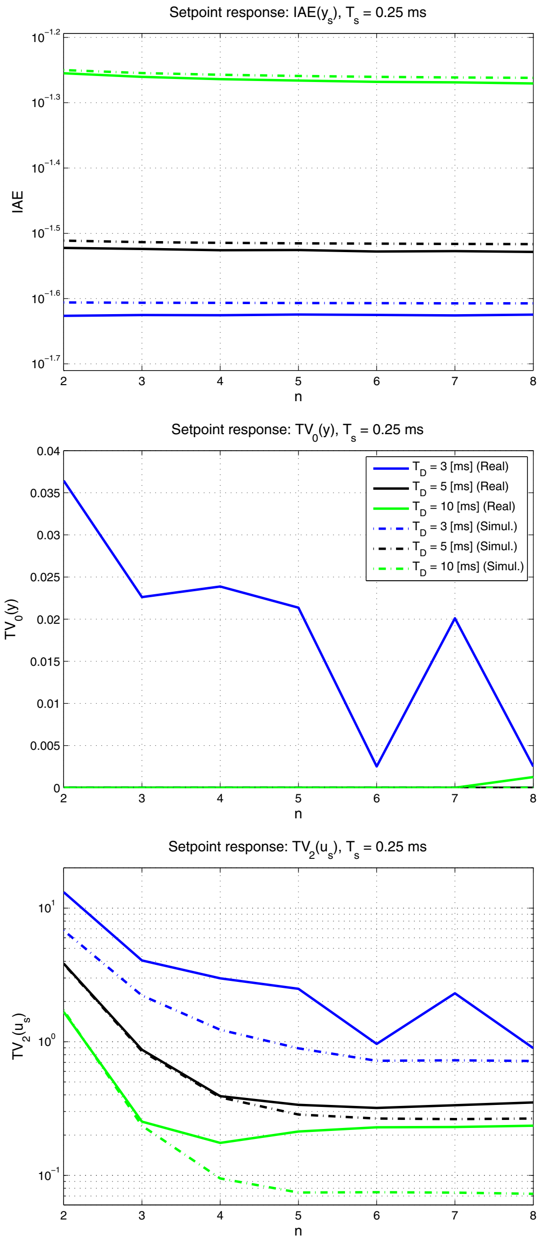

Fig. 2. Setpoint responses: $IAE(y_s)$, $TV_0(y_s)$ and $TV_2(u_s)$ versus n for three different values $T_D = \{3, 5, 10\}ms$

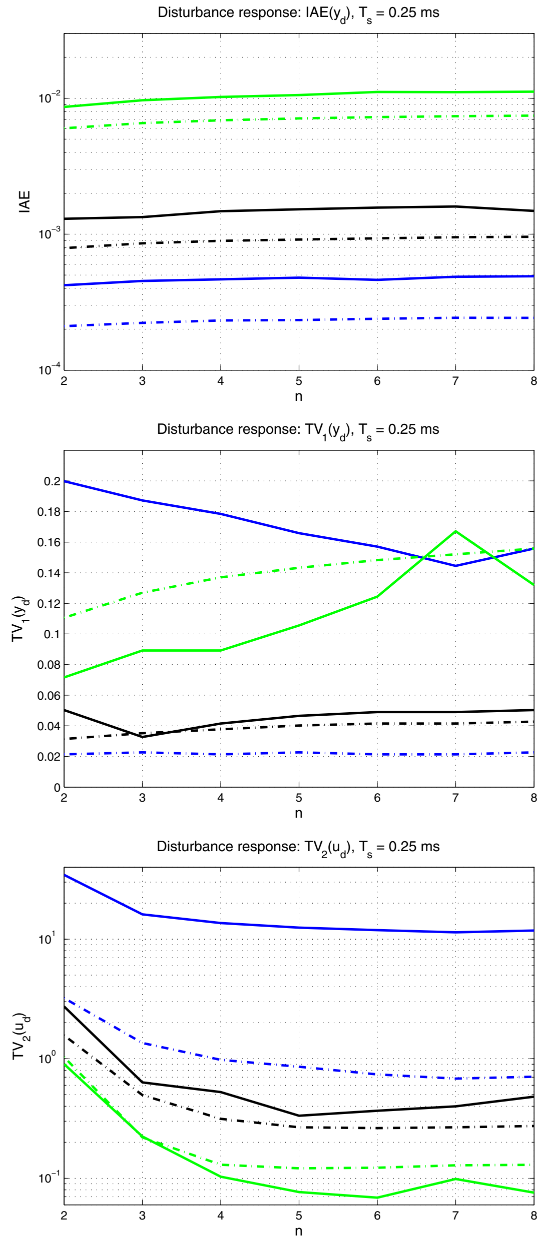

Fig. 3. Disturbance responses: $IAE(y_d)$, $TV_1(y_d)$ and $TV_2(u_d)$ versus n for the values $T_D = \{3, 5, 10\}ms$

When considering the plant model (1) one gets $K_s = 1/J = 8333.3$, $a_1 = B/J = 1.3333$ and $a_0 = 0$. All internal delays (including the torque generator time constant T_{GM}, the electrical time constant and the sampling period $T_s = 0.25ms$) will be approximated by the dead time $T_d = T_{GM} + T_s = 0.5ms$.

For a chosen n, the equivalent dead time allocated to filtration (23) is used instead of T_D in calculating the filter time constant T_n according to (22). The remaining FPID, or DO-FPID controller parameters K_P and K_D are set for T_D according to (19).

Dependences of the simulated and really measured IAE, $TV_0(y_s), TV_1(y_d)$ and $TV_2(u)$ values (for the setpoint and disturbance steps) on n are shown in Figs 2-3. Examples of transient responses for two different n are shown in Fig. 4.

6. DISCUSSION

The IAE curves in Figs 2-3 fully confirm that the equivalence of loop delays (22) allows to tune the control loop for a different noise attenuation by changing n without influencing significantly dynamics of the setpoint and disturbance responses. This holds despite the fact that the used equivalence of the loop delays has been derived for a double integrator, i.e. by neglecting the plant coefficient $a_1 \neq 0$ (Huba, 2014). Thereby, in terms of $TV_2(u_s)$, the noise impact may be significantly reduced (nearly 10 times), but not so strongly as indicated by the simulation results. By shortening T_s this improvement factor increases. Discrepancy between the simulated and really measured setpoint responses could be explained by an additional noise source from the moment generator (PWM, commutator effect). The much larger differences appearing for the disturbance

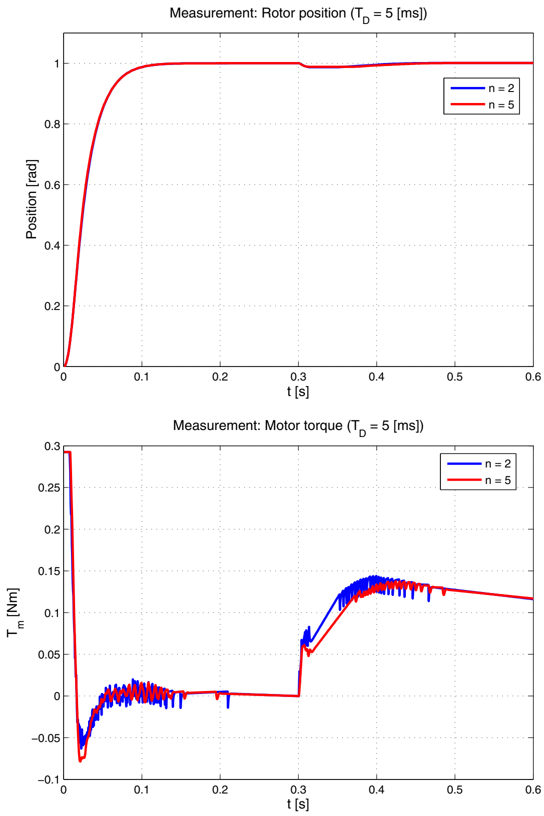

Fig. 4. Setpoint and disturbance step responses of the servo output (above) and input for two different values $n = 2$ and $n = 5$, $T_s = 0.5ms$

responses are obviously caused by a noise produced by the second motor used as a load torque generator. These differences increase by weakening the primary motor control. This noise impact contributes also to increased IAE values of real responses. For a more detailed study, one should identify these up to now nonmodelled noise sources and thus to increase a matching between the simulation and real time control.

Regarding the choice of an optimal filter order n the measured results do not offer a simple interpretation. In each case it is obvious that, when working with a minimal filter order $n = 2$ required for a proper inversion of the plant dynamics, one gets a relatively high noise impact. This may be significantly reduced by $n > 2$. To get a deeper insight into the loop properties, a loop sensitivity analysis Åström and Hägglund (2006) might be helpful. For a noise attenuation evaluated at the control output, from the "gang of four", the noise-to-control frequency gains are important. For $T_d << T_D$, when $F_d(s) = exp(-T_d s) \approx 1/(1 + T_d s)$, one gets

$$\begin{aligned} &C(s) = K_P + K_D s; F_m(s) = K_s/(s^2 + a_{1m} s) \\ &S_F(s) = F_n(s)/(1 - F_n(s)F_d(s)) \\ &L(s) = C(s)S_F(s)F_d(s)F(s)\left(1 + \frac{1}{C(s)F_m(s)}\right) \\ &F_{\delta un}(s) = \frac{U_r(s)}{\delta(s)} = \frac{C(s)S_F(s)}{1 + L(s)}\left(1 + \frac{1}{C(s)F_m(s)}\right) \end{aligned} \quad (24)$$

Fig. 5 shows that for the minimal DO order $n = 2$ the high frequency measurement noise amplification is the highest and for $\omega \to \infty$ it converges to a $|F_{\delta u2}(\infty)| > 0$. For

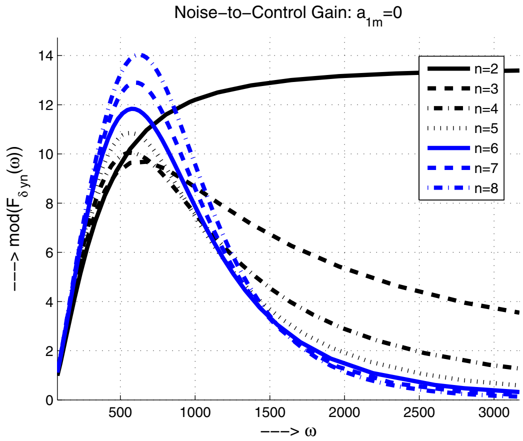

Fig. 5. DO-FPID: Noise-to-control frequency gains (24) for $n = 2 \div 8$, $a_{1m} = 0$ and $T_D = 5$ms

$n > 2$ it already holds $|F_{\delta un}(\infty)| = 0$ and the measurement noise attenuation at high frequencies improves. However, increasing resonant peaks at middle frequencies occur. This indicates that an increase of n will improve the overall filtration just to some degree. Though, since a noise is also generated by the torque generator imperfections (PWM, commutator issues, etc.) that are not covered by the model used, without an experiment it is not evident, up to which moment the trend of improved quantization noise filtering holds, or at which values of n and T_D one may get minimal $TV_2(u)$ values corresponding to the setpoint and disturbance steps. Nevertheless, for a practical use, in majority of situations the default values $n = 3 - 5$ may be considered.

Since the noise impact of other alternative solutions, as e.g. different cascaded, or PID controllers (Žabiňski and Trybus, 2010; Kim, 2009; Huba and Bélai, 2014a) is slightly worse than for the DO-FPID controller with $n = 2$, this structure shows its basic advantage in possibility to achieve a much better noise attenuation.

In the case of general second order systems, the DO-FPID tuning based on TRDP has disadvantage of complex formulas (Huba, 2014). On the other side, it is well known that many practicall approaches use the possibility to approximate a complex plant dynamics by the double integrator models (Fliess and Join, 2008; Han, 2009).

As it was already discussed in Huba (2013b), when specifying a final bound on particular shape related deviations, by the numerical performance portrait method one may achieve an improved control performance compared to the TRDP approach. By this method, an optimal equivalence of the loop delays derivation discussed in Sec. 4.5 could be accomplished fully numerically. But then one looses advantages of an analytical design (Skogestad, 2003).

7. CONCLUSIONS

The carried out experiments fully confirm expectation that by using DO-FPID control one may significantly reduce influence of a measurement and quantization noise. Since this conclusion seems to be in a contrast to the analysis in Sariyildiz and Ohnishi (2013), to get a deeper insight into this problem, a more detailed sensitivity loop analysis, or

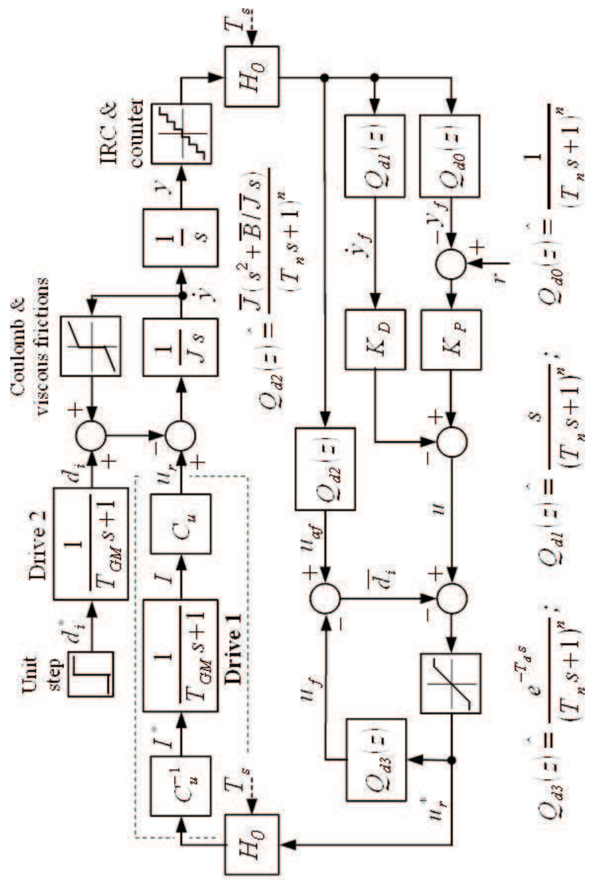

Fig. 6. DO-FPID: Implementation scheme

an analysis by the performance portrait method should be carried out and confronted with these experimental results. Recently, results of this paper have been augmented by a comparison with several traditional PID control structures (Huba and Bélai, 2014a) and extended to a case of constrained control (Huba and Bélai, 2014c).

REFERENCES

Åström, K.J. and Hägglund, T. (2006). *Advanced PID Control.* ISA, Research Triangle Park, NC.

Fliess, M. and Join, C. (2008). Intelligent PID Controllers. In *Proc. 16th Mediterrean Conference on Control and Automation*, 326–331. IEEE, Ajaccio, France.

Hägglund, T. (2012). Signal Filtering in PID Control. *IFAC Conf. Advances in PID Control PID'12*, 2, Part 1.

Han, J. (2009). From PID to Active Disturbance Rejection Control. *Industrial Electronics, IEEE Transactions on*, 56(3), 900 –906.

Huba, M. (2012a). Modular disturbance observer based constrained PI ontroller design. In *Int. Conf. Advances in Motion Control.* IEEE, Sarajevo, BIH.

Huba, M. (2012b). Open and flexible P-controller design. In *Int. Conf. Advances in Motion Control.* IEEE, Sarajevo, BIH.

Huba, M. (2012c). Setpoint versus disturbance responses of the IPDT plant. *IFAC Conf. Advances in PID Control PID'12.*

Huba, M. (2013a). Comparing 2DOF PI and Predictive Disturbance Observer Based Filtered PI Control. *Journal of Process Control*, 23, 10, 1379–1400.

Huba, M. (2013b). Modular PID-controller design with different filtering properties. In *39th Annual Conference of the IEEE Industrial Electronics Society (IECON).* IEEE, Vienna, Austria.

Huba, M. (2013c). Open flexible PD-controller design for different filtering properties. In *39th Annual Conference of the IEEE Industrial Electronics Society (IECON).* IEEE, Vienna, Austria.

Huba, M. (2013d). Performance measures, performance limits and optimal PI control for the IPDT plant. *Journal of Process Control*, 23, 4, 500–515.

Huba, M. (2014). Tuning of a Filtered Pole Assignment Controller for an Integral Plant. In *15th Int. Carpathian Control Conference - ICCC.* Velké Karlovice, Czech Republic.

Huba, M. and Bélai, I. (2014a). Comparison of Two Approaches to a Positional Servo Control. In *15th Int. Carpathian Control Conference - ICCC.* Velké Karlovice, Czech Republic.

Huba, M. and Bélai, I. (2014b). Noise attenuation motivated controller design. Part I: Speed control. In *Speedam Symposium 2014.* Capri, Italy.

Huba, M. and Bélai, I. (2014c). Noise attenuation motivated controller design. Part Part II: Position control. In *Speedam Symposium 2014.* Capri, Italy.

Kim, J.S. (2009). A Feasible Approach for The Unified PID Position Controller Including Zero-Phase Error Tracking Performance for Direct Drive Rotation Motor. *JPE*, 9(1), 74–84.

Larsson, P. and Hägglund, T. (2012). Comparison Between Robust PID and Predictive PI Controllers with Constrained Control Signal Noise Sensitivity. *IFAC Conf. Advances in PID Control PID'12*, 2, Part 1, 175–180.

Ohishi, K., Nakao, M., Ohnishi, K., and Miyachi, K. (1987). Microprocessor-controlled dc motor for load-insensitive position servo system. *IEEE Trans. Industrial Electronics,*, IE-34(1), 44 –49.

Oldenbourg, R. and Sartorius, H. (1944, 1951). *Dynamik selbsttätiger Regelungen. 2nd Ed.* R.Oldenbourg-Verlag, München.

Radke, A. and Gao, Z. (2006). A survey of state and disturbance observers for practitioners. In *American Control Conference, 2006.*

Sariyildiz, E. and Ohnishi, K. (2013). Performance and Robustness Trade-off in Disturbance Observer Design. In *39th Annual Conference of the IEEE Industrial Electronics Society (IECON)*, 3679–3684. IEEE, Vienna, Austria.

Schrijver, E. and van Dijk, J. (2002). Disturbance observers for rigid mechanical systems: Equivalence, stability, and design. *ASME, J. Dyn. Sys., Meas., Control*, 124 (4), 539–548.

She, J.H., Xin, X., and Pan, Y. (2011). Equivalent-input-disturbance approach;analysis and application to disturbance rejection in dual-stage feed drive control system. *Mechatronics, IEEE/ASME Transactions on*, 16(2), 330 –340.

Skogestad, S. (2003). Simple analytic rules for model reduction and PID controller tuning. *J. Process Control*, 13, 291–309.

Žabiński, T. and Trybus, L. (2010). Tuning P-PI and PI-PI controllers for electrical servos. *Bulletin of the Polish Academy of Sciences: Technical Sciences*, 58(1), 51–58.