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Model-based higher-order PID control design

M. Huba

STU FEI Bratislava (e-mail: mikulas.huba@stuba.sk)

Abstract:
Higher order (HO) proportional-integrative-derivative (PID) control aims to fill the gap between traditional and fractional order PID control. Thereby, it allows to fulfill more complex requirements on the target loop performance than it is possible in traditional PID control. This paper illustrates, how HO PID controllers may be derived by generalizing the constructive simple/Skogestad (SIMC) method for analytical model-based controller tuning. Controllers with HO derivative actions are necessarily based on improved design of noise attenuation filters. To evaluate their impact, shape related performance measures of the input and output step responses will be used based on concept of piece wise monotonic signals. Deviations from the ideal shapes due to the noise, design imperfections and plant-model mismatch, establishing additional constraints in designing as fast as possible transients, are quantified in terms of modified total variations. Tuning scenarios based on modified "half-rule" and their superiority in comparison to the IAE optimization based controller design are demonstrated by simulation and by real time experiments. Pros and cons of more complex HO PID controllers are discussed.

Keywords: Filtration, PID control, measurement noise, derivative action, monotonicity.

1. INTRODUCTION

The high interest in fractional-order PID controllers has been apparent for a long time and was also documented by the contributions of the IFAC PID'2018 conference. It should be remembered here that under this "exotic" name there are actually considered controllers, which are ultimately realized by higher-order (HO) filters. Their main benefit in comparison with traditional PID controllers (Tepljakov et al., 2018) is the possibility of obtaining a wider range of performance properties. On the other hand, in the original paper by Skogestad (2003) devoted to simple internal model (SIMC) method for constructive analytical PID controller design, a focus has been paid to the simplest PI and PID control. For first and second order time delayed plant model $F(s)$, a controller $R(s)$ has been designed to get a required closed loop model

$$F_{cl}(s) = \frac{R(s)F(s)}{1+R(s)F(s)} \tag{1}$$

specified by the transfer function $F_{cl}(s)$

$$F_{cl}(s) \stackrel{!}{=} {}^1F_{cl}(s) = \frac{1}{1+T_{c1}s}e^{-T_d s} \tag{2}$$

This may be accomplished by the controller

$$R(s) = \frac{F_{cl}(s)}{1-F_{cl}(s)}\frac{1}{F(s)} \tag{3}$$

When applied to a 2nd order time delayed (SOTD) model

$${}^2F(s) = \frac{K_s e^{-T_d s}}{(s+1/T_{12})(s+1/T_{22})} \tag{4}$$

the method resulted into a not implementable ideal PID controller. Design of an appropriate implementation and noise-attenuation filter has not been treated in details. Thus, the bad reputation of PID control in dealing with noisy loops remained. In a recent paper by Grimholt and Skogestad (2018), the authors made a step towards higher order controllers, however, by presenting integral of absolute error (IAE) optimization based PID controller modifications that should increase the achievable performance limits. And, furthermore, the filtration aspects have not been deeper analyzed again. At the same time, it has been shown that by proper filtration enabling work with higher order derivative actions (see, for example Huba et al. (2018a)) significant performance improvements may be achieved. This raised the question whether also the SIMC method could continue by being extended to design of HO controllers, thus filling the gap between the simplest traditional PID regulators and their substantially more complex generalizations of non-integer orders.

1.1 FOTD based PI controller design

For the first order time delayed (FOTD) plant model with a dead time T_d, time constant T_1 and the plant gain K_s

$${}^1F(s) = \frac{K_s e^{-T_d s}}{s+1/T_1} = \frac{K_s e^{-T_d s}}{s+a} \tag{5}$$

after approximating the exponential term in denominator of (3) by its first-order Taylor series

$$e^{-T_d s} \approx 1 - T_d s \tag{6}$$

for stable FOTD systems ($a > 0$) with

$$T_i = T_1 > 0;\ K_c = \frac{1}{K_s(T_{c1}+T_d)} \tag{7}$$

the SIMC design yields PI controller

$$R_{PI}(s) = \frac{s+1/T_1}{K_s(T_{c1}+T_d)s} = K_c\left(1+\frac{1}{T_i s}\right) \tag{8}$$

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In Skogestad (2003) it has been recommended to choose

$$T_{c1} \geq T_d \tag{9}$$

Next, to explain the controller tuning by stressing possible differences between the identified plant model and the really considered system, parameters of the model (5) T_1 and T_d will also be denoted as T_{1m}, T_m and K_m.

1.2 Example of HO PID controller design

In case of HO processes, as, for example the fourth order time delayed model with a quadruple time constant T_4

$$^4F(s) = \frac{K_s e^{-T_d s}}{(s + 1/T_4)^4} \tag{10}$$

(see E4 in Skogestad (2003)), the original system has been firstly reduced by the "half-rule" method. Formulated for a mix of different time constant, when applied for simplifying plant transfer functions including several delays

- the largest neglected (denominator) time constant (lag) has to be distributed evenly to the effective delay and the smallest retained time constant,
- the effective delay summarizes (besides of above contribution) the original plant delay and different shorter loop delays.

The required closed loop (2), of course, does not respect the causality conditions and brings uncertainty about the results achieved throughout the approach. Therefore, next we pay attention to the situation arising in full respect of the order of the controlled system by the choice

$$^4F_{cl}(s) = \frac{1}{(1 + T_{c4}s)^4} e^{-T_d s} \tag{11}$$

Then, (6) yields a solution of (3) in form of a controller

$$\begin{aligned}
&^4PID(s) = K_c \frac{1 + T_i s}{T_i s} \frac{1 + T_{D1}s + T_{D2}s^2 + T_{D3}s^3}{1 + T_{f1}s + T_{f2}s^2 + T_{f3}s^3} \\
&T_i = T_4;\ T_{D1} = 3T_4;\ T_{D2} = 3T_4^2;\ T_{D3} = T_4^3 \\
&K_c = \frac{1}{K_s(4T_{c4} + T_d)T_4^3};\ T_{f1} = \frac{6T_{c4}^2}{4T_{c4} + T_d}; \\
&T_{f2} = \frac{4T_{c4}^3}{4T_{c4} + T_d};\ T_{f3} = \frac{T_{c4}^4}{4T_{c4} + T_d}
\end{aligned} \tag{12}$$

In difference to the ideal PID considered in Skogestad (2003), the proper ^4PID transfer function simplifies its implementation and evaluation of noise attenuation filters. Although the approximation of monotonic step responses by (10) is nearly as simple as for FOTD models (5) (Strejc, 1959), next we are going to show, how this controller based on a HO approximation may be used to get as fast as possible, but thereby also smooth transients based on simple integral plus dead time (IPDT) plant models.

2. CONTROL OF INTEGRAL PLANTS

Integral models are frequently used as approximations of more general systems, as those with long time constants, or nonlinear systems, where they represents the simplest linear approximation and play central role in several popular approaches as Model Free Control (Fliess and Join, 2014), Active Disturbance Rejection Control (Gao, 2014), or feedback linearization (Isidori, 1995). Tuning of PI and PID controllers for IPDT models achieved with appropriate model reduction techniques, as e.g. the "half-rule" (Skogestad, 2003) is frequently treated in all control

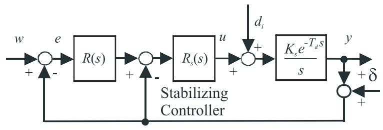

Fig. 1. Cascade control of IPDT plants with a stabilizing controller $R_s(s)$ and a SIMC controller $R(s)$; δ - measurement noise

areas for a broad range of processes. In the simplest case, even those approximated originally by FOTD model (5) (by putting $a = 0$). Consequently, high number of "optimal" tuning rules based on this model may be found (O'Dwyer, 2009). Among the first ones we could mention the experimental controller tuning by Ziegler and Nichols (1942), analyzed in details in transition to sampled-data control (Takahashi et al., 1971) and still giving inspiration to many new approaches (Hägglund and Aström, 2002; Åström and Hägglund, 2004; Šekara and Mataušek, 2010). However, they may not be directly treated by the approach presented in Section 2.1. Namely, when $T_1 \to \infty$

$$F(s) = \frac{Y(s)}{U(s)} = F_0(s)e^{-T_d s};\ F_0(s) = \frac{K_s}{s} \tag{13}$$

the PI design degrades into the proportional control with a poor rejection of input disturbances (Skogestad, 2003). In case of integral models, Skogestad abandoned the constructive model-based design and instead addressed the question of an optimal tuning of add-hoc chosen PI controller. Its tuning has been proposed by analyzing conditions of the critically damped closed loop with the PI controller and integral delay-free plant ($T_d = 0$), when the double real dominant pole method yields

$$T_i = 4/(K_s K_c) \tag{14}$$

To consider T_d in above equations, $T_{c1} = T_d$ was firstly chosen which yields simple and easy to remember SIMC tuning rules $K_c = 1/(2K_s T_d)$; $T_i = 8T_d$, which, in comparing with IMC, resulted into a reasonable improvement of the input-disturbance response. Yet simpler they may be expressed by dimensionless parameters

$$^1\kappa = K_c K_s T_d;\ \tau_i = T_i/T_d;\ i_{ae} = IAE/(K_s T_d^2); \tag{15}$$

$$\text{SIMC } ^1\text{PI:}\ \ ^1\kappa = 0.5;\ \ \tau_i = 8;\ i_{ae} = 16 \tag{16}$$

And still, the analytical double real dominant pole derivation dominated in the applied design.

2.1 Stabilization of integral plants by cascaded control

Instead of the simplified tuning applied above, an alternative approach may be proposed by applying a two-step stabilization-based design In the first step, the integral system (13) may be stabilized by a P, PD, or PDA (proportional-derivative-accelerative) control proposed by the multiple real dominant pole method with possible inclusion of the filtration aspects (e.g. by the half-rule method). In the second step, dominant dynamics of such stabilized loops may be approximated by stable second order, third order, or fourth-order models. Then, the analytical SIMC design may be directly applied. In order to illustrate such more complex situations, an ideal stabilizing PDA controller has been chosen defined as

$$R_s(s) = K_P + K_{D1}s + K_{D2}s^2 \tag{17}$$

For IPDT plants it yields the characteristic quasi-polynomial

$$A(s) = se^{T_d s} + K_s K_{D2} s^2 + K_s K_{D1} s + K_s K_P \quad (18)$$

The condition of quadruple real dominant pole s_o in $A(s) = (s - s_o)^4 A_{red}(s)$ requires to fulfill

$$\left\{ A(s);\ \frac{d}{ds}A(s);\ \frac{d^2}{ds^2}A(s);\ \frac{d^3}{ds^3}A(s) \right\}_{s=s_o} = \mathbf{0} \quad (19)$$

From $d^3A(s)/ds^3 = [sT_d^3 + 3T_d^2]e^{T_d s} = 0$ it follows

$$s_o = -3/T_d; \quad T_o = -1/s_o = T_d/3 \quad (20)$$

$$K_{Po} = \frac{27}{2K_s T_d e^3};\ K_{D1o} = \frac{5}{K_s e^3};\ K_{D2o} = \frac{T_d}{2K_s e^3} \quad (21)$$

Thus, since the stabilized loop has a quadruple dominant pole s_o and a unity steady-state gain, it may well be approximated by the transfer function (10) with parameter estimates $\overline{T}_4 = T_o = T_d/3$ and $\overline{K}_s = 1/T_o^4 = 81/T_d$. For cascade PDA-^{4}PID control with $T_f \to 0$

$$\begin{aligned} IAE &= \lim_{s\to 0} E(s) = \left[\frac{T_i K_s}{K_P K_s[K_c(1+T_i s) + T_i s]}\right]_{s=0} = \\ &= \frac{T_i}{K_c K_P} = \frac{2}{27}e^3 K_s T_d(4T_{c4} + T_d);\ i_{ae} = \frac{2}{27}e^3(4\tau_{c4} + 1) \end{aligned} \quad (22)$$

Since in dimensionless parameters it is realistic to set $\tau_{c4} \approx 0.5$, the achievable performance $i_{ae} \approx 4.5$ may nearly be up to 4 times better than for PI control with $i_{ae} = 16$.

2.2 Noise attenuation filters

Due to the proportional term, the high frequency noise is not attenuated even in the simplest case of PI control. Whereas some authors add in such situations 2nd order Butterworth filters (Segovia et al., 2014), other recommend the simplest binomial filters (Rivera et al., 1986)

$$Q_n(s) = 1/(T_f s + 1)^n \quad (23)$$

The binomial filters will be included into the plant delay by a modification of the "half-rule" (MHR).

Definition 1. (Modified Half Rule MHR). When working with a combination of FOTD system (5) with $n$tuple filter time constant T_f of (23), the plant parameters considered in design should be modified according to

$$T_1 = T_{1m} + nT_f/2;\ T_d = T_m + nT_f/2 \quad (24)$$

For the IPDT model (13) it is enough to modify T_d and with respect to the stabilizing controller (17) to guarantee the stabilizing loop causality by choosing $n \geq 2$.

3. PERFORMANCE MEASURES

Higher emphasis on the measurement noise attenuation is being put today by numerous authors. In the new setup of the optimal control design (Segovia et al., 2014), one has to deal with a trade-off between control error attenuation (measured usually in terms of integral of absolute error)

$$IAE = \int_0^\infty |e(t)|\, dt\ ;\ e = w - y\ ;\ w = setpoint \quad (25)$$

measurement noise injection influencing primarily the "excessive control effort" (controller activity, input usage, but including possibly also the "output wobbling" Huba (2019a)) and the "robustness". In this paper we combine the IAE values (used for evaluating speed of the transients, or, more precisely, the loop retardation which is inverse to the considered speed) with the shape related constraints. These may be applied both in analytical derivations and in experimental evaluation. Since the setpoint step responses may also be optimized by an appropriate feedforward, the analysis will preferably focus on the (input/load) disturbance step responses given fully by the feedback controller. Furthermore, since the achieved responses strongly depend on possible uncertainty of the considered plant model (with an uncertainty impact similar to external disturbances (Chen et al., 2016)), this gives additional motivation to deal with the disturbance responses as an indispensable part of the robustness analysis.

3.1 Useful and excessive output and input increments

Together with the requirement to have the output responses as fast as possible (wit as low as possible IAE), it is also necessary to consider shapes of actual output and input signals. The corresponding measures based on deviations from monotonicity represent modifications of the total variation (the total sum of absolute increments (Skogestad, 2003)). For one monotonic interval (with an initial value y_0 and a final value y_∞) this yields

$$\begin{aligned} TV_0(y) &= \textstyle\int_0^\infty \left(\left|\frac{dy}{dt}\right| - sign(y_\infty - y_0)\frac{dy}{dt} \right) dt \approx \\ &\approx \textstyle\sum_i (|y_{i+1} - y_i|) - |y_\infty - y_0| \end{aligned} \quad (26)$$

An ideal input disturbance step response of first order plants (as e.g. in Fig. 2) has always the shape of a one-pulse (1P) curve (Huba, 2019a,b) consisting of two monotonic intervals separated by an extreme point $y_m \notin (y_0, y_\infty)$ (the monotonicity evaluation according to (26) has to be applied twice) with the deviation evaluated according to

$$TV_1(y) = \sum_i |y_{i+1} - y_i| - |2y_m - y_\infty - y_0| \quad (27)$$

As shown e.g. in Huba (2019a), in case of the single integrator also the input responses corresponding to the setpoint and input disturbance steps have to ideally consist of two monotonic intervals forming a one-pulse (1P) shape. Similarly as above, deviations of the plant input $u(t)$ from an ideal 1P step response should be constrained in terms of $TV_1(u)$ measures. Basically, for control of the FOTD plants it might be meaningful to consider also input with lower or higher number of control pulses, but such situations represent already marginal options.

3.2 Optimization problem

Let us start with summarizing basic facts (Huba, 2019a):

(1) Traditional optimization based on quadratic cost functions (LQ control design) does not distinguish useful and excessive signal increments which significantly limits effectiveness of its application.
(2) Separation of the excessive and useful increments (at the input and output) enables to focus fully on an effective minimization of the superfluous changes.
(3) In the new setup of the optimal control design (Segovia et al., 2014), one has to deal with a trade-off between control error attenuation (IAE), measurement noise injection resulting into "excessive control effort" ("controller activity/input usage" (Grimholt and Skogestad, 2018), or the "output wobbling") and "robustness".

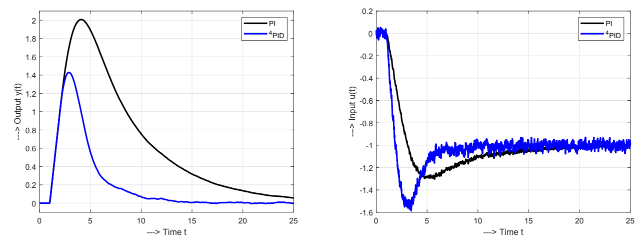

Fig. 2. Transients of the loops with IPDT plant and filtered SIMC PI (8) and cascade PDA-^{4}PID control for a measurement noise $|\delta| \leq 0.25$ generated in Matlab/Simulink with a Uniform Random Number block corresponding to parameters (30): $K_s = 1$; $T_m = 1$; $T_s = 0.001$; $t_{sim} = 30$

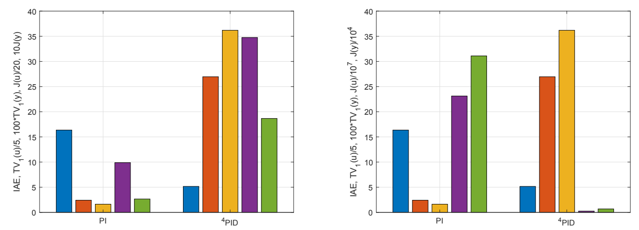

Fig. 3. IAE (blue), $TV_1(u)/5$, $100TV_1(y)$, $J(u)/20$ and $10J(y)$ (green), $k = 1$ (left) and the same measures with $J(u)/10^7$ and $J(y)/10^4$, $k = 6$ (right) of the loops with IPDT plant from Fig. 2

(4) Optimal controller and filter tuning is expected to depend on the noise parameters. Thus, without considering filtration properties, a "generally" optimal PID tuning becomes questionable.

For the loop optimization, different cost functions and optimization constraints may be defined. A "holistic" optimization considering both the plant input and output may be looking for a minimal value of the cost function

$$J(u) = IAE^k \, TV_1(u) \tag{28}$$

By the parameter k it is possible to weight contributions of IAE (speed of control) into the resulting product. To minimize the output wobbling, for the disturbance step responses the cost function may be defined as

$$J(y) = IAE^k \, TV_1(y) \tag{29}$$

3.3 Example 1: Simulation of IPDT system

In this example, 3rd order binomial filter (23) has been added to both SIMC PI and to the inner loop of cascade PDA-^{4}PID controller based on IPDT models. Then, the dead time of the identified plant model T_m has been modified according to (24). Thereby, for $T_m = 1; K_s = 1$ the controller parameters have been specified as:

$$\begin{aligned} &{}^1\text{PI:} \quad T_{c1} = 1;\ T_f = 0.01 \\ &{}^4\text{PID: } T_{c4} = 0.4;\ T_f = 0.1 \end{aligned} \tag{30}$$

The required closed loop time constant T_{c1} has been chosen according to (9). Since PI controller does not include "aggressive" derivative terms, T_f may be chosen relatively short. PDA-^{4}PID control usually allows faster transients, which is reflected by shorter T_{c4}. However, due to expectation of an increased noise impact, T_f has been intuitively increased. Transients in a loop with an external noise with an amplitude $|\delta| \leq 0.25$ generated in Matlab/Simulink by a Uniform Random Number block are in Fig. 2. The corresponding performance measures and combined cost functions (28) and (29) (with $k = 1$ and $k = 6$) are in Fig. 3. At the output, both controllers yield nearly ideal 1P responses. Whereas the IAE value of the cascade PDA-^{4}PID control is (despite the differences in filtration) still more than 3x lower than for SIMC PI control, $TV_1(u)$ and $TV_1(y)$ are more than 10x higher. For $k = 1$, the combined cost functions (28)-(29) indicate advantages of simple PI control (Fig. 3 left). However, when penalizing slower responses by $k = 6$ (Fig. 3 right), PI control is essentially worse than the much faster cascade control. Obviously, evaluation of the chosen controller + filter tuning and its optimization are far from being trivial and require to develop a systemic approach. The up to now carried out considerations have shown that we can design controllers by taking into account the effects of measurement noise. The considered PDA-^{4}PID controller represents only a fraction of many possible solutions. Others could be obtained by considering P or PD stabilizing regulators, replacing transport delay by Padés approximations of different degrees, etc. In other words, in the yet relatively little explored gap between traditional PID controllers and fractional order controllers there exist a wide range of other solutions with an acceptable degree of complexity. These can be based on the most widespread 3 and 2 parameter models of the FOTD or IPDT systems, but also on the widely used 4 parameter models with multiple time constants and dead time. In order to fully grasp the issue, however, there is no greater emphasis on evaluating the optimum solutions obtained. In each case, it is not enough to consider only the speed of the transients evaluated eg. using IAE. Substantially rarely used deviations from the ideal shapes of input and output based on modifications of TV introduced in Skogestad (2003) may also not be considered separately. Ultimately, however, it would be desirable to use combined purpose-based cost functions that holistically take into account several performance indicators at the same time. It should also be noted that:

- Obtaining a controlled system model is always associated with a trial-and-error approach to achieve the highest match between theoretical expectations and experimental results underpinned by identification.
- Its evaluation is usually based on a multicriterial cost function, whose multiplicity is virtually unlimited and excludes substitution by a single general purpose function and a single all-encompassing optimization.
- Given requirements are much easier to cope with by simple analytical approaches, especially when they are built using a modular approach.
- Of course, standard solutions for the most common situations can also be obtained. Well, that does not mean that they will cover at a high level also less numerous but still important situations, in which other presented approaches may yield better results.
- From this perspective, openness and flexibility of adaptation can be considered as the main advantages of the SIMC method. But we pay for it by navigating within a wide range of existing options and finding the optimal alternative.

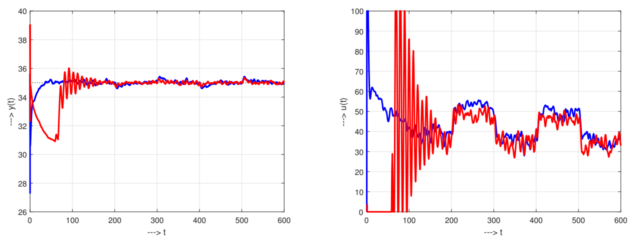

Fig. 4. TOM1A thermal channel: left-plant output, right-plant input step responses; ^{1}PI-blue, PDA-^{4}PID-red

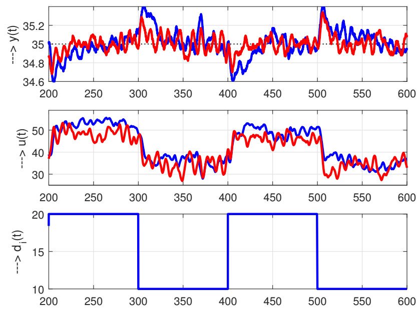

Fig. 5. Details of transients at the input and output of the TOM1A thermal channel and the fan input as a disturbance; ^{1}PI-blue, PDA-^{4}PID-red

4. EXAMPLE 2: THERMAL PLANT CONTROL

Possible balance of all three aspects of the controller design: speed of the transients, noise impact and robustness will now be illustrated experimentally. We will deal with the thermal channel of the laboratory Arduino-based thermo-opto-mechanical system TOM1A used already for testing of method for filtered PI and PID controller tuning based on equivalence of delays, or for testing different approaches to dead time compensator design (Huba et al., 2017). Although the design could be based on standard SIMC design for stable plants, the experiments will be carried out by a simplified approach based on approximating just initial part of the step responses by IPDT models, which essentially shortens time required for identification. Simultaneously, such an approach demonstrates robustness of the applied design: The model parameters taken from Huba et al. (2017) ($K_s = 0.011, T_m = 0.3$s) [1] will be increased by an additional communication delay $T_{mc} = 1$s helping to achieve for $T_s = 0.02$s [2] and (24) higher ratio T_d/T_s and thus allow a simplified design in continuous-time domain with

$$\begin{array}{rl} {}^{1}\text{PI:} & T_f = 0.4,\ n = 2 \\ PDA-{}^{4}\,\text{PID: } T_{c4} = 0.5; & T_f = 0.4,\ n = 4 \end{array} \qquad (31)$$

Higher aggressiveness of PDA-^{4}PID control will be considered by an increased filter order. Thanks to $T_d/T_s \in [85, 105]$, we can expect that the design in a continuous time domain will give sufficiently concise results. Overall transients in Fig. 4 show that (after controller installation), both solutions yield nearly equivalent solutions. However, for PDA-^{4}PID control the initialization period requires much longer time. Finer details may be illustrated

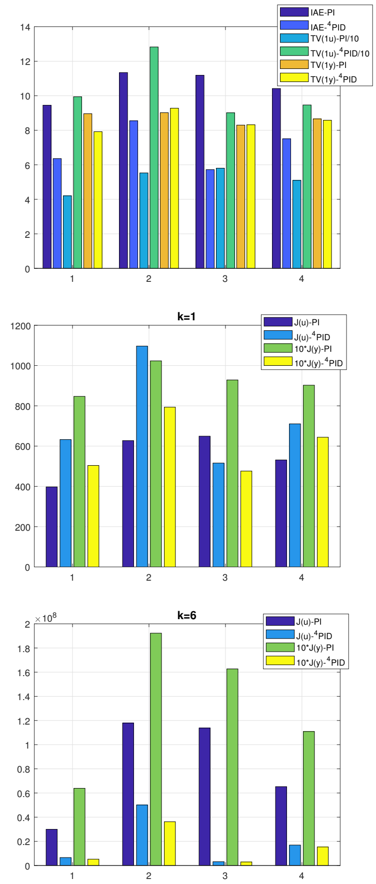

Fig. 6. Performance measures (above) and combined cost functions for $k = 1$ (middle) and $k = 6$ (below) evaluated over four disturbance intervals

by Fig. 5 showing correlation of the transients with the fan producing control disturbances. They demonstrate that when properly combining filtration, which yields smoother signals, with prediction (derivative action) allowing faster dynamics, we may get faster transients without impairing their shapes by noise impact. This documents that once we accept possibility to upgrade PI to PID, we should continue with upgrading PI to control with higher order derivatives. In time of the software-implemented controllers, the argument that PI control is preferably used due to its simplicity (Grimholt and Skogestad, 2018) might seem to be questionable. However, higher order controllers also bring complexity related to high number of initial conditions which may require special start-up procedures. Furthermore, by enabling possibility of tighter control working with higher gains, special attention has to be paid to control constraints (Huba, 2019a). Since the transients are influenced also by slow modes of heat transfer with the time constants longer than experiment duration, four disturbance periods have been considered to identified possible changes in performance. Evaluation of performance measures and combined cost functions (28) and (29) in Fig. 6 illustrates lower IAE and higher shape related deviations of PDA-^{4}PID control at the plant input. However, with higher accent on speed of the transients, for $k = 6$ this control is definitely better in all combined cost functions. In the carried out example, the controller tunings have not been chosen to show limits of the considered

[1] the identified plant time constant $T_{1m} \approx 20$s will not be used
[2] limited by Arduino-Simulink communication speed

approaches, but to illustrate their features. Impact of a chosen tuning parameter on the trade-off between speed of control and the shape related deviations at the input and output may be systematically analyzed by using several types of characteristics. Dependence of two basic basic measures of the closed loop performance displaying the shape related deviations at the input, or output from their ideal values (variable ξ) and the control error attenuation (IAE characterizing speed of the transients, variable η)

$$\xi = TV_1(u), \eta = IAE, \text{ or } \xi = TV_1(y), \eta = IAE \quad (32)$$

define the *speed-effort* (SE) and *speed-wobbling* (SW) characteristics (Huba, 2019b).

5. CONCLUSIONS

SIMC controller design represents a powerful approach which may further be extended by considering control including HO derivative actions and noise filters. Thereby, the original, rather "speculative" design for integral plants may be replaced by a more rigorous and at the same time powerful cascade control, when the integral (or unstable) system will firstly be stabilized by P, PD, or PDA controllers. Then, by using "half rule" and its modifications, the original approach yields excellent results, which enable to use it also in high demanding applications. This paper has shown that by taking into account the filtration aspects, huge space available for performance improvements may be created. In this newly established framework, optimal PID design depends significantly on the measurement noise level and on the required filtration degree. Due to its analytical core, the introduced method remains flexible in a huge range of possible situations. For further research, we see several lines suitable, as numerical verification of the analytically proposed equivalence of the time delays by modifications of the"half-rule" and other types of delay equivalence (Huba et al., 2018a), use of alternative performance measures and tuning strategies (for example, to keep a constant standard deviation), or extension to higher order plant models. It is also to remember that any good technology we use, the success of implementation greatly depends on the tuning of the controller and filter. Thereby, the non-ideal state of the art in the PID control fully confirms this comments (Åström and Hägglund, 2006). Since the staff dealing with control application has to focus on numerous other tasks, the only realistic way to improve this situation seems to be use of intelligent technologies combining the presented approaches with the plant identification and evaluation tools. As a step towards such a computer support, we have started to develop tools for such a design (Bistak, 2018; Bisták and Huba, 2019).

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