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Another Step Towards the Renaissance of Automatic Reset Based Control

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Abstract: This work extends list of recently published constrained automatic reset based controllers by optimizing one particular "problematic" solutions with the 4th-order output derivative. Although the proposal itself is derived for integral plus dead-time (IPDT) process models, thanks to the universality of the used ultra-local model, the field of applicability can be very wide and include a number of stable and unstable, linear and non-linear processes. With regard to the progress achieved in the design of generalized automatic reset controllers, the diversity of their various modifications and also with respect to the genius of original automatic reset and hyper reset controllers and their inventors, the contribution further discusses related terminological aspects, especially the misleading designation of these solutions by the term series PI and PID controllers, which was the source of many misunderstandings, problems and contributed to the significant lag in the further development of this type of control. Similar comments it also adds regarding the use of the term anti-reset-windup controller.

Keywords: Automatic reset, higher-order derivatives, disturbance observer, multiple real dominant pole, constrained control.

1. INTRODUCTION

When pneumatic controllers with disturbance compensation marked as automatic reset appeared about a century ago, it was far from clear in advance that they were going to dominate the field of automatic control. They were just one of many solutions, most of which have since disappeared from industrial practice (Bennet, 1996). Even before the World War II, their generalization emphasized in the name as hyper reset was established. In the following decades, there were several waves of innovation of these two successful controllers brought by the introduction of vacuum tubes, transistors, and operational amplifiers. Although the basic scheme of original solutions was preserved in principle and still represents the basis of industrial automation, their designation was changed. Today, however, only a few people know them under the original designation, when a new name emerged from the work of Minorsky (1922) as an abbreviation of adjectives proportional-integral-derivative (PID) Bennet (1996). For a long decades, it seemed that the three corresponding components related to the present, past and future of control error best reconciled the requirements of the automatic control. However, from the point of view of automatic reset and hyper reset controllers, such an interpretation diverted attention from the further development of their essence. They are equivalent to PIDs with an explicit integrator only around steady-states of the system. At the time of dominance of the linear theory of automatic control, their significantly different properties during transients with a limited value of the control action were simply neglected. Instead of automatic reset and hyper reset, PI and PID controllers started to be used and another adjective, series (serial) PIDs, was sometimes used to distinguish the reset based solutions. But many users forgot about it, which caused problems with the application of older methods for setting controllers, such as proposed in Ziegler and Nichols (1942). Another, even more important consequence of the new interpretation and terminology was the emergence of the windup problem, which was able to completely invalidate the activity of controllers with explicit integrator in the case of control signal restrictions (Glattfelder and Schaufelberger, 2003). Although automatic reset controllers do not have an explicit integrator, and therefore no redundant integration, the inertia of human thinking created a completely meaningless abbreviation ARW (from anti reset windup), which can be encountered in many publications (Glattfelder and Schaufelberger, 1983, 1987; Huba et al., 1995; Rundqwist, 1990). In this context, one of the goals of this contribution is to show the meaninglessness of this concept, when in a correctly set circuit with automatic reset, the limited control signal does not have any negative effect on the dynamics of the circuit.

Another important moment is that generalizations of PIDs with higher-order derivatives are increasingly used to control more complex processes with difficult-to-control dynamics (see e.g. Jung and Dorf (1996); Ukakimaparn et al. (2009); Sahib (2015); Oladipo et al. (2021); Zandavi et al. (2022); Boskovic et al. (2020); Veinović et al. (2023); Kumar and Hote (2021); Ferrari and Visioli (2022); Visioli and Sánchez-Moreno (2022)). However, with the increase in the number of used controller components, traditional explanations based on the presence, past and future of transients are simply not enough, and even anti-windup schemes derived from PIDs can be overcome by solutions returning to structures with automatic reset.

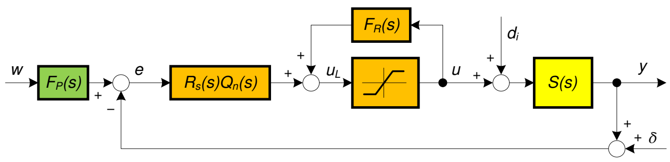

Fig. 1. 2DoF Series controller for a plant $S(s)$ (1) with a filter $Q_n(s)$ (4), the automatic reset $F_R(s)$ (9) established as a positive feedback from a constrained controller output u, a stabilizing controller $R_s(s)$ (2), and a prefilter $F_p(s)$ (16); w=setpoint, d_i=input disturbance, δ=measurement noise.

What is surprising, however, is that none of these new contributions (with exception of Huba et al. (2023b,d)) is based on the generalization of widely proven automatic reset based schemes. In addition to contributions that directly address the design of controllers with higher order derivatives, there is also a huge number of works where the use of complex controllers is hidden behind the exotic designation of fractional order PIDs. Such works were, for example, the focus of the previous conference IFAC PID'18 in Ghent (Tepljakov et al., 2018). Higher-order controllers are created here by Oustaloup approximation of fractional order operators. Increase in robustness of the circuit by using higher-order controllers was, however, shown also by another contribution from the PID'18 conference (Huba and Vrančič, 2018). It should be remembered here that the progress in the design of higher-order controllers was achieved thanks to the simplified design of filters. These are not solved for each derivative separately, which would be unbearable when using higher derivatives. Instead, one binomial filter is proposed for all control components (Huba, 2015; Huba et al., 2021b, 2023d). However, a quasi-continuous-time design must be used, because discrete-time higher-order filters calculated on the basis of H_0 equivalence may contain unstable numerator zeros (Åström et al., 1984). Other important features of the implemented development included the evaluation of the obtained step responses using monotonicity based performance measures and the optimization of controllers using the multiple real dominant pole (MRDP) method.

On the other hand, the declining scope of teaching focused on the basics of automatic control (Rossiter et al., 2020), lower readiness of students in mathematics and physics, lower perseverance and resilience of students, growing range of theoretical approaches popular in practice focused on optimality and robustness: PID controllers belong to them for a long time, see e.g. Skogestad (2003); Víteček and Vítečková (2019)), but also new approaches as active disturbance rejection control (ADRC), trying to combine needs on simplification with improved performance, are increasingly used (Nie et al., 2021; Wu et al., 2021)). All this leads to the fact that, despite the growing possibilities of application of theory in practice created by the mass expansion of various applications of automatic control, their quality can hardly be characterized globally as growing. And once we accept the legitimacy of such a statement, another question arises as to whether and what can be done with such a situation. Our proposal, which has not yet been tested in practice, is that we need to make available to users solutions that would enable them to create better designs for regulators than has been the case with conventional methods. Of course, even here we can doubt whether we have such better alternative solutions available, and especially whether we can persuade users to work with them.

Progress in the design of controllers with higher-order derivatives was demonstrated for parallel PIDs in Huba et al. (2021b), which, for reasons of space limitations, presented solutions up to derivative degree $m = 5$. Their practical application required additional anti-windup modifications, the complexity of which increased with increasing m, but they still did not ensure the achievement of ideal shapes of transients. This brought attention to the careful examination of the so-called series PIDs, i.e. automatic reset and hyper reset controllers and their generalizations (Huba et al., 2021a). The further achieved results were published in Huba et al. (2023d), which, with regard to space limitations, dealt again with solutions up to $m = 5$. They showed that with regard to removing the deformation of setpoint responses arising due to control signal saturation, MRDP controllers should be readjusted optimally so that the smallest time constant of the controller numerator is chosen as the automatic reset time constant. Huba et al. (2023b,a) dealt with solving the problem for even value of $m = 2$, when the controller numerator has complex conjugate zeros. Now, these results will be generalized for the case with $m = 4$.

The rest of the paper will be structured as follows. Section 2 repeats basic steps of the MRDP-optimal two degree of freedom (2DoF) controller design for the IPDT system considering the output derivatives up to $m = 4$. It shows how to modify the MRDP controller with the aim of achieving nearly ideal step responses in the control loop with saturated controller output. Section 3 continues the discussion about the necessity of a separate treatment of higher-order generalizations of PID and automatic reset controllers and development of a concise terminology.

2. OPTIMAL CONTROLLER FOR IPDT MODELS

For a process with the output $y(t)$ and the input $u(t)$, a two-parameters integrator plus dead-time (IPDT) model

$$S(s) = \frac{Y(s)}{U(s)} = S_0(s)e^{-T_{dp}s};\ S_0(s) = \frac{K_{sp}}{s} \tag{1}$$

will be considered, specified with a gain K_{sp} and a dead-time T_{dp}. In a "nominal" case, the simplified symbols K_s and T_d will be used. The "ideal" stabilizing controller using output derivatives up to the 4th order is

$$\begin{aligned} R_s(s) &= K_c + K_{d1}s + K_{d2}s^2 + K_{d3}^3 s + K_{d4}s^4 = \\ &= K_c(1 + T_1 s + T_2 s^2 + T_3 s^3 + T_4 s^4) \end{aligned} \tag{2}$$

It operates on the control error

$$e = w - y \tag{3}$$

defined as the difference among the setpoint w and the output y. To get access to the particular output derivatives, $R_s(s)$ has to be combined with a low-pass filter

$$Q_n(s) = 1/(T_f s + 1)^n = 1/P_n(s) \tag{4}$$

Its delay can already be included in the process model (1) (Huba et al., 2021a; Bisták et al., 2023), or be approximated by a delay equivalence (Huba and Vrančić, 2021; Huba et al., 2023b), adding an "equivalent" filter delay T_e to the process delay T_{dp}, when the "total" dead-time is

$$T_d = T_{dp} + T_e \tag{5}$$

T_e can be simply determined (Huba et al., 2023d) from the parameters n and T_f of the filter (4) according to

$$T_e = nT_f \quad (6)$$

Output u_s of $R_s(s)$ can be modified by an offset u_{off}

$$u_L = u_s + u_{off} \quad (7)$$

used for compensation of acting disturbances and limited by a saturation nonlinearity block (Fig. 1) defined as

$$u(t) = sat\{u_L\} = \begin{cases} U_{max}\,; & u > U_{max} \\ u_L\,; & U_{min} \le u_L \le U_{max}, \\ U_{min}\,; & u < U_{min} \end{cases} \quad (8)$$

By a positive feedback loop with a low-pass filter $F_R(s)$

$$F_R(s) = 1//(1 + T_i s) \quad (9)$$

with a time constant T_i, used for calculating the offset signal u_{off} from u Huba et al. (2023d), it is constructed the simplest known disturbance observer (DOB) (Huba and Gao, 2022). In steady states, the controller output acting on integral models has to be equal to the negative value of input disturbances d_i. Thus, this DOB can be interpreted as being based on steady states of integral models. In the zone of proportional control defined by the limit values U_{min} and U_{max}

$$u \in [U_{min}, U_{max}], \quad (10)$$

the positive feedback called originally as automatic reset establishes an integral component of control, when

$$\begin{aligned} R(s) &= \frac{U(s)}{E(s)} = \frac{R_s(s)}{1 - F_R(s)} = \\ &= \frac{(K_c + K_{d1}s + K_{d2}s^2 + K_{d3}^3 s + K_{d4}s^4)(1 + T_i s)}{T_i s} \end{aligned} \quad (11)$$

For a nominal plant (1) with parameters K_s and T_d, the controller (11) yields the closed loop transfer function

$$\begin{aligned} F_c(s)) &= \frac{Y(s)}{W(s)} = \frac{R(s)S(s)}{1 + R(s)S(s)} = \frac{B(s)}{A(s)} \\ B(s) &= K_s(K_c + K_{d1}s + K_{d2}s^2 + K_{d3}^3 s + K_{d4}s^4)(1 + T_i s) \\ A(s) &= s^2 T_i e^{T_d s} + B(s) \end{aligned} \quad (12)$$

Remember that the positive feedback yields (11) with an integral term just in the proportional zone of control, when the control does not exceeds the (always existing) limits. Outside the proportional zone $[U_{min}, U_{max}]$ the controller with automatic reset behaves completely differently than the controllers with explicit parallel integral (I) term. To prevent redundant integration (controller windup), which occurs after exceeding the output limitations of parallel controllers, various anti-windup measures must be used (Kothare et al., 1994). In the case of the automatic reset based controller, the correct operation under constraints can be simply guaranteed by an appropriate tuning of the controller parameters (Huba et al., 2023b, 2021b).

2.1 MRDP-optimal controller design

Application of the multiple real dominant pole (MRDP) method avoids existence of slow modes which would prolong the transient responses. Its calculation is based on simple analytical conditions: The dominant pole multiplicity follows from the number of unknown controller parameters (six) that has to be increased by the position of the pole itself. Thus, for the controller design, there are a total of 7 unknowns. To achieve the given dominant pole, it is necessary to ensure that for the characteristic quasi-polynomial $A(s)$ (12) the equations are fulfilled

$$\left[\frac{d^i}{ds^i}A(s)\right]_{s=s_o} = \mathbf{0}; \quad i = 0, 1, ..., 6 \quad (13)$$

From $d^6 A(s)/ds^6$ (Huba et al., 2023d) one gets the dominant pole s_o, or the equivalent time constant T_o as

$$\begin{aligned} s_o &= -(m + 2 - \sqrt{m+2})/T_d = -3.551/T_d; \\ T_o &= -1/s_o = 0.282T_d \end{aligned} \quad (14)$$

Solution of (13) then yields (Huba et al., 2023d)

$$\begin{aligned} &K_c = \frac{\kappa}{K_s T_d};\ T_i = \tau_i T_d; \\ &T_1 = \tau_1 T_d;\ T_2 = \tau_2 T_d^2;\ T_3 = \tau_3 T_d^3;\ T_4 = \tau_4 T_d^4; \\ &\kappa = 1.2702;\ \tau_i = 1.8283;\ \tau_1 = 0.55149; \\ &\tau_2 = 0.1227;\ \tau_3 = 0.01297;\ \tau_4 = 0.000546 \end{aligned} \quad (15)$$

To remove high overshooting after setpoint step responses appearing under the one-degree-of-freedom (1DoF) control, a prefilter can be proposed to cancel the zeros of $F_c(s)$ (12) given by the polynomial $B(s)/(K_c K_s)$

$$\begin{aligned} F_p(s) &= \frac{N_p(s)}{(1 + T_i s)(1 + T_1 s + T_2 s^2 + T_3 s^3 + T_4 s^4)} \\ N_p(s) &= b_5 s^5 + b_4 s^4 + b_3 s^3 + b_2 s^2 + b_1 s + 1 \end{aligned} \quad (16)$$

The simplest and the most robust prefilter tuning is

$$b_i = 0;\ i = 0, 1, ..., 5 \quad (17)$$

According to Víteček and Vítečková (2019), faster setpoint step responses correspond to $N_p(s)$ canceling one dominant closed loop time constant T_o (14)

$$b_5 = 0;\ b_4 = 0; b_3 = 0;\ b_2 = 0;\ b_1 = T_o \quad (18)$$

Real contribution of such a design should always be verified, as the cancellation of one, or even several poles can reduce the robustness of the control. Without control signal limitations, the MRDP-PIDA controller gives ideal shapes with a minimum number of input and output monotonic sections after both setpoint and input disturbance steps. The problem is that the MRDP-optimal parameters lead under constrained control to output and input overshooting both in the setpoint and disturbance responses (see Fig. 2). Because such overshoots also occur when using higher-order PID controllers in anti-windup setup based on the conditioning technique (Huba et al., 2021b), the first question therefore was whether it is possible to avoid them at all with some modified controller.

2.2 Constrained controller tuning

With $p = T_d s$ the MRDP controller numerator can be factorized as follows

$$\begin{aligned} N_R(p) &= 1 + \tau_1 p + \tau_2 p^2 + \tau_3 p^3 + \tau_4 p^4 = \\ &= \tau_4(p^2 + 13.3641p + 46.0994)(p^2 + 10.3920p + 39.7253) \end{aligned} \quad (19)$$

It has the complex conjugate zeros

$$\begin{aligned} p_{1,2} &= -5.1960 \pm j3.5675, \\ p_{3,4} &= -6.6821 \pm j1.2039 \end{aligned} \quad (20)$$

By choosing the new feedback time constant as

$$\tau = 1/5.1960 = 0.1925 \quad (21)$$

or

$$\tau = 1/6.6821 = 0.1497 \quad (22)$$

calculated by neglecting the imaginary part of $p_{1,2}$, or $p_{3,4}$, one could ensure a faster settling down of the automatic reset than with the calculated MRDP value $\tau_i = T_i/T_d = 0.55149$. As in Huba et al. (2023c,a), an "optimal" value of $\overline{\tau}_i$ could also be derived by replacing the complex zeros $p_{1,2}$, or $p_{3,4}$ with their modulus. In order to verify the properties of such modifications of the MRDP controller for a wider range of the optional parameter $\overline{\tau}_i = \tau$, let us consider the controller transformation defined by the modified transfer function expressed for $p = T_d s$ as

$$\begin{aligned}&\kappa\frac{\tau_4(p+1/\tau)^2(p^2+10.3920p+39.7253)(1+\tau_i s)}{\tau_i s} = \\ &= \frac{\kappa\tau_4}{\tau_i\tau}\frac{(\tau p+1)(p^2+10.3920p+39.7253)(1+\tau_i s)(\tau p+1)}{\tau p} = \\ &= \overline{\kappa}\frac{(1+\overline{\tau}_1 p+\overline{\tau}_2 p^2+\overline{\tau}_3 p^3+\overline{\tau}_4 p^4)(1+\tau p)}{\tau p}\end{aligned} \tag{23}$$

From the requirement to preserve the controller transfer function after the replacement of the complex zero by a double real zero (22), the parameter transformations result

$$\begin{aligned}&\overline{\kappa} = 0.01507/\tau;\ \overline{\tau}_i = \tau \\ &\overline{\tau}_1 = 2.08989 + \tau;\ \overline{\tau}_2 = 2.08989\tau + 0.503449;\ ; \\ &\overline{\tau}_3 = 0.046023 + 0.503449\tau;\ \overline{\tau}_4 = 0.0460236\tau.\end{aligned} \tag{24}$$

When changing the parameter $\tau \in [0.1, 0.21]$, which corresponds to the range from approximately 70 to 100% of (22), we get setpoint and disturbance step responses as in Fig. 2 ($n = 5$, red). The evaluation of achieved IAE values and shape related deviations from the ideal shapes of responses with a minimal number of monotonic intervals (introduced e.g. in Huba et al. (2023d,a)) in Fig. 3 confirm the suitability of adjusting the controller transfer function at least around the value (22). The most interesting is the course of deviations of the control signal $TV_1(u_d)$, where the influence of the complex zero with a possible real approximation (21) seems to be visible. In the future, it therefore seems interesting to verify also the modification of the controller based on the approximation of this zero, or to verify the possibility of the simultaneous approximation of both complex zeros. To illustrate the meaning of automatic reset in disturbance reconstruction and compensation, Fig. 2 also shows the reconstructed disturbance responses. Since the disturbance observer (DOB) used is based on steady-state output values of the controller output, in MRDP controller it is manifested by the time constant $\tau_i = 1.8283$ that is longer than the dominant time constants $\tau_o = T_o/T_d = 0.282$ (14) of the equivalent stabilizing controllers. In the modified constrained controller it can still be observed that the reconstructed disturbance settles down just after the output reaches the neighborhood of the required reference value. The τ_i values can't be further decreased, because then the DOB would react significantly to stabilization interventions in the first part of the transients. Therefore, there is no point in looking for more perfect optimization methods that would achieve this. But what makes sense is to shorten the values of τ_o by using controllers with higher order derivatives, which will create space for further reduction of τ_i.

3. DISCUSSION

The realized derivation can be used for several purposes. First of all, it showed MRDP design as an effective tool for designing high-quality linear and constrained controllers, whose performance can be modified in several simple ways. If we add to the above treated constrained controller modification also several possibilities of specifying the closed loop dynamics by differently grouped sets of real poles (Huba and Vrančić, 2022), the use of this optimal design methodology will be further expanded. It is interesting that the MRDP methodology was already used by one of the first textbooks of automatic control Oldenbourg and Sartorius (1944), but in the further development it largely disappeared from the design of automatic controllers.

However, from the point of view of the discussion of interest to the wider control community, the derived constrained modification is extremely important as a new counterexample, which shows the used term "anti reset windup" as an inherently meaningless concoction. Automatic reset can be optimally set even in tasks with saturating controller output. Thus, no further anti-windup adjustments are needed. And that brings us to the main point of discussion, namely the need for separate naming and designing of constrained PID and automatic reset controller families. It is not only a question of respect for the inventors of these still dominantly used industrial solutions, but also of avoiding their mutual confusion, which are committed not only by a large number of students, but also often by the academics themselves. However, the most important thing is the terminological chaos caused by the fact that in the family of PID controllers, which shows an increasing number of members (see e.g. Huba et al. (2021b)), it is senseless to derive the name of the entire family from the name of one specific family member. This creates meaningless formulations of the type "the most frequently used PID controllers are PI controllers". And there is a lack of uniform terminology for controllers with higher order derivatives. We have, for example, PIDA controllers, but also PIDD2, or PIDD^2. Because with higher order controllers it is no longer practical to enumerate all the components (as e.g. PIDD2D3 controller), in Huba et al. (2021b) we proposed the use of the term PID$_n^m$ control, which would take into account not only the maximum order of the used derivative, but also the order of the used filter, which represents one of the most important aspects of the design. With regard to use in abstracts and other texts that do not allow indexes, however, it might be more appropriate to make an agreement on using the designation of PIDmn controller. Of course, a similar problem awaits in the family of generalized automatic reset controllers. In Huba et al. (2023d), we used the abbreviation PD$_n^m$PI to distinguish it from PID$_n^m$. But, by its very nature, it was not a good proposal, because there are no historical or functional specifics for it - in automatic reset there is no explicit integral term. From this point of view, to be as short as possible, it might be more appropriate to use the acronym PD$_n^m$R, or PDmnR, expressing the use of reset. In any case, due to the rapidly growing interest in higher order controllers, it would be helpful to solve these terminology problems as soon as possible.

4. CONCLUSIONS

The article showed how it is possible in a simple way to modify MRDP optimal automatic reset controllers with output derivatives of the fourth order, so that even with saturating controller output, the shapes of closed loop variables still have ideal waveforms. Because by solving

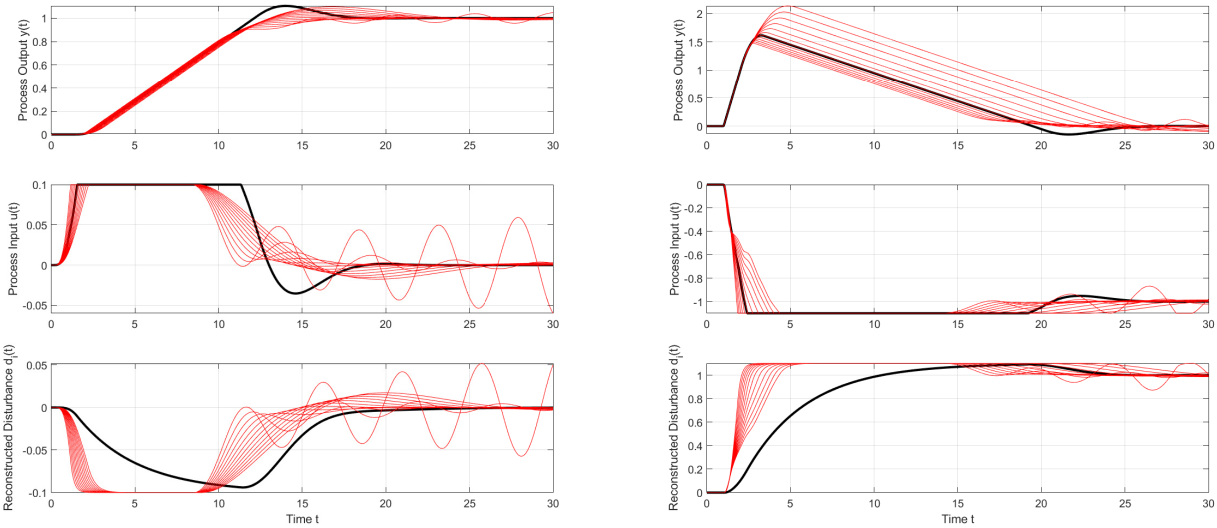

Fig. 2. Setpoint (left) and input disturbance step responses (right) of constrained series MRDP-controller (black) and controllers approximated for $\tau \in [0.1, 0.21]$ according to (24) (red), $\Delta\tau = 0.01$; $U_{max} = 0.1, U_{min} = -0.1$ for the setpoint steps and $U_{max} = 0.1, U_{min} = -1.1$ for the input disturbance steps; $K_s = 1$; $T_{dp} = 1$; $T_e = 1$; $n = 5$; $T_f = T_e/n$; $T_s = 0.001$

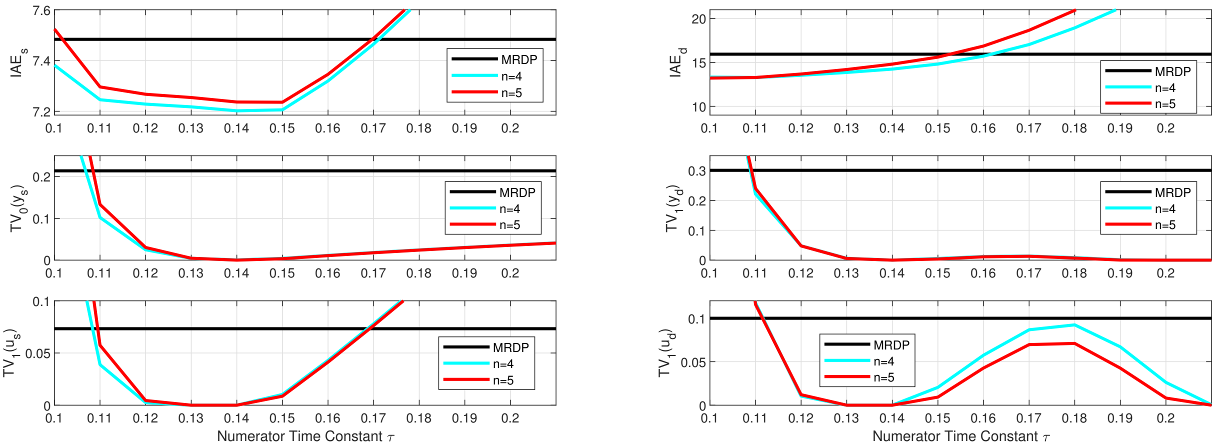

Fig. 3. Impact of numerator time constant τ of the parameter transformation (24) for $n = 4$ and $n = 5$ ($\tau \in [0.1, 0.21], \Delta\tau = 0.01$) on performance of setpoint responses ($U_{max} = 0.1, U_{min} = -0.1$, left) and disturbance responses ($U_{max} = 0.1, U_{min} = -1.1$, right); $K_s = 1$; $T_{dp} = 1$; $T_e = 1$; $T_f = T_e/n$; $T_s = 0.001$

several important problems, it is possible to expect in the near future the use of a large number of members of these two families of controllers with an explicit integral term and with automatic reset, the article also proposes to use the PID'24 conference for a wider discussion regarding the appropriate terminology for the higher-order controllers.

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