scientific reports

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Measurement noise attenuation in modified Smith predictor and automatic offset controllers for integrator plus dead-time system

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The paper compares the recently modified Åström-Smith predictor (ASP) developed for IPDT models with an automatic offset controller (AOC). While an excellent performance can be achieved with ASP in idealized conditions, unacceptable transients with extraordinary high excessive controller effort and a steady-state control error result in the presence of measurement noise. AOC combines a possibly higher-order (HO) stabilizing controller (SC) with compensation of disturbances using a full disturbance observer (DOB). Nominal full DOB includes both the model dead-time and inversion of the integral mode of the model. By increasing the number of output derivatives used in the SC of the AOC, together with increasing the order of the low-pass filter used both in the SC and the DOB, it is possible to significantly increase the speed and robustness of responses in time-delayed processes, together with decreasing the measurement noise impact. The AOC based on the ultralocal IPDT model can be used to replace the higher-order PID in an universal controller for a wide class of processes with dominant first-order dynamics. Significant improvements in measurement noise attenuation can be demonstrated also in application of equivalent low-pass filters to the ASP. But even after such a modification of the noise attenuation, the filtered ASP can exhibit permanent control error or even instability at higher noise amplitudes. Hence, although even the ASP can be used as a universal controller for processes with dominant first-order dynamics, the benefit of its use should always be verified depending on the amplitude of the measurement noise. From this point of view, the use of AOC is simpler and more reliable. Despite the need for appropriate selection of the degree of derivatives used in SC and the tuning of the low-pass filters used. The conclusions of the article are illustrated by simulation experiments of unstable process control and real-time thermal process control.

Keywords Disturbance observer, Automatic offset controller, Higher-order derivatives, Multiple real dominant pole method, Measurement noise attenuation, Filtered Åström-Smith predictor

Smith predictor (SP)[1] represents the first model-based controller that attempted to eliminate output disturbances in time-delayed processes using a parallel process model. At the time of its introduction, the first significant problem in this control concept was related to the analog dead-time implementation used. However, since SP uses a combination of setpoint feedforward with reconstruction and compensation of output disturbance, a compelling question also concerns whether the given concept can be extended to the control of marginally stable and unstable processes. Furthermore, due to the observability and stability problems[2], reconstruction of output disturbances is not feasible in the case of integral and unstable processes. There, it must be replaced by reconstruction and compensation of input disturbances.

In[3] SP was modified for integrator plus dead-time (IPDT) systems. The authors stated that the advantage of the new design is that the setpoint response is decoupled from the load response and hence can be independently optimized. Denoted as Åström-Smith predictor (ASP), this solution was recently modified in[4]. This paper corrected some imperfections in[3] and tried to explain all the pitfalls of the given proposal. It also pointed out the connections with several other modifications of the solution[5–10]. By presenting the stability proof, the paper[4]

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also tried to eliminate the objection[2] that the setpoint response cannot be decoupled from the load response. Hence, the circuit cannot be optimized independently. According to[2], it is usable only after being supplemented with a reference disturbance model and a superior stabilizing controller. However, the problem with the proof presented in[4] is that it only applies to factors included in the description of the system. And it lacked, for example, measurement noise, the influence of which is discussed in more detail in this paper. Simulation and real-time control experiments will show that the ASP is unable to guarantee zero permanent control error. The experimental observations will finally be confirmed by a theorem.

The importance of the discussed issue is still extremely relevant for practice, despite the advanced age of the SP. Control of time delayed systems focused on different modifications of Smith predictors, disturbance observers, and fractional-order controllers still represents an active area of research (see for example[11–20] and the references therein). Applications in areas such as vehicle dynamics control[21] require new, more efficient feedback mechanisms than PID control provides. In exploring the existing context, ASP is compared with the newly proposed model-based controller design based on ultra-local time-delayed models that include the full DOB. Denoted as automatic offset control (AOC), it has several interesting properties[22]. Similarly as in modern control theory (MCT,[23–26]), the design of a nominal stabilizing controller (SC) and a nominal observer of the state and disturbances can be dealt with independently. In PID control, loops with or without integral (I) action needed to be tuned separately[27]. Thus, the addition of the I action increases the noise sensitivity and slows down the transients. The introduction of automatic offset controllers brings about the advantage[22,28] of adjustable disturbance reconstruction without influencing the setpoint responses. In AOC, by increasing the number of output derivatives used in the SC, it is possible to increase the speed and robustness of responses in time-delayed processes. Furthermore, it contributes to enhancing measurement noise attenuation and broadening possibilities of shaping the closed-loop responses. The use of higher-order derivatives appears to be a starting point in the fight against process uncertainties and time-delays that limit the acquisition of updated information[29]. As a result, the AOC derived for the ultralocal IPDT model can be used as a universal controller for a wide class of processes with dominant first-order dynamics. The advantages of this new solution stand out compared to the family of previously published higher-order parallel and series proportional-integral-derivative controllers (HO PID)[30,31] especially with regard to solving the control signal saturation problem, measurement noise attenuation, and the speed of reconstruction and disturbance compensation. Since the proposed controller can be successfully used to control a wide range of processes with dominant first-order dynamics and possibly saturated input, the AOC proposed is especially relevant and welcomed for unstable time-delayed processes, for which the design of a suitable controller is still a current problem.

The simplification of controller design by using ultralocal models is now widely used in several approaches to so-called model-free control. The most well-known are the active disturbance rejection control (ADRC)[32–35], and the flatness-based intelligent PID (iPID) control[36–41]. The present paper shows that the simplifications achieved by using ultralocal models can also be applied to both AOCs and modified ASPs, but it points to significant differences in the behavior achieved.

In its consequence, the AOC proposed allows replacing the existing paradigms of automatic control based on PID control with more advanced solutions that provide improvement of the achieved performance with simplification of the controller design even in the case of systems with difficult-to-handle dynamics. The discovery of these new possibilities, together with the partial solution of the problem of measurement noise attenuation in ASP, can be considered to be the main contributions of the article.

The rest of the paper will be structured as follows. Section 2 summarizes the design of AOC for integrator plus dead-time (IPDT) models. Section 3 introduces the Åström-Smith predictor (ASP) for the IPDT model modified in[4], compares its application with the AOC in ideal and noisy conditions, and proposes expanding the ASP with low-pass filters to increase noise attenuation. Section 4 deals with illustrative examples of application of both considered approaches to control of an unstable first-order time-delayed (UFOTD) process and real time control of a thermal process. Both ASP and AOC are discussed in more detail and compared in Section 4 and summarized in Conclusions.

Higher-order automatic offset controller (AOC)

The integral (I) action of the automatic reset controller (ARC)[31] transfer function appears due to the simplified description valid only in the proportional zone of control[42]. In PID control, such a transfer function represents a starting point to explain its operation. In the end, it led to the emergence of the windup effect. This transfer function was neither an ideal starting point when looking for possibilities of improved performance. Whether it was taking into account nonlinear process properties, time delays, uncertainties, or acting measurement noise. In this context, it shows to be much easier to explain ARC as a disturbance-observer-based control (DOBC) with a simplified DOB. By omitting the inversion of the process model in the DOB and its possible dead-time element, it was possible to decrease the impact of measurement noise and simplify the controller construction. For the first analog (pneumatic) implementations, these aspects could be essential. However, from the point of view of implementation using current programmable devices, the advantages resulting from the use of a full DOB may already prevail. These apply not only to the aforementioned simplicity of the design, but also to the improvement of the achieved closed-loop properties. The AOC structure is derived from the reconstruction and compensation of input disturbances. Although it essentially follows on from the historically long-used ARC, this area has long been overshadowed by the generally popular PID controller methodology.

Process modeling

The "user manual" for ARC implementation published by Ziegler and Nichols[43] introduced the advantages of process approximation by an integrator plus dead-time (IPDT) models

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

With dead-time T_{dp} and gain K_{sp}, this two-parameter-model means that the dead-time of the process T_{dp} could be taken into account in the controller design even when the dead-time model was omitted from the DOB structure. Historically, IPDT models are among the most widely used models for PID control design[44]. However, they are also crucial for the design of "model-free" approaches such as active disturbance rejection control (ADRC)[32–35], or intelligent PID control (iPID)[36,37]. The term "model-free" originated as a not entirely accurate designation of simplified modeling, in which internal process feedback is neglected. From this point of view, the term ultra-local models is more accurate, which expresses their limited accuracy in the vicinity of operating points compared to the generally more widely accurate linear "local" models. About higher precision, one can speak at least with respect to the more detailed approximation of internal process feedback. When solving some problems with the nominal setting of integral time-delayed processes, when it makes sense to assume an identical correspondence between the process and the model, we omit the index *p* in the model parameters.

In general, model-based solutions consider the inversion of the dynamics of the integrator $S_0(s)$ in DOB with the use of (minimally) first-order implementation filters (see Fig. 1). With respect to marginally stable model dynamics, the loop stabilization has to be designed using at least a P controller. The design of the P controller is the simplest, but from the point of view of noise attenuation, it may be associated with increased excessive controller effort.

Historical ARCs used automatic "resetting" of the controller offset[31] to eliminate the permanent deviation caused by the difference in the dynamics of the model and the real process. Since the word "reset" is now frequently associated with other functionality in the field of engineering, the newer designation "automatic offset controller (AOC)" was used when innovating the original ARC scheme[22]. The change of designation is also associated with the innovated functionality, in the design of which the inversion of the (ultra-local) process model and the dead-time model of the given loop are used in combination with HO filters.

Stabilizing controller with two degrees of freedom

The first AOC presented in[22] was designed for DIPDT systems. It used a combination of a stabilizing controller (SC) that included possibly HO derivatives with DOB using HO filters. The full predictive DOB applied instead of its simplification used in ARC makes the nominal SC tuning independent of the possible extension by reconstruction and compensation of disturbances. Furthermore, compared to ARC, it increases noise attenuation. The AOC modification applied to the control of IPDT models in[28], will now be extended to the design of a controller with two degrees of freedom (2DOF).

As shown in[28], for the delay-free integral process $S_0(s)=Y(s)/U(s)=K_s/s$, the proportional (P) controller $U(s)=K_P[W(s)-Y(s)]$ produces the closed-loop transfer function $F_{wy}(s)=K_PK_s/(s+K_PK_s)$. For $K_PK_s>0$, the closed-loop time constant $T=1/(K_PK_s)>0$ guaranties stability and can be chosen practically arbitrarily long. However, in a closed loop with a transport delay T_d (contributing by the term e^{-T_ds}), the P controller without compensation of T_d can lead to oscillations and even instability. The use of a finite number of terms of expansion of $e^{T_ds}=1+T_ds+T_d^2s^2/2!+\ldots$ does not give good results without further modification[20]. However, it can be modified further by the MRDP. Additionally, by moving the compensation to the feedforward path, a controller with two degrees of freedom can be achieved, which can improve setpoint responses. Suppose first an ideal feedback controller $C^m(s)$,

$$C^m(s)=\frac{Y^m(s)}{Y(s)}=1+T_1s+\ldots+T_ms^m=N_c(s) \tag{2}$$

where $T_1>0$ to $T_m>0$ are some appropriately calculated positive coefficients. The calculated output $Y^m(s)=C^m(s)Y(s)$ is then substituted instead of Y(s) in the P controller $U(s)=K_P[W(s)-Y(s)]$ with

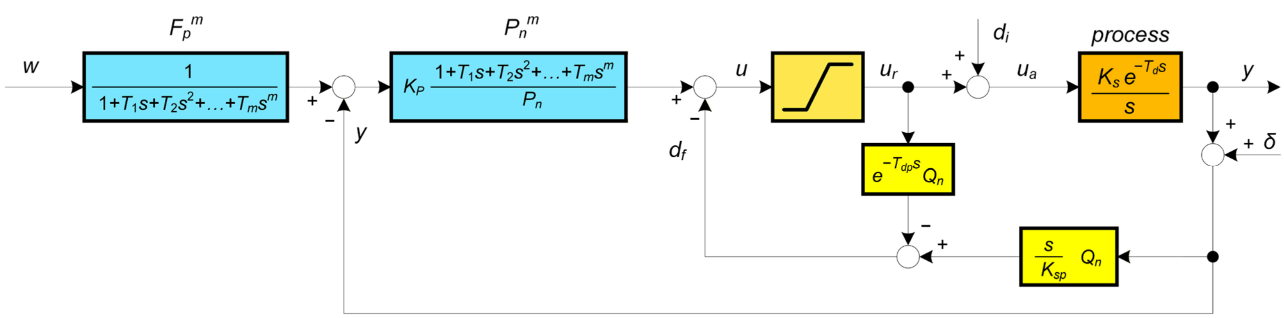

Figure 1. AOC as a 2DOF stabilizing controller $P_n^m=K_PN_c(s)/P_n(s)$, $N_c(s)=1+T_1s+\ldots+T_ms^m$ compensation of the dead-time T_{dp} of the IPDT model (1) and with the prefilter $F_p^m=1/N_c(s)$. The stabilizing loop is extended by the DOB-based disturbance reconstruction and compensation d_f with a dead-time model T_{dp}, gain model K_{sp} and low-pass filter $Q_n(s)=1/P_n(s)$; δ- measurement noise.

the expectation, allowing a higher gain K_P and achieving faster transients than when the loop is closed directly with the delayed output $Y(s)$. Moving $C^m(s)$ to the direct control path, while combining it with the P controller, creates the P^m controller

$$P^m(s) = \frac{U^m(s)}{W_p(s) - Y(s)} = K_P(1 + T_1 s + \ldots + T_m s^m) \tag{3}$$

and a prefilter

$$F_p^m(s) = \frac{W_p(s)}{W(s)} = \frac{1}{1 + T_1 s + \ldots + T_m s^m} \tag{4}$$

The transfer function of the resulting control loop can be expressed as

$$F_w^m(s) = \frac{Y(s)}{W(s)} = \frac{C^m(s)S(s)}{1 + C^m(s)S(s)} = \frac{K_P K_s(1 + T_1 s + \ldots + T_m s^m)}{se^{T_d s} + K_P K_s(1 + T_1 s + \ldots + T_m s^m)} \tag{5}$$

The closed loop has the fastest responses for the quasi-polynomial $A(s)$

$$A(s) = se^{T_d s} + K_P K_s(1 + T_1 s + \ldots + T_m s^m) = (s - s_o)^{m+2} \tag{6}$$

with multiple negative real dominant poles s_o. It is calculated to satisfy conditions[22]

$$\left[A(p); \ \frac{dA\,(p)}{dp}; \ \frac{dA^2\,(p)}{dp^2}; \ \ldots \ \frac{dA^{m+1}\,(p)}{dp^{m+1}}\right]_{p=p_o} = 0 \tag{7}$$

Calculation of the m constants $T_i, i = 1, \ldots, m$ and of the gain $K_P = K_P^m$ begins with the calculation of $p_o = T_d s_o$ from condition (7) as

$$\frac{dA^{m+1}\,(p)}{dp^{m+1}} = T_d^m e^{T_d s}(T_d s + m + 1) = 0 \tag{8}$$

which is yielding

$$p_o = s_o T_d = -(m+1); \ m = 0, 1, 2, \ldots \tag{9}$$

The remaining parameters required for the AOC design calculated in[28] by a triangular system of equations[22] are in Table 1. The table uses “the dimensionless model parameters” introduced with $T_d = T_{dp}$ and $K_s = K_{sp}$ as

$$\kappa_0 = K_s K_P T_d; \ \tau_j = \frac{T_j}{T_d^j}; \ \tau_o = -\frac{1}{p_o} = \frac{1}{m+1}; \ j \in [1, m]. \tag{10}$$

The advantage of the feedback controller $C^m(s)$ used in[28] is the greater simplicity and clear interpretation of its role. The use of the controller $P^m(s)$ with the prefilter $F_p^m(s)$ extended by canceling $r \in Z^0$ closed-loop time constants T_o according to

$$F_{pr}^m(s) = \frac{W_p(s)}{W(s)} = \frac{(1 + T_o s)^r}{1 + T_1 s + \ldots + T_m s^m}; \ r \in [0, m]; T_0 = \tau_o T_d \tag{11}$$

allows us to speed up the setpoint responses.

	$m=0$	$m=1$	$m=2$	$m=3$	$m=4$	$m=5$
$-p_o$	1	2	3	4	5	6
τ_o	1	1/2	1/3	1/4	1/5	1/6
κ_0	0.367879	0.541341	0.672125	0.781467	0.877336	0.963738
τ_1	0	0.25	0.370370	0.445313	0.497920	0.537551
τ_2	0	0	0.037037	0.070313	0.097920	0.120885
τ_3	0	0	0	0.003906	0.008960	0.014146
τ_4	0	0	0	0	0.000320	0.000857
τ_5	0	0	0	0	0	0.000021

Table 1. Optimal normed parameters (10) of C^m and P^m controllers corresponding to IPDT model (1) for $m \in [0, 5]$.

Design of the controller filter

For implementation, $P^m(s)$ has to bee extended by a stable denominator $P_n(s)$ to guarantee closed-loop stability, performance, and causality of the calculation. Specified by a stable polynomial $P_n(s)$ with some $n \geq m$ and a time constant T_f

$$P_n(s) = (T_f s + 1)^n \ ; \ n \geq m. \tag{12}$$

It yields the binomial low-pass filter

$$Q_n(s) = \frac{1}{P_n(s)} = \frac{1}{(T_f s + 1)^n} \ ; \ n \geq m \tag{13}$$

that can be included in the controller setting by adding an equivalent dead time T_e to the process delay T_{dp} according to

$$T_d = T_{dp} + T_e \tag{14}$$

The simplest expression of T_e by the filter time constant T_f can be achieved by

$$T_e = nT_f. \tag{15}$$

The "total closed-loop dead time T_d" is then used in the calculation of the AOC parameters according to Table 1, (10) and (11). The extension of $P^m(s)$ by $Q_n(s) = 1/P_n(s)$ gives a fully realizable controller transfer function

$$P_n^m(s) = P^m(s) Q_n(s); \ m \leq n. \tag{16}$$

allowing us to modify the measurement noise attenuation. As was already mentioned above, nominally, the DOB using the same low-pass filter $P_n(s)$ does not influence the stabilizing loop setting. The filtered input disturbance d_f can be calculated by evaluating the input to the process $U_a(s) = sY(s)/K_{sp}$. Since it is estimated by inversion of the delay-free process dynamics, it has to be compared with the estimate of the delayed controller output $e^{-T_{dp}s}U(s)$. The estimation can be made causal by adding a low-pass filter, which yields

$$d_f = \frac{1}{(1 + T_f s)^n} \left(\frac{sY(s)}{K_{sp}} - e^{-T_{dp}s} U(s) \right); \ n \geq 1; \ T_f > 0. \tag{17}$$

When we finally include the DOB (17), together with the low-pass filter of the stabilizing controller and the DOB (13) and the prefilter (11) into the input-output transfer functions of the closed circuit, we get

$$\begin{aligned} F_{wn}^{mr}(s) &= \frac{Y(s)}{W(s)} = \frac{K_P K_s (1 + T_o s)^r}{s e^{T_{dp}s}(1 + T_f s)^n + K_P K_s (1 + T_1 s + \ldots + T_m s^m)} \\ F_{in}^{m}(s) &= \frac{Y(s)}{D_i(s)} = \frac{K_s \left(e^{-T_{dp}s} - (1 + T_f s)^n \right)}{s e^{T_{dp}s}(1 + T_f s)^n + K_P K_s (1 + T_1 s + \ldots + T_m s^m)}. \end{aligned} \tag{18}$$

These give that, while guaranteeing stability and $t \to \infty$, when $s \to 0$, we achieve with $F_{wn}^{mr}(0) = 1$ a faithful tracking of the constant setpoint and with $F_{in}^{m}(0) = 0$ a full rejection of constant disturbances.

In the simulation, the AOC scheme according to Fig. 1 was implemented in Matlab/Simulink software according to Fig. 2. In the next section, AOC will be compared with ASP, first using the example of controlling the IPDT process (1) with parameters

$$\begin{aligned} &K_{sp} = K_s = 1; \ T_{dp} = L = 0.5; \ U_{max} = 1.15; U_{min} = -0.15; \\ &m \in [0, 5]; \ n = m + 2; \ T_e = T_{dp}; \ T_f = T_e/n; \ r = 0. \end{aligned} \tag{19}$$

However, before doing so, it is necessary to state the performance measures used and clarify the constraints on the choice of the equivalent filter delay value T_e.

Performance measures used

The speed of the setpoint and disturbance responses will be evaluated in terms of the integral of the absolute error

$$IAE = \int_0^\infty |e(t)| \, dt \ ; \ \ e = w - y \ , \tag{20}$$

and denoted as IAE_s, or IAE_d. Hence, *w* denotes the reference setpoint, *y* the output of the process, and *e* the control error. Pure optimization based on IAE leads to transient responses with moderate overshooting. To eliminate this overshooting, it is appropriate to combine IAE-based optimization with monotonicity-based constraints. These can be based on deviations from the monotonicity calculated as the total variation (the total sum of absolute increments[45]) exceeding the net variation of the signal considered. In the case of the output with net change $|y_\infty - y_0|$ calculated from the initial and final values y_0 and y_∞, it will be calculated as

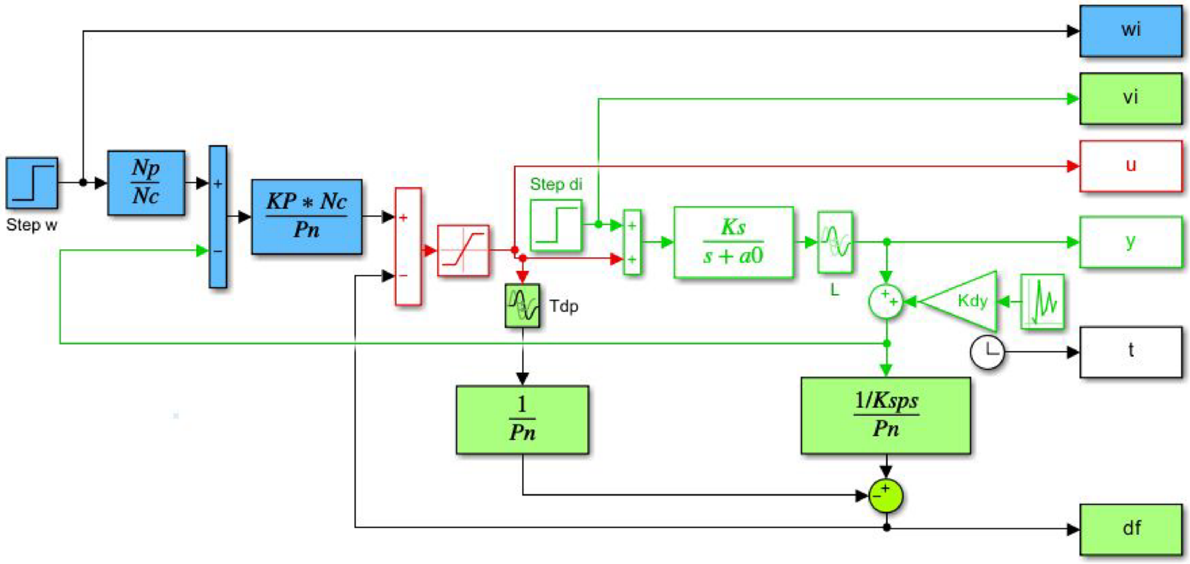

Figure 2. AOC: Simulation scheme in Matlab/Simulink generalized for the first-order time delayed process $K_s e^{-Ls}/(s+a_0)$ with measurement noise generated by Uniform Random Number block with amplitudes ± 0.01, $K_{dy} = 1$.

$$TV_0(y_s) = \int_0^\infty \left(\left| \frac{dy}{dt} \right| - sign(y_\infty - y_0) \frac{dy}{dt} \right) dt \approx \\ \approx \sum_i \left(|y_{i+1} - y_i| \right) - |y_\infty - y_0| \tag{21}$$

Thus, excessive increments characterize the "smoothness" of the output step response, with $TV_0(y_s) = 0$ corresponding to an ideal smooth monotonic output. Otherwise, $TV_0(y_s) > 0$.

The input (controller output) of the integral process model consists of two monotonic intervals, separated by an extreme point u_m that is outside the interval specified by the initial value u_0 and the final u_∞. Therefore, it can be denoted as the one-pulse (1P) signal. From evaluation of two such monotonic intervals it is possible to calculate the deviation from 1P input according to

$$TV_1(u) = \sum_i \left(|u_{i+1} - u_i| \right) - |2u_m - u_\infty - u_0| \,. \tag{22}$$

The trade-off between transient speed (expressed in terms of IAE_s, or IAE_d) and excessive controller activity (input usage $TV_1(u_s)$, or $TV_1(u_d)$) can be calculated using "a holistic cost function"

$$J_k = IAE^k * TV_1(u). \tag{23}$$

The parameter *k* can be used as a weighting factor to stress the role of *IAE* with respect to the excessive controller effort $TV_1(u)$ in the performance evaluation.

Minimal feasible values of the equivalent time delay

Reducing the filter time constant T_f leads to an acceleration of transient responses, but also to an increase in the effects of neglected aspects of emulation of continuous-time circuits by discrete ones operating with a small sampling period T_s. Therefore, when designing AOC controllers, it is necessary to determine the threshold values of T_e that are still sufficiently suitable for use. By simulation we found the minimum values of the equivalent delay $T_e = T_{emin}$ in Table 2 that correspond to the same controller and DOB filters and selected values of *m* and *n* to stable setpoint and disturbance responses with a limited value of excessive controller effort fulfilling the requirement $TV_1(u) < 0.1$ for $r = 0$. Alternatively, with regard to the presence of a saturation block in the loop, it would also be possible to limit the range of filter settings using hyperstability and absolute stability methods[30].

Modified Åström-Smith predictor

The Smith predictor (SP) was designed to reconstruct and compensate for output disturbances of stable first-order time-delayed systems[1]. Subsequently, it has been the focus of research for about seven decades, with mainly attempts to extend it to other types of processes. It also became the basis for a more advanced model-based approach to controller design called internal model control (IMC)[46–48]. Despite the fact that its use in connection with simple ultralocal models has long faced several problems.

The SP proposed in[1] can be interpreted as a feedforward control extended by reconstruction and compensation of output disturbances. This open-loop control concept can be used only for open-loop stable

T_{emin} - setpoint response						
–	$m=0$	$m=1$	$m=2$	$m=3$	$m=4$	$m=5$
$n=1$	0.001	0.002	–	–	–	–
$n=2$	0.001	0.003	0.086	–	–	–
$n=3$	0.001	0.004	0.099	0.227	–	–
$n=4$	0.001	0.005	0.112	0.244	0.370	–
$n=5$	0.002	0.006	0.124	0.267	0.388	0.500
$n=6$	0.002	0.007	0.136	0.290	0.416	0.526
$n=7$	0.002	0.008	0.147	0.314	0.447	0.557
$n=8$	0.003	0.009	0.158	0.338	0.480	0.594
T_{emin} - disturbance response						
–	$m=0$	$m=1$	$m=2$	$m=3$	$m=4$	$m=5$
$n=1$	0.001	0.002	–	–	–	–
$n=2$	0.001	0.003	0.083	–	–	–
$n=3$	0.001	0.004	0.098	0.187	–	–
$n=4$	0.001	0.005	0.113	0.204	0.326	–
$n=5$	0.002	0.006	0.126	0.224	0.334	0.502 -
$n=6$	0.002	0.007	0.139	0.245	0.354	0.513
$n=7$	0.002	0.008	0.152	0.266	0.378	0.527
$n=8$	0.003	0.009	0.164	0.287	0.405	0.554

Table 2. Minimal values T_{emin} required for filter time constant $T_f = T_e/n$ to guarantee stable transients with limited maximal shape deviations $max(TV_1(u)) = 0.1$ at the process input calculated for $r = 0$, $T_s = 0.001$, $T_d = 1$ and $K_s = 1$.

time-delayed processes[49]. For integral models, the output disturbance is unobservable. Also, with regard to marginally stable dynamics, instead of output disturbances, it is necessary to reconstruct and compensate for the input disturbances, which requires significant structural transformations. If we began the AOC design by criticizing the inconsistent terminology used in PID control, the situation would not be better with regard to the Smith predictor. When defining the SP as a structure for reconstructing and compensating output disturbances by changing the setpoint variable, then compensating input disturbances by changing the stabilizing controller offset is apparently a completely different structure. This should also be clearly respected by using an appropriate name for it. The Åström-Smith predictor modified for ultralocal models according to[3,4,50] and hereinafter denoted ASP satisfies this requirement. Many other names, such as the filtered Smith predictor (FSP), ignored it[49]. Hence, they contributed to the not-so-great reputation of work in this field. Moreover, SP was often interpreted just as an "improvement" of the PI controller, which remained as part of the primary SP loop. Since the primary loop task was only to generate feedforward control, using the PI controller with various optimal settings for stable, integral, and unstable systems[51], although a much simpler P controller with two degrees of freedom (2DoF P) should suffice, just continued spreading misleading information. After repeated criticism, the authors of the criticized works never put this matter in proper perspective in their papers. After a series of works using the primary loop PI controller, without any explanation and citation of the correcting papers, suddenly the 2DoF P control was used instead[52]. The reaction of practice to the state of research in the field of the SP was a criticism[53] and a lack of interest in the given issue regarding inclusion in the basic topics of teaching in the field of bachelor's studies[54].

Noise attenuation problem of ASP

In the first step, ASP will be simulated according to Fig.3 extended by blocks to generate uniform random noise with amplitude ± 0.01. We will try to show the problems of the modified version of ASP by comparing it with AOC. Although the notation used mostly respects the paper[4], when comparing it with AOC, it is impossible to avoid the need for some adjustments in the labeling. Thus, the gain of the process and its model will be simultaneously denoted as $K_s = K_{sp} = b$ and the model of the process dead-time as $T_{dp} = L$. The process parameter a_0 will first be chosen as $a_0 = 0$. The controller is tuned by the gain k_0/K_{sp} of the feedforward control (the part of the schema in Fig. 3 marked with a light blue background color) in order to obtain a simple first-order transfer function

$$T(s) = \frac{k_0}{s + k_0} e^{-sL}, \tag{24}$$

with the dead-time L and a time constant

$$T_{ff} = 1/k_0. \tag{25}$$

With parameters k_p, k_i, and

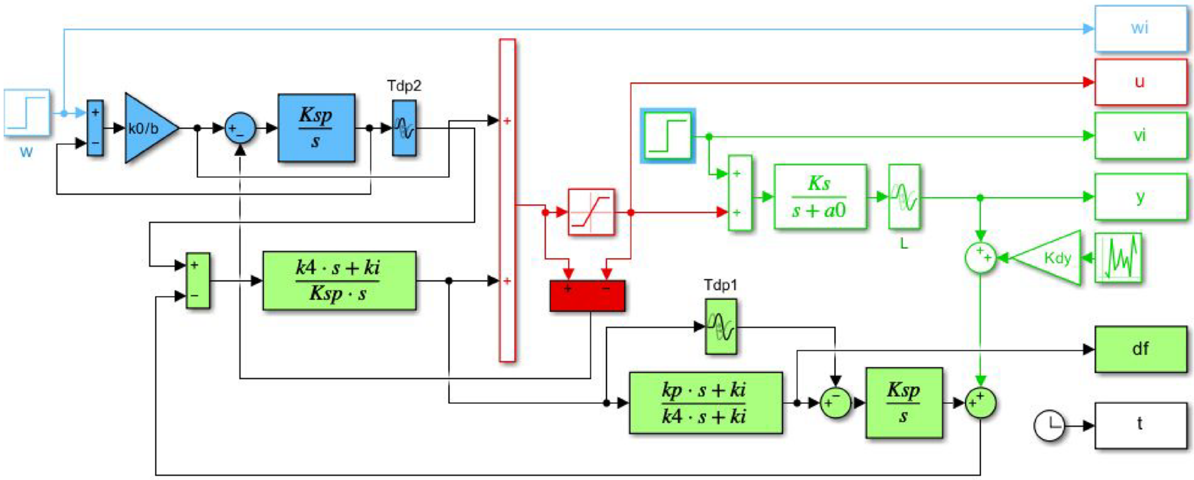

Figure 3. ASP: Åström-Smith predictor modified according to[3] and[4] with the measurement noise generated in Matlab/Simulink by Uniform Random Number block with amplitudes ± 0.01, $K_{dy} = 1$.

$$k_4 = k_p + k_i L, \tag{26}$$

it is possible to obtain a disturbance response

$$F_{dy} = \frac{Y(s)}{D_i(s)} = K_s \frac{s^2 + k_p s + k_i - (k_4 s + k_i)e^{-sL}}{s[s^2 + k_p s + k_i]} \tag{27}$$

corresponding to the part of the schema in Fig. 3 marked with a light green background color. It can be specified with a polynomial

$$P(s) = s^2 + k_p s + k_i = (s + \alpha)^2. \tag{28}$$

$T_{di} = 1/\alpha$ can be interpreted as the time constant of the response to the input disturbance. While guaranteeing stability and $t \to \infty$, when $s \to 0$ we achieve rejection of constant disturbances thanks to $F_{dy}(0) = 0$. However, the limit $\lim_{s \to 0} F_{dy}(s) = 0$ must be taken into account by applying L'Hospital's rule. This is the first difference compared to the calculation of the steady-state disturbance response value from (18) when using AOC.

ASP versus AOC

The process parameters (19) of the comparative experiment will also be stated using the notation given in[4] as

$$K_{sp} = K_s = b = 1; \; T_{dp} = L = 0.5; \; U_{max} = 1.15; U_{min} = -0.15; \tag{29}$$

Two sets of controller parameters will be used.

$$\begin{aligned} &\text{ASP1: } k_0 = 2; \; T_{ff} = 1/(K_s k_0) = 0.5; \; \alpha = 4/L = 4/T_{dp} = 8; \; k_p = 2\alpha = 16; \; k_i = \alpha^2 = 64 \\ &\text{ASP2: } k_0 = 0.5; \; T_{ff} = 1/(K_s k_0) = 2; \; \alpha = 1/L = 1/T_{dp} = 2; \; k_p = 2\alpha = 4; \; k_i = \alpha^2 = 4 \end{aligned} \tag{30}$$

In ASP1, the choice of the feedforward P controller gain $k_0 = 2$ gives the setpoint feedforward time constant $T_{ff} = 1/(K_s k_0) = 0.5$. By parameter $\alpha = 1/L = 1/T_{dp} = 8$ and $k_p = 2\alpha = 16$, $k_i = \alpha^2 = 64$, the double time constant $T_{12} = 0.125$ is given that determines the dynamics of the disturbance reconstruction. The calculation of DOB parameters is concluded by the specification $k_4 = k_p + k_i L = 48$. In ASP2, by choosing $k_0 = 0.5$, the feedforward setpoint time constant is set to $T_{ff} = 1/(K_s k_0) = 2$. By choosing $\alpha = 1/L = 1/T_{dp} = 2$, the parameters $k_p = 2\alpha = 4$ and $k_i = \alpha^2 = 4$ are derived, which give a double time constant for disturbance reconstruction $T_{12} = 0.5$. The next parameter $k_4 = k_p + k_i L = 6$ is then uniquely specified and, together with k_p and k_i, determines the disturbance response dynamics (27).

The AOCs proposed for a given process with the derivative degree $m \in [0, 5]$ (19), with the order of the applied low-pass filter $n = m + 2$ and with the choice of $T_e = T_{dp}, T_f = T_e/n$ are compared with the ASP in Fig. 4.

Without measurement noise (see Table 3), the response achieved by ASP1 is faster than P0-P3 from the point of view of IAE_s and faster than all P0-P5 from the point of view of IAE_d (see Fig. 5). From the setpoint responses point of view, ASP2 is slower than all the AOC step responses P0-P5 and the disturbance response P3 - P5. Formally, both the setpoint and the disturbance responses of ASP could be made significantly faster by a tighter controller tuning. However, even considering the influence of measurement noise with an amplitude of ± 0.01, that is, with a relatively low level of up to 1% of the setpoint value, the excessive controller effort of the

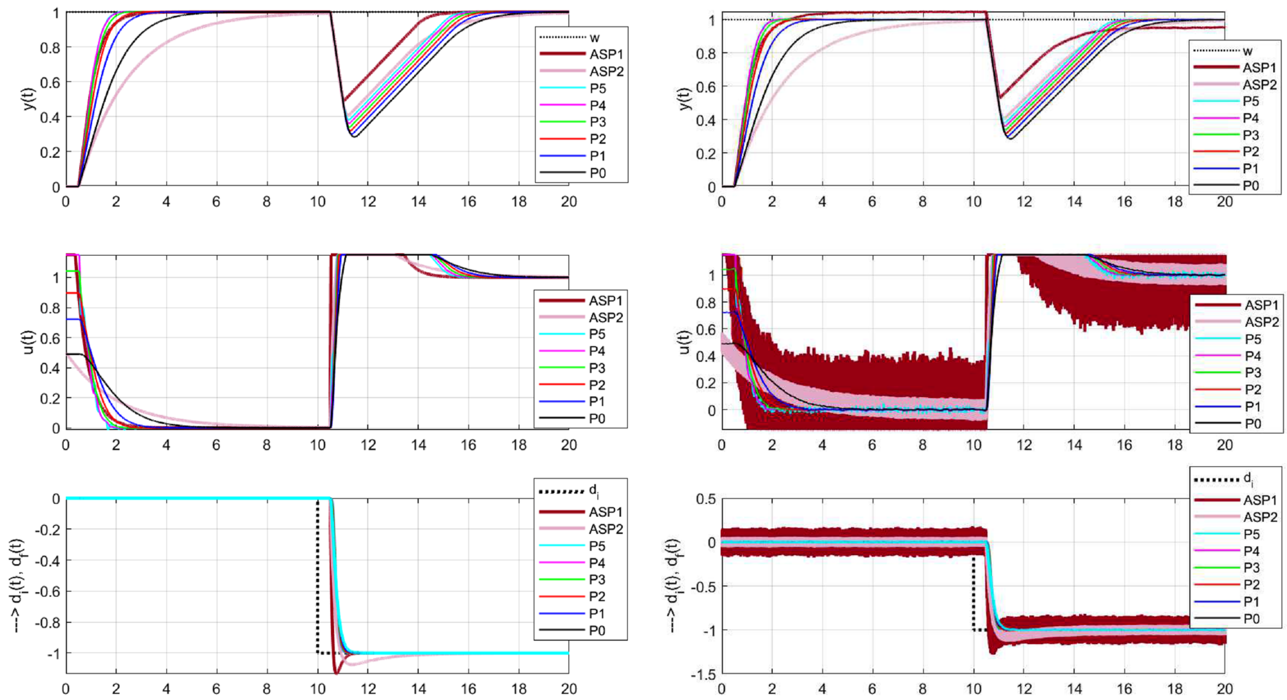

Figure 4. Setpoint and input disturbance step responses of ASP and MRDP-optimal AOC controllers $P_0 - P_5, m \in [0,5], n = m+2, r = 0$ for IPDT system (19); $T_s = 0.001$, left-no measurement noise, $K_{dy} = 0$ in Simulink schemes in Figs1-3; right - measurement noise with amplitude $\delta = 0.01$, $K_{dy} = 1$.

	ASP1	ASP2	P0	P1	P2	P3	P4	P5
IAE_s	1.079	2.483	1.789	1.324	1.146	1.048	0.991	0.989
IAE_d	1.025	1.776	2.265	2.031	1.887	1.763	1.653	1.557
$TV_1(u_s)$	0.00	0.00	0.00	0.00	0.00	0.00	0.01	0.40
$TV_1(u_d)$	0.30	0.29	0.30	0.00	0.00	0.00	0.00	0.30
J_{1s}	0.000	0.000	0.000	0.000	0.000	0.000	0.006	0.401
J_{1d}	0.000	0.001	0.677	0.000	0.001	0.000	0.001	0.471

Table 3. Performance measures of ASP1, ASP2 and Pm controllers from Fig. 4 left for $m \in [0,5]$, $\delta = 0$, $K_{dy} = 0$.

ASP is significantly higher than for the AOC responses (see Fig. 4 right and Table 4). While AOC reaches values ≈ 1, with ASP values of

$$\text{ASP1: } TV_1(u_s) = 2197.68;\ TV_1(u_d) = 1537.80,;\ \text{ASP2: } TV_1(u_s) = 400.98;\ TV_1(u_d) = 297.28, \quad (31)$$

that means up to 2000 times larger were measured. Such an increase of the excessive effort is unacceptable for many applications. Among the AOC options, only the use of a stabilization controller P0 without additional filtering is not recommended. The choice of the optimal solution from options P1-P5 will then depend on the priorities of the given application. Moreover, as in the case of ASP1 in Fig. 4 right, high measurement noise typically leads to permanent control error in conjunction with control signal saturation[55]. Also worth noting is the faster reconstruction of the resulting external disturbance in the AOC.

ASP augmented by a low-pass filter

To achieve a sufficient measurement noise attenuation, ASP can be used in the same way as AOC with included binomial low-pass filters (see Fig. 6). Again, it can be specified by the polynomial $P_n(s)$ (12) with some degree $n \geq 1$ and a time constant T_f. In the controller design, the filter can be included by adding an equivalent dead time T_e to the process delay T_{dp} to obtain the "total" closed-loop dead time (14). The simplest expression of T_e by the filter time constant T_f is according to (15).

Transients corresponding to the ASP1 controller with $n \in [1,4]$ and parameters

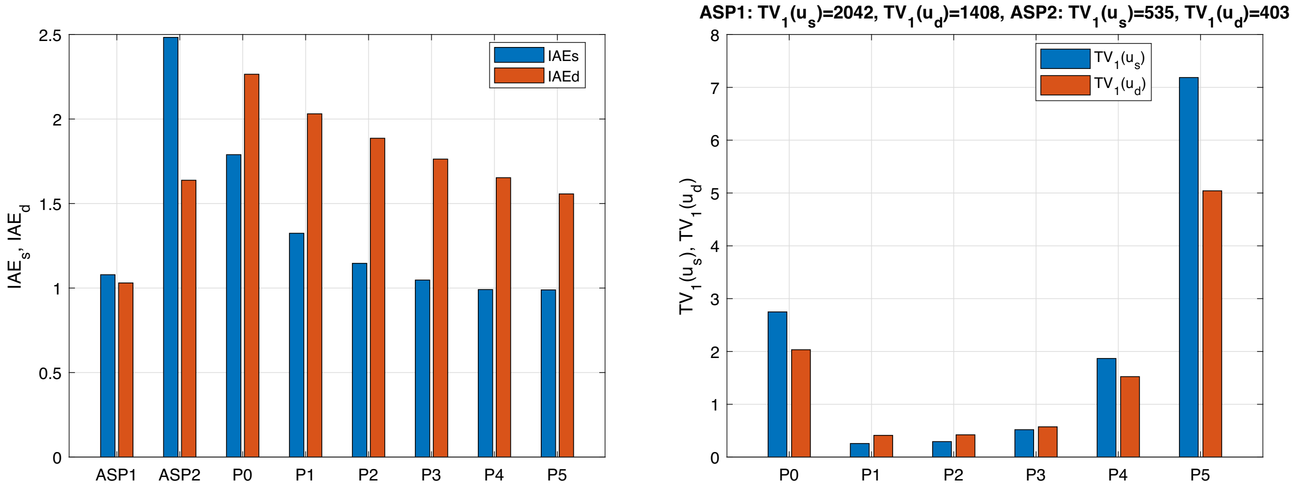

Figure 5. Left - IAE values of setpoint and input disturbance step responses of ASP and MRDP-optimal AOC controllers from Fig. 4 (no noise); right - the corresponding excessive control $TV_1(u)$ values for measurement noise generated by the Uniform random number block in Simulink with amplitude $\delta = \pm 0.01$, $T_s = 0.001$, $K_{dy} = 1$.

	ASP1	ASP2	P0	P1	P2	P3	P4	P5
IAE_s	1.488	2.484	1.79	1.325	1.149	1.051	0.995	0.993
IAE_d	1.363	1.786	2.263	2.027	1.885	1.763	1.654	1.557
$TV_1(u_s)$	2197.68	400.98	2.75	0.26	0.29	0.52	1.87	7.19
$TV_1(u_d)$	1537.80	297.28	2.03	0.41	0.42	0.57	1.52	5.04
J_{1s}	3.270	0.996	4.921	0.341	0.337	0.545	1.857	7.136
J_{1d}	2.095	0.531	4.600	0.835	0.796	1.012	2.519	7.863

Table 4. Performance measures of ASP1, ASP2 and Pm controllers from Fig. 4 right for $m \in [0, 5]$, $\delta = \pm 0.01$, $K_{dy} = 1$.

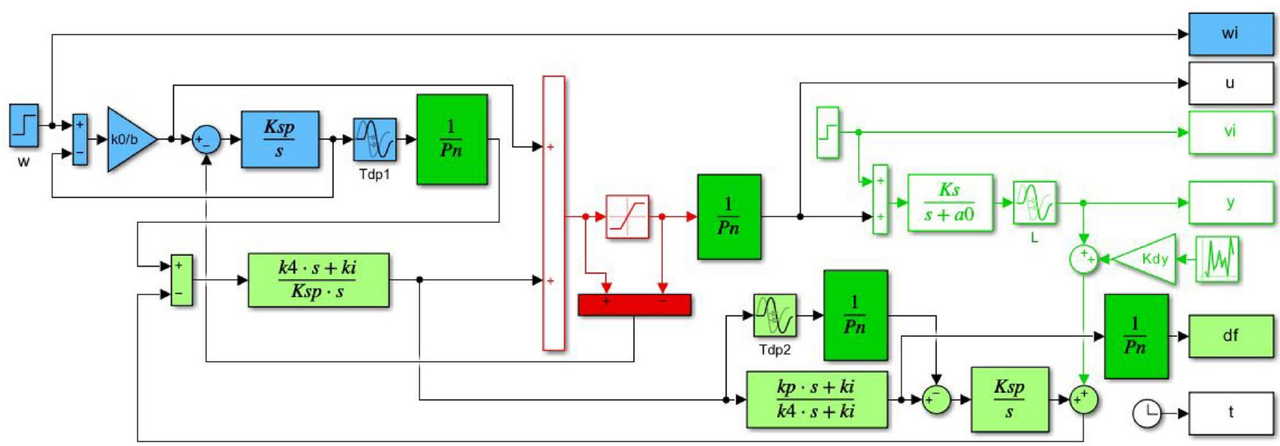

Figure 6. ASP extended by a low-pass filter $Q_n = 1/P_n$ (dark green background color).

$$k_0 = 2;\ T_{ff} = 1/(K_s k_0) = 0.5; T_e = 0.8L;\ \ \alpha = 4/(L + T_e);\ k_p = 2\alpha;\ k_i = \alpha^2 \tag{32}$$

are in Fig. 7. The corresponding performance measures are shown in Fig. 8 and Table 5. For comparison with the AOC design, a P_5^3 controller with the prefilter F_{pr}^3 (11), $r \in [0, 3]$ (options P30-P33) and the same value of $T_e = 0.8L$ as above was used.

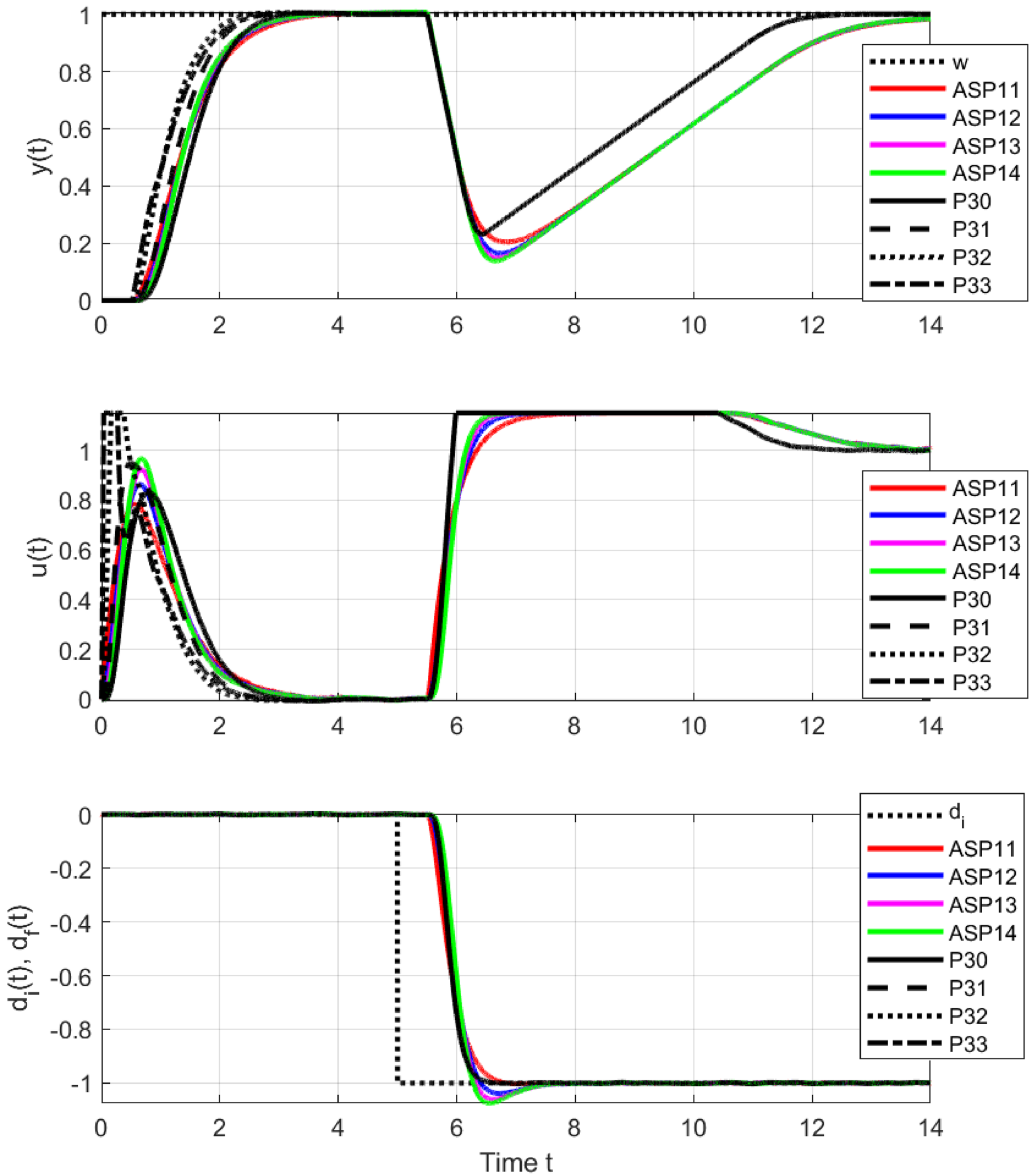

Figure 7. Setpoint and input disturbance step responses of ASP11-ASP14 and AOC P30-P33 controllers for IPDT system (29) designed as P_5^3 controller with the prefilter F_{pr}^3 (11), $r \in [0, 3]$; $T_s = 0.001$; measurement noise with amplitude $\delta = 0.01$, $K_{dy} = 1$.

From the point of view of IAE_s, the values achieved by including filters with $P_n(s)$ of the first to fourth order with equivalent delay $T_e = 0.8T_m$ increase. compared to equally set ASP1, by less than 50%. In the case of IAE_d, they increase, too. Compared to ASP1, by about 2x. However, excessive controller effort $TV_1(u)$ is significantly reduced. For the second-order filter it already reaches below the level corresponding to the modifications P30-P33 with a prefilter that cancels 0 to 3 closed-loop time constants T_o. The ASP variants with an additional low-pass filter of the first order and P33 with a prefilter canceling with $r = m = 3$ a triple closed-loop time constant T_o are therefore the least recommended. The last observation can be generalized for another

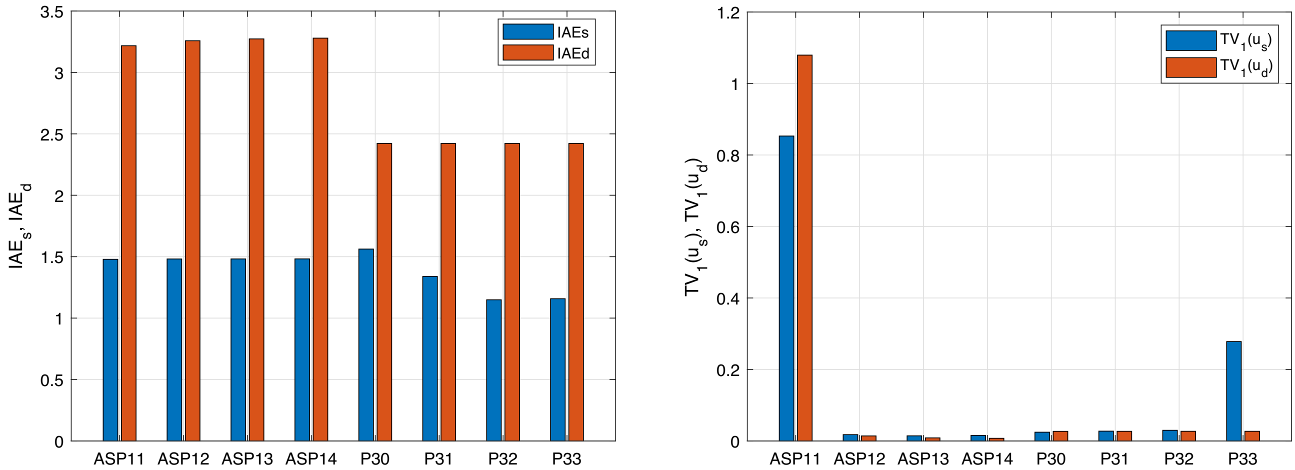

Figure 8. Left IAE values of setpoint and input disturbance step responses of ASP11-ASP14 and AOC P30-P33 controllers from Fig. 7 left and the corresponding excessive control $TV_1(u)$ values (right); measurement noise with amplitude $\delta = \pm 0.01$, $T_s = 0.001$, $K_{dy} = 1$.

	ASP11	ASP12	ASP13	ASP14	P30	P31	P32	P33
IAE_s	1.479	1.481	1.482	1.482	1.562	1.340	1.149	1.158
IAE_d	3.217	3.258	3.273	3.279	2.422	2.422	2.422	2.422
$TV_1(u_s)$	0.85	0.02	0.01	0.02	0.02	0.03	0.03	0.28
$TV_1(u_d)$	1.08	0.02	0.01	0.01	0.03	0.03	0.03	0.03
J_{1s}	1.261	0.026	0.021	0.023	0.038	0.037	0.034	0.322
J_{1d}	3.473	0.046	0.028	0.024	0.065	0.065	0.065	0.065

Table 5. Performance measures of transient responses with ASP11-14, and P30-33 controllers from Fig. 7 for $\delta = \pm 0.01$, $K_{dy} = 1$.

AOC design, where it is not recommended to work with the prefilter designed for $r > 2$ when choosing MRDP-optimal controller parameters. The reason is the emergence of high control signal peaks after the setpoint signal step. These quickly decay after saturation. Only then does a longer-term growth in the process input occur. This increases the excessive controller effort and values of IAE_s.

Let us also recall that the permanent error of the ASP controller due to control signal saturation disappeared already with the inclusion of a first-order filter, even without further changes in its settings or structure. The reason is the decrease in amplitude of the noisy control signal in the vicinity of steady-states below the saturation limits.

From the point of view of the comparison carried out with ASP and AOC for the given T_e, we could state the most obvious faster compensation of the disturbance steps in AOC, together with a faster reconstruction of the value of d_i. Of course, the comparison of both solutions with a larger number of adjustable parameters will be possible only on the basis of a more extensive study. In such a comparison, it can be expected that ASP will not be as sensitive to a decrease in the equivalent filter constant T_e as AOC, for which T_e should not be reduced below the value T_{emin}. And it should also take into account the applications of both considered controllers for numerous other processes that can be approximated by the IPDT model. We will be mainly interested in the case of unstable processes for which it is not enough to use SP in its original simpler version.

Dependence of the filtered ASP on the measurement noise amplitude

The different disturbance rejection properties achieved with AOC and ASP under noise impact are evident when applying random noise of different amplitudes. It is shown in Fig. 9 for ASP with setting (32) when choosing $T_e = 0.8L, n = 2$ and measurement noise values

$$\delta = \{0.001;\ 0.01;\ 0.02;\ 0.04\} \tag{33}$$

The corresponding responses ASP0.001-ASP0.04 with increasing noise amplitude slow down the dynamics. Despite the other controller parameters remaining unchanged, for $\delta = 0.04$ the disturbance compensation of the filtered ASP is practically ineffective.

The three responses P30-P32 of AOC with $m = 3, n = 5, T_e = 0.9T_{dp}$ and the prefilter parameter $r \in [0, 2]$ correspond to the highest amplitude of measurement noise $\delta = 0.04$. The achieved values of *IAE* and J_1 (see

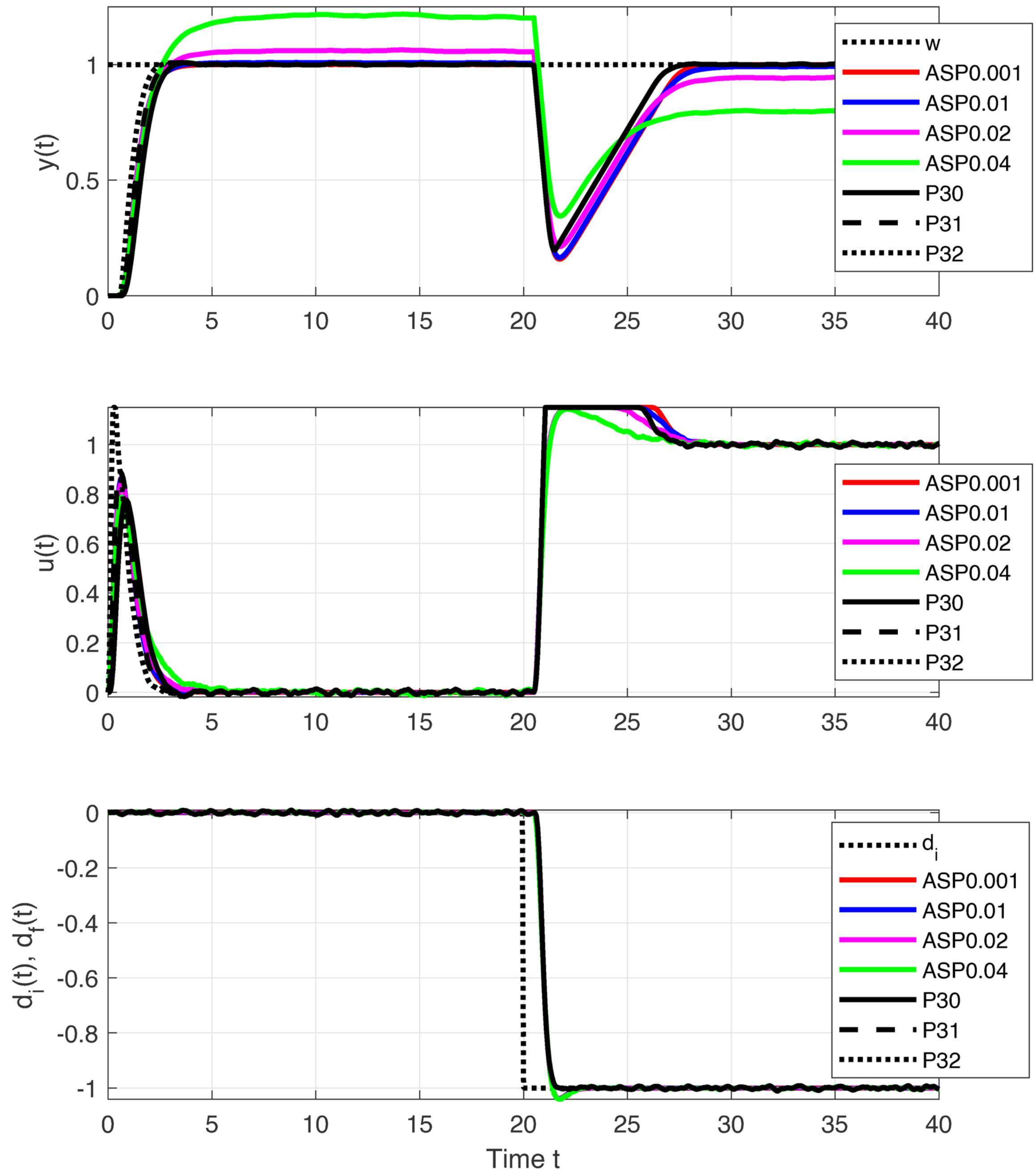

Figure 9. Setpoint and input disturbance step responses of ASP0.001-ASP0.04 for the IPDT process (29) tuned according to (32), extended by a filter with $T_e = 0.8L, n = 2$ for the measurement noise (33) and AOC P30-P32 controllers with $T_e = 0.9T_{dp}$ and the measurement noise amplitude $\delta = 0.04$ designed as P_5^3 controller with the prefilter F_{pr}^3 (11), $r \in [0, 2]$; $T_s = 0.001$; $K_{dy} = 1$.

Fig. 10, or Table 6) of filtered ASP0.001-ASP0.04 control, achieved even for lower values of noise amplitudes, can exceed those corresponding to AOC. Furthermore, with permanent control error, the values of *IAE* and J_1 for ASP would continue to increase with increasing time.

The simulation results show that while more efficient filtering can significantly reduce excessive controller effort. However, for higher noise amplitudes, persistent deviation of the filtered ASP cannot be eliminated even by choosing significantly higher values of T_e.

Application of AOC and ASP to different systems

Unstable process control

In[4] the author mentions the possibility of extending the presented modification of ASP to control unstable processes, but so far we are not aware of work with such content. Since the control of unstable processes is still an important topic, next we are going to show how the problem can be solved using the above presented

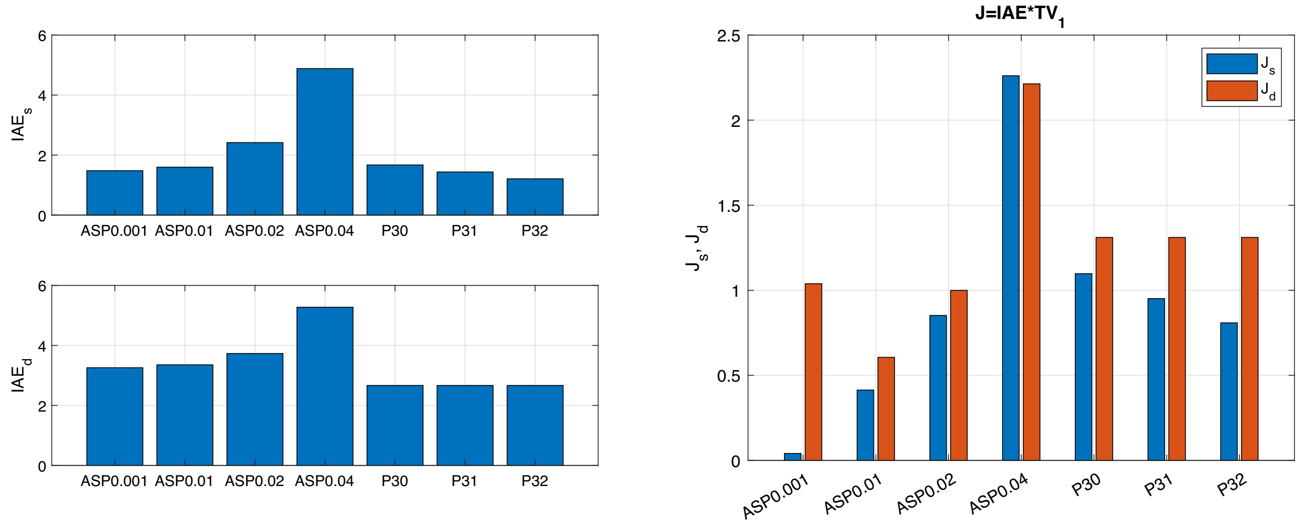

Figure 10. Left IAE values of setpoint and input disturbance step responses of ASP0.001-ASP0.04 and AOC P30-P32 controllers from Fig. 9 and the corresponding values of the combined cost function J_1 (23) (right); measurement noise with amplitude (33), $T_s = 0.001$, $K_{dy} = 1$.

	ASP0.001	ASP0.01	ASP0.02	ASP0.04	P30	P31	P32
IAE_s	1.480	1.597	2.417	4.884	1.670	1.437	1.208
IAE_d	3.256	3.353	3.731	5.274	2.666	2.666	2.666
$TV_1(u_s)$	0.03	0.26	0.35	0.46	0.66	0.66	0.67
$TV_1(u_d)$	0.32	0.18	0.27	0.42	0.49	0.49	0.49
J_{1s}	0.041	0.413	0.852	2.261	1.097	0.951	0.809
J_{1d}	1.038	0.606	0.999	2.213	1.311	1.311	1.311

Table 6. Performance measures of transient responses with ASP0.001-0.04, and P30-33 controllers from Fig. 9 for δ (33), $K_{dy} = 1$.

solutions derived for the IPDT model. That is, after approximating unstable processes with the dominant first-order dynamics with an IPDT model. This approximation can be achieved using the procedure for linearizing nonlinear processes using two types of linear models formulated recently in[42]. The ultralocal models are equivalent to replacing a process nonlinearity using the zeroth term of its Taylor expansion. The local models correspond to historically used procedures that work with both the zeros and the first term of such an expansion. This statement only explicitly expressed the long-known "intuitive practice" of using integral process models also in situations, when the process dynamics corresponds to a more complex system. Among the postmodern methods that imply the use of ultralocal models, let us mention at least ADRC or MFC. However, they can also be met in the PID control framework (for example, in[56]). Thus, in the context of this paper, the above proposed AOC and ASP can therefore be used to control the entire range of time-delayed processes with dominant first-order dynamics. The most interesting will probably be to illustrate the validity of this conclusion first with a control of an unstable process.

Approximation of an initial part of process step response by an IPDT model

The work[55] deals with the control of a nonlinear chemical reactor, which was approximated by linearization around the operating point by an unstable transfer function

$$S(s) = \frac{Y(s)}{U(s)} = \frac{3.433e^{-20s}}{103.1s - 1} \approx \frac{0.0333e^{-20s}}{s} \tag{34}$$

Obviously, for $s \to \infty$ that is, for $t \to 0$, this process can be approximated by IPDT model (1) with parameters

$$K_{sp} = 0.0333; \; T_{dp} = 20. \tag{35}$$

In[55] the control specifications considered a step change of the setpoint variable $w(t) = 0.3$ from the initial output value $y(0) = 0$ and a symmetrical saturation limit $|u(t)| \le 0.03$. However, to achieve a steady output $y(\infty) = 0.3$ of the system (34), a steady value of the control signal $u(\infty) = -0.3/3.433 = -0.0874$ is required.

This, in absolute value, is significantly higher than the saturation limit value 0.03 given in the article. The article obviously contains an error, so comparison of our results with the results presented in[55] is meaningless. The saturation limits for further experiments will increase to the value ± 0.3.

With a relatively small number of samples taken during one transient response, it probably makes no sense to base ourselves on their normal distribution. To emulate measurement noise, we will instead use a random noise signal with different amplitudes. For the sampling period $T_s = 0.02$ and $L = 20$, the modified filtered ASP controller has been designed with parameters

$$k_0 = 2;\ T_{ff} = 1/(K_s k0) = 0.5;\ T_e = 0.4L;\ \alpha = 4/(L + T_e);\ k_p = 2\alpha;\ k_i = \alpha^2 \tag{36}$$

Compared to the previously treated control of the IPDT system, the only change in controller parameters (32) was reflected in a decrease in the value of T_e. Unstable processes are very sensitive to circuit delays; therefore, the intensity of measurement noise filtration specified by the choice of T_e must always be considered. Another change concerns the amplitudes of random noise, when due to the change of model parameters K_s and T_d, problematic situations occur for higher noise amplitudes. The highest verified value is now $\delta = 0.6$

$$\delta = \{0.01;\ 0.1;\ 0.2;\ 0.6\}. \tag{37}$$

$\delta = 0.6$ was also applied to all three responses measured with P30-P32.

The transient responses ASP0.01-ASP0.6 with a low-pass filter order $n = 2$ and the noise amplitudes (37) are shown in Fig. 11. The control signal does not attack saturation limits in the vicinity of the steady state, so the undesirable permanent error (such as in[55]) does not occur. Thus, it must be caused by the filtered ASP structure itself. The reconstructed disturbance also appears in the initial phase of transients, where no external disturbances are applied to the process. It corresponds to the contribution of the internal feedback of the process. Similarly, as the so-called lumped/total disturbances that occur in the case of ADRC[34].

The transient responses corresponding to the P30-P33 controllers with the prefilter parameter $r \in [0, 2]$ in Fig. 11 have no problems with permanent error. They correspond to $T_e = 0.9L$ (chosen with respect to Table 2 slightly larger than $T_e = 0.4L$ used for ASP in (36)). However, the MRDP-optimal controller tuning specified for two-parameter IPDT models leads this three-parameter unstable process to overregulation. The excessive controller effort values are shown in Fig. 11 (see also Table 7).

Again, the achieved values of *IAE* and J_1 (see Fig. 12, or Table 7) of the filtered ASP control, achieved even for lower values of noise amplitudes, can exceed those that correspond to AOC. Furthermore, with permanent control error, the values of *IAE* and J_1 for ASP would continue to increase with increasing time.

Real-time control of a thermal process

The robustness and feasibility of the proposed AOC and ASP algorithms can also be illustrated by real time control of a stable thermal process. It brings an expansion of the review dedicated to dead-time compensators[49] considered in the discrete-time domain. The use of the same process allows us to shorten the detailed description of the Arduino-based laboratory plant TOM1A[57–59] (Fig. 13) and the control experiment. It also allows to use the previously identified parameters of the IPDT model

$$K_{sp} = 0.01;\ T_{dp} = 0.3;\ T_s = 0.02;\ U_{min} = 0;\ U_{max} = 100. \tag{38}$$

By evaluating the open-loop step responses of the process, it would also be possible to obtain the parameter $a_0 \approx 0.05\mathrm{s}^{-1}$ of the local linear model. This corresponds to the dominant time constant of the fast process mode $T_p = 1/a_0 \approx 20\mathrm{s}$[60]. However, the use of local linear model will usually not bring any further improvement in the closed-loop responses. Therefore, we again prefer the possibility to work with a simpler ultralocal model considering $a_0 = 0$. Compared to older experiments in[49] carried out in the summer, the first difference to be noted is that due to the lower ambient temperature in the laboratory during the winter season, the responses run with slightly lower setpoint values.

In short, we will recall that the thermal process is formed by a heat source (bulb), a temperature sensor, and a cooling fan. Due to several modes of heat transfer, the internal dynamics is significantly different from the IPDT approximation. Nevertheless, by fitting the model parameters to the fast mode (heat transfer by radiation), it is possible to achieve control responses that are very close to the ideal shapes mentioned above when defining the monotonicity-based performance measures.

Two sets of parameters were proposed for setting up the filtered ASP:

$$\begin{array}{l} \text{ASP1: } k_0 = 2;\ T_{ff} = 1/(K_s k_0) = 0.5;\ \alpha = 4/(L + T_e) = 8.8889; \\ k_p = 2\alpha = 17.7778;\ k_i = \alpha^2 = 79.0123;\ k_4 = k_p + k_i(L + T_e) = 53.3333 \\ \text{ASP2: } k_0 = 0.5;\ T_{ff} = 1/(K_s k_0) = 2;\ \alpha = 1/(L + T_e) = 2.2222; \\ k_p = 2\alpha = 4.4444;\ k_i = \alpha^2 = 4.9383;\ k_4 = k_p + k_i(L + T_e) = 6.6667. \end{array} \tag{39}$$

When using ASP (39) with $T_e = T_{dp}/2 = 0.15, n = 2, T_f = 0.075$, a steady-state control error occurs already in the first phase of transient responses. The initial phase (see the experiment scenario in[49]) was included to set the output temperature at the value $y_0 = w_0 = 28^o$C. The permanent error also dominates when the setpoint increases to a new level $w_1 = 32^o$C (see Fig. 14). The experiments show that the permanent control error is higher at higher noise amplitudes, which corresponds to a more aggressive setting of the ASP1 controller. It is possible that by further slowing down the responses, the resulting permanent error could still be reduced.

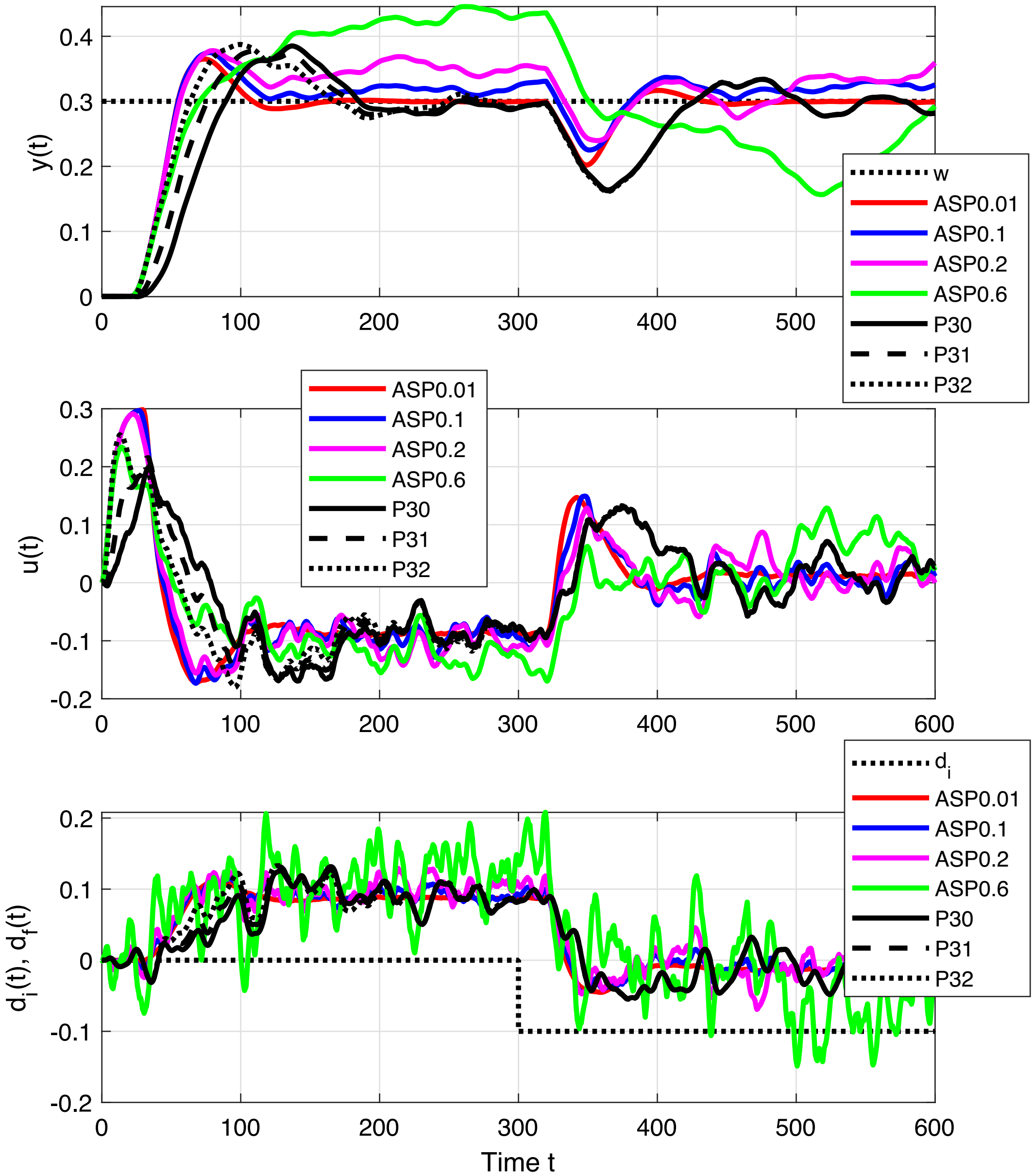

Figure 11. Setpoint and input disturbance step responses of ASP0.01-ASP0.6 tuned according to (36), extended by a filter with $T_e = 0.4L, n = 2$ for the measurement noise (37) and AOC P30-P32 controllers with $T_e = L$ and the measurement noise amplitude $\delta = 0.6$ designed as P_5^3 controller with the prefilter F_{pr}^3 (11), $r \in [0, 2]; T_s = 0.01;, K_{dy} = 1$.

	ASP0.01	ASP0.1	ASP0.2	ASP0.6	P30	P31	P32
IAE_s	14.952	17.973	24.259	37.061	24.604	22.168	19.937
IAE_d	3.990	7.647	8.685	20.998	10.658	10.758	10.865
$TV_1(u_s)$	0.31	0.87	0.94	0.89	0.88	0.92	0.97
$TV_1(u_d)$	0.45	1.05	1.17	1.18	1.12	1.12	1.12
J_{1s}	4.609	15.548	22.700	33.158	21.761	20.330	19.321
J_{1d}	1.80951	8.013	10.123	24.771	11.934	12.055	12.186

Table 7. Performance measures of transient responses with ASP0.01-0.6, and P30-33 controllers from Fig. 11 for δ (37), $K_{dy} = 1$.

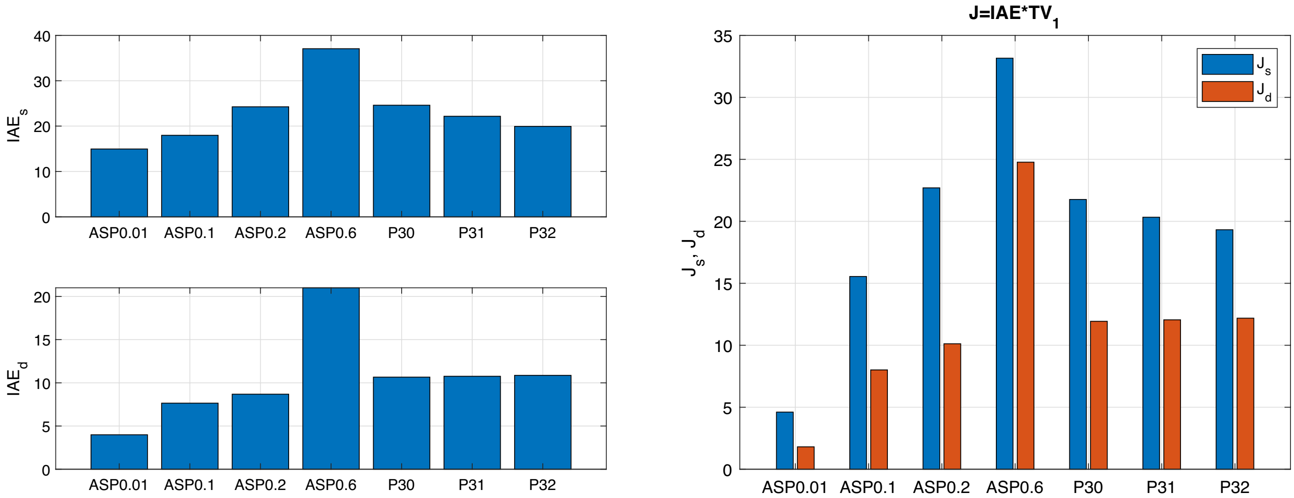

Figure 12. Left IAE values of setpoint and input disturbance step responses from Fig. 11 with ASP0.01-ASP0.6 (36) achieved under the measurement noise amplitudes (37) and AOC P30-P32 controllers with the noise amplitudes $\delta = 0.6$; right - the corresponding values of the combined cost function J_1 (23); $T_s = 0.01$, $K_{dy} = 1$.

Figure 13. The Arduino-based laboratory plant TOM1A.

However, already the transient responses in Fig. 14 are relatively slow. The maximum control signal barely reaches the saturation limit. So there was no motivation to experiment further with such control.

To illustrate the application of AOC, we will limit ourselves to two examples of transient responses in Fig. 15. These were obtained using controllers designated for simplicity as P30 and P54. The first of them corresponds to the use of P_5^3 with the simplest prefilter $F_p^3(s)$ (4), that is, with a prefilter $F_{pr}^3(s)$ (11) designed for $r = 0$. The filter $P_n(s)$ of order $n = 5$ was set for $T_e = T_{dp}$. In the case of controller P54, it is a controller P_7^5 with a prefilter $F_{pr}^5(s)$ (11) designed for $r = 4$. The low-pass filter $P_n(s)$ with $n = 7$ was designed for $T_e = 3T_{dp}$. The waveforms obtained show that the AOC gives nearly ideal setpoint and disturbance responses for a wide range of possible settings expressed by the parameters m, n, r and T_{dp}. Improvement (smoothing) of transients occurred even when considering relatively high values of $m = 5$.

Discussion

Based on the experiments carried out, the revision of the oldest ARC and SP controllers led to AOC and filtered ASP designed on the basis of IPDT models. The attempt to generalize them to control a wider class of time-delayed processes with dominant first-order dynamics can lead to the following comments. The transient responses of the AOC controllers P30-P32 in Fig. 11 achieved with dominant UFOTD dynamics (34) approximated by their ultralocal IPDT models (35), show some overshooting with $T_e = T_{dp} = L$, $n = m + 2, r \in [0, 2]$ and $\delta = 0.6$. Hence, when wishing to avoid the emergence of overregulation in this particular application, the MRDP controller tuning needs to be modified. The overshooting problem is a consequence of the effort to achieve the fastest possible transient responses without taking into account special features of the internal dynamics of the process.

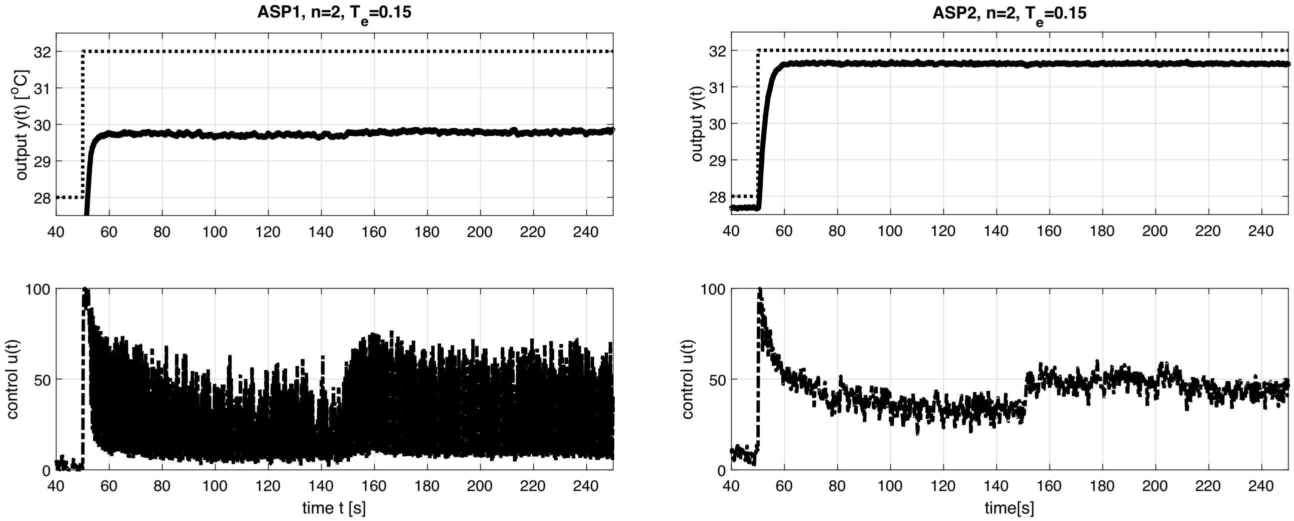

Figure 14. Thermal process with ASP1 and ASP2 controllers (39): setpoint step at time $t = 50$s from $w_0 = 28^o$C to $w_1 = 32^o$C followed by an input (load) disturbance at $t = 150$s produced by a setpoint step applied to a cooling fan.

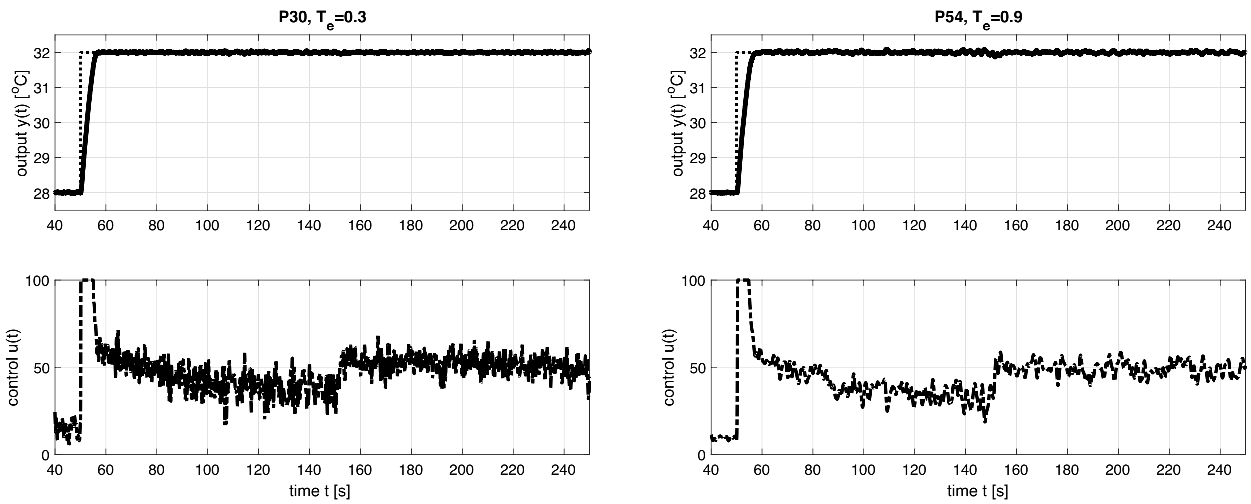

Figure 15. Thermal process with AOC controllers P30 (tuned with $T_e = T_{dp}$) and P54 (with $T_e = 3T_{dp}$): setpoint step at time $t = 50$s from $w_0 = 28^o$C to $w_1 = 32^o$C followed by an input (load) disturbance at $t = 150$s produced by a setpoint step of a cooling fan.

Problems with overshooting do not arise with AOC applied to the control of stable processes. There, the AOC derived for the IPDT model and applied to the control of a stable thermal process with slow and fast modes gives transient responses close to the time-optimal ones. In contrast to relay time-optimal processes, however, the AOC gives the possibility of achieving relatively smooth input and output shapes with limited values of excessive controller effort and output wobbling. The transient responses in Fig. 15 and Fig. 16 refute the old dogma that the use of derivative terms is not suitable for noisy processes. Quite the opposite, in an attempt to smooth out noise, it is necessary to increase the time constants, or possibly the order of low-pass filters used. The controllers with higher order derivatives are then the only way to prevent the deterioration of process dynamics. In addition to excellent input and output process curves, AOC also provides information about the external disturbance. From the reconstructed disturbance curve in Fig. 16, it is possible to clearly identify the change in the power supply of the cooling fan. The slow changes in the reconstructed disturbance correspond to the slow mode of heat transfer by conduction.

Although SP has a reputation as the most effective structure for controlling time-delayed processes, its more detailed examination in the presence of measurement noise undermines this reputation. More precisely, doubts arise when it comes to the application based on controllers for integral and unstable processes. Integral models do not have observable output disturbances. Therefore, the Smith predictor for the reconstruction and

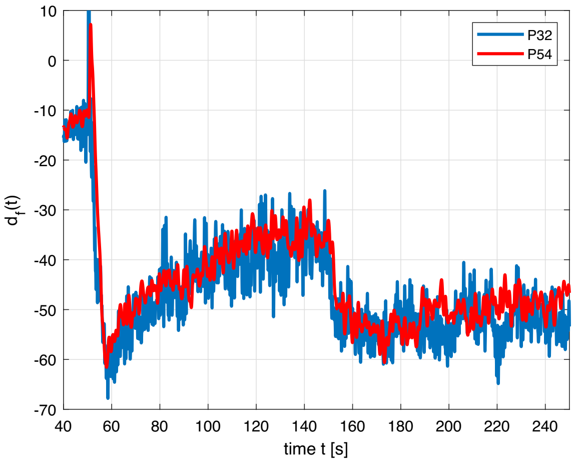

Figure 16. Reconstructed disturbance $d_f(t)$ of the thermal process with AOC controllers P32 (tuned with $T_e = T_{dp}, r = m-1, n = m+2$) and P54 (with $T_e = 3T_{dp}, r = m-1, n = m+2$): setpoint step at time $t = 50$s from $w_0 = 28^o$C to $w_1 = 32^o$C followed by an input (load) disturbance at $t = 150$s produced by a setpoint step change of a cooling fan input from 5 to 15.

compensation of output disturbances cannot be directly applied to them. And that is also the reason why the review[49] emphasized that the so-called filtered Smith predictor, in which the feasibility of the control structure used is achieved by excluding the reconstructed disturbance signal from the scheme, should use a different name. Its functionality is essentially different: it is just a controller without explicit disturbance reconstruction. In the case of ASP, the unobservability of the output disturbance is manifested in a different way. We formulate it in the following theorem:

Theorem 1 (**Ambiguity of ASP steady-states**) ASP based on IPDT model does not guarantee a unique steady state $y = w$.

Proof The ASP ambiguity is caused by the fact that two marginally stable blocks (integrators) are used in the reconstruction of the disturbance. If external interference (for example, the influence of noise) disrupts the ideal integration of signals, the ASP can no longer ensure the accuracy of setpoint tracking in steady states. In steady state it is not needed to consider delays and saturation. Thus, the block diagram (see Fig. 3 used with $a_0 = 0$) can be simplified. All derivatives must be equal to 0 and all integrator inputs must be equal to 0. If we admit that the output of the integrator K_{sp}/s is a nonzero value Δ then the output y of the process K_s/s must be $y = w - \Delta$. Then, the integrator $(k_4 s + k_i)/(K_{sp}s)$ will have zero input and output $-d_i$ needed to eliminate the disturbance d_i and keep 0 at input of the process K_s/s. However, this proves that the output $y = w - \Delta$ can end up with a steady state error Δ.

Thus, all the simulation problems above can be explained by the fact that ASP does not have a single steady state $y = w$. If we want to show what impact this has on its use, we can recall another similar case from the control of stepper motors. Their use in applications is interesting because of the possibility of working without angle sensors. Therefore, they are frequently used in practice despite their possible permanent control error. However, they require an adequate approach. Reducing the amplitude of the noise with the use of low-pass filters and appropriate ASP settings could be useful. However, it may happen that with such a setup, ASP will no longer be more advantageous than alternative solutions. In the specific case of controlling a stable thermal process, we have shown only that simplifying the entire design by using ultralocal integral models may be disadvantageous because of the unobservability of the output disturbance of the integral models. However, there still exists a simpler option for the SP design based on the use of a stable local model.

Theorem 1 together with the simulations in Fig. 9, Fig. 11 and Fig. 14 point out the consequences of long-term ignoring the problem of unobservable output disturbance of integral models for the design of various SP modifications. We will leave a more detailed examination of the question of to what extent it is possible to achieve ASP functionality by tuning it depending on the measurement noise level to other publications.

Conclusions and future work

The work showed a new design of automatic offset controllers (AOC) based on an IPDT system model. The design of a stabilizing controller (SC) uses an approximation of the loop-delay inversion by a finite number of terms of its development into a series. With respect to the stability and performance of the closed-loop, it is re-calculated by the multiple real dominant pole (MRDP) method. When supplementing the SC with a full disturbance observer (DOB) of the input disturbance, in the nominal case, there is no need to change the parameters of the stabilizing circuit. This is the first advantage compared to the use of a simplified observer in the design of ARC and PID controllers. To ensure causality and achieve an appropriate measurement noise attenuation, it is necessary to supplement the DOB and the SC with low-pass filters of sufficiently high orders. Compared to ARC and PID controllers, smoother and faster transient responses can be achieved.

In comparative experiments with an IPDT, an unstable and stable thermal process, the design based on AOC was alternated with a modified Åström-Smith predictor (ASP). AOC proved to be advantageous in terms of universality in achieving high performance and robustness, but also with regard to other aspects such as measurement noise attenuation, impact of control signal saturation, or usability for fault diagnostics. The differences between AOC and ASP in terms of noise attenuation and excessive controller effort in Table 4 range up to $10^3 - 10^4$. Hence, a realistic comparison was only possible after supplementing ASP[4] with a low-pass filter to attenuate the high-frequency measurement noise. Although such an extended ASP can even be more advantageous, it holds only at low levels of measurement noise. Higher noise amplitudes in the ASP lead to permanent control error and even instability. Problems with achieving the desired steady states appeared to be a consequence of the use of marginally stable terms in the DOB and the resulting inability of ASP to provide a single steady state. The contribution of the article can thus be considered as the design of two new alternative approaches to the control of time-delayed systems with dominant first-order dynamics under the possible impact of measurement noise and control saturation constraints.

One of the crucial moments in comparing AOC and ASP is the fact that, while AOC can be proposed for systems approximated by double integrator plus dead-time models (see[22]), ASP for such models is not known. However, there exist many other modifications inspired by the SP scheme (such as, for example,[61–63]), or even generalized PID and double-loop controllers[64,65], which makes the evaluation of other approaches to the given issue a more extensive, but also more interesting problem. However, such a comparison and evaluation require a separate contribution.

Future tasks related to the development of entire families of controllers with possibly HO derivatives include the design of modifications necessary to eliminate performance imperfections caused by the difference between the IPDT model used and the dynamics of the unstable controlled process, comparison with other alternatives to control difficult-to-manage processes[4,50,66–74], and automation of the selection of the optimal solution from among many existing alternatives[75].

Data Availability

The programs and datasets generated and analyzed during the current study are available in the Zenodo repository, https://www.doi.org/10.5281/zenodo.17364578.

Received: 3 October 2025; Accepted: 6 February 2026
Published online: 11 February 2026

References

1. Smith, O. Closser control of loops with dead time. *Chem. Eng. Prog.* **53**, 217–219 (1957).
2. Huba, M., Bisták, P., Vrančić, D. & Zakova, K. Asymmetries in the disturbance compensation methods for the stable and unstable first order plants. *Symmetry***2020, 12, 1595** (2020).
3. Åström, K. J., Hang, C. C. & Lim, B. C. A new Smith predictor for controlling a process with an integrator and long dead-time. *IEEE Trans. Autom. Control* **39**(2), 343–345 (1994).
4. Hölzl, S. L. The case for the Smith-Åström predictor. *J. Process Control* **128**, 103026 (2023).
5. Watanabe, K. & Ito, M. A process-model control for linear systems with delay. *IEEE Trans. AC* **26**(6), 1261–1266 (1981).
6. W. Zhang, Y. S. & Xu, X. Two degree-of-freedom smith predictor for processes with time delay. *Automatica* 1279–1282 (1998).
7. Mataušek, M. & Micič, A. A modified Smith predictor for controlling a process with an integrator and long dead-time. *IEEE Trans. Autom. Control* **41**, 1199–1203 (1996).
8. Garcia, P. & Albertos, P. A new dead-time compensator to control stable and integrating processes with long dead-time. *Automatica* **44**(4), 1062–1071 (2008).
9. Garcia, P. & Albertos, P. Robust tuning of a generalized predictor-based controller for integrating and unstable systems with long time-delay. *J. Process Control* **23**(8), 1205–1216 (2013).
10. Lima, T. A., de Almeida Filho, M. P., Torrico, B. C., Nogueira, F. G. & Correia, W. B. A practical solution for the control of time-delayed and delay-free systems with saturating actuators. *European Journal of Control* (2019).
11. Badavath, A. & Hote, Y. V. Internal Model Control-Based Design of PID, PIDA, and PID-II Controllers for Networked Control Systems. *IETE Tech. Rev.* **0**, 1–24 (2025).
12. Mejia, C., Salazar, E. & Camacho, O. A comparative experimental evaluation of various smith predictor approaches for a thermal process with large dead time. *Alex. Eng. J.* **61**, 9377–9394 (2022).
13. Alarcón, J. et al. The impact of delays in actuators and transmitters on control system performance. In *2024 7th IEEE Biennial Congress of Argentina, ARGENCON 2024*, 2024 7th IEEE Biennial Congress of Argentina, ARGENCON 2024 (Institute of Electrical and Electronics Engineers Inc., 2024).
14. Vega, S. et al. Comparative analysis and evaluation of fractional order controllers for systems with variable delay. In *2024 IEEE International Conference on Automation/XXVI Congress of the Chilean Association of Automatic Control (ICA-ACCA)*, 1–6 (2024).
15. Wang, L. Disturbance observer-based PID control with constraints: Design and robustness analysis. *IFAC-PapersOnLine* (2024). 4th IFAC Conference on Advances in Proportional-Integral-Derivative Control PID (2024).
16. Wang, Z. Research on dead-time compensation algorithm for PMSM drive systems based on stacked LSTM. In Jiang, Y. (ed.) *Third International Conference on Mechatronics and Mechanical Engineering (ICMME 2024)*, vol. 13220, 132201V. International Society for Optics and Photonics (SPIE, 2024).
17. Wang, S. et al. A fractional-order active disturbance rejection control for permanent magnet synchronous motor position servo system. *Asian Journal of Control* **26**, 3137–3147 (2024).
18. Wang, H., Zuo, Y., Zhao, C. & Lee, C. H. T. Developing a Unified Framework for PMSM Speed Regulation: Active Disturbance Rejection Control via Generalized PI Control. *World Electr. Veh. J.* **16** (2025).
19. Bistak, P., Belai, I., Belai, I., Jr., Vrancic, D. & Huba, M. Application of a Fractional Order PI Controller for a Speed Servo. *Symmetry*, **16**(11), 1543 (2024).
20. Huba, M., Bistak, P., Skrinarova, J. & Vrancic, D. Performance Portrait Method: Robust Design of Predictive Integral Controller. *Biomimetics* **10**(2), 74 (2025).
21. Liu, D., Baldi, S., Yang, K. & Astolfi, A. Equipping vehicular platoons with partial-state immersion and invariance adaptation. *IEEE Trans. Control Syst. Technol.* **33**, 2378–2392 (2025).

22. Huba, M., Bistak, P., Vrancic, D. & Sun, M. PID versus Model-Based Control for the Double Integrator Plus Dead-Time: Noise Attenuation and Robustness Aspects. *Mathematics* **13**(4), 664 (2025).
23. Luenberger, D. Observers for multivariable systems. *IEEE Trans. Autom. Control* **11**(2), 190–197 (1966).
24. Brogan, W. L. *Modern control theory* (Pearson Education India, 1991).
25. Ackermann, J. *Abtastregelung* (Springer, Berlin, 1988).
26. Ogata, K. *Modern Control Engineering. 3rd Ed.* (Marcel Dekker, 1997).
27. Huba, M. Disturbance Observer in PID Controllers for First-Order Time-Delayed Systems. In *13th IFAC Symposium Advances in Control Education* (Hamburg Bergedorf, Germany, 2022).
28. Huba, M., Bisták, P. & Vrančić, D. Saturated controllers for integrator plus dead-time model. In *International Conference on Electrical Drives and Power Electronics. 12th Joint Croatia-Slovakia Conference EDPE* (Dubrovnik, Croatia, 2025).
29. Huba, M. & Vrančič, D. Comparing filtered PI, PID and PIDD2 control for the FOTD plants. *IFAC-PapersOnLine* **51**, 954–959 (2018).
30. Huba, M., Bisták, P. & Vrančić, D. Series PID Control with Higher-Order Derivatives for Processes Approximated by IPDT Models. *IEEE TASE* **21**, 4406–4418 (2024).
31. Hu, Y. & Gao, Z. On mason reset based control and the scale of integration. *Control Theory Technol.* 104426 (2025).
32. Han, J. From PID to Active Disturbance Rejection Control. *Industrial Electronics, IEEE Transactions on* **56**, 900–906 (2009).
33. Radke, A. & Gao, Z. A survey of state and disturbance observers for practitioners. In *American Control Conference, 2006* (2006).
34. Gao, Z. On the centrality of disturbance rejection in automatic control. *ISA Trans.* **53**, 850–857 (2014).
35. Stankovic, M., Ting, H. & Madonski, R. From PID to ADRC and back: Expressing error-based active disturbance rejection control schemes as standard industrial 1DOF and 2DOF controllers. *Asian J. Control* 1–11 (2024).
36. Fliess, M. & Join, C. Model-free control. *Int. J. Control* **86**, 2228–2252 (2013).
37. Jiang, N., Zhang, S., Xu, J. & Zhang, D. Model-free control of flexible manipulator based on intrinsic design. *IEEE/ASME Trans. Mechatron.* **26**(5), 2641-2652, (2021).
38. He, D., Wang, H., Tian, Y. & Precup, R.-E. Model-free global sliding mode control using adaptive fuzzy system under constrained input amplitude and rate for mechatronic systems subject to mismatched disturbances. *Inf. Sci.* **697**, 121769 (2025).
39. Roman, R.-C., Precup, R.-E. & Petriu, E. Hybrid data-driven fuzzy active disturbance rejection control for tower crane systems. *Eur. J. Control* **58** (2020).
40. Roman, R.-C., Precup, R.-E. & Petriu, E. Higher-Order Model-Free Control Tuned by Fictitious Reference Iterative Tuning for 3D Cranes. *Rom. J. Inf. Sci. Technol.* **28**, 89–102 (2025).
41. Rigatos, G., Zouari, F. & Siano, P. A flatness-based adaptive fuzzy control method for an endogenous economic growth model. *Artif. Intell. Auton. Syst.* **2** (2025).
42. Huba, M., Bistak, P. & Vrancic, D. Gain-Scheduled Disturbance Observer-Based Saturated Controllers for Non-Linear First-Order System. *Appl. Sci.* **15**, 2812 (2025).
43. Ziegler, J. G. & Nichols, N. B. Optimum settings for automatic controllers. *Trans. ASME* 759–768 (1942).
44. O'Dwyer, A. *Handbook of PI and PID controller tuning rules. 3rd Ed.* (Imperial College Press, 2009).
45. Skogestad, S. Simple analytic rules for model reduction and PID controller tuning. *J. Process Control* **13**, 291–309 (2003).
46. Morari, M. & Zafiriou, E. *Robust Process Control* (Prentice Hall, Englewood Cliffs, N. Jersey, 1989).
47. Kumar, M. & Hote, Y. V. PIDD2 Controller Design Based on Internal Model Control Approach for a Non-Ideal DC-DC Boost Converter. In *2021 IEEE Texas Power and Energy Conference, TPEC 2021* (2021).
48. Kumar, M., Hote, Y. V. & Sikander, A. A novel cascaded CDM-IMC based PIDA controller design and its application. In *2023 IEEE IAS Global Conference on Renewable Energy and Hydrogen Technologies (GlobConHT)*, 1–7 (2023).
49. Huba, M., Bisták, P., Vrančić, D. & Zakova, K. Dead-Time Compensation for the First-Order Dead-Time Processes: Towards a Broader Overview. *Mathematics* **9**, 1519 (2021).
50. Hölzl, S. L. Globally asymptotically stable control of integrators with long dead time in the presence of actuator constraints. In *2024 IEEE 63rd Conference on Decision and Control (CDC)*, 5789–5794 (2024).
51. Normey-Rico, J. & Camacho, E. Unified approach for robust dead-time compensator design. *J. Process Control* **19**(1), 38–47 (2009).
52. Torrico, B. C., Cavalcante, M. U., Braga, A. P. S., Normey-Rico, J. E. & Albuquerque, A. A. M. Simple tuning rules for dead-time compensation of stable, integrative, and unstable first-order dead-time processes. *Ind. Eng. Chem. Res.* **52**, 11646–11654 (2013).
53. Skogestad, C. G. S. Should we forget the Smith predictor? *IFAC-PapersOnLine* **51**, 769 – 774. 3rd IFAC Conference on Advances in Proportional-Integral-Derivative Control PID 2018. (2018)
54. Rossiter, A., Serbezov, A., Visioli, A., Zakova, K. & Huba, M. A survey of international views on a first course in systems and control for engineering undergraduates. *IFAC J. Syst. Control* **13**, 100092 (2020).
55. da Silva, L. R., Flesch, R. C. & Normey-Rico, J. E. Analysis of anti-windup techniques in pid control of processes with measurement noise. *IFAC-PapersOnLine* **51**, 948 – 953. 3rd IFAC Conference on Advances in Proportional-Integral-Derivative Control PID 2018. (2018).
56. Grimholt, C. & Skogestad, S. Optimal PID control of double integrating processes. *IFAC-PapersOnLine* **49**, 127–132 (2016).
57. Huba, M., Bisták, P. & Huba, T. Filtered PI and PID control of an Arduino based thermal plant. In *14th IFAC PDeS* (Lednice, Brno, Czech Rep., 2016).
58. Huba, M., Bisták, P., Huba, T. & Vrančič, D. Comparing filtered PID and Smith Predictor Control of a Thermal Plant. In *19th International Conference on Electrical Drives and Power Electronics (EDPE)*, 324-329, (Dubrovnik, Croatia, 2017).
59. Huba, M. Improving noise attenuation in modified filtered Smith predictor. In *Cybernetics & Informatics, IEEE 30th International Conference on* (Velke Karlovice, Czech Rep., 2020).
60. Huba, M., Chamraz, S., Bisták, P. & Vrančić, D. Making the PI and PID Controller Tuning Inspired by Ziegler and Nichols Precise and Reliable. *Sensors* **18**, 6157 (2021).
61. Lu, X., Yang, Y., Wang, Q. & Zheng, W. A double two-degree-of-freedom control scheme for improved control of unstable delay processes. *J. Process Control* **15**, 605–614 (2005).
62. Rao, A. S., Rao, V. S. R. & Chidambaram, M. Set point weighted modified Smith predictor for integrating and double integrating processes with time delay. *ISA Trans.* **46**(1), 59–71 (2007).
63. Tavakoli, S. Enhanced control of double integrating plus time delay systems using a two-degree-of-freedom Smith predictor. *Electron. Lett.* **60**, e13309 (2024).
64. Kumar, D., Aryan, P. & Raja, G. L. Decoupled double-loop FOIMC-PD control architecture for double integral with dead time processes. *Canadian J. Chem. Eng.* **100**, 3691–3703 (2022).
65. Li, Y., Bi, J., Han, W. & Tan, W. Tuning of PID/PIDD2 controllers for integrating processes with robustness specification. *ISA Trans.* **140**, 224–236 (2023).
66. Campregher, F., Milanesi, M., Schiavo, M. & Visioli, A. Generalized Haalman tuning of PIDA controllers. *IFAC-PapersOnLine* **58**, 406–411. 4th IFAC Conference on Advances in Proportional-Integral-Derivate Control PID 2024. (2024).
67. Li, Z., Zhang, H. & Tan, W. Comparison of PI/PID/PIDD2 controllers for higher order processes. *IFAC-PapersOnLine* **58**, 340–345. 4th IFAC Conference on Advances in Proportional-Integral-Derivate Control PID 2024. (2024)
68. Milanesi, M., Visioli, A. & Chen, Y. From PID to PIDD2α: Performance improvement with a fractional double derivative action. *Annu. Rev. Control.* **60**, 101006 (2025).
69. Kumar, V. & Hote, Y. V. New approach of series-PID controller design based on modern control theory: Simulations and real-time validation. *IFAC J. Syst. Control* 100295 (2025).

70. Serdar Ekinci, D. I., Gider, V., Abualigah, L., Bajaj, M. & Zaitsev, I. Efficient control strategy for electric furnace temperature regulation using quadratic interpolation optimization. *Sci. Rep.* **15** (2025).
71. Ekinci, S., Izci, D., Jabari, M., Bajaj, M. & Rubanenko, O. Design of a novel and robust 2-DOF PIDA controller based on enzyme action optimizer for ball position regulation in magnetic levitation systems. *Sci. Rep.* **15** (2025).
72. Sevcik, J., Chamraz, S. & Zakova, K. Compensator with weighted input parameters for automatic ball-balancing mechatronic system. *Electronics* **14**(9), 1695 (2025).
73. Veinović, S., Stojić, D. & Ivanović, L. Optimized PIDD2 controller for AVR systems regarding robustness. *Int. J. Electr. Power Energy Syst.* **145**, 108646 (2023).
74. Boskovic, M. C., Sekara, T. B., Stojic, D. M. & Rapaic, M. R. Novel tuning rules for PIDC controllers in automatic voltage regulation systems under constraints on robustness and sensitivity to measurement noise. *Int. J. Electr. Power Energy Syst.* **157**, 109791 (2024).
75. Huba, M., Skrinarova, J., Vranic, D. & Bistak, P. New Generation of Higher-Order Controllers: Leveraging PID Control for Improved System Response. *IEEE Access* **13**, 164567–164591 (2025).

Author contributions
Writing-original draft preparation, M.H. and D.V. Experiments and simulations, M.H. Figures M.H., D.V., and P.B. Editing, D.V., P.B., and M.H. Project administration, M.H., D.V., and P.B. All authors have read and agreed to the published version of the manuscript.

Funding
This research was supported in part by the following grants: Grant No. 1/0821/25 financed by the Scientific Grant Agency of the Ministry of Education, Research, Development and Youth of the Slovak Republic (VEGA); Grant No. APVV-24-0390 financed by the Slovak Research and Development Agency (APVV); Research Program P2-0001 (Systems and Control) financed by the Slovenian Research and Innovation Agency; Grant Agreement No 101007175 (project REACTT) financed by Clean Hydrogen Partnership (EU Horizon 2020).

Declarations

Competing interests
The authors declare no competing interests.

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