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PID$_n^m$ Control for IPDT Plants. Part 1: Disturbance Response

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***Abstract*—The paper introduces generalized PID control with the derivative action degrees $m \in [0, 4]$ and with $n \geq m$th order series binomial filters. Together with the traditional PI ($m = 0$) and PID control ($m = 1$), this setup covers also solutions with higher order derivatives denoted previously as the PIDD^2, or PIDD^{2}D^3 controllers and remains open to more complex combinations. The novel contribution includes explicit tuning formulas for the integral-plus-dead-time (IPDT) plant models with $m = 3$ and $m = 4$ as well as the systemic approach to design of such a long sequence of controllers. Derived by the multiple real dominant pole method they enable to significantly reduce the measurement noise impact by simultaneously speeding up transients and increasing the closed loop robustness against plant uncertainties. With respect to the necessary extent of the problem description, solution and of the corresponding performance evaluation, the contribution has been split into two parts. The first part focuses on the disturbance response evaluation by simulation. This is limited to some degree by the numerical problems of the available Simulink solvers. To overcome the simulation problems, the second part dealing with design and evaluation of the setpoint responses will consider also real time experiments.**

***Index Terms*—Filtration, multiple real dominant pole method, derivative action**

I. Introduction

Many dynamical processes to be controlled by the PID control may be well approximated by an integrator plus dead time (IPDT) system (see, for example, [1]–[5]). The simple "ultra local" IPDT models are important especially in a robust design of a broad class of more complex and possibly nonlinear systems, when the model simplifications make the controller design executable. This simplicity may be preferred also in demonstration of novel approaches, where it enables to keep transparency of the main ideas. In this way, the pole assignment control design may be further simplified by considering just the dominant closed loop poles [6], [7], which may shadow theoretically infinitely many closed loop poles of time delayed closed loop systems. This represents the core idea of the multiple real dominant pole (MRDP) method which may be considered as one of the oldest analytical tools for the controller tuning [8] and still attracts attention of a broad research community. It has been successfully applied to design of PD, PI, or PID control [9]–[11]. Application of such controllers in a rough industrial environment frequently requires noise attenuation filters, which are unavoidable in controllers with derivative action [12]–[15].

This contribution develops an idea presented in [5], [16]–[19] that the time constant of a filter with an arbitrarily chosen order nth may be specified to keep a constant position of the dominant closed loop poles for any chosen n. From such a requirement of a fixed dominant closed loop pole position guaranteeing a nearly constant closed loop dynamics, a delay equivalence may be derived enabling for any filter order n a simple integrated tuning of the controller and of the filter parametrized with the plant model dead time T_{dm} and an equivalent filter delay T_e (specified usually as a fraction of T_{dm}). In the situation, when the standard textbooks as [20], [21] state that the derivative action is usually swished off because it is the most difficult to tune, this simple integrated tuning procedure made it possible to deal easily even with controllers using higher order derivative action, where the design is even more complicated by the fact that such controllers can not be used without properly designed filters.. The performance improvements may then be achieved also in dealing with the time-delayed systems [22]–[24]. Since the applied approach offers generalization to higher orders m and n, the study carried out in [5], [18], [19], [25], [26] for $m \leq 2$ is going to be extended for IPDT systems with a broader range of the derivative action degrees $m \in [0, 4]$. Thereby, the controller denotation has to be simplified to PID$_n^m$. It will be investigated with the aim to find the rools and limits in improving the closed loop performance and robustness. Since the aim to consider such a broad spectrum of derivative orders and the increasing complexity of the PID$_n^m$ controllers open numerous problems the topic has beens split into two standard conference papers. These aim to show that by a filter choice carried out together with the controller design and by their integrated tuning the resulting undesired control activity may be significantly reduced by simultaneously boosting the closed loop performance. The optimal tuning rules for the broad family of filtered PID$_n^m$ controllers equipped with the nth order series binomial filters (11) will be embodied into a uniform approach based on the MRDP method. Thereby, the discussion opens dominant question important for filtering in control applications: *may the use of higher order derivative actions increase the speed of transients with simultaneously decreasing the adverse noise impact without threating the system robustness?* In our best knowledge, the above question will be carefully formulated and sufficiently answered in such an extent for the first time in this work.

The first part of these contributions is structured as follows. Section II introduces the time and shape related performance measures used in the transients evaluation. Section III deals with the PID^m controllers for the IPDT plants and with filtered PID$_n^m$ controllers for purely integral plants tuned by the MRDP method. Results of these two approaches are then generalized in Section IV into an integrated filter & controller tuning by the equivalent filter time delay T_e. Section V deals with evaluation of the proposed PID$_n^m$ controllers tuning by simulation in Matlab/Simulink. The achieved results are discussed in Section VI and summarized in Conclusions.

The 2nd part will then deal also with the setpoint responses and extend the loop evaluation by real time experiments.

II. Performance measures

Since the setpoint step responses do not depend just on a feedback and may be improved by an appropriate feedforward term, due to the limited space, the paper will focus on the disturbance responses with the setpoint $w = 0$. The responses will be evaluated in terms of the absolute integral error

$$IAE_d = \int_0^\infty |e(t)|\, dt \; ; \;\; e = w - y \tag{1}$$

As shown in the Theorem 1 in [27], an ideal disturbance response of integral first order plants has shape of a one-pulse (1P) transition consisting of two monotonic intervals separated by an extreme point $y_m \notin (y_0, y_\infty)$. This is lying out of the interval specified by the initial and the final values y_0 and y_∞. Deviations of the output $y(t)$ from an ideal one-pulse (1P) behavior may be measured in terms of a relative total variance of the output denoted as $TV_1(y)$

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

Similarly, deviations of the plant input $u(t)$ from an ideal one-pulse (1P) step response [27] should be constrained in terms of $TV_1(u)$ measure. Basically, for control of the IPDT plants it might be meaningful to consider also input with a higher number of control pulses, but such a situation represents already a more advanced and seldom used option.

III. PID^m and PID$_n^m$ controllers for the IPDT plant tuned by the MRDP method

It is always advantageous to deal with analytical methods for the controller tuning [28]. For an IPDT plant model

$$S(s) = \frac{Y(s)}{U(s)} = S_0(s)e^{-T_{dm}s}; \; S_0(s) = \frac{K_{sm}}{s} \tag{3}$$

with a model gain K_{sm} and a model dead time T_{dm} it is possible by a modification of the multiple real dominant pole (MRDP) method. Where appropriate, the model index "m" will be omitted, which corresponds to the model parameters equal to the plant parameters K_s and T_d.

A. *PID^m control*

A generalization of the PID control with possible derivative terms up to an integer degree "m" may be proposed as

$$\begin{aligned} C_m(s) &= K_c\left[\frac{1+T_is}{T_is} + T_{D1}s + ... + T_{Dm}s^m\right] = \\ &= K_c + \frac{K_i}{s} + K_{D1}s + ... + K_{Dm}s^m; \; m = 0,1,2,... \end{aligned} \tag{4}$$

For $m = 0$ it corresponds to PI, for $m = 1$ to PID, for $m = 2$ to PIDD^2 control, etc. Obviously, due to the improper $C_m(s)$, for $m > 0$ the controller may not be implemented, but this problem will be treated later. Nominally, the input-disturbance-to-output relation corresponding to $C_m(s)$ and $S(s)$ is described by

$$\begin{aligned} F_{dy}(s) &= \frac{Y(s)}{D_i(s)} = \\ &= \frac{K_sT_is}{T_is^2e^{T_ds} + K_cK_s[1 + sT_i(1 + T_{D1}s + ... + T_{Dm}s^m)]} \end{aligned} \tag{5}$$

It yields the characteristics quasi-polynomial

$$P(s) = T_is^2e^{T_ds} + K_cK_s[1 + sT_i(1 + T_{D1}s + ... + T_{Dm}s^m)] \tag{6}$$

A multiple real dominant pole (MRDP) s_o of the characteristic quasi-polynomial $P(s)$ (6) may be calculated by a generalization of [10]. From the last row of the equation system

$$\begin{aligned} &[P(s)]_{s=s_o} = 0 \\ &\left[\frac{dP(s)}{ds}\right]_{s=s_o} = 0 \\ &\vdots \\ &\left[\frac{d^{m+2}P(s)}{ds^{m+2}}\right]_{s=s_o} = e^{T_ds_o}T_iT_d^m. \\ &.\left[T_d^2s_o^2 + 2(m+2)T_ds_o + (m+2)(m+1)\right] = 0 \end{aligned} \tag{7}$$

in which the second term of $P(s)$ vanishes, follows the wanted (dimensionless) root as

$$p_o = s_oT_d = \sqrt{m+2} - (m+2) \tag{8}$$

By substituting s_o into the first $m + 2$ equations of (7) it is then possible to derive the corresponding ("optimal") dimensionless controller parameters

$$\begin{aligned} K_o &= K_{co}K_sT_d \\ \tau_{io} &= \frac{T_{io}}{T_d} \\ \tau_{jo} &= \frac{T_{Dj}}{T_d^j}\; ; \;\; j = 1,2,...,m \end{aligned} \tag{9}$$

summarized for $m \in [0, 4]$ in Tab. I.

Reasons for introducing higher order derivative actions may be demonstrated by the optimal IAE_d values

$$IAE_d = T_i/K_c \tag{10}$$

TABLE I
OPTIMAL PID^m PARAMETERS, $m \in [0,4]$

$Parameter$	$m=0$	$m=1$	$m=2$	$m=3$	$m=4$
K_o	0.4612	0.78361	1.08268	1.37114	1.65330
τ_{io}	5.8284	3.73205	3.00000	2.61803	2.37980
τ_{1o}	0	0.26289	0.37500	0.43673	0.47525
τ_{2o}	0	0	0.04167	0.07492	0.09972
τ_{3o}	0	0	0	0.00474	0.01020
τ_{4o}	0	0	0	0	0.00042

TABLE II
OPTIMAL IAE_d VALUES CORRESPONDING TO INPUT DISTURBANCE STEPS FOR OPTIMAL PID^m PARAMETERS FROM TAB. I

-	$m=0$	$m=1$	$m=2$	$m=3$	$m=4$
$IAE_d/(K_sT_d^2)$	12.639	4.763	2.771	1.909	1.439

corresponding to an unit input disturbance step [27] (which may be derived under assumption of a not changing sign of the control error as $IAE_d = IE_d$), see Tab. II. From Tab II it is then clear that the IAE values decrease with increasing parameter m. One of the most important question which still remains is whether the improved disturbance rejection performance can be achieved with PID^m controllers in practice. This will be tested in the following subsection by generalizing the filtered solutions proposed in [18].

B. PID$_n^m$ controllers with nth order filters for the IPDT plant tuned by the MRDP method

Let us now consider a PID^m controller for an IPDT plant (3) combined with an nth order binomial filter

$$Q_n(s) = 1/(T_fs+1)^n\,;\; n = 1,\,2,\,... \tag{11}$$

For $n \geq m$ the open-loop transfer functions become

$$F_n^m(s) = \left[\frac{1+T_is}{T_is} + T_{D1}s + ... + T_{Dm}s^m\right]\frac{K_cK_se^{-T_ds}}{s(1+T_fs)^n} \tag{12}$$

For practical implementation, it is enough to consider $n=m$, since the resulting transfer function $R_n^m(s) = C_m(s)Q_n(s)$ becomes proper. However, as was demonstrated by simulation and by real time experiments [13], [16], [18], [29], the measurement (or quantization) noise impact is significantly lower for strictly proper $R_n^m(s)$.

One possible solution to get analytical results is to replace all n time constants T_f by an equivalent dead time T_e. However, since the model represents an integrating process (3), the half rule introduced by Skogestad [28] cannot be applied. As it was shown in [17], the delay equivalence $T_e \approx nT_f$ based on the average residence time [20] does not yield sufficiently precise results. It is therefore more appropriate to apply a delay equivalence based on an equal position of the dominant closed loop poles corresponding under considered PID^m control to particular type of delays.

C. Filtered PID$_n^m$ tuning

To derive such an equivalence, consider PID_n^m control of a loop with the plant $S_0(s)$ (3) and a delay introduced just by the filter (11). The closed loop transfer functions have now the characteristic polynomial

$$\begin{aligned}P(s) &= T_is^2(1+T_fs)^n + \\ &+ K_cK_s[1+T_is(1+T_{D1}s+...+T_{Dm}s^m)]\end{aligned} \tag{13}$$

A multiple real dominant pole s_n of $P(s)$ (13) follows from

$$\begin{aligned}&P^{m+2}(s_n) = (1+T_fs_n)^{n-m-2}n(n-1)...(n-m+1).\\ &.T_iT_f[(n+1)(n+2)T_f^2s_n^2 + 2(m+2)(n+1)T_fs_n + \\ &+(m+2)(m+1)]\end{aligned} \tag{14}$$

as

$$s_n = \frac{\sqrt{(m+2)(n+1)(n-m)} - (m+2)(n+1)}{(n+1)(n+2)T_f} \tag{15}$$

Since in the controller tuning several loop dead time elements may be replaced by their sum, the filter delay will now be represented by the equivalent dead time T_d denoted as T_e yielding according to (8) the same dominant pole position as (15). Such an equivalence based on a requirement of a constant position of the dominant closed loop poles $s_o = s_n$ formulated for $T_d = T_e$ and T_f yields equation

$$\begin{aligned}&\frac{\sqrt{m+2}-(m+2)}{T_e} = \\ &= \frac{\sqrt{(m+2)(n+1)(n-m)} - (m+2)(n+1)}{(n+1)(n+2)T_f}\end{aligned} \tag{16}$$

Solving (16) for T_f yields

$$T_f = \frac{\sqrt{(m+2)(n+1)(n-m)} - (m+2)(n+1)}{(n+1)(n+2)[\sqrt{m+2}-(m+2)]}T_e \tag{17}$$

A dead time T_e has the same impact on the dominant closed loop position as an equivalent time constant T_f of the n-th order filter (11). It enables a simplified tuning in loops containing simultaneously plant dead time and noise attenuation filters.

Remark 1 (Constant average residence time equivalence): Overview of the particular ratios T_f/T_e are given in Tab. III. It is to note that with the exception of the PID$_2^2$ controller do not correspond to an equivalence $T_e = nT_f$ following from the requirement of a constant average residence time [20].

TABLE III
EQUIVALENT TIME DELAYS RATIOS T_f/T_e, $m \in [0,4]$, $n \in [m,6]$

m	$n=1$	$n=2$	$n=3$	$n=4$	$n=5$	$n=6$
0	0.5690	0.3608	0.2647	0.2092	0.1729	0.1474
1	0.7887	0.3943	0.2800	0.2180	0.1787	0.1514
2	-	0.5000	0.3000	0.2279	0.1847	0.1555
3	-	-	0.3618	0.2412	0.1917	0.1599
4	-	-	-	0.2279	0.2012	0.1651

IV. TUNING BY THE EQUIVALENT TIME DELAY T_e

After identifying the parameters K_{sm} and T_{dm} and taking into account the level of required noise filtration, we choose the tuning parameter $T_e > 0$ specifying a new total loop dead time

$$T_d = T_{dm} + T_e \tag{18}$$

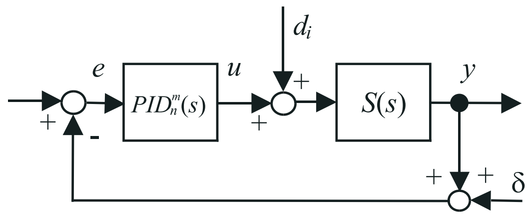

Fig. 1. Considered loop with the PID^m_n controller with an input disturbance d_i and a measurement noise δ

By substituting into (9), the controller tuning depends both on the plant parameters and on the tuning parameter T_e. Furthermore, for an arbitrarily chosen $n \geq m$ it offers the filter time constants T_f calculated according to (17). For a particular m, all these formulas keep a constant location of the dominant poles and guarantee a nearly equal closed loop dynamics. With an increasing m, the speed of transients for a constant T_d increases. It is also to note that if the controller tuning guarantees a monotonic error decay, the resulting IAE (10) does not change with n, just with T_d determining K_i. When taking into account the equivalence of filter delays $Q_n(s) \Leftrightarrow e^{-T_e s}$ and the PID^m_n controller tuning (9) based on the dead time T_d, after introducing $T_e = T_d - T_{dm}$ as a parameter characterizing an equivalent dead time spent on filtration, the tuning problem may be formulated as follows:

1) Choose an appropriate $T_e = T_d - T_{dm} > 0$ corresponding to a required filtration degree;
2) Specify the derivative degree m and the controller parameters (9) corresponding to T_d;
3) Choose the filter order n and by a delay equivalence $T_f = f(m, n, T_e)$ (17) specify the filter time constant T_f;
4) If the controller is implemented digitally using a short sampling period T_s, check, if the calculated T_f fulfills the requirement $T_f >> T_s$. If not, you should either decrease T_s, or n, which should still fulfill the condition $n \geq m$. In the worst case you may still use the non filtered PI control.
5) By experimentally evaluating the loop properties for different m and n, choose an optimal controller guaranteeing under the hardware constraints the optimal loop performance.

We may expect that by increasing n the sensitivity of the loop increases, but since also the noise attenuation increases [30], an optimum is usually at $n > m$, i.e. with a higher filter order than the derivative order, which corresponds to a full roll off.

V. Evaluation by simulation

A. Numerical issues

It is well known that in simulation of time delayed systems the available Matlab/Simulink solvers may exhibit numerical problems [31], [32]. Therefore, before evaluating the newly proposed controllers under realistic conditions with a measurement noise and possible plant uncertainties, it may be helpful to find the application limits due to the numerical

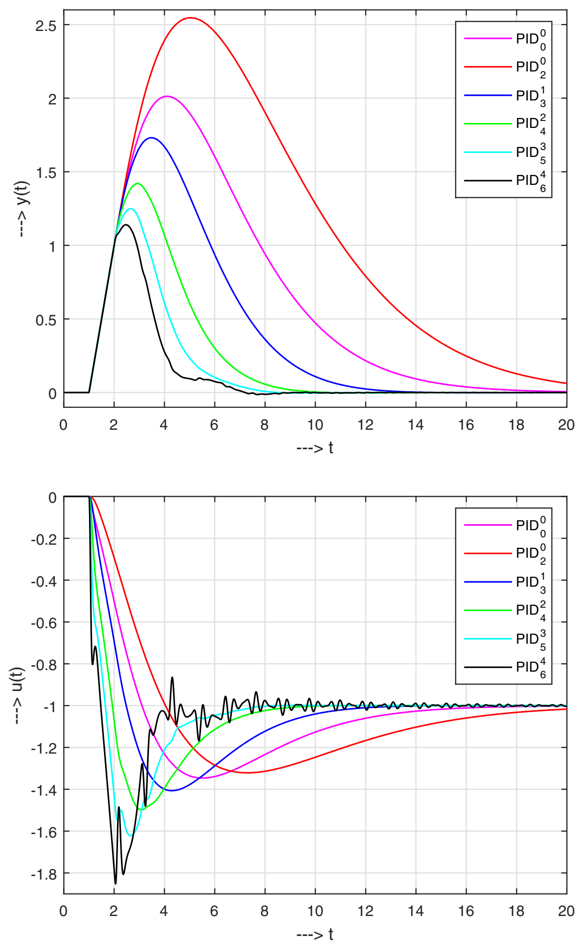

Fig. 2. Step responses of the PID^m_n controllers for $T_e = 0.3T_{dm}$, no measurement noise, $T_s = 0.001$,$K_s = K_{sm} = 1, T_{dm} = 1$

simulation issues. As, for example, apparent from the simulation results on disturbance steps for chosen controllers tuned with $T_e = 0.3T_d$ (Fig. 2), some deviations of the plant input and output signals may occur due to the numerical problems. In situations, where such imperfections dominate (as for PID^4_6 with $T_e \leq 0.3T_d$), the simulation results become useless. A more comprehensive picture of the loop numerical properties may be given by Fig. 3. Obviously, without an external noise the traditional PI controller (denoted now as PID^0_0) yields the lowest shape related deviations. All additional numerical operations necessary for the considered filtration and derivative actions manifest themselves as a source of numerical imperfections. For example, for PID^3_5 the numerical instability appears for $T_e = 0.2$, which is signalized by an apparently deviating sample. Similar situation, but with a significantly higher sample deviation, appeared also for PID^4_6, which has not been displayed.

B. Noise charateristics

Performance achievable by particular controllers has been demonstrated for $n = m + 2$, which (according to the previous works [16], [18]) enabled to optimally suppress

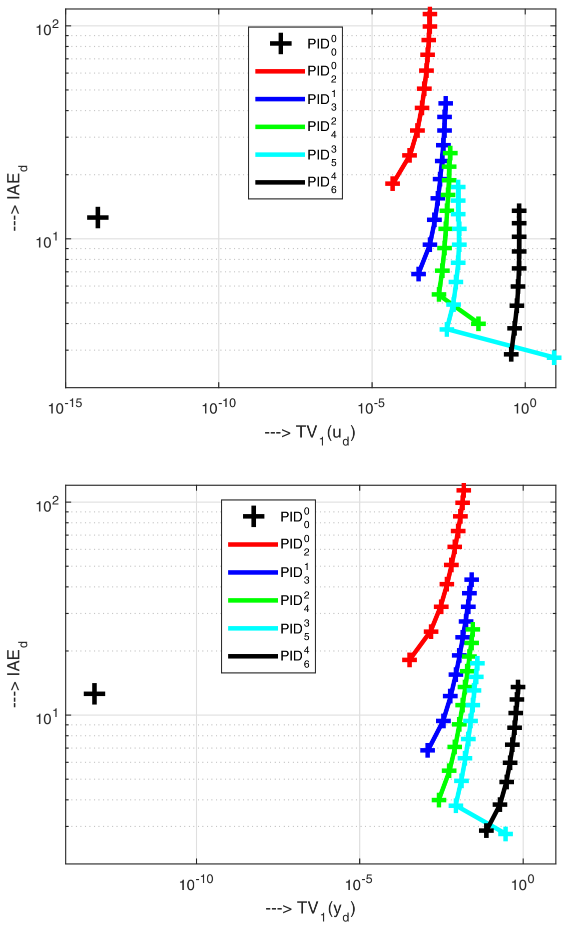

Fig. 3. IAE versus shape related deviations at the input and output for the plant (3) and different controllers, no noise, $T_s = 0.001$, $T_e \in [0.2T_{dm}, 2T_{dm}])$, for PID$^4_6$ $T_e \in [0.4T_{dm}, 2T_{dm}])$, $\Delta T_e = 0.2T_{dm}$, $T_{dm} = 1$

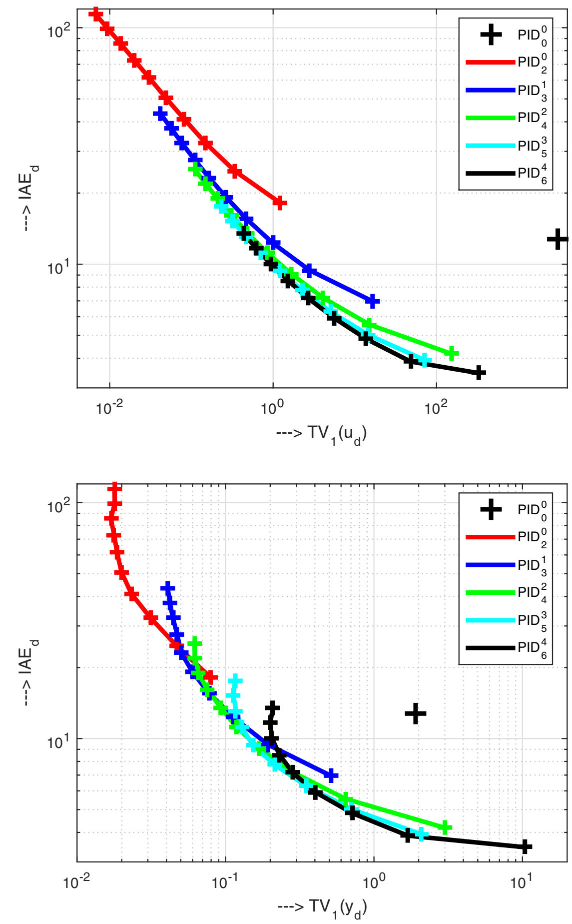

Fig. 4. IAE versus shape related deviations at the input and output for the plant (3) and different controllers, noise with the amplitudes $|\delta| \leq 0.1$ generated in Matlab/Simulink by the block "Uniform Random Number", $T_s = 0.001$, $T_e \in [0.2T_{dm}, 2T_{dm}])$, $\Delta T_e = 0.2T_{dm}$., for PID$^4_6$ $T_e \in [0.4T_{dm}, 2T_{dm}])$

the excessive noise-induced control effort at the controller output. For the same controllers, filters and plant as above, measurement noise has been generated in Matlab/Simulink by the block "Uniform Random Number" with sampling period $T_s = 0.001$ and amplitudes $|\delta| \leq 0.1$ (Fig. 1).

The IAE_d values corresponding to a noisy measurement are just slightly increased (vertical shift), which depends also on the length of the evaluated intervals ($t_{max} = 100$). Significant changes, however, appear in the performance measures (horizontal shifts). As the reference option, the (nonfiltered) PID$^0_0$ control (i.e. the PI control) is used showing relatively high $TV_1(u)$ and $TV_1(y)$ values.

There exist areas of the $TV_1(u)$ and $TV_1(y)$ values, where, under the same noise impact, PID$^1_3$ control gives faster and less distorted transients than the PID$^0_0$ and PID$^0_2$ control. The trend continues with PID$^2_4$, PID$^3_5$ and PID$^4_6$ control. The results stresses importance of the newly introduced solutions.

As already mentioned in [18], it is important to know that filtration performance dominantly depends on the applied sampling period. For the higher-order filters, tuned by the above given expressions, it holds that $nT_f \approx T_e$. In order to keep the ratio $T_f/T_s >> 1$ also for higher n, by keeping $T_e = const$, the sampling time should be $T_s << T_e/n$.

VI. Discussion

The numerical imperfections of the continuous time solvers, which determine reliability borders of this simulation based analysis, are supposed not to occur in real time experiments. Thus, the test of the newly proposed controllers in the real time situations may not only be interesting from the practical point of view, but also to understand the numerical issues appearing in simulation tests.

As shown, for example, in [5], the numerical imperfections of the continuous-time solvers applied to dead-time compensators may be avoided by developing the discrete-time controllers [26]. However, the improvement range is limited by the increased controller complexity. Though, it is to remember that such controllers are necessary with respect to their implementation by different embedded computers.

VII. Conclusions

The paper has demonstrated possibilities of designing controllers with higher order derivatives, which were enabled due to the progress in an integrated controller + filter tuning. Novel formulas for generalized optimal PID$_n^m$ controllers have been elaborated for $m \in [0, 4], n \geq m$. This sequence could be extended also to higher m. The loop performance has been evaluated by the integral of the absolute error (IAE) and by the relative total variance measures $TV_1(u)$ and $TV_1(y)$ representing excessive input and output changes and thus exhibiting impacts of the measurement noise and of other control imperfections at the plant input and output.

Similarly as in [5], [18], [19], [25], the controller parameters considered for the IPDT plant model (3) have been expressed as functions of the plant model dead time T_{dm}. This represents a key parameter influencing the speed of transient responses, used in discussions about the filtration intensity and in determining the filter time constants.

All the loop parameters and tuning formulas have been analytically determined by the MRDP method. Since the previous papers have shown that a significant performance improvement at the controller output may be achieved by keeping the filter order $n = m+2$ (i.e. higher than necessarily required for the controller implementation by 2, or for the full roll-off [13] by 1), all experiments have been carried out for PID$_{m+2}^m$ controllers. They may be summarized by the conclusions that the use of a higher order derivative action combined with a proper filtration may speed up transients and simultaneously decrease the adverse impact of the measurement noise.

Future research in this area will focus on use of alternative simulation tools, development of discrete time solutions [26], experiments on real plants and generalization to higher order plant models.

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