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Dynamic Modification of Reference Value for Reduction of Negative Effects of Actuator Constraints in Single-loop PID Control Systems

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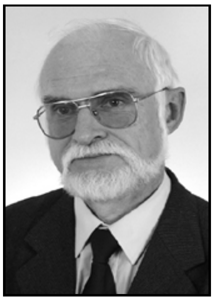

Działalność naukowa w zakresie identyfikacji, modelowania i sterowania procesów. Obszary badań aplikacyjnych: identyfikacja procesów przemysłowych, synteza algorytmów sterowania dla pneumatycznych układów pozycjonujących, hydraulicznych napędów wyporowych, zespołu napędu hybrydowego o 2 stopniach swobody, oraz pakiety oprogramowania dla identyfikacji układów dynamicznych (IDCAD), i modelowania, symulacji działania i sterowania procesów przemysłowych (PExSim).

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Abstract

A modification of the PID digital controller algorithm, based on the introduction of a virtual reference value that never exceed active constraints in the actuator output is presented and investigated. This idea derived from virtual modification of a control error can be used in digital control systems subjected by both: magnitude and rate constraints. The adaptation (to the actuator constraints) is performed by a special transformation of the control error, and is equivalent to introduction of a new, virtual reference value for the control system. Simulation results for dynamic, linear SISO systems are presented for comparison of the proposed approach with used techniques.

Keywords: Actuator constraints, digital PID control algorithm, wind-up.

Dynamiczna modyfikacja wartości zadanej w celu redukcji negatywnego wpływu ograniczeń elementu wykonawczego w jedno-obwodowych układach regulacji PID

Streszczenie

W pracy została przedstawiona modyfikacja działania cyfrowego algorytmu PID, oparta na wprowadzeniu dynamicznie zmiennej wartości zadanej, która nie prowadzi do przekroczenia ograniczeń możliwości elementu wykonawczego. Metoda jest oparta na dynamicznej modyfikacji odchyłki regulacji i może być stosowana w układach regulacji o ograniczeniach amplitudy i prędkości zmian zespołu wykonawczego. Adaptacja układu (do ograniczeń aktuatora) jest równoważna dynamicznej modyfikacji wartości zadanej. Wyniki badań symulacyjnych, porównujące efekty działania metody z innymi modyfikacjami stosowanymi dla algorytmu PID są zaprezentowane dla dwóch liniowych układów dynamicznych.

Słowa kluczowe: Cyfrowe algorytmy PID, ograniczenia elementu wykonawczego, wind-up.

1. Introduction

Constraints of an actuator have always involved many problems in a practical application of different control techniques. They have affected the basic controller action, prolonged control time and sometimes can induce oscillations that are hard to be dumped. Most known, widely commented and investigated, is a wind-up effect observed in the PID control. Commonly used, efficient PID algorithm suffers from two major drawbacks caused by the actuator output constraints: limitation of the differential action at significant variations of the control error and the prolonged integration of large control errors. The first effect, resulting in less efficient disturbance compensation, can be partly reduced by the introduction of inertia in the differential term, or the control algorithm modification, Fig. 1. The second effect is more embarrassing – the actuator output saturation yields the integration of the control error in a prolonged time interval. A practical approach to the reduction of this effect is either to form another shape of the reference value (e.g. a time ramp instead of step change), modifications of the integral action such as the conditional integration, reduced or even reversed integration. These methods have been introduced to digital PID controllers. Saturation of the actuator output has been considered by many authors [2, 3, 4, 10]. Observed deterioration of the control action, due to the limited actuator output has forced manufacturers of controllers to introduce some alternations of the basic digital PID algorithm, e.g., [1, 8]. The negative effect of the rate constraint in an actuator has also been noticed, and some countermeasures have been proposed [5].

2. Investigated control system and compared PID algorithms

The proposed approach can be used in digital control systems (not only the PID algorithm), with the same sampling interval Δ for all signals. Sampled values of signals are represented by the corresponding subscript $x_t \equiv x(t*\Delta)$. Considered control is used for a linear, SISO, dynamic discrete-time model of a plant

$$A(q^{-1})y_t = B(q^{-1})q^{-d}v_t \tag{1}$$

where y_t is a value of the plant output, v_t is a value of the actuator output and $d \geq 0$ is a discrete plant delay. Polynomials $A(q^{-1})$ and $B(q^{-1})$ determined by a shift operator q^{-1} ($q^{-p}x_r = x_{r-p}$)

$$\begin{aligned} A(q^{-1}) &= 1 + a_1 q^{-1} + \ldots + a_{nA} q^{-nA} \\ B(q^{-1}) &= 1 + b_1 q^{-1} + \ldots + b_{nB} q^{-nB} \end{aligned} \tag{2}$$

are proper and have real coefficients. The actuator output v_t in (1) is subjected to constraints: interval of admissible magnitudes

$$v_t \in U_M = < u_{low}, u_{high} > \tag{3}$$

and actuator output change rate (within one sampling interval)

$$\begin{aligned} & v_t - v_{t-1} = \sigma v_t \in \Delta U, \\ & \Delta U = \begin{cases} -\Delta * V_{neg} \leq \sigma v_t \leq s_{off-} \leq 0 \\ = 0 \Leftrightarrow \sigma v_t \in (s_{off-}, s_{off+}) \\ 0 \leq s_{off+} \leq \sigma v_t \leq \Delta * V_{pos} \end{cases} \end{aligned} \tag{4}$$

with extreme values V_{neg}, V_{pos}. The interval of minimum actuator variations (s_{off-}, s_{off+}) used for the chattering effect reduction does not alter the controller efficiency and will be not considered in further investigation. The proper controller design yields always reasonable surplus of the actuator output. A prompt reaction on a disturbance impact demands a small value of the sampling interval Δ, and then actuator can alter its output by only a fraction $\Delta*(V_{neg}$,or $V_{pos})$.

A general form of the PID-algorithm can be presented by the following equation

$$M(q^{-1})u_t = R(q^{-1})r_t - P(q^{-1})y_t \,. \tag{5}$$

The output signal u_t of the ideal digital I-PID algorithm contains three forms of the control error processing (expressed in continuous mode), Fig. 1: proportional to the control error, derivative action and integral term for the control error asymptotic removal

$$
\begin{aligned}
u_t &= G_{I-PID}(q^{-1})e_t = \frac{P(q^{-1})}{M(q^{-1})}e_t, \qquad e_t = r_t - y_t \\
G_{I-PID}(q^{-1}) &= k_P + k_I\frac{q^{-1}}{1-q^{-1}} + k_D(1-q^{-1}) = \\
&= K_P\left[1 + \frac{\Delta(1+q^{-1})}{2T_I(1-q^{-1})} + \frac{T_D}{\Delta}(1-q^{-1})\right]
\end{aligned} \tag{6}
$$

where K_P, T_D and T_I are optimized algorithm parameters for the defined sampling value Δ.

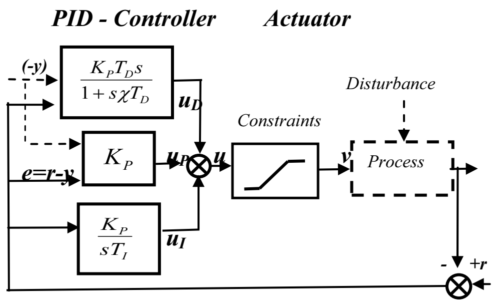

Fig. 1. Basic structure of a PID closed-loop control system
Rys. 1. Podstawowa struktura zamkniętego układu regulacji PID

The ideal difference operation at a short sampling interval Δ cannot be adequately processed by the actuator, hence a "real" form of the discrete-time derivation filtered by the first-order inertia χT_D [2, 10], Fig. 1, is used. The control algorithm with optimized χ is called R-PID, and is defined by

$$
\begin{aligned}
G_{R-PID}(q^{-1}) &= \frac{p_0 + p_1 q^{-1} + p_2 q^{-2}}{(1-q^{-1})(1-c_1 q^{-1})} = \frac{P(q^{-1})}{M(q^{-1})} \\
p_0 &= \frac{K}{1+T_V/\Delta}\left[1 + \frac{T_D+T_V}{\Delta} + \frac{\Delta+T_V}{2T_I}\right], \qquad T_V = \chi T_D\,. \\
p_1 &= \frac{K}{1+T_V/\Delta}\left[-1 + \frac{\Delta}{2T_I} - 2\frac{(T_D+T_V)}{\Delta}\right] \\
p_2 &= \frac{K}{1+T_V/\Delta}\left[\frac{T_D+T_V}{\Delta} - \frac{T_V}{2T_I}\right], \qquad c_1 = \frac{T_V}{T_V+\Delta}
\end{aligned} \tag{7}
$$

The PID control action can be expressed in the form of one transfer function (6) or (7) but industrial solutions [1, 8] prefer a distinct implementation, Fig. 1. It is more suitable for the controller actions separate tuning and has another advantage – the controller actions u_P, u_D or u_I can be separately on or off, and allow us the bump-less control switching between hand/automatic modes.

$$
R(q^{-1}) = \rho P(q^{-1}), \qquad \rho \in <0,\ 1>\,. \tag{8}
$$

where the polynomial $P(q^{-1})$ is defined by (7). The factor ρ creates a second degree of freedom (the first is a choice χT_D) in algorithm design. $R(q^{-1})$ in (7) is determined by the following formula

$$
\begin{aligned}
&R(q^{-1}) = r_0 + r_1 q^{-1} + r_2 q^{-2}, \\
&r_0 = \gamma\left(1 + \frac{\Delta+T_V}{2T_I}\right), \qquad r_1 = \gamma\left(-1 + \frac{\Delta}{2T_I}\right), \qquad r_2 = -\gamma\left(\frac{T_V}{2T_I}\right), \\
&\gamma = \frac{\rho K}{1+T_V/\Delta}.
\end{aligned} \tag{9}
$$

This type of algorithm, with optimized χ and ρ, will be called the TF-PID algorithm. All applied PID algorithms suffer from the above mentioned wind-up effect resulting from the prolonged integration of the control error. The most used technique for the wind-up is a conditional integration reduction, i.e. the suspension of the control error integration when the actuator output saturation is observed [4, 20]. Used conditions are different but usually the integration is suspended when the actuator output meets extreme values, and the process gain sign is equal to that of control error.

A more active attitude to the wind-up problem is presented by the "back-calculation" (BC-PID) approach, e.g., [11]. It is based on an idea of the integral action u_I decrease, when the saturation of actuator is active. The input into the integral part of the controller

$$
e_I = \frac{K_P}{T_I}e + \frac{1}{T_t}(v-u) \tag{10}
$$

is reduced. A user has to choice a constant T_t, that together with the inertia χT_D yields two degrees of freedom (after optimization of the controller parameters). In the case of the rate saturation, this technique needs an additional measurement of the actuator output.

3. Dynamic modification of the reference input for removing of wind-up effects

Let us consider a compact form of the basic PID controller equation derived from (7) with $M(q^{-1}) = 1 - q^{-1}$

$$
u_t(1-q^{-1}) = e_t P(q^{-1}) = e_t(p_0 + p_1 q^{-1} + p_2 q^{-2})\,. \tag{11}
$$

The output value u_t combines the control error e_t with the controller parameters p_0, p_1 and p_2 and the previous output value u_{t-1}

$$
e_t = r_t - y_t = (u_t - u_{t-1} - p_1 e_{t-1} - p_2 e_{t-2})/p_0\,. \tag{12}
$$

A change of the controller output is determined by

$$
\Delta u_t = u_t - u_{t-1} = p_0 e_t + p_1 e_{t-1} + p_2 e_{t-2}\,. \tag{13}
$$

Output changes Δu_+ and Δu_- that can be realized by the actuator are determined by

$$
\begin{aligned}
\Delta u_+ &= \min(u_{high} - u_{t-1}, V_{pos} * \Delta) \\
\Delta u_- &= \max(u_{low} - u_{t-1}, -V_{neg} * \Delta)
\end{aligned} \tag{14}
$$

With the above increments of the controller output, the actuator will never exceed constraints (3), (4). It will be satisfied when the control error is determined in respect to a new reference value r'_t

$$
r'_t = \begin{cases} r_t & \Leftrightarrow \Delta u_t \in <\Delta u_-, \Delta u_+> \\ y_t + [\Delta u_+ - e_{t-1}p_1 - e_{t-2}p_2]/p_0 \Leftrightarrow & \Delta u_t > \Delta u_+ \\ y_t + [\Delta u_- - e_{t-1}p_1 - e_{t-2}p_2]/p_0 \Leftrightarrow & \Delta u_t < \Delta u_- \end{cases} \tag{15}
$$

If a change of the controller output Δu_t exceeds the acceptable area (3)-(4), a new virtual reference value r'_t (15) is generated. This modification reduces the control error $e'_t = r'_t - y_t$ value that is used in the algorithm at next time instants $t+1$, $t+2$, etc. It reduces the integral action u_I (now integrating the control errors e'_t) and prevents the windup approach. Known constraints of the actuator (3), (4) can be integrated in the controller PLC program. The algorithm preserves its efficiency at the disturbance compensation. Output changes Δu_- and Δu_+ will profit of the entire actuator output range. The controller output is now determined by a virtual reference signal r'_t, what justifies its proposed name - Dynamic Modification of the Reference value (DMR-PID) algorithm.

4. Comparison of different digital PID algorithms

Testing was performed in a closed-loop system shown in Fig. 2 that provides the possibility of comparison of investigated algorithms both in the case of the reference value change and the disturbance compensation as well.

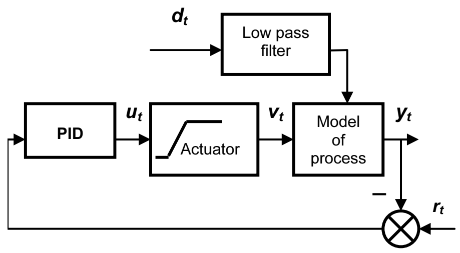

Fig. 2. Closed-loop system used for testing of compared PID algorithms
Rys. 2. Układ regulacji do testowania porównywanych algorytmów PID

In presented examples, the closed-loop control quality is defined by the absolute error index

$$I_R = \sum_{t=N_1}^{N_2} | r_t - y_t | \tag{16}$$

where N_1, N_2 determine the simulation interval. The variations of r_t are high enough to induce actuator saturations. The disturbance compensation efficiency is investigated with the constant reference signal r_0 =0.35y_{max} in the presence of an additive disturbance signal that is introduced by a low pass filter

$$F(s) = \frac{K}{1+sT_1} \qquad K = 2, \qquad T_1 = 1.45s \tag{17}$$

The performance index for the compensation of disturbances is defined as

$$I_D = \sum_{t=N_1}^{N_2} | r_0 - y_t | . \tag{18}$$

The reference for all tested algorithms are results achieved by the I-PID algorithm (6) with parameters K, T_I and T_D optimized for the minimum of (16) and no constraints (ideal actuator). These results are denoted in Tables as I-PID. Results achieved for the controller with the same parameters but subjected constraints (3), (4) are denoted as SAT-PID. All other mentioned algorithms – R-PID (7), TF-PID (9), BC-PID (10) and DMR-PID (15) have used the same parameters K, T_I and T_D. In all investigated cases, additional free parameters are optimized for the minimum value of I_R index (16) (except the DMR-PID algorithm). For easy comparison of results, indices I_R (16) and I_D (18) are determined as a ratio $I_{XX\text{-}PID}$/I_I_PID. Additionally for each algorithm a control time T_C at step response and overshoot η is presented.

The comparison of the closed-loop control is presented with the use of examples of two SISO linear systems: third-order system investigated in [5] and an oscillating system considered in [3].

4.1. Multi-inertia, third-order system

A dynamics of the system was represented by the following linear transfer function

$$G_1(s) = \frac{k}{(1+sT)^3}, \qquad k = 1, \qquad T = 1s \tag{19}$$

In [5] a state space controller was used for closed-loop control with a compensator of the control error in the steady state that can cope with wind-up effects. Table 1 presents the results determined for the optimal parameters for I-PID controller: K=1.89, T_I=2.45s, T_D=1.12s and the sampling interval Δ=0.25s. Algorithms R-PID, TF-PID and BC-PID have free parameters tuned for the minimum of index (16) at step response r_t =1.0. For this experiment, a control time T_C and overshoot η have been determined. In case of the saturation u∈<-2,2>, results determined for the I-PID, are the best, while the ones resulted from the SAT-PID are the worst. Within algorithms subjected constraints, the best is the DMR-PID algorithm with a remarkable advantage over other algorithms, in all considered parameters.

Tab. 1. Control results for the process $G_1(s)$
Tab. 1. Wyniki regulacji dla procesu $G_1(s)$

Controller / index	I-PID	SAT-PID	R-PID	TF-PID	BC-PID	DMR-PID
	Actuator constraints: u∈<-2, 2>					
$I_{R\ XX}/I_{R\ I\text{-}PID}$	1	3.50	1.61	1.71	1.75	1.45
T_C [s]	4.8	11.2	9.0	8.2	8.5	5.6
η [-]	0.12	0.22	0.17	0.11	0.1	0.04
	Actuator constraints: u∈<-2, 2>, V=0.25					
$I_{R\ XX}/I_{R\ I\text{-}PID}$	1	2.81	2.12	2.08	2.10	2.0
T_C [s]	4.8	10.8	9.7	8.0	8.3	6.8
η [-]	0.12	0.26	0.06	0.08	0.09	0.08

More realistic are results achieved in the case of both amplitude and rate constraints, with transients shown in Fig. 3. The SAT-PID result was the worst; other algorithms presented more or less alike results (Fig. 3b), however, the DMR-PID is again the most alert.

a) b)

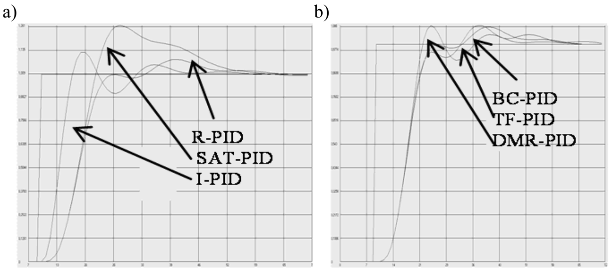

Fig. 3. Step responses of investigated PID-algorithms: a) I-PID, R-PID, and SAT-PID, b) TF-PID, BC-PID, DMR-PID in the case of the constraints v∈<-2,2>, V=0.25 1/s.
Rys. 3. Odpowiedzi skokowe dla regulatorów przy ograniczeniach v ∈<-2,2>, V=0.25 1/s.

The controller output is at first, Fig. 3b, of the same shape as the output of other algorithms (BC-PID and TF-PID) due to saturations, but then it has used full range of possible output values, when the competitors have used only a fraction of the possible output. This is an effect of prolonged derivation action in the controller, and the modified reference value r'_t (15) has had quite another shape than r_t, Fig. 4b.

a) b)

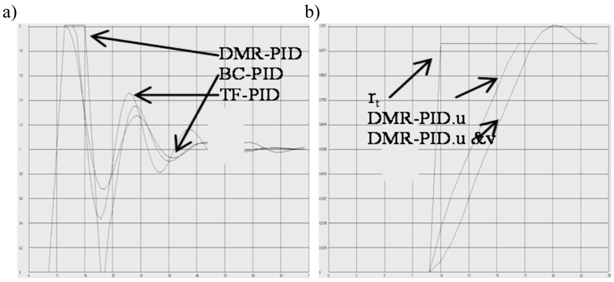

Fig. 4. a) Controller responses, b) transients of the original and modified reference value for the DMR-PID: amplitude (u.) and both amplitude and rate saturation (u&v)
Rys. 4. a) Odpowiedzi regulatorów, b) zmodyfikowana wartość zadana regulatora DMR-PID dla ograniczeń amplitudy (u.) i prędkości (u.& v.)

This effect can explain why this algorithm can gain the lowest control time and performance index $I_{DMR\text{-}PID}/I_{I\text{-}PID}$. Transients of the modified reference value, Fig. 4b, are quite alike as the ramp function used in the drive controllers.

4.2. Oscillatory second-order system with a proportional term

In [3], a SISO second-order oscillatory system has been tested with a weak damping factor

$$G_2(s) = 0.5 + \frac{b_0 + b_1 s}{s^2 + 2\xi\omega_0 s + \omega_0^{\;2}}, \qquad . \quad (20)$$
$$b_0 = 0.9, \qquad b_1 = 0.4, \qquad \xi = 0.223, \qquad \omega_0 = 0.447$$

The system is troublesome to be controlled with the constrained actuator output. For the actuator output within <-0.5,0.5> it can induce strong wind-up effect (the overshoot close to 80%, and control time 170s), while results achieved in [3] for the continuous PID controller fitted to the known process representation with the LMI approach have presented an overshoot of 25% and the control time of appr. 17s. In Table 2 the results are gathered of the closed-loop control with the sampling interval Δ=0.25s and the *I-PID* controller with parameters K=0.026, T_I = 0.16s and T_D = 36.8s (optimized for the minimum of I_R and the gain margin of 6dB). The *I-PID* can achieve quite good control performance. Transients of the step responses for r_t = 2, are presented in Fig. 5a. In the case of the actuator output saturation with no countermeasures (SAT-PID) a very strong wind-up effect is visible. A comparison of the results determined for the actuator output saturation u∈<-0.5, 0.5> has shown again a superiority of the DMR-PID algorithm, both in the case of the reference value transients and for the disturbance compensation. In the case of the actuator output and the rate saturation, the results are alike but this time the BC-PID algorithm is a little better than the DMR-PID. A close investigation of this effect has concluded in the following remark: in case of a very weakly damped oscillatory process behavior (20) rate limitations can induce an oscillatory behavior of the virtual reference r'_t. This effect can be reduced with an additional dumping of r'_t, when oscillations of generated signal r'_t are detected in the control system. An efficient optimization of the I-PID algorithm parameters can be performed when a representation (1),(2) is known. In the case of the Ziegler-Nichols experiment, parameters K, T_I, T_D of PID algorithm are determined by known rules of hand-tuning for the minimum control time [2]. For the investigated process $G_2(s)$, these parameters were equal K=0.1026, T_I = 4.7s and T_D=1.2s. These values are then used in the above mentioned algorithms as fixed parameters, and free parameters have been optimized (for algorithms R-PID, BC-PID and TF-PID). Results of this approach of the controller design are gathered in the last rows of Table 2. The results of BC-PID, TF-PID and DMR-PID are very close, both for the reference value change and the disturbance compensation.

a) b)

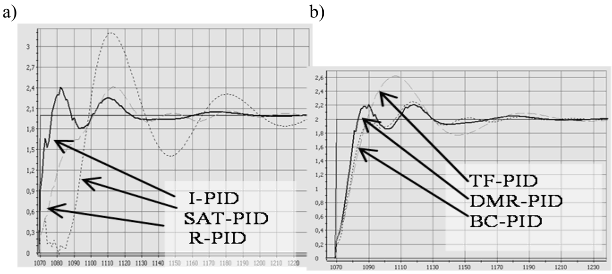

Fig. 5. Results of the control testing for the process $G_2(s)$ with constrains u∈<-0.5, 0.5>: a) system output with the I-PID algorithm, SAT-PID R-PID, b) DMR-PID, BC-PID and TF-PID

Rys. 5. Wyniki modelowania dla procesu $G_2(s)$ przy ograniczeniach v∈<-0.5, 0.5>

Tab. 2. Control results for process $G_2(s)$

Tab. 2. Wyniki regulacji procesu $G_2(s)$

Controller/ Index	I-PID	SAT-PID	R-PID	TF-PID	BC-PID	DMR-PID
	Actuator constraints: u∈<-0.5, 0.5>,					
$I_{R\,XX}/I_{R\,I\text{-}PID}$	1	20.21	4.81	4.85	3.53	2.95
T_C [s]	11.7	59.2	15.7	23.7	13.7	13.5
$I_D/I_{D\,I\text{-}PID}$	1	1.52	1.32	1.16	1.28	1.11
	Actuator constraints: u∈<-0.5,0.5>, V=0.25					
$I_{R\,XX}/I_{R\,I\text{-}PID}$	1	13.25	6.04	7.43	4.97	5.21
T_C [s]	11.7	36.3	13.7	25.5	14.2	15.7
$I_D/I_{D\,I\text{-}PID}$	1	1.48	1.30	1.25	1.28	1.22
Ziegler-Nichols	Actuator constraints: u∈<-0.5, 0.5>					
$I_{R\,X}/I_{R\,I\text{-}PID}$	1	2.54	1.24	1.08	1.09	1.07
T_C [s]	31	47	35	32	35	34
$I_D/I_{D\,I\text{-}PID}$	1	1.22	1.21	1.11	1.12	1.09

5. Conclusions

The proposed approach of reference value dynamic modification, DMR-PID, based on the idea of creating for system controller-actuator, a new reference value signal seems to be quite prospective. In all cases it is better than the R-PID, it requires little calculation effort, and has no tuned parameters. The virtual reference value r'_t is like a ramp function applied in many control systems but its shape has not to be fixed before the control start, what made the proposed approach to be more flexible. It is adapted to the actual state of the control and is most efficient in the sense of a fast control and the disturbance compensation.

The introduced modification preserves a structure of the basic PID-controller. The only innovation is a calculation of a new virtual reference value r'_t (15) which can be made without any problems by the CPU controller. The only new information are constraints (3), (4) induced by the actuator, and these have to be known by the design engineer. The proposed modification does not require any additional tuning.

An additional advantage of the proposed approach is a separation of the controller parameter design and the reduction actuator constraints effects. Once optimized, the controller parameters can be used for different actuator constraints (different actuators) used in the closed-loop control.

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Artykuł recenzowany