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CONTRIBUTION TO THE THEORETICAL ANALYSIS OF THE ZIEGLER–NICHOLS METHOD

Mikuláš Huba — Katarína Žáková *

The high practical significance of the I_1T_d-plant approximations has been shown by the perhaps most frequently used method for controllers tuning by Ziegler and Nichols, modified later for sampled data systems by Takahasi *et al.* This paper extends previous works by analysing also control structures that include I_1T_1 and more general $I_1T_1T_d$-system. All these approximations are used for tuning of the P and PI controllers designed for single integrator system with constrained input.

Keywords: P (PD) controllers, controller tuning, Ziegler-Nichols method

1 INTRODUCTION

One of the most widespread tuning rules was introduced by [13] already at the beginning of 40's in the last century. They have presented two different methods of the controller setting. While in the first one, the controller parameters are determined on the basis of the transient response measurement, in the "ultimate sensitivity method" it is necessary to bring the system to the stability boundary by raising the gain of a P controller. Ziegler and Nichols have designed controllers to guarantee the "quarter amplitude ratio" (quarter amplitude damping). This criterion was also followed by *eg* [9] and [2]. Ziegler-Nichols method creates a starting point for numbers of other tuning rules.

2 SETTING OF THE PROPORTIONAL ACTION

For the sake of simplicity, Ziegler and Nichols have not analyzed the types of time lag involved in the control loop. In order to show the possible consequences, the behavior of an integrator system combined with two basic parasitic delays: the dead time and the time constant delay will be analyzed. A similar approach was used previously by [8].

2.1 I_1T_d system

2.1.1 Optimal setting of P controller

Let us consider the integral + dead time system

$$F(s) = \frac{K_S\, e^{-T_d s}}{s} \tag{1}$$

that is going to be controlled by an analog P controller. The suggested design of the optimal controller setting that corresponds to the multiple dominant pole follows a requirement of monotonic transient response at the presence of possible maximal value of controller gain. Therefore, its setting can be found by looking for a controller gain extreme. It means the optimal gain of the P controller K_C has to satisfy the condition

$$\frac{\mathrm{d}K_C}{\mathrm{d}s} \stackrel{!}{=} 0 \tag{2}$$

whereby K_C can be expressed from the characteristic equation of the system closed loop:

$$K_C K_S\, e^{-T_d s} + s = 0\,. \tag{3}$$

Solving (2) and (3) leads to the optimal controller gain

$$K_{Copt} = \frac{1}{eK_S T_d} \tag{4}$$

and dominant real pole

$$s_{opt} = -1/T_d\,. \tag{5}$$

These formulae are known for long time — see *eg* [8], [4] or [12]. The derived gain can be used for setting of P controller in the structures with constrained control signal (Fig. 1), too. The blocks "Plant" and "Delay" together represent any process that can be approximated by I_1T_d model. The dominant $1st$ order dynamics presented by a single integrator is supposed to be deployed in the feedforward path. The transport delay can be present at any place within the feedback.

2.1.2 Critical controller gain

For comparison, in addition to the optimal gain of the P controller there was analytically derived also a gain at the stability margin. For this one a term "critical controller gain" was introduced. To derive the critical closed loop gain we can substitute

$$s = j\omega \tag{6}$$

* Department of Automation and Control, Faculty of Electrical Engineering and Information Technology, Slovak University of Technology, Ilkovičova 3, 812 19 Bratislava, Slovakia Email: [huba, katarina.zakova]@elf.stuba.sk

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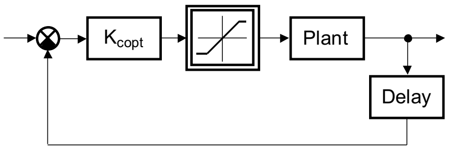

Fig. 1. P controller with constrained output

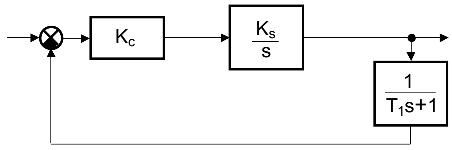

Fig. 2. Control structure with I_1T_1 system

into (3). In this way, there arises a set of two equations

$$\begin{aligned} K_C K_S \cos(T_d\omega) &= 0 \\ K_C K_S \sin(T_d\omega) &= \omega \end{aligned} \tag{7}$$

whose solution leads to the critical controller gain. A trivial solution of (7) $\omega = 0$ leads to the minimum critical value

$$K_{Ccr1} = 0\,. \tag{8}$$

The second solution gives the maximum critical value of the stability area

$$K_{Ccr2} = \frac{\pi}{2K_S T_d}\,. \tag{9}$$

The corresponding period of oscillation is

$$P_u = \frac{2\pi}{\omega} = 4T_d\,. \tag{10}$$

2.2 I_1T_1 system

2.2.1 Optimal controller gain

In the case of I_1T_1-system

$$F(s) = \frac{K_S}{s\,(T_1 s + 1)} \tag{11}$$

whose dynamics is deployed in the control structure according to Fig. 2, the optimal controller gain can be derived from the requirement of double real pole of the closed loop characteristic equation

$$T_1 s^2 + s + K_C K_S \ = 0\,. \tag{12}$$

This requirement corresponds to

$$K_{Copt} = \frac{1}{4K_S T_1} \tag{13}$$

whereby

$$s_{opt} = -\frac{1}{2T_1}\,. \tag{14}$$

The optimal controller gain (13) can also be applied in the structure shown in Fig. 1. Here, the block "Plant" represents the dominant single integrator dynamics and the block "Delay" represents the identified time constant T_1.

2.2.2 Critical controller gain

The critical controller gain for this system is going to infinity, since all coefficients of the characteristic polynomial are positive and according to Routh-Schur test of stability it means that the system is stable.

$$K_{Ccr} = \infty\,. \tag{15}$$

2.3 $I_1T_1T_d$ system

2.3.1 Optimal controller gain

The last integral model that will be considered is an integral system with both dead time and time constant

$$F(s) = \frac{K_S}{s(T_1 s + 1)} e^{-T_d s}\,. \tag{16}$$

The optimal control gain can be determined from the characteristic equation

$$s(T_1 s + 1) + K_C K_S e^{-T_d s} = 0\,. \tag{17}$$

After taking a derivative of (17) according to the complex variable s

$$2T_1 s + 1 - K_C K_S T_d e^{-T_d s} \overset{!}{=} 0 \tag{18}$$

and expressing

$$K_C K_S T_d e^{-T_d s} = -(T_1 s^2 + s) \tag{19}$$

we get the double real pole value by means of a simple substitution (19) into (18):

$$s_{1,2} = -\frac{1}{2}\frac{T_d + 2T_1 \pm \sqrt{T_d^2 + 4T_1^2}}{T_1 T_d}\,. \tag{20}$$

Substituting the value of s_1 corresponding to the sign "+" into (19) we get the optimal gain of the proportional controller

$$K_{Copt} = \frac{-2T_1 + \sqrt{T_d^2 + 4T_1^2}}{K_S T_d^2\, e^{\frac{2T_1 + T_d - \sqrt{T_d^2 + 4T_1^2}}{2T_1}}}\,.$$

After denoting the total time lag

$$L = T_d + T_1 \tag{21}$$

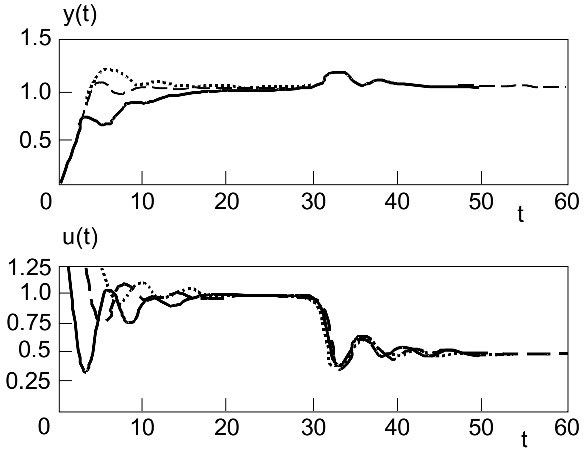

Fig. 3. Optimal and critical P controller gain. Solid line: $I_1T_1T_d$ system with $\varepsilon = 1$; dashed line: I_1T_d system ($\varepsilon \to \infty$); dotted line: I_1T_1 system ($\varepsilon = 0$).

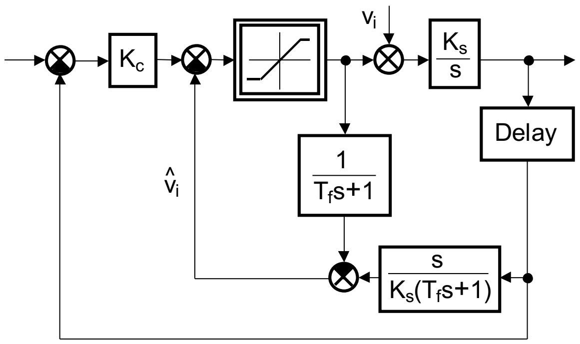

Fig. 4. Reconstruction and compensation of the input system disturbance (windupless PI controller).

and the ratio

$$\varepsilon = \frac{T_d}{T_1} \tag{22}$$

there follow

$$T_d = \frac{L\varepsilon}{\varepsilon + 1}\,, \qquad T_1 = \frac{L}{\varepsilon + 1}\,. \tag{23}$$

Then

$$K_{Copt} = \frac{-2 + \sqrt{\varepsilon^2 + 4}}{K_S \varepsilon^2 L} \frac{\varepsilon + 1}{e^{\frac{2+\varepsilon-\sqrt{\varepsilon^2+4}}{2}}}\,. \tag{24}$$

The graphical dependence of the normed gain $K_{opt} = K_{Copt}K_S$ on the total lag L for $\varepsilon = 1$ is presented in Fig. 3. It is interesting to note that this optimal value lies between the values corresponding to the limit situations with $\varepsilon = 0$ and $\varepsilon \to \infty$. This conclusion regarding $I_1T_1T_d$ system can be generalized also for other values $\varepsilon \in (0, \infty)$.

The dependence of the optimal and critical gain on the ratio T_d/T_1 can be found *eg* in [8].

2.3.2 Critical controller gain

The derivation of the critical closed loop gain comes out from the characteristic equation (17). Substitution $s = j\omega$ leads to the pair of equations

$$K_C K_S \cos(\omega T_d) = \omega^2 T_1$$
$$K_C K_S \sin(\omega T_d) = \omega$$

Unfortunately, it is not possible to compute the value of critical gain analytically and therefore only a graphical representation of K_{Ccr} corresponding to $\varepsilon = 1$ is introduced in Fig. 3. This is close to that corresponding to $\varepsilon \to \infty$.

3 SETTING OF THE PROPORTIONAL AND INTEGRAL ACTIONS

Up to now, only the design of the P controller for the single integrator has been considered. The additional time lags were compensated by decreasing the optimal controller gain. Due to this, an increase of time delays causes an increase of the possible permanent control error. One possibility is to compensate the permanent control error by the feedforward control. The other possibility is to use a windupless integral action. The basic control loop (1^{st} order system + delay + P controller) can be extended by blocks that serve for reconstruction and compensation of an input system disturbance v_i. It creates the integral action of the controller.

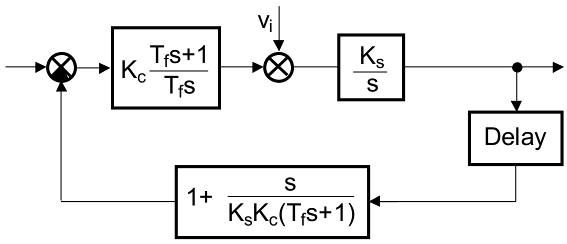

Fig. 5. Equivalent structure for the proportional band to the scheme in Fig. 4

The new control structure (see also [5], [7] and [6]) can be seen in Fig. 4. The included filter serves for a reduction of possible noise in the structure. For simplicity, it is connected with reconstruction blocks. In the case when only a linear part of saturation is considered, the whole control structure can be modified to the form that is presented in Fig. 5.

It is evident that the control corresponds to PI controller with the transfer function

$$C(s) = K_C\Big(1 + \frac{1}{T_f s}\Big)\,. \tag{25}$$

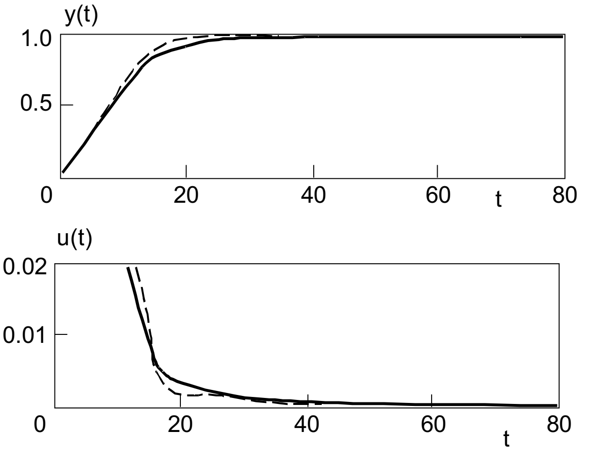

Fig. 6. Transient responses and control signals for I_1T_d system with the windupless PI controller according to Fig. 4 ($K_S = 3$, $T_d = 1.5$)

3.1 Structures with transport delay

The controller design follows the control structure in Fig. 4 where the block of delay represents the transport delay in the loop. Arising from the characteristic polynomial of the closed loop

$$chp(s) = T_f s^2 + (1 + K_S K_C T_f)se^{-T_d s} + K_S K_C e^{-T_d s} \tag{26}$$

the setting of controller and filter can be determined from the requirement of its dominant triple real pole. It can be achieved by solving a following set of equations

$$chp(s) = 0\,; \quad \frac{\mathrm{d}}{\mathrm{d}s}chp(s) = 0\,; \quad \frac{\mathrm{d}^2}{\mathrm{d}s^2}chp(s) = 0\,. \tag{27}$$

Hence

$$s_{opt} = \frac{-2+\sqrt{2}}{T_d}\,, \tag{28}$$

$$K_{Copt} = \frac{\sqrt{2}-1+I(5\sqrt{2}-7)\tau}{K_S T_d}e^{-2+\sqrt{2}}\,, \tag{29}$$

where

$$\tau = \sqrt{(14+10\sqrt{2})e^{2-\sqrt{2}} - 12\sqrt{2} - 17}\,.$$

The time constant of the filter is

$$T_{f\,opt} = \frac{3+2\sqrt{2}+I\tau}{2}T_d\,. \tag{30}$$

However, the achieved solutions are complex numbers and the controller and the filter as well are supposed to work with real parameters. Experimentally, it can be shown that the loop dynamics that corresponds to parameters, which are derived by approximating the computed complex values by the real part or by the module, is close to the required one. Then

$$\begin{aligned} T_{f\,re} &= \frac{2\sqrt{2}+3}{2}T_d \doteq 2.914T_d\,, \\ T_{f\,\mathrm{mod}} &= T_d\frac{\sqrt{10\sqrt{2}+14}}{2}e^{\frac{2-\sqrt{2}}{2}} \doteq 3.555T_d\,, \end{aligned} \tag{31}$$

$$\begin{aligned} K_{C\,re} &= \frac{\sqrt{2}-1}{K_S T_d}e^{-2+\sqrt{2}} \doteq \frac{0.23}{K_S T_d}\,, \\ K_{C\,\mathrm{mod}} &= \frac{\sqrt{10\sqrt{2}-14}}{K_S T_d}e^{\frac{\sqrt{2}-2}{2}} \doteq \frac{0.281}{K_S T_d}\,. \end{aligned} \tag{32}$$

For comparison, the controller gain of the P controller was $K_{Copt} \doteq 0.367/K_S T_d$ (4), *eg* approximately 1.6 times bigger that the gain K_{Cre} of the PI controller.

In Fig. 6 it is possible to compare transient responses with the corresponding control signals for both types of approximations. Simulations are done for I_1T_d system controlled by PI controller according to the structure in Fig. 4. It can be seen that there is not a big difference between both approximations and therefore the simpler approximation by the real part will be used.

The correctness of the chosen approximation is also observable from Fig. 7. Let us follow the approximation by the real part. Firstly, the simulation was accomplished with the controller parameters that were set according to formulae given by (31) and (32). Then, the simulation was compared with simulations where controller parameters were increased or decreased by 20%. The quality of these modified transient responses and the corresponding control signals is slightly going down.
Finally, it is to note that resulted formulae (31) and (32) remind of formulae achieved by Ziegler and Nichols. Their form is the same, they slightly differ only in constants.

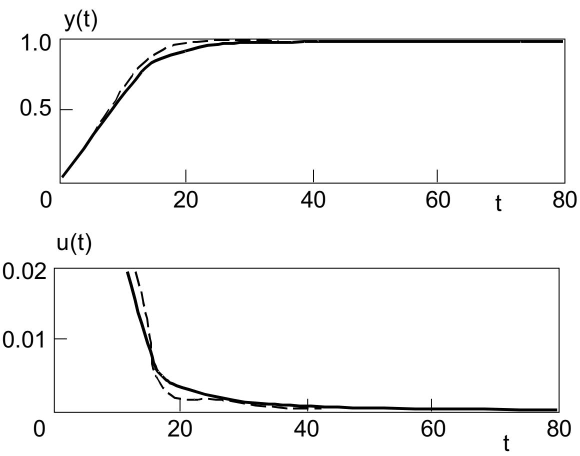

Fig. 7. Transient responses and control signals for I_1T_d system controlled by PI controller. Solid line: controller parameters according to (31) and (32); dashed line: controller parameters decreased by 20%; dotted line controller parameters increased by 20%.

3.2 Structures with time constant delay

The controller design follows the control structure in Fig. 4 where the block of delay represents the time constant delay in the loop. The optimal setting of the controller can be found from the closed loop characteristic polynomial

$$chp(s) = T_1 T_f s^3 + T_f s^2 + +(1 + K_C K_S T_f)s + K_C K_S \quad (33)$$

Following (27) one gets the optimal controller setting

$$K_{Copt} = \frac{3 \pm I\sqrt{3}}{18 K_S T_1}$$
$$T_{f\,opt} = \frac{3T_1}{2}(3 \pm I\sqrt{3})$$

that correspond to the dominant triple pole

$$s_{opt} = -\frac{1}{3T_1}\,. \quad (34)$$

After approximation of the complex solution by the real part or by the module one gets the following PI controller settings

$$K_{C\,re} = \frac{1}{6K_S T_1} \doteq \frac{0.1667}{K_S T_1}$$
$$K_{C\,\mathrm{mod}} = \frac{\sqrt{3}}{9K_S T_1} \doteq \frac{0.192}{K_S T_1} \quad (35)$$

and

$$T_{f\,re} = \frac{9}{2}T_1 \doteq 4.5T_1$$
$$T_{f\mathrm{mod}} = 3\sqrt{3}T_1 \doteq 5.196T_1 \quad (36)$$

For comparison the P controller gain for the I_1T_d system was $K_{Copt} = 0.25/K_S T_1$.
The graphical presentation of the achieved solution can be found in Fig. 9.

3.3 Structures with transport and time constant delay

Following the control structure (Fig. 4) with a transport and a time constant delay ($I_1T_1T_d$ system) one gets the characteristic polynomial

$$chp(s) = T_1 T_f s^3 + T_f s^2 + [(1 + K_C K_S T_f)s + K_C K_S]e^{-T_d s}\,. \quad (37)$$

As it was mentioned already before, the setting of controller and filter can be determined from the requirement of its dominant triple pole that can be found from equations (27). Unfortunately the analytical solution is not available. The controller parameters can be computed only numerically as depending on the total lag L (21) and the ratio ε (22). For example for $\varepsilon = 0.5$ one gets $K_{C\,opt} = 0.193/K_S L$ and $T_{f\,opt} = 3.717L$ and for $\varepsilon = 2$ it can be found that $K_{C\,opt} = 0.219/K_S L$ and $T_{f\,opt} = 3.127L$. The controller parameters for $\varepsilon = 1$ are illustrated in Fig. 9 and there it holds $K_{C\,opt} = 0.207/K_S L$ and $T_{f\,opt} = 3.386L$.

4 EXAMPLE

Let us consider the system described by the transfer function

$$F(s) = \frac{1}{(s+1)^3} \quad (38)$$

Firstly, the controller design is based on the I_1T_d, I_1T_1 and $I_1T_1T_d$ approximations. Experiments show that such integrator based approximation model should concentrate mainly on the initial part of the step response [14]. Following this requirement one can receive

model	transfer function
I_1T_d	$\frac{0.257}{s}e^{-0.745s}$
I_1T_1	$\frac{0.429}{s(1.95s+1)}$
$I_1T_1T_d$	$\frac{0.32}{s(0.9s+1)}e^{-0.215s}$

The method described in [11], [3] and [13] arising from I_1T_d approximation are compared with the proposed PI controller. All controller parameters can be found below

	Controller parameters	
	$K_C T_I$	(T_f)
I_1T_d	1.2065	2.1709
I_1T_1	0.1993	8.7750
$I_1T_1T_d$	0.5093	4.4964
Tyreus-Luyben	2.5435	6.5187
Fruehauf et al.	2.6114	3.7250
Ziegler-Nichols	4.7006	2.4809

The control signal was limited to $U_{\min} = -1.25$ and $U_{\max} = 1.25$ and at $t = 30$ there was considered an input disturbance with the value $v_i = 0.5$.

The simulation results are presented in Fig. 8. The proposed PI controller based on I_1T_d approximation was able to reach the required value without an overshoot. The non presented simulations following I_1T_1 and $I_1T_1T_d$ approximations had a too long settling time whereby the settling time of transient response based on I_1T_1

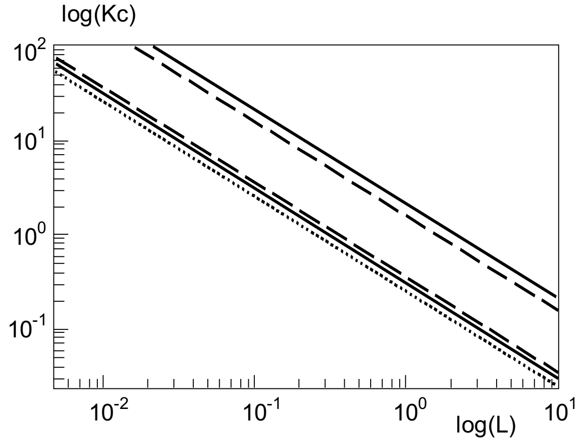

Fig. 8. Transient responses and control signals for the given example. Setting of PI controller is done according to approximation by I_1T_d model (solid line), Tyreus-Luyben method (normal dashed line), Fruehauf-Chien-Lauritsen method (short dashed line).

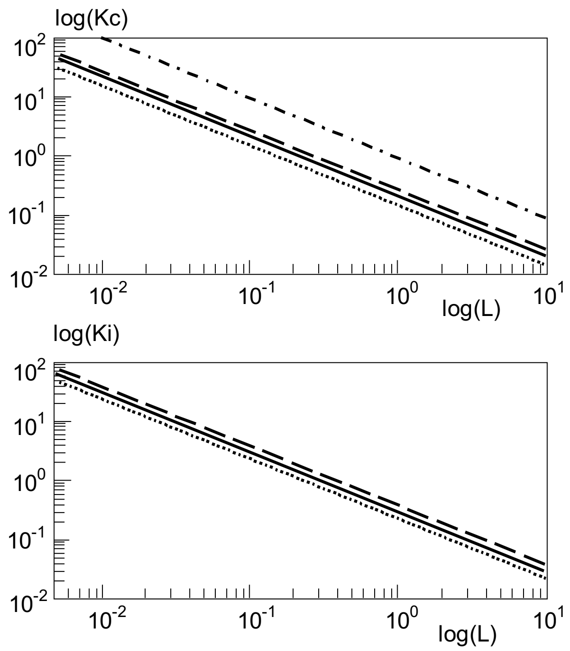

Fig. 9. Optimal parameters of the PI controller ($K_i = 1/T_f$). Solid line: $I_1T_1T_d$ system ($\varepsilon = 1$); dashed line: I_1T_d system; dotted line: I_1T_1 system; dash-and-dot line: Ziegler-Nichols setting

model was longer than at the transient response following $I_1T_1T_d$ approximation.

5 DISCUSSION

The paper was dedicated to the controller design following the control structure with the $1st$ order dynamics in the feedforward path and possible additional parasitic delays in the feedback. On the basis of such distribution of the dynamical elements, P controller following the dominant pole design was derived. The design procedure also takes into account always present limitation in the control signal. The disturbance reconstruction and compensation is solved by introducing an I action to the control structure. It leads to the establishment of PI controller.

In contrast to the "quarter amplitude damping" introduced by Ziegler and Nichols, the controller setting corresponds to the double (triple) real dominant pole and guarantees the fastest possible transients without overshoot.

The comparison of the optimal P controller gains (4) and (13) with that one given by the Ziegler-Nichols "transient response method" shows that the approximations by the integral+dead time system or integral+time constant system yield the gains $K_{Copt} = 0.368/K_SL$ $(L = T_d)$ and $K_{Copt} = 0.25/K_SL$ $(L = T_1)$ which are 2.71 or 4 times smaller than the values recommended by Ziegler and Nichols. This is caused by the different performance criteria: The double real pole corresponds to transients without overshoot, while Ziegler and Nichols tried to set the gain as high as possible in order to reduce the offset and influences of possible load changes. However, this also creates a closed loop system that is very poorly damped and that has very poor stability margins. Then, as also mentioned in [1], the closed loop gain is typically 2–3 times higher. It is to note that analytical solutions for optimal and critical gain of P controller were discussed for I_1T_1 and I_1T_d model already by Oldenbourg and Sartorius, [8].

For the PI controller it is possible to compare parameter $K_I = 1/T_i$ that for the approximation by I_1T_d model with the dead time $L = T_d$ gives

$$K_{Iopt} = 0.343/L\,. \tag{39}$$

For I_1T_1 model where $L = T_1$ one gets

$$K_{Iopt} = 0.222/L \tag{40}$$

The above computed parameters are very close to the value determined by Ziegler and Nichols that is $K_I = 0.3/L$. As far as the setting of the proportional gain of the PI controller is considered, Ziegler and Nichols proposed to decrease its value in comparison to the "pure" P controller. This was due to the increased inclination of the loop to oscillations in the presence of the I part of the controller. The tendency of the proportional parameter decrease can also be seen at the presented controller design.

Following the previous analysis, to accept only one type of the dominant time lag could lead to not sufficiently precise results. A broad class of dynamical systems can be approximated by a model containing both types of the dominant time lags (16). Then, new relations for the optimal and the critical controller gain depending on the ratio T_d/T_1 can be found. The previous analysis shows that the P controller optimal gain computed on the basis of $I_1T_1T_d$ model lies between values that were computed for above mentioned I_1T_d and I_1T_1 models. For example, for $T_d/T_1 = 0.5$ the optimal gain is $K_{Copt} = 0.297/K_SL$

($L = T_d + T_1$), for $T_d/T_1 = 1$, $K_{Copt} = 0.322/K_S L$ and for $T_d/T_1 = 2$, $K_{Copt} = 0.346/K_S L$. The gains are 3.371, 3.103 and 2.891 times smaller than the values recommended by Ziegler and Nichols. The conclusion can be confirmed by the graphical dependence in Fig. 3 that is drawn for $\varepsilon = 1$.

Similar results were also achieved for PI controller setting. As it can be seen from Fig. 9 both controller parameters (K_{Copt} and K_{Iopt}) computed on the basis of $I_1T_1T_d$ model are again situated between values corresponding to I_1T_d and I_1T_1 models. In the figure the optimal values for selected models are compared with the values determined by Ziegler and Nichols. The analytically computed parameter K_I that follows $I_1T_1T_d$ model is practically equivalent to the value that was found by Ziegler and Nichols. The coincidence is reached for $\varepsilon = T_d/T_1 \doteq 1.13$.

However, it is necessary to mention that Ziegler and Nichols used the classical structure of PI controller, whereas the PI controller introduced here arose on the basis of reconstruction and compensation of the input system disturbance. Such modification of control structure ensures smaller overshoot of the transient response than the parallel PI controller does.

One of the main advantages of this concept of the windupless PI controller and its tuning is that they can be generalized also for the higher order approximative models, which allow to achieve higher control dynamics.

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References

[1] ASTROM, K. J.—HAGGLUND, T.: PID Controllers: Theory, Design, and Tuning, 2nd ed., Instrument Society of America, 1995.

[2] CORRIPIO, A. B.: Tuning of Industrial Control Systems, Instrument Society of America, Research Triangle Park, North Carolina.

[3] FREUHAUF, P. S.—CHIEN, I. L.—LAURITSEN, M. D.: Simplified IMC-PID Tuning Rules, ISA/93 Advances in Instrumentation and Control **48** (1993), 1745–1766.

[4] HUBA, M.: Using microcomputers as special control and measuring devices (Použitie mikropočítačov ako špeciálnych riadiacich a meracích prístrojov), EF SVŠT. (in Slovak)

[5] HUBA, M.—KULHA, P.: Linear and Nonlinear (Saturating) Structures of PID Control, Summer Course BEST "Computed Aided Control Design", Bratislava 1994.

[6] HUBA, M.—KULHA, P.: Saturating Control for the Dominant First Order Plants, Proceeding of the IFAC Workshop "Motion Control", Munich 1995, pp. 197–204.

[7] HUBA, M.—KULHA, P.—BISTAK: Synthese der Systeme mit Begrenzungen: Dynamik der PI- und PID- Regler, Elektrotechnisches Kolloquium FAU Erlangen 1995. (in German)

[8] OLDENBOURG, R. C.—SATORIUS, H.: Selbsttätiger Regelnugen, 2. Auflage, R. Oldenbourg Verlag, München, 1951.

[9] PAAR, E. A.: Industrial Control Handbook, Vol. 3, BSP Professional Books 1989.

[10] TAKAHASHI, Y.—CHAN, C. S.—AUSLANDER, D. M.: Parametereinstellung Bei Linearen DDC-Algorithmen, Regelungstechnik und Prozess-Datenverarbeitung **19** (1971), 237–244.

[11] TYREUS, B. D.—LUYBEN, W. L.: Tuning PI Controllers for Integrator/Dead Time Processes, Industrial Engineering Chemistry Research **31** (1992), 2625–2628.

[12] VITECKOVA, M.—VITECEK, A.—SMUTNY, L.: Controller Tuning for Controlled Plants with Time Delay, Preprints of IFAC Workshop on Digital Control (Past, Present and Future of PID Control), Terrassa (Spain) 2000, pp. 283–288.

[13] ZIEGLER, J. G.—NICHOLS, N. B.: Optimum Settings for Automatic Controllers, Transaction of the A.S.M.E. (1942), 759–768.

[14] ZAKOVA, K.: Design of Controllers for Time-Delayed Systems, Slovak University of Technology, Faculty of Electrical Engineering and Information Technology, Bratislava 2003.

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Mikuláš Huba (Doc, Ing, CSc) born 1951 in Ružomberok, graduated from the Slovak University of Technology in Technical Cybernetics in 1974. From 1989 an associated professor at the Department of Control and Automation, Faculty of Electrical Engineering and Information Technology, Slovak University of Technology, appointed from 1996 also by leading the Local Centre of Distance Education. Active in the field of nonlinear control and modern learning technologies, like Internet based interactive education, computer aided/based control design, multimedia, *etc*.

Katarína Žáková (Ing, PhD) graduated from the Slovak University of Technology in Technical Cybernetics in 1991. Since 1991 she is with the Department of Automation and Control, Faculty of Electrical Engineering and Information Technology, Slovak University of Technology. Her current research activities are connected with synthesis of nonlinear systems, computer aided control design and e-learning.