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2DOF PID for Current Driven Servo

Leszek Trybus, Andrzej Bożek
Rzeszów University of Technology, Department of Computer and Control Engineering, al. Powstańców Warszawy 12, 35-959 Rzeszów

Abstract: A design method for 2DOF PID continuous or discrete controller for a current driven servo described by a double integrator is presented. The closed-loop time constant is the single design data. The method results in the rules for three PID settings and two PID weights b, c that reduce gains for P and D modes for the set-point input. The PID settings provide a triple real pole of the closed-loop system to assure smooth transients. The weights b, c eliminate overshoot of the step response. In the case of continuous control, the weights are fixed, independent of the design data. In the discrete case, the weights decrease when the discretization step increases. Laboratory experiments confirm fairly good correspondence of system response to simulations, but also demonstrate the effects of friction when the output is close to a set-point. Role of additional parameters of an industrial PID function block is also examined.

Keywords: current driven servo, double integrator, 2DOF PID, set-point prefiltering, pole placement, friction effects

1. Introduction

2DOF PID (two-degree-of-freedom) controller allows for shaping the set-point response of a feedback system without altering the disturbance response. Conventional way of achieving this is to apply a prefilter on the set-point input. In practical solutions, the PID controller is first tuned for satisfactory disturbance response and then the prefilter is added to get satisfactory set-point response, usually with a small or no overshoot. The prefilter and PID may be replaced by a single 2DOF PID controller which by means of two weights, denoted typically by b, c, decreases the set-point input for P and D modes, respectively [1, 2]. Since the weights determine zeros of the control system, they may be adjusted to shape the set-point response. The 2DOF PID controller is particularly convenient for integrating plants which for satisfactory disturbance response and no prefilter exhibit excessive overshoots [3, 4].

Servo-drives, indispensable in robotics, machine tools, etc. [5], are typical examples of integrating plants. Since oscillatory transients are forbidden, tuning a servo controller for multiple real poles of the closed-loop system is one of possible options. In the case of integrator plus time constant plant, being a model of voltage driven servo [6], the system with PID controller becomes of the third order, so may be tuned for a triple pole. This has already been presented for 2DOF PID in [7] with such selection of the weights b, c, so as to remove two of the three poles. Then the system for the set-point input becomes of the first order, so provides smooth transients. Discrete PID for such a servo is presented in [8], however without referring to 2DOF. As far as the current driven servo modeled by a double integrator is concerned, continuous and discrete PID controllers for the triple pole are designed in [9], also without referring to 2DOF. Therefore, the natural objective of this paper is to extend the results from [9] on 2DOF PID, motivated by the approach from [7].

We add that other design methods for a double integrator are reviewed in [10]. Advanced design involving a constrained state feedback controller for a permanent magnet synchronous servo-drive is presented in [11]. For practical applications conventional frequency designs are used, given natural frequency and damping ratio [6, 12].

The paper is organized as follows. Design of the continuous 2DOF PID controller for a double integrator given a closed-loop time constant is presented in Secs. 2 and 3. Next two sections deal with the discrete case, ended up with analytic expressions and nomograms for the weights b, c in terms of the closed-loop system triple pole. Experimental and simulated responses recorded in a lab set-up are compared in Sec. 6 taking also into account features of an industrial PID controller. Last section summarizes the results.

2. Triple pole placement

Consider a feedback system of Fig. 1 involving a double integrator k_o/s^2 as the plant, PID controller, and a set-point prefilter F. The double integrator may represent a servo-drive with current (torque) driven motor and mechanical load, resulting in the effective gain k_o. The standard PID controller has the form

$$PID(s) = k_P + \frac{k_I}{s} + k_D s. \tag{1}$$

Autor korespondujący:
Andrzej Bożek, abozek@prz.edu.pl

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Hence, the closed-loop transfer function for the no prefilter case (NF, i.e. F = 1) becomes

$$G_{cl}(s)=\frac{k_o\left(k_Ds^2+k_Ps+k_I\right)}{s^3+k_o\left(k_Ds^2+k_Ps+k_I\right)}. \tag{2}$$

We want G_{cl} to have a triple pole $-p$. By equating corresponding coefficients in the G_{cl} denominator with $(s+p)^3$ polynomial one obtains the following controller settings [9]

$$k_P=\frac{3}{\lambda^2k_o},\ k_I=\frac{1}{\lambda^3k_o},\ k_D=\frac{3}{\lambda k_o}, \tag{3}$$

here expressed in terms of a given closed-loop time constant $\lambda=1/p$. Step response of G_{cl} shown in Fig. 2 for $\lambda=0.075$ s (as in Sec. 6) exhibits 20 % overshoot (NF plot).

To eliminate the overshoot, note that for the settings (3) the controller transfer function becomes

$$PID(s)=\frac{3}{\lambda k_o}\frac{(s+\alpha_c)(s+\bar{\alpha}_c)}{s}, \tag{4a}$$

with the complex zero

$$-\alpha_c=-\frac{1}{2\lambda}\left(1+j\frac{1}{\sqrt{3}}\right). \tag{4b}$$

The overshoot may be eliminated by applying a first order prefilter determined by the real part of the zero, so

$$F_1(s)=\frac{\mathrm{Re}\,\alpha_c}{s+\mathrm{Re}\,\alpha_c}=\frac{1}{2\lambda s+1}. \tag{5}$$

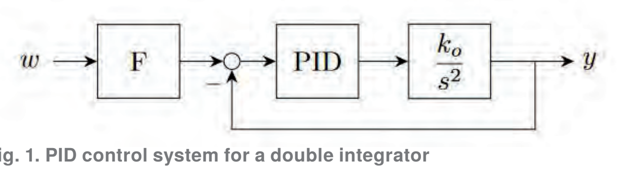

Fig. 1. PID control system for a double integrator
Rys. 1. Układ z regulatorem PID dla podwójnego integratora

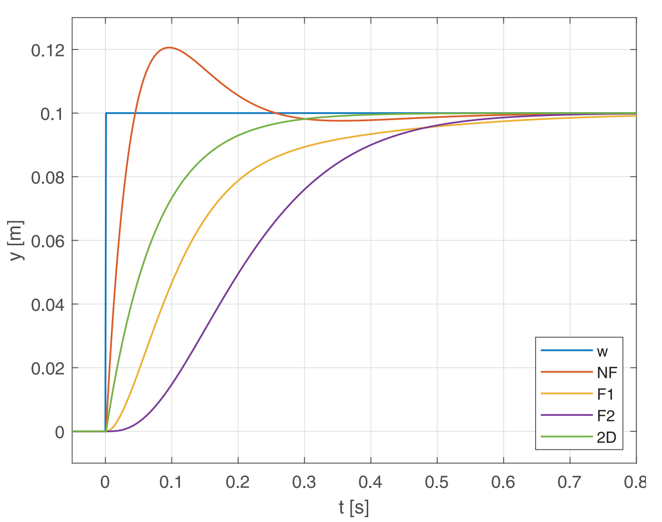

Fig. 2. Step responses of the control system
Rys. 2. Odpowiedzi skokowe układu sterowania

Settling time is estimated as $t_s\cong 8\lambda$, i.e. $t_s\cong 0.6$ s for $\lambda\cong 0.075$ s (Fig. 2). Eliminating the zeros from G_{cl} by

$$F_2(s)=\frac{k_I}{k_Ds^2+k_Ps+k_I} \tag{6}$$

is another filtering option.

3. 2DOF PID for servo

Following [1, 2] introduce a reduced gain PID controller in the form

$$PID_{bc}(s)=bk_P+\frac{k_I}{s}+ck_Ds \tag{7}$$

with the weights b, c. Such a controller is a component of the 2DOF PID control system in Fig. 3. The diagram can be transformed to the original one in Fig. 1 by using the filter

$$F_{2D}(s)=\frac{PID_{bc}(s)}{PID(s)}=\frac{ck_Ds^2+bk_Ps+k_I}{k_Ds^2+k_Ps+k_I}. \tag{8}$$

After cancellation of the terms $k_Ds^2+k_Ps+k_I$ and using $(s+p)^3$ in the denominator the system transfer function $F_{2D}G_{cl}$ becomes

$$F_{2D}G_{cl}(s)=\frac{\left(ck_Ds^2+bk_Ps+k_I\right)k_o}{(s+p)^3}. \tag{9}$$

To decrease the settling time the idea proposed in [7] is to express the numerator $ck_Ds^2+bk_Ps+k_I$ as $ck_D(s+p)^2$, so as to reduce the order of the system to one. Using the settings (3) and equating coefficients of the two polynomials yield the following fixed values of the 2DOF weights

$$b=\frac{2}{3},\quad c=\frac{1}{3}. \tag{10}$$

So they do not depend either on the plant's k_o or the design data λ.

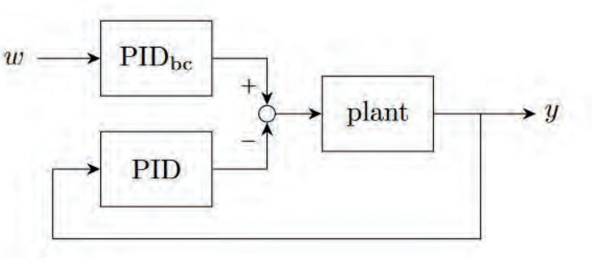

Fig. 3. 2DOF PID control system with two controllers
Rys. 3. Struktura układu sterowania 2DOF PID z dwoma regulatorami

The system transfer function simplifies to

$$F_{2D}G_{cl}(s)=\frac{p}{s+p}=\frac{1}{\lambda s+1}, \tag{11}$$

so with the settling time 4λ, twice shorter than before (see Fig. 2). The 2DOF filter can be expressed as

$$F_{2D}(s)=\frac{(\lambda s+1)^2}{3\lambda^2s^2+3\lambda s+1}. \tag{12}$$

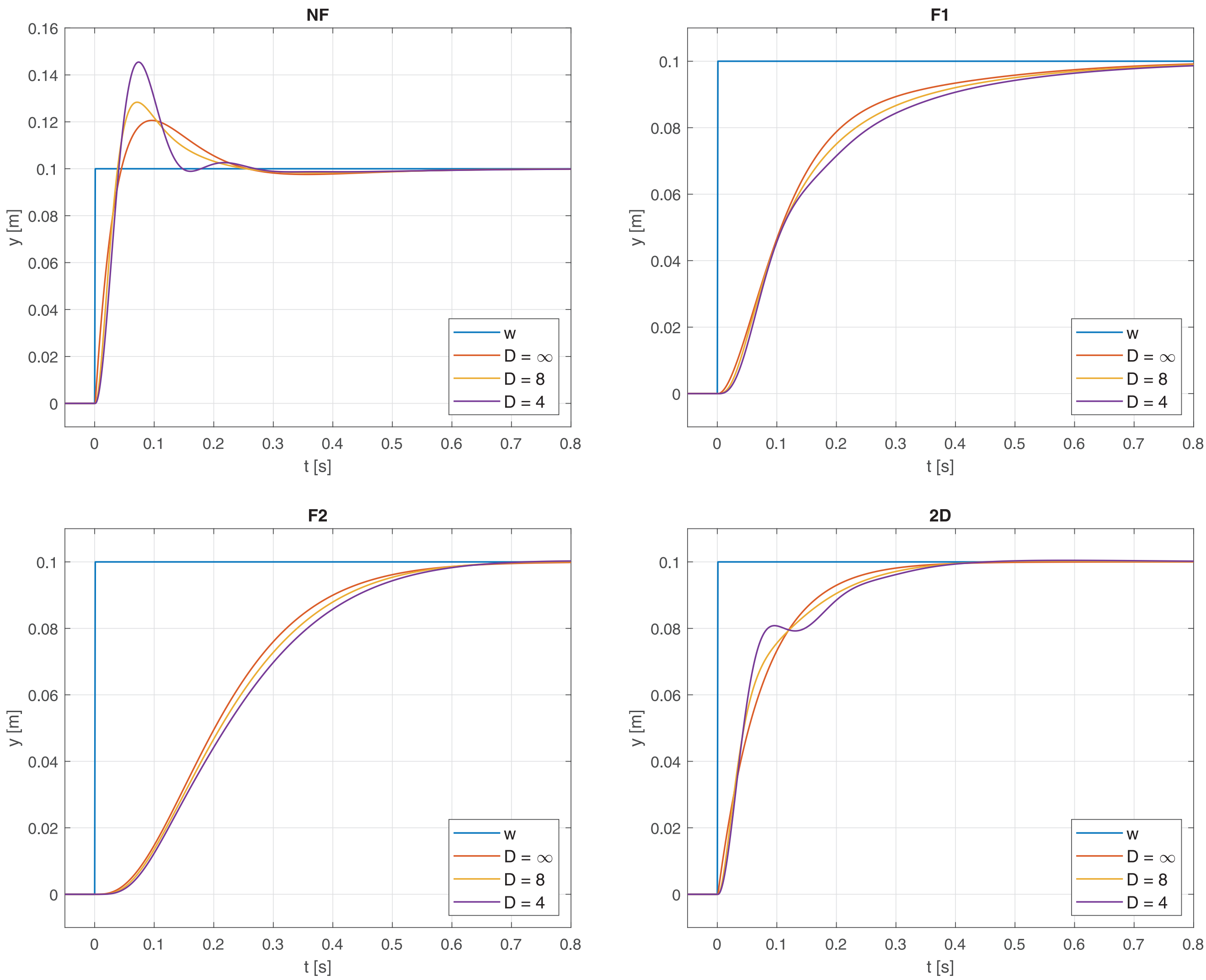

Fig. 4. Simulated step responses for the controller with derivative filtering
Rys. 4. Symulowane odpowiedzi skokowe dla regulatora z filtracją w członie różniczkującym

The 2DOF PID continuous controller that replaces F and PID in Fig. 1 operates according to the algorithm (s-dependence in U, W, Y dropped)

$$U = k_P\left(\frac{2}{3}W - Y\right) + \frac{k_I}{s}(W - Y) + k_D s\left(\frac{1}{3}W - Y\right). \tag{13}$$

Finally we quote [7] that for the integrator plus time constant plant, i.e. for the voltage driven servo, the weights $b = 2/3$, $c = 1/(9\alpha)$ with $\alpha = T_D / T_I \le 0.25$ have been obtained. So, in the continuous case, the reduction of P mode is the same for both types of the servo.

In practical implementations, the PID controller involves a filtered derivative term, here in the form $k_D s/(T_D s/D + 1)$, $T_D = k_D/k_P$, with the divisor D typically from 4 to 8. Simulated step responses of the four systems for $D = 4$, 8 and ∞ are shown in Fig. 4. Whereas the responses for F1 and F2 filters with $D = 4$, 8 remain quite close to the one with $D = \infty$ (i.e. to Fig. 2), the response for 2D filter exhibits a „saddle" in the middle part for $D = 4$. Thus small values of D adversely affect the 2DOF PID system designed without taking D into account, what may be viewed as a cost of the short settling time. As could be expected, a decrease of D increases the overshoot for the NF case.

4. Pole placement for discrete control

We shall briefly repeat essential results from [9], regarding discrete-time servo control running with a discretization step Δ. Zero-order-hold discretization of the double integrator results in

$$G_o(z) = k_o \frac{\Delta^2}{2} \frac{z+1}{(z-1)^2}. \tag{14}$$

The typical discrete PID controller has the form [3]

$$PID(z) = k_P + k_I \Delta \frac{z}{z-1} + \frac{k_D}{\Delta}\frac{z-1}{z} = \frac{k_1 z^2 - k_2 z + k_3}{z(z-1)}, \tag{15a}$$

where

$$k_1 = k_P + k_I \Delta + \frac{k_D}{\Delta}, \quad k_2 = k_P + 2\frac{k_D}{\Delta}, \quad k_3 = \frac{k_D}{\Delta}. \tag{15b}$$

Hence, the closed-loop transfer function becomes

$$G_{cl}(z) = \frac{(z+1)(K_1 z^2 - K_2 z + K_3)}{z(z-1)^3 + (z+1)(K_1 z^2 - K_2 z + K_3)} \tag{16a}$$

with

$$K_i = k_o k_i \frac{\Delta^2}{2}, \quad i = 1, 2, 3. \tag{16b}$$

Following the continuous case, we want the denominator of G_{cl} to have the triple pole r given by

$$r = e^{-p\Delta}, \quad p = \frac{1}{\lambda}. \tag{17}$$

Such a denominator may be written as $(z - r)^3(z - z_1)$ with a fourth single pole z_1. The triple pole r is provided by the following gains [9]

$$\begin{aligned} K_1 &= C(3r^3 + 8r^2 + 5r - 4), \\ K_2 &= C(3r^4 + 12r^3 + 14r^2 - 4r - 1), \\ K_3 &= C(r^2 + 4r + 7), \end{aligned} \tag{18a}$$

where

$$C = \frac{1 - r}{(r + 1)^3}. \tag{18b}$$

Having the gains one can calculate the controller settings

$$\begin{aligned} k_P &= \frac{2(K_2 - 2K_3)}{k_o \Delta^2}, \\ k_D &= \frac{2K_3}{k_o \Delta}, \\ k_I &= \frac{2(K_1 - K_2 + K_3)}{k_o \Delta^3}. \end{aligned} \tag{19}$$

Nomograms expressing the settings directly in terms of r are given in [9].

The fourth pole z_1 does not exceed r, so it does not affect system dynamics. In particular, for $r = z_1$ one obtains the unique quadruple pole $r_4 = r = z_1 = \sqrt[4]{8} - 1 \cong 0.682$. The condition $r \geq r_4$ corresponds to requirement $\Delta \leq 0.383\lambda$ on the discretization step. For $\lambda = 0.075$ s the requirement means $\Delta \leq 0.0287$ s. Here $\Delta = 0.02$ s is taken for testing (close to the limit 0.0287) to distinguish the discrete case from continuous. Step responses of the discrete implementation are quite similar to the continuous ones in Fig. 2, except that the overshoot for the NF case increases to 50 %.

5. Discrete 2DOF PID

The discrete reduced gain PID controller becomes

$$PID_{bc}(z) = bk_P + k_I \Delta \frac{z}{z - 1} + c\frac{k_D}{\Delta}\frac{z - 1}{z} = \frac{k_1' z^2 - k_2' z + k_3'}{z(z - 1)} \tag{20a}$$

with

$$\begin{aligned} k_1' &= bk_P + k_I \Delta + c\frac{k_D}{\Delta}, \\ k_2' &= bk_P + 2c\frac{k_D}{\Delta}, \\ k_3' &= c\frac{k_D}{\Delta}, \end{aligned} \tag{20b}$$

so the discrete F_{2D} filter has the form

$$F_{2D}(z) = \frac{PID_{bc}(z)}{PID(z)} = \frac{k_1' z^2 - k_2' z + k_3'}{k_1 z^2 - k_2 z + k_3} = \frac{\Delta^2}{2} k_o \frac{k_1' z^2 - k_2' z + k_3'}{K_1 z^2 - K_2 z + K_3} \tag{21}$$

by using (16b). The system transfer function $F_{2D} G_{cl}$ after cancellation of the terms $K_1 z^2 - K_2 z + K_3$ (see (16a)) and inserting the triple pole denominator becomes

$$F_{2D} G_{cl}(z) = \frac{\Delta^2}{2} k_o \frac{(z + 1)(k_1' z^2 - k_2' z + k_3')}{(z - r)^3 (z - z_1)}. \tag{22}$$

As before, two poles can be eliminated by equating $k_1' z^2 - k_2' z + k_3'$ to $k_1'(z - r)^2$. Using the expressions (20b) one obtains two linear equations for the weights b, c which yield

$$b = 2\frac{k_I}{k_P}\frac{\Delta r}{1 - r}, \quad c = \frac{k_I}{k_D}\left(\frac{\Delta r}{1 - r}\right)^2, \tag{23}$$

where k_P, k_I, k_D are given by (18) and (19). Applying some symbolic calculations [13] one finally obtains the weights as the following rational functions of the given pole r

$$\begin{aligned} b(r) &= \frac{2r(r^3 + 3r^2 + 3r - 3)}{2r^4 + 7r^3 + 9r^2 - 5r - 1}, \\ c(r) &= \frac{r^3 + 3r^2 + 3r - 3}{r(r^2 + 4r + 7)}. \end{aligned} \tag{24}$$

The nomograms $b(r)$, $c(r)$ for $r \geq r_4 \cong 0.682$ are shown in Fig. 5. For r close to 1, so for a very small step Δ, b and c approach 2/3 and 1/3, respectively, so as (10) for the continuous case. Increase of Δ decreases r, what also decreases the weights. For $r = \exp(-0.02/0.075) \cong 0.75$ used in the tests we get $b = 0.52$ and $c = 0.17$ instead of the continuous case values 2/3 and 1/3.

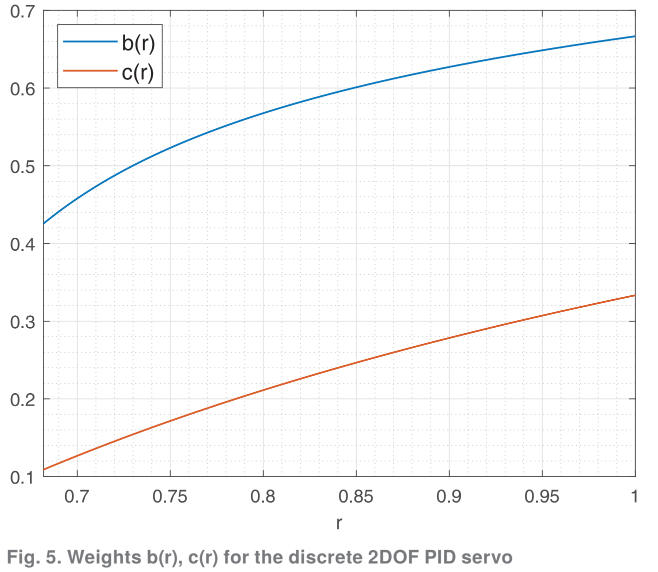

Fig. 5. Weights b(r), c(r) for the discrete 2DOF PID servo
Rys. 5. Wagi b(r), c(r) dla serwo-napędu z regulatorem 2DOF PID

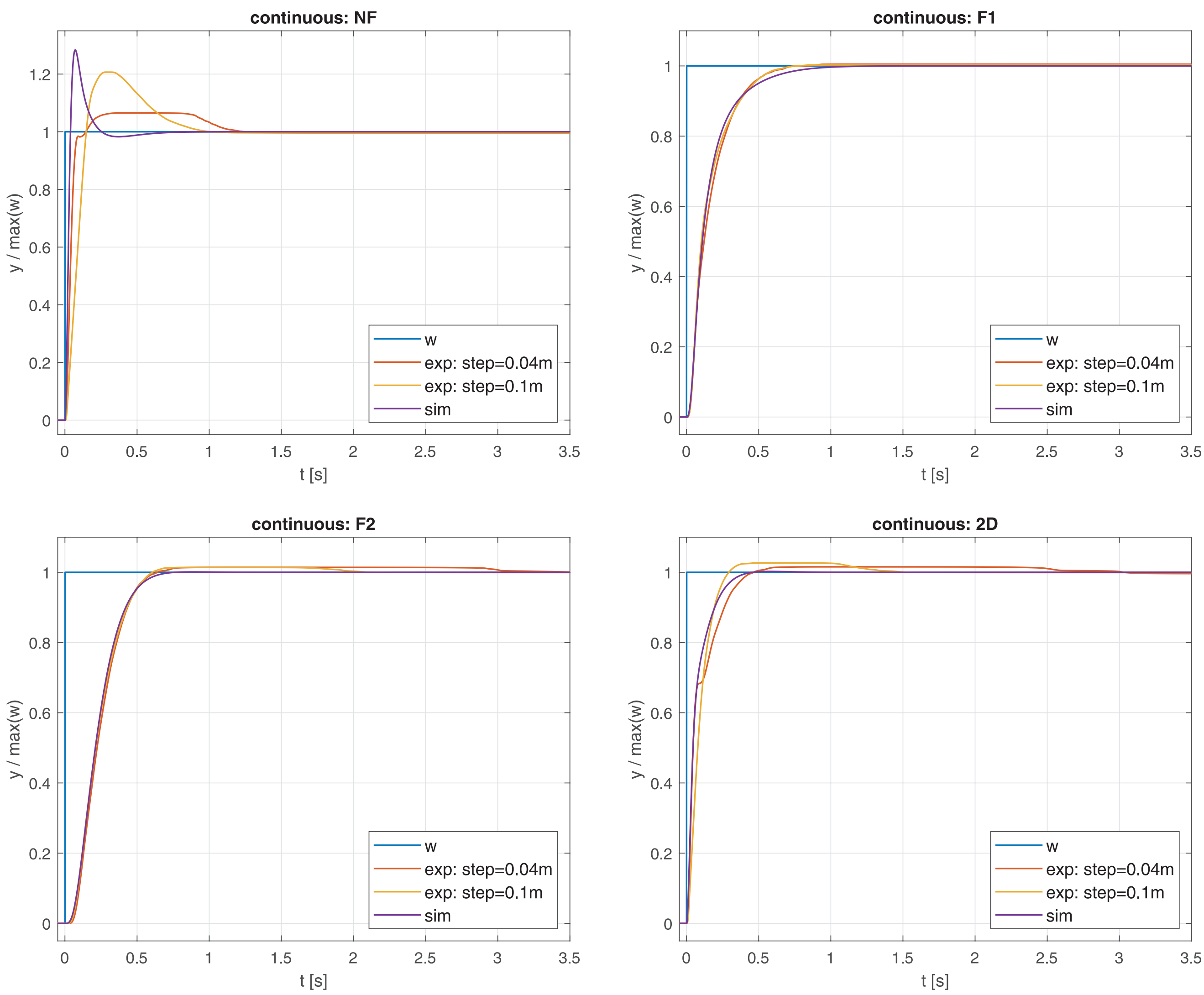

Fig. 6. Experimental and simulated responses for continuous control
Rys. 6. Odpowiedzi eksperymentalne i symulacyjne dla sterowania ciągłego

6. Experimental verification

The servo-system already presented in [9] has been used for experiments. The system consists of an ESTUN AC motor (400 W, 3000 RPM) with a dedicated servo-drive [14], ball screw linear actuator (0.5 m range) as the mechanical load, Beckhoff C6920 industrial PC [15] as a real-time controller, and standard PC with TwinCAT 3 software for programming and recording. The servo-drive and controller are connected by EtherCAT, whereas the controller and PC by Ethernet. The drive operates in the current (torque) control mode, so the plant dynamics is described by a double integrator. The continuous control has been implemented by FB_CTRL_PID function block from TwinCAT 3, whereas the filters and discrete control by means of recursive functions generated by *c2dm-ZOH* MATLAB instruction. The controller may also limit the control signal to some prescribed values. To avoid the limits in most runs the set-point for the linear actuator motion is restricted to 0.1 m or 0.04 m.

Experiments with the real system have been affected by a few phenomena not encountered in the simulations, particularly by friction of the plant, dynamics of the drive not taken into account in the k_o/s^2 model, limits of the maximum torque. Role of the PID function block additional parameters has also been examined. The experimental step responses for the tested cases are presented in Figs. 6, 7, 8, where the first two compare the responses with simulations, while the third one deals with the PID additional parameters. Details are described below.

1. Time of the step responses has been considerably extended over several closed-loop time constants λ (design data) to show that exceeding the set-point does not mean a steady-state error but rather small overshoots caused by imperfect correspondence of the system dynamics to the model. The overshoots are later slowly compensated by integral action of the controller. If small overshoots (up to 1 % of the range) are not acceptable in a practical application one may try to increase stiffness of the controller by decreasing λ, or refer to nonlinear control methods. This is however beyond the scope of this paper.
2. Comparison of the responses for the set-points 0.1 m and 0.04 m in Figs. 6, 7 highlights the nonlinear effects in the real system. Basically the responses of the systems with F1 and F2 filters are very similar, differences are small and rather random, owing to stochastic nature of friction. In the case of the NF response, the flattening is a typical effect of friction (loss of motion continuity) when the control signal goes down.

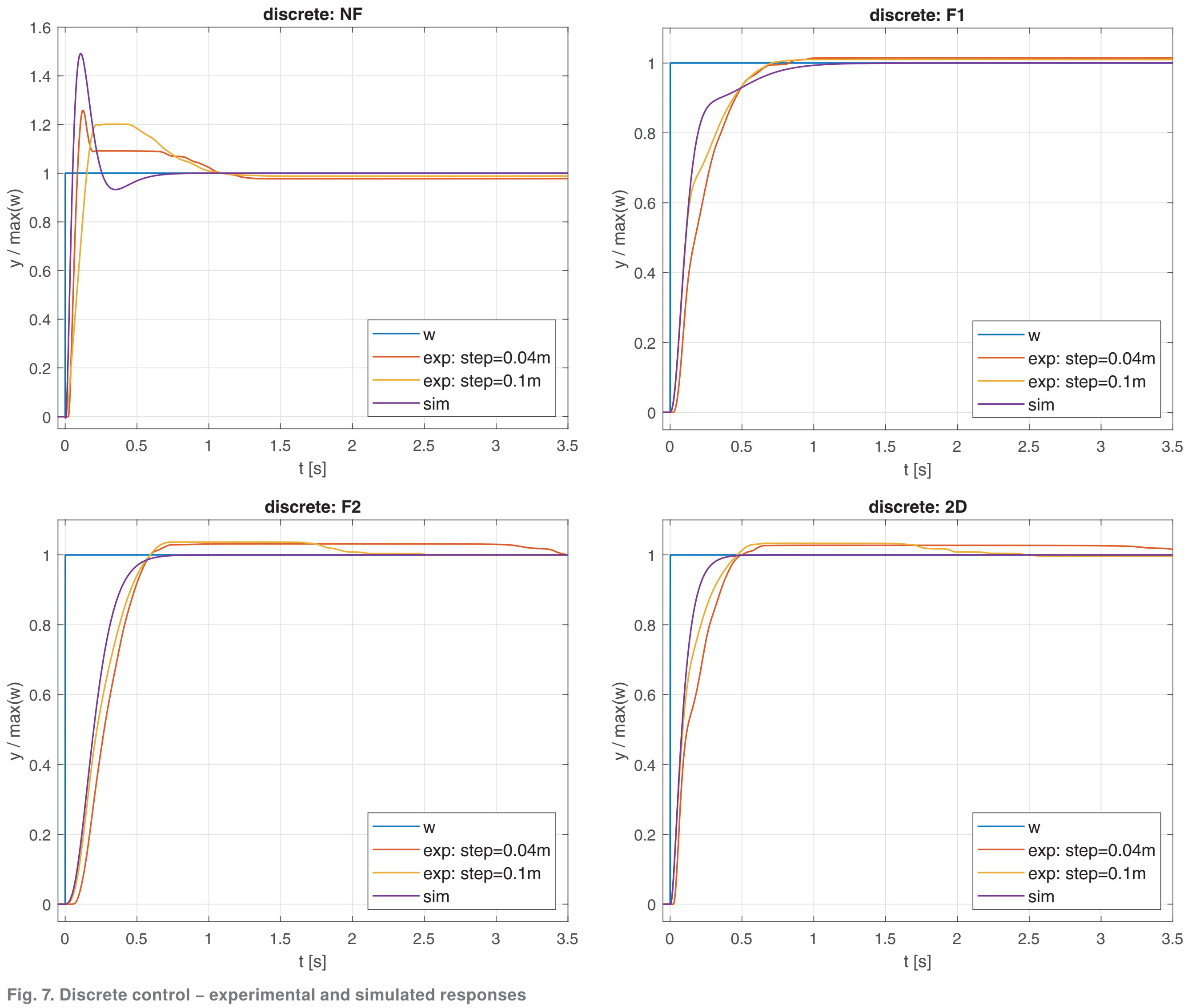

Fig. 7. Discrete control – experimental and simulated responses

Rys. 7. Sterowanie dyskretne – odpowiedzi eksperymentalne i symulacyjne

3. Responses for different values of the FB CTRL PID function block additional parameters, such as derivative filtering divisor D (see Sec. 3), control limit $uLim$, and antiwindup AWDP protection (ON or OFF) are shown in Fig. 8.
   - For the filters F1, F2 and the set-point $w = 0.04$ m, the control signal does not reach 10 % of its nominal range, so the responses do not depend on $uLim$ or AWDP. Also the divisor $D = 8$ or 4 affects the responses only slightly, hence the systems with F1 or F2 filters are fairly resistant to the changes of D.
   - Nominal value (100 %) of the control signal is reached only for the NF system what activates AWDP (if ON) and decreases the overshoot.
   - Setting $uLim$ to 10 % significantly slows down raising the responses in NF and 2D systems, increasing the overshoot in 2D. However, the control process remains stable and correct.
   - Setting $D = 4$ for NF and 2D alters considerably the system dynamics what triggers oscillations in the responses. Therefore, to get short settling time provided by 2D, small values of the derivative filtering divisor D should be avoided.

7. Summary

Current driven servos are standard components of robots, machine tools, conveyors and other equipment involving motion control. Conventional frequency designs are applied in practice to tune PID controllers for such servos. Since position of the current driven motor is described by a double integrator, a new design method has been proposed in [9] based on a given triple pole of the closed-loop system, both for continuous and discrete cases. First or second order prefilters eliminate the overshoot of a set-point step response, however with the settling time equal to eight closed-loop time constant. Here this method has been upgraded for a 2DOF PID controller, so with the weights b, c on P, D modes for the set-point input. The proposed 2DOF PID cancels out two of three poles what reduces the settling time to four time constants, i.e. twice. It turns out that for the continuous case the weights b, c are fixed numbers, independent on the plant and design data. For the discrete case the weights depend on discretization step and decrease when the step increases. Experimental verification has shown reasonably good correspondence of system responses to simulations, but also demonstrated that friction of the plant considerably slows down the responses in the vicinity of a set-point. Small values of the derivative filtering divisor D

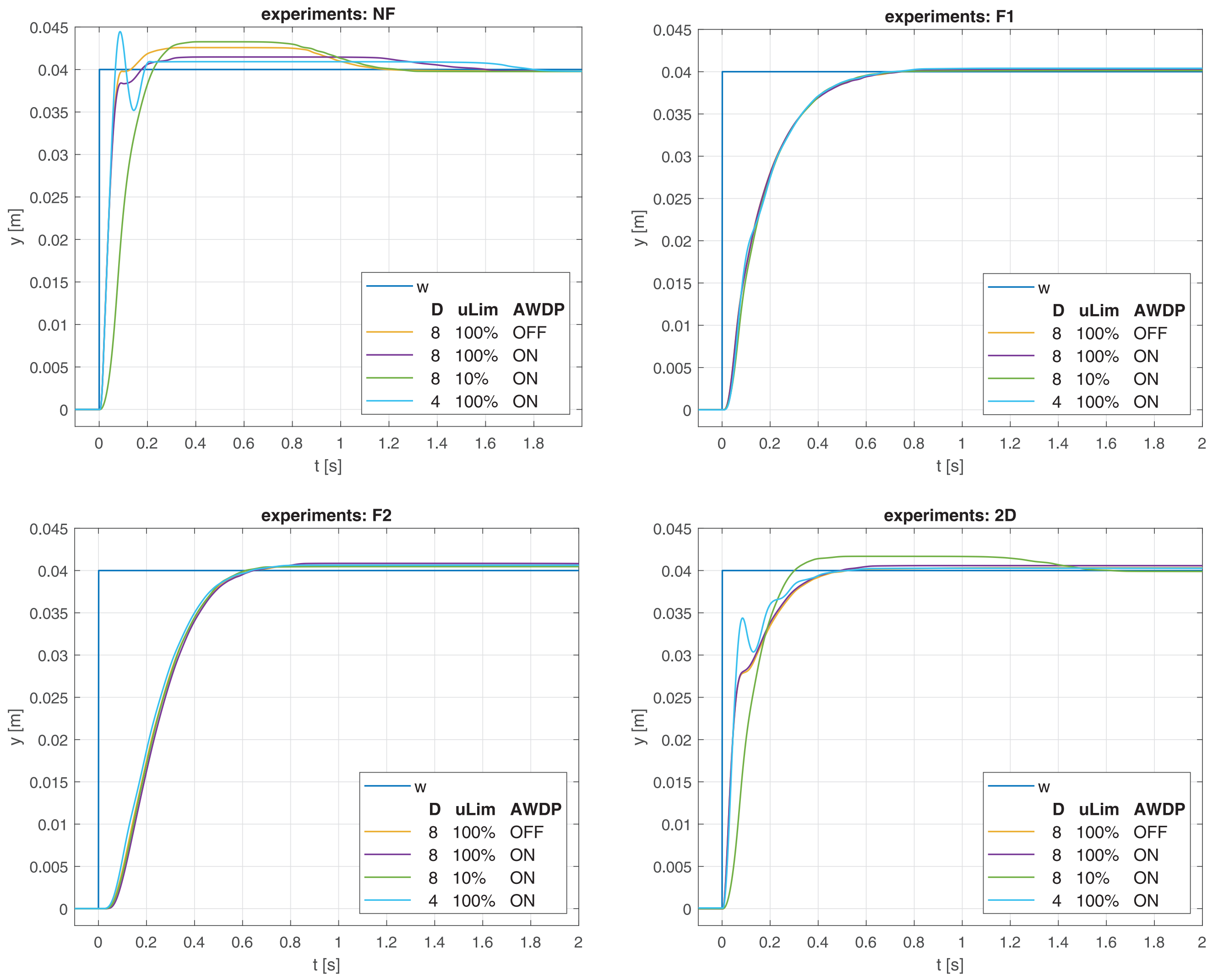

Fig. 8. Dependance of experimental responses on the PID function block parameters: D, uLim, AWDP

Rys. 8. Zależność odpowiedzi eksperymentalnych od parametrów bloku funkcyjnego PID: D, uLim, AWDP

(strong filtering) in the 2DOF PID controller should be avoided as they inflict oscillations.

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Regulator 2DOF PID dla serwomechanizmu prądowego

Streszczenie: Przedstawiono metodę projektowania ciągłego i dyskretnego regulatora 2DOF PID dla serwonapędu sterowanego prądowo, opisanego jako podwójny integrator. Stała czasowa układu zamkniętego jest jedyną daną projektową. Wynikiem metody są reguły dla trzech nastaw PID oraz dwóch wag b, c w 2DOF, które redukują wzmocnienia torów P, D dla wielkości zadanej. Nastawy PID dają potrójny rzeczywisty biegun układu zamkniętego, co zapewnia gładkie przebiegi. Wagi b, c eliminują przeregulowanie odpowiedzi na skok wielkości zadanej. W przypadku sterowania ciągłego wagi są ustalonymi liczbami, niezależnymi od danych projektowych. W przypadku dyskretnym wagi maleją przy wzroście kroku próbkowania. Eksperymenty laboratoryjne potwierdzają dobrą zgodność odpowiedzi systemu z symulacjami, ale także demonstrują efekty tarcia, gdy wyjście jest bliskie wartości zadanej. Zbadano również rolę dodatkowych parametrów przemysłowego bloku funkcyjnego PID.

Słowa kluczowe: serwomechanizm prądowy, podwójny integrator, 2DOF PID, filtr wielkości zadanej, lokalizacja biegunów, efekty tarcia

Prof. Leszek Trybus, DSc PhD Eng.
trybus@prz.edu.pl
ORCID: 0000-0002-1415-3679

He graduated from AGH University (1970), where also received PhD, DrSC and professor title. Since graduation he has been working at the Department of Computer and Control Engineering, Rzeszow University of Technology. Author of over 200 publications, supervisor of 20 PhD thesis. Since 30 years member of Control and Robotics Committee, Polish Academy of Sciences. His R&D interests cover programmable controllers, distributed control systems and applications of control theory.

Andrzej Bożek, PhD Eng.
abozek@prz.edu.pl
ORCID: 0000-0003-3015-7474

He received MSc in Electrical Engineering (2008) and PhD in Computer Engineering (2015) at the Rzeszow University of Technology. He has been working at the Department of Computer and Control Engineering, Rzeszow University of Technology, since 2008, currently as an assistant professor. His research interests lie in the fields of discrete optimization, task scheduling, mechatronics and control algorithms.