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A new control concept based on process self-feedback

Damir Vrančić*,**, Mikuláš Huba**, Pavol Bisták** and Paulo Moura Oliveira*****

**J. Stefan Institute, Ljubljana, Slovenia; ****Faculty of Industrial Engineering, Novo mesto, Slovenia (e-mail: damir.vrancic@ijs.si).*

***Institute of Automotive Mechatronics, Faculty of Electrical Engineering and Information Technology,*

Slovak University of Technology in Bratislava, SK-812 19 Bratislava, Slovakia (e-mail: mikulas.huba@stuba.sk, pavol.bistak@stuba.sk)

**** INESC-TEC, UTAD-Universidade de Trás-os-Montes e Alto Douro, Vila Real, Portugal, (e-mail: oliveira@utad.pt)*

Abstract: The paper presents a new control concept based on the control of the process moment instead of the process states or the process output signal. The control scheme is based on separate control of reference tracking and disturbance rejection. The tracking control is achieved by additionally feeding the input of the process model by the scaled output signal of the process model. The advantage of such self-feedback is that the final state of the process output can be calculated analytically and used for control instead of the actual process output value. The disturbance rejection, including model imperfections, is controlled by feeding back the filtered difference between the process output and the model output to the process input. The tracking and disturbance rejection performance is simply controlled by two user-defined gains. Several examples have shown that the new control method provides very good and stable tracking and disturbance rejection performance.

Keywords: self-feedback control, tracking, disturbance-rejection, process model, 2-DOF controller.

1. INTRODUCTION

The classic feedback strategy is based on comparison between the desired and the actual process output signals (or process states) and feeding back this difference to the controller which manipulates the process input signal (Åström and Hägglund, 1995; Vilanova and Visioli, 2012; Visioli, 2006). The controller tuning is based either directly on the process input and output signals in time-domain or frequency-domain (non-parametric tuning) or indirectly on identified process model (parametric tuning).

While there are many tuning methods developed so far, an inherent deficiency of the mentioned control approaches is that they are based on the measured process output (or state) value. Therefore, any control action from the controller, due to the process dynamics, takes some time before it results in the changed process output value.

In order to decrease the mentioned lag, the predicted process output value can be used for control instead of the actual one (Vilanova and Visioli, 2012; Schwenzer *et al*., 2021). This approach is used in predictive control algorithms (where the process output is predicted several steps ahead) and, in some extent, also by the PID controller (it is essentially the PI controller controlling the predicted process output by time interval equivalent to the derivative time constant).

The proposed approach herein is to extent the mentioned prediction even further by estimating and controlling the final process output value. This can be done by monitoring the process input and output signals.

2. THE MAIN PRINCIPLE OF PROCESS SELF-FEEDBACK

The innovative control principle is based on measuring the so-called process moment (A). The received process moment can be calculated by integrating the difference between the scaled process input (u) and output (y) signal (Åström and Hägglund, 1995; Vrančić and Huba, 2021):

$$A = \int_{\tau=0}^{\tau=t} (K_{PR}u(\tau) - y(\tau))d\tau \tag{1}$$

where K_{PR} is the process steady state gain. The nominal stable and strictly proper process transfer function in Laplace form is given below:

$$G_P(s) = K_{PR}\frac{(1 + b_1 s + b_2 s^2 + \cdots)e^{-sT_{delay}}}{(1 + a_1 s + a_2 s^2 + \cdots)} \tag{2}$$

The final process output value depends on the measured area A and the chosen process as follows

$$y(\infty) = \frac{A}{A_1} \tag{3}$$

where A_1 stands for the first process moment/area, which is defined as (Vrančić *et al*., 2001; Vrančić and Huba, 2021):

$$A_1 = K_{PR}(a_1 - b_1 + T_{delay}) \tag{4}$$

Therefore, the proposed control method estimates and controls the final process output signal (3) by measuring the integral A (1). In this case, the process delays and lags can be avoided in the main control loop. This can make the entire closed-loop control more robust and the process input limitations are implicitly taken into account without the need for additional anti-windup protection.

To successfully control the process in practice, a separate control of setpoint tracking and disturbance rejection has to be designed.

3. REFERENCE TRACKING CONTROL BASED ON MOMENTS

According to equations (1) and (3), the area A and therefore the final process output value (3) does not change when the process input signal equals to:

$$u = u_F = \frac{y}{K_{PR}} \quad (5)$$

Therefore, the final value of the process output can be controlled by the block scheme given in Fig. 1.

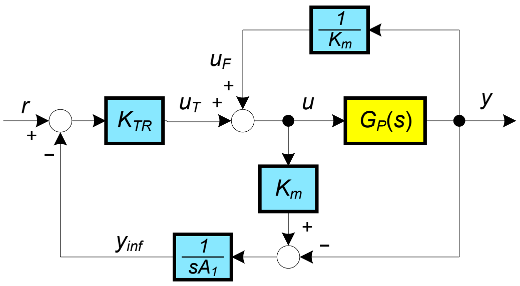

Fig. 1. The basic principle of the closed-loop tracking control based on moments. Signal y_{inf} represents the estimated process output signal in steady-state. Consider that parameter K_m equals to K_{PR}.

The process input signal u is calculated as the sum of the self-feedback signal u_F and the control signal u_T, where u_T is the amplified (by K_{TR}) difference between the desired (r) and the estimated (y_{inf}) final process output value. Note that Fig. 1 only depicts the main control principle, since the actual control scheme is slightly modified, as will be shown later.

Let us illustrate the proposed control principle on the following process:

$$G_{P1}(s) = \frac{e^{-0.2s}}{(1+s)^2} \quad (6)$$

The closed-loop control configuration according to Fig. 1 was used. The parameters K_m=K_{PR}=1 and, according to (4), A_1=2.2. The closed-loop responses were obtained with different gains K_{TR} (from K_{TR}=1 to K_{TR}=30). In Fig. 2 it can be seen that by increasing gain K_{TR}, the closed-loop response becomes faster. However, the speed does not improve significantly for $K_{TR} \geq 10$. It is clear that the closed-loop responses are all smooth without visible overshoots for any chosen K_{TR}. Stability does not seem to be deteriorated by increased gain K_{TR}.

Naturally, if the estimated process moment/area A_1 does not correspond to the actual one, an error of the closed-loop steady-state value can be expected. Moreover, if the estimated process steady-state gain (K_m) is even slightly smaller or larger than the actual gain (K_{PR}), then the final process closed-loop output value might steadily increase or decrease, instead.

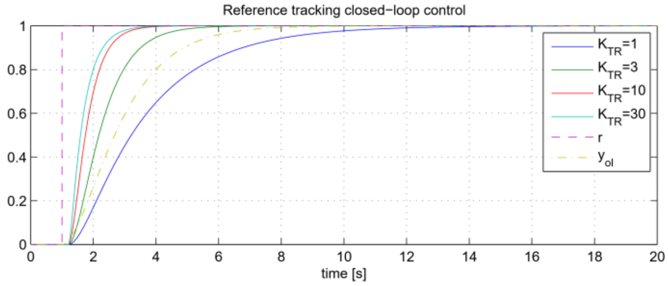

Fig. 2. The reference tracking for process G_{P1}. Signal r is the reference and y_{ol} is the open-loop response.

In order to alleviate the mentioned problems, the control algorithm should control the *process model* instead of the actual process. The control signal, fed to the model, should be fed to the actual process as well. Such approach has another advantage, since only the non-delayed part of the process model can be controlled and therefore the closed-loop performance can be increased, as will be shown in section 4.

However, a mismatch between the actual process and the model, as well as the process disturbances, may still produce some steady-state error. This can be avoided by adding the disturbance-rejection controller.

4. DISTURBANCE REJECTION CONTROL

Disturbance rejection control has two goals. The first is to reject process disturbances and the second one is to smooth out the differences between the actual process and the process model.

The proposed disturbance-rejection controller is given in Fig. 3. The $G_M(s)$ represents the second-order process model with delay:

$$G_M(s) = K_m \frac{e^{-sT_{delaym}}}{1 + a_{1m}s + a_{2m}s^2} = G_{M0}(s)e^{-sT_{delaym}} \quad (7)$$

where $G_{M0}(s)$ is the process model transfer function without delay. $G_F(s)$ is the filter transfer function containing the inverse process model without delay. The suggested filter transfer function in Fig. 3 is:

$$G_F(s) = \frac{1 + a_{1m}s + a_{2m}s^2}{K_m(1 + T_{FDR}s)^2} \quad (8)$$

where the filter time constant is:

$$T_{FDR} = \sqrt{\frac{a_{2m}}{K_{DR}K_m}} \quad (9)$$

Parameter K_{DR} represents the desired high-frequency gain of the disturbance rejection controller and is the tuning parameter for disturbance rejection. If the model and the process match perfectly, the estimated disturbance signal d_m becomes:

$$d_m(s) = \frac{d(s)}{(1 + T_{FDR}s)^2} \tag{10}$$

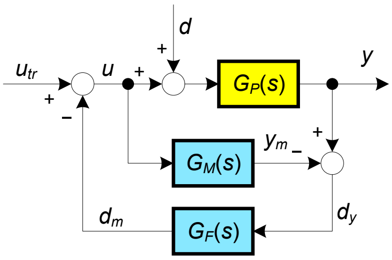

Fig. 3. The basic principle of the disturbance-rejection control. Signal u_{tr} represents the signal of the tracking controller (considered to be constant (e.g. 0) for pure disturbance-rejection control).

The higher is the gain K_{DR}, the faster is compensation (d_m) of disturbance (d) and the higher is the high-frequency noise level in signal d_m. Naturally, if parameter a_{2m}=0, the first-order filter is used in denominator of (8) and $T_{FDR}=a_{1m}/(K_{DR}K_m)$.

The second-order process model (7) can be estimated from the process time-responses or directly from the higher-order actual process by the moment method (Vrečko *et al.*, 2001; Kos *et al.*, 2021). In our case the process model corresponds to the actual process. The disturbance rejection has been tested on the same process model as before (6) by choosing K_{DR} from 1 to 30. The closed-loop responses are given in Fig. 4.

The values of $K_{DF} \geq 10$ slightly increase the disturbance rejection performance, but also increase controller high-frequency noise. Naturally, if the actual process and the process model differ, high values of K_{DR} may also destabilise the closed-loop response.

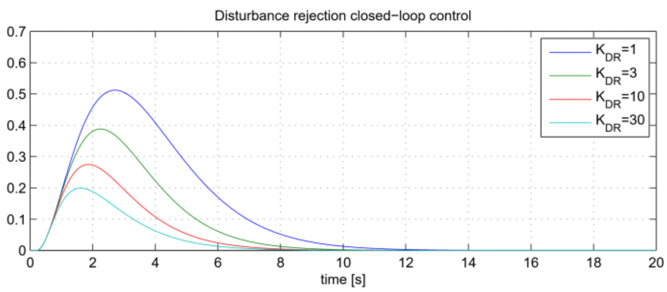

Fig. 4. The disturbance-rejection responses for process G_{P1}. Disturbance d=1 appears at t=0.

5. CLOSED-LOOP CONTROL BASED ON MOMENTS

When combining tracking controller with disturbance-rejection controller, the overall controller scheme is obtained (see Fig. 5). Note that the reference tracking controller controls the process model without delay (G_{M0}(s)) instead of the actual process. Also note that in practical realisation, the tracking controller block scheme can be simplified. Due to the limited space, the stability, noise and constrained control analyses are also not presented herein, but will be carried out in more detail in the extended journal version.

The controller will be tested on the following process models:

$$G_{P2}(s) = \frac{e^{-2s}}{(1+s)^2} \tag{11}$$

$$G_{P3}(s) = \frac{1}{(1+s)^3} \tag{12}$$

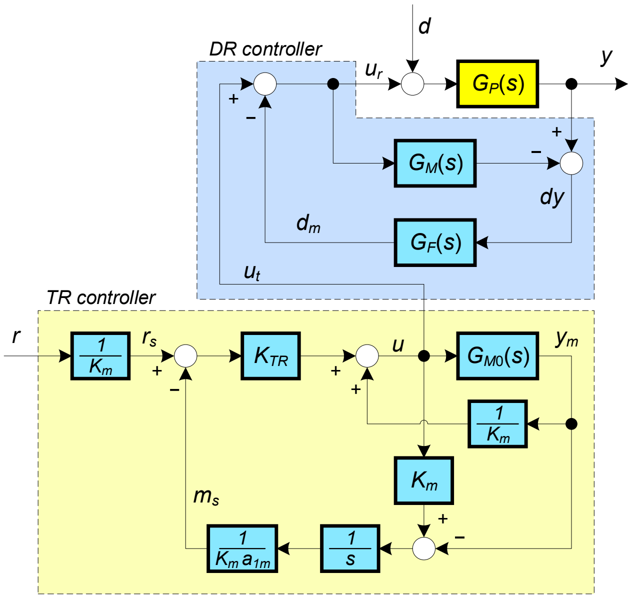

Fig. 5. Block scheme of the entire controller.

The process model for the process G_{P2}(s) matches the actual process, while the second-order process model for G_{P3}(s), according to Vrečko *et al.* (2001) and Kos *et al.* (2021), is:

$$G_{M3}(s) = \frac{e^{-0.416s}}{1 + 2.584s + 1.839s^2} \tag{13}$$

The proposed method was compared to Magnitude Optimum Multiple Integration (MOMI) tuning method for PID controllers, since it offers fast and stable closed-loop responses (Vrančić *et al.*, 2001, Vrančić and Huba, 2021). The PID controller transfer function is:

$$G_{PID}(s) = \frac{K_I + K_P s + K_D s^2}{s(1 + sT_F)} \tag{14}$$

where the PID filter time constant was chosen as T_F=0.05 s. The PID controller parameters were:

$$G_{P2}(s): K_I = 0.306, K_P = 0.739, K_D = 0.486$$
$$G_{P3}(s): K_I = 0.869, K_P = 2.15, K_D = 1.40 \tag{15}$$

The tracking gain of the proposed method for both processes was K_{TR}=10. In order to perform fair comparison, the disturbance rejection gain K_{DR} was chosen to be similar to high-frequency noise amplification (K_D/T_F) of the PID controller. We, therefore, chose K_{DR}=10 for G_{P2}(s) and K_{DR}=30 for G_{P3}(s). The closed-loop responses are shown in Figs. 6 and 7.

It can be seen that the proposed method gives favourable closed-loop tracking and disturbance-rejection responses. Moreover, the initial kick of the control output signal is much smaller with the proposed method for G_{P3}. Tracking and control responses can be adjusted separately by two parameters. When the process and the model differ, the gain K_{DR} should be limited. For example, the actual process G_{P3} (12) and the model G_{M3} (13) differ considerably and the closed-loop would become unstable if K_{DR} is higher than about 10000. However, such high gains should never be used in practice due to the exaggerated controller noise.

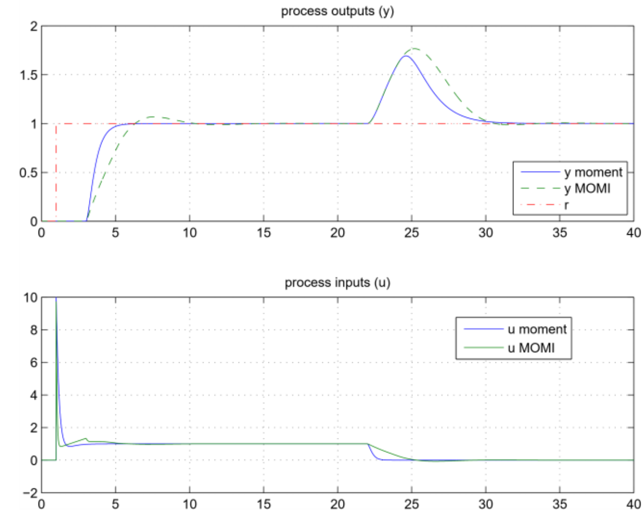

Fig. 6. The process G_{P2}(s) output and input signals when using the proposed controller (y moment, u moment) and PID controller tuned by the MOMI method (y MOMI, u MOMI)

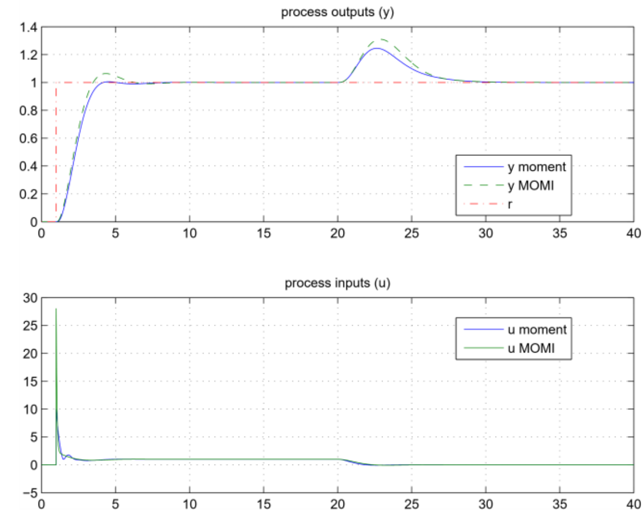

Fig. 7. The process G_{P3}(s) output and input signals when using the proposed controller (y moment, u moment) and PID controller tuned by the MOMI method (y MOMI, u MOMI)

6. CONCLUSIONS

The proposed method controls the measured process moment (the anticipated final steady-state) instead of the process output. It is shown that this approach significantly stabilises the control loop for reference tracking. The disturbance rejection was achieved by feeding back the inversed estimated disturbance signal.

The proposed control scheme has two independent gains, one for reference tracking (K_{TR}) and the other one for disturbance rejection (K_{DR}). The separation of tracking and disturbance-rejection responses is ideal for perfect process models. The user-defined gain K_{DR} also sets the amplification of the high-frequency process noise.

The examples on a few different process models showed that the proposed control method may be suitable for control of stable processes. The comparison to MOMI tuning method with PID controller showed that the proposed method can also have faster closed-loop responses with smaller overshoots.

In our further work we are planning to make stability, noise, robustness and constrained control analysis. Another direction of future research will be additional optimisation of disturbance rejection response and control of integrating processes.

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REFERENCES

Åström, K.J. Hägglund, T. (1995). *PID Controllers: Theory, Design, and Tuning*, 2nd ed.; Instrument Society of America: Pittsburgh, PA, USA.

Kos, T., Vrančić, D. (2021). A Simple Analytical Method for Estimation of the Five-Parameter Model: Second-Order with Zero Plus Time Delay. Mathematics, Vol. 9, 1707.

Schwenzer, M., Ay, M., Bergs, T. *et al.* (2021). Review on model predictive control: an engineering perspective. Int J. Adv. Manuf. Technol. 117, pp. 1327–1349.

Vilanova, R., Visioli, A. (2012). *PID Control in the Third Millennium: Lessons Learned and New Approaches*; Springer: London, UK.

Visioli, A. (2006). *Practical PID Control*; Springer: London, UK.

Vrančić, D., Strmčnik, S., Juričić, Đ. (2001). A magnitude optimum multiple integration method for filtered PID controller. *Automatica*, Vol. 37, pp. 1473–1479.

Vrančić, D., Huba, M. (2021). High-Order Filtered PID Controller Tuning Based on Magnitude Optimum. *Mathematics*, Vol. 9, 1340.

Vrečko D., Vrančić D., Juričić Đ. and Strmčnik S. (2001). A new modified Smith predictor: the concept, design and tuning. *ISA transactions*, vol. 40, pp. 111–121.