IDENTIFICATION AND DIGITAL CONTROL OF HIGHER-ORDER PROCESSES USING PREDICTIVE STRATEGY

Vladimír Bobál, Marek Kubalčík, Petr Chalupa and Petr Dostál
Tomas Bata University in Zlín
Faculty of Applied Informatics
Nad Stráněmí 4511
760 05 Zlín
Czech Republic
E-mail: bobal@fai.utb.cz

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ABSTRACT

In technical practice often occur higher order processes when a design of an optimal controller leads to complicated control algorithms. One of possibilities of control of such processes is their approximation by lower-order model with time-delay (dead time). The contribution is focused on a choice of a suitable experimental identification method and a suitable excitation input signals for an estimation of process model parameters with time-delay. The further contribution is design of an algorithm for digital control of high-order processes which are approximated by second-order model of the process with time-delay. The designed control algorithms are based on a predictive control strategy. The controller's algorithm uses the digital modification of the Smith Predictor (SP). The program system MATLAB/SIMULINK was used for simulation verification of these algorithms.

INTRODUCTION

Some technological processes in industry are characterized by high-order dynamic behaviour or large time constants and time-delays. For control engineering, such processes can often be approximated by the FOTD (first-order-time-delay) model. Time-delay in a process increases the difficulty of controlling it. However using the approximation of higher-order process by lower-order model with time-delay provides simplification of the control algorithms.

Let us consider a continuous-time dynamical linear SISO (single input $u(t)$ – single output $y(t)$) system with time-delay T_d. The transfer function of a pure transportation lag is $e^{-T_d s}$ where s is a complex variable. Overall transfer function with time-delay is in the form

$$G_d(s) = G(s)e^{-T_d s} \tag{1}$$

where $G(s)$ is the transfer function without time-delay.

Processes with time-delay are difficult to control using standard feedback controllers. When a high performance of the control process is desired or the relative time-delay is very large, a predictive control strategy must be used. The predictive control strategy includes a model of the process in the structure of the controller. The first time-delay compensation algorithm was proposed by (Smith 1957). This control algorithm known as the Smith Predictor (SP) contained a dynamic model of the time-delay process and it can be considered as the first model predictive algorithm.

Historically first modifications of time-delay algorithms were proposed for continuous-time (analogue) controllers. On the score of implementation problems, only the discrete versions are used in practice in this time.

The digital pole assignment SP was designed using a polynomial approach in (Bobál et al. 2011a). The design of this controller was extended by a method for a choice of a suitable pole assignment of the characteristic polynomial. The designed digital SP was verified by simulation control of the fifth-order system which was identified by a second-order model with time-delay.

IDENTIFICATION OF TIME-DELAY PROCESSES

In this paper, the time-delay is obtained separately from an off-line identification using the least squares method (LSM). The measured process output $y(k)$ is generally influenced by noise. These nonmeasurable disturbances cause errors $\boldsymbol{e}$ in the determination of model parameters and therefore real output vector is in the form

$$\boldsymbol{y} = \boldsymbol{F\Theta} + \boldsymbol{e} \tag{2}$$

It is possible to obtain the LSM expression for calculation of the vector of the parameter estimates

$$\hat{\boldsymbol{\Theta}} = \left(\boldsymbol{F}^T\boldsymbol{F}\right)^{-1}\boldsymbol{F}^T\boldsymbol{y} \tag{3}$$

The matrix $\boldsymbol{F}$ has dimension (N-n-d, $2n$), the vector $\boldsymbol{y}$ (N-n-d) and the vector of parameter model estimates

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$\hat{\boldsymbol{\Theta}}$ (2*n*). *N* is the number of samples of measured input and output data, *n* is the model order.
Equation (3) serves for calculation of the vector of the parameter estimates $\hat{\boldsymbol{\Theta}}$ using *N* samples of measured input-output data. The individual vectors and matrices in Equations (2) and (3) have the form

$$\boldsymbol{F} = \begin{bmatrix} -y(n+d) & -y(n+d-1) & \cdots & -y(d+1) & u(n) & u(n-1) & \cdots & u(1) \\ -y(n+d+1) & -y(n+d) & \cdots & -y(d+2) & u(n+1) & u(n) & \cdots & u(2) \\ \vdots & \vdots & \cdots & \vdots & \vdots & \vdots & \cdots & \vdots \\ -y(N-1) & -y(N-2) & \cdots & -y(N-n) & u(N-d-1) & u(N-d-2) & \cdots & u(N-d-n) \end{bmatrix} \quad (4)$$

$$\boldsymbol{y}^T = \begin{bmatrix} y(n+d+1) & y(n+d+2) & \cdots & y(N) \end{bmatrix} \quad (5)$$

$$\boldsymbol{e}^T = \begin{bmatrix} \hat{e}(n+d+1) & \hat{e}(n+d+2) & \cdots & \hat{e}(N) \end{bmatrix} \quad (6)$$

$$\hat{\boldsymbol{\Theta}}^T = \begin{bmatrix} \hat{a}_1 & \hat{a}_2 & \cdots & \hat{a}_n & \hat{b}_1 & \hat{b}_2 & \cdots & \hat{b}_n \end{bmatrix} \quad (7)$$

Most of higher-order industrial processes can be approximated by a model of reduced order with pure time-delay. Let us consider the following second order linear model with a time-delay

$$G_d\left(z^{-1}\right) = \frac{B\left(z^{-1}\right)}{A\left(z^{-1}\right)} z^{-d} = \frac{b_1 z^{-1} + b_2 z^{-2}}{1 + a_1 z^{-1} + a_2 z^{-2}} z^{-d} \quad (8)$$

The term z^{-d} represents the pure discrete time-delay. The time-delay is equal to dT_0 where T_0 is the sampling period.
Our experience proved that quality of system identification when the higher-order process is identified by the lower-order model is very dependent on the choice of an input excitation signal $u(k)$. The best results were achieved using a Random Gaussian Signal (RGS). The MATLAB code

```
u=idinput(N,'rgs',[0 B],[Umin, Umax])
```

generates an RGS of the length N, where [0 B] determines the frequency passband. Umin, Umax defines the minimum and maximum values of u. The signal level is such that Umin is the mean value of the signal, minus one standard deviation, while Umax is the mean value plus one standard deviation. Gaussian white noise with zero mean and variance one is thus obtained for levels [-1, 1], which are also the default values.

Consider that model (8) is the deterministic part of the stochastic process described by the ARX (regression) model

$$\begin{aligned} y(k) &= -a_1 y(k-1) - a_2 y(k-2) + \\ &+ b_1 y(k-1-d) + b_2 y(k-2-d) + e_s(k) \end{aligned} \quad (9)$$

where $e_s(k)$ is the random nonmeasurable component.
The vector of parameter model estimates is computed by solving equation (3)

$$\hat{\boldsymbol{\Theta}}^T(k) = \begin{bmatrix} \hat{a}_1 & \hat{a}_2 & \hat{b}_1 & \hat{b}_2 \end{bmatrix} \quad (10)$$

and is used for computation of the prediction output.

$$\begin{aligned} \hat{y}(k) &= -\hat{a}_1 y(k-1) - \hat{a}_2 y(k-2) + \\ &\hat{b}_1 u(k-1-d) + \hat{b}_2 u(k-2-d) \end{aligned} \quad (11)$$

The quality of identification can be considered according to error, i.e. the deviation

$$\hat{e}(k) = y(k) - \hat{y}(k) \quad (12)$$

In this paper, the error was used for suitable choice of the time-delay dT_0. The LSM algorithm (3) – (7) is computed for several time-delays dT_0 and the suitable time-delay is chosen according to quality of identification based on the prediction error (12).

Stable process

Consider the following fifth order linear system

$$G_A(s) = \frac{2}{(s+1)^5} = \frac{2}{s^5 + 5s^4 + 10s^3 + 10s^2 + 5s + 1} \quad (13)$$

System (13) was identified by discrete model (11) using off-line LSM (3) – (6) for different time-delay dT_0; $T_0 = 0.5\,\mathrm{s}$. A criterion of the identification quality is based on sum of squares of error

$$J_{\hat{e}^2}(d) = \sum_{k=1}^{N} \hat{e}^2(k) \quad (14)$$

This criterion represents accuracy of process identification. From Fig. 1, it is obvious that value of the criterion (14) decreases when the number of time-delay steps *d* increases in the interval $d \in [0, 4]$ (it is obvious, that criterion (14) has minimum for some higher *d*). This is caused by the fact that the increase of the number of time-delay steps in the above-mentioned interval improves estimation of the static gain

$$\hat{K}_g = \frac{\hat{b}_1 + \hat{b}_2}{1 + \hat{a}_1 + \hat{a}_2} \quad (15)$$

The difference between estimates of the static gain $\hat{K}_g$ of the discrete model (8) and the continuous-time model (13) plays important role for the quality of identification because the identification time was relatively long (300 s) with regard to the response time (about 15 s).

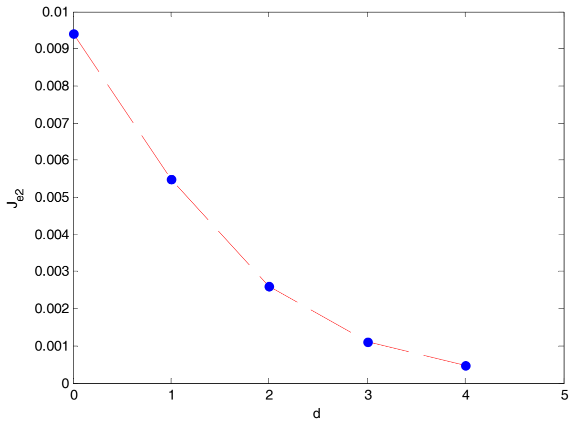

Figure 1: Criterion of Quality Identification for $d \in [0,4]$

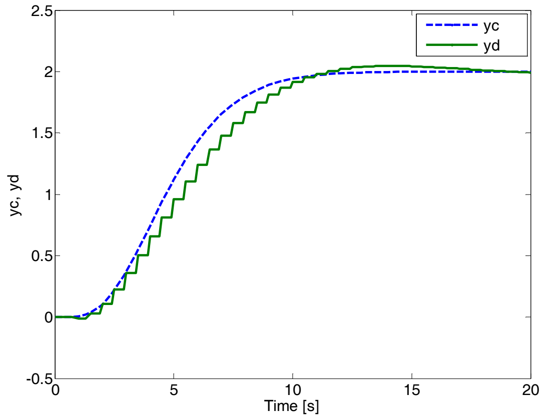

Figure 2: Comparison of step responses yc, yd for $d = 0$ (process (13))

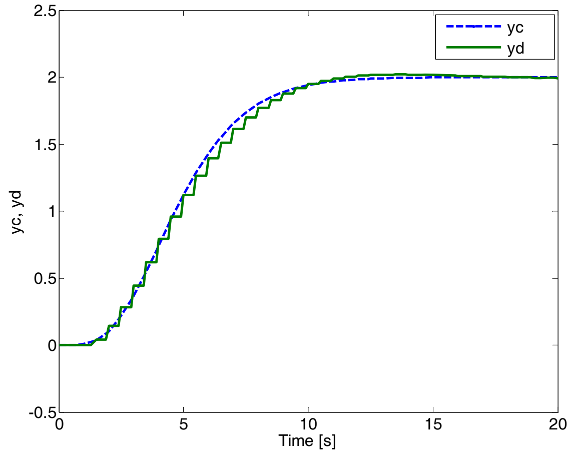

Figure 3: Comparison of step responses yc, yd for $d = 2$ (process (13))

$$G_A\left(z^{-1}\right) = \frac{0.0424z^{-1} + 0.0296z^{-2}}{1 - 1.6836z^{-1} + 0.7199z^{-2}} z^{-d} \qquad (16)$$

for sampling period $T_0 = 0.5$ s (16) with different d are shown in Figs. 2 – 4, where yc is the step response of the model (13) and yd are step responses of the discrete models (16) for individual numbers of time-delay steps d.

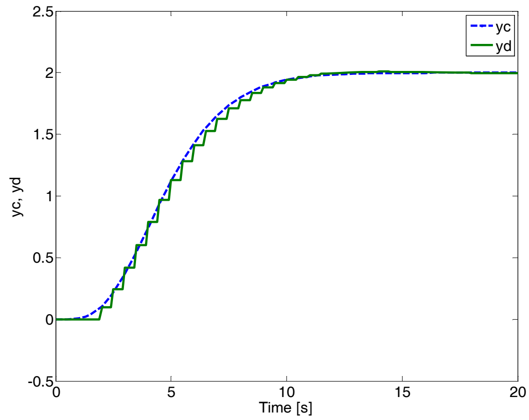

Figure 4: Comparison of step responses yc, yd for $d = 3$ (process (13))

From Figs. 2 – 4 it results that a suitable model (16) for the design of the predictive controller is the model with $d = 2$. Its structure is simple and it relatively well approximates the dynamic behaviour of the continuous-time model (13).

Non-minimum phase process

Consider the following fifth-order linear system with non-minimum phase

$$G_B(s) = \frac{2(1 - 5s)}{s^5 + 5s^4 + 10s^3 + 10s^2 + 5s + 1} \qquad (17)$$

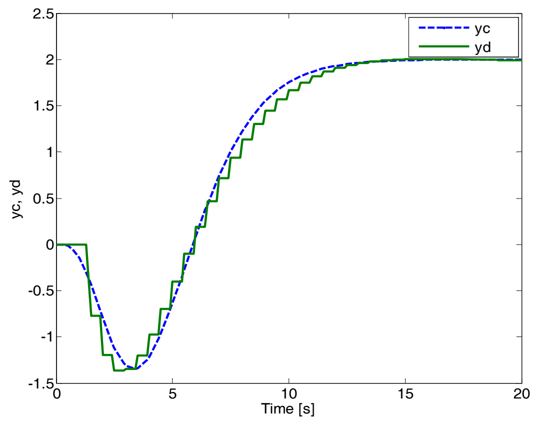

Figure 5: Comparison of step responses yc, yd for $d = 2$ (process (17))

Process (17) was identified by model (8) with a time-delay $d = 2$ and sampling period $T_0 = 0.5$ s. The discrete model is in the following form

$$G_B\left(z^{-1}\right) = \frac{-0.7723z^{-1} + 0.8514z^{-2}}{1 - 1.6521z^{-1} + 0.6920z^{-2}} z^{-2} \tag{18}$$

The comparison of the step responses of the continuous-time model (17) and the discrete model (18) is shown in Fig. 5.

DIGITAL SMITH PREDICTOR

Although time-delay compensators appeared in the mid 1950s, their implementation with analogue technique was very difficult and these were not used in industry. Since 1980s digital time-delay compensators can be implemented. The digital time-delay compensators are presented e.g. in (Palmor and Halevi 1990, Normey-Rico and Camacho 1998). The discrete versions of the SP and its modifications are suitable for time-delay compensation in industrial practice.

Structure of Digital Smith Predictor

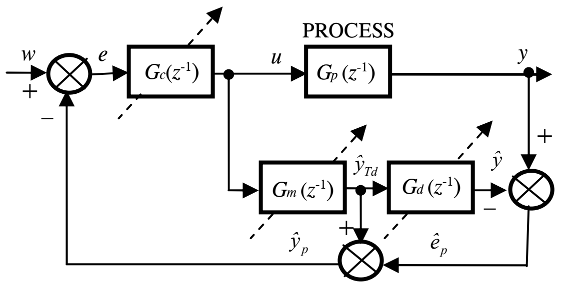

Figure 6: Block Diagram of a Digital Smith Predictor

The block diagram of a digital SP (see Hang et al. 1989, Hang et al. 1993) is shown in Fig. 6. The function of the digital version is similar to the classical analogue version. The block $G_m\left(z^{-1}\right)$ represents process dynamics without the time-delay and is used to compute an open-loop prediction. The difference between the output of the process y and the model including time-delay $\hat{y}$ is the predicted error $\hat{e}_p$ as shown is in Fig. 1 where u, w and e are the control signal, the reference signal and the error. If there are no modelling errors or disturbances, the error between the current process output and the model output will be null and the predictor output signal $\hat{y}_p$ will be the time-delay-free output of the process. Under these conditions, the controller $G_c(s)$ can be tuned, at least in the nominal case, as if the process had no time-delay. The primary (main) controller $G_c\left(z^{-1}\right)$ can be designed by the different approaches (for example digital PID control or methods based on algebraic approach). The outward feedback-loop through the block $G_d\left(z^{-1}\right)$ in Fig. 1 is used to compensate for load disturbances and modelling errors. The dash arrows indicate the tuned parts of the Smith Predictor.

Digital PID Smith Predictor

Hang *et al.* (1989, 1993) used the Dahlin PID algorithm (Dahlin 1968) for the design of the main controller $G_c\left(z^{-1}\right)$. This algorithm is based on the desired close-loop transfer function in the form

$$G_e\left(z^{-1}\right) = \frac{1 - e^{-\alpha}}{1 - z^{-1}}; \quad \alpha = \frac{T_0}{T_m} \tag{19}$$

where T_m is a desired time constant of the first order closed-loop response. It is not practical to set T_m to be small since it will demand a large control signal $u(k)$ which may easily exceed the saturation limit of the actuator. Then the individual parts of the controller are described by the transfer functions

$$G_c\left(z^{-1}\right) = \frac{\left(1 - e^{-\alpha}\right)}{\left(1 - z^{-1}\right)} \frac{\hat{A}\left(z^{-1}\right)}{\hat{B}(1)}; \quad G_m\left(z^{-1}\right) = \frac{z^{-1}\hat{B}(1)}{\hat{A}\left(z^{-1}\right)}$$

$$G_d\left(z^{-1}\right) = \frac{z^{-d}\hat{B}\left(z^{-1}\right)}{z^{-1}\hat{B}(1)} \tag{20}$$

where $B(1) = \hat{B}\left(z^{-1}\right)\Big|_{z=1} = \hat{b}_1 + \hat{b}_2$.

Since $G_m\left(z^{-1}\right)$ is the second order transfer function, the main controller $G_c\left(z^{-1}\right)$ becomes a digital PID controller having the following form:

$$G_c\left(z^{-1}\right) = \frac{U(z)}{E(z)} = \frac{q_0 + q_1 z^{-1} + q_2 z^{-2}}{1 - z^{-1}} \tag{21}$$

where $q_0 = \gamma, q_1 = \hat{a}_1\gamma, q_2 = \hat{a}_2\gamma$ using by the substitution $\gamma = \left(1 - e^{-\alpha}\right) / \hat{B}(1)$. The PID controller output is given by

$$u(k) = q_0 e(k) + q_1 e(k-1) + q_2 e(k-2) + u(k-1) \tag{22}$$

Some simulation experiments using this digital SP are introduced in (Bobál et al. 2011).

Digital Pole Assignment Smith Predictor

The digital pole assignment SP was designed using a polynomial approach in (Bobál et al. 2011). Polynomial control theory is based on the apparatus and methods of linear algebra (see e.g. Kučera 1991, Kučera 1993). The design of the controller algorithm is based on the general block scheme of a closed-loop with two degrees of freedom (2DOF) according to Fig. 7.

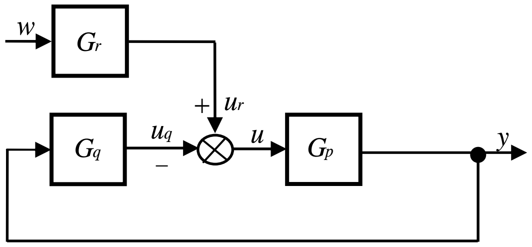

Figure 7: Block Diagram of a Closed Loop 2DOF Control System

The controlled process is given by the transfer function in the form

$$G_p(z^{-1}) = \frac{Y(z)}{U(z)} = \frac{B(z^{-1})}{A(z^{-1})} \tag{23}$$

where A and B are the second order polynomials. The controller contains the feedback part G_q and the feedforward part G_r. Then the digital controllers can be expressed in the form of discrete transfer functions

$$G_r\left(z^{-1}\right) = \frac{R\left(z^{-1}\right)}{P\left(z^{-1}\right)} = \frac{r_0}{1+p_1 z^{-1}} \tag{24}$$

$$G_q\left(z^{-1}\right) = \frac{Q\left(z^{-1}\right)}{P\left(z^{-1}\right)} = \frac{q_0 + q_1 z^{-1} + q_2 z^{-2}}{\left(1+p_1 z^{-1}\right)\left(1-z^{-1}\right)} \tag{25}$$

According to the scheme presented in Fig. 7 and Equations (21) – (23) it is possible to derive the characteristic polynomial

$$A(z^{-1})P(z^{-1}) + B(z^{-1})Q(z^{-1}) = D(z^{-1}) \tag{26}$$

where

$$D\left(z^{-1}\right) = 1 + d_1 z^{-1} + d_2 z^{-2} + d_3 z^{-3} + d_4 z^{-4} \tag{27}$$

The feedback part of the controller is given by solution of the polynomial Diophantine equation (26). The procedure leading to determination of controller parameters in polynomials Q, R and P (24) and (25) is in (Bobál et al. 2005). The asymptotic tracking is provided by the feedforward part of the controller given by solution of the polynomial Diophantine equation

$$S(z^{-1})D_w\left(z^{-1}\right) + B(z^{-1})R(z^{-1}) = D(z^{-1}) \tag{28}$$

For a step-changing reference signal value $D_w\left(z^{-1}\right) = 1 - z^{-1}$ holds and S is an auxiliary polynomial which does not enter into controller design. For a step-changing reference signal value it is possible to solve Equation (27) by substituting z = 1

$$R\left(z^{-1}\right) = r_0 = \frac{D(1)}{B(1)} = \frac{1 + d_1 + d_2 + d_3 + d_4}{b_1 + b_2} \tag{29}$$

The 2DOF controller output is given by

$$\begin{aligned} u(k) &= r_0 w(k) - q_0 y(k) - q_1 y(k-1) - q_2 y(k-2) + \\ &+ (1+p_1) u(k-1) + p_1 u(k-2) \end{aligned} \tag{30}$$

The control quality is very dependent on the pole assignment of the characteristic polynomial

$$D(z) = z^4 + d_1 z^3 + d_2 z^2 + d_3 z + d_4 \tag{31}$$

inside the unit circle. The simple method for choice of individual poles is based on the following approach. Consider 1DOF control loop where controlled process (23) with second-order polynomials A and B is controlled using PID controller which is given by transfer function

$$G_q\left(z^{-1}\right) = \frac{Q\left(z^{-1}\right)}{P\left(z^{-1}\right)} = \frac{q_0\left(1 + a_1 z^{-1} + a_2 z^{-2}\right)}{\left(1 - z^{-1}\right)} \tag{32}$$

Substitution of polynomials A, B, Q, P into Equation (26) yields the following relation

$$\begin{aligned} &\hat{A}(z^{-1})\left(1-z^{-1}\right) + \hat{B}(z^{-1}) q_0 \hat{A}(z^{-1}) = \\ &= \hat{A}(z^{-1})\left[\left(1-z^{-1}\right) + \hat{B}(z^{-1}) q_0\right] = D(z^{-1}) \end{aligned} \tag{33}$$

where

$$\hat{A}\left(z^{-1}\right) = 1 + \hat{a}_1 z^{-1} + \hat{a}_2 z^{-2}; \quad \hat{B}\left(z^{-1}\right) = \hat{b}_1 z^{-1} + \hat{b}_2 z^{-2} \tag{34}$$

are polynomials with model parameter estimates.
From Equation (33) it is obvious that polynomial $A(z) = z^2 + a_1 z + a_2$ is included in polynomial $D(z)$ (31). Its parameter estimates are known from process identification. The second two poles are dependent on the parameter (see expressions (19, 20))

$$q_0 = \frac{\left(1 - e^{-\alpha}\right)}{\hat{b}_1 + \hat{b}_2}; \quad \alpha = \frac{T_0}{T_m} \tag{35}$$

which is function of time constant T_m (free setting parameter of the controller). By increasing T_m, the control response is slower (respective without overshoot).

SIMULATION VERIFICATION DIGITAL SP CONTROLLER ALGORITHM

As simulation examples of digital SP controller algorithm, the processes (13) and (17) were chosen. By the identification procedure, the discrete models (16) and (18) for sampling period $T_0 = 0.5$ s were obtained.

A simulation verification of the designed controller was performed in MATLAB/SIMULINK environment. A typical used control scheme is depicted in Fig. 8.

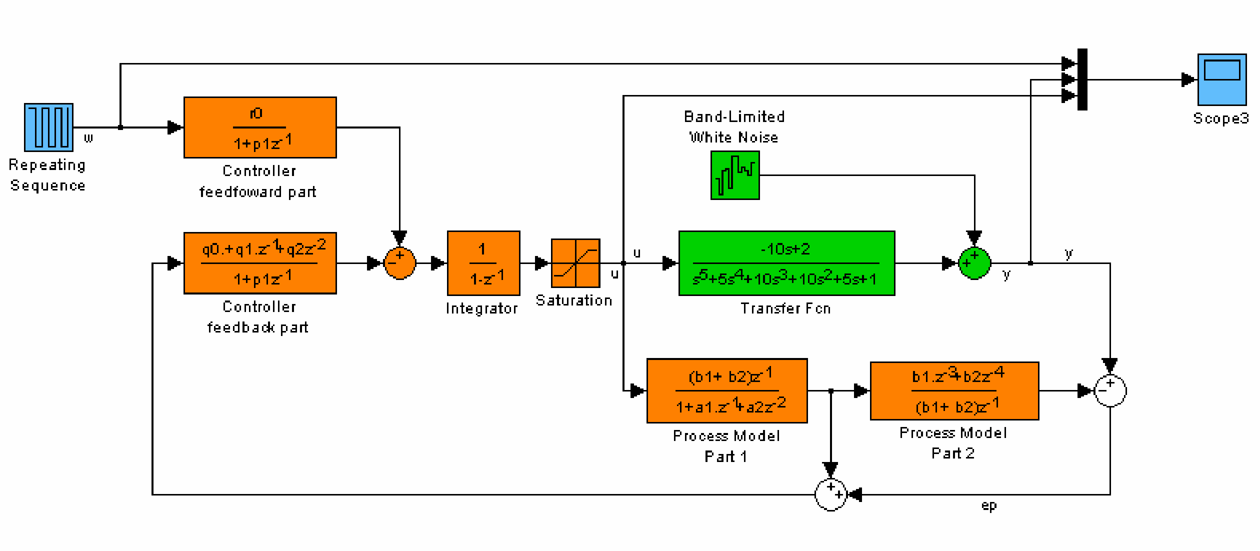

Figure 8: SIMULINK Control Scheme

Control of stable process (13)

1. Simulation conditions:

Time constant $T_m = 1.5$, characteristic polynomial

$D_{A1}(z) = z^4 - 2.5167z^3 + 2.2390z^2 - 0.7959z + 0.0839$

Simulation control results of the model $G_A(s)$ with $T_m = 1.5$ are shown in Figs. 9 and 10.

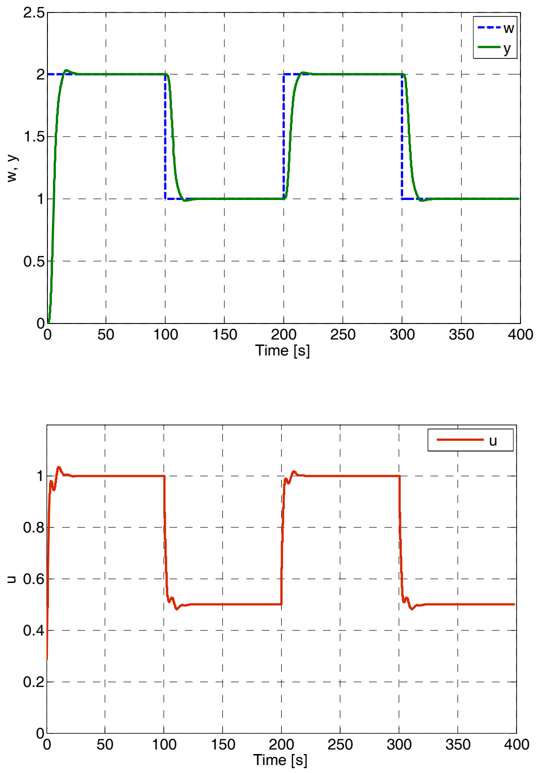

Figure: 9 Control of the Model $G_A(s)$; $T_m = 1.5$

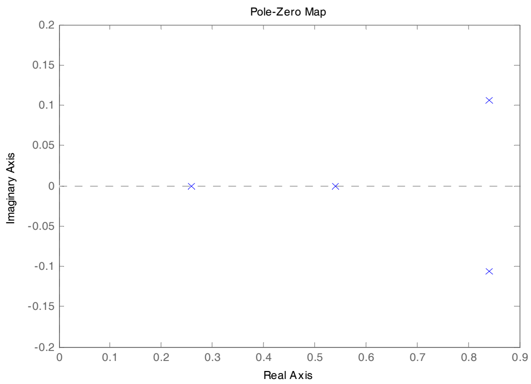

Figure: 10 Pole Map of the Polynomial D_{A1}

2. Simulation conditions:

Time constant $T_m = 3$, characteristic polynomial

$D_{A2}(z) = z^4 - 2.5932z^3 + 2.3144z^2 - 0.7611z + 0.0454$

Simulation control results of the model $G_A(s)$ with $T_m = 3$ are shown in Figs. 11 and 12.

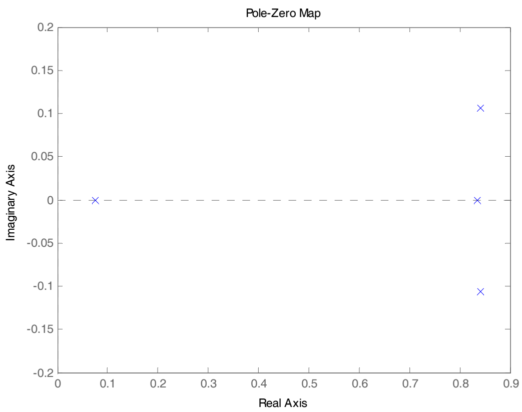

Figure: 11 Pole Map of the Polynomial D_{A2}

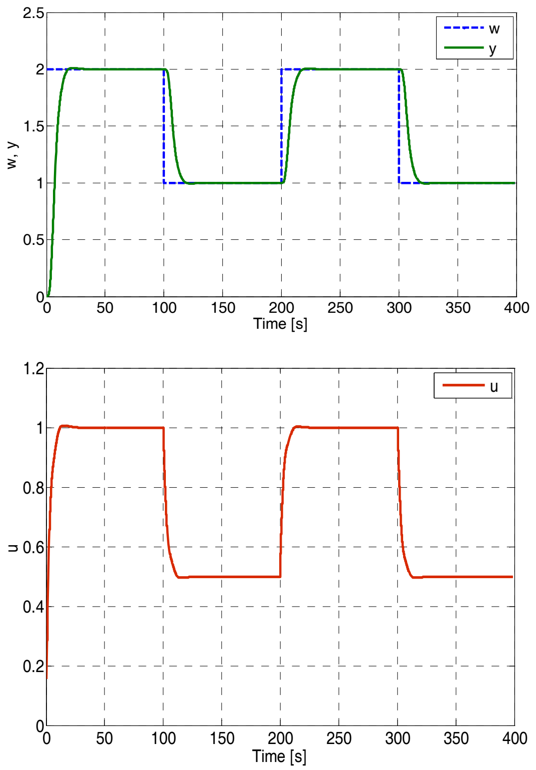

Figure: 12 Control of the Model $G_A(s); T_m = 3$

Control of non-minimum phase process (17)

Simulation conditions:
Time constant $T_m = 7$, characteristic polynomial

$$D_B(z) = z^4 - 3.3252z^3 + 4.1981z^2 - 2.3836z + 0.5135$$

Simulation control results of the model $G_B(s)$ with $T_m = 7$ are shown in Figs. 13 and 14.

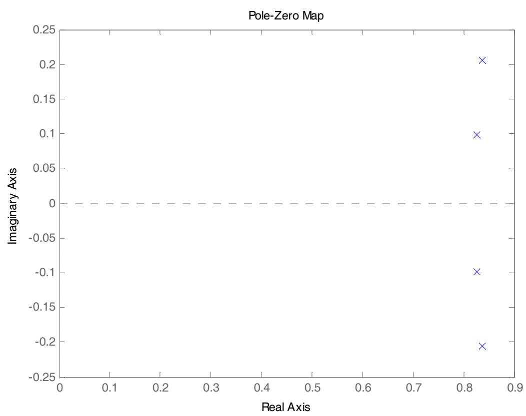

Figure: 13 Pole Map of the Polynomial D_B

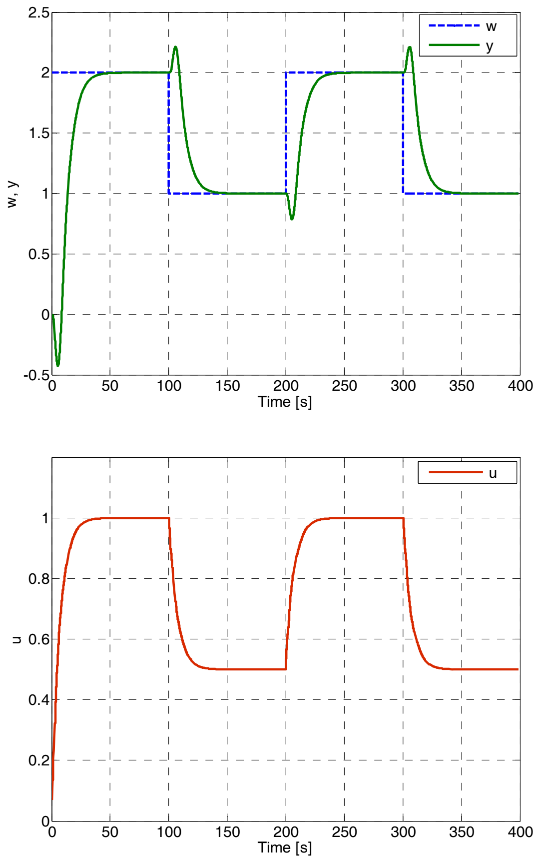

Figure: 14 Control of the Model $G_B(s); T_m = 3$

For small the time constant T_m, the control process could be unstable.

CONCLUSION

Digital Smith Predictor algorithm for control of the higher-order processes was designed. The higher-order process was identified by the second-order model with time delay. For the process identification of the time-delay was used the off-line least squares method, as excitation signal was generated Random Gaussian Signal using MATLAB code idinput( ). The controller algorithm is based on polynomial design (pole assignment approach). The method for a choice of suitable poles of the characteristic polynomial was designed. The polynomial controller was derived purposely by analytical way (without utilization of numerical methods) to obtain algorithm with easy implementability in industrial practice. The control of two modifications of the firth-order processes (stable and non-minimum phase) were verified by simulation. Results of simulation verification in both cases demonstrated very good of control quality. Unfortunately, digital Smith Predictor is not suitable for the control of unstable processes. The proposed digital Smith Predictor will be verified in real-time laboratory conditions for the control of the heat exchanger. Adaptive versions of digital Smith Predictors are designed in (Bobál et al. 2011b).

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AUTHOR BIOGRAPHIES

VLADIMÍR BOBÁL graduated in 1966 from the Brno University of Technology, Czech Republic. He received his Ph.D. degree in Technical Cybernetics at Institute of Technical Cybernetics, Slovak Academy of Sciences, Bratislava, Slovak Republic. He is now Professor at the Department of Process Control, Faculty of Applied Informatics of the Tomas Bata University in Zlín. His research interests are adaptive control and predictive control, system identification and CAD for automatic control systems. You can contact him on email address bobal@fai.utb.cz.

MAREK KUBALČÍK graduated in 1993 from the Brno University of Technology in Automation and Process Control. He received his Ph.D. degree in Technical Cybernetics at Brno University of Technology in 2000. From 1993 to 2007 he worked as senior lecturer at the Faculty of Technology, Brno University of Technology. From 2007 he has been working as an associate professor at the Department of Process Control, Faculty of Applied Infromatics of the Tomas Bata University in Zlín, Czech Republic. Current work cover following areas: control of multivariable systems, self-tuning controllers, predictive control. His e-mail address is: kubalcik@fai.utb.cz.

PETR CHALUPA was born in Zlin, Czech Republic in 1976. He graduated from Brno University of Technology in 1999. He obtained his Ph.D. in Technical Cybernetics at Tomas Bata University in Zlin in 2003. He works as a researcher at Faculty of Applied Informatics at Tomas Bata University in Zlin. His research interests are modeling, adaptive control and predictive control of real-time systems. You can contact him on email address chalupa@fai.utb.cz.

PETR DOSTÁL studied at the Technical University of Pardubice, Czech Republic, where he obtained his master degree in 1968 and PhD. degree in Technical Cybernetics in 1979. In the year 2000 he became professor in Process Control. He is now head of the Department of Process Control, Faculty of Applied Informatics of the Tomas Bata University in Zlín. His research interests are modelling and simulation of continuous-time chemical processes, polynomial methods, optimal and adaptive control. You can contact him on email address dostalp@fai.utb.cz.