

# VERIFICATION OF ROBUST PROPERTIES OF DIGITAL CONTROL CLOSED-LOOP SYSTEMS

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## KEYWORDS

Digital Control, Polynomial Methods, Robustness, Robustness Margins, LQ Method, Simulation of Control Loop Systems.

## ABSTRACT

Robustness is specific property of closed-loop systems when the designed controller guarantees control not only for one nominal controlled system but also for all predefined class of systems (perturbed models). The robust theory is mainly exploited for design of the continuous-time systems. This paper deals with an experimental simulation investigation of robust properties of digital control closed-loop systems. Minimization of the Linear Quadratic (LQ) criterion was used for the design of control algorithm. Polynomial approach is based on the structure of the controller with two degrees of freedom (2DOF). Four types of process models (stable, non-minimum phase, unstable and integrating) were used for controller design. The Nyquist plot based characteristics of the open-loop transfer function (gain margin, phase margin and modulus margin) served as robustness indicators. The influence of change of process gain was chosen as a parametric uncertainty. The experimental results demonstrated that a robustness of examined digital control closed-loop systems could be improved by addition of user-defined poles (UDP).

## INTRODUCTION

One of possible approaches to digital control systems is the polynomial theory. Polynomial methods are design techniques for complex systems (including multivariable), signals and processes encountered in control, communications and computing that are based on manipulations and equations with polynomials, polynomial matrices and similar objects. Systems are described by input-output relations in fractional form and processed using algebraic methodology and tools (Šebek and Hromčík 2007). The design procedure is thus reduced to algebraic polynomial equations. Controller design consists of solving polynomial (Diophantine) equations. The Diophantine equations

can be solved using the uncertain coefficient method – which is based on comparing coefficients of the same power. This is transformed into a system of linear algebraic equations (Kučera 1997).

It is obvious that the majority processes met in industrial practice are influenced by uncertainties. The uncertainties suppression can be solved by implementation of either adaptive control or robust control. The robust control and the adaptive control are viewed as two control techniques, which are used for controller design in the presence of process model uncertainty - process model variations (Landau 1999; Landau et al. 2011).

The design of a robust controller deals in general with designing the controller in the presence of process uncertainties. This can be simultaneously: parameter variations (affecting low- and medium-frequency ranges) and unstructured model uncertainties (often located in high-frequency range).

The aim of this paper is the experimental examination of the robustness of digital controllers based on LQ method. Robustness is the property when the dynamic response of control closed loop (including stability of course) is satisfactory not only for the nominal process transfer function used for the design but also for the entire (perturbed) class of transfer functions that expresses uncertainty of the designer in dynamic environment in which a real controller is expected to operate. The design of a robust digital pole assignment controller is investigated in (Landau and Zito 2006), the robust stability of discrete-time systems with parametric uncertainty is analysed in (Matusš 2014). A more comprehensive discussion of robustness is taken when the design based on frequency methods is considered. One can readily compare the system gain at the desired operating point and the point(s) of onset of instability to determine how much gain change is acceptable. Only this method will be used for investigation of the robustness of digital control stable, unstable, non-minimum phase and integrating processes.

The paper is organized in the following way. The fundamental principle of the robustness of digital

control-loop systems with basic concepts are illustrated in Section 2. Design of 2DOF digital LQ controller is presented in Section 3. The simulation verification of robust properties of digital closed-loop control with their results are presented in Section 4. Section 5 concludes this paper.

## ROBUSTNESS PROBLEM AND BASIC CONCEPTS

### Digital Control Loops

The “degrees of freedom” concept is frequently used in the development of control strategies and control system design. The degree of freedom of a control system is defined as the number of closed-loop transfer functions that can be adjusted independently. The design of control systems is a multi-objective problem, so a **Two-Degree-of-Freedom (2DOF)** control system naturally has advantages over a **One-Degree-of-Freedom** (abbreviated as 1DOF) control system. This fact was already stated by (Horowitz 1963). Typical structure of the digital control-loop of 2DOF modification is depicted schematically in Fig. 1.

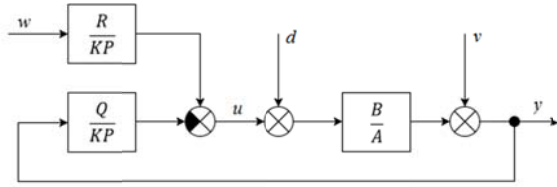


Figure 1: Block diagram of a closed loop 2DOF control system

The specification of individual discrete transfer functions and signals in Fig. 1:

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

is the process model and

$$G_r(z^{-1}) = \frac{R(z^{-1})}{P(z^{-1})K(z^{-1})} \quad (2)$$

$$G_q(z^{-1}) = \frac{Q(z^{-1})}{P(z^{-1})K(z^{-1})} \quad (3)$$

are the feedback part  $G_q$  and the feedforward part  $G_r$  of individual controllers;  $y$ ,  $u$ ,  $w$  are the process output, the controller output and the reference signal;  $d$  and  $v$  are disturbances and

$$K(z^{-1}) = 1 - z^{-1} \quad (4)$$

### Robustness

The mathematical model of the process is just inaccurate interpretation of the real system, therefore

the discrepancies between the model and the real system do not enable the optimal controller design. A number of factors may be responsible for modelling errors. A generic term for this modelling error is *model uncertainty* which can itself be represented mathematically in different ways. Two major classes of uncertainty are: structured uncertainty (parametric uncertainty) and unstructured uncertainty (specified in the frequency domain). The model used for controller design will be termed the *nominal model*. Robustness is a property of the controller that was designed for control of the nominal model and is suitable for control of the family (similar class) models.

Robust control methods are analyzed e.g. in (Morari and Zafirov 1989; Doyle et al. 1990; Sánchez-Peña and Sznajner 1998; Skogestad and Postlethwaite 2005; Matušů and Prokop 2011).

### Robustness Margins

The open-loop transfer function of 2DOF is

$$G_{OL}(z^{-1}) = \frac{B(z^{-1})Q(z^{-1})}{A(z^{-1})K(z^{-1})P(z^{-1})} \quad (5)$$

The frequency characteristic can be obtained by substitution  $z = e^{j\omega}$  where  $\omega$  is the normalized frequency. The Nyquist plot is a parametric plot of a frequency response of the open-loop transfer function  $G_{OL}(e^{-j\omega})$ .

Robustness of closed loops is closely related to the distance between Nyquist plot of open loop transfer function and critical point  $(-1, j0)$ . There are some elements, which help to evaluate this distance: gain margin, phase margin, delay margin and modulus margin.

The *gain margin*  $G_m$  is defined as the change in open-loop gain required to make the system unstable. Systems with greater gain margins can withstand greater changes in system parameters before becoming unstable in closed loop. It can be considered as an inverse value of the gain corresponding to a phase shift that occurs at the critical frequency  $\omega_{pc}$ .

$$G_m = \frac{1}{|G_{OL}(e^{-j\omega_{pc}})|} \quad (6)$$

Minimal recommended value of the gain margin is 0.5 (Landau and Zito 2006). The *phase margin*  $P_m$  is defined as the change in open-loop phase shift required to make a closed-loop system unstable. The phase margin also measures the system's tolerance to time-delay. If there is a time-delay greater than  $180/\omega_{pc}$  in the open-loop, the system will become unstable in closed-loop (Messner and Tilbury 2011). The phase margin is the additional phase that it must be add at the

crossover frequency  $\omega_{cr}$ , for which the gain of the open-loop system equals 1. It is defined (in degrees) as

$$P_m = 180^\circ - \varphi(\omega_{cr}) \quad (7)$$

where  $\varphi(\omega_{cr})$  is phase shift at crossover frequency. Typical values for a good phase margin are  $30^\circ \leq P_m \leq 60^\circ$  (Landau and Zito 2006).

The *delay margin*  $\tau_m$  is amount of added time delay that the system can tolerate before it becomes unstable.

$$\tau_m = \frac{P_m}{\omega_{cr}} \quad (8)$$

The *modulus margin*  $M_m$  is minimum distance in the Gauss plane between Nyquist plot of open-loop transfer function and critical point  $(-1, j0)$

$$M_m = \left| 1 + G_{OL}(z^{-1}) \right|_{\min} \quad (9)$$

Notice that the modulus margin is a much more reliable robustness indicator than gain and phase margin. It is possible to consider this indicator as modern approach so-called infinity norm  $H_\infty$  (Kwakernaak 1993). It can be shown that good gain margin does not guarantee good phase margin, and vice versa. On the other side, values of modulus margin bigger than 0.5 guarantee  $G_m \geq 2$  and  $P_m > 29^\circ$  (Landau 1998).

The gain, phase and modulus margins in the Nyquist plot are depicted in Fig. 2.

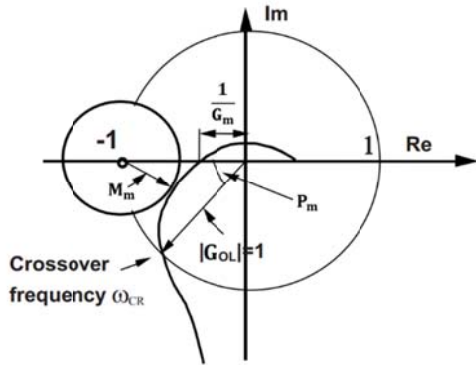


Figure 2: Gain, phase and modulus margins

## DESIGN OF POLYNOMIAL 2DOF LQ CONTROLLER

The design of the control algorithm is based on a general block scheme of closed-loop with two degrees of freedom (2DOF) as seen in Fig. 1.

The second-order discrete process model is considered

$$G_p(z^{-1}) = \frac{B(z^{-1})}{A(z^{-1})} = \frac{b_1 z^{-1} + b_2 z^{-2}}{1 + a_1 z^{-1} + a_2 z^{-2}} \quad (10)$$

Then the individual parts of digital controller can be expressed in the form of discrete transfer functions

$$G_r(z^{-1}) = \frac{R(z^{-1})}{P(z^{-1})K(z^{-1})} = \frac{r_0}{(1 + p_1 z^{-1})(1 - z^{-1})} \quad (11)$$

$$G_q(z^{-1}) = \frac{Q(z^{-1})}{P(z^{-1})K(z^{-1})} = \frac{q_0 + q_1 z^{-1} + q_2 z^{-2}}{(1 + p_1 z^{-1})(1 - z^{-1})} \quad (12)$$

The controller synthesis consists of solving linear polynomial (Diophantine) equations. From first polynomial equation

$$A(z^{-1})K(z^{-1})P(z^{-1}) + B(z^{-1})Q(z^{-1}) = D(z^{-1}) \quad (13)$$

it is possible to calculate 4 feedback controller parameters – coefficients of the polynomials  $Q$ ,  $P$ . Polynomial  $D(z^{-1})$  is the characteristic polynomial.

Asymptotic tracking of the reference signal  $w$  is provided by the feedforward part of the controller which is given by solution of the following polynomial Diophantine equation

$$S(z^{-1})D_w(z^{-1}) + B(z^{-1})R(z^{-1}) = D(z^{-1}) \quad (14)$$

For a step-changing reference signal value, polynomial  $D_w(z^{-1}) = 1 - z^{-1}$  and  $S$  is an auxiliary polynomial which does not enter into the controller design. Then it is possible to derive the polynomial  $R$  from equation (14) by substituting  $z = 1$

$$R = r_0 = \frac{D(1)}{B(1)} \quad (15)$$

The 2DOF controller output is given by

$$u(k) = \frac{r_0}{K(z^{-1})P(z^{-1})} w(k) - \frac{Q(z^{-1})}{K(z^{-1})P(z^{-1})} y(k) \quad (16)$$

Polynomial LQ controllers are derived in this paper using minimization of LQ criterion (Kučera 1991).

### Minimization of LQ Criterion

The linear quadratic control methods try to minimize the quadratic criterion which uses penalization of the value of the controller output

$$J = \sum_{k=0}^{\infty} \left\{ [w(k) - y(k)]^2 + q_u [u(k)]^2 \right\} \quad (17)$$

where  $q_u$  is so-called penalization constant, which relates the controller output to the value of the criterion. In this paper, criterion minimization (17) will be realized through the spectral factorization for an input-output description of the system

$$A(z)q_u A(z^{-1}) + B(z)B(z^{-1}) = D(z)\delta D(z^{-1}) \quad (18)$$

where  $\delta$  is a constant chosen so that  $d_0 = 1$ .

Because the experimental process model (10) is second-order with second-degree polynomials  $A(z^{-1}), B(z^{-1})$ , the second-degree polynomial is also obtained from the spectral factorization:

$$D(z^{-1}) = 1 + d_1 z^{-1} + d_2 z^{-2} \quad (19)$$

Spectral factorization of polynomials of the first- and the second- degree can be computed by analytical way; the procedure for higher degrees must be performed iteratively (Bobál et al. 2005). The MATLAB Polynomial Toolbox (Šebek 2014) can be used for a computation of spectral factorization of the higher degree polynomials using file *spf.m* by command

$$d = \text{spf}(a^* \text{qu}^* a' + b^* b') \quad (20)$$

It is known that by using the spectral factorization (18), it is possible to compute only two suitable poles ( $\alpha, \beta$ ). It is obvious from equation (13) that in this case a choice of the fourth-degree polynomial  $D(z)$  is suitable

$$D_4(z^{-1}) = 1 + d_1 z^{-1} + d_2 z^{-2} + d_3 z^{-3} + d_4 z^{-4} \quad (21)$$

Therefore, the other poles ( $\gamma_1, \gamma_2$ ) are needed to be user-defined. The application of user-defined poles is significant mainly in nonstandard process models (e.g. non-minimum phase, unstable, integrating or with time-delay). The usage of user-defined poles makes it possible to improve the robustness of these processes. Then the digital 2DOF controller (16) can be expressed in the form

$$u(k) = r_0 w(k) - q_0 y(k) - q_1 y(k-1) - q_2 y(k-2) + (1-p_1)u(k-1) + p_1 u(k-2) \quad (22)$$

where

$$r_0 = \frac{1 + d_1 + d_2 + d_3 + d_4}{b_1 + b_2} \quad (23)$$

and parameters  $q_0, q_1, q_2, p_1$  are computed from (13).

## SIMULATION VERIFICATION AND RESULTS

A simulation verification of designed algorithms was performed in MATLAB/SIMULINK environment. The robustness of individual control loops was experimentally investigated by changing characteristic polynomial degree by adding user-defined poles. In addition, the influence of a change of the gain of the nominal process model was examined. From the point of view of robust theory, it is possible to consider these experiments as determination of individual robust stability margins that are caused by parametric uncertainty influence.

This paper presents robustness properties of the digital second-order control systems that can be described by the following continuous-time transfer function:

1) Stable system:

$$G_1(s) = \frac{1}{10s^2 + 7s + 1} \quad (24)$$

2) Non-minimum phase stable system:

$$G_2(s) = \frac{1-2s}{10s^2 + 7s + 1} \quad (25)$$

3) Unstable system:

$$G_3(s) = \frac{1}{8s^2 - 2s - 1} \quad (26)$$

4) Integrating system:

$$G_4(s) = \frac{1}{s(2s+1)} \quad (27)$$

Let us now discretize (24) - (27) using a sampling period  $T_0 = 2$  s. Discrete forms of these transfer functions are (see (10))

$$G_{N1}(z^{-1}) = \frac{0.1281z^{-1} + 0.0803z^{-2}}{1 - 1.0382z^{-1} + 0.2466z^{-2}} \quad (28)$$

$$G_{N2}(z^{-1}) = \frac{-0.0736z^{-1} + 0.2820z^{-2}}{1 - 1.0382z^{-1} + 0.2466z^{-2}} \quad (29)$$

$$G_{N3}(z^{-1}) = \frac{0.3104z^{-1} + 0.3656z^{-2}}{1 - 3.3248z^{-1} + 1.6487z^{-2}} \quad (30)$$

$$G_{N4}(z^{-1}) = \frac{0.7385z^{-1} + 0.5285z^{-2}}{1 - 1.3679z^{-1} + 0.3679z^{-2}} \quad (31)$$

Transfer functions (28) - (31) represent nominal models. From the second-order model (10) and two parts (11), (12) of digital 2DOF controller, the open-loop transfer function can be expressed using individual process and controller parameters

$$G_{OL}(z) = \frac{e_1 z^3 + e_2 z^2 + e_3 z + e_4}{z^4 + f_1 z^3 + f_2 z^2 + f_3 z + f_4} \quad (32)$$

where

$$\begin{aligned} e_1 &= b_1 q_0; & e_2 &= b_1 q_1 + b_2 q_2; & e_3 &= b_1 q_2 + b_2 q_1; \\ e_4 &= b_1 q_0; & f_1 &= p_1 + a_1 - 1; & f_2 &= p_1(a_2 - a_1) - a_2; \\ f_3 &= p_1(a_1 - 1) - a_1 + a_2; & f_4 &= -a_2 p_1 \end{aligned} \quad (33)$$

Experimental process models (28) - (31) were used for simulation experiments. Individual simulation experiments are realized subsequently:

- The individual nominal models (28) - (31)  $G_{Ni}$  for ( $i = 1, 2, 3, 4$ ) were multiplied by the parameter  $K_{Pi}$ , then the perturbed models are given as

$$G_{Pi} = K_{Pi} G_{Ni} \quad (34)$$

The parameter  $K_{Pi}$  was increased (decreased) as far as control closed-loops were in the stability

boundary - the critical gain  $K_{ci}$  was determined.

- These experiments were realized for the case when the poles  $\alpha, \beta$  were computed by spectral factorization and user-defined poles (UDF) were  $\gamma_1 = \gamma_2 = 0$ . The penalization factor  $q_u = 10$  was used for all experiments.
- Obtained critical gains were used for simulation in control-loops when the poles  $\alpha, \beta$  were computed by spectral factorization and  $\gamma_1 = 0.2, \gamma_2 = 0$ .
- The same experimental conditions were used as in the second case, only  $\gamma_2 = 0.4$ .
- Individual control behaviours of models (28) – (31) are shown in Figs. 3, 5, 7a, 7b and 11.
- Frequency plots of open-loops with the nominal models  $G_{Ni}$  were depicted, see (Figs. 4, 6, 8 and 12).
- Values of individual robustness margins -  $G_m, P_m, M_m$  were computed (see Tabs. 1 – 4).

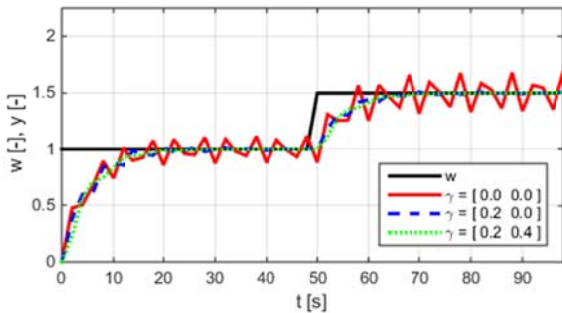


Figure 3: Control of stable model (28) and using UDP  $K_{c1} = 3.55$

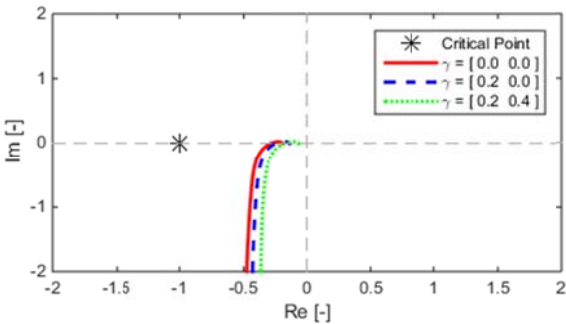


Figure 4: Nyquist plots of open loop with model (28)

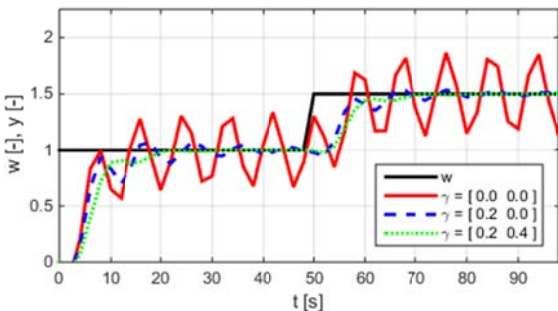


Figure 5: Control of non-minimum phase model (29) and using UDP,  $K_{c2} = 1.72$

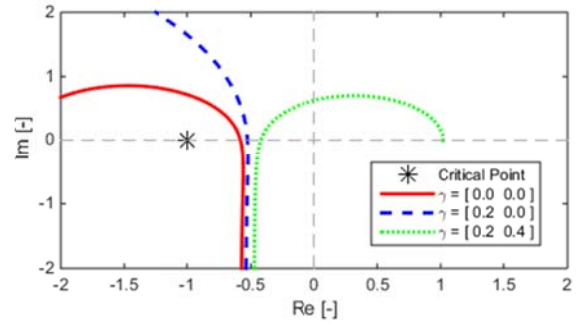


Figure 6: Nyquist plots of open loop with model (29)

In the case of the unstable model (30), the Nyquist plot crosses the real axis in two points (see Fig. 8). Therefore, two critical gains exist and two cases of control closed-loops are in the stability boundary (see Figs. 7a, 7b). It is obvious from Fig. 7a that the addition of UDP does not stabilize the control process, on the contrary the control process is unstable by addition  $\gamma_1, \gamma_2$ .

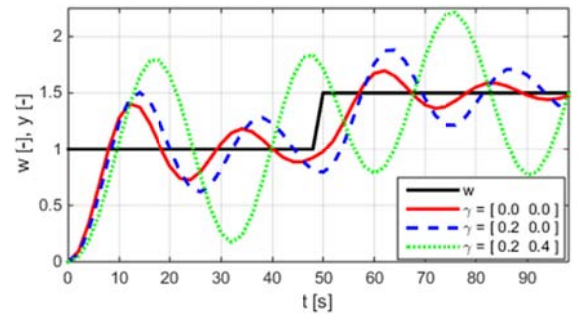


Figure 7a: Control of unstable model (30) and using UDP,  $K_{c3a} = 0.78$

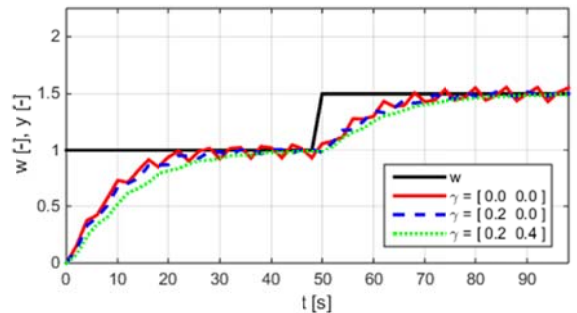


Figure 7b: Control of unstable model (30) and using UDP,  $K_{c3b} = 1.29$

Figs. 9 ( $K_{c3a} = 0.78$ ) and 10 ( $K_{c3b} = 1.29$ ) show where Nyquist plots cross the real axis relatively to critical points.

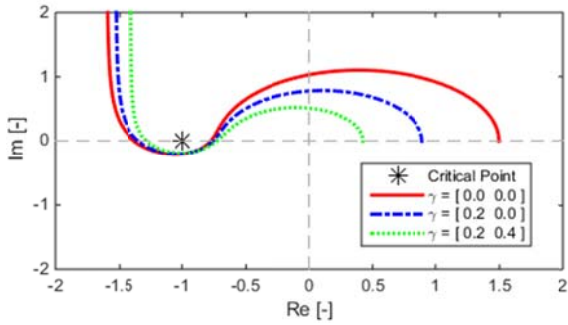


Figure 8: Nyquist plots of open loop with model (30)

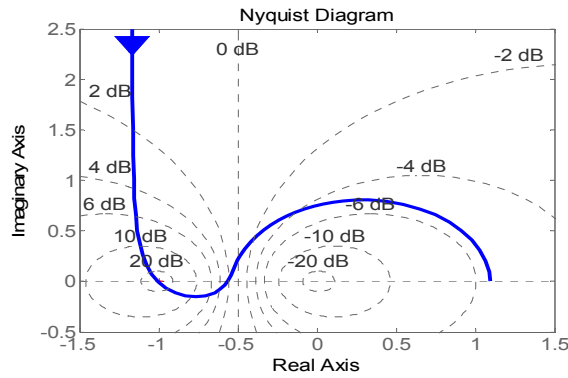


Figure 9: Cross point of Nyquist plot with real axis is located on the right side from critical point  $(-1, j0)$

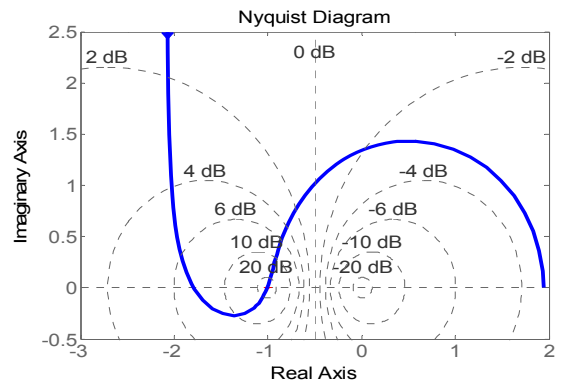


Figure 10: Cross point of Nyquist plot with real axis is located on the left side from critical point  $(-1, j0)$

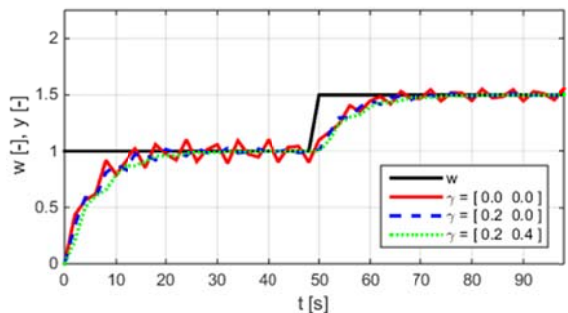


Figure 11: Control of integrating model (31) and using UDP,  $K_{c2} = 2.37$

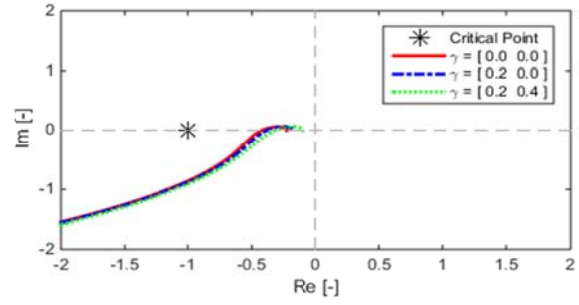


Figure 12: Nyquist plots of open loop with model (31)

Table 1: Robustness Margins, Model (28)

UDP	$G_m$	$P_m$	$M_m$
$\gamma_1 = \gamma_2 = 0$	3.55	63.40	0.64
$\gamma_1 = 0.2; \gamma_2 = 0$	4.18	65.96	0.68
$\gamma_1 = 0.2; \gamma_2 = 0.4$	5.84	69.84	0.74

Table 2: Robustness Margins, Model (29)

UDP	$G_m$	$P_m$	$M_m$
$\gamma_1 = \gamma_2 = 0$	1.72	-34.56	0.41
$\gamma_1 = 0.2; \gamma_2 = 0$	1.90	-50.87	0.47
$\gamma_1 = 0.2; \gamma_2 = 0.4$	2.37	62.18	0.58

Table 3: Robustness Margins, Model (30)

UDP	$G_m$	$P_m$	$M_m$
$\gamma_1 = \gamma_2 = 0$	*	11.42	0.19
$\gamma_1 = 0.2; \gamma_2 = 0$	**	11.37	0.20
$\gamma_1 = 0.2; \gamma_2 = 0.4$	***	10.82	0.19

\* Interval (0.73 – 1.29)

\*\* Interval (0.74 – 1.32)

\*\*\* Interval (0.78 – 1.37)

Table 4: Robustness Margins, Model (31)

UDP	$G_m$	$P_m$	$M_m$
$\gamma_1 = \gamma_2 = 0$	2.37	38.26	0.50
$\gamma_1 = 0.2; \gamma_2 = 0$	2.61	39.31	0.52
$\gamma_1 = 0.2; \gamma_2 = 0.4$	3.15	40.88	0.57

Typical values for stability margins in a robust design are recommended in (Landau and Zito 2006):

1. Gain margin:  $G_m \geq 2$ , [min: 1.6]
2. Phase margin:  $30^\circ \leq P_m \leq 60^\circ$
3. Modulus margin:  $M_m \geq 0.5$ , [min: 0.4]

It is obvious from Tabs. 1 – 4 that these recommendations are fulfilled subsequently:

*Recommendation 1* is fulfilled for stable system (28) and integrating system (31) – their Nyquist plots are located on the left side of the complex plane (see Figs.

4 and 12) and they cross the real axis relatively near the point (0, 0).

*Recommendation 2* is almost fulfilled for system (28) and fully for system (31).

*Recommendation 3* can be fulfilled for all systems except of unstable system (30).

*Remarks:* The negative  $-P_m$  in the system (29) is caused by an unstable zero (see the first and second Nyquist plot – Fig. 6). The unstable system (30) does not fulfil any recommended value. The open-loop of this system has two cross points of Nyquist plot with real axis  $(-0.73, j0)$  and  $(-1.29, j0)$  – see Figs. 8 -10. The perturbed system is at the stability boundary for  $K_{p3} = K_{c3a}$  and  $K_{p3} = K_{c3b}$  and is stable between these points  $(K_{c3a} < K_{p3} < K_{c3b})$  - see Tab.3.

## CONCLUSION

It is well known that the robust theory is applied mainly for design of continuous-time controllers for control of stable systems. The paper presents an experimental simulation investigation of robustness of algorithms for digital LQ control. Individual control algorithms were verified using 2DOF modification in the MATLAB/Simulink environment. Four types of process models (stable, non-minimum phase, unstable and integrating) were used for controller design. The influence of change of process gain (parametric uncertainty) was used for determination of the stability boundary. The gain, phase and modulus margins of the Nyquist plot of open-loop sampled-data system were used as robustness indicators. Values of these indicators showed that only stable and integrating processes are in recommended intervals. Simulation experiments demonstrated that robustness margins of examined digital control closed-loop systems can be improved by addition of user-defined poles.

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