# DESIGN OF IMPACT CONTROLLING STRUCTURE WITH CONVENTIONAL DIGITAL CONTROLLERS

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**Summary** – *This paper presents the design of a simplified IMPACT (Internal Model Principle and Control Together) structure with conventional digital PI control law. The design procedure is accomplished to enable the extraction of a known class of immeasurable external disturbances and easy setting of controller parameters. In the proposed controlling structure, the set point transient response and speed of disturbance rejection can be adjusted independently. The efficiency and robust properties of the proposed controlling structure are verified and tested by simulation and on an experimental setup.*

## 1. INTRODUCTION

The concept of internal models means the inclusion of the nominal plant model and/or the model of immeasurable external disturbance into the control portion of the system. In the IMC (Inernal Model Control) approach, the internal plant model is used to achieve a high sistem performance [1-2]. In the IMP (Internal Model Principle), the model of external disturbance is incorporated into the minor local loop of control system in order to eliminate or to reject the influence of disturbance on the steady state value of system conrolled variable [1-5]. Since the real plant differs from its nominal model, in both the IMC and IMP the system robustness must be provided [4]. The IMPACT conrolling structure incorporates both the internal nominal plant model explicitely and internal model of disturbance implicitely. The structure has been recently developed by Ya.Z. Tsypkin independently of the IMC and IMP. The IMPACT structure enables easy achievement of the desired set-point transient response, rejection of the known class of disturbances, and a high robustness of system with respect to interval changes of plant parameters [6]. Unlike the classical IMC, which is applicable for stable plants, the IMAPCT structure may be applied for all kind of control plants including unstable plants and plants of nonminimal phase [7].

In many control applications, particularly in controlled of slow varying industrial processes, the conventional P, PI, and PID control laws are applied [8-11]. In designing single-loop control system with the conventional control laws, the control plant is approximated by typical relatively simple nominal plant model developed in a low frequency range. This paper shows the design of IMPACT structure of the system with PI controller in the main control loop and the internal models of nominal plant and disturbance in the local minor loop. The structure enables the set point transient response and speed of disturbance rejection to be adjusted independently by setting a small number of parameters having clear physical meanings.

## 2. PRINCIPLE OF ABSORPTION

Suppose that *k*th sample of external disturbance $w(t)$ may be determined by finite number $m_0$ of previous samples. Then, the disturbance is regular and may be described by extrapolation equation [1,3,7]

$$w(kT) = D_w(z^{-1})w((k-1)T) \qquad (1)$$

where $D_w(z^{-1})$ is the prediction polynomial of order $m_o$ - 1. Relation (1) is called the equation of extrapolation or prediction [3,7] and it may be rewritten as

$$\left(1 - z^{-1}D_w(z^{-1})\right)w(z^{-1}) = 0 \qquad (2)$$

where $w(z^{-1})$ denotes the *z*-transform of disturbance. Relation (2) is called compensation equation and FIR filter having the pulse transfer function $\left(1 - z^{-1}D_w(z^{-1})\right)$ is the absorption filter or the compensation polynomial [3].

Absorption filter $\Phi_w(z^{-1}) = 1 - z^{-1}D_w(z^{-1})$ is designed for a known class of disturbances and its impulse response becomes identically equal to zero after *n* sampling instants, where $n \geq m_o$. Hence, the compensation equation (2) may be considered as the absorption condition of a given class of disturbances. The condition can be expressed as

$$\Phi_w(z^{-1})w(z^{-1}) = 0, \quad \text{za } t = kT \geq (\deg \Phi_w)T \qquad (3)$$

The extrapolation polynomial $D_w(z^{-1})$ is determined by an apriori information about disturbance $w(t)$ [1,3,7]; nevertheless, it is simply resolved from as

$$\Phi_w(z^{-1}) = w_{den}(z^{-1}), \qquad w(z^{-1}) = \frac{w_{num}(z^{-1})}{w_{den}(z^{-1})} \qquad (4)$$

In the case of a stochastic disturbance *s*(*t*), absorption filter (4) should suppress as much as possible the influence of disturbance on the system output. Thus, for a law frequency disturbance *s*(*t*), that can be generated by double integration of the white noise, an appropriate choice of absorption filter is $\Phi_s(z^{-1}) = (1 - z^{-1})^2$ or prediction polynomial $D(z^{-1}) = 2 - z^{-1}$ that corresponds to absorption of linear (ramp) disturbance [1,3,5].

## 3. IMPACT STRUCTURE

In the IMPACT structure shown in Fig.1, the controlling process is given by its pulse transfer function or by polynomials $P_u(z^{-1})$ and $Q(z^{-1})$, and process dead-time given by integer *k*. Within the control portion of the structure in Fig.1 (shaded part) two internal models are included: the two-input nominal plant model

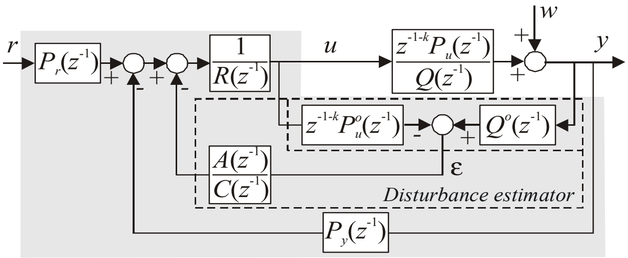

Fig.1. IMPACT controlling structure

$$W^0(z^{-1}) = \frac{z^{-1-k}P_u^0(z^{-1})}{Q^0(z^{-1})} \tag{5}$$

explicitly and the disturbance model embedded into the discrete filter $A(z^{-1})/C(z^{-1})$. Both the internal nominal plant model and disturbance model is treated as a disturbance estimator. The controlling structure of Fig.1 has two control loops that can be designed independently. The minor local control loop is designed by the proper choice of polynomials $A(z^{-1})$ and $C(z^{-1})$ while polynomials $P_r(z^{-1})$ and $P_y(z^{-1})$ in the main control loop are determined to achieve the desired system set point response. For a minimal phase plant, the proper choice of polynomial $R(z^{-1})$ is $R(z^{-1}) = P_u^o(z^{-1})$ [1,2,5].

Under the nominal case ($P_u(z^{-1}) \equiv P_u^0(z^{-1})$, $Q(z^{-1}) \equiv Q^0(z^{-1})$) and for $R(z^{-1}) = P_u^o(z^{-1})$, the closed-loop transfer functions $y(z^{-1})/r(z^{-1})$ and $y(z^{-1})/w(z^{-1})$ are easily derived from Fig.1 as

$$\frac{y(z^{-1})}{w(z^{-1})} = \frac{Q^0(z^{-1})\left[C(z^{-1}) - z^{-1-k}A(z^{-1})\right]}{C(z^{-1})\left[Q^0(z^{-1}) + z^{-1-k}P_y(z^{-1})\right]} \tag{6}$$

and

$$\frac{y(z^{-1})}{r(z^{-1})} = \frac{z^{-1-k}P_r(z^{-1})}{Q^0(z^{-1}) + z^{-1-k}P_y(z^{-1})} \tag{7}$$

In virtue of (7), the system set-point response can be adjusted by determining appropriate polynomials $P_r(z^{-1})$ and $P_y(z^{-1})$ according to the desired system closed loop transfer function $y(z^{-1})/r(z^{-1}) = G_{de}(z^{-1})$. Then, the absorption of an external disturbance, speed of disturbance transient response, and the system robustness with respect to uncertainties of plant parameters are adjusted by choosing the structure and parameters of the disturbance estimator.

From (6), the steady-state error in the presence of a known class of external disturbance $w(t)$ becomes zero if

$$\lim_{z \to 1}(1 - z^{-1})\frac{Q^0(z^{-1})\left[C(z^{-1}) - z^{-1-k}A(z^{-1})\right]}{C(z^{-1})\left[Q^0(z^{-1}) + z^{-1-k}P_y(z^{-1})\right]}w(z^{-1}) = 0 . \tag{8}$$

In the case of stable polynomial $C(z^{-1})$ and the plant of nonminimal phase

$$\lim_{z \to 1}\frac{Q^0(z^{-1})}{C(z^{-1})\left[Q^0(z^{-1}) + z^{-1-k}P_y(z^{-1})\right]} \neq 0 \tag{9}$$

the relation (8) is reduced to

$$\lim_{z \to 1}(1 - z^{-1})\left[C(z^{-1}) - z^{-1-k}A(z^{-1})\right]w(z^{-1}) = 0 . \tag{10}$$

The stable polynomial $C(z^{-1})$ is to be chosen first according to the desired speed of disturbance rejection and degree of system robustness and then polynomial $A(z^{-1})$ is determined to satisfy relation (10).

According to the principle of absorption, it is possible to design the observer estimator that reject any kind of expected disturbances. To this end, suppose the class of disturbances having the *z*-transform $w(z^{-1}) = N(z^{-1})/D(z^{-1})$. Then, relation (10) is satisfied if the following Diophantine equation holds

$$z^{-1-k}A(z^{-1}) + B_1(z^{-1})\Phi(z^{-1}) = C(z^{-1}) \tag{11}$$

where $\Phi(z^{-1})$ represents the absorbtion filter determined by $\Phi(z^{-1}) \equiv D(z^{-1})$. For example, to the polynomial and sinusoidal disturbances ($w(t) = \sum_{i=1}^{m} d_i t^{i-1}$ and $w(t) = \sin \omega t$) correspond respectively $\Phi(z^{-1}) = (1 - z^{-1})^{m+1}$ and $\Phi(z^{-1}) = 1 - 2z^{-1}\cos \omega T_s + z^{-2}$, where $T_s$ is the sampling period.

A single solution of the Diophantine equation, which plays a crucial role in the design procedure of the observer estimator, proposed in this paper, does not exist [12]. Relation (11) is a linear equation in polynomials $A(z^{-1})$ and $B_1(z^{-1})$. Generally, the existence of solution of the Diophantine equation is given in [11]. According to [11], there always exists the solution of (11) for $A(z^{-1})$ and $B_1(z^{-1})$ if greatest common factor of polynomials $z^{-1-k}$ and $\Phi(z^{-1})$ divides polynomial $C(z^{-1})$; then, the equation has many solutions. The particular solution of (11) is constrained by the fact that the control law must be causal, i.e., $\deg A(z^{-1}) \le \deg C(z^{-1})$. Hence, after choosing a stable polynomial $C(z^{-1})$ and degrees of polynomials $A(z^{-1})$ and $B_1(z^{-1})$, and inserting the absorption polynomial $\Phi(z^{-1})$ that corresponds to an expected external disturbance, polynomials $A(z^{-1})$ and $B_1(z^{-1})$ are calculated be equating coefficients of equal order from the left- and right-hand sides of equation (11).

Polynomial $A(z^{-1})$ obtained by solving (11) guarantees the absorption of expected class of disturbances, while the choice of $C(z^{-1})$ affects the speed of disturbance rejection, system robustness and sensitivity with respect to measuring noise. Good filtering properties and the system efficiency in disturbance rejection are opposite requirements. Therefore, to reduce the noise contamination, the low-pass digital filter may be introduces to modify the internal model of disturbance into

$$\frac{A(z^{-1})}{C(z^{-1})} = \frac{A_f(z^{-1})A_1(z^{-1})}{C(z^{-1})} \tag{12}$$

where $A_f(z^{-1})/C(z^{-1})$ represents the pulse transfer function of low-pass filter and $A_1(z^{-1})$ is a polynomial that satisfies (11) and thus includes the internal model of disturbance, implicitly. The lower bandwidth of low-pass filter corresponds to higher degree of system robustness and vice versa [1,5]. According to [1,4], complex disturbances require higher order of polynomial $A(z^{-1})$ and it will further reduce system robustness with respect to mismatches of plant parameters.

## 4. CONVENTIONAL CONTROLLER WITH INTERNAL MODELS

In the IMPACT structure, the main control loop and disturbance estimator may be designed independently. Hence, the control structure with conventional digital PI or PID controllers often used in control of slow varynig indusrial processes may be modified by including the local control loop with internal models in order to improve the system robustness and to extract an expected class of immeasurable external disturbances. The modified controlling structure is shown in Fig. 2

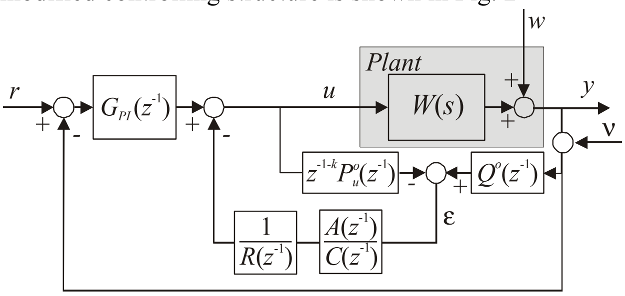


Fig. 2. Modified control structure with conventional digital controller

The design procedure will be illustrated by the example of control system having the plant described by

$$W(s) = \frac{e^{-10s}}{(1+s)(1+0.6s)(1+0.15s)(1+0.1s)}. \tag{13}$$

In the low frequency band, the nominal plant model is identified as [13,14]

$$W^o(s) = \frac{e^{-10.5s}}{1.5s+1}. \tag{14}$$

The zero-hold equivalence pulse transfer function of the nominal plant, with sampling time $T_s = 0.1875s$, is calculated as

$$W^o(z^{-1}) = \frac{z^{-1-k}P_u^o(z^{-1})}{Q^o(z^{-1})} = \frac{0.1175z^{-57}}{1-0.8825z^{-1}}. \tag{15}$$

The sampling time is chosen to be $T_s = \tau/56$ where $\tau = 10.5$ is the process dead-time. Since the control plant is of minimal phase, $R(z^{-1}) = P_u^o(z^{-1}) = 0.1175$ is to be selected [5].

In the main control loop of system in Fig.2, the conventional digital PI controller

$$G_{PID}(z^{-1}) = K_p\left(1 + \frac{T_s/T_I}{1-z^{-1}} + \frac{T_D}{T_s}(1-z^{-1})\right) \tag{16}$$

is applied and its parameters are set by using Dahlin's algorithm to obtain [15,16]

$$K_p = \frac{1-e^{-\lambda T_s}}{K(e^{T_s/T_1}-1)(1+N(1-e^{-\lambda T_s}))}. \tag{17}$$

$$T_s/T_I = e^{T_s/T_1} - 1$$

With $k = 56$, $T_1 = 1.5s$, $K = 1$, and Dahlin's tuning parameter $\lambda = 1/1.5$ one obtains $K_p = 0.1164$ and $T_s/T_1 = 0.1331$.

The inner control loop of the system in Fig. 2 is designed by nominal plant model (15), polynomial $R(z^{-1}) = P_u^o(z^{-1}) = 0.1175$, and digital filter (12). In (12), polynomial $A_1(z^{-1})$ represents the implicit model of disturbance obtained by solving the Diophantine equation (11), and $A_f(z^{-1})/C(z^{-1})$ is the adopted low-pass digital filter which should improve the system robustness and reduce the system sensitivity to measuring noise.

The solution of Diophantine equation (11) with relatively large dead-time $k$ is rather difficult. To simplify the solution, the alternative approach has been proposed [5]. Namely, if we assume the prediction polynomial $D(z^{-1}) = 2 - z^{-1}$ or prediction filter $\Phi(z^{-1}) = (1-z^{-1})^2$ corresponding to extraction of ramp disturbances, then polynomial $A_1(z^{-1})$ in (12) may be split into

$$A_1^{(N)}(z^{-1}) = A_1^{(0)}(z^{-1}) + N(1-z^{-1}) \tag{18}$$

where $A_1^{(N)}(z^{-1}) = A_1^{(0)}(z^{-1}) + N(1-z^{-1})$ and $A_1^{(N)}(z^{-1})$ are the solutions of Diophantine equation for $k = 0$ and arbitrary $k$, respectively. Of course, the solution of equation

$$z^{-1-k}A(z^{-1}) + B_1(z^{-1})\Phi(z^{-1}) \equiv C(z^{-1}) \tag{19}$$

depends upon the assumed absorption filter $\Phi_s(z^{-1}) = (1-z^{-1})^2$ and low-pass filter or polynomial $z^{-1-k}A(z^{-1}) + B_1(z^{-1})\Phi(z^{-1}) \equiv C(z^{-1})$. Thus, if we assume $\Phi(z^{-1}) = (1-z^{-1})^2$ and Butterworth filter $A_f(z^{-1})/C(z^{-1})$ of third order having the bandwidth of $f_0 = 0.05/2T_s = 0.05/2 \cdot 0.1875 = 0.1333$ Hz, the implicit internal model of disturbance (12) is derived as

$$\frac{A(z^{-1})}{C(z^{-1})} = \frac{0.0295 + 0.0593z^{-1} + 0.0012z^{-2} - 0.0577z^{-3} - 0.0290z^{-4}}{1 - 2.6862z^{-1} + 2.4197z^{-2} - 0.7302z^{-3}} \tag{20}$$

In all simulation runs that follow, the reference signal $r(t) = 1 \cdot (t-10)$ is applied and the system is subjected by the slow varying disturbance contaminated by the measuring noise. With PI controller (16) and implicit model of disturbance (20), the system in Fig. 2 was simulated and the results of simulation are shown in Fig. 3. Trace 1 shows that, dispite of the I-action in the main controller, the system cannot reach the required steady-state value in the presence of disturbance. After introducing of local loop, the system eliminates the disturbance in the steady-state (Trace 2). However, the derivative prediction polynomial $D(z^{-1}) = 2 - z^{-1}$ imbeded into the implicit model of disturbance (20) produces significant fluctuations of the system output.

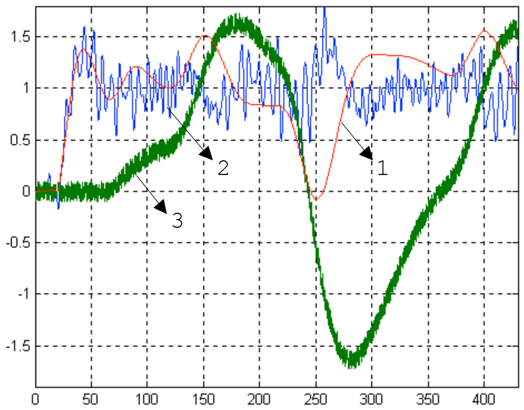


Fig. 3. (1) Output of the system without the local loop. (2) Output of the system with the local loop designed by the internal podel of ramp disturbance. (3) Disturbance.

In the second simulation, the same sampling time is applied and the local loop is designed with the same low-pass filter and predicton polynomial $\Phi(z^{-1}) = 1 - z^{-1}$ that corresponds to a constant disturbance.

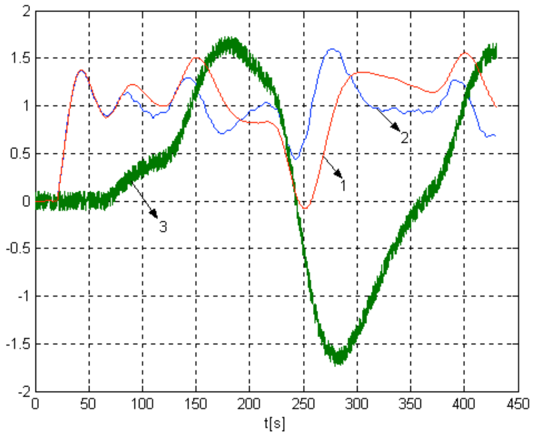


Fig. 4. (1) Output of the system without the local loop. (2) Output of the system with the local loop designed by the internal podel of constant disturbance. (3) Disturbance.

In this case, the system is not able to eliminate the disturbance completely (Fig. 4). However, fluctuations of the system output are notably suppressed.

Fig. 5 shows the simulation of the system in which the local loop is designed by the internal model of ramp disturbance and low-pass filter with reduced bandwidth of $f_0 = 0.025 / 2T_s = 0.025 / 2 \cdot 0.1875 = 0.0667$ Hz. Comparing Figs. 3 and 5, one can notice that fluctuations of of the system output in Fig. 5 are radically reduced.

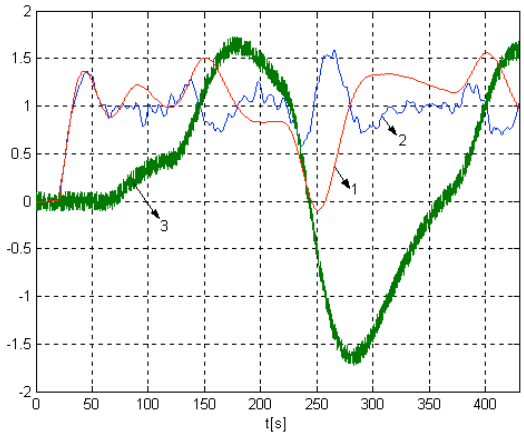


Fig. 5. (1) Output of the system without the local loop. (2) Output of the system with the local loop designed by the internal podel of ramp disturbance. (3) Disturbance.

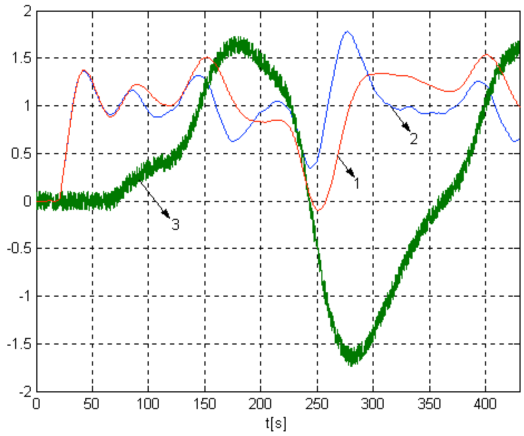


Fig. 6. (1) Output of the system without the local loop. (2) Output of the system with the local loop designed by the internal model of ramp disturbance. (3) Disturbance.

Finally, the local loop is designed by the low-pass filter of $f_0 = 0.025 / 2T_s = 0.025 / 2 \cdot 0.1875 = 0.0667$ and prediction polynomial corresponding to constant disturbances. Simulation results are shown in Fig. 6. Notice that now fluctuations of the system output disapeared but the disturbance is not completely eliminated.

## 5. EXPERIMENTAL SETUP

Eksperimentalna analiza ponašanja predloženog algoritma upravljanja je izvršena na Process Trainer Feedback PT-326, tj. laboratorijskom modelu procesa sušenja materije gde je potrebno izvršiti regulaciju temperature vazduha. Korišćena eksperimentalna insta-lacija je šematski prikazana na Sl.7. Digitalni zakon upravljanja je implementiran korišćenjem PC računara, sa periodom odabiranja od 0.11s, pri čemu je obezbeđena A/D i D/A konverzija sa 12-bitnom rezolucijom. Snaga na rešetkastom grejaču, otpornosti 120 Ω, se kontroliše tiristorskim blokom, odnosno naponskim ulazom od 0V do 10V, u opsegu od 15W do 85W. Protok vazduha kroz cev se određuje otvaranjem ili zatvaranjem segmentnog zatvarača ventilatora u opsegu $\alpha \in [10^o, 170^o]$. Senzor temperature u cevi je termistor koji je ugrađen u plastični čep (osetljivost termistora zavisi od ugla zaokrenutosti čepa u odnosu na struju vazduha (β)).

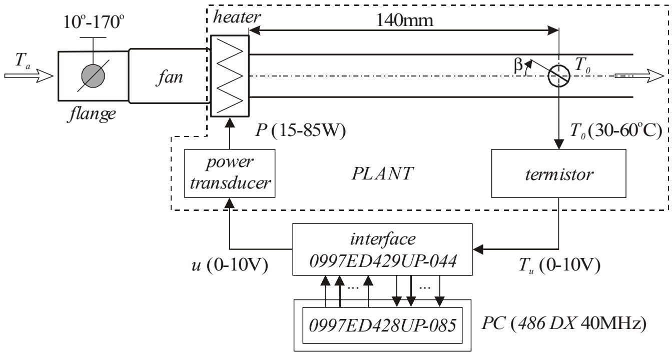


Fig. 7. *Experimental setup*

Objekat upravljanja je nelinearan i nestacionaran [5], ali pripada klasi industrijskih procesa čije se dinamičko ponašanje u okolini radne tačke može dobro aproksimirati funkcijom prenosa prvog reda

$$W_{ou}(s)=\frac{Ke^{-\tau s}}{Ts+1}$$

Svi parametri procesa su promeljivi i zavise od nominalnih radnih uslova i poremećaja (vrednosti protoka, ambijentalne temperature, zagrejanosti cevi, ugla zakretanja kućišta senzora temperature, upravljane temperature vazduha, temperature rešetkastog grejača, itd.) [5].

Za izabranu strukturu modela objekta upravljanja, i radni režim (nominalne uslove) $u_{nom}$=5V, $\alpha$=50°, $\beta$=30°, sprovedena je parametarska identifikacija modela [14]. Usrednjavanjem rezultata identifikacije, dobijenih iz devet setova snimljenih podataka, usvojeni su sledeći parametri

$$K=0.33,\; T=0.75 \;\text{ i }\; \tau=0.33 \qquad (21)$$

koji su korišćeni za podešavanje parametara upravljačkog algoritma.Usvojen je parametar $\lambda$=0.25 Dahlinovog PI regulatora. A kao NF filtar koji je implementiran u unutrašnjem modelu, izabran je Butterworthov filtar trećeg reda i propusnog opsega $f_o$=0.05$f_s$/2=0.05/(2$T_s$)=0.23Hz. Korišćeni su unutrašnji modeli odskočnog i nagibnog (linearnog) poremećaja, a dobijeni su i eksperimentalni rezultati i za slučaj kada unutrašnji model uopšte nije korišćen. U cilju ponovljivosti rezultata i adekvatnog poređenja različito podešenih struktura, neke klase determinističkih poremećaja su softverski setovane. Naime, na izmerenu vrednost izlaza, posle AD konverzije, superponira se vrednost efekata poremećaja na izlaz sistema *w*(*t*) (videti Sl. 1 i 8). Ovako setovan poremećaj je ekvivalentan stvarnoj situaciji koja može nastati u sistemu (ako zanemarimo dodatne perturbacije parametara koje mogu nastati usled nelinearnosti procesa, jer ovakvim "setovanjem poremećaja" zapravo menjamo radnu tačku procesa uprkos zadatoj referentnoj vrednosti). To je potpuno ekvivalentno tome da se temperatura vazduha u cevi promeni za superponirani iznos *w*(*t*), što se moglo desiti kao rezultat promene temperature ulazne struje vazduha i/ili uticaja temperature cevi. Na ovaj način, stvorena je mogućnost da se na najneposredniji način razmotri saglasnost eksperimentalnih rezultata sa rezultatima simulacije (pošto su funkcije i referentne trajektorije i poremećaja poznate). Pri tome, simulira se rad sistema u okolini nominalne radne tačke, pod pretpostavkom da nema parametarskih perturbacija, ni mernog šuma, dok se linearnost sistema podrazumeva. Eksperimentalni rezultati su dati zajedno sa rezultatima simulacije. Razlika od realnih eksperimentalnih krivih je dovoljno očigledna te je na slikama koje slede izostao komentar te prirode. Sagledava se efikasnost tri upravljačka algoritma: 1) algoritma bez unutrašnjeg modela, koga čini PI regulator (Fig. 9 – A1,B1); 2) algoritma sa unutrašnjim modelom odskočnih poremećaja (Fig. 9 – A2,B2); i 3) algoritma sa unutrašnjim modelom linearnih (nagibnih poremećaja) (Fig. 9 – A3,B3).

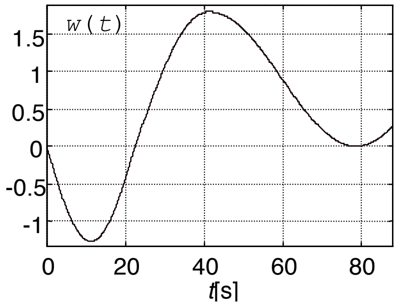


Fig. 8. *Efekat poremećaja w(t)*

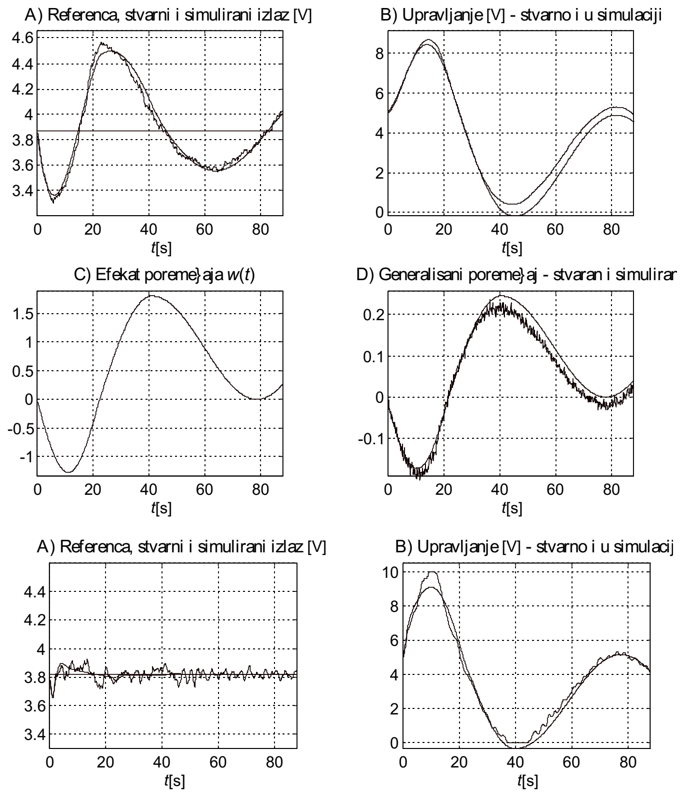


Fig. 9. *Rad sistema u uslovima dejstva poremećaja w(t):* 1–(A1,B1) *bez unutrašnjeg modela* 2-(A2,B2) *sa unutrašnjim modelom odskočnog poremećaja* 3-(A3,B3) *sa unutrašnjim modelom nagibnog poremećaja*

Korišćenje unutrašnjeg modela linearnih poremećaja (Sl.9. A3) daje očigledno bolji rezultat od slučaja kada je upotrebljen unutrašnji model odskočnih poremećaja (Sl.9. A2), dok su oba rezultata daleko bolja u odnosu na slučaj kada unutrašnji model nije ni korišćen (Sl.9. A1). Širi pregled eksperimentalnih rezultata dat je u [5], pri čemu svi dobijeni rezultati u potpunosti verifikuju upotrebnu vrednost predloženog algoritma.

## 6. CONCLUSION

U radu je predstavljena Tsypkinova IMPACT struktura, sa osnovnim konceptom njene parametarske sinteze, koji je sada osetno unapre|en. Sinteza strukture se realizuje re{avanjem dve Dipophantineove jedna~ine, {to mo`e biti zametan zadatak. Samim tim, prepode{avanje strukture nije jednostavan zadatak u tipi~nim industrijskim primenama. Zato su razmotrene mogu}nosti modifikacije strukture u cilju prilago|avanja konkretnim primenama i pojednostavljenja postupka njene parametarske sinteze i pode{avanja parametara. Struktura je svedena na jednostavniju, koja mo`e biti shva}ena kao PI/PID algoritam oboga}en unutra{njim modelom. Dobijeni algoritam mo`e biti realizovan sa razli~itim varijacijama, a sama parametrizacija je pogodna za implementaciju raznih modifikacija (mogu}e kori{}enje razli~itih formi PI/PID algoritama, mehanizama samopode{avanja, adaptacije unutar unutra{njeg modela, itd...). Dakle, postoje}i algoritam, sem svoje upotrebne vrednosti u upravljanju industrijskih procesa, pru`a odli~nu osnovu za razvoj novih algoritama visokih performansi. Algoritam poseduje mali broj pode{ljivih parametara sa jasnim fizi~kim zna~enjem. Parametar Dahlinovnog PI/PID algoritma $\lambda$ ima uticaja na filtarske i robustne osobina sistema, ali i na dinamiku apsorpcije poreme}aja. Unutra{nji model se sastoji iz unutra{njeg modela objekta upravljanja, koji je odre|en parametrima objekta, i unutra{njeg modela poreme}aja, u kome je implementiran NF filtar adekvatnog propustnog opsega $f_o$ i model poreme}aja. Manja frekvencijska propustnost NF filtra omogu}ava manju osetljivost sistema na {um, {iru oblast robustne stabilnosti, mirnije promene upravlja~kog signala, ali i manju efikasnost u pogledu apsorpciije efekata spoljnih poreme}aja. Dana{nja digitalna tehnika omogu}ava realizaciju upravlja~kog softvera koji bi korisniku stavio na raspolaganje izbor adekvatnog unutra{njeg modela poreme}aja, a da on ne mora da razume principe njegove sinteze. Pri tom, upravlja~ki parametri bi bili frekventna propustnost $f_o$ (izbor iz skupa diskretnih vrednosti) i model poreme}aja (od prakti~nog zna~aja bi bio izbor izme|u odsko~nog i nagibnog (linearnog) karaktera poreme}aja). Kako postoji integralno dejstvo u direktnoj grani, sa unutra{njim modelom odsko~nog/nagibnog poreme}aja mogu}a je apsorpcija uticaja nagibnog/paraboli~nog poreme}aja na izlaznu promenljivu. Izborom unutra{njeg modela odsko~nog poreme}aja, pored skromnije apsorpcije efekata poreme}aja na izlaz sistema, dobijaju se ne{to mirnije promene upravlja~kog signala, manja osetljivost sistema na {um i {ira oblast robustne stabilnosti. Ipak, generalno, adekvatnim izborom NF filtra mogu}e je posti}i zadovoljavaju}e performanse sistema kori{}enjem i slo`enijih tipova modela poreme}aja u unutra{njem modelu. Eksperimentalni rezultati su verifikovali upotrebnu vrednost predlo`enog algoritma. PID pro{iren unutra{njim modelom ostvaruje dobru robustnu performansu i ima o~iglednu prednost u pogledu apsorpcije proizvoljne klase poreme}aja.

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