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Research on First Order Delays System Automation

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Abstract

Many of industrial plant require high performance and linear operation; higher density position and/or incremental PID can be used to integrate large amounts of control methodology in a single methodology. This work, proposes a developed method to design PID controller (PID) with optimal-tunable gains method using PC-based method. Many industrial processes can be represented by a first order model. The time delay occurs when a sensor or an actuator are used with a physical separation. The method used to design a PID is to design it as Proportional – derivative controller (PDC) and proportional – integral controller (PIC) connected in parallel through a summer. PIC is designed by accumulating the output of PDC. This method contributes to avoid writing a huge number of fuzzy rules and to reduce the memory considerations in digital design.

Keywords: *First order delays system, position PID controller, PID incremental controller, online gain tuning method*

1. Introduction and Background

In recent years, linear controller has been successfully applied to a large number of linear and nonlinear systems. Linear controller provides an alternative to PID controller since it is a good tool to control the systems that are difficult in modeling. Proportional-Integral-Derivative (PID) controllers are more sufficient than classical Proportional-Derivative (PD) or Proportional-Integral (PI) controllers because they can cover a much wider range of operating conditions. Control action in PID controllers can be expressed with simple model-free techniques. Given the dominance of conventional PID control in industrial control, it is significant both in theory and in practice if a controller can be found that is capable of outperforming the PID controller with comparable ease of use. Some of PID controllers are quite close to this dream. There are several types of control systems that use linear PD, PI and PID controllers as an essential system component. The majority of applications during the past two decades belong to the class of PIDC in industries. These controllers can be further classified into three types: the direct action (DA) type, the gain scheduling (GS) type and a combination of DA and GS types. The majority of PIDC applications belong to the DA type; here the PID controller is placed within the feedback control loop, and computes the PID coefficients through trial and error. In GS type controllers, supervisory technique is used to compute the individual PID gains. From the recent years, the majority of the research work on PID controllers focuses on the two-input PI or PD type controller. However, PID controller design is still a complex task due to the involvement of a large number of parameters in defining the coefficients and the rates of $K_p, K_v\ and\ K_i$ with each others. By expressing the coefficients in different forms, each PID structure is distinctly identified. The simple analytical procedure has developed to deduce the closed form solution for a three-input

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controller. This solution is used to identify the PID action of each structure type in the dissociated form [1-6]. The solution for SISO nonlinear system illustrates the effect of nonlinearity tuning. The design of a PID controller is then treated as a two-level tuning problem. The first level tunes the PID gains and the second level tunes the initial on-line tuning gains, including scale factors of PID variables [7-9]. The two type gains are deduced and explicitly have been presented by assigning a minimum time. Tuning of the characteristics of different PID structures is evaluated with respect to their functional behaviors. Proportional type control is used to responds immediately to difference of control input variables by immediately changing its influences variables, but this type of control is unable to eliminate the control input difference. PD controller is widely used in control process where the results are sensitive to exceeded of set point. This controller, like Proportional controller, has permanent variation in presence of self-limitation control. The Derivative component in this type of methodology is used to cancel outs the change process variables change in presence of quick change in controllers input. Integral term category, integrate the input signal deviation over a period of time. This part of controller is used to system stability after a long period of time. In contrast of Proportional type of controller, this type of controller used to eliminate the deviation. According to integral type of controller, it takes relatively long time [10-13]. The proportional type controller used to immediately response to the input variations. The proportional-integral (PI) controller has the advantages of both proportional and integral controller; it is rapid response to the input deviation as well as the exact control at the desired input. The combination of proportional (P) component, integral (I) component with a derivative (D) controller offered advantages in each case. This type of controller has rapid response to the input deviation, the exact control at the desired input as well as fast response to the disturbances. The PID controller takes the error between the desired joint variables and the actual joint variables. A proportional-derivative integral control system can easily be implemented. This method does not provide sufficient control for systems with time-varying parameters or highly nonlinear systems. An Important question which comes to mind is that why this proposed methodology should be used when lots of control techniques are accessible? Answering to this question is the main objective in this part. First order delay system is nonlinear and delay system. The problem of nonlinearity can be reduced in linear control technique, with the following two methods [14-15]:

- Limiting the performance of the system
- System linearization

Therefore linear type of controller, such as PD or PID cannot be having a good performance. Consequently, to have a good performance, linearization and decoupling without using many gears, online tuning control methodologies is presented and applied to linear control technique [16-18].

This paper is organized as follows; second part focuses on the system modeling dynamic formulation. Third part is focused on the methodology. Simulation result and discussion is illustrated in forth part. The last part focuses on the conclusion and compare between this method and the other ones.

2. Theory

Delay First Order Plant:

Many industrial processes can be represented by a first order model; equation (1) shows the mathematical plant model (in *s-plane*). Discrete transfer function of this model has obtained using ZOH method, and the selected sampling period (T) is 0.1, equation (2) shows the discrete transfer functions, (in *z-plane*).

$$CS_1(s) = \frac{1}{S+1} \quad (1)$$

And;

$$CS_1(z) = \frac{0.09516}{Z - 0.9048}, T = 0.1 \quad (2)$$

The time delay occurs when a sensor or an actuator are used with a physical separation. Equation (3) shows the mathematical plant model (in *s-plane*). Discrete transfer functions of this model has been obtained using ZOH method, and the selected sampling period (T) is 0.1, equation (4 and 5) show the discrete transfer functions, (in *z-plane*).

$$CS_2(s) = \frac{1}{S^2 \times (S+1)} \quad (3)$$

$$CS_2(z) = Z^{-2} \times CS_1(z) \quad (4)$$

$$CS_2(z) = Z^{-2} \times \frac{0.09516}{Z - 0.9048}, T = 0.1 \quad (5)$$

3. Methodology

In a P controller the control deviation $e(t)$ is produced by forming the difference between the process variable $y_{p(t)}$ and the desired output$y_{d(t)}$; this is then amplified to give the manipulating variable, which operates a suitable actuator. The P controller simply responds to the magnitude of the deviation and amplifies it. As far as the controller is concerned, it is unimportant whether the deviation occurs very quickly or is present over a long period. Beside P component, there are other control components that behave in the same way mentioned above: “D” component responds to changes in the process variable, and “I” component responds to the duration of the deviation. It sums the deviation applied to its input over a period of time. The D and I components, are often combined with a P component to give PI, PD or PID controllers. The PID control law could be represented in two forms, positional form and incremental form.

Assume:

$$e(t) = y_d(t) - y_p(t) \quad (6)$$

$$e(n) = Sample\ (e(t)) \quad (7)$$

$$r(n) = e(n) - e(n-1) \quad (8)$$

$$a(n) = r(n) - r(n-1) \quad (9)$$

The continuous-time linear PID controller in position form is described by the following equation:

$$U(t) = K(e(t) + \frac{1}{T_i}\int_0^t e(\tau)d\tau + T_d\frac{de(t)}{dt} \quad (10)$$

With time$t$, being continuous instead of discrete, K is a gain, T_i is integration time, and T_d is derivative time. The corresponding discrete-time position form is:

$$U(n) = K\left\{e(n) + \frac{T}{T_i}\sum e(i) + \frac{T_d}{T} r(n)\right\} = Ke(n) + \frac{K.T}{T_i}\sum e(i) + \frac{KT_d}{T} r(n) = K_p e(n) + K_i \sum e(i) + K_d\, r(n) = \quad (11)$$

Where T is the sampling period, $Kp, Ki, and\ Kd$ are the proportional gain, integral gain and derivative gain of the PID controller, respectively.

The above PID control algorithms are in position form because they directly compute the controller output itself. The PID controller is often used in the incremental form, in which the controller calculates change of the controller output. Note that at sampling time *n*-1,

$$U(n-1) = K_p e(n-1) + K_i \sum e(i) + K_d\, r(n-1) \quad (12)$$

Hence, the incremental form of the PID controller corresponding to Equation (11) is:

$$\Delta U(n) = U(n) - U(n-1) = K_p r(n) + K_i e(n) + K_d a(n) \quad (13)$$

On the other hand, the integral term can cause slower system response and larger system overshoot; it should not be included in certain applications of PID control. When Kd is set to zero in Equation (13), the PID controller becomes a PI controller in incremental form:

$$\Delta U(n) = K_p r(n) + K_i e(n) \quad (14)$$

Whereas when $Ki = 0$ in Equation (13), the PID controller reduces to a PD controller in incremental form:

$$\Delta U(n) = K_p r(n) + K_d a(n) \quad (15)$$

A PI controller in incremental form is related to a PD controller in position form. Letting $Ki = 0$ in Equation (13), a PD controller is obtained in position form:

$$U(n) = K_p e(n) + K_d r(n) \quad (16)$$

Now by comparing Equation (16) with Equation (14), it could see that the PD controller in position form becomes the PI controller in incremental form if (1) $e(n)\ and\ r(n)$ exchange positions, (2) Kd is replaced by Ki, and (3) $u(n)$ is replaced by $\Delta u(n)$ as shown in Figure(1).

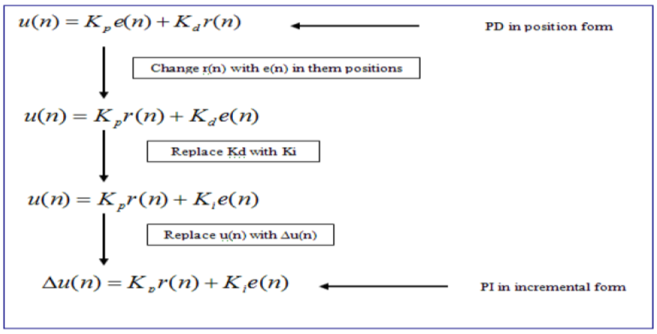

Figure 1. Steps Required Making the PD in Position Form Works as PI in Incremental Form

In PD controller the control law is given by the following equation;

$$\boldsymbol{\tau} = \boldsymbol{K_{p_i} e} + \boldsymbol{K_{v_i} \dot{e}} \tag{17}$$

Where $\boldsymbol{e} = \boldsymbol{q_{i_d}} - \boldsymbol{q_{i_a}}$ and $\boldsymbol{\dot{e}} = \boldsymbol{\dot{q}_{i_d}} - \boldsymbol{\dot{q}_{i_a}}$.

In this theory $\boldsymbol{K_{p_i}}$ **and** $\boldsymbol{K_{v_i}}$ are positive constant. To show this controller is stable and achieves zero steady state error, the Lyapunov function is introduced;

$$\boldsymbol{V} = \frac{\boldsymbol{1}}{\boldsymbol{2}} \left[\boldsymbol{\dot{q}^T S \dot{q}} + \boldsymbol{e^T K_p e}\right] = \frac{\boldsymbol{1}}{\boldsymbol{2}} \frac{\boldsymbol{d}}{\boldsymbol{dt}} [\boldsymbol{\dot{q}^T S \dot{q}}] = \boldsymbol{\dot{q}\, U} \tag{18}$$

If the conversation energy is written by the following form:

$$\frac{\boldsymbol{1}}{\boldsymbol{2}} \frac{\boldsymbol{d}}{\boldsymbol{dt}} [\boldsymbol{\dot{q}^T S \dot{q}}] = \boldsymbol{\dot{q}\, U} \tag{19}$$

Where $(\boldsymbol{\dot{q}\, U})$ shows the power inputs and $\frac{\boldsymbol{1}}{\boldsymbol{2}} \frac{\boldsymbol{d}}{\boldsymbol{dt}} [\boldsymbol{\dot{q}^T S \dot{q}}]$ is the derivative of the kinetic energy

$$\boldsymbol{\dot{V}} = \boldsymbol{\dot{q}^T}\left[\boldsymbol{U} + \boldsymbol{K_p e}\right] \tag{20}$$

Based on $\boldsymbol{U} = -\boldsymbol{K_{p_i} e} - \boldsymbol{K_{v_i} \dot{e}}$ **,** we can write:

$$\boldsymbol{\dot{V}} = \boldsymbol{\dot{q}^T K_p \dot{q}} \leq \boldsymbol{0} \tag{21}$$

If $\boldsymbol{\dot{V}} = \boldsymbol{0}$, we have

$$\boldsymbol{\dot{q}} = \boldsymbol{0} \rightarrow \boldsymbol{\ddot{q}} = \boldsymbol{0} \rightarrow \boldsymbol{\ddot{q}} = \boldsymbol{A^{-1} K_p e} \rightarrow \boldsymbol{e} = \boldsymbol{0} \tag{22}$$

In this state, the actual trajectories converge to the desired state.

To design on-line tuning linear methodology three parameters are introduced: θ_j, σ_j^l , α_j^l and E_j. The adaptation laws are expressed as

$$\boldsymbol{\dot{\theta}_j} = \boldsymbol{\eta_{j2} S_j \varphi_j} \tag{23}$$

$$\boldsymbol{\dot{\alpha}_j} = \boldsymbol{\eta_{j3} S_j B_j^T \theta_j} \tag{24}$$

$$\boldsymbol{\dot{\sigma}_j} = \boldsymbol{\eta_{j4} s_j C_j^T \theta_j} \tag{25}$$

$$\boldsymbol{f_{cpj}(s_j)} = \boldsymbol{E_j(s_j)} \tag{26}$$

$$\boldsymbol{\dot{E}_j} = \boldsymbol{\eta_{j1}} |\boldsymbol{s_j}| \tag{27}$$

Where $\boldsymbol{\eta_{j1}}, \boldsymbol{\eta_{j2}}, \boldsymbol{\eta_{j3}}$ and $\boldsymbol{\eta_{j4}}$ are positive constants; $\boldsymbol{\theta_j} = [\theta_j^1,\ \theta_j^2, \dots, \theta_j^M]^T$, $\boldsymbol{\alpha_j} = [\alpha_j^1,\ \alpha_j^2, \dots, \alpha_j^M]^T$, $\boldsymbol{\sigma_j} = [\sigma_j^1,\ \sigma_j^2, \dots, \sigma_j^M]^T$; B_j , C_j are given in

$$B_j = \begin{bmatrix} \frac{\partial\varphi_j^1}{\partial\alpha_j^1} & \frac{\partial\varphi_j^2}{\partial\alpha_j^1} & \cdots & \frac{\partial\varphi_j^M}{\partial\alpha_j^1} \\ \frac{\partial\varphi_j^1}{\partial\alpha_j^2} & \frac{\partial\varphi_j^2}{\partial\alpha_j^2} & \cdots & \frac{\partial\varphi_j^M}{\partial\alpha_j^2} \\ \vdots & & & \\ \frac{\partial\varphi_j^1}{\partial\alpha_j^M} & \frac{\partial\varphi_j^2}{\partial\alpha_j^M} & \cdots & \frac{\partial\varphi_j^M}{\partial\alpha_j^M} \end{bmatrix},$$

$$C_j = \begin{bmatrix} \frac{\partial\varphi_j^1}{\partial\alpha_j^1} & \frac{\partial\varphi_j^2}{\partial\alpha_j^1} & \cdots & \frac{\partial\varphi_j^M}{\partial\alpha_j^1} \\ \frac{\partial\varphi_j^1}{\partial\alpha_j^2} & \frac{\partial\varphi_j^2}{\partial\alpha_j^2} & \cdots & \frac{\partial\varphi_j^M}{\partial\alpha_j^2} \\ \vdots & & & \\ \frac{\partial\varphi_j^1}{\partial\alpha_j^M} & \frac{\partial\varphi_j^2}{\partial\alpha_j^M} & \cdots & \frac{\partial\varphi_j^M}{\partial\alpha_j^M} \end{bmatrix}$$

; $\boldsymbol{f_{cpj}(s_j)}$ is the compensation term.
If the following Lyapunov function candidate defined by:

$$\boldsymbol{V = \frac{1}{2}s^T U s + \frac{1}{2}\sum_{j=1}^{m}\left(\frac{\tilde{E}_j^2}{\eta_{j1}} + \frac{\tilde{\theta}_j^T\tilde{\theta}}{\eta_{j2}} + \frac{\tilde{\alpha}_j^T\tilde{\alpha}}{\eta_{j3}} + \frac{\tilde{\sigma}_j^T\tilde{\sigma}}{\eta_{j4}}\right)} \quad \textbf{(28)}$$

The derivative of V is defined by:

$$\boldsymbol{\dot{V} = s^T U\dot{s} + \frac{1}{2}s^T\dot{U}s + \sum_{j=1}^{m}\left(\frac{\tilde{E}_j\dot{\tilde{E}}_j}{\eta_{j1}} + \frac{\tilde{\theta}_j^T\dot{\tilde{\theta}}}{\eta_{j2}} + \frac{\tilde{\alpha}_j^T\dot{\tilde{\alpha}}}{\eta_{j3}} + \frac{\tilde{\sigma}_j^T\dot{\tilde{\sigma}}}{\eta_{j4}}\right)} \quad \textbf{(29)}$$

where $\boldsymbol{\tilde{E}_j = E_j^* - E_j}$, $\boldsymbol{\tilde{\theta}_j = \theta_j^* - \theta_j}$, $\boldsymbol{\tilde{\alpha}_j = \alpha_j^* - \alpha_j}$, $\boldsymbol{\tilde{\sigma}_j = \sigma_j^* - \sigma_j}$.
Then $\dot{V}$ becomes

$$\boldsymbol{\dot{V} = s^T(U\dot{s} + C_1 s) + \sum_{j=1}^{m}\left(\frac{\tilde{E}_j\dot{\tilde{E}}_j}{\eta_{j1}} + \frac{\tilde{\theta}_j^T\dot{\tilde{\theta}}}{\eta_{j2}} + \frac{\tilde{\alpha}_j^T\dot{\tilde{\alpha}}}{\eta_{j3}} + \frac{\tilde{\sigma}_j^T\dot{\tilde{\sigma}}}{\eta_{j4}}\right)} \quad \textbf{(30)}$$

$$\boldsymbol{= s^T(F - F^{\wedge}(s) - F_{cp}(s)) + \sum_{j=1}^{m}\left(\frac{\tilde{E}_j\dot{\tilde{E}}_j}{\eta_{j1}} + \frac{\tilde{\theta}_j^T\dot{\tilde{\theta}}}{\eta_{j2}} + \frac{\tilde{\alpha}_j^T\dot{\tilde{\alpha}}}{\eta_{j3}} + \frac{\tilde{\sigma}_j^T\dot{\tilde{\sigma}}}{\eta_{j4}}\right)}$$

$$\boldsymbol{= \sum_{j=1}^{m} s_j(f_j - f^{\wedge}{}_j(s_j) - f_{cpj}) + \sum_{j=1}^{m}\left(\frac{\tilde{E}_j\dot{\tilde{E}}_j}{\eta_{j1}} + \frac{\tilde{\theta}_j^T\dot{\tilde{\theta}}}{\eta_{j2}} + \frac{\tilde{\alpha}_j^T\dot{\tilde{\alpha}}}{\eta_{j3}} + \frac{\tilde{\sigma}_j^T\dot{\tilde{\sigma}}}{\eta_{j4}}\right)}$$

$$\boldsymbol{= \sum_{j=1}^{m} s_j(\theta_j^T B_j\tilde{\alpha}_j + \theta_j^T C_j\tilde{\sigma}_j + \tilde{\theta}_j^T\varphi_j + \varepsilon_j - f_{cpj}) - \sum_{j=1}^{m}\left(\frac{\tilde{E}_j\dot{\tilde{E}}_j}{\eta_{j1}} + \frac{\tilde{\theta}_j^T\dot{\tilde{\theta}}}{\eta_{j2}} + \frac{\tilde{\alpha}_j^T\dot{\tilde{\alpha}}}{\eta_{j3}} + \frac{\tilde{\sigma}_j^T\dot{\tilde{\sigma}}}{\eta_{j4}}\right)}$$

$$\boldsymbol{= \sum_{j=1}^{m}\left[\tilde{S}_j^T\left(s_j\varphi_j - \frac{\dot{\theta}_j}{\eta_{j2}}\right) + \tilde{\alpha}_j^T\left(s_j B_j^T\theta_j - \frac{\dot{\alpha}_j}{\eta_{j3}}\right) + \tilde{\sigma}_j^T\left(s_j C_j^T\theta_j - \frac{\dot{\sigma}_j}{\eta_{j4}}\right)\right] + \sum_{j=1}^{m}\left(s_j\varepsilon_j - s_j f_{cpj}\frac{\tilde{E}_j\dot{\tilde{E}}_j}{\eta_{j1}}\right)}$$

$$\begin{aligned}\dot{V} &= \sum_{j=1}^{m}\left[s_j\varepsilon_j - s_jE_j(s_j) - \tilde{E}_j|s_j|\right] \\ &= \sum_{j=1}^{m}\left[s_j\varepsilon_j - s_jE_j(s_j) - (E_j^* - E_j)s_j(s_j)\right] \\ &= \sum_{j=1}^{m}\left[s_j\varepsilon_j - E_j^* s_j(s_j)\right] \\ &= \sum_{j=1}^{m}\left[|s_j||\varepsilon_j| - E_j^*|s_j|\right] \\ &= \sum_{j=1}^{m}\left[|s_j|\left(|\varepsilon_j| - E_j^*\right)\right] \le 0\end{aligned} \quad (31)$$

Where $\dot{V}$ is negative semidefinite. We define $\dot{V}_j = |s_j(t)|\left(|\varepsilon_j| - E_j^*\right)$. From $\dot{V}_j \le 0$, we can get $s_j(t)$ is bounded. We assume $|s_j(t)| \le \eta_s$ and rewrite $|s_j(t)|\left(E_j^* - |\varepsilon_j|\right) \le -\dot{V}_j$ as

$$s_j(t) \le \frac{1}{E_j^*}|s_j(t)||\varepsilon_j| - \frac{1}{E_j^*}\dot{V}_j \le \frac{\eta_s}{E_j^*}|\varepsilon_j| - \frac{1}{E_j^*}\dot{V}_j \quad (32)$$

$$\int_0^t |s_j(v)|dv \le \frac{\eta_s}{E_j^*}\int_0^t |\varepsilon_j|dv + \frac{1}{E_j^*}\left(V_j(0) - V_j(t)\right) \le \frac{\eta_s}{E_j^*}\int_0^t |\varepsilon_j|dv + \frac{1}{E_j^*}\left(|V_j(0)| - |V_j(t)|\right) \quad (33)$$

4. Result and Discussion

Figure 2 shows the result of first order system and first order delay system. Regarding to the following Figure the rise time in control-free first order system is about 2.3 second and in control-free delay first order system is about 3.9 second.

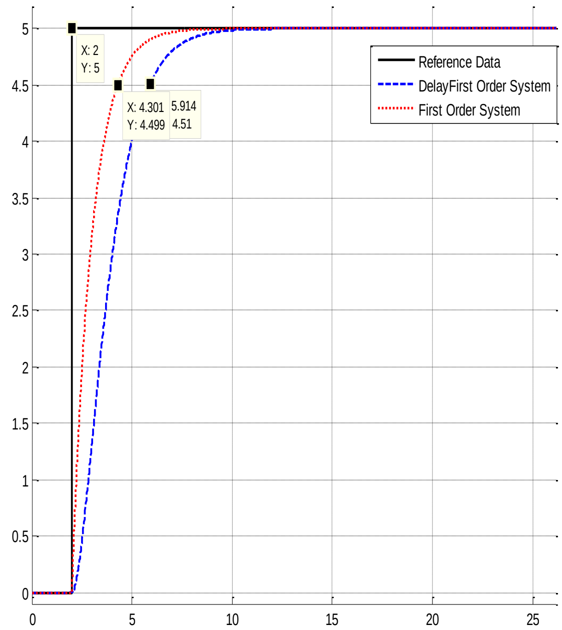

Figure 2. Control free Delay First Order System (DFOS) and Control free First Order System (FOS)

Reduce the rise time: regarding to Fig 3, control delay first order system help to reduce the rise time from 3.9 second to 0.23 second. Regarding to following Figure this method can reduce the rise time about 170%.

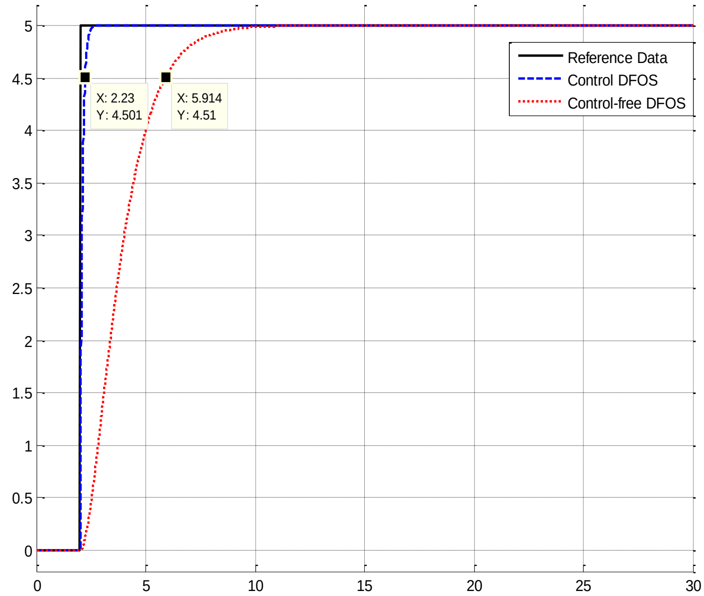

Figure 3. Control Delay First Order System (CDFOS) and Control free Delay First Order System (DFOS)

Power of Disturbance rejection: Figure 4 has shown the power disturbance elimination in control-free first order system and control-free delay first order system. Regarding to the following Figure both of systems has fluctuations in presence of uncertainty because they are control-free.

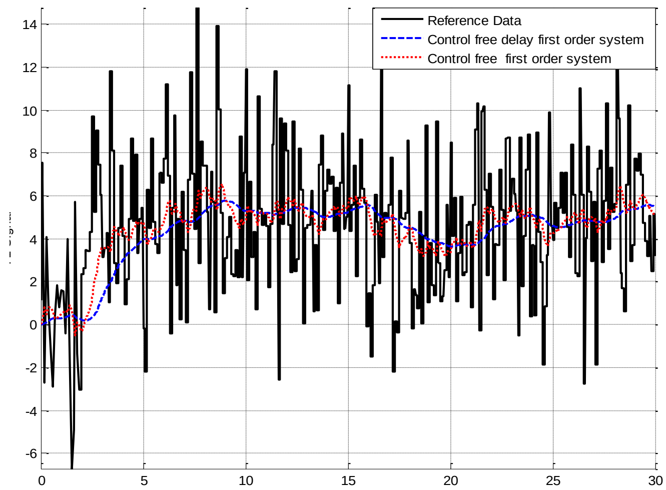

Figure 4. Control Free Delay First Order System (DFOS) and Control Free First Order System (FOS) in Presence of Uncertainty

Figure 5 shows the power of disturbance rejection in control first order delay system. However, this type of controller reduce the rise time (delay) in certain condition but it has fluctuation in presence of uncertainty.

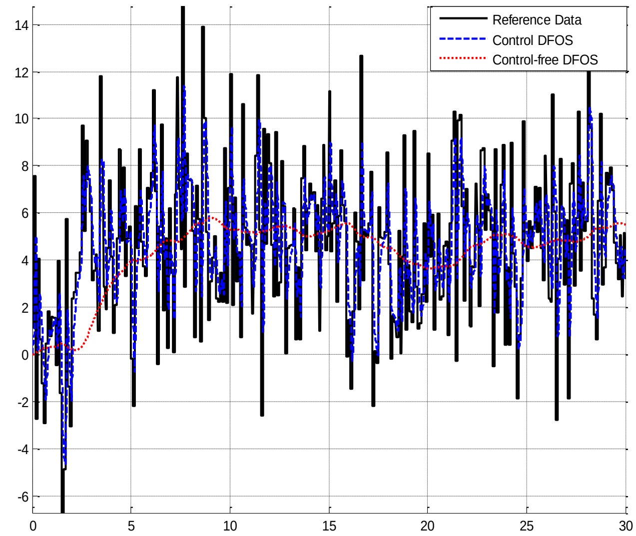

Figure 5: Control Delay First Order System (CDFOS) and Control Free Delay First Order System (DFOS) in Presence of Uncertainty

Regarding to Fig 5 it can be seen that, off-line tuning control technique has fluctuations in presence of uncertainty. To solve this challenge classical on-line tuning method is introduced in this research. Figure 6 shows the power of disturbance rejection in proposed method.

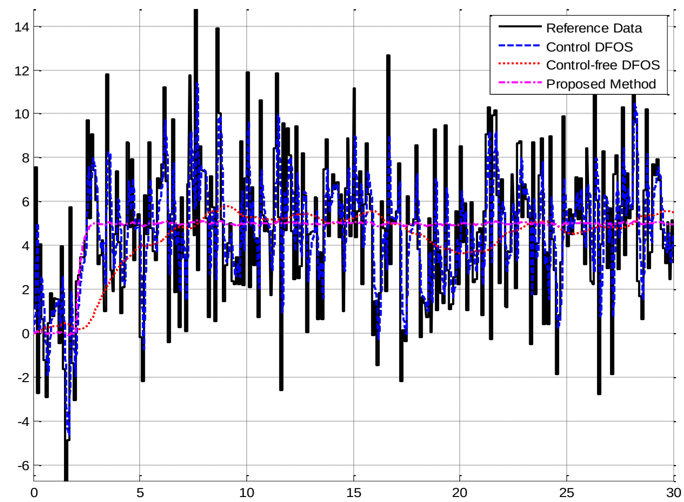

Figure 6. Control DFOS and Control-free DFOS Vs. Proposed method in Presence of Uncertainty

Regarding to above graph, we can reduce the fluctuation more than 200% compare to control DFOS and the overshoot is also reduce more than 100%. On-line tuning control of delay first order system improves the rise time and overshoot in certain and uncertain condition.

5. Conclusion

Classical on-line tuning linear control for first order delay system is investigated in this research. Proposed algorithm utilizes single input-single output classical coefficient tuner to estimate the cross-coupling effects in first order delay system and gets perfect accuracy. However, the classical control has stability in certain condition but it has high frequency fluctuations in presence of uncertainty and external disturbance. To solve this challenge in first order delay system online tuning is applied and reduce the overshoot and fluctuations.

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Iranian center of Advance Science and Technology (IRAN SSP) is one of the independent research centers specializing in research and training across of Control and Automation, Electrical and Electronic Engineering, and Mechatronics & Robotics in Iran. At IRAN SSP research center, we are united and energized by one mission to discover and develop innovative engineering methodology that solve the most important challenges in field of advance science and technology. The IRAN SSP Center is instead to fill a long standing void in applied engineering by linking the training a development function one side and policy research on the other. This center divided into two main units:

- Education unit
- Research and Development unit

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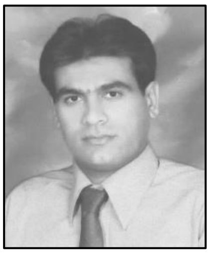

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