Controller Design Methods for Linear Systems with Emphasis on Integrating Processes

A

Thesis Submitted

in Partial Fulfilment of the Requirements

for the Degree of

DOCTOR OF PHILOSOPHY

By

AHMAD ALI

Department of Electronics and Communication Engineering

Indian Institute of Technology Guwahati

Guwahati - 781039, INDIA.

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Certificate

This is to certify that the thesis entitled "**Controller Design Methods for Linear Systems with Emphasis on Integrating Processes**", submitted by Ahmad Ali (04610203), a research scholar in the *Department of Electronics and Communication Engineering, Indian Institute of Technology Guwahati*, for the award of the degree of **Doctor of Philosophy**, is a record of an original research work carried out by him under my supervision and guidance. The thesis has fulfilled all requirements as per the regulations of the Institute and in my opinion has reached the standard needed for submission. The results embodied in this thesis have not been submitted to any other University or Institute for the award of any degree or diploma.

Dated:
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Acknowledgements

First of all, I would like to thank my supervisor Prof. S. Majhi, for his valuable guidance, moral support and endless encouragement. I would also like to thank my doctoral committee members, Prof. P. K. Bora, Dr. C. Mahanta and Dr. H. Nemade. Their suggestions and assistance have been valuable. I am thankful to Mr. Sanjib Das and Mr. L. N. Sarma, research engineers in the department, Mr. Goswami, Mr. Sidananda and Mrs. Jharna for their help during the entire course of this work. I am grateful to all the other faculty members and staff of the department of ECE.

Among my friends, I would like to extend my special thanks to Babusena, Dakua, Ganesh, Jayanna, Katiyar, Karthikeyan, Krishnamoorthy, Manikandan, Mehta, Mrinal, Padhi, Senthil, Shyam, Rupaban and Vinod. I would like to express my heartfelt gratitude to Zaheer, Khan, Shahzad, Monowar and his family for their steadfast support without which this dissertation would not have been completed.

Whenever I was feeling down, my younger brother Safdar was always a call away patiently listening to me cribbing. My parents, brother, and sister were always behind me, regardless of being miles away. Their love and confidence gave me strength when I needed it most.

(Ahmad Ali)

Abstract

Certain new approaches for designing efficient controllers for linear systems, in particular, for integrating processes, are proposed in this thesis. The parameters of the proposed proportional-integral-derivative (PID) controllers and its variants are optimized by minimizing the integral square error (ISE) using the so called bacterial foraging algorithm (BFA). Although ISE is a standard minimization criterion as well as BFA has certain advantages such as better solution quality and good convergence, both of them suffer from some drawbacks. To obtain the robust closed loop performance, the ISE criterion is initially modified by appending time parameters to the said criterion. Similarly, to ensure the convergence of BFA, an adaptive strategy for the run length vector along with two modifications in the standard BFA are also proposed. Next, the parameters of the PID and its variants are estimated using the proposed 'slope of the Nyquist curves' technique. The approach of the said slope of the Nyquist curves method is also extended to design controllers for the modified Smith predictors. Finally, the parameters of the PID and its variants are derived analytically using the user-defined percentage overshoot and process model parameters.

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Nomenclature

BFA	Bacterial foraging algorithm
DIPTD	Double integrating plus time delay process model
FOPTD	First order plus time delay process model
IAE	Integral of absolute error
IFT	Iterative feedback tuning
IFOPTD	Integrating plus first order plus time delay process model
IMC	Internal model control
IPTD	Integrating plus time delay process model
ISE	Integral of error squared
ISTE	Integral of time error squared
IST^{2}E	Integral of time square error squared
IST^{3}E	Integral of time cube error squared
ITAE	Integral of time weighted absolute error
KPGM	Kookos gain and phase margin method
MBFA	Modified bacterial foraging algorithm
P	Proportional controller
PD	Proportional derivative controller
PI	Proportional integral controller
PID	Proportional integral derivative controller
PI-D	Proportional integral - derivative controller
PI-PD	Proportional integral - proportional derivative controller
PSO	Particle swarm optimization
SIMC	Skogestad's IMC tuning rules
SOPTD	Second order plus time delay process model
SOPTDLD	Second order plus time delay with lead process model

Mathematical Notations

A_m	Gain margin
d	Decay ratio
$e(t)$	Error signal
e_{ss}	Steady state error
$F_{sp}(s)$	Setpoint filter
$G_c(s)$	Transfer function of a controller
$G_{mp}(s)$	Minimum phase part of the plant transfer function
$G_{nmp}(s)$	Non minimum phase part of the plant transfer function
$G_p(s)$	Transfer function of a plant
J	Objective function to be optimized
K	Steady state gain of a process
K_p	Proportional gain
K_{pm}	Mean of the proportional gain obtained in various runs of evolutionary algorithm
L	Load disturbance
$L(s)$	Loop transfer function
M_s	Maximum sensitivity
n_g	Slope of the open loop gain at gain crossover frequency
N_c	Number of chemotaxis steps
N_{ed}	Number of elimination and dispersal events
N_s	Swimming length
N_{re}	Number of reproduction steps
P_{ed}	Elimination-dispersal probability
$r(t)$	Setpoint input
S	Number of bacteria used for searching the global minimum
t	Time
t_r	Rise time
t_s	Settling time

T	Time constant of a plant model transfer function
T_d	Derivative time constant
T_{dm}	Mean of the derivative time constant obtained in various runs of evolutionary algorithm
T_i	Integral time constant
T_{im}	Mean of the integral time constant obtained in various runs of evolutionary algorithm
$u(t)$	Control signal
$y(t)$	Output of linear plant
α	Derivative filter constant
ξ	Damping ratio
η	Set of controller parameters
θ	Time delay of a process
λ	IMC filter time constant
ϕ_m	Phase margin
ψ	Slope of the Nyquist curve at any frequency
$\bar{w}$	Weighting factor
ω_g	Gain crossover frequency
ω_n	Natural frequency
ω_u	Phase crossover frequency
$\%os$	Percentage overshoot

List of Publications

Journals

Published:

1. A. Ali and S. Majhi, "PI/PID controller design based on IMC and percentage overshoot specification to controller setpoint change," *ISA Transactions*, vol. 48 (1), pp. 10–15, Jan 2009.

2. A. Ali and S. Majhi, "Controller tuning by an adaptive bacterial foraging strategy," *Journal of Systems Science and Engineering*, vol. 17 (1), pp. 44–51, 2008.

Revision Submitted:

1. A. Ali and S. Majhi, "Maximum sensitivity based controller tuning for stable and integrating plants," Revision submitted to *ISA Transactions.* (ISATRANS-D-08-00063)

2. A. Ali and S. Majhi, "Optimal tuning of PI/PID controller for integrating processes," Revision submitted to *Asian Journal of Control.* (KR-08-086)

3. A. Ali and S. Majhi, "PID controller tuning for integrating processes," Revision submitted to *ISA Transactions.* (ISATRANS-D-08-000136)

Conferences

1. A. Ali and S. Majhi, "PI controller tuning for integrating processes," in *National Systems Conference*, Roorkee, India, Dec. 2008.

2. A. Ali and S. Majhi, "Controller design for unstable FOPTD plants based on sensitivity," in *Proc. of the 17th World Congress, IFAC*, Seoul, Korea, July 2008, pp. 5837-5841.

3. A. Ali and S. Majhi, “A new objective function for controller tuning,” in *Proc. of the 17th World Congress, IFAC*, Seoul, Korea, July 2008, pp. 6120-6124.

4. A. Ali and S. Majhi, “ Controller tuning by an adaptive bacterial foraging strategy,” in *National Systems Conference*, Manipal, India, Dec. 2007.

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Chapter 1

Introduction

1.1 Introduction

Control engineering deals with dynamic systems such as aircraft, spacecraft, ships, trains, automobiles and industrial processes like distillation columns and rolling mills etc. Electrical systems such as motors, generators, and power systems, and machines such as numerically controlled lathes and robots are a few other areas where control engineering also plays an important role. In each case, there are certain dependent variables called outputs which are to be controlled to behave in a prescribed way. For instance, it may be necessary to assign the temperature and pressure at various points in a process, or the position and velocity of a vehicle, or the voltage and frequency in a power system, to given desired fixed values despite uncontrolled and unknown variations at other points in the system. The independent variables called inputs are used to regulate and control the behavior of the system and these may be voltage applied to the motor terminals or valve mechanism. Also, there are unknown and unpredictable disturbances impacting the system which could be for example, the fluctuations of load in a power system, disturbances such as wind gusts acting on a vehicle, external weather conditions acting on an air conditioning plant, or the fluctuating load torque on an elevator motor as passengers enter and exit. The equations describing the plant dynamics and the parameters contained in these equations are not known at all or at best known imprecisely. These considerations suggest the following general representation of the plant or system to be controlled.

In Fig. 1.1 the inputs or outputs shown could actually be representing a vector of signals. In such cases the plant is said to be a *multivariable plant* as opposed to the case where the signals are scalar, in which case the plant is said to be a *scalar or monovariable plant*.

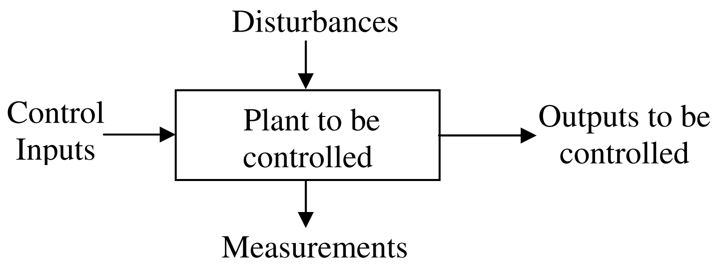


Figure 1.1: A general plant

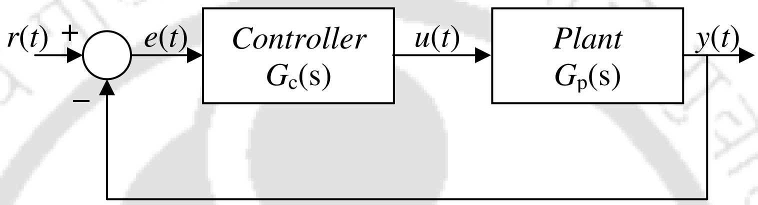


Figure 1.2: A feedback control system

Control is exercised by feedback, which means that the corrective control input to the plant is generated by a device that is driven by the available measurements. Thus the controlled system can be represented by the feedback or closed-loop system shown in Fig. 1.2 where $r(t)$, $e(t)$, $u(t)$ and $y(t)$ are the reference, error, control and output signals. The controller design problem is to find the tuning constants so that the controlled outputs can be

1. Set to prescribed values called *references*

2. Maintained at the reference values despite the unknown disturbances

3. Achieved despite the inherent uncertainties and changes in the plant dynamic characteristics

The first operation is called *tracking*, the second, *disturbance rejection*, and the third, *robust performance* of the system. The simultaneous satisfaction of all three is called *robust tracking and disturbance rejection* and control systems designed to achieve this are called *robust servomechanisms*.

1.2 Types of Feedback Controllers

A proportional, 'P', controller with increased gain typically increases the speed of response but at the same time yields a much larger percentage overshoot. Especially for higher order systems, large values of the proportional controller gain cause instability. For most systems there is an upper limit on the proportional gain in order to achieve a well-damped stable response, and this limit may still have an unacceptable steady state error. One way to improve the steady state accuracy of response without adding extremely high proportional gains is to introduce an additional integral term to the controller. In this case, the controller is called proportional plus integral, 'PI', controller. The steady state error of a system depends on the number of poles at the origin of the s-domain feedforward loop transfer function. Since a PI controller introduces an integrator or a pole at the origin, any steady state error of the system to a step input will be eliminated by the controller. Thus, PI controllers are usually preferred for use with a plant, which has no integrator. A PI controller has the form

$$G_c(s) = K_p + \frac{K_p}{sT_i} = K_p(1 + \frac{1}{sT_i}) \tag{1.1}$$

where K_p is the proportional gain and T_i is the integral time constant. The value of T_i affects the speed of the system response. The control algorithm can be described in the time domain by the following equation

$$u(t) = K_p e(t) + \frac{K_p}{T_i}\int e(t)d(t) \tag{1.2}$$

Equation (1.2) shows that as long as error exists, the PI controller removes the error by integrating it. As a result, a PI controller reduces the steady state error to zero, but this benefit typically comes at the cost of a worse transient response. To eliminate this problem, a derivative, 'D', term is usually added to the PI controller. This controller is called proportional plus integral plus derivative, 'PID', controller. PID controllers are usually used as PI controllers but the derivative term is also used for stabilization of the system. A PID controller has a transfer function

$$G_c(s) = K_p(1 + \frac{1}{T_i s} + T_d s) \tag{1.3}$$

In the time domain, the control algorithm becomes

$$u(t) = K_p\left[e(t) + \frac{1}{T_i}\int e(t)d(t) + T_d\frac{de(t)}{dt}\right] \tag{1.4}$$

The control signal is a linear combination of the error, the time integral of the error, and the time rate of change of error. The derivative part of the controller usually improves the system's dynamic performance, but at the same time increases gain at high frequencies. This means that the high frequency noise effects are accentuated by the derivative term. To eliminate this problem, usually a noise filter is used with the derivative action of the controller and the transfer function of the PID controller becomes

$$G_c(s) = K_p(1 + \frac{1}{T_i s} + \frac{T_d s}{T_f s + 1}) \tag{1.5}$$

where $T_f = \alpha T_d$. Generally, a small value of α is considered in the literature. In this work, α is assumed as 0.1 in all the simulations. When the classical PID controller is implemented in the forward path of the closed loop system, the control action can be written as

$$U(s) = K_p(1 + \frac{1}{T_i s} + T_d s)E(s) \tag{1.6}$$

in Laplace transforms. The derivative part can be written as

$$\frac{de(t)}{dt} = \frac{d[r(t) - y(t)]}{dt} = \frac{dr(t)}{dt} - \frac{dy(t)}{dt} \tag{1.7}$$

If there is a sudden change in the reference input to the controller, the first part of the derivative action in (1.7) yields a sudden abrupt change in the control signal, which is referred to as derivative kick and it may drive the plant to saturation. To avoid this, the reference input may be excluded from the derivative term. In practice, the implementation of the scheme is done as follows; the derivative part of the controller is implemented in an inner feedback loop, which yields the derivative of the output rather than the error, so that

$$u(t) = K_p[e(t) + \frac{1}{T_i}\int e(t)dt] - T_d\frac{dy(t)}{dt} \tag{1.8}$$

The implementation is shown in Fig. 1.3 and the scheme is called proportional integral plus derivative, 'PI-D', controller. A powerful form of controller design is to introduce state feedback to modify the poles of the uncompensated plant and then add a PI controller, particularly where the initial plant model did not include an integrator. In its simplest form for a second order plant this is equivalent to proportional plus derivative feedback and a PI controller. This design approach is called proportional integral proportional derivative, 'PI-PD', controller. In this scheme, a PI controller is implemented in the forward path acting upon the error $e(t)$ and a PD controller acting on the measurement signal $y(t)$, in the inner feedback loop, as shown in Fig. 1.4 where L is the static load disturbance. The setpoint filter $F_{sp}(s)$ is often used to add or cancel the poles or zeros of the closed loop system transfer function.

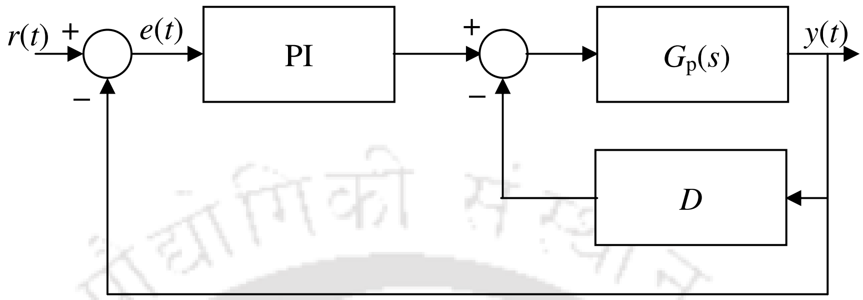


Figure 1.3: Derivative in the inner feedback

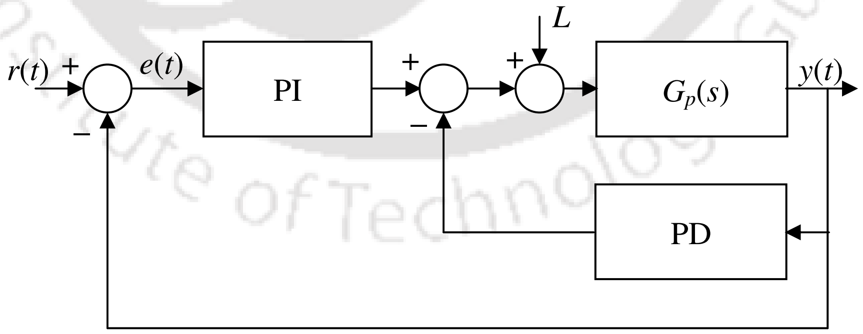


Figure 1.4: PI-PD Controller

1.3 Process Control Requirements

Typical specifications on a control system may include attenuation of load disturbances, robustness to model uncertainty and setpoint following. These are discussed below in brief.

1. *Load Disturbances*: Load disturbances are disturbances that drive the process variables away from their desired values and attenuation of load disturbances is of primary concern for regulatory problems where the processes are running in steady state with constant setpoint for a long time. For analysis, load disturbances are often represented by low frequency signals entering the system at the process input as shown in Fig. 1.4. For a plant in action with the PID controller given by (1.3), if a unit step load disturbance is applied at the process input, the integrated error (IE) is given by [16]

$$IE = \int_0^\infty e(t)dt = \frac{T_i}{K_p} = \frac{1}{k_i} \tag{1.9}$$

The effect of the load disturbance on the output can be minimized by maximizing the integral gain (k_i).

2. *Sensitivity to Process Characteristics*: The controller parameters are typically matched to the process characteristics. Since the process may change it is important that the controller parameters are chosen in such a way that the closed loop system is not too sensitive to variations in process dynamics. There are many ways to specify the sensitivity. Many different criteria are conveniently expressed in terms of the Nyquist curve of the loop transfer function, $L(s) = G_p(s)G_c(s)$. The maximum sensitivity (M_s) is the inverse of the shortest distance from the Nyquist curve to the critical point -1 and is given by

$$M_s = \max_{0\leq\omega<\infty} \left| \frac{1}{1 + G_p(j\omega)G_c(j\omega)} \right| \tag{1.10}$$

Reasonable values of M_s are in the range from 1.3 to 2 [16]. Gain margin (A_m) and phase margin (ϕ_m) are other commonly used sensitivity measures and are given by

$$\begin{aligned} A_m &= \frac{1}{|L(j\omega_u)|} \\ \phi_m &= \pi + \arg L(j\omega_g) \end{aligned} \tag{1.11}$$

where the ultimate frequency ω_u is the frequency where $\arg L(j\omega_u) = -\pi$ and the gain cross over frequency ω_g is the frequency where $|L(j\omega_g)| = 1$. Gain and phase margins are related to M_s by the following equations

$$A_m > \frac{M_s}{M_s - 1}$$
$$\phi_m > 2 \arcsin \frac{1}{2M_s} \tag{1.12}$$

Typical values of ϕ_m range from 30° to 60° whereas gain margin could vary from 2 to 5 [16].

3. *Setpoint Following*: Specifications on setpoint following include requirements on rise time, settling time, decay ratio, overshoot and steady state offset for step changes in setpoint.

- The *rise time* (t_r) is either defined as the inverse of the largest slope of the step response or the time it takes the step to pass from 10% to 90% of its steady state value.
- The *settling time* (t_s) is the time it takes before the step response remains within $p\%$ of its steady state value. The values of p commonly used are 2 and 5.
- The *decay ratio* (d) is the ratio between two consecutive maxima of the error for a step change in setpoint or load. The value $d = 1/4$ which is called quarter amplitude damping has been used traditionally.
- The *overshoot* is the ratio between the difference between the first peak and the steady state value and the steady state value of the step response. In industrial control applications it is common to specify an overshoot of $8 - 10\%$. However, in many situations it is desirable to have an overdamped response with no overshoot.
- The *steady state error* (e_{ss}) is the value of control error e in steady state. The steady state error is always zero if there is an integral action in the controller.

Criteria like integral of absolute error (IAE), IE and ISE are often used to characterize setpoint responses. There will always be a large initial error for step changes in the setpoint and hence, it is useful to have criteria that put little weight on the initial error.

$$\text{Integral of time weighted absolute error (ITAE)} = \int_0^\infty t\,|e(t)|dt$$
$$\text{Integral of time error (ITE)} = \int_0^\infty te(t)dt$$
$$\text{Integral of error squared time (ITSE)} = \int_0^\infty te^2(t)dt$$

Criteria of the type given above are therefore, more suitable to judge performance for set-point following.

1.4 Controller Design

Proportional, integral, and derivative (P, PD, PI, PI-P, PID, PI-PD) controllers are widely used in industry because of their simplicity, robustness, and successful practical applications. They have few tuning parameters and provide good control performance in most situations. The controller parameters can be determined from the identified transfer function models in a number of ways. Zhuang and Atherton [17] have obtained the PID parameters based on the minimization of integral performance criteria in the frequency domain. The use of an all-pole standard form for different performance criteria provides a simple algebraic way for control system design for a system with no zero as shown by Graham and Lathrop [18]. They have obtained standard forms with a zero for ISE, IAE and ITAE criteria, but in this case the zero is constrained so that the zero must be equal to the penultimate coefficient of the denominator so that there is no steady state error to ramp input. Madhuranthakam et al. [19] have proposed tuning rules for a class of stable processes by minimizing IAE performance criterion. It can be observed from [20] that the number of tuning rules for stable overdamped processes are more as compared to integrating processes and hence, recently a number of strategies for obtaining the controller parameters for integrating processes have been reported in the literature [7, 10, 21, 22]. Using the direct synthesis method, Rao et al. [10] have designed a PID controller in series with a lead/lag compensator for control of open loop integrating processes with time delay. Shamsuzzoha and Lee [22] have proposed an analytical design method based on the internal model control (IMC) principle for a PID controller cascaded with a first order lead/lag filter for integrating and first order unstable processes with time delay.

It has been pointed out by Kwak et al. [23] that the conventional PID controller is structurally unsuitable for efficient control of an integrating process. They use an internal feedback loop to convert the integrating process to an open loop stable process. Park et al. [24] have proposed an enhanced PID control strategy including an inner feedback loop for unstable processes. Therefore, it is obvious that one must convert the integrating or unstable process to an open loop stable process to control it effectively using a PID controller. This strategy results in a four parameter controller, PID-P. The advantages of PI-PD controller for controlling integrating and unstable plant transfer functions has also been reported in the literature [25].

In process control, when the time delay is small, a conventional parallel PID controller is commonly used; when the time delay is large, the Smith predictor configuration is common. However, the Smith predictor strategy cannot be used directly for integrating processes with long time delay. Watanabe and Ito [26], Åström et al. [27] and Majhi and Atherton [28] presented a modification of the Smith predictor to control the process with an integrator and a long time delay. Also De Paor [29] and De Paor and Egan [30] proposed an integrated design procedure for a modified Smith predictor and controller for a class of unstable processes with time delay. However, their control strategy is applicable for a limited range of the values of the normalized dead time for which closed loop asymptotic stability can be achieved besides giving a poor setpoint response in terms of overshoot and settling time. Recently, Rao et al. [11] have proposed a simple method of designing the controllers for a modified form of Smith predictor for integrating and double integrating processes with time delay. The setpoint tracking controller is designed using the classical direct synthesis method and the PD controller considered for load disturbance rejection is tuned based on gain and phase margin specifications.

1.5 Motivation of the Present Work

Several tuning methods for PID and its variants for a class of integrating processes have been reported in the literature. However, none of the methods give the settings for integrating plus time delay (IPTD), integrating plus first order plus time delay (IFOPTD) and double integrating plus time delay (DIPTD) process models, respectively. Keeping in view this objective, the thesis proposes explicit expressions for the controller parameters in terms of plant model parameters for the above mentioned class of integrating processes by adjusting the slope of the Nyquist curve at gain crossover frequency and by minimizing IST^{2}E and IST^{3}E integral criterions. Furthermore, as PID and its variants fail to give satisfactory performance for plants with long time delays, it is shown improved closed loop performances can be achieved if the controllers in the modified Smith predictor are tuned properly. In control applications such as manufacture of plastic gloves and PCR analysis, it is desirable to have no overshoot at all. This motivated the researchers to propose control strategies such that the response of the compensated system has overshoot close to a prescribed value. An attempt has been made on similar lines to derive closed form expressions for the controller parameters for stable first/second order plus time delay and IPTD process models.

1.6 Outline of the Thesis

This thesis is divided into six chapters and this section gives a brief outline of the thesis. In Chapter 2, the inability of the classical gradient based search method to find the minimum of an integral performance criterion is first illustrated to justify the use of random search methods for controller tuning. The newly proposed evolutionary algorithm namely, bacterial foraging algorithm is then used to search the optimal parameters of a PID controller by minimizing the modified integral of time error squared (ISTE) criterion and the results are compared with those obtained by particle swarm optimization (PSO) to illustrate the superiority of the proposed method. Also, the widely used ISTE criterion results in step responses with 10% overshoot and hence, time weighted derivative of the control signal is added to the objective function and it is shown that closed loop step responses with desired percentage overshoot can be obtained if the two terms of the objective function are added with a proper weighting factor. Finally, a new objective function is proposed by modifying the ISE criterion to obtain step responses with no overshoot for transfer function models commonly encountered in the process industries.

Chapter 3 presents tuning formulas for PID and its variants for a class of integrating processes. The controller parameters are obtained by minimizing integral performance criteria (ISTE, IST^{2}E, IST^{3}E) by the modified bacterial foraging algorithm discussed in chapter 2. Tuning rules are also proposed by adjusting the slope of the Nyquist curve at ω_g and guidelines are provided for selecting ω_g and the desired slope. Lastly, tuning formulas are given for the PI-PD controller by minimizing the new objective function proposed in chapter 2.

The controller design problem for integrating processes with large time delay is addressed in Chapter 4. The modified Smith predictor proposed in [28] is considered and a proportional controller with unity gain is used for stabilization. The parameters of the PD controller used for load disturbance rejection are obtained to achieve a specified slope of the Nyquist curve at ω_g. The controller for setpoint tracking is designed so that the delay free component of the system output follows the response of a second order plant, assuming a perfect matching between the actual plant and model in both the dynamics and time delay. The provided simple tuning formulae have physically meaningful parameters namely, the damping ratio and the natural frequency. The damping ratio is assumed as unity to obtain a critically damped response with no overshoot and the natural frequency is obtained based on the user specified rise time.

The parameters of the PID controller are obtained to achieve a user specified M_s in Chapter 5 and it is observed that the percentage overshoot remains constant for a particular value of M_s. Hence, analytical expressions correlating the controller parameters to percentage overshoot and plant model parameters are presented for stable first order plus time delay and integrating plus time delay process models, respectively. The approach is also extended to the IMC framework to obtain the normalized IMC filter time constant as a function of percent overshoot for stable first/second order plus time delay models, respectively.

Chapter 6 concludes the research work and gives some general remarks. Also, some ideas about possible future work related to the research is included in this chapter.

Chapter 2

Controller Design based on Bacterial Foraging Algorithm

2.1 Introduction

The proportional-integral-derivative controller and its variants which are the workhorse of the control industry have simple structures and give satisfactory performances for a wide range of industrial plants. The tuning of a PID controller is basically finding the three controller parameters namely, the proportional gain, the integral and the derivative time constants for satisfactory closed loop operation of the plants. The process of tuning can vary from a trial and error attempt to an elaborate optimization problem, based on a model of the plant and a specific criterion of optimal control. Some methods of PID tuning proposed in the literature are the Ziegler-Nichols method, the refined Ziegler-Nichols method, the Cohen-Coon method, the internal model control, the gain and phase margin method [16]. One of the methods for obtaining the optimal PID parameters is by minimizing an integral performance criterion (integral square error, integral absolute error etc.) [17, 19, 31]. Zhuang and Atherton [17] have shown that the integral of time error squared (ISTE) criterion gives good step response results with a small overshoot and short settling time for stable first order plus time delay (FOPTD) plant models. Explicit expressions for the PID parameters in terms of plant model parameters have been proposed in [31] for second order plus time delay (SOPTD) plant models by minimizing integral of time weighted absolute error (ITAE). Madhuranthakam et al. [19] have proposed a set of generalized expressions relating the controller parameters to the process model parameters for FOPTD, SOPTD and second order plus time delay with lead (SOPTDLD) by minimizing IAE performance criterion.

The classical methods used for optimization in [17,19,31] are highly sensitive to starting points and may fail to locate or identify the global minimum in the absence of a good initial guess. Also, the classical methods require the gradient of the objective function and hence, they cannot be used for optimizing multi-dimensional non-linear, non-differentiable problems. These drawbacks of the classical methods are taken care of by evolutionary algorithms. Evolutionary algorithms are stochastic search methods that mimic the natural biological evolution and/or the social behavior of species. These algorithms use the performance index to guide the search in the problem space and can minimize non-differentiable objective functions.

It is assumed in foraging theory that the objective of the animals is to search for and obtain nutrients so that the energy intake per unit time is maximized. This foraging strategy has been formulated as an optimization problem by employing optimal foraging theory. The foraging behavior of *E. coli* bacteria present in our guts has been mimicked to propose an evolutionary algorithm namely bacterial foraging in [32]. The newly proposed evolutionary algorithm has been used successfully to solve real world optimization problems [33–36]. Mishra [33, 34] has proposed a novel strategy for estimating the harmonic components present in power system voltage/current waveforms by employing bacterial foraging algorithm. The simultaneous design of multiple optimal power system stabilizers in two power systems using two bio-inspired algorithms, small-population based particle swarm optimization (SPPSO) and BFA has been reported in [35]. In [36], the authors have proposed an optimal multilevel thresholding algorithm for multimodal image histograms for segmentation of brain magnetic resonance image by employing BFA.

The use of evolutionary algorithms for controller tuning is also reported in the literature. To mention just a few, Biswas et al. [37] have used the differential evolution to design a fractional order PID controller and the optimal design of a PID controller for synchronization of two identical delayed discrete chaotic systems using PSO is reported in [38]. In [39], the evolutionary programming based PID design is presented by minimizing IAE for synchronization of chaotic systems with application in secure communication.

The BFA was used to search the optimal PID parameters by minimizing ISTE criterion in [40] and it was observed that the convergence of BFA depends on the proper selection of the variables associated with the newly proposed algorithm mainly, the run length vector. Hence, in this work the BFA is made adaptive and a strategy to find the suitable run length vector that ensures convergence is proposed. Also, the basic BFA is modified to expedite convergence and the results are compared with those obtained by PSO to illustrate the superiority of the proposed method. The widely used ISTE criterion results in step responses with 10% overshoot [41] and hence, time weighted derivative of the control signal is added to the objective function as suggested in [42]. It is shown that closed loop step responses with desired percentage overshoot can be obtained if the weighting factor is chosen judiciously. A new integral performance criterion is proposed and it is observed that the obtained PID/PI-PD controllers result in smooth step responses with no overshoot for transfer function models commonly encountered in the process industries.

The plant transfer function is assumed to be known *apriori* throughout in this work. The organization of the chapter is as follows. Section 2.2 presents the modified integral criterion and a brief overview of the evolutionary algorithms is given in section 2.3. Adaptive strategies to find the run length vector and the weighting factor are presented in sections 2.4 and 2.5, respectively. The simulation results are discussed in section 2.6. A new objective function is proposed in section 2.7 and conclusions are drawn in section 2.8.

2.2 Integral Performance Criterion

A controller is designed to minimize the error $e(t)$ between the reference and the controlled variable. Hence, a criterion worth characterizing the time response of a system is usually given as an integral function of the error or its weighted products [43]. Graham and Lathrop [18] in 1953 suggested two of the most frequently used criteria: ISE and IAE. As ISE results in oscillatory step responses, time weighted versions of ISE were introduced in [17] and it was shown that ISTE gives good step response results with less overshoot and short settling time. Mathematically, the performance criterion ISTE can be expressed as

$$J_{ISTE} = \int_0^\infty t^2 e(\eta, t)^2 dt \tag{2.1}$$

where η denotes the set of controller parameters. In this work, the integral of $t\dot{u}(t)$ squared is added to the ISTE criterion to penalize excessive excursion of control effort. The modified objective function therefore, becomes

$$J = \int_0^\infty t^2 e(\eta, t)^2 dt + \bar{w} \int_0^\infty t^2 \dot{u}(\eta, t)^2 dt \tag{2.2}$$

where $\bar{w}$ is the weighting factor.

2.2.1 Evaluation of Modified Objective Function in the Frequency Domain

A PID controller with the derivative filter (1.5) in the feedback configuration shown in Fig. 1.2 is considered in the following analysis. Let the transfer function of the process be expressed in the form $G_p(s) = \frac{N(s)}{D(s)}$, where $N(s)$ and $D(s)$ are the numerator and denominator polynomials of the proper $G_p(s)$. A third order Padé approximation is used for any dead time associated with the plant model. The error signal can be expressed in the frequency domain as $E(s) = \frac{B(s)}{A(s)}$, where $A(s)$ and $B(s)$ are polynomials with real coefficients. Thus for $J_{ISE} = \int_0^\infty e^2(\eta, t)dt$, the corresponding s-domain realization is given by $J_{ISE} = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} E(s)E(-s)ds$. Using the Laplace transform, the expression for J_{ISTE} becomes

$$J_{ISTE} = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} F(s)F(-s)ds \tag{2.3}$$

where $F(s)$ is given by

$$F(s) = \frac{dE(s)}{ds} = \frac{B(s)\frac{d}{ds}A(s) - A(s)\frac{d}{ds}B(s)}{\{A(s)\}^2} \tag{2.4}$$

The expression for $sU(s)$ is given by

$$sU(s) = E(s)\frac{K_p}{T_i}\frac{s^2T_i(T_d + T_f) + s(T_i + T_f) + 1}{sT_f + 1} \tag{2.5}$$

The second term of the modified objective function can be evaluated from the derivative of (2.5) which is given by

$$H(s) = \frac{K_p}{T_i}\left(\frac{s^2T_i(T_d + T_f) + s(T_i + T_f) + 1}{sT_f + 1}F(s) - E(s)\frac{s^2T_fT_i(T_f + T_d) + 2sT_i(T_d + T_f) + T_i}{(sT_f + 1)^2}\right) \tag{2.6}$$

In frequency domain, the modified objective function can be written as

$$J = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} \{F(s)F(-s) + \bar{w}H(s)H(-s)\}ds \tag{2.7}$$

The above integral can be optimized for a set of controller parameters (η) values with the help of Åström's recursive algorithm [44] and an optimization algorithm. Furthermore, it is observed that closed loop step responses with desired percentage overshoot can be obtained by a proper choice of the weighting factor, $\bar{w}$. In this work, $\bar{w}$ is selected to obtain transient response with overshoot less than 3%.

2.3 Bio-Inspired Algorithms

The two bio-inspired algorithms namely PSO and BFA are described in brief in this section. The strength of PSO lies in its ability to explore and exploit the search space by varying its parameters (inertia weight and acceleration constants). On the other hand, BFA due to its unique operator (elimination-dispersal event) can find favorable regions during search. These unique features of the algorithms overcome the premature convergence and enhance the search capability.

2.3.1 Particle Swarm Optimization

Particle swarm optimization is a population-based, self adaptive search optimization technique first introduced by Kennedy and Eberhart [45] in 1995. In PSO, a member in the swarm called a particle represents a potential solution which is a point in the search space. The particles fly around in a multi-dimensional search space looking for the global minimum. Each particle keeps track of the previous best position and the corresponding fitness. The previous best value of the particle is called p_{best} and is represented as p_{ip}. Amongst all the particles p_{best} in the swarm, the best value is called g_{best} and is represented as p_{gp}. The basic concept of PSO lies in accelerating each particle towards its p_{ip} and the p_{gp} locations with a random weighted acceleration at each time step. The following steps explain the procedure involved in the classical PSO algorithm.

1. Initialize a population of particles with random positions and velocities in p dimensions of the problem space.

2. For each particle, evaluate the desired fitness function.

3. Compare each particle's fitness evaluation with its p_{best} value p_{ip}. If the current value is better than p_{ip}, set p_{ip} equal to the current location in p-dimensional space.

4. Compare the updated p_{best} values with the population's previous g_{best} value. If any of the p_{best} values are better than p_{gp}, update p_{gp} and its parameters.

5. Compute the new positions of the particles according to

$$x_{ip}(k+1) = x_{ip}(k) + w \times v_{ip}(k) + c_1 \times rand_1 \times (p_{ip}(k) - x_{ip}(k)) + c_2 \times rand_2 \times (p_{gp}(k) - x_{ip}(k)) \tag{2.8}$$

where v_{ip} and x_{ip} represent the velocity and position of ith particle in the pth dimension, respectively. c_1 and c_2 are constants known as acceleration coefficients and $rand_1$, $rand_2$ are two separately generated uniformly distributed random numbers in the range $[0, 1]$.

6. Repeat from step 2 until a specified termination condition is met, usually a sufficiently good fitness or a maximum number of iterations.

The inertia weight w maintains the balance between global and local search. A considerably high diversity is required during the early part of the search to explore the full range of the search space whereas during the latter part of the search, when the algorithm is converging to the optimal solution, fine-tuning of the solutions is important to find the global optima efficiently. Considering these concerns, Shi and Eberhart [46] have proposed a linearly varying inertia weight over the generations. Also, a maximum velocity (v_{maxp}) for each modulus of the velocity vector of the particles (v_{ip}) is defined to control excessive roaming of particles outside the user defined search space. Whenever v_{ip} exceeds the defined limit, it is set equal to v_{maxp}.

2.3.2 Bacterial Foraging Algorithm

The Bacterial Foraging Algorithm mimics the foraging behavior of the *E. coli* bacterium present in our intestines. The foraging process consists of four stages: chemotaxis, swarming, reproduction and elimination-dispersal [32]. Each of these processes are described below in brief.

- Chemotaxis: This stage mimics the bacteria's ability to climb to regions of nutrient concentration, avoiding noxious substances and searching for a way out of neutral media. The bacterium usually takes a tumble, followed by a tumble or a swim to carry out this search. For N_c number of chemotactic steps, the direction of movement after a tumble is given by

$$\beta^i(j+1, k, l) = \beta^i(j, k, l) + C(i) \times \phi(j) \tag{2.9}$$

  where $C(i)$ is the step size taken in the direction of the tumble by the ith bacterium (run length vector). j, k, l are the corresponding indices for the chemotactic step, reproduction step and elimination-dispersal event whereas $\phi(j)$ represents the unit length random direction taken at each step. If the cost at $\beta^i(j+1, k, l)$ is better than the cost at $\beta^i(j, k, l)$, then the bacterium takes another step of size $C(i)$ in that direction (swimming). This process is continued until the number of steps taken is not greater than N_s (counter for the number of swim steps).

- Swarming: The bacteria in times of stress release attractants to signal other bacteria to swarm together. It however also releases a repellant to signal others to be at a minimum distance from it. Thus, all of them have a cell to cell attraction via the attractant and cell to cell repulsion via the repellant. The following equation represents the swarming behavior:

$$\begin{aligned} J_{cc}(\beta, P(j,k,l)) &= \sum_{i=1}^{S} J_{cc}^{i}(\beta, \beta^{i}(j,k,l)) \\ &= \sum_{i=1}^{S}\left[-d_{attract}\exp(-w_{attract}\sum_{m=1}^{p}(\beta_m-\beta_m^i)^2)\right] \\ &\quad+\sum_{i=1}^{S}\left[h_{repellant}\exp(-w_{repellant}\sum_{m=1}^{p}(\beta_m-\beta_m^i)^2)\right] \end{aligned} \tag{2.10}$$

  where $d_{attract}$, $w_{attract}$, $h_{repellant}$ and $w_{repellant}$ are user defined values. p is the number of parameters to be optimized and S is the number of bacteria. Equation (2.10) is to be added to the cost function to get the optimal parameters of the PID controller.

- Reproduction: After all the N_c chemotactic steps have been covered, a reproduction step takes place. $S_r(S_r = S/2)$ bacteria having a lower survival value (less healthy) die, and the remaining S_r's are split into two to maintain a constant population size.

- Elimination-Dispersal: As the environment changes, bacteria are either destroyed or moved to different parts of the intestine resulting in positive/negative influences on their lives. This process is incorporated in the BFA. For an elimination-dispersal event, each bacterium is eliminated with a probability of P_{ed}. A low value of N_{ed} (number of elimination-dispersal events) dictates that the algorithm will not rely on elimination-dispersal events to find favorable regions. On the other hand, a higher value increases computational complexity but allows the bacteria to find favorable regions. The P_{ed} should not be large either, or else, it will lead to an exhaustive search. The number of reproduction and elimination-dispersal events is problem specific and the values used in this work are decided by trial and error.

The pseudo-code of the complete algorithm is given below.

Step 1: Initialization

1. Number of parameters (p) to be optimized

2. Number of bacteria (S) to be used for searching the global minimum

3. Number of chemotactic steps, N_c

4. Swimming length, N_s

5. Number of reproduction steps, N_{re}

6. Number of elimination and dispersal events, N_{ed}

7. Elimination-dispersal probability, P_{ed}

8. Location of each bacterium

Step 2: Iterative Algorithm for Optimization

1. Elimination-dispersal loop: $l = l + 1$

2. Reproduction loop: $k = k + 1$

3. Chemotaxis loop: $j = j + 1$

   (a) For $i = 1, 2....S$, take a chemotactic step for bacterium i as follows.

   (b) Compute fitness function, $J(i, j, k, l) = J(i, j, k, l) + J_{cc}(\beta^i(j, k, l), P(j, k, l))$.

   (c) Let $J_{last} = J(i, j, k, l)$ to save this value since we may find a better cost via a run.

   (d) Tumble: Generate a random vector $\Delta(i) \in \Re^p$ with each element in the range $[-1, 1]$.

   (e) Move: $\beta^i(j + 1, k, l) = \beta^i(j, k, l) + C(i)\frac{\Delta(i)}{\sqrt{\Delta^T(i)\Delta(i)}}$

   (f) Compute $J(i, j + 1, k, l)$ and let

$$J(i, j + 1, k, l) = J(i, j + 1, k, l) + J_{cc}(\beta^i(j + 1, k, l), P(j + 1, k, l)) \tag{2.11}$$

   (g) Swim

      i. $m = 0$

      ii. While $m < N_s$

      - $m = m + 1$
      - If $J(i, j + 1, k, l) < J_{last}$
      - Let $J_{last} = J(i, j + 1, k, l)$

- $\beta^i(j+1,k,l) = \beta^i(j+1,k,l) + C(i)\frac{\Delta(i)}{\sqrt{\Delta^T(i)\Delta(i)}}$
- Use this $\beta^i(j+1,k,l)$ to compute new $J(i,j+1,k,l)$ with cell to cell attraction effect
- Else $m = N_s$

(h) Go to next bacterium $(i+1)$ if $i \neq S$.

4. If $j < N_c$, continue chemotaxis since the life of the bacteria is not over.
5. Reproduction
   - For the given k and l and for each $i = 1, 2...S$ calculate

$$J^i_{health} = \sum_{j=1}^{N_c+1} J(i,j,k,l) \tag{2.12}$$

   - Sort bacteria in order of ascending cost values of J_{health}.
   - Destroy S_r bacteria with the highest values of J_{health}.
   - Split each of the remaining S_r bacteria into two at the same location.
6. If $k < N_{re}$ go to step 2.
7. Elimination and Dispersal: Eliminate and disperse each bacterium with probability P_{ed} keeping population of bacteria constant.
8. If $l < N_{ed}$, go to step 1 else End.

The following two modifications are incorporated in the BFA for better results.

- As all the bacteria are subjected to elimination and dispersal stage in [32]; the bacteria with lower J_{health} values may be lost during this stage. Therefore, in this work after the reproduction stage when two bacteria occupy the same location, the elimination and dispersal event is carried out on only one set of bacteria ensuring that the healthy bacteria are not lost.
- For the final elimination and dispersal step, as the global exploration stage is over and only fine search is needed, the N_s, N_c, $C(i)$ are reduced to $N_s/2$, $N_c/2$ and $C(i)/4$ for better convergence.

2.4 An Adaptive Strategy for Run Length Vector

The run length vector $C(i)$ plays an important role in the convergence of BFA. A small value of $C(i)$ causes slow convergence whereas a large value may fail to locate the minima by swimming through them without stopping. Therefore, the value of $C(i)$ should be chosen properly. In this work, an adaptive scheme to update $C(i)$ is proposed which ensures the convergence of BFA. Initially, with the choice of $\bar{w} = 0$, $N_{re} = 2$, $N_{ed} = 4$, $C(1) = C(2) = C(3) = 0.01$, the optimal J values corresponding to five trial runs of the modified BFA (MBFA) are obtained. If the same J value is not obtained in all the runs, it implies that one or more controller parameters are still in the stage of global exploration and the corresponding run lengths should be increased. To update $C(i)$, the mean of the obtained controller parameters are calculated as K_{pm}, T_{im} and T_{dm} and the run length for the controller parameter with the maximum mean value is multiplied by 10. The procedure is repeated till the same J value is obtained in all the five trials. The following steps summarize the proposed adaptive strategy.

1. Assume $C(1) = C(2) = C(3) = 0.01$.
2. Run the MBFA five times with the settings of $N_{re} = 2$, $N_{ed} = 4$ and $\bar{w} = 0$.
3. Calculate the minimum of the J values obtained in five runs as J_{min}.
4. Calculate $J_{diff} = J - J_{min}$.
5. The mean of the J_{diff} values is obtained as J_{mean}.
6. If $J_{mean} > 0.01$
   (a) Calculate K_{pm}, T_{im}, T_{dm} where K_{pm}, T_{im}, T_{dm} are the mean of the controller parameters obtained in the five runs.
   (b) Find the maximum of K_{pm}, T_{im}, T_{dm} and multiply the corresponding $C(i)$ by 10.
   (c) Go to step 2.

   else
7. Return $C(i)$ and stop.

2.5 Selection of Weighting Factor

In this work, the weighting factor ($\bar{w}$) is selected to get closed loop step responses with overshoot less than 3% . Once the run length vector $C(i)$ is obtained, the MBFA is used to get the optimal controller parameters assuming $\bar{w} = 0.0001$ and the overshoot is calculated for the obtained set of controller parameters. The weighting factor is increased by a factor of 10 if the overshoot is more than 3% to give more weightage to the control signal which reduces the proportional gain thereby, decreasing the percent overshoot. This procedure is repeated till the desired $\bar{w}$ is obtained.

2.6 Simulation Study

In this section, two typical examples are considered to illustrate the usefulness of the MBFA (without swarming). The sensitivity of *fminsearch* command of MATLAB to the initial guess is illustrated for the first example to justify the use of heuristic based methods for PID tuning. Owing to the randomness involved in these approaches, twenty five trials are performed for each optimization problem to observe the variation during the evolutionary processes and to compare their solution quality, convergence characteristic and computational efficiency. The performance of the two evolutionary algorithms is compared using three criteria:

- The average of the solution obtained in all trials, J_{mean}
- The minimum value obtained, J_{min}
- The processing time taken in seconds for 25 trials

The initial controller parameters for both PSO and BFA are taken randomly within $(0 - 5)$ range such that the closed loop system is stable. A brief discussion regarding the selection of various variables for the evolutionary algorithms is presented hereafter.

1. **Bacterial Foraging Algorithm**

   (a) As three parameters of the controller are to be obtained, $p = 3$.

   (b) N_s in a chemotactic loop with N_c iterations should not be very large to prevent the solution being trapped in a local minima. On the other hand, a small value of N_s may lead to unsatisfactory results. Therefore, to have a trade-off between the two, N_s is chosen approximately one fifth of N_c. Choice of $N_c = 50$ gives $N_s = 10$.

(c) $N_{re} = 2$, $N_{ed} = 4$ and $P_{ed} = 0.25$ give satisfactory performance for our problem.

(d) The population size (S) is taken as 10.

The above parameters remain the same for both BFA and the MBFA. The run length vector $C(i) = [0.01, 0.01, 0.01]$ is considered for BFA whereas for the modified BFA, the run length vector is obtained by the method outlined in Section 2.4.

2. **Particle Swarm Optimization**

(a) The velocity of all the particles is initialized randomly within $[-5 \quad 5]$ range.

(b) Clerc et al.'s [47] equation is used for velocity calculation.

(c) The maximum velocity (v_{max}) is specified as 5 in all the three dimensions of the search space.

(d) The number of generations is set equal to 300.

2.6.1 Example 1

The higher order plant considered by Wang et al. [1] is $G_p(s) = \frac{1}{(s+1)(s+5)^2}e^{-0.5s}$. Assuming $\bar{w} = 0$, the *fminsearch* command is used to get the optimal controller parameters. K_p, T_i and T_d are varied from 0.1 to 5 in steps of 0.5 and the first combination such that the closed loop system is stable is taken as the initial guess. The controller parameters obtained with the initial guess $[0.1, 0.1, 1.1]$ are $K_p = 0.15$, $T_i = 0.02$, $T_d = 1.37$ with $J = 11.86$. The *fminsearch* command is used again but with the initial guess $[5.1, 0.1, 2.1]$. The global minimum is obtained this time with $J = 0.27$. Thus, we can see that the classical method is highly sensitive to the starting point being provided by the user.

Fig. 2.1 shows the best values of J and the controller parameters at the end of each chemotaxis loop by BFA and the MBFA. It can be observed that the BFA fails to find the minimum J and the controller parameter K_p is still in the stage of global exploration and hence, its corresponding run length unit is increased. The run length vector and the weighting factor obtained by the method outlined in Sections 2.4 and 2.5 are $\bar{w} = 0.001$ and $C(i) = [0.1, 0.01, 0.01]$, respectively.

For a fair comparison with PSO, the population size (N) is varied to get the best results. The minimum and mean values of J obtained by varying the population size from 10 to 50 are given in Table 2.1. The best results are obtained for $N = 30$ and increasing the population size thereafter does not leads to any improvement in the results. Hence, N is taken as 30. Table 2.2 shows the

J_{min}, J_{mean}, mean values of the controller parameters and the time taken in seconds for 25 trial runs for the BFA, MBFA and PSO. The premature convergence in some of the runs results in different J_{min} and J_{mean} values for PSO and BFA. The mean values of the controller parameters are used to obtain the step responses shown in Fig. 2.2. A step load disturbance of magnitude 5 is applied at time $t = 20$ seconds. The controller parameters obtained by PSO gives an unstable response and hence, the corresponding step response is not shown. The BFA gives a slow response with large settling time and poor load disturbance rejection. Step response with lower overshoot and less settling time as compared to Wang et al.'s PID is obtained by the MBFA.

Table 2.1: J_{min}, J_{mean} for various N for Example 1

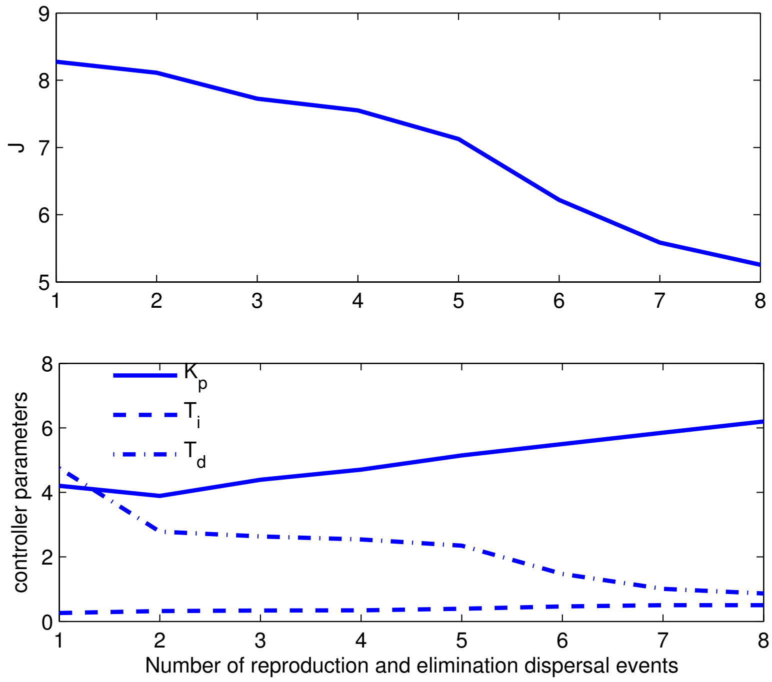
Population size (N)	J_{min}	J_{mean}
10	0.96	1.97
20	0.96	1.16
30	0.96	1.01
40	0.96	1.01
50	0.96	1.02

Table 2.2: J_{min}, J_{mean}, controller parameters and time taken by various evolutionary algorithms for Example 1

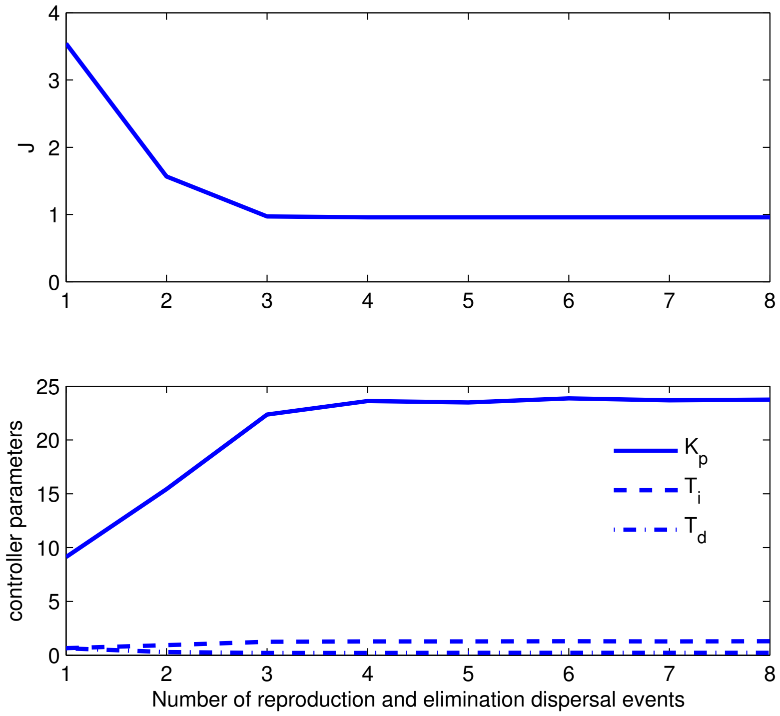
	J_{min}	J_{mean}	Controller parameters			Time (seconds)
			K_{pm}	T_{im}	T_{dm}	
BFA	3.47	4.77	6.71	0.58	0.36	231.92
MBFA	0.96	0.96	23.63	1.29	0.22	204.59
PSO	0.96	1.01	21.01	1.15	101.86	127.28

2.6.2 Example 2

The controller design for a non-minimum phase plant with the transfer function $G_p(s) = \frac{1-5s}{(1+10s)(1+20s)}$ is considered. The performance index (J) and the controller parameters corresponding to the best individual after each generation of PSO are shown in Fig. 2.3. Fig. 2.4 shows that the run length unit for T_i should be increased and hence, the run length vector is taken as $[0.01, 0.10, 0.01]$. The proposed method gives the weighting factor ($\bar{w}$) as 0.1. Table 2.3 shows that the best results for PSO are obtained for a population size of 30. The results obtained by the various evolutionary algorithms are given in Table 2.4. The MBFA and PSO give same J_{min} and J_{mean} values whereas BFA once again gets trapped in the local minima in several of the runs. The servo and the regulatory responses for a unit step change are shown in Fig. 2.5 and the results



(a) BFA



(b) MBFA

Figure 2.1: Best values of J and the corresponding controller parameters at the end of each chemotaxis loop for Example 1 by a) BFA and b) MBFA

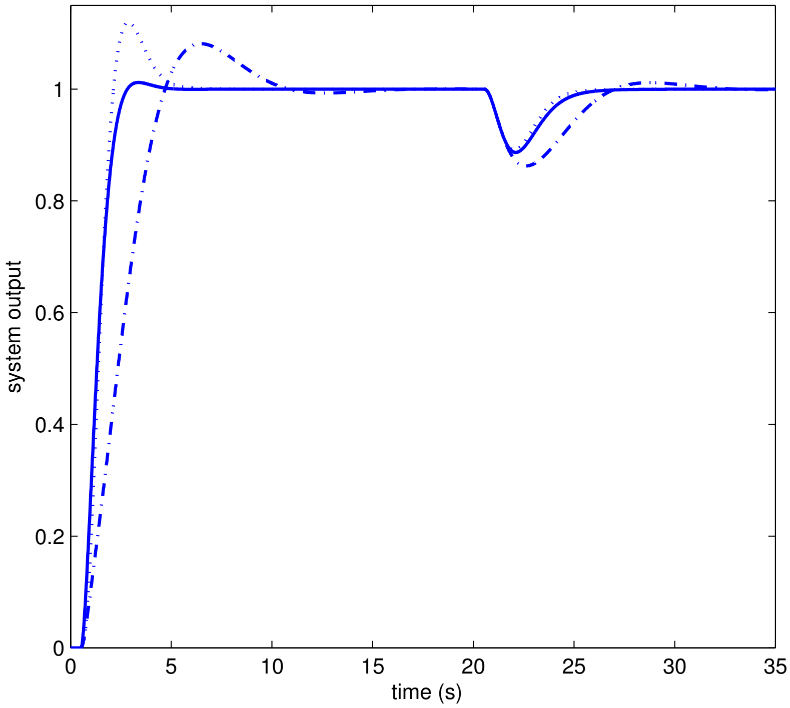


Figure 2.2: Step responses for Example 1: (−) MBFA, (−.−) BFA and (...) Wang et al. [1]

are compared with the method proposed in [2]. Once again, the BFA gives unacceptable closed loop step responses. The MBFA gives a faster response with better load disturbance rejection as compared to the IFT (iterative feedback tuning) controller.

Table 2.3: J_{min}, J_{mean} for various N for Example 2

Population size (N)	J_{min}	J_{mean}
10	184.47	2082.11
20	184.47	1844.95
30	184.47	184.47

2.7 A New Objective Function for Controller Tuning

Tuning methods based on the minimization of ISE guarantee small error and very fast response, which is particularly useful in the case of regulatory control. However, the closed-loop step response is oscillatory with excessive controller output swings. A new objective function is

Table 2.4: J_{min}, J_{mean}, controller parameters and time taken by various evolutionary algorithms for Example 2

	J_{min}	J_{mean}	Controller parameters			Time (seconds)
			K_{pm}	T_{im}	T_{dm}	
BFA	3.24×10^4	3.71×10^4	0.19	6.57	2.51	193.78
MBFA	184.47	184.47	3.16	29.55	6.01	238.30
PSO	184.47	184.47	3.16	29.55	6.01	127.45

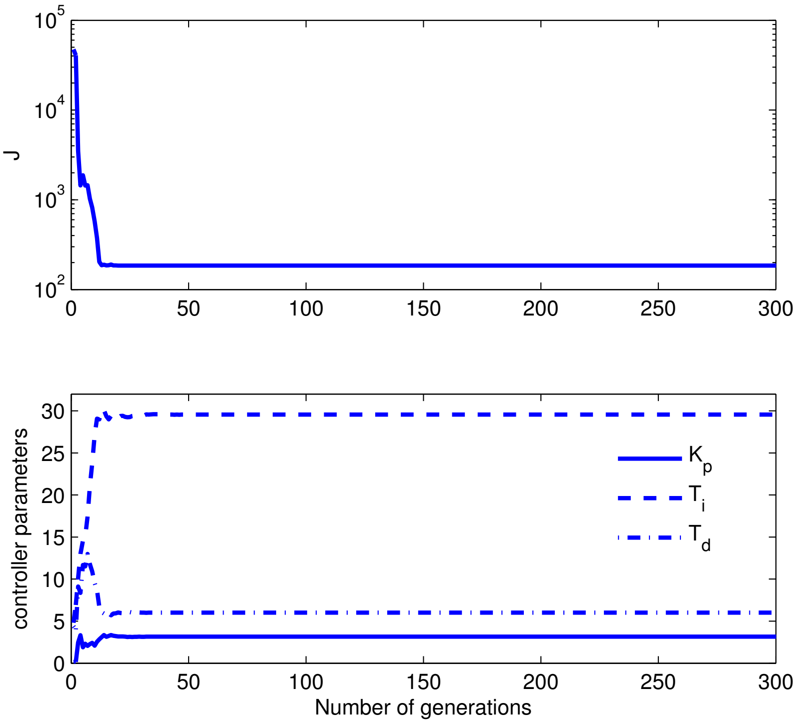
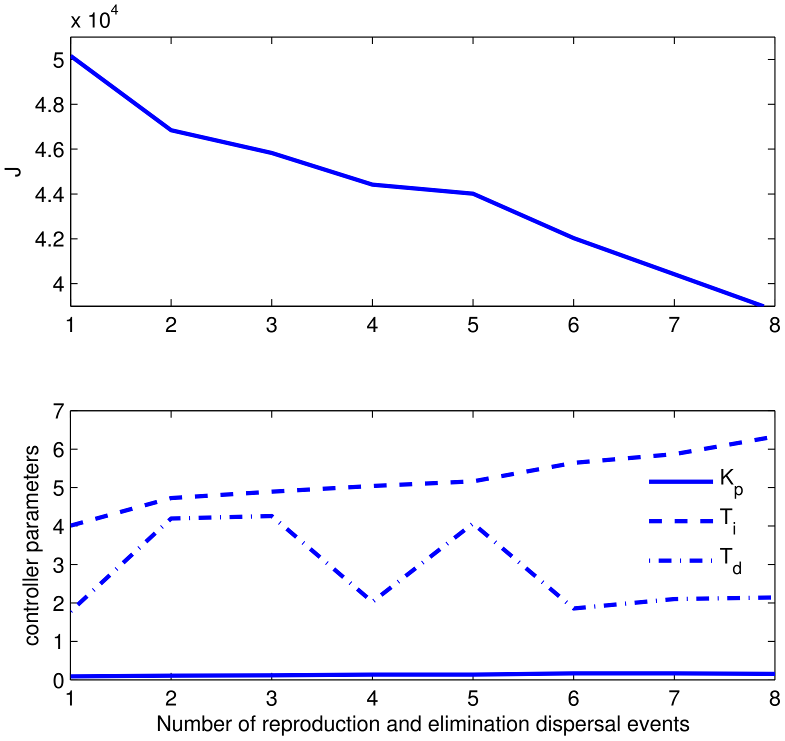
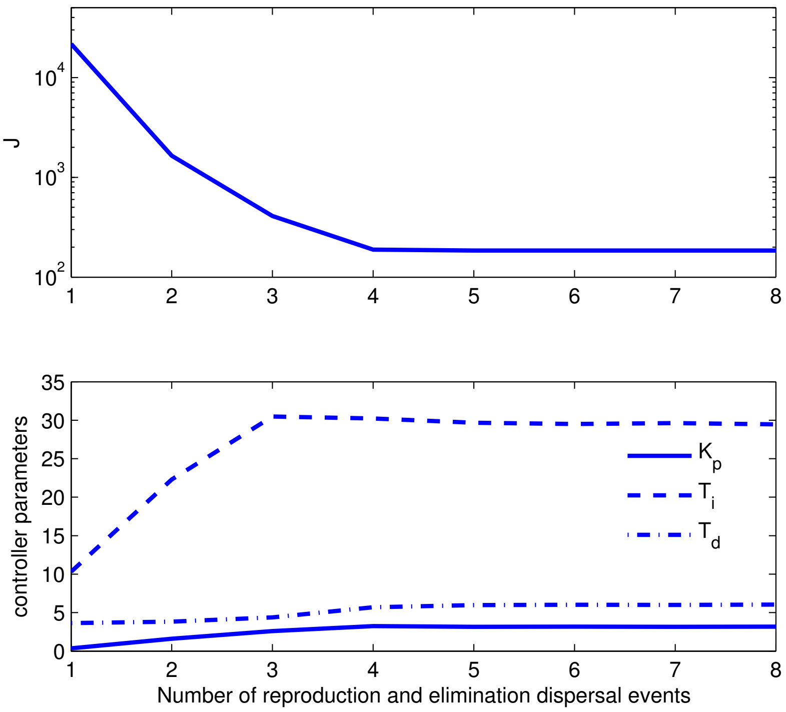


Figure 2.3: Best values of J and the corresponding controller parameters at the end of each generation for Example 2

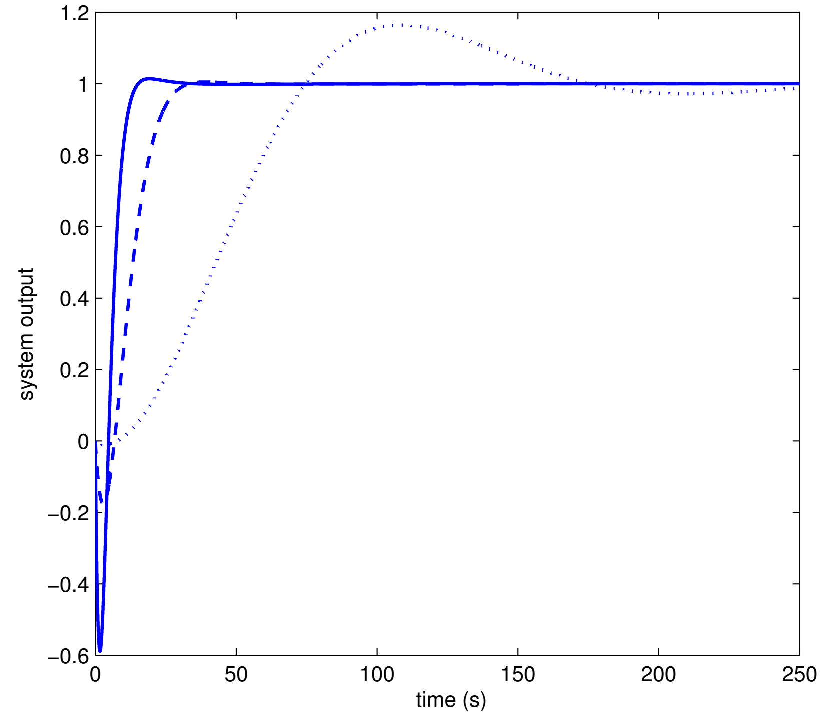


(a) BFA

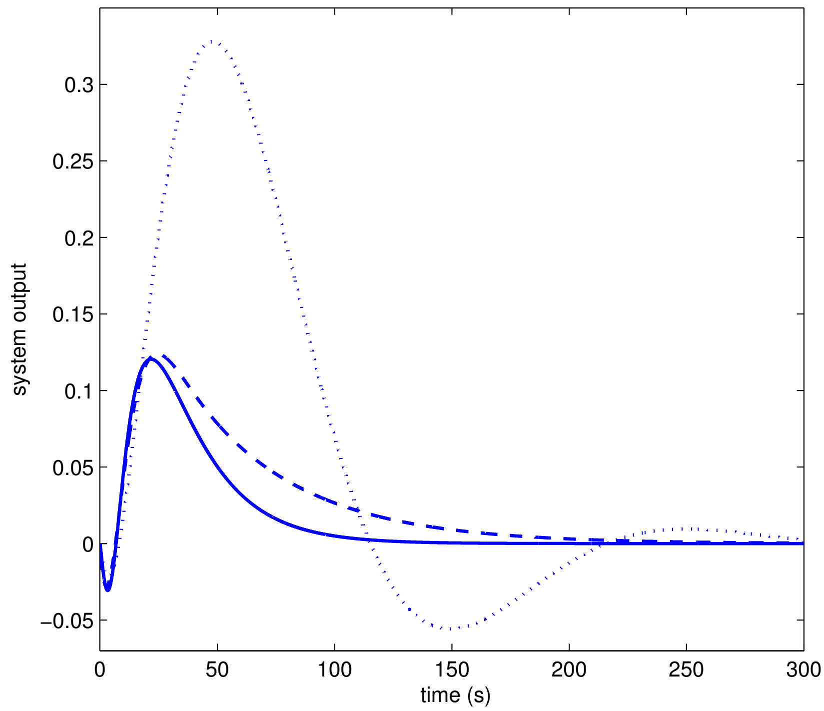


(b) MBFA

Figure 2.4: Best values of J and the corresponding controller parameters at the end of each chemotaxis loop for Example 2 by a) BFA and b) MBFA



(a) servo control performances



(b) regulatory control performances

Figure 2.5: Step responses for Example 2: (−) MBFA , (...) BFA and (−−) IFT [2]

therefore proposed by modifying the ISE criterion and is given by

$$J_{new} = \int_0^\infty (e(\eta,t)^2 + t^4 \dot{u}(\eta,t)^2)dt \tag{2.13}$$

The results are compared with those reported in [3] where the authors have tuned setpoint weighted PID controller by a numerical method based on optimization of load disturbance rejection with constraints on M_s and / or complementary sensitivity to obtain smooth step responses. The PI and PD controllers in the forward and inner feedback paths in the PI-PD configuration are given by:

$$G_{c1}(s) = \frac{U(s)}{E(s)} = K_p(1 + \frac{1}{sT_i}) \tag{2.14}$$

$$G_{c2}(s) = K_{pb}(1 + \frac{sT_{db}}{0.1sT_{db} + 1}) \tag{2.15}$$

2.7.1 Simulation Study

$$\begin{aligned}
G_1(s) &= \frac{1}{s(s+1)^3} \\
G_2(s) &= \frac{e^{-5s}}{(s+1)^3} \\
G_3(s) &= \frac{1}{(s+1)(1+0.2s)(1+0.04s)(1+0.008s)} \\
G_n(s) &= \frac{1}{(s+1)^n} \qquad \forall \; n = 4,5,6,7 \\
G_8(s) &= \frac{1-2s}{(s+1)^3}
\end{aligned}$$

The above transfer functions considered in [3] capture the dynamics of the commonly encountered processes in the control industry and are therefore, used to illustrate the effectiveness of the new objective function. $G_1(s)$ represents an integrating plant model whereas $G_2(s)$ models a process with a long dead time. A non-minimum phase plant ($G_8(s)$) is also included to demonstrate the wide applicability of the proposed method. The parameters of the PID and PI-PD controller are obtained by minimizing (2.13) and are given in Tables 2.5 and 2.6, respectively. The system outputs for a unit step change in the setpoint and the load disturbance at the plant input are shown in Figs. 2.6, 2.7, 2.8 and 2.9, respectively. It can be observed that even though the considered plant transfer functions represent processes with large variations in plant dynamics, the resulting setpoint responses are very much similar with no overshoot as desired in the control industry [3, 48].

The proposed objective function results in a PD controller for the integrating plant model ($G_1(s)$) which fails to reject the load disturbance and hence a PI-PD configuration is recommended for the integrating plant models. Even though the PID and PI-PD controllers give almost similar setpoint responses for $G_3(s)$, the PI-PD gives improved load disturbance rejection performances. Table 2.5 shows that the value of the maximum sensitivity for the considered plant models lies between 1.3 to 1.8 except $G_3(s)$ for which it is 1.2. Also, it can be observed that the proposed PID/PI-PD performs better than the two degree of freedom PID controller proposed in [3] for all the examples. One point to be observed is that similar results can be obtained by the modified objective function proposed in section 2.2. However, the iterative procedure involved in the selection of $\bar{w}$ limits its use.

Table 2.5: PID parameters for various process models

	K_p	T_i	T_d	M_s
$G_1(s)$	0.24	–	1.86	1.3
$G_2(s)$	0.46	3.97	1.27	1.7
$G_3(s)$	5.19	0.88	0.15	1.2
$G_4(s)$	0.86	2.78	0.81	1.3
$G_5(s)$	0.72	3.21	0.94	1.4
$G_6(s)$	0.64	3.65	1.09	1.4
$G_7(s)$	0.60	4.10	1.25	1.5
$G_8(s)$	0.52	2.47	0.68	1.8

Table 2.6: PI-PD parameters for various process models

	K_p	T_i	K_{pb}	T_{db}	M_s
$G_1(s)$	0.18	3.56	0.30	2.37	1.4
$G_2(s)$	0.33	3.42	0.03	11.77	1.7
$G_3(s)$	6.49	0.19	9.57	0.18	1.4
$G_4(s)$	0.62	2.50	0.14	4.56	1.4
$G_5(s)$	0.51	2.88	0.08	6.72	1.4
$G_6(s)$	0.46	3.24	0.06	8.53	1.4
$G_7(s)$	0.43	3.61	0.05	10.23	1.5
$G_8(s)$	0.37	2.09	0.06	4.20	1.8

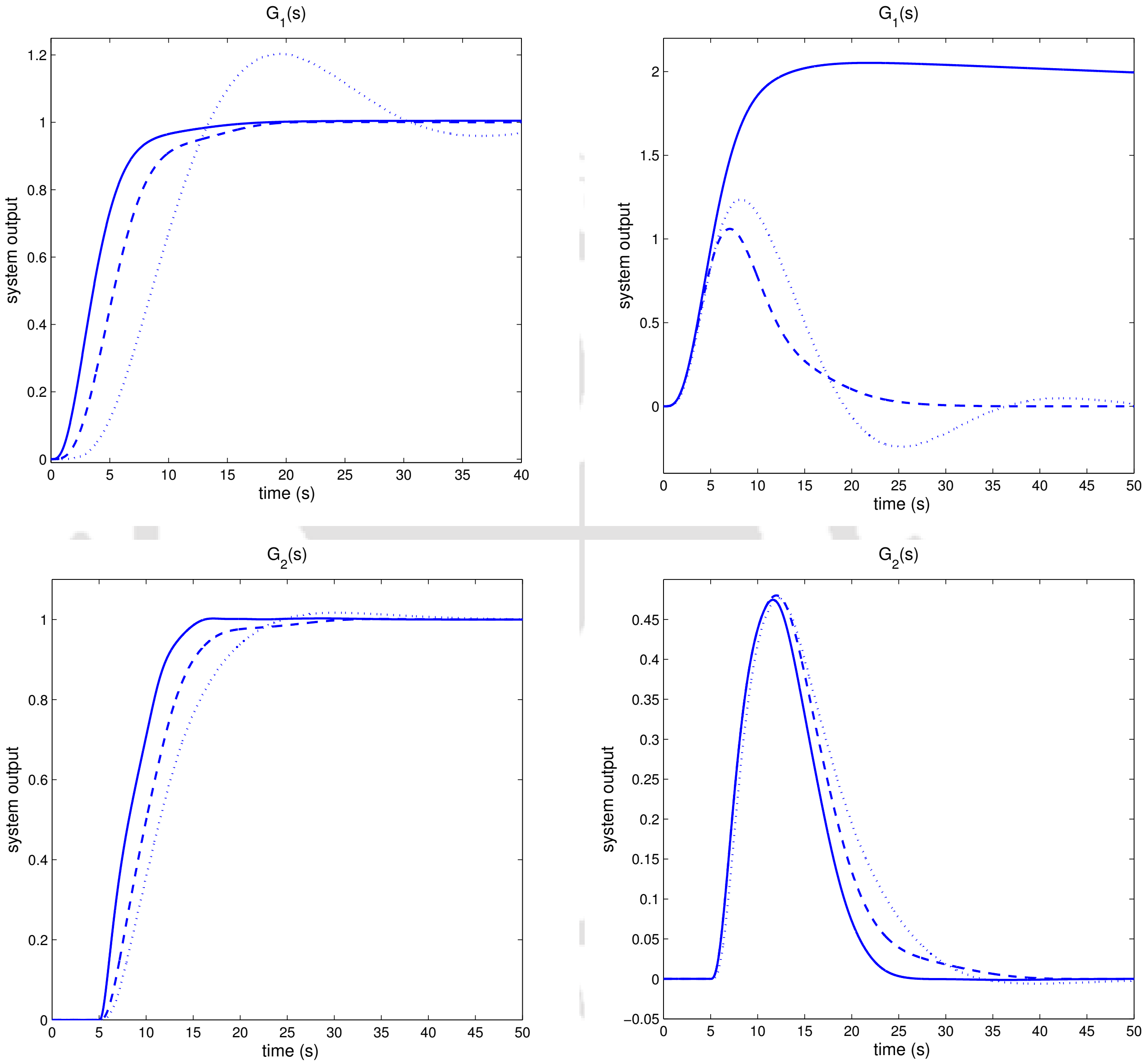


Figure 2.6: Step responses for $G_1(s)$ and $G_2(s)$: (—) Proposed PID, (−−) Proposed PI-PD and (...) Panagopoulos et al. [3]

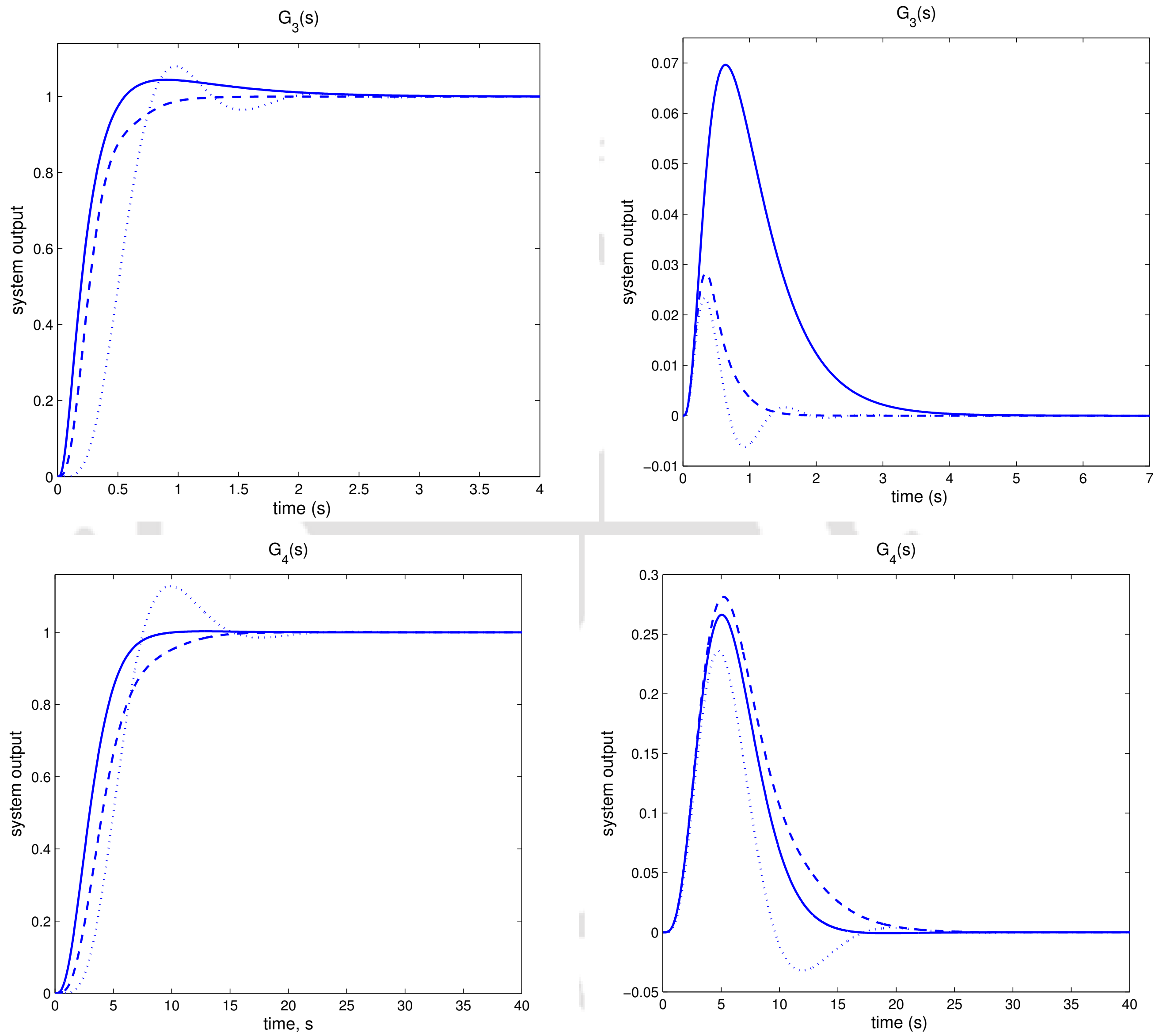


Figure 2.7: Step responses for $G_3(s)$ and $G_4(s)$: (—) Proposed PID, (– –) Proposed PI-PD and (...) Panagopoulos et al. [3]

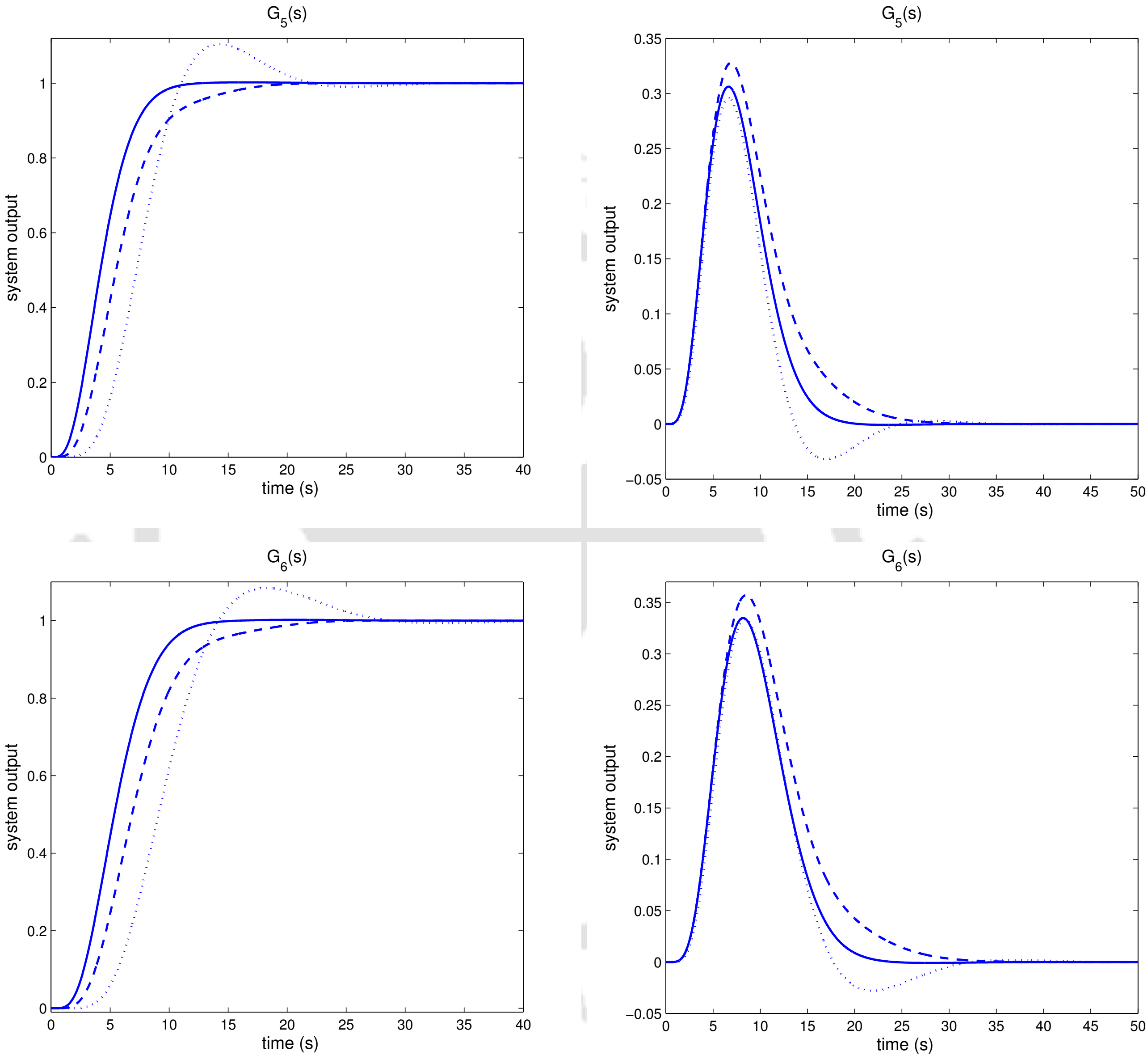


Figure 2.8: Step responses for $G_5(s)$ and $G_6(s)$: (–) Proposed PID, (– –) Proposed PI-PD and (...) Panagopoulos et al. [3]

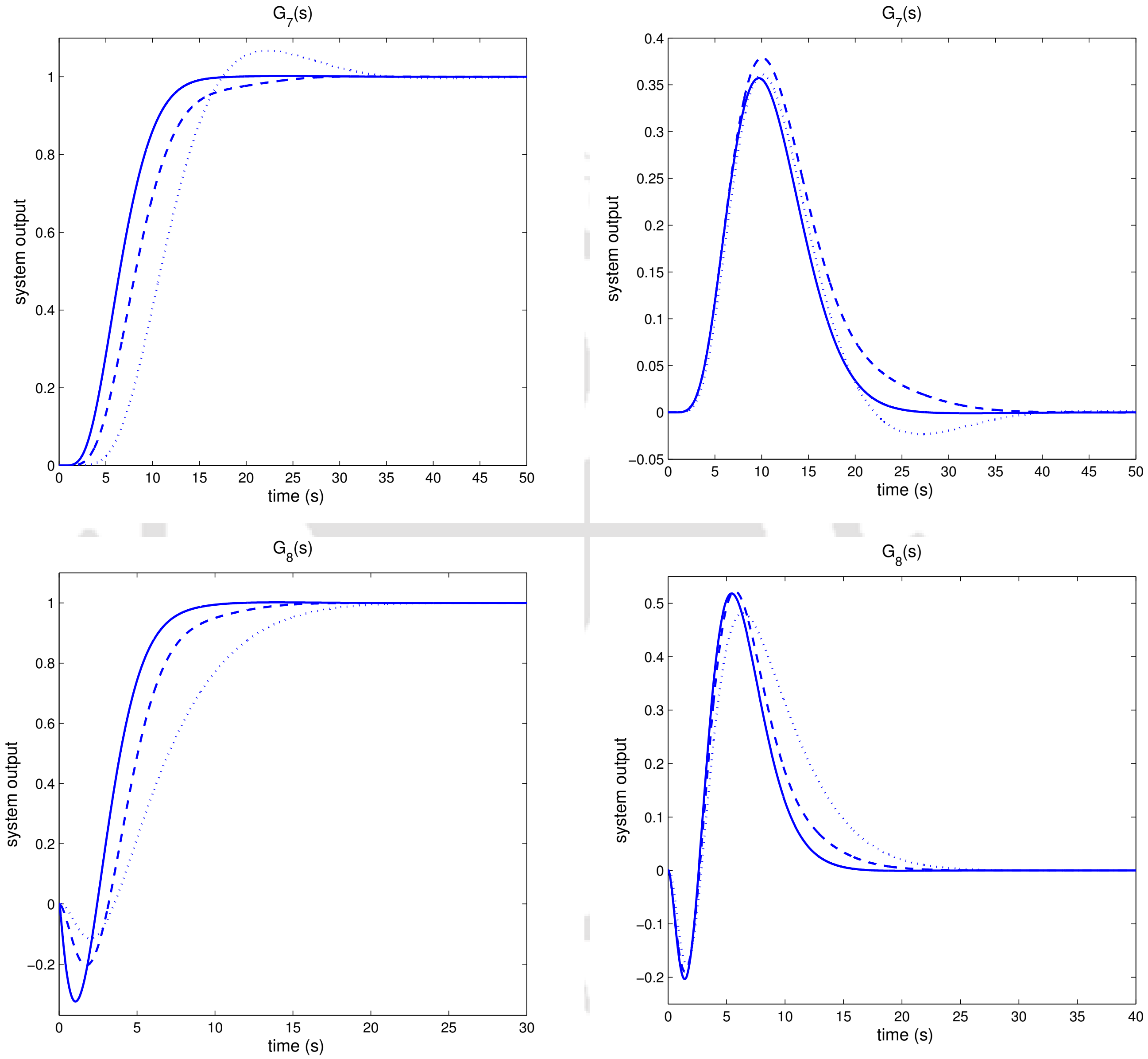


Figure 2.9: Step responses for $G_7(s)$ and $G_8(s)$: (–) Proposed PID, (– –) Proposed PI-PD and (...) Panagopoulos et al. [3]

2.8 Conclusions

A novel method for determining the optimal PID parameters using the modified bacterial foraging algorithm is presented and the results are compared with that obtained by the well established evolutionary algorithm namely, PSO to illustrate the superiority of the proposed method. Simulation results show the better convergence of the MBFA. This is because the randomness involved in the MBFA is much more because of the elimination-dispersal event which reduces the probability of getting trapped in the local minima. Also, the ISTE criterion is modified by adding the integral of the time $\dot{u}(t)$ squared to obtain closed loop step responses with desired percent overshoot. As the run length vector plays an important role in the convergence of BFA, a strategy to obtain the run length vector that ensures convergence is proposed. Two modifications in the original BFA are also proposed for better results. Finally, a new objective function is proposed by modifying the ISE criterion and it is observed that the obtained PID/PI-PD controller gives smooth step responses with almost zero percent overshoot for transfer function models commonly encountered in the process industry.

Chapter 3

Controller Tuning for Integrating Processes

3.1 Introduction

A number of controller design methods for integrating processes have been reported in the literature [4, 8, 13, 21, 49–51]. Tyreus and Luyben [49] have proposed tuning rules based on the classical frequency response method to achieve maximum closed loop log modulus of 2 dB. Kookos et al. [4] have obtained the parameters of the PI controller so as to achieve user defined gain and phase margins (KPGM) and by minimizing weighted ISE (KWISE). Tuning rules for PI controller based on the maximum peak resonance specification is reported in [8].

The method proposed in [49] is extended for designing PID controllers in [50] and the desired control signal trajectory was used as a performance specification to design a PID controller in [51]. Chidambaram and Sree [21] have obtained the parameters of a PI, PID and PD controller by matching the coefficients of corresponding powers of s in the numerator and that in the denominator of the closed loop transfer function for a servo problem. The method proposed in [21] is extended using two tuning parameters to obtain robust closed loop performances in [7]. Skogestad [5] has proposed SIMC (Skogestad's internal model control) tuning rules for a class of processes in the IMC (internal model control) framework by choosing the process model's equivalent time delay as the IMC filter parameter. However, none of the above reported works give settings for both PI and PID controllers for integrating plus time delay (IPTD), integrating plus first order plus time delay (IFOPTD) and double integrating plus time delay (DIPTD) process models, respectively.

Recently, Karimi et al. [52] have used the Bode's integrals to approximate the derivatives of amplitude and phase of a system with respect to frequency at a given frequency and these derivatives are then used to adjust the slope of the Nyquist curve at the gain crossover frequency to obtain the parameters of a PID controller using the modified Ziegler-Nichols method. However, they have not provided an adequate method for selecting the gain crossover frequency and the slope of the Nyquist curve. In this work, Karimi et al.'s approach is extended to design PI and PID controllers for integrating processes in section 3.3.

The ISTE criterion is widely used for controller tuning as it results in step responses with a relatively small overshoot and a reasonable settling time [17, 41, 53–55]. However, only a few works have been reported where the commonly used ISTE criterion is used for controller tuning for integrating processes [6, 13]. Tuning rules for PI and PID controllers are therefore proposed for a class of integrating processes by minimizing ISTE, IST^{2}E and IST^{3}E performance indices in section 3.4. Furthermore, it is shown that even though the commonly used ISTE criterion fails to give satisfactory closed loop performances, good nominal and robust control performances are obtained by IST^{2}E and IST^{3}E criteria. The MBFA proposed in the previous chapter is used for minimizing the three integral criteria for step changes in the setpoint. Simulation results are presented in section 3.5. The new objective function proposed in the previous chapter is used to obtain tuning formulas for IPTD and IFOPTD plant models in section 3.6.

3.2 Integrating Process Models

The dynamics of the commonly encountered integrating processes can be described by the following transfer function models:

1. Integrating plus time delay (IPTD): $G_p(s) = \frac{Ke^{-\theta s}}{s}$

2. Integrating plus first order plus time delay (IFOPTD): $G_p(s) = \frac{Ke^{-\theta s}}{s(Ts+1)}$

3. Double integrating plus time delay (DIPTD): $G_p(s) = \frac{Ke^{-\theta s}}{s^2}$

3.3 Loop Slope Adjustment

The controller parameters are obtained by adjusting the slope of the Nyquist curve at gain crossover frequency in this section. Guidelines are provided for selecting ω_g and the desired slope for satisfactory closed loop operation of the considered integrating plant models.

3.3.1 Selection of Gain Crossover Frequency

It is reported in [52] that the slope of the Nyquist curve at ω_g drastically affects the performance and robustness of the closed loop system and hence, an upper bound on ω_g is first obtained in this subsection. To figure out the best achievable control performance for a process, the process model is factorized into

$$G_p(s) = G_{mp}(s)G_{nmp}(s) \tag{3.1}$$

where $G_{mp}(s)$ and $G_{nmp}(s)$ are the minimum-phase and non-minimum phase parts with $|G_{nmp}(j\omega)| = 1$. Åström [56] has shown that the gain crossover frequency satisfies

$$\arg G_{nmp}(j\omega_g) \geq -\pi + \phi_m - n_g\pi/2 \tag{3.2}$$

where ϕ_m is the required phase margin and n_g is the slope of the open loop gain at the crossover frequency. The open loop gain should have a slope of about -1 around the crossover frequency, with preferably steeper slopes before and after the crossover and hence n_g is assumed as -1 [57]. For the class of integrating plant models given in section 3.2, (3.2) reduces to

$$\theta\omega_g \leq \pi/2 - \phi_m \tag{3.3}$$

Typical values of ϕ_m range from 30° to 60° [16]. In this work, three values of ϕ_m, i.e. $\pi/6$, $\pi/4$ and $\pi/3$ are considered and the corresponding values of $\theta\omega_g$ (considering the equality sign in (3.3)) are $1.05, 0.78$ and 0.52, respectively.

3.3.2 PID Controller

A relationship between T_i and T_d is obtained to achieve a user specified slope at any given frequency in this subsection. The slope of the Nyquist curve of the loop transfer function $L(s) = G_p(s)G_c(s)$ at any frequency ω_o is equal to the phase of the derivative of $L(j\omega)$ at ω_o. Differentiating the loop transfer function, we get

$$\frac{dL(j\omega)}{d\omega} = G_p(j\omega)\frac{dG_c(j\omega)}{d\omega} + G_c(j\omega)\frac{dG_p(j\omega)}{d\omega} \tag{3.4}$$

Also, we have

$$\ln G_p(j\omega) = \ln|G_p(j\omega)| + j\angle G_p(j\omega) \tag{3.5}$$

Differentiating the above equation, we get

$$\frac{dG_p(j\omega)}{d\omega} = G_p(j\omega)\left\{\frac{d\ln|G_p(j\omega)|}{d\omega} + j\frac{d\angle G_p(j\omega)}{d\omega}\right\} \tag{3.6}$$

The derivative of the PID controller transfer function (neglecting T_f) with respect to ω gives

$$\frac{dG_c(j\omega)}{d\omega} = jK_p(T_d + \frac{1}{\omega^2 T_i}) \tag{3.7}$$

Substituting $G_c(j\omega)$, (3.6) and (3.7) in (3.4), we get

$$\frac{dL(j\omega)}{d\omega} = G_p(j\omega)K_p\left\{j(T_d + \frac{1}{\omega^2 T_i}) + \left(1 + j(T_d\omega - \frac{1}{\omega T_i})\right) \times \left(\frac{d\ln|G_p(j\omega)|}{d\omega} + j\frac{d\angle G_p(jw)}{d\omega}\right)\right\} \tag{3.8}$$

The slope of the Nyquist curve at ω_o is given by

$$\psi = \varphi_o + \arctan\left(\frac{1 + T_iT_d\omega_o^2 + s_p(\omega_o)\omega_oT_i + (T_iT_d\omega_o^2 - 1)s_a(\omega_o)}{s_a(\omega_o)\omega_oT_i - (T_iT_d\omega_o^2 - 1)s_p(\omega_o)}\right) \tag{3.9}$$

where

$$\begin{aligned} \varphi_o &= \angle G_p(j\omega_o) \\ s_a(\omega_o) &= \omega_o \left.\frac{d\ln|G_p(j\omega)|}{d\omega}\right|_{\omega_o} \\ s_p(\omega_o) &= \omega_o \left.\frac{d\angle G_p(j\omega)}{d\omega}\right|_{\omega_o} \end{aligned} \tag{3.10}$$

Rearranging (3.9), we get

$$T_d = \frac{s_a(\omega_o) - 1 + s_p(\omega_o)\tan(\psi - \varphi_o) - T_i\omega_o(s_p(\omega_o) - s_a(\omega_o)\tan(\psi - \varphi_o))}{\omega_o^2 T_i(1 + s_a(\omega_o) + s_p(\omega_o)\tan(\psi - \varphi_o))} \tag{3.11}$$

Using the condition that $|L(j\omega_g)| = 1$, we obtain

$$\begin{aligned} KK_p &= \frac{\omega_g^2 T_i}{\sqrt{(1 - \omega_g^2 T_i T_d)^2 + \omega_g^2 T_i^2}} && \text{IPTD} \\ KK_p &= \frac{\omega_g^2 T_i\sqrt{1 + \omega_g^2 T^2}}{\sqrt{(1 - \omega_g^2 T_i T_d)^2 + \omega_g^2 T_i^2}} && \text{IFOPTD} \\ KK_p &= \frac{\omega_g^3 T_i}{\sqrt{(1 - \omega_g^2 T_i T_d)^2 + \omega_g^2 T_i^2}} && \text{DIPTD} \end{aligned} \tag{3.12}$$

The proposed method of obtaining the controller parameters can be therefore summarized as: Given any integrating plant model find T_i such that ISE is minimum subject to (3.11) and (3.12), where $\omega_o = \omega_g$ and ω_g, ψ are to be specified by the user. Guidelines for choosing ω_g and ψ are outlined below.

3.3.3 IPTD Process Model

The values of ω_g and ψ are selected based on load disturbance rejection perfromances. To illustrate the procedure adopted for selecting ω_g and ψ, an IPTD process model is considered with both K and θ assumed as unity. The settling time and the maximum value of the system output (y_{max}) for various ω_g and ψ for a unit step change in the load disturbance at the system input are given in Table 3.1. Regarding selection of ω_g, it can be observed that the minimum value of y_{max} is obtained for $\omega_g = 1.05$. Also $\psi = 60°$ results in least settling time and hence, ω_g and ψ are taken as $1.05/\theta$ and $60°$, respectively. The controller parameters obtained for various θ are given in Table 3.2 and the recommended PID settings are:

- $K_p = 1.03/(\theta K)$
- $T_i = 3.17\theta$
- $T_d = 0.49\theta$

Table 3.1: Performance summary for load disturbance rejection

ω_g	$\psi = 55°$		$\psi = 60°$		$\psi = 65°$	
	y_{max}	$t_s(s)$	y_{max}	$t_s(s)$	y_{max}	$t_s(s)$
0.52	1.49	27.04	1.54	20.81	1.56	25.50
0.78	1.19	16.27	1.19	12.12	1.20	14.60
1.05	1.13	10.58	1.14	8.04	1.15	9.42

Table 3.2: Controller parameters for IPTD process models

θ	KK_p	T_i	T_d
0.5	2.05	1.59	0.24
1	1.03	3.17	0.49
2	0.51	6.35	0.98
3	0.34	9.52	1.47
4	0.26	12.70	1.96

3.3.4 IFOPTD Process Model

For IFOPTD plant models, ω_g is assumed as $0.78/\theta$ and the values of ψ that give satisfactory closed loop performance and the corresponding controller parameters are given in Table 3.3. Analytical expressions correlating the controller parameters to plant model parameters obtained using the curve fitting toolbox are:

- $KK_pT = 0.78(\frac{\theta}{T})^{-1.53} + 0.10$
- $\frac{T_i}{T} = 5.44(\frac{\theta}{T})^{1.02}$
- $\frac{T_d}{T} = 0.96(\frac{\theta}{T})^{0.77}$

Table 3.3: Controller parameters for IFOPTD process models

θ/T	KK_pT	T_i/T	T_d/T	ψ
0.5	2.37	3.01	0.58	45°
1	0.88	5.13	0.98	60°
2	0.39	11.21	1.60	70°
3	0.25	16.57	2.22	70°
4	0.19	22.50	2.80	70°

3.3.5 DIPTD Process Model

Table 3.4: Controller parameters for DIPTD process models

θ	KK_p	T_i	T_d
0.5	0.67	4.25	1.43
1	0.17	8.51	2.87
2	0.04	17.02	5.73
3	0.02	25.53	8.60
4	0.01	34.05	11.46

The values of ω_g and ψ are assumed as $0.52/\theta$ and $15°$, and the recommended controller settings are:

- $K_p = 0.17/(K\theta^2)$
- $T_i = 8.51\theta$
- $T_d = 2.87\theta$

3.3.6 PI Controller

Substituting $T_d = 0$ in (3.11) and rearranging, we get the following equation for the integral time constant of a PI controller

$$T_i = \frac{s_a(\omega_o) - 1 + s_p(\omega_o)\tan(\psi - \varphi_o)}{\omega_o(s_p(\omega_o) - s_a(\omega_o)\tan(\psi - \varphi_o))} \tag{3.13}$$

Once T_i is obtained, K_p can be calculated using the following equations

$$KK_p = \frac{\omega_g^2 T_i}{\sqrt{1+\omega_g^2 T_i^2}} \qquad \text{IPTD}$$

$$KK_p = \frac{\omega_g^2 T_i \sqrt{1+\omega_g^2 T^2}}{\sqrt{1+\omega_g^2 T_i^2}} \qquad \text{IFOPTD}$$

Based on extensive simulation studies, the following parameters are recommended for IPTD and IFOPTD plant models, respectively.

1. **IPTD Process Model**
   - $\omega_g = 0.52/\theta$
   - $\psi = 50°$
   - $KK_p = 0.48/\theta$
   - $T_i = 5.24\theta$
2. **IFOPTD Process Model**
   - $\omega_g = 0.35/\theta$
   - $\psi = 50°$
   - $TKK_p = 0.37(\theta/T)^{-1.18}$
   - $T_i/T = 3.85(\theta/T)^{1.35} + 5.85$

3.4 Integral Performance Criteria

In this section, the controller parameters are obtained by minimizing the integral performance criteria given by

$$J_n = \int_0^\infty [t^n e(\eta, t)]^2 dt \qquad n = 1, 2, 3 \tag{3.14}$$

It is to be noted that J_1 is denoted as the ISTE criterion, whilst J_2 and J_3 are known as the IST^{2}E and IST^{3}E criteria respectively. For each process model (assuming the steady state gain and the time constant as unity), the plant delay is varied and for each value of the considered plant delay, the MBFA is run 30 times to obtain the optimal controller parameters which are then fitted into simple algebraic equations.

3.4.1 IPTD Process Model

The tuning rules for PI/PID controllers for IPTD process models are given in Tables 3.5 and 3.6, respectively. It can be observed from Table 3.5 that the controller parameters decrease with increase in n in the objective function. Also, the IST^{2}E and IST^{3}E criterion give almost similar results for PID controller and hence the tuning rules are not given for IST^{3}E criterion in Table 3.6.

Table 3.5: Tuning rules for PI controller for IPTD process model

		ISTE	IST^{2}E	IST^{3}E
$KK_p = a_1/\theta$	$a_1 =$	0.79	0.73	0.66
$T_i = a_2\theta$	$a_2 =$	8.93	4.71	4.18

Table 3.6: Tuning rules for PID controller for IPTD process model

		ISTE	IST^{2}E
$KK_p = a_1/\theta$	$a_1 =$	1.03	1.10
$T_i = a_2\theta$	$a_2 =$	4.42	2.72
$T_d = a_3\theta$	$a_3 =$	0.41	0.38

3.4.2 IFOPTD Process Model

For IFOPTD process model, the tuning rules depend on the normalized time constant (θ/T) in addition to the time constant (T) and are given in Tables 3.7 and 3.8, respectively.

Table 3.7: Tuning rules for PI controller for IFOPTD process model

		ISTE	IST^{2}E	IST^{3}E
$TKK_p = a_1(\theta/T)^{b_1}$	$a_1 =$	0.46	0.40	0.34
	$b_1 =$	-0.64	-0.62	-0.62
$T_i/T = a_2(\theta/T)^{b_2} + c_2$	$a_2 =$	5.20	3.93	3.92
	$b_2 =$	1.28	1.09	1.03
	$c_2 =$	15.15	6.10	4.84

3.4.3 DIPTD Process Model

A PI controller fails to stabilize DIPTD processes and the PID tuning rules are given in Table 3.9. It is observed that the ISTE criterion results in a PD controller.

Table 3.8: Tuning rules for PID controller for IFOPTD process model

		ISTE	IST^{2}E	IST^{3}E
$KK_pT = a_1(\theta/T)^{b_1}$	$a_1 =$	0.80	0.86	0.85
	$b_1 =$	-0.98	-1.00	-0.99
$T_i/T = a_2(\theta/T)^{b_2} + c_2$	$a_2 =$	5.26	3.42	3.38
	$b_2 =$	0.94	0.91	0.90
	$c_2 =$	2.09	1.15	1.13
$T_d/T = a_3(\theta/T)^{b_3} + c_3$	$a_3 =$	0.43	0.42	0.36
	$b_3 =$	0.94	0.91	0.94
	$c_3 =$	0.62	0.53	0.51

Table 3.9: Tuning rules for PID controller for DIPTD process model

		ISTE	IST^{2}E	IST^{3}E
$KK_p = a_1\theta^{b_1}$	$a_1 =$	0.08	0.12	0.13
	$b_1 =$	-2.00	-2.00	-1.98
$T_i = a_2\theta^{b_2} + c_2$	$a_2 =$	$-$	20.65	5.11
	$b_2 =$	$-$	1.00	1.66
	$c_2 =$	$-$	$-$	2.67
$T_d = a_3\theta^{b_3} + c_3$	$a_3 =$	5.57	4.13	4.54
	$b_3 =$	1.00	1.00	0.72
	$c_3 =$	$-$	$-$	-0.60

3.5 Simulation Study

The closed loop transfer function zero results in large overshoot for setpoint tracking for integrating processes. A first/second order setpoint filter for PI/PID controller with the following transfer function is therefore considered in all the simulations.

$$F_{sp}(s) = \frac{1}{sT_i + 1} \tag{3.15}$$

$$F_{sp}(s) = \frac{1}{s^2 T_i T_d + sT_i + 1} \tag{3.16}$$

After a comparative study of the closed loop performances that can be achieved by the three integral criteria, guidelines for selecting the integral criterion is discussed in the Appendix. Furthermore, the results are compared with those obtained by adjusting the slope of the Nyquist curve at gain crossover frequency (slope method) and some of the latest reported approaches to illustrate the usefulness of the proposed tuning formulas.

Example 1. PI-control for IPTD process

Distillation is the most widely used separation technique in the chemical process industries for the separation of fluid mixtures. The distillation column separates a small amount of a low boiling material from the final product. The bottom level of the distillation column is controlled by adjusting the steam flow rate. The process model for the level control system is given by $G_p(s) = 0.2e^{-7.4s}/s$ [58]. The controller parameters obtained by various methods are given in Table 3.10. The step responses and the corresponding control signals are shown in Figs. 3.1 and

Table 3.10: PI parameters for Example 1

	SIMC	Slope method	IST^{3}E	KPGM [4]
K_p	0.34	0.32	0.44	0.30
T_i	59.20	38.78	30.93	49.36

3.2, respectively. A load disturbance of magnitude -0.20 is injected into the system at time 400 s. The slope method outperforms SIMC and KPGM for both setpoint tracking and load disturbance rejection. However, the IST^{3}E criterion gives better regulatory performances as compared to the slope method as is evident from Fig. 3.1.

Furthermore, as the parameters of the physical system vary with operating conditions and time, 20% change in the dead time θ is considered and the aforementioned design parameters are used to compare the robust performances of the design methods. Related results are shown in Fig. 3.3

and it is observed that the robustness to the assumed parameter perturbation is very satisfactory. The corresponding control signals are shown in Fig. 3.4.

Example 2. PID-control for IPTD process

An IPTD process model with the transfer function $0.0506e^{-6s}/s$ is considered and the corresponding controller parameters are given in Table 3.11. Figs. 3.5 and 3.6 show the step responses and the corresponding control signals for a unit step change in the setpoint and a disturbance of magnitude -0.50 introduced at time 100 s. The slope method gives almost similar load disturbance rejection performances as compared to Sree and Chidambaram's method. However, improved regulatory performances as compared to the method reported in [6] is obtained by the slope method. The IST^{2}E criterion gives the best performance for both the setpoint change and load disturbance rejection. Even though Kaya's method gives a faster response, the overshoot and settling time are more as compared to IST^{2}E criterion. The settling time for both servo and regulatory performances is more for Sree and Chidambarams's method [7] as compared to IST^{2}E criterion, thereby proving the value of the proposed tuning formulas.

Table 3.11: Controller parameters for Example 2

	Kaya [6]	Slope method	IST^{2}E	Sree and Chidambaram [7]
K_p	1.62	3.39	3.62	2.95
T_i	17.60	19.02	16.32	15
T_d	2.69	2.94	2.28	3
K_f	1.09	–	–	–

Example 3. PI-control for IFOPTD process

The design of PI controller for $G_p(s) = e^{-s}/s(s+1)$ is considered in this example. Table 3.12 shows the controller parameters obtained by various design methods. The step responses and the control signals to a unit setpoint and disturbance change, which is of magnitude -0.3 at time 80 s are given in Figs. 3.7 and 3.8. Almost similar performance is obtained by slope method and IST^{3}E criterion and both the methods outperform the approach reported in [8] with better load disturbance rejection as is evident from Fig. 3.7.

Table 3.12: PI parameters for Example 3

	Slope method	IST^{3}E	Poulin and Pomerleau [8]
K_p	0.37	0.34	0.29
T_i	9.70	8.76	7.62

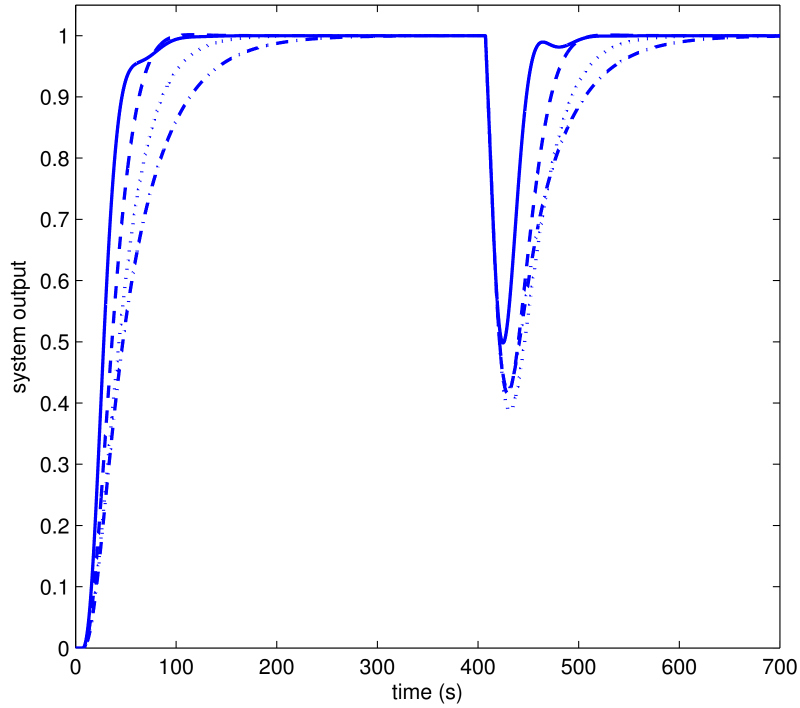


Figure 3.1: Step responses for Example 1: (–) IST^{3}E, (...) KPGM [4], (– –) Slope method and (–.–) SIMC [5]

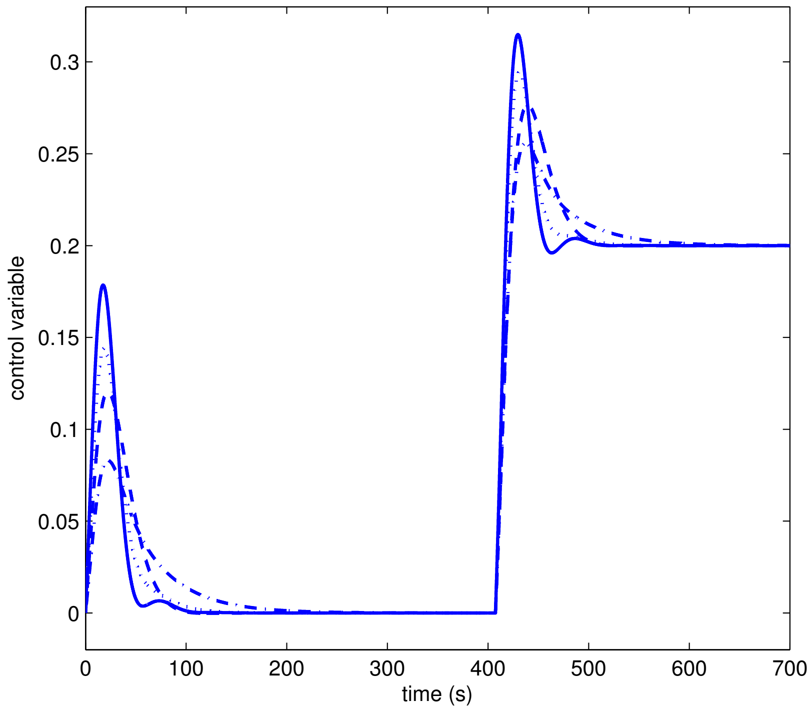


Figure 3.2: Control variables for step responses for Example 1: (–) IST^{3}E, (...) KPGM [4], (– –) Slope method and (–.–) SIMC [5]

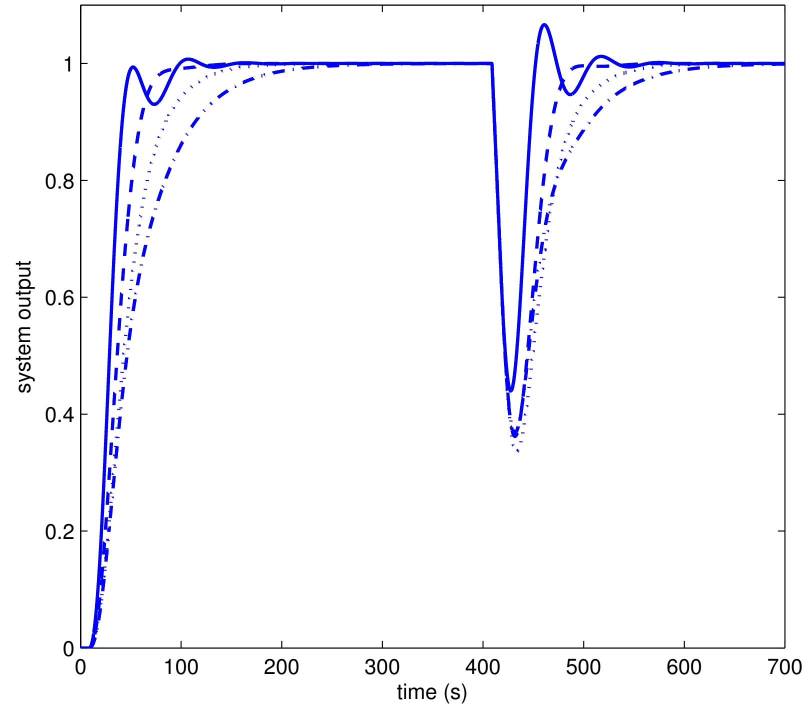


Figure 3.3: Robust responses for Example 1: (−) IST^{3}E, (...) KPGM [4], (−−) Slope method and (−.−) SIMC [5]

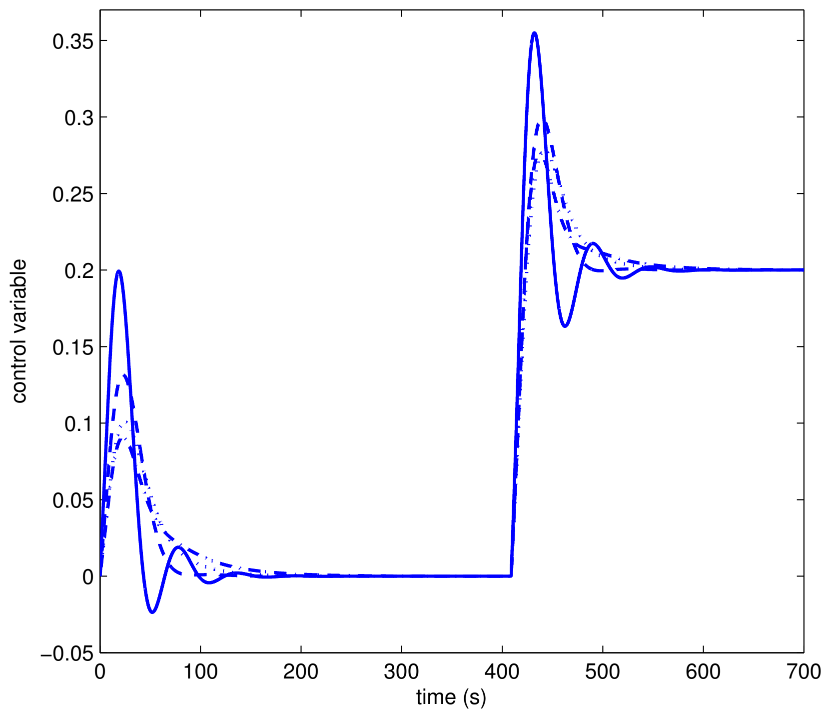


Figure 3.4: Control variables for robust responses for Example 1: (−) IST^{3}E, (...) KPGM [4], (−−) Slope method and (−.−) SIMC [5]

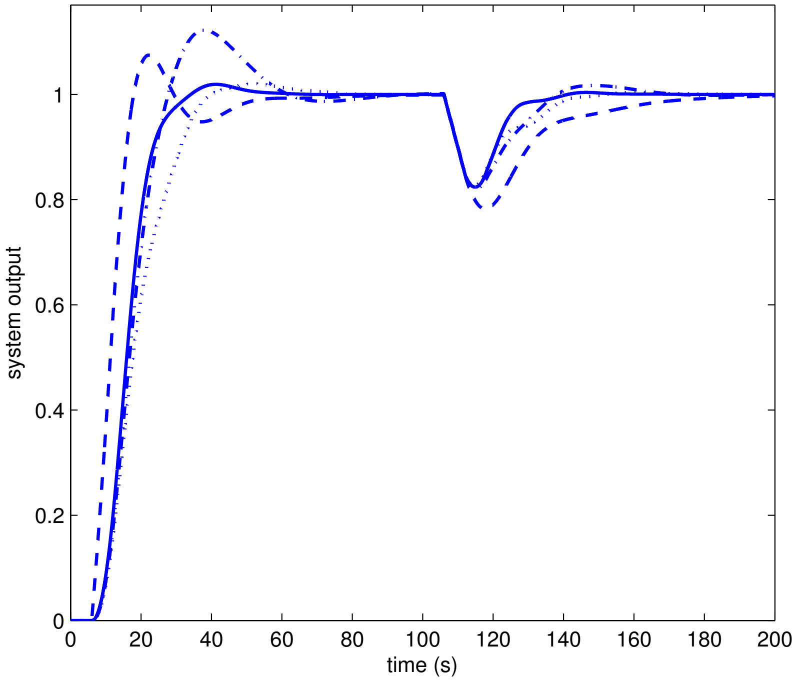


Figure 3.5: Step responses for Example 2: (—) IST^{2}E, (...) Slope method, (– –) Kaya [6] and (–.–) Sree and Chidambaram [7]

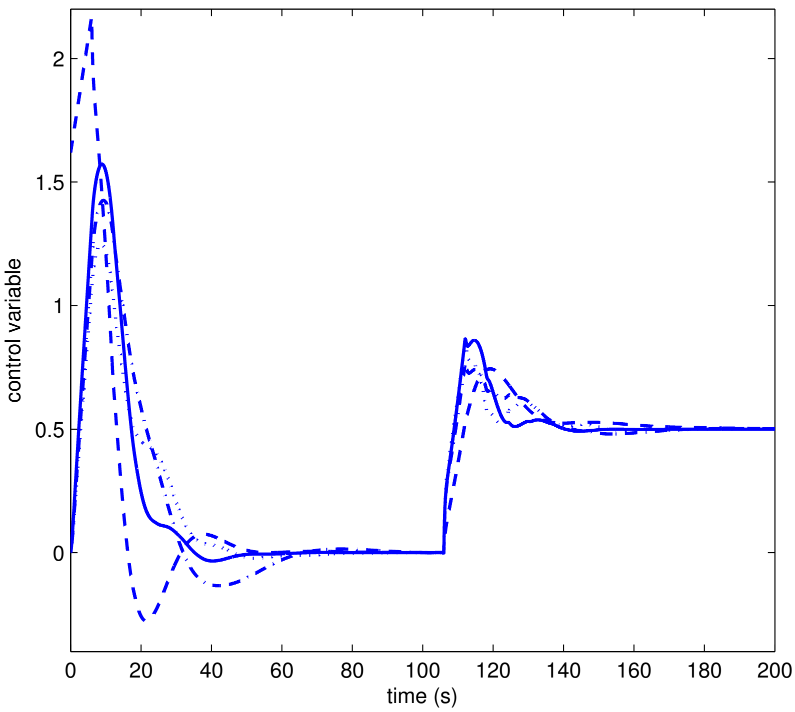


Figure 3.6: Control variables for step responses for Example 2: (—) IST^{2}E, (...) Slope method, (– –) Kaya [6] and (–.–) Sree and Chidambaram [7]

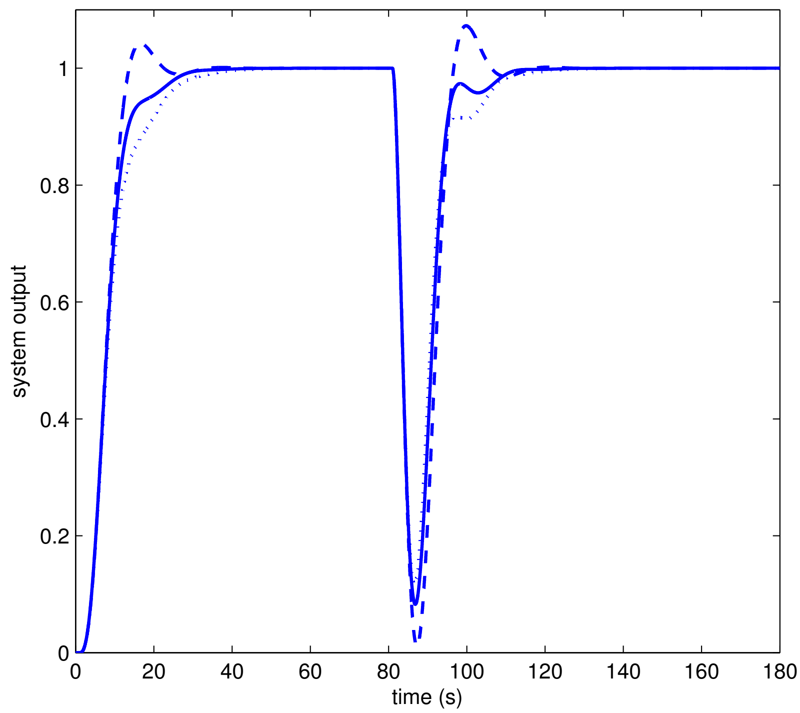


Figure 3.7: Step responses for Example 3: (–) IST^{3}E, (...) Slope method and (– –) Poulin and Pomerleau [8]

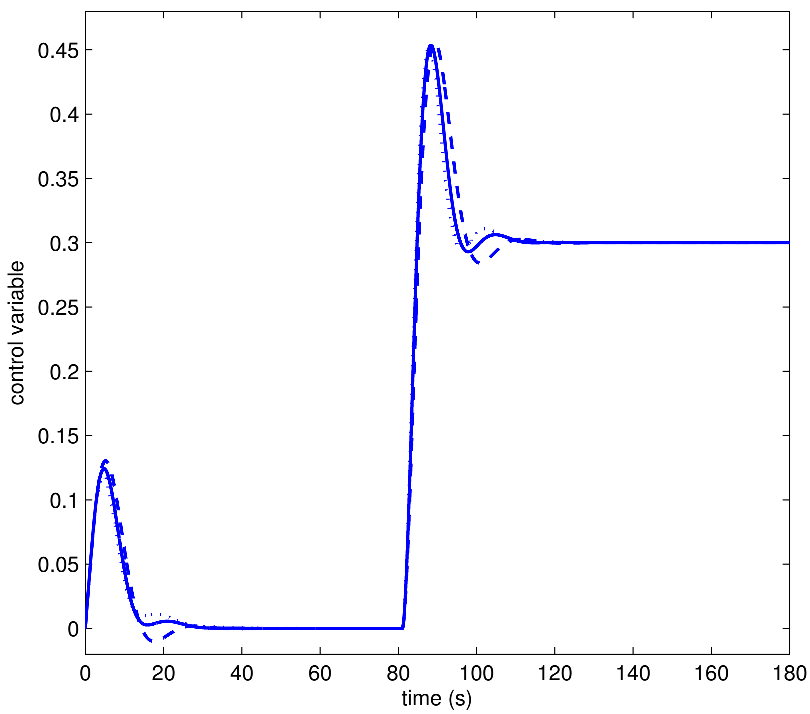


Figure 3.8: Control variables for step responses for Example 3: (–) IST^{3}E, (...) Slope method and (– –) Poulin and Pomerleau [8]

Example 4. PID-control for IFOPTD process

A higher order plant transfer function of $G_p(s) = e^{-0.5s}/s(s+1)(s+2)(s+3)$ is considered and the IFOPTD model obtained by Kaya [59] is $G_p(s) = 0.167e^{-1.077s}/s(1.863s+1)$. The results are compared with the SIMC tuning rules and the approach proposed in [9]. For Tan et al. method [9], the design parameter (λ) is taken as 0.25 and a setpoint filter with time constant equal to 0.80 times the integral time is considered. Figure 3.9 shows the responses for a unit change in setpoint and load disturbance introduced at time 75 s. The IST^{2}E settings and the slope method give almost similar and better control performances for load disturbance rejection as compared to SIMC and Tan et al.'s [9] controller settings.

Example 5. PID-control for DIPTD process

A DIPTD plant model given by $G_p(s) = e^{-\theta s}/s^2$ with the plant time delay assumed as unity is considered and the PID parameters are given in Table 3.13. The system outputs and the corresponding control variables for a unit step setpoint change and load disturbance, introduced at time 130 s are shown in Figs. 3.11 and 3.12. The magnitude of the load disturbance is assumed as -0.1. From the presented results, it can be observed that the slope method results in less settling time for both setpoint tracking and disturbance rejection as compared to SIMC and IST^{3}E integral criterion.

Also, it is to be noted that there always exists plant model mismatch and thus, the control performances for an assumed 20% change in the process dead time is shown in Fig. 3.13 and the corresponding control variables are shown in Fig. 3.14. As seen from the Figure, the proposed method gives satisfactory closed loop performance for the assumed parameter perturbations.

Table 3.13: Controller parameters for Example 5

	SIMC	Slope method	IST^{3}E
K_p	0.06	0.17	0.13
T_i	8.00	8.51	7.78
T_d	8.00	2.87	4.13

3.6 A New Objective Function

The new objective function proposed in the previous chapter (Section 2.7) results in robust closed loop step responses with no overshoot for many transfer function models commonly encountered in the process industries. Furthermore, it was shown that it resulted in a PD controller for integrating plant models which fails to reject the load disturbances and hence a PI-PD controller

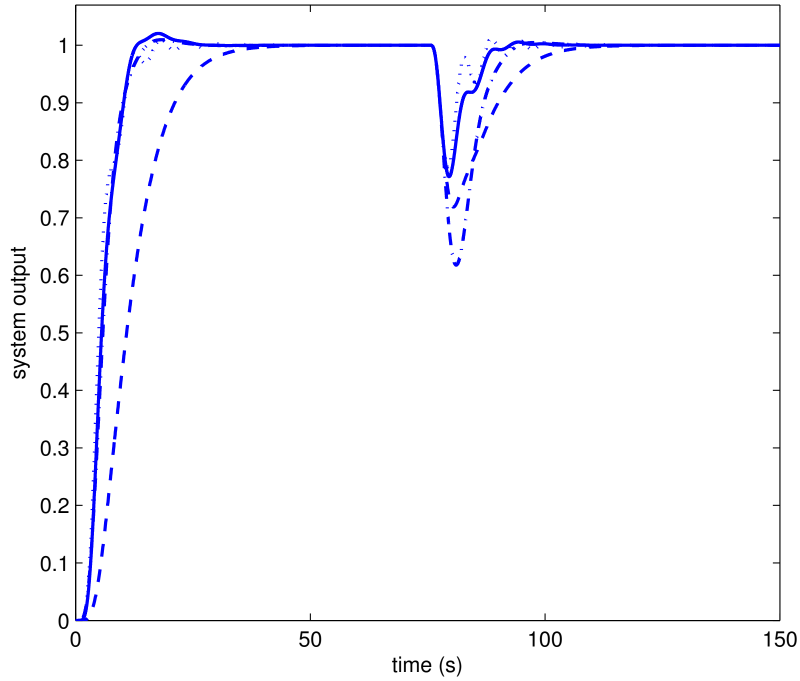


Figure 3.9: Step responses for Example 4: (—) IST^{2}E, (...) Slope method, (−−) SIMC [5] and (−.−) Tan et al. [9]

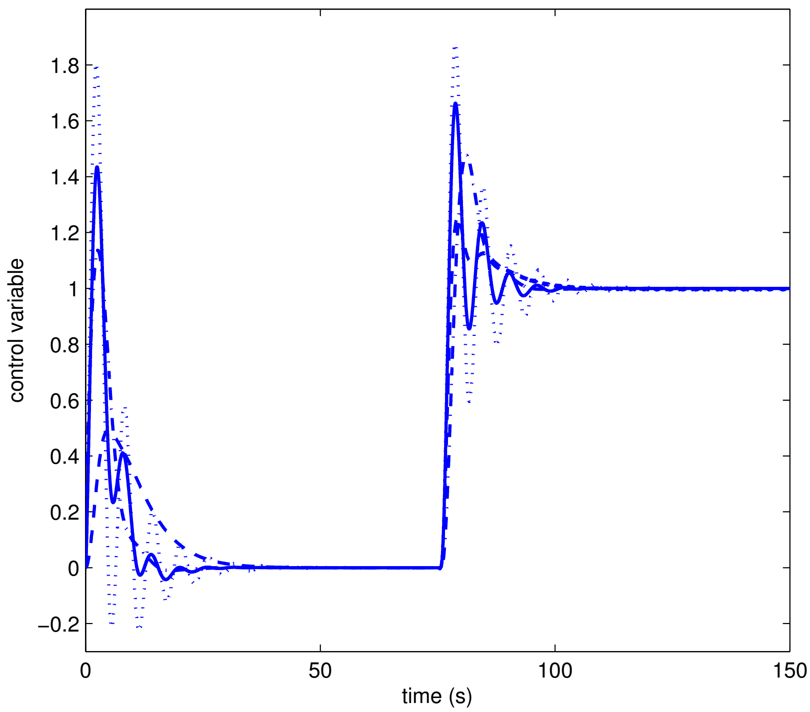


Figure 3.10: Control variables for step responses for Example 4: (—) IST^{2}E, (...) Slope method, (−−) SIMC [5] and (−.−) Tan et al. [9]

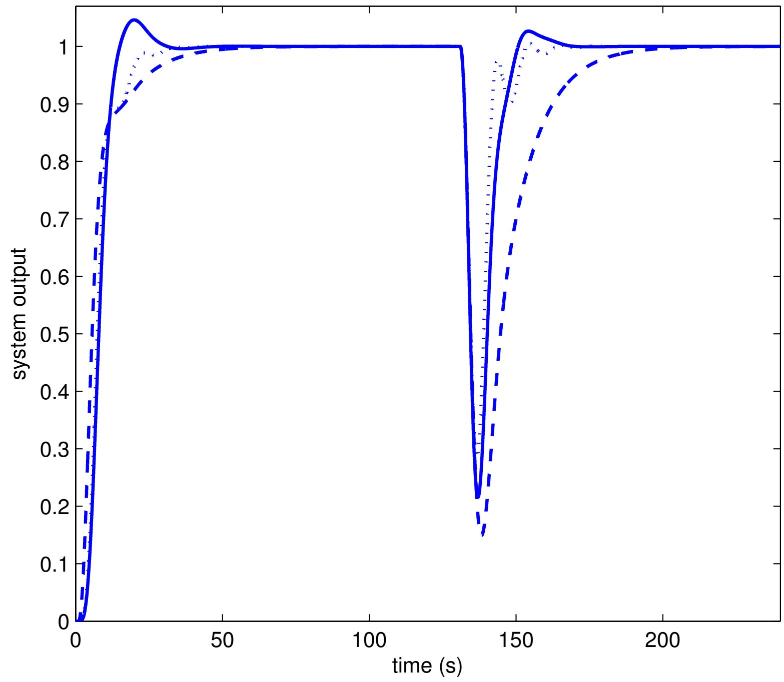


Figure 3.11: Step responses for Example 5: (–) IST^{3}E, (...) Slope method and (– –) SIMC [5]

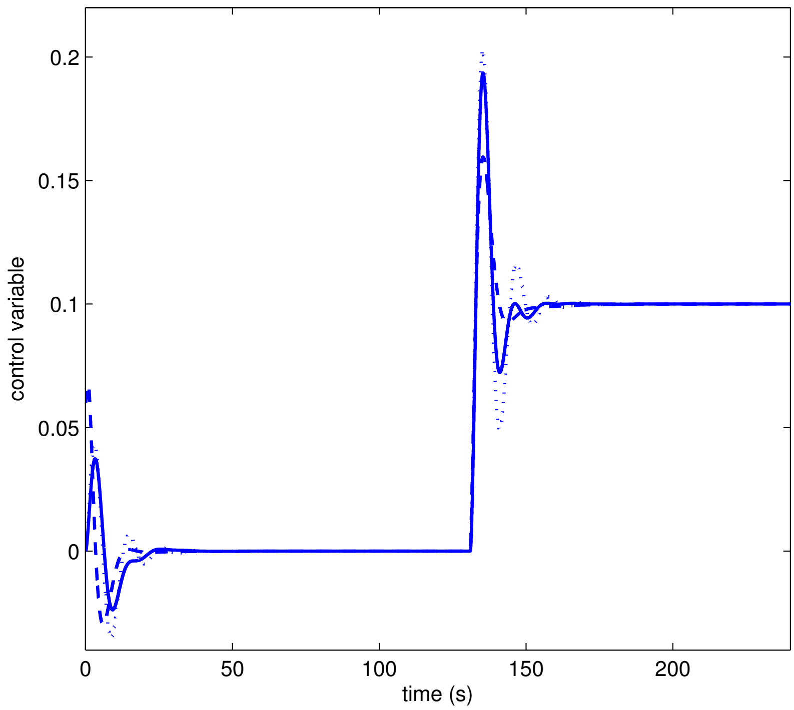


Figure 3.12: Control variables for step responses for Example 5: (–) IST^{3}E, (...) Slope method and (– –) SIMC [5]

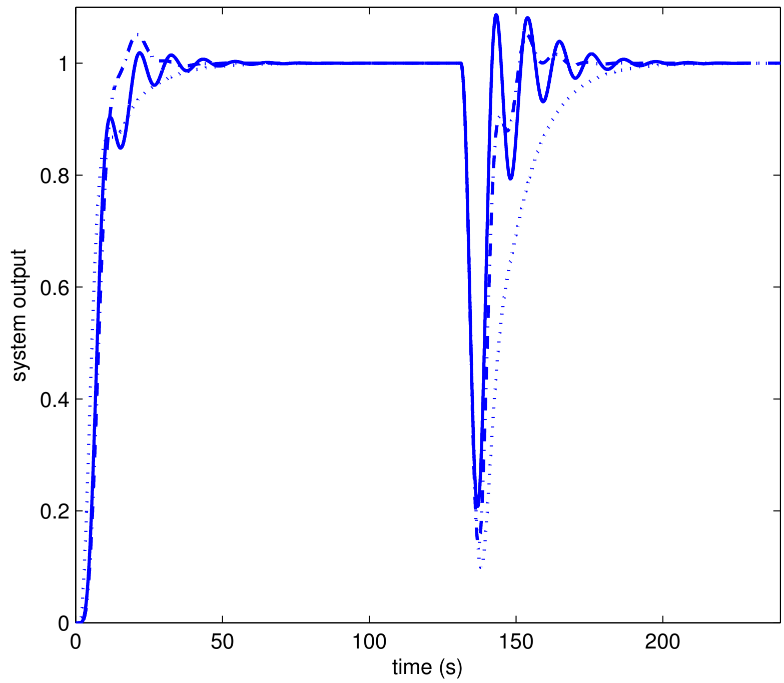


Figure 3.13: Robust responses for Example 5: (—) Slope method, (...) SIMC [5] and (—.—) IST^{3}E

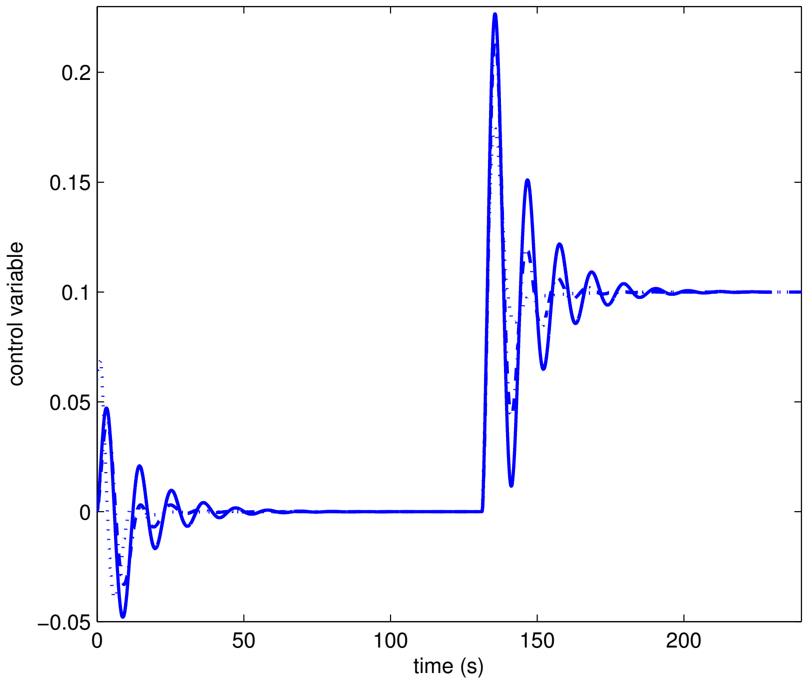


Figure 3.14: Control variables for robust responses for Example 5: (—) Slope method, (...) SIMC [5] and (—.—) IST^{3}E

was recommended for integrating plant models. In this section, analytical expressions correlating controller parameters to plant model parameters are obtained for IPTD and IFOPTD process models by minimizing the new objective function.

3.6.1 IPTD Process Model

The controller parameters obtained for various θ are given in Table 3.14 and the recommended controller settings are:

- $K_p = 0.32/(K\theta)$
- $T_i = 1.24\theta$
- $K_{pb} = 0.60/(K\theta)$
- $T_{db} = 0.51\theta$

Table 3.14: PI-PD parameters for IPTD process model

θ	KK_p	T_i	KK_{pb}	T_{db}
0.50	0.70	0.76	1.09	0.30
1	0.32	1.24	0.60	0.51
2	0.16	2.49	0.30	1.02
3	0.11	3.73	0.20	1.53

3.6.2 IFOPTD Process Model

Table 3.15 shows the controller parameters for IFOPTD process model and the recommended controller settings are:

- $K_p = 0.24/(K\theta)$
- $T_i = 1.40\theta + 0.71T$
- $K_{pb} = 0.44/(K\theta)$
- $T_{db} = 0.46\theta + 0.90T$

The step responses corresponding to the proposed controller settings for IPTD and IFOPTD plant models are shown in Figs. 3.15 and 3.16, respectively. It can be observed that satisfactory closed loop performances are achieved by the proposed tuning formulas.

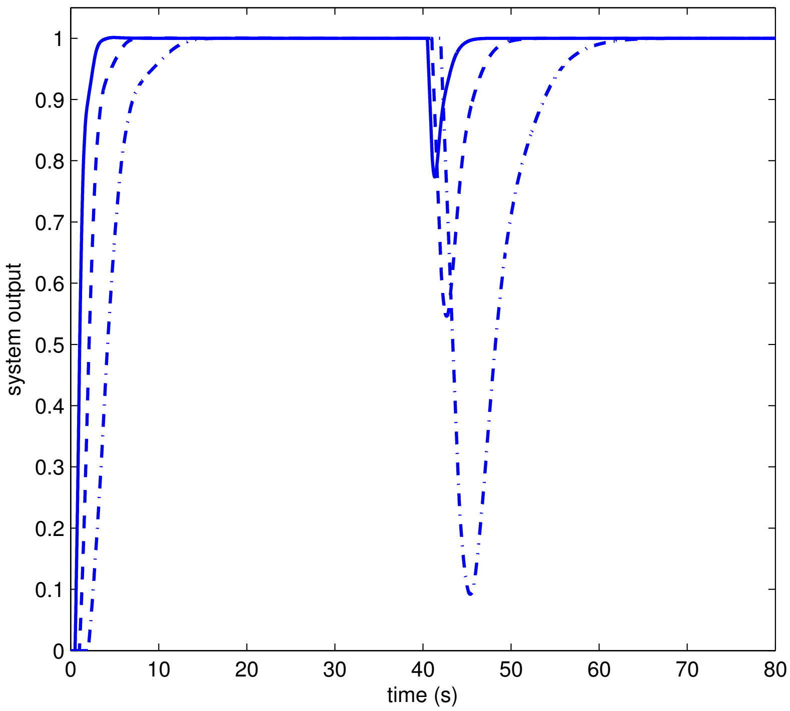


Figure 3.15: Step responses for $\frac{e^{-\theta s}}{s}$: $(-)$ $\theta = 0.5$, $(--)$ $\theta = 1$ and $(-.-)$ $\theta = 2$

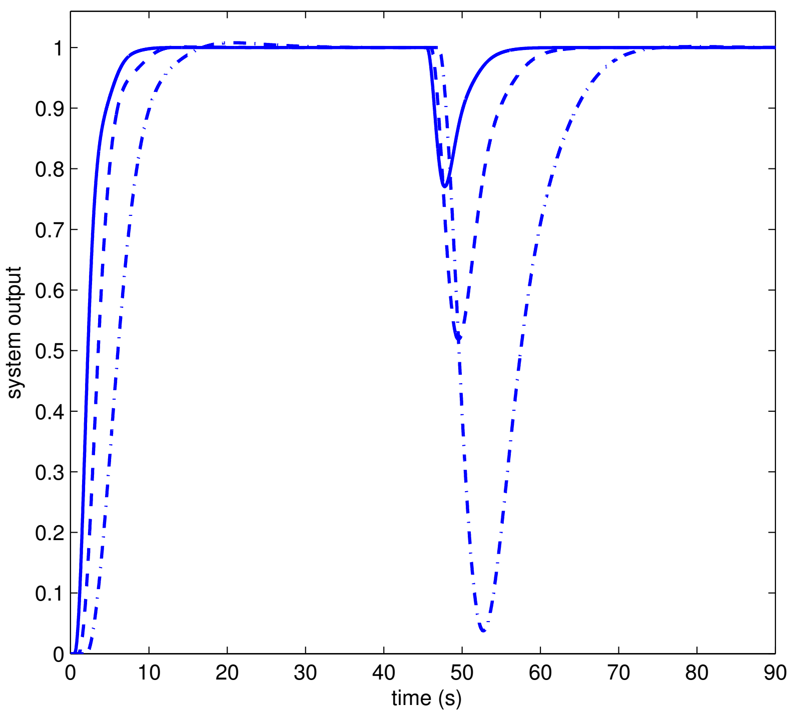


Figure 3.16: Step responses for $\frac{e^{-\theta s}}{s(s+1)}$: $(-)$ $\theta = 0.5$, $(--)$ $\theta = 1$ and $(-.-)$ $\theta = 2$

Table 3.15: PI-PD parameters for IFOPTD process model

θ/T	KK_pT	T_i/T	$KK_{pb}T$	T_{db}/T
0.50	0.49	1.35	0.89	1.11
1	0.24	2.16	0.44	1.38
2	0.12	3.57	0.23	1.82
3	0.09	4.88	0.16	2.27

3.6.3 Simulation Study

The IPTD plant model considered in [10] is $G_p(s) = \frac{e^{-5s}}{s}$. The proposed controller parameters are $K_p = 0.06$, $T_i = 6.20$, $K_{pb} = 0.12$ and $T_{db} = 2.55$. The controller settings recommended in [10] are $K_p = 0.20$, $T_i = 17.50$, $T_d = 2.14$ and $T_f = 0.07$. The setpoint weighting parameter is taken as 0.40. With these controller settings, the performances of the closed loop control system are evaluated by giving a unit step input in the setpoint and a negative step input of 0.1 in the load at $t = 100$ s. The step responses are shown in Fig. 3.17. It is observed that both the methods give almost similar setpoint and load disturbance rejection performances. To illustrate the robustness of the proposed tuning formulas, 30% perturbation is given in both K and θ and the corresponding step responses are shown in Fig 3.18. Rao et al.'s method gives unstable step responses whereas the proposed method gives satisfactory closed loop performances.

3.7 Conclusions

In this chapter, tuning formulas for a class of integrating processes are proposed by adjusting the slope of the Nyquist curve at gain crossover frequency. The controller parameters are also obtained by minimizing ISTE, IST^{2}E and IST^{3}E criteria. A comparative study of the performances that can be achieved by the proposed tuning methods is done to help the user in selecting the tuning rules. The results are compared with some of the latest reported approaches to prove the value of the proposed tuning formulas. Also, it is shown that the proposed design methods give satisfactory closed loop performances for perturbations in the process parameters. Tuning formulas for PI-PD controller that ensure robust closed loop performances are also given for IPTD and IFOPTD plant models.

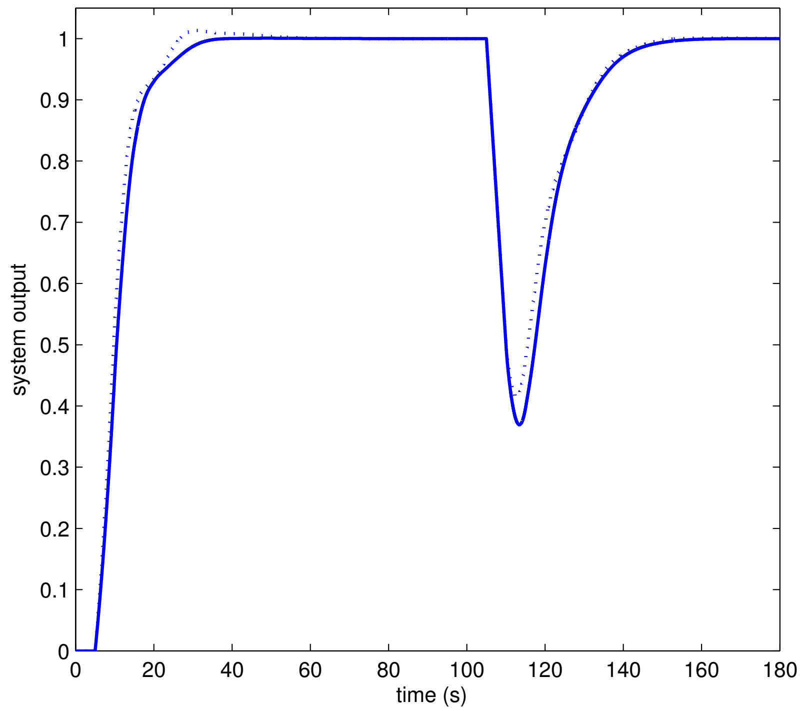


Figure 3.17: Step responses for $\frac{e^{-5s}}{s}$: (–) proposed PI-PD and (...) Rao et al. [10]

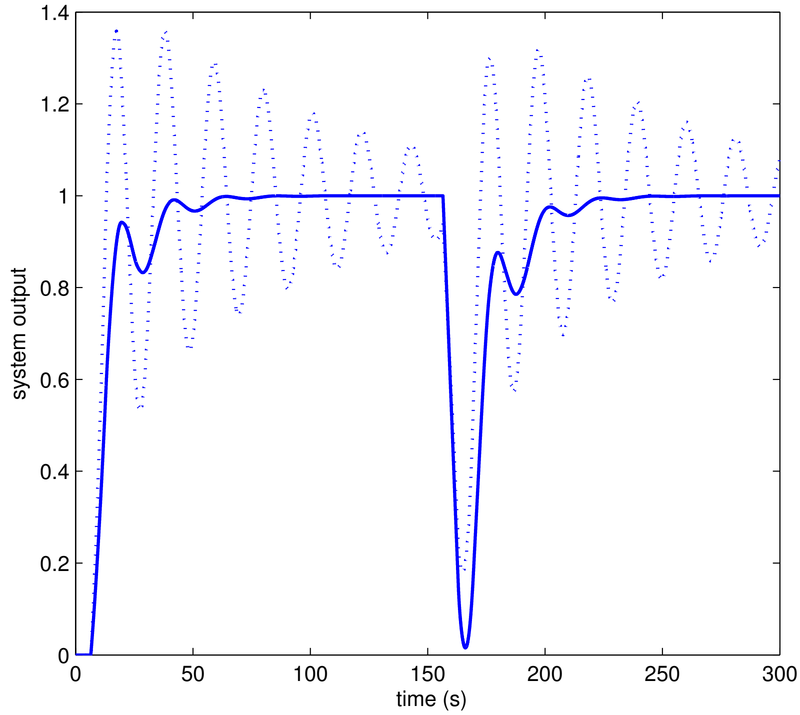


Figure 3.18: Robust responses for $\frac{e^{-5s}}{s}$: (–) proposed PI-PD and (...) Rao et al. [10]

Chapter 4

Modified Smith Predictor for Integrating Processes

4.1 Introduction

Good control of plants with a long time-delay may be difficult using a PID controller. This is because the additional phase lag contributed by the time delay tends to destabilize the closed loop system. The stability problem can be solved by decreasing the controller gain which results in a sluggish closed loop response. The Smith predictor is a popular and very effective long dead-time compensator for stable processes. The main advantage of the Smith predictor method is that time delay is effectively taken outside of the control loop in the transfer function relating process output to setpoint. However, this approach fails in a very significant way for an unstable process due to the problem of stabilization. Furukawa and Shimemura [60] augmented the scheme with an observer, Watanabe and Ito [26] deliberately replaced the known process by a mismatched process model and De Paor's [29] approach has a severe constraint on the ratio of dead-time to the time constant even for the setpoint response of the predictor. Later on, De Paor and Egan [30] extended and partially optimized the modified Smith predictor and controller [29] to relax the constraint on the ratio, but there is no significant extension to the ranges of that ratio for which closed loop asymptotic stability can be achieved. These cited authors have thus solved the stabilization problem by involving greater complexity than the Smith predictor.

Watanabe and Ito [26] pointed out that if the process has poles near the origin in the left half plane, then the responses may be sluggish enough to be unacceptable. Since then, a number of methods have been proposed to overcome the problem of controlling a process with an integrator and long dead time. A new Smith predictor that isolates the setpoint response from the load disturbance response and gives improved closed loop performances was proposed in [27]. Mataušek and Micič [61, 62] have shown that the disturbance response of the Smith predictor can be controlled independently of the setpoint response for integrating processes by adding a P/PD controller in the feedback path from the difference of the plant output and the model output to the reference input. Majhi and Atherton [28] modified the original Smith predictor by providing an internal feedback loop across the delay free part of the process model and sending the same amount of signal to the original process to stabilize the unstable process. Three controllers: a PI controller for setpoint tracking, a P/PD controller for disturbance rejection and another P controller for stabilizing the unstable and integrating processes are used. The P/PD controller used for disturbance rejection was designed using the approach presented in [63].

Recently, several modifications of the basic Smith predictor scheme have been proposed to enhance the closed loop performance for integrating processes with dominant time delay [64–72]. Liu et al. [73] have proposed a modified Smith predictor scheme with three controllers: a setpoint tracking controller, a disturbance estimator and a third controller for stabilizing the integrating process without time delay. A double two degree of freedom control scheme with four controllers for controlling integrating and unstable processes has been proposed in [74]. Rao et al. [11] have proposed an analytical method for designing the controllers for a modified Smith predictor for a class of integrating processes. The setpoint tracking controller is designed using the classical direct synthesis method whereas the PD controller used for disturbance rejection is tuned to achieve user defined gain and phase margins. However, the selection of tuning parameters namely, the closed loop time constant, setpoint weighting parameter, gain and phase margins is a difficult task for the user.

In this chapter, the modified Smith predictor proposed in [28] is considered and the parameters of the PD controller employed for load disturbance rejection are obtained by adjusting the slope of the Nyquist curve at gain crossover frequency. Guidelines are provided for selecting the gain crossover frequency and the slope of the Nyquist curve. Tuning formulas which have physically meaningful parameters like the damping ratio, ζ and the natural frequency, ω_n are also given for tuning the setpoint tracking controller for a class of integrating processes.

4.2 Modified Smith Predictor

The modified Smith predictor proposed by Majhi and Atherton [28] is shown in Fig. 4.1. It is similar to the structure suggested in [27] apart from G_{c2} which has a major role for an unstable and integrating plant and has three controllers which are designed for different objectives. Of the three controllers, G_{c2} in the inner loop is provided to stabilize an unstable or integrating process. The other two controllers G_{c1} and G_{c3} are then used to take care of servo-tracking and disturbance rejection respectively by considering the inner loop as an open-loop stable process. Also the signal given by the G_{c3} feedback can be interpreted as an estimate of the input disturbance L. Based on the assumption that the model used perfectly matches the plant dynamics i.e. $G_m = G$ and $\theta_m = \theta$, the closed loop setpoint and load disturbance responses are given by

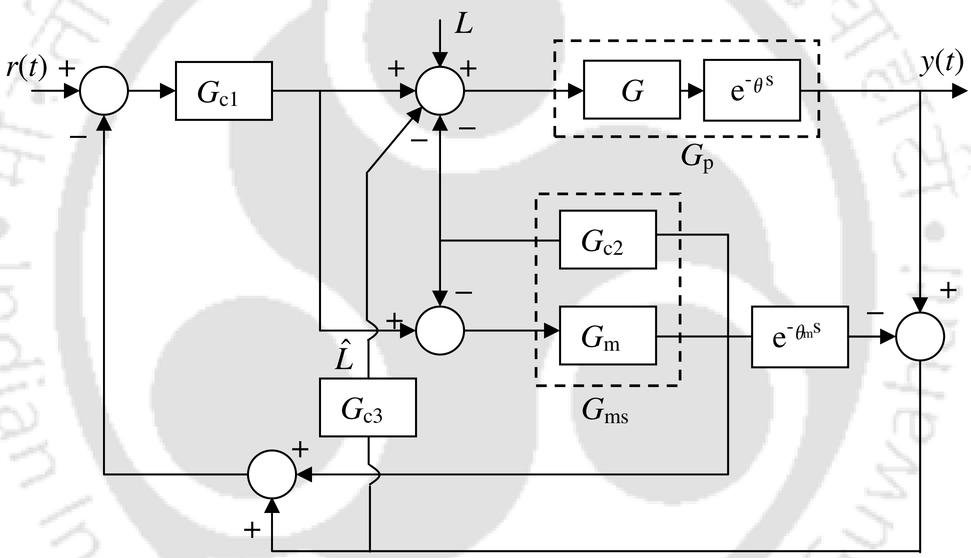


Figure 4.1: Modified Smith predictor

$$Y(s) = Y_r(s)R(s) + Y_L(s)L(s) \tag{4.1}$$

where

$$Y_r(s) = \frac{GG_{c1}e^{-\theta s}}{1 + G(G_{c1} + G_{c2})} = T_r(s)e^{-\theta s} \tag{4.2}$$

$$Y_L(s) = \frac{Ge^{-\theta s}}{1 + G(G_{c1} + G_{c2})} \frac{1 + G(G_{c1} + G_{c2}) - GG_{c1}e^{-\theta s}}{1 + GG_{c3}e^{-\theta s}} \tag{4.3}$$

It can be observed that the setpoint responses are decoupled from the load disturbance rejection responses. From (4.2), it can be observed that the setpoint tracking depends on the controllers G_{c1} and G_{c2}. Once these controllers have been designed for satisfactory setpoint tracking, the load disturbance rejection is taken care of by G_{c3}.

4.3 Controller Design for Integrating Processes

The modified Smith predictor shown in Fig. 4.1 requires design of three controllers (G_{c1}, G_{c2} and G_{c3}) for a class of integrating processes. G_{c2} is chosen as a proportional controller with unity gain and the design of G_{c1} and G_{c3} are explained hereafter.

4.3.1 Design of G_{c1}

Case (i): IPTD process model

The delay free part of the IPTD model is

$$G_m(s) = \frac{K}{s} \tag{4.4}$$

Assuming G_{c1} to be a PI controller and using (4.2), the delay free part of the closed loop transfer function for servo tracking is obtained as

$$T_r(s) = \frac{\frac{KK_p}{T_i}(1 + sT_i)}{s^2 + sK(1 + K_p) + \frac{KK_p}{T_i}} \tag{4.5}$$

The characteristic equation is given by the denominator of (4.5). By setting the closed-loop characteristic equation to match a desired critically damped closed loop system such as

$$s^2 + 2\omega_n s + \omega_n^2 = 0 \tag{4.6}$$

we get the following equations

$$\begin{aligned} K_p &= \frac{2\omega_n}{K} - 1 \\ T_i &= \frac{KK_p}{\omega_n^2} \end{aligned} \tag{4.7}$$

The parameters of the PI controller can be obtained from the above equations for a given ω_n. Time domain specification namely, rise time is used to obtain ω_n in this work. The relation between the rise time and the natural frequency [75] for a given damping ratio ζ is given by

$$\begin{aligned} t_r &= \frac{1 - 0.4167\zeta + 2.917\zeta^2}{\omega_n} \\ &= \frac{3.5003}{\omega_n} \qquad (\text{for} \qquad \zeta = 1) \end{aligned} \tag{4.8}$$

For a given t_r, ω_n can be calculated using (4.8) which is then substituted in (4.7) to obtain the parameters of the PI controller.

Case (ii): IFOPTD plant model

The delay free part of the closed loop transfer function for servo control assuming G_{c1} as a parallel PID controller is given by

$$T_r(s) = \frac{\frac{KK_p}{T_iT}(s^2T_iT_d + sT_i + 1)}{s^3 + s^2(\frac{1+KK_pT_d}{T}) + sK(\frac{1+K_p}{T}) + \frac{KK_p}{T_iT}} \tag{4.9}$$

By setting the closed-loop characteristic equation to match (4.10),

$$(s + \bar{\beta}\omega_n)(s^2 + 2\omega_n s + \omega_n^2) = 0 \tag{4.10}$$

we obtain the following equations by equating the coefficients of s^0, s^1 and s^2 terms

$$\begin{aligned} \frac{KK_p}{T_iT} &= \bar{\beta}\omega_n^3 \\ \frac{K(1+K_p)}{T} &= \omega_n^2(1+2\bar{\beta}) \\ \frac{1+KK_pT_d}{T} &= \omega_n(\bar{\beta}+2) \end{aligned} \tag{4.11}$$

The characteristic equation can be approximated to that of the second order system by a proper choice of $\bar{\beta}$. The natural frequency for a given rise time can be calculated using (4.8). Once, ω_n and $\bar{\beta}$ are known, the PID parameters can be obtained from (4.11).

Case (iii): DIPTD plant model

Considering G_{c1} to be a parallel PID controller, the following equations are obtained for DIPTD plant model

$$\frac{KK_p}{T_i} = \bar{\beta}\omega_n^3 \tag{4.12}$$

$$K(1+K_p) = \omega_n^2(1+2\bar{\beta}) \tag{4.13}$$

$$KK_pT_d = \omega_n(\bar{\beta}+2) \tag{4.14}$$

Then, the parameters of the PID controller are estimated following the procedure given in the above *case (ii)*.

4.3.2 Design of G_{c3}

G_{c3} is assumed as a PD controller of the form:

$$G_{c3} = K_{p3}(1+sT_{d3}) \tag{4.15}$$

It can be observed that the denominator of (4.3) has the term $(1+GG_{c3}e^{-\theta s})$. Once the controllers G_{c1} and G_{c2} are designed for satisfactory setpoint tracking, the parameters of G_{c3} are obtained by adjusting the slope of the Nyquist curve of the loop transfer function $L(s) = GG_{c3}e^{-\theta s} = G_pG_{c3}$ at gain crossover frequency (ω_g). The derivative of the loop transfer function with respect to ω is given as

$$\frac{dL(j\omega)}{d\omega} = G_p(j\omega)\frac{dG_{c3}(j\omega)}{d\omega} + G_{c3}(j\omega)\frac{dG_p(j\omega)}{d\omega} \tag{4.16}$$

Furthermore, one has

$$\ln G_p(j\omega) = \ln|G_p(j\omega)| + j\angle G_p(j\omega) \tag{4.17}$$

Differentiating the above equation gives

$$\frac{dG_p(j\omega)}{d\omega} = G_p(j\omega)\left\{\frac{d\ln|G_p(j\omega)|}{d\omega} + j\frac{d\angle G_p(j\omega)}{d\omega}\right\} \tag{4.18}$$

The derivative of the controller transfer function with respect to ω gives

$$\frac{dG_{c3}(j\omega)}{d\omega} = jK_pT_d \tag{4.19}$$

Substituting (4.15), (4.18) and (4.19) in (4.16) gives

$$\frac{dL(j\omega)}{d\omega} = G_p(j\omega)K_p\left\{jT_{d3} + (1+jT_{d3}\omega)\left(\frac{d\ln|G_p(j\omega)|}{d\omega} + j\frac{d\angle G_p(j\omega)}{d\omega}\right)\right\} \tag{4.20}$$

The slope of the Nyquist curve at ω_o is given by

$$\psi = \phi_o + \arctan \frac{s_p(\omega_o) + s_a(\omega_o)\omega_o T_d + \omega_o T_d}{s_a(\omega_o) - s_p(\omega_o)\omega_o T_d} \tag{4.21}$$

where

$$\begin{aligned} \phi_o &= \angle G_p(j\omega_o) \\ s_a(\omega_o) &= \omega_o \left.\frac{d \ln |G_p(j\omega)|}{d\omega}\right|_{\omega_o} \\ s_p(\omega_o) &= \omega_o \left.\frac{d\angle G_p(j\omega)}{d\omega}\right|_{\omega_o} \end{aligned} \tag{4.22}$$

Rearranging the above equation, we get

$$T_d = \frac{\bar{a}s_a(\omega_o) - s_p(\omega_o)}{\omega_o(1 + \bar{a}s_p(\omega_o) + s_a(\omega_o))} \tag{4.23}$$

where $\bar{a} = tan(\psi - \phi_o)$. The derivative time constant of the PD controller can be obtained from (4.23) for a given $\omega_o = \omega_g$ and ψ. Using the condition $|L(j\omega_g) = 1|$, we get the following equations

$$KK_p = \frac{\omega_g}{\sqrt{1 + \omega_g^2 T_d^2}} \tag{4.24}$$

$$KK_p = \frac{\omega_g \sqrt{1 + \omega_g^2 T^2}}{\sqrt{1 + \omega_g^2 T_d^2}} \tag{4.25}$$

$$KK_p = \frac{\omega_g^2}{\sqrt{1 + \omega_g^2 T_d^2}} \tag{4.26}$$

for IPTD, IFOPTD and DIPTD process models, respectively. Once T_d is obtained, the proportional gain K_p can be calculated using the above equations.

4.3.3 Selection of Tuning Parameters

The main advantage of the proposed method is that the design is simple with only two controllers G_{c1} and G_{c3} since $G_{c2} = 1$. There are four tuning parameters: rise time (t_r), gain cross over frequency (ω_g), slope of the Nyquist curve, and $\bar{\beta}$. A smaller value of t_r implies faster response with better load disturbance rejection but at the cost of robustness. A smaller value of t_r implies more robustness with degraded closed loop performance. Thus, the value of t_r can be obtained depending on the system requirements. The gain crossover frequency is selected based on Åström's inequality proposed in [56] and it is observed that $\omega_g = 0.78/\theta$ which corresponds to a phase margin of $45°$ gives satisfactory closed loop performances for the integrating process models considered in this work. Furthermore, $\psi = 90°$ is recommended for IPTD and IFOPTD plant models whereas ψ should be taken as $45°$ for DIPTD plant models. Regarding the choice of $\bar{\beta}$, it is observed that a large $\bar{\beta}$ results in oscillatory control signal and hence $\bar{\beta}$ is taken as 5 in this work. The appropriateness of the selection is assessed in the following section.

4.4 Simulation Study

In this section, one example each for an IPTD, IFOPTD and DIPTD process models are considered to illustrate the use of the proposed design method. The results are compared with Rao et al.'s [11] method as the authors of the cited work have shown that their method gives better control performances as compared to the recently reported approaches in the literature.

Example 1

Consider an IPTD plant with the transfer function model

$$G_p(s) = \frac{e^{-5s}}{s} \tag{4.27}$$

The parameters of the setpoint tracking controller proposed in [11] are $K_p = 0.4$ and $T_i = 10$. The setpoint weighting parameter is chosen as 0.6 and the PD controller designed for load disturbance rejection using gain and phase margin criteria is $(0.12 + 0.45s)$. Rao et al. have shown that their method gives better results than Zhong and Normey-Rico [70], Liu et al. [73] and Lu et al. [74].

For a rise time of 4.0 seconds, we get $\omega_n = 0.87$ and the corresponding PI parameters obtained using (4.7) are $K_p = 0.74$ and $T_i = 0.98$. The PD controller designed for load disturbance rejection is $G_{c3} = 0.15(1 + 1.74s)$. With these controller settings, the performances of the closed loop system is evaluated by giving a unit step input in the setpoint and a negative step input of 0.1 in the load at $t = 60$ seconds. The step responses for no mismatch in the plant model are shown in Fig. 4.2. It can be observed that the proposed method outperforms Rao et al.'s method for both setpoint tracking and load disturbance rejection. The corresponding control variables are shown in Fig. 4.3.

Now suppose that there exists a 10% error in estimating the process time delay θ such that it is actually 10% larger. The perturbed system responses and the corresponding control variables are shown in Figs. 4.4 and 4.5, respectively. It can be observed that the robustness of the proposed design technique towards assumed parameter perturbation is satisfactory.

The effect of rise time on system robustness is studied by simulating the system responses for the perturbed plant model for two values of t_r. The corresponding controller parameters are given in Table 4.1. The system outputs and the corresponding control signals are shown in Figs. 4.6 and 4.7, respectively. Fig. 4.6 shows that there are less oscillations in the system output for a larger t_r with poor load disturbance rejection performances. The rise time should be therefore chosen to achieve the best tradeoff between the nominal performance and robust stability of the closed loop in the proposed control structure so as to satisfy the disturbance rejection requirement in practice.

Table 4.1: Controller parameters for perturbed process model of Example 1

t_r (s)	G_{c1}		G_{c3}	
	K_p	T_i	K_{p3}	T_{d3}
3	1.34	0.98	0.14	1.91
5	0.40	0.82	0.14	1.91

Example 2

Consider an IFOPTD process with the transfer function model

$$G_p(s) = \frac{1}{s(3.4945s+1)}e^{-6.567s} \tag{4.28}$$

Rao et al. have already shown that their method gives improved closed loop performances as compared to Kaya's method [59] and Liu et al.'s method proposed in [73]. The controller parameters proposed by Rao et al. are: $K_p = 0.124$, $T_i = 27.58$ and $T_d = 1.130$ for setpoint tracking and a PD controller with transfer function $(0.035 + 1.12s)$ for load disturbance rejection.

Table 4.2: G_{c1} for Example 2 for various $\bar{\beta}$

$\bar{\beta}$	G_{c1}		
	K_p	T_i	T_d
5	0.11	1.28	28.71
10	1.12	6.52	5.47
15	2.13	8.27	4.27

The proposed controller parameters for G_{c1} for various $\bar{\beta}$ for $t_r = 20$ seconds are given in Table 4.2. The PD controller corresponding to $\omega_g = 0.78/\theta$ and $\psi = 90°$ is $(0.11 + 4.72s)$. A unit step input is added at $t = 0$ and an inverse step load disturbance of magnitude 0.1 is added to the process input at $t = 100s$. Simulation results are shown in Figs. 4.8 and 4.9, respectively. The proposed method gives improved closed loop performances for both setpoint tracking and load disturbance rejection for the considered values of $\bar{\beta}$.

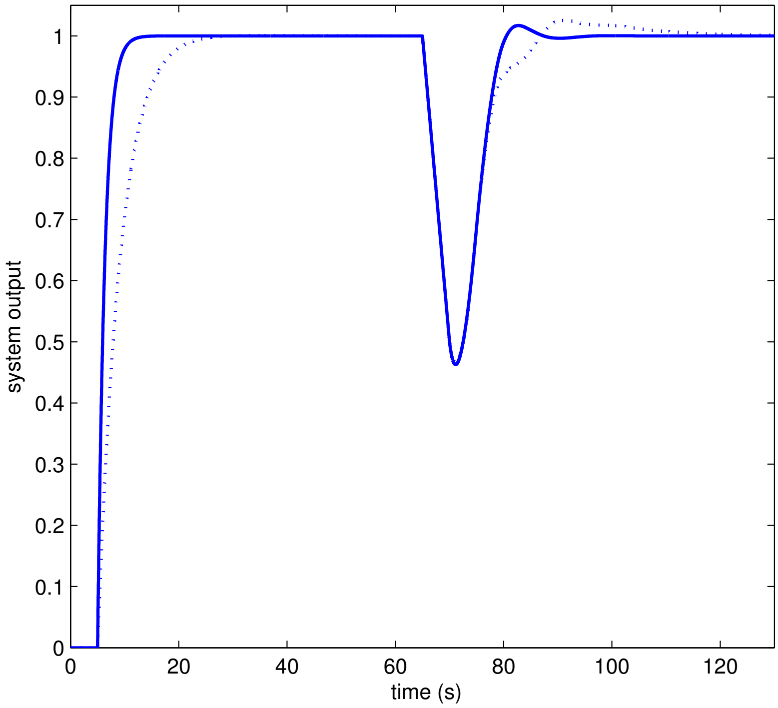


Figure 4.2: Nominal responses for Example 1: (−) proposed and (...) Rao et al. [11]

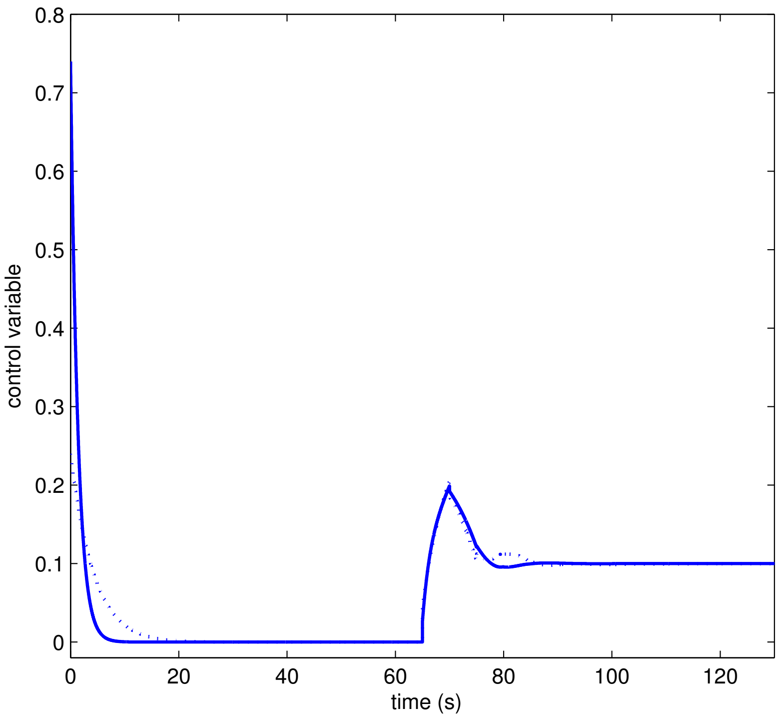


Figure 4.3: Control variables for nominal responses for Example 1: (−) proposed and (...) Rao et al. [11]

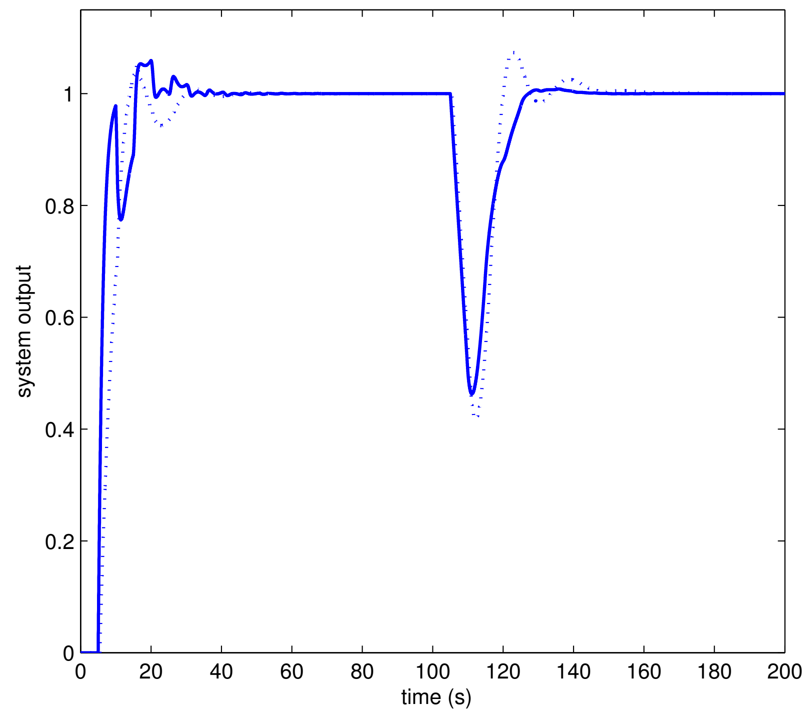


Figure 4.4: Perturbed system responses for Example 1: (−) proposed and (...) Rao et al. [11]

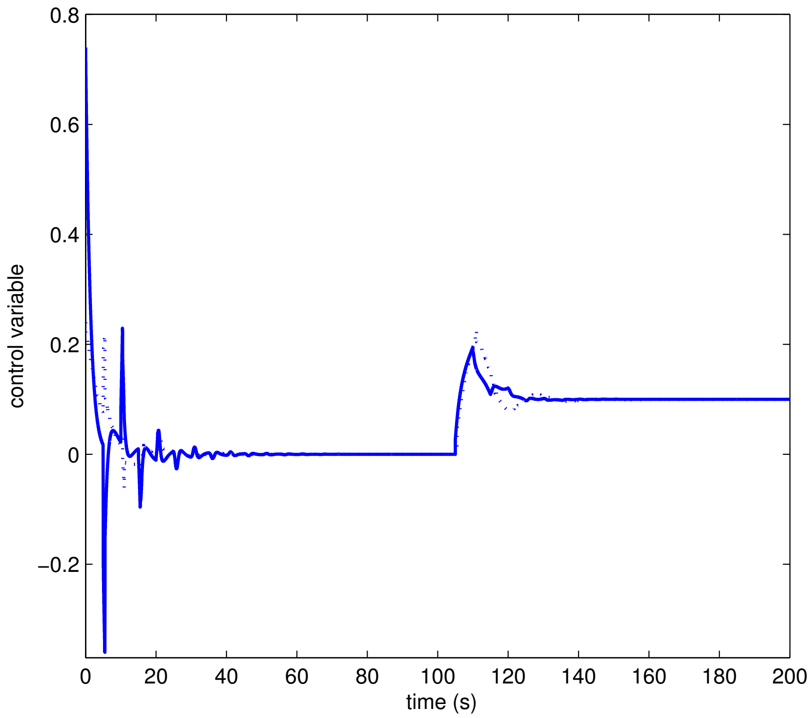


Figure 4.5: Control variables for perturbed system responses for Example 1: (−) proposed and (...) Rao et al. [11]

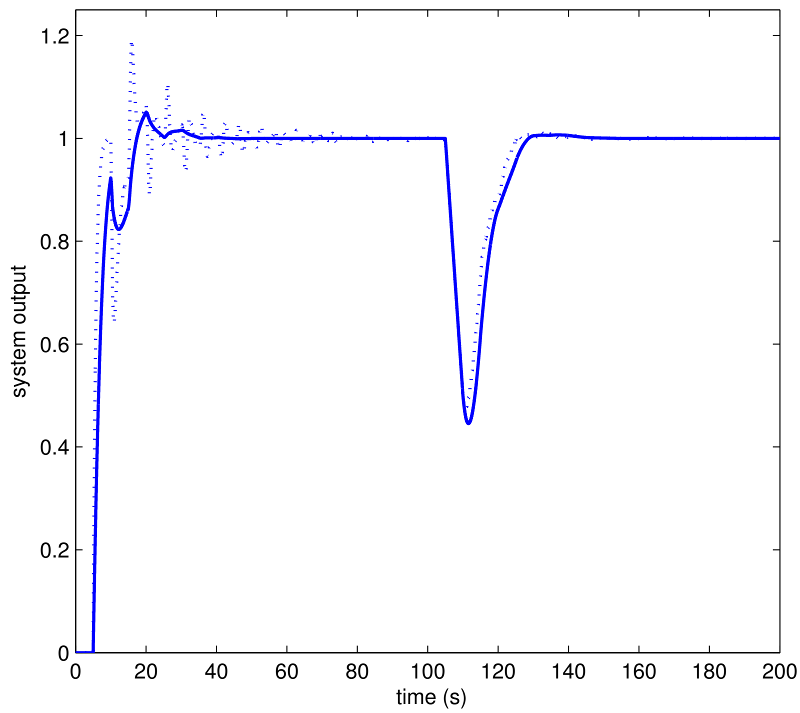


Figure 4.6: Perturbed system responses for Example 1: $(-)\ t_r = 5$ and $(...)\ t_r = 3$

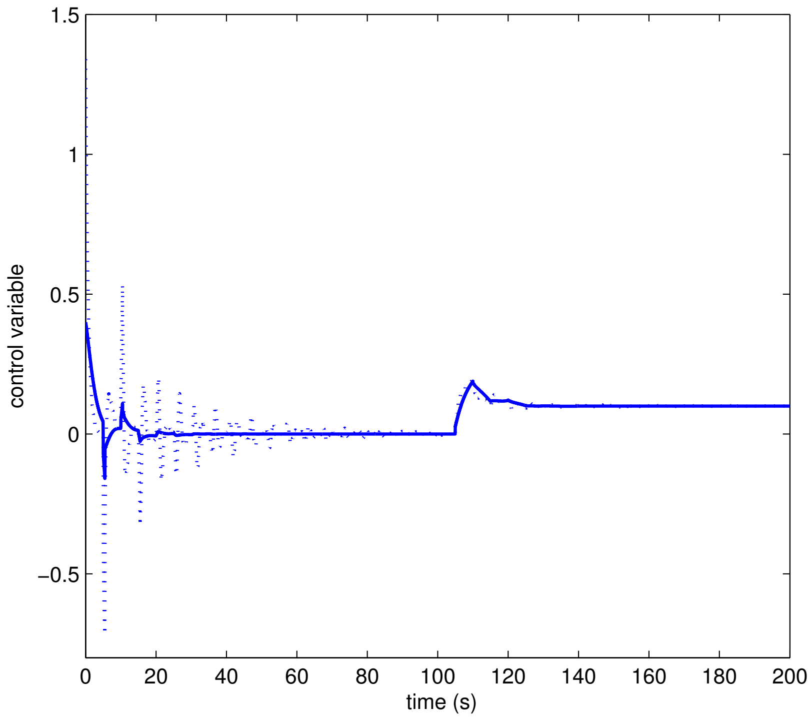


Figure 4.7: Control variables for perturbed system responses for Example 1: $(-)\ t_r = 5$ and $(...)\ t_r = 3$

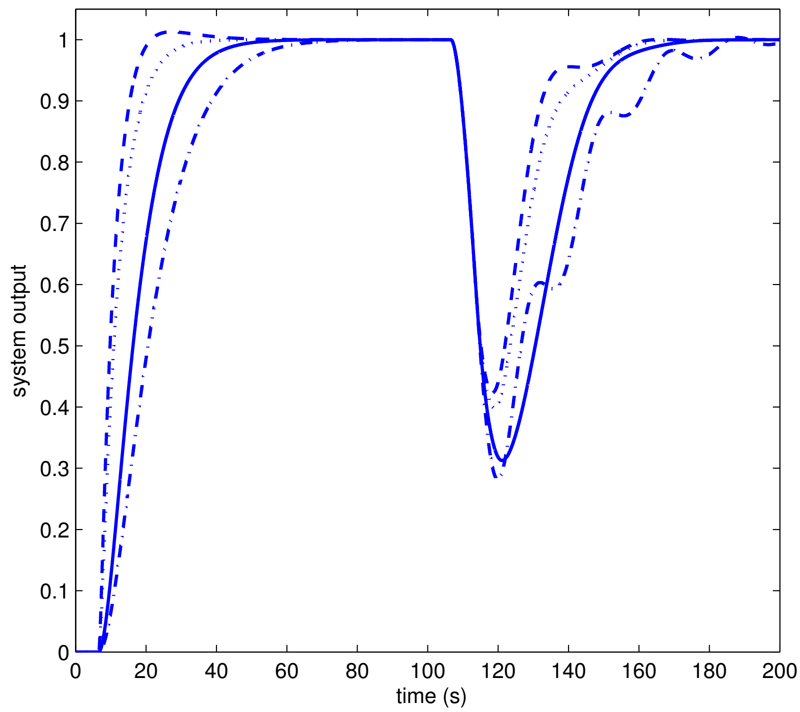


Figure 4.8: Nominal responses for Example 2: (−) $\bar{\beta} = 5$, (...) $\bar{\beta} = 10$, (−−) $\bar{\beta} = 15$ and (−.−) Rao et al. [11]

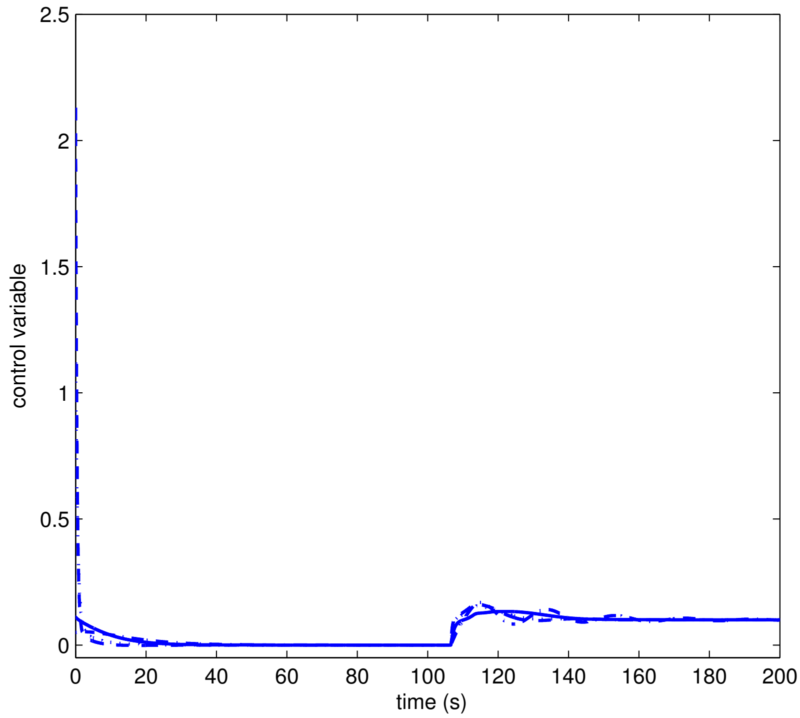


Figure 4.9: Control variables for nominal responses for Example 2: (−) $\bar{\beta} = 5$, (...) $\bar{\beta} = 10$, (−−) $\bar{\beta} = 15$ and (−.−) Rao et al. [11]

To show the effect of $\bar{\beta}$ on the robust stability and performance of the closed loop system, simulations have been done by giving a perturbation of 10% in the time delay and the corresponding step responses and the control signals for $\bar{\beta} = 5, 10$ and 15 are shown in Figs. 4.10 and 4.11, respectively. From Fig. 4.10, one can infer that the robust stability margin decreases and the control signal becomes oscillatory with increase in $\bar{\beta}$. Hence, $\bar{\beta}$ is taken as 5 in this work.

To show the robustness of the proposed method, the results are also compared with Rao et al.'s method and the corresponding step responses and the corresponding control signals are shown in Figs 4.12 and 4.13, respectively. It can be observed that the proposed method is robust towards perturbation in plant model parameters.

Example 3

A DIPTD plant model is considered to have the transfer function

$$G_p(s) = \frac{1}{s^2}e^{-5s} \tag{4.29}$$

The PID controller with $K_p = 0.245$, $T_i = 10.50$ and $T_d = 3.50$ is considered for setpoint tracking in [11]. The setpoint weighting parameter is chosen as 0.5. The PD controller considered for setpoint tracking is $0.008 + 0.13s$. Rao et al. have shown the superiority of their method over [69] and [76]. Hence, we compare the proposed method with Rao et al.'s method.

For a rise time $t_r = 7s$ and $\bar{\beta} = 5$, the controller parameters obtained are $K_p = 1.75$, $T_i = 2.80$ and $T_d = 2.0$. The PD parameters for various values of ψ are given in Table 4.3. Using these values the performance is evaluated by giving a unit step change in the setpoint and a negative step input of 0.01 in the load at $t = 60$ seconds, respectively. The step responses are shown in Fig. 4.14. Comparing the load disturbance responses, we can see that the peak value of the system output for $\psi = 40°$ is more whereas oscillatory step responses are obtained for $\psi = 50°$. Hence, ψ is taken as $45°$. The step responses and the control signals are shown in Figs. 4.15 and 4.16, respectively. For setpoint tracking the proposed method gives fast response with less settling time whereas both the methods give almost similar performances for load disturbance rejection. Fig. 4.17 shows the responses for a perturbation of 5% in process time delay. It can be observed that both the methods are robust towards the assumed perturbation in the plant model parameters.

4.5 Conclusions

In this work, it is shown that the three controllers of the modified Smith predictor proposed in the literature if tuned properly gives improved closed loop performances for a class of integrating

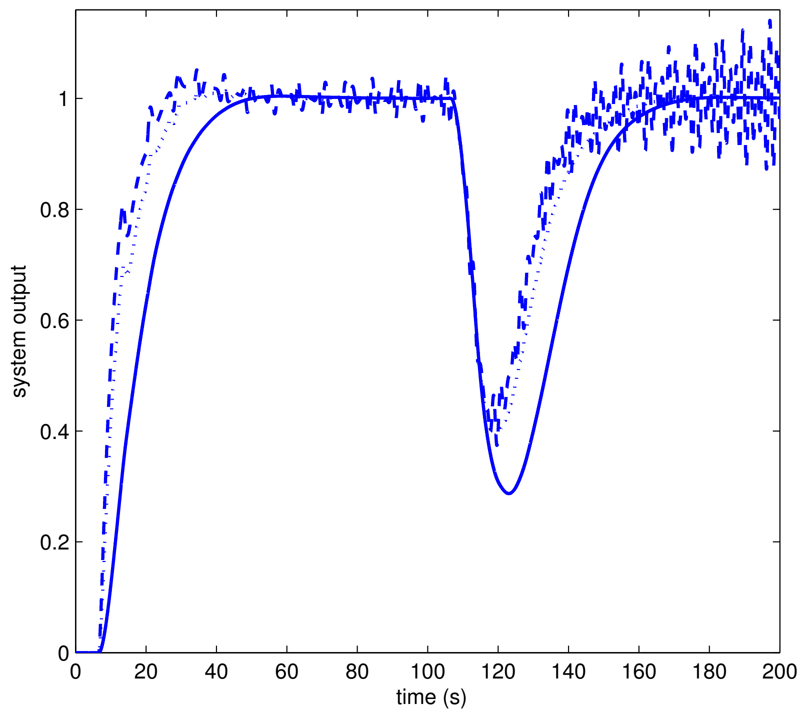


Figure 4.10: Perturbed responses for Example 2 : (–) $\bar{\beta} = 5$, (...) $\bar{\beta} = 10$ and (– –) $\bar{\beta} = 15$

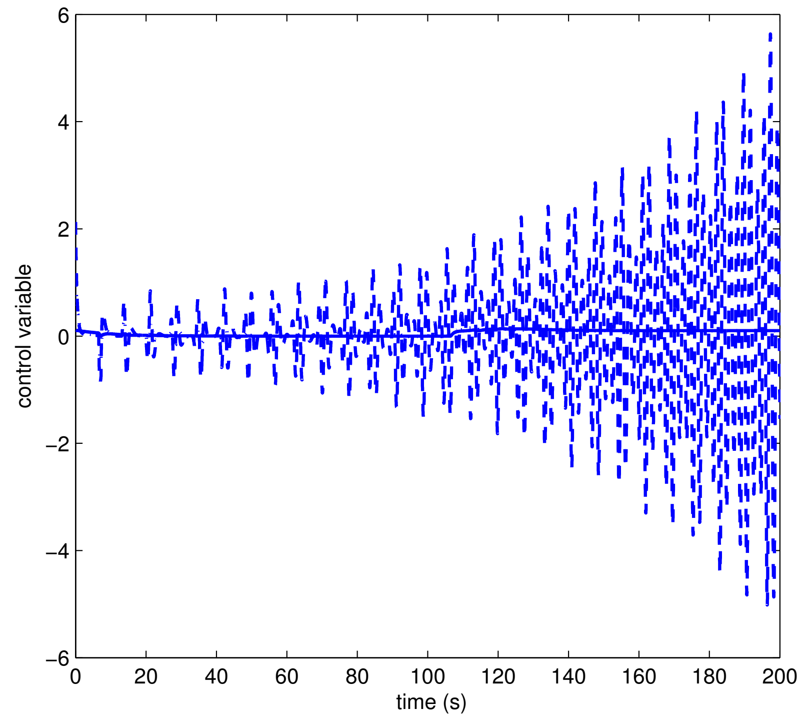


Figure 4.11: Control variables for perturbed responses for Example 2: (–) $\bar{\beta} = 5$, (...) $\bar{\beta} = 10$ and (– –) $\bar{\beta} = 15$

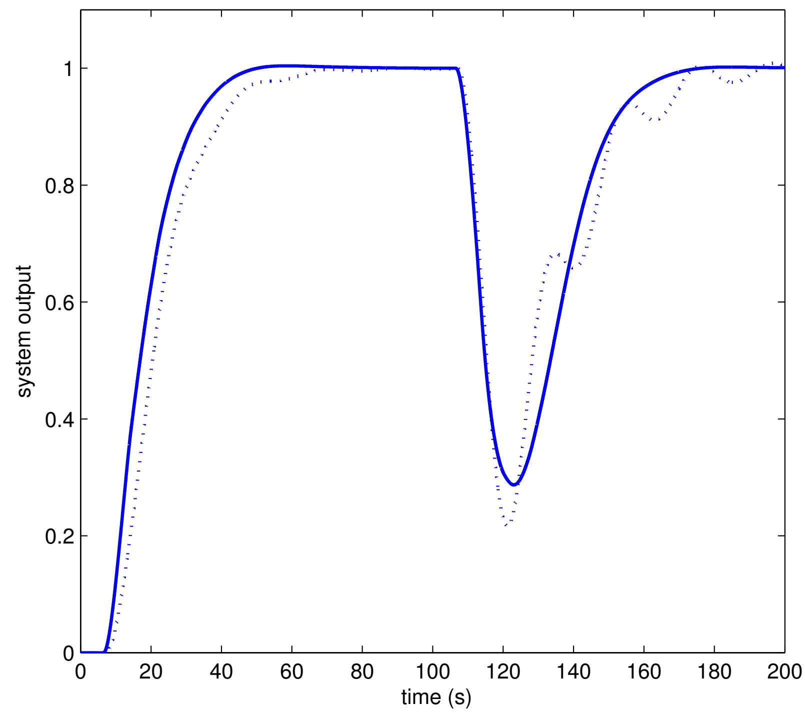


Figure 4.12: Perturbed responses for Example 2: (−) proposed and (...) Rao et al. [11]

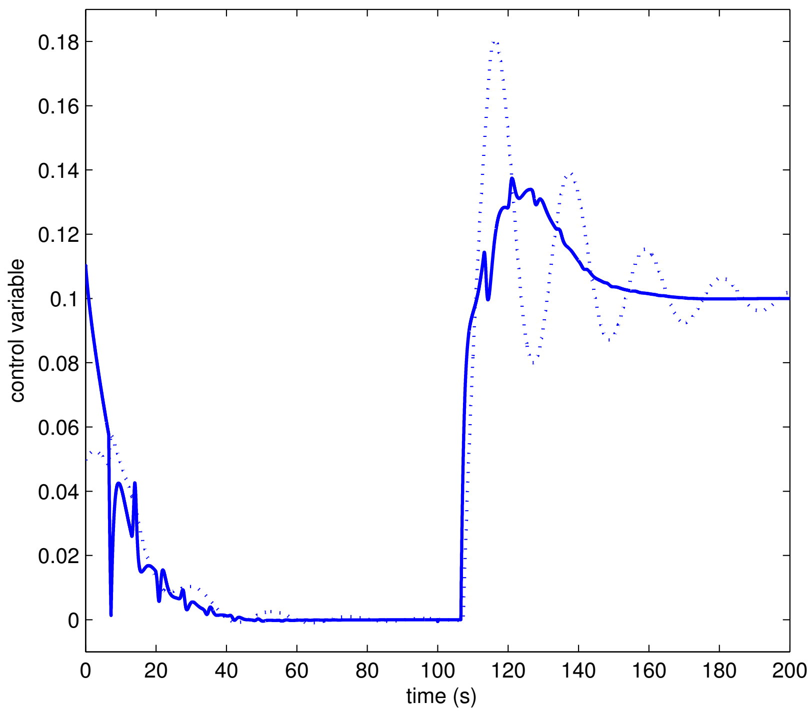


Figure 4.13: Control variables for perturbed responses for Example 2: (−) proposed and (...) Rao et al. [11]

Table 4.3: G_{c3} for Example 3 for various values of ψ

ψ	K_{p3}	T_{d3}
$40°$	0.01	14.15
$45°$	0.01	16.29
$50°$	0.01	18.96

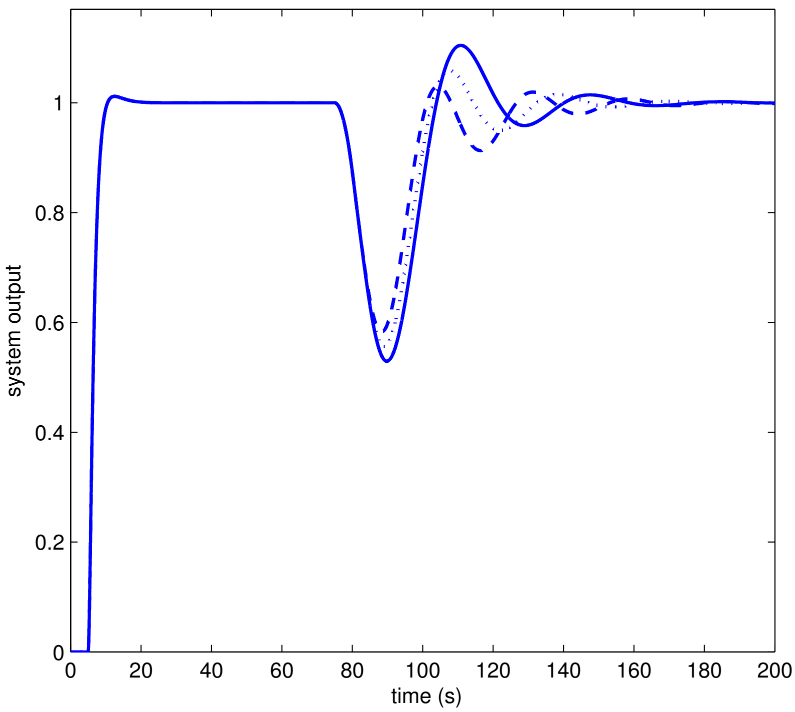


Figure 4.14: Nominal responses for Example 3: (−) $\psi = 40°$, (...) $\psi = 45°$ and (−−) $\psi = 50°$

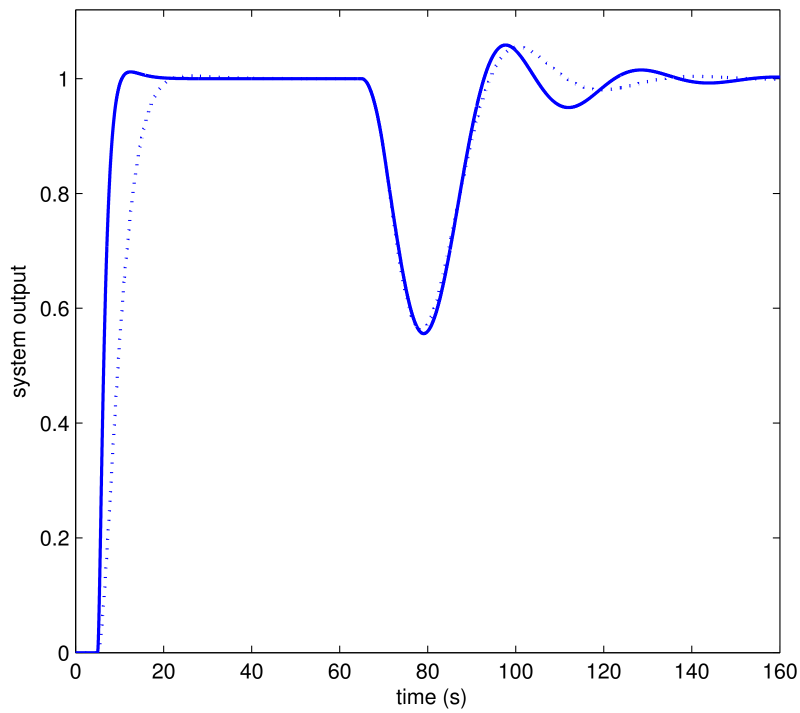


Figure 4.15: Nominal responses for Example 3: (—) proposed and (...) Rao et al. [11]

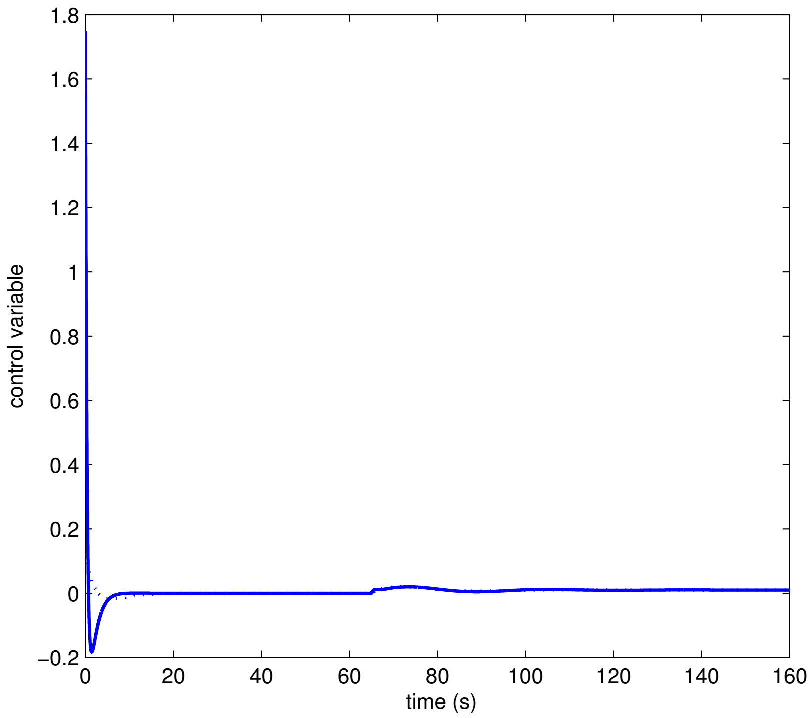


Figure 4.16: Control variables for nominal responses for Example 3: (—) proposed and (...) Rao et al. [11]

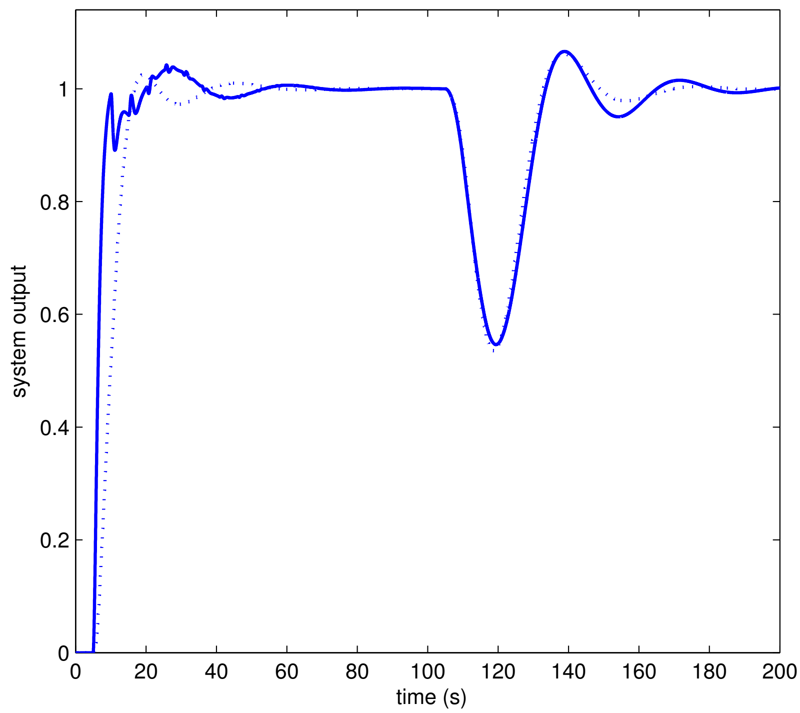


Figure 4.17: Perturbed responses for Example 3: (−) proposed and (...) Rao et al. [11]

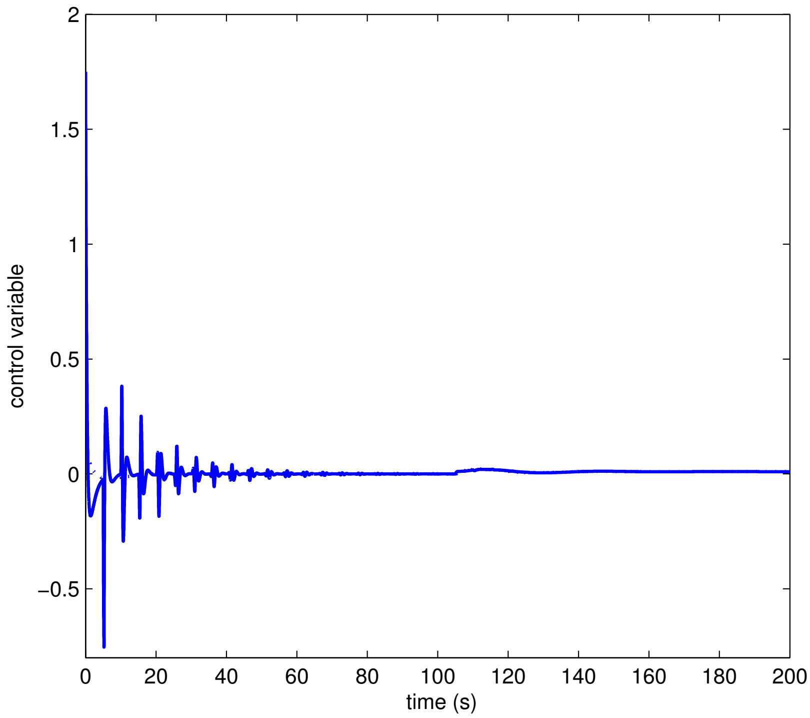


Figure 4.18: Control variables for perturbed responses for Example 3: (−) proposed and (...) Rao et al. [11]

processes as compared to some of the recently reported approaches in the literature. The controller for setpoint tracking is designed so that the delay free component of the system output follows the response of a second order plant, assuming a perfect matching between the actual plant and model in both the dynamics and time delay. The provided simple tuning formulae have physically meaningful parameters namely, the damping ratio and the natural frequency and the values of the natural frequency are obtained based on the user specified rise time. The PD controller considered for load disturbance rejection is tuned to achieve a user specified slope of the Nyquist curve at gain crossover frequency. A proportional controller with unity gain is considered for stabilizing the integrating process models and simulation examples are given to demonstrate the value of the proposed control method.

Chapter 5

Controller Tuning Based on Percentage Overshoot Specification

5.1 Introduction

In a PID controller, the P action (mode) adjusts the controller output according to the size of the error. The I action (mode) eliminates the steady state offset and the future trend is anticipated via the D action (mode). These three functions are sufficient for a large number of process applications and the transparency of the features leads to wide acceptance by the users. Furthermore, the internal model control (IMC) framework leads to PID controllers for virtually all the transfer function models common in industrial practice. Also, the PID controller is a building block in control of complex systems where thousands of PID controllers may by connected through cascade control, feed-forward control, ratio control, selectors, limiters, etc. to perform the overall control of the plant. The PID controller is often used when a top-down approach is taken. Most Model-Predictive Controllers (MPCs), for example, are used only at a higher level and deliver setpoints to PID controllers that perform the low-level control. A number of strategies for obtaining the tuning constants of a PID controller have been therefore proposed in the literature [20].

Lennartson and Kristiansson [77] have proposed an optimization based approach to tune PI/PID controllers with low pass filters on the derivative term to reject the load disturbances while optimizing the control effort and bounding the sensitivity functions. The maximum sensitivity and the corresponding frequency were used as the design parameters to tune a PI controller in [78]. Åström and co-workers [3, 44] have obtained the parameters of the setpoint weighted PI/ PID controllers by a numerical method based on optimization of load disturbance rejection with constraints on maximum sensitivity (M_s) and / or complementary sensitivity. The system output corresponding to two values of the maximum sensitivity (1.4 and 2.0) were investigated and it was observed that $M_s = 1.4$ gives smooth closed loop step responses with little or no overshoot whereas the step responses corresponding to $M_s = 2.0$ are faster with better load disturbance rejection. As fast response is oscillatory with large overshoot, $M_s = 1.4$ was recommended as a sufficient condition for controller tuning. The classical gradient based search algorithm used for optimization in [3] needs a good initial guess and a suitable search interval for convergence and hence, a PI controller has to be first designed to get the initial conditions.

A large percentage overshoot is not a problem in several chemical processes, provided that the system returns rapidly to the neighborhood of the steady state value. However, it is desirable to have no overshoot at all in control applications such as: manufacture of plastic gloves, PCR analysis etc. Thus, attempts have been made to propose control startegies such that the response of the compensated system has overshoot close to a prescribed value [12, 15, 79]. In this chapter, the ideas of [3, 44] are explored further by studying the step responses over a wide range of M_s for stable FOPTD and IPTD process models. It is observed that the PD controller gives improved setpoint responses for IPTD process model. However, as the PD controller fails to reject the load disturbances, PID settings are also provided for satisfactory load disturbance rejection. Analytical expressions correlating the controller parameters to percentage overshoot and plant model parameters are presented for stable FOPTD and IPTD process models, respectively.

The IMC based approach for controller design proposed in [80] has gained widespread acceptance because a clear tradeoff between the closed loop performance and robustness is achieved by a single tuning parameter namely the IMC filter time constant (λ). Several methods for obtaining λ have been proposed in the literature [5, 14, 81]. The IMC filter parameter was obtained by minimizing a weighted function of integral square error and the maximum of the complementary sensitivity function in [81]. Chen et al. [14] have proposed PI/PID tuning formula for stable first and second order plus time delay process models based on the IMC principle to achieve a $+2$ dB maximum closed loop amplitude ratio. In a recent article, Skogestad [82] reports that there are two approaches for controller design: tight control and smooth control. The fastest possible control with acceptable robustness is achieved by tight control whereas smooth control results in the slowest possible control subject to achieval of acceptable disturbance rejection. In this work, the normalized IMC filter time constant for smooth and tight control (where smooth and tight control corresponds to step response with 0 and 10 percent overshoot) for stable FOPTD and SOPTD process models are presented.

The chapter is organized as follows. The controller design for stable FOPTD and IPTD process models is presented in section 5.2. Simulation results are discussed in section 5.3. The normalized IMC filter time constant for smooth and tight control is derived in Section 5.4. A comparison of the closed loop achievable performances by PI/PID controller is given in section 5.5. Finally, conclusions are drawn in section 5.6.

5.2 Controller Design

In this section, analytical expressions are derived for the controller parameters in terms of plant model parameters for stable FOPTD and IPTD process models. Consider the model transfer function

$$G_p(s) = \frac{Ke^{-\theta s}}{Ts+1} \tag{5.1}$$

A stable first order plus time delay model represented by (5.1) is considered as this transfer function model can represent the dynamics of many stable industrial processes. An IPTD plant model which is generally related to level control problems and is frequently encountered in the processing industries is considered next.

5.2.1 Stable FOPTD Process Model

The controller design for a stable FOPTD process model is considered in this section. The general series form of a PID controller is given by

$$G_c(s) = K_p\left(1+\frac{1}{sT_i}\right)\left(\frac{1+sT_d}{1+\alpha sT_d}\right) \tag{5.2}$$

The derivative filter term of (5.2) is neglected for ease in the following analysis. The loop transfer function $L(s)$ is given by

$$\begin{aligned} L(s) &= G_p(s)G_c(s) \\ &= \frac{KK_pe^{-\theta s}(1+sT_d)(1+sT_i)}{sT_i(1+sT)} \end{aligned} \tag{5.3}$$

Replacing the delay term by a 1/1 Padé approximation, the sensitivity function can be written as

$$
\begin{aligned}
S(s) &= \frac{1}{1+L(s)} \\
&= \frac{\theta T T_i s^3 + \theta T_i s^2 + 2TT_i s^2 + 2T_i s}{\begin{array}{c}[\theta T T_i s^3 + \theta T_i s^2 + 2TT_i s^2 + 2T_i s + KK_p(2 + 2T_d s + 2T_i s \\ -\theta s + 2T_i T_d s^2 - \theta T_d s^2 - \theta T_i s^2 - \theta T_i T_d s^3)]\end{array}}
\end{aligned}
\tag{5.4}
$$

Setting $T_i = T$, the sensitivity function in the frequency domain becomes

$$
S(j\omega) = \frac{(-\theta T_i \omega^2 + j2T_i\omega)}{(-\theta T_i\omega^2 + KK_p\theta T_d\omega^2 + j2KK_pT_d\omega - jKK_p\theta\omega + j2T_i\omega + 2KK_p)} \tag{5.5}
$$

Squaring the magnitude of both sides of (5.5) and simplifying gives

$$
\theta^2(a^2 - v^2T_i^2)\omega^4 + (4T_i^2c^2 + K^2K_p^2b^2 + 4T_iKK_pb + 4KK_pa\theta)\omega^2 + 4K^2K_p^2 = 0 \tag{5.6}
$$

where $a = KK_pT_d - T_i$, $b = 2T_d - \theta$ and $c = \sqrt{1-v^2}$. $v = \frac{1}{M_s^*} = \frac{1}{\|S(jw)\|_\infty}$ is the inverse of the user specified maximum sensitivity value. For v to be the minimum distance from the critical operating point to the Nyquist curve, the Nyquist curve should lie outside the circle with center $(-1, 0)$ and radius v and should touch this circle at only one frequency. This is evident from the various Nyquist plots given in [44]. Therefore, using the condition that ω^2 should have repeated roots gives

$$
\begin{aligned}
4T_i^2c^2 + K^2K_p^2b^2 + 4T_iKK_pb + 4KK_p\theta(KK_pT_d - T_i) &= 4KK_p\theta\sqrt{a^2 - v^2T_i^2} \\
&\approx 4KK_p\theta(KK_pT_d - T_ic)
\end{aligned}
\tag{5.7}
$$

Simplification of (5.7) yields

$$
K^2K_p^2b^2 + 4KK_p(bT_i - \theta T_i + \theta cT_i) + 4c^2T_i^2 = 0 \tag{5.8}
$$

5.2.2 Selection of T_d

To explain the procedure involved in selecting T_d, Equation (5.8) is solved for $M_s = 1.30, 1.35$ and 1.40 for $T_d = \gamma\theta$ (where γ is the fraction by which the plant time delay is multiplied to obtain the derivative time constant) for the plant model given by (5.1) with K, T, θ assumed as unity. The results are given in Table 5.1.

Table 5.1: K_p for various γ for $M_s = 1.30, 1.35, 1.40$

γ	K_p		
	$M_s = 1.30$	$M_s = 1.35$	$M_s = 1.40$
0.10	0.37	0.42	**0.48**
0.15	0.40	0.46	0.52
0.20	0.44	0.51	0.58
0.25	0.49	**0.56**	0.64
0.30	**0.55**	0.64	0.73
0.35	0.63	0.74	0.84

As in this work, the controller parameters are obtained to achieve user defined percent overshoot, the values of K_p that result in no overshoot are shown in bold in Table 5.1. The corresponding step responses are shown in Fig. 5.1. It can be observed that K_p corresponding to $T_d = 0.25\theta$ and $M_s = 1.35$ gives the best response with least settling time and hence, T_d is taken as $\theta/4$.

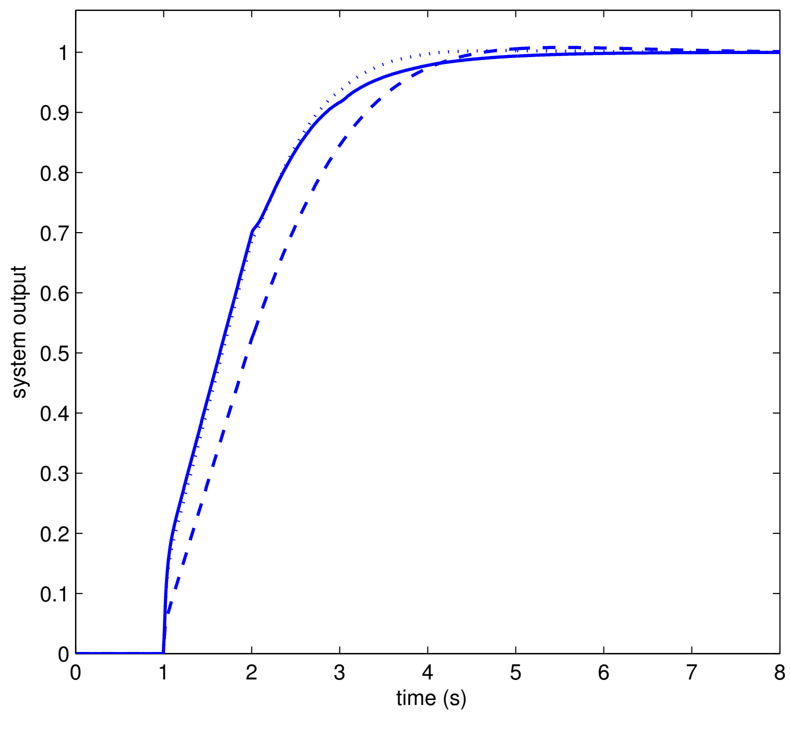


Figure 5.1: Setpoint responses for $\frac{e^{-s}}{s+1}$: (—) $M_s = 1.30$, (...) $M_s = 1.35$ and (−−) $M_s = 1.40$

The solution of (5.8) for the above value of the derivative time constant gives

$$q = KK_p\frac{\theta}{T} = 4(-2c + 3 - \sqrt{3(c^2 - 4c + 3)}) \tag{5.9}$$

Note that the percentage overshoot ($\%os$) remains the same for a particular value of v for a wide range of θ/T. The desired maximum sensitivity (M_s^*) is varied from 1.10 to 1.55 in steps of 0.05 and the values of q and the corresponding percentage overshoots are given in the third and fourth columns of Table 5.2. M_s represents the true maximum sensitivity corresponding to the obtained set of controller parameters and is calculated using the robust control toolbox of MATLAB. The difference between M_s and M_s^* is because of the Padé approximation of the time delay term. Analytical expressions relating M_s and q to $\%os$ obtained using the curve fitting toolbox of MATLAB are given by

$$M_s = 18.70 \times 10^{-6}(\%os)^3 - 12.97 \times 10^{-4}(\%os)^2 + 44.54 \times 10^{-3}(\%os) + 1.47 \quad (5.10)$$

$$q = 23.41 \times 10^{-6}(\%os)^3 - 10.13 \times 10^{-4}(\%os)^2 + 25.30 \times 10^{-3}(\%os) + 0.5579 \quad (5.11)$$

Table 5.2: M_s^*, M_s, controller parameters and percentage overshoots for stable FOPTD and IPTD process models

M_s^*	M_s	FOPTD (PID)		IPTD (PD)	
		q	$\%os$	$KK_p\theta$	$\%os$
1.10	1.10	0.16	0	0.16	0
1.15	1.20	0.25	0	0.25	0
1.20	1.30	0.33	0	0.33	0
1.25	1.30	0.41	0	0.41	0
1.30	1.40	0.49	0	0.49	0
1.35	1.50	0.57	0.51	0.57	0
1.40	1.60	0.64	3.70	0.64	2.55
1.45	1.80	0.72	9.06	0.72	7.89
1.50	1.90	0.79	15.09	0.79	13.93
1.55	2.00	0.85	20.35	0.85	19.17

Table 5.3: PID tuning rules for stable FOPTD process model

$\%os$	M_s	q	T_i	T_d
0	1.47	0.56	T	$\theta/4$
5	1.66	0.66	T	$\theta/4$
10	1.8	0.73	T	$\theta/4$

The values of q and M_s for 0, 5 and 10 percent overshoots are given in Table 5.3. The PID parameters for a stable FOPTD process model can be therefore obtained as follows:

- Assume $T_i = T$ and $T_d = \theta/4$
- Calculate K_p corresponding to any user specified percentage overshoot using (5.11) since $q = KK_p\theta/T$ as given in (5.9)

5.2.3 IPTD Process Model

The PD controller considered for setpoint tracking is given by

$$G_c(s) = K_p(1 + \frac{sT_d}{\alpha s T_d + 1}) \tag{5.12}$$

Neglecting the derivative filter constant, the loop transfer function becomes

$$L(s) = \frac{KK_p e^{-\theta s}(1 + sT_d)}{s} \tag{5.13}$$

The loop transfer function for a FOPTD plant subjected to a feed forward series PID controller is same as (5.13) for $T_i = T = 1$. Therefore, incorporating this condition in the analysis of previous section gives

$$K^2K_p^2b^2 + 4KK_p(b - \theta + \theta c) + 4c^2 = 0 \tag{5.14}$$

The derivative time constant (T_d) is taken as $\theta/4$ and the proportional gain is obtained by solving (5.14) for a given v. Once again, it is observed that the percentage overshoot remains constant for a particular value of the maximum sensitivity. $KK_p\theta$ for various values of v and the corresponding percentage overshoot are given in the last two columns of Table 5.2. The following equations correlate M_s and $KK_p\theta$ to $\%os$.

$$M_s = 54.17 \times 10^{-6}(\%os)^3 - 24.37 \times 10^{-4}(\%os)^2 + 53.14 \times 10^{-3}(\%os) + 1.493 \tag{5.15}$$

$$KK_p\theta = 38.99 \times 10^{-6}(\%os)^3 - 14.73 \times 10^{-4}(\%os)^2 + 28.43 \times 10^{-3}(\%os) + 0.5721 \tag{5.16}$$

Table 5.4: PD settings for setpoint tracking

$\%os$	M_s	$KK_p\theta$	T_d
0	1.49	0.57	$\theta/4$
5	1.70	0.68	$\theta/4$
10	1.83	0.75	$\theta/4$

Table 5.5: PID settings for load disturbance rejection

$KK_p\theta$	T_i	T_d	M_s
0.9	1.7θ	0.5θ	3.3

The controller settings for setpoint tracking for 0, 5 and 10 percent overshoot are given in Table 5.4. The PD controller however fails to reject the load disturbances and hence the PID

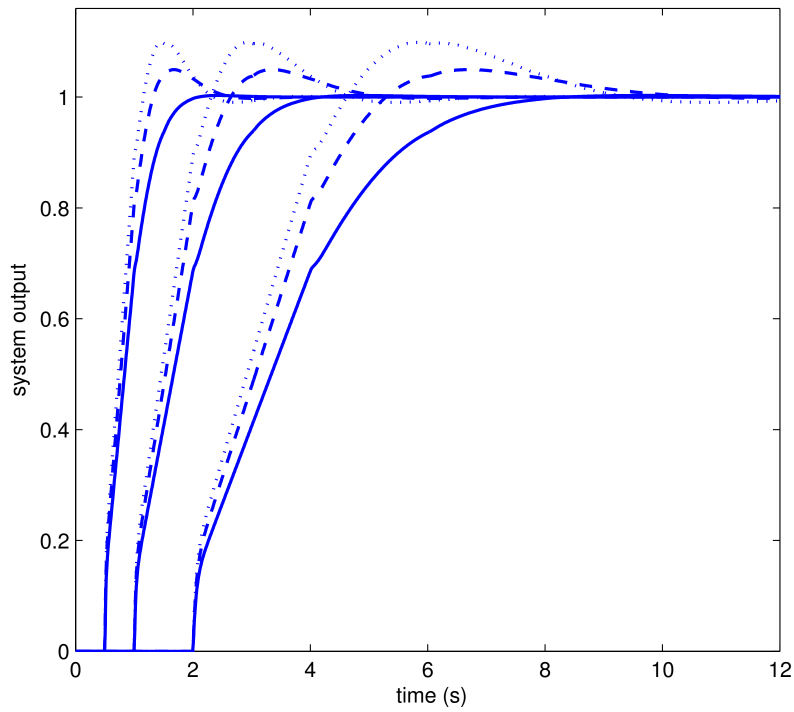


Figure 5.2: Setpoint responses for FOPTD process models for $\theta = 0.5, 1, 2$ for $0(-)$, $5(--)$ and $10(...)$ percent overshoots

settings of Table 5.5 found intuitively using the technique of subsection 5.2.1 are recommended for the overall satisfactory performance of the integrating plant models.

5.2.4 Properties of the Tuning Method

An interesting property of the proposed method is that it gives controller parameters corresponding to any user specified percentage overshoot. The FOPTD plant model represented by (5.1) is considered with $T = 1$ for various values of θ and the corresponding setpoint responses are shown in Fig. 5.2. The setpoint responses for integrating plant models are shown in Fig. 5.3. It can be observed from the step response plots that the percentage overshoots obtained are in close proximity to the specified values. Also, it can be observed that the proposed tuning methods give consistent performance over a wide range of θ/T and θ for stable FOPTD and IPTD process models, respectively.

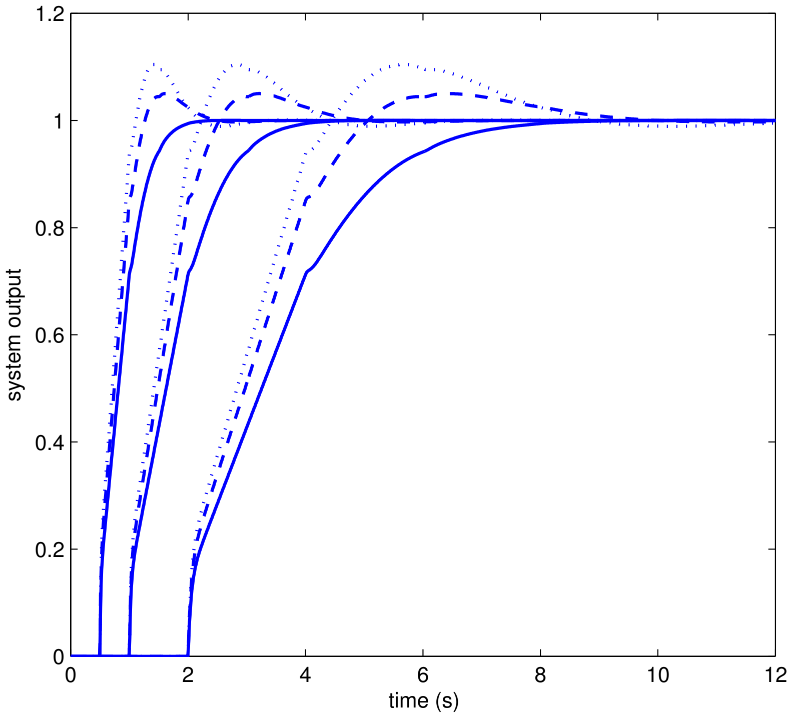


Figure 5.3: Setpoint responses for IPTD process models for $\theta = 0.5, 1, 2$ for $0(-)$, $5(--)$ and $10(...)$ percent overshoots

5.3 Simulation Study

In this section, few examples from the literature are considered to illustrate the usefulness of the proposed tuning rules. The first example presents the simulation results for a stable FOPTD plant with θ/T ratio of 1.5. A higher order plant is then considered to show that the tuning method has applicability for any stable process whose dynamics can be described by the FOPTD model. Finally, an IPTD plant model is considered to demonstrate the wide applicability of the proposed tuning rules.

5.3.1 Examples

Example 1

A stable FOPTD plant model with transfer function $G_p(s) = \frac{e^{-1.5s}}{s+1}$ is considered and the results are compared with Abbas's [12] method as both the methods give controller parameters such that the closed loop response of the system has a user defined percent overshoot. The controller parameters obtained by the proposed and Abbas's methods are given in Table 5.6. With these controller settings, the performance of the closed loop system is evaluated by giving a unit step input in the setpoint and a negative step input of 1 in the load at time $t = 20$ seconds. The simulation results are shown in Figs 5.4 and 5.5, respectively. From the step response plots, it can be observed that the proposed method gives 0, 4.95 and 10.31 percent overshoots whereas the percentage overshoot obtained by Abbas's method are 3.86, 10.16 and 13.83, respectively. This illustrates the usefulness of the proposed tuning rules.

Table 5.6: Controller parameters for Example 1

$\%os$	Proposed			Abbas [12]		
	K_p	T_i	T_d	K_p	T_i	T_d
0	0.37	1.00	0.37	0.77	1.75	0.43
5	0.44	1.00	0.37	0.84	1.75	0.43
10	0.49	1.00	0.37	0.88	1.75	0.43

To illustrate the robustness of the proposed method, a 30% perturbation is given in both the steady state gain and the plant delay and the corresponding step responses are shown in Fig. 5.6. It can be observed that the proposed method is more robust.

Example 2

The higher order plant considered in [83] is $G_p(s) = \frac{1.08}{(s+1)^2(2s+1)^3}e^{-10s}$ and the identified equivalent FOPTD plant model is $\hat{G}(s) = \frac{1.08}{4.41s+1}e^{-13.8146s}$. The plots of the system output are shown in Fig. 5.7 and it can be observed that the proposed tuning rules give satisfactory performance for the equivalent FOPTD transfer function model of the higher order process.

Example 3

The integrating plant model considered in [13] is $G_p(s) = \frac{0.0506}{s}e^{-6s}$. The controller parameters for setpoint tracking and load disturbance rejection are given in Tables 5.7 and 5.8, respectively. As Visioli's method gives 7.63% overshoot, the PD settings for setpoint tracking for no overshoot and 7.63 percent overshoot are given. The plots of the system output for a unit change in the step input for setpoint tracking and disturbance rejection are shown in Figs. 5.8 and 5.10, whilst the corresponding control signals are plotted in Figs. 5.9 and 5.11 respectively. Fig. 5.8 shows that the proposed method has the same speed of response as Visioli's method for same percent overshoot but the latter gives an oscillatory response. One point to be noted is that the proposed method gives the controller settings corresponding to any user specified value of percentage overshoot which is not the case with the method proposed in [13]. The load disturbance response plot shows that the proposed method has larger peak amplitude but less settling time as compared to Visioli's method. Furthermore, the control signal corresponding to Visioli's method exhibits significant oscillations before reaching the steady state which is not desirable in the context of process control.

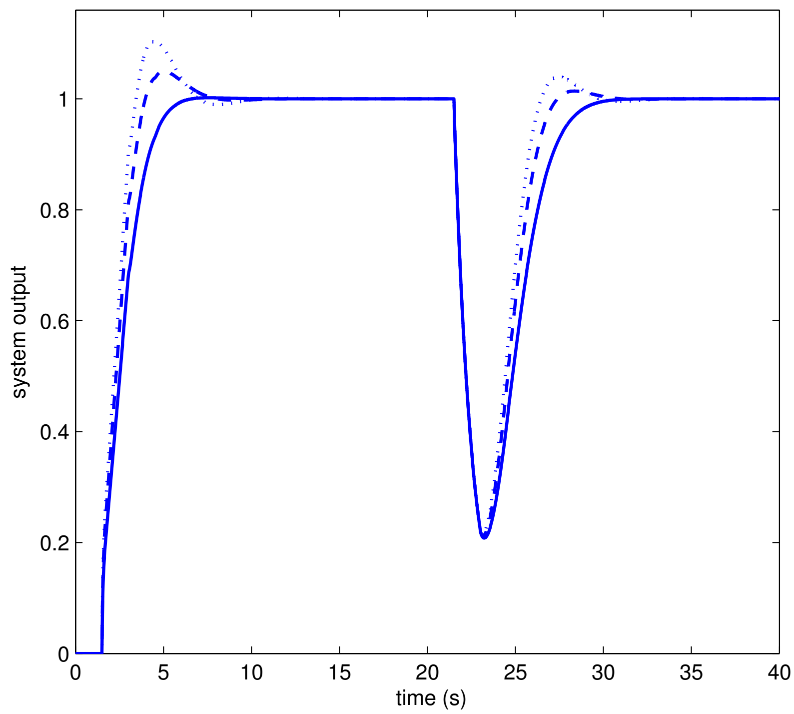


Figure 5.4: Step responses for Example 1 for 0, 5 and 10 percent overshoots by the proposed method

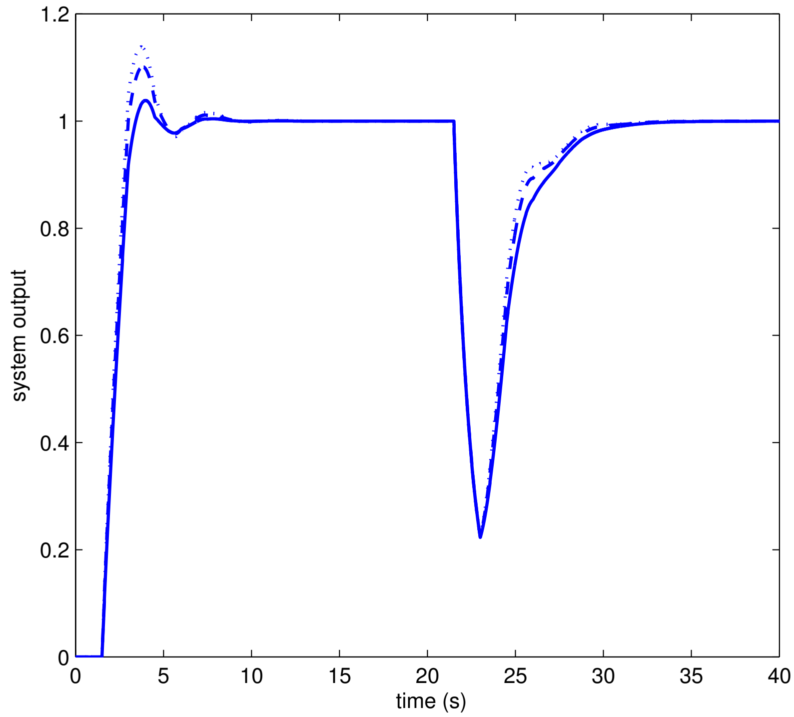


Figure 5.5: Step responses for Example 1 for 0, 5 and 10 percent overshoots by Abbas method [12]

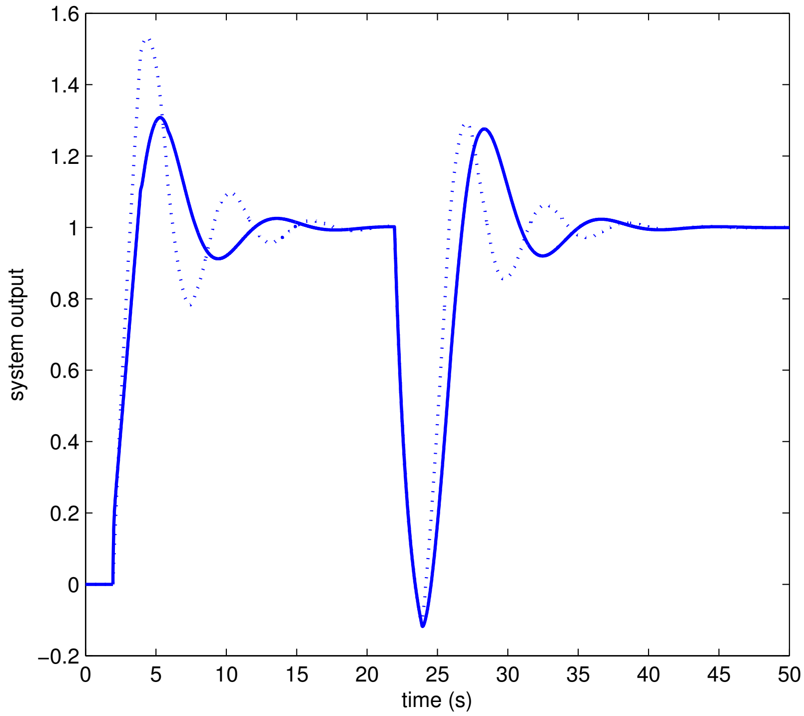


Figure 5.6: Robust responses for Example 1: (—) proposed and (...) Abbas [12]

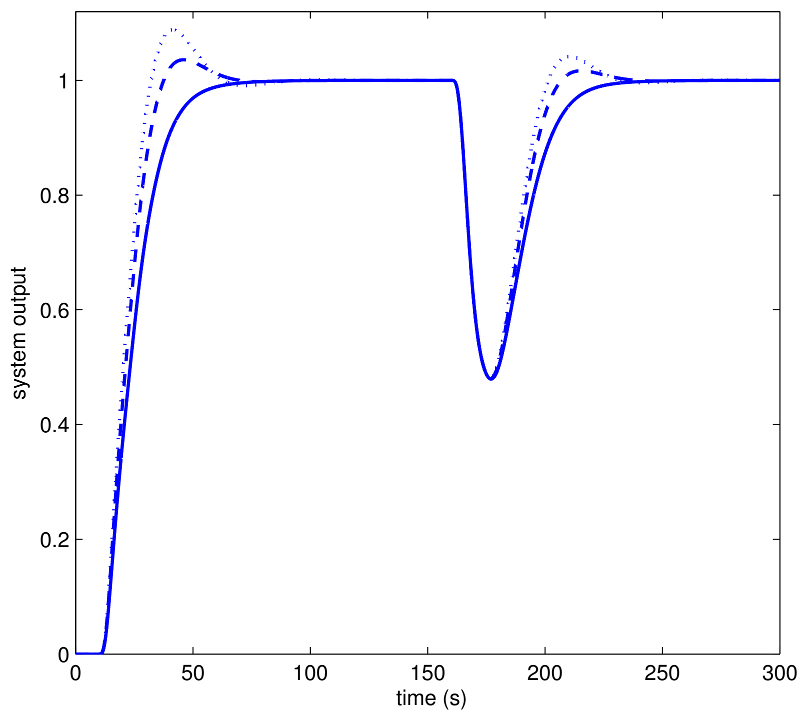


Figure 5.7: Step responses for Example 2 for 0, 5 and 10 percent overshoots

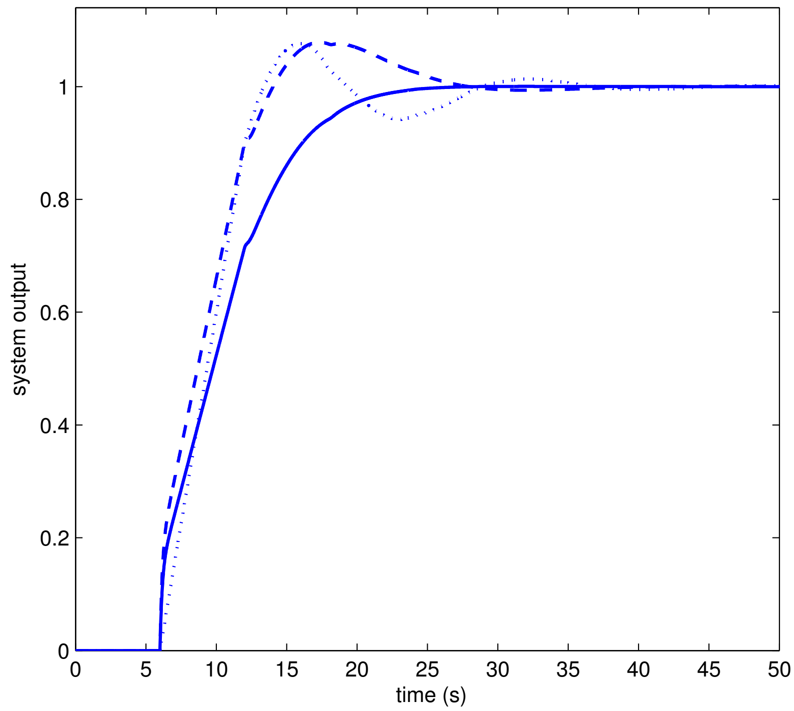


Figure 5.8: Setpoint responses for Example 3: proposed method for no (—) and 7.63 (−−) percent overshoots, (...) Visioli's method [13]

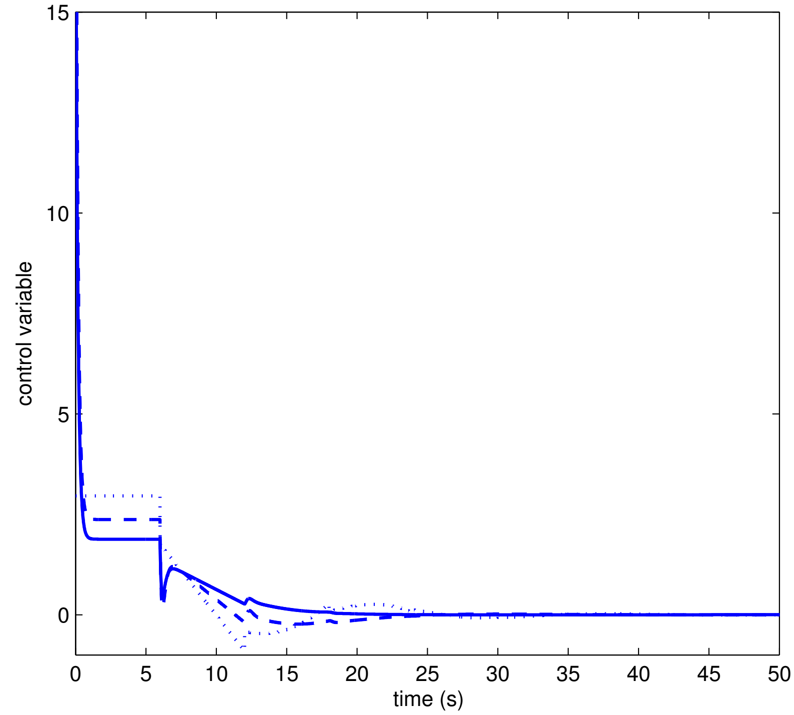


Figure 5.9: Control variables for setpoint responses for Example 3: proposed method for no (—) and 7.63 (−−) percent overshoots, (...) Visioli's method [13]

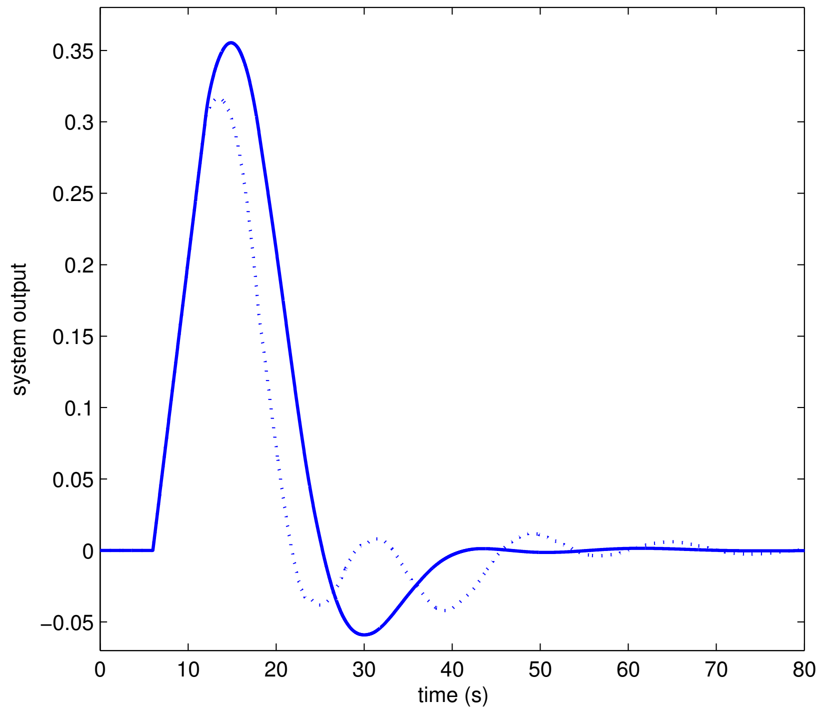


Figure 5.10: Load disturbance responses for Example 3: (−) proposed and (...) Visioli [13]

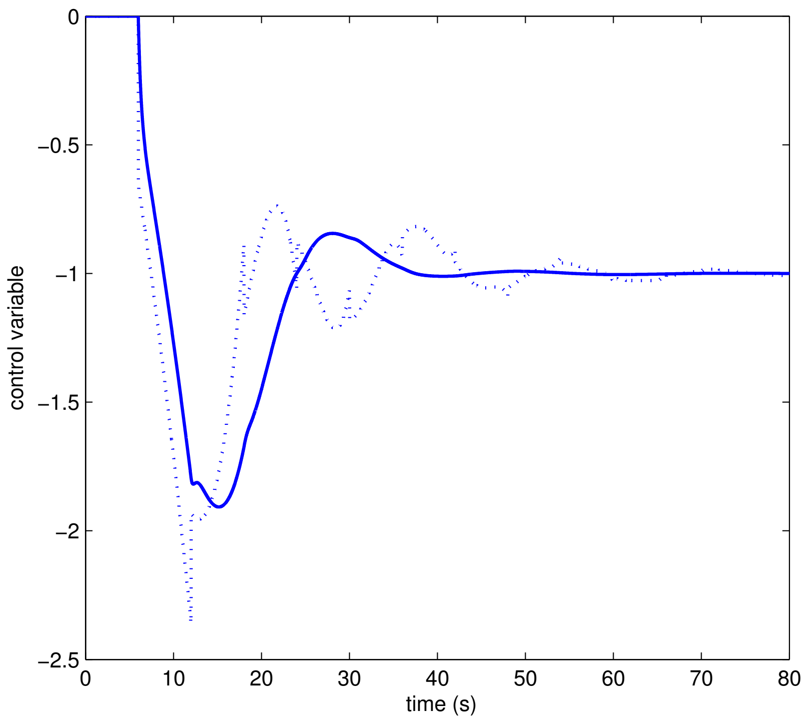


Figure 5.11: Control variables for load disturbance responses for Example 3: (−) proposed and (...) Visioli's method [13]

Table 5.7: Controller parameters for setpoint response for Example 3

	K_p	T_d	$\%os$
Proposed	1.88	1.50	0
	2.37	1.50	7.85
Visioli [13]	2.96	2.70	7.63

Table 5.8: Controller parameters for load disturbance rejection for Example 3

	K_p	T_i	T_d
Proposed	2.96	10.2	3
Visioli [13]	4.41	10.98	2.94

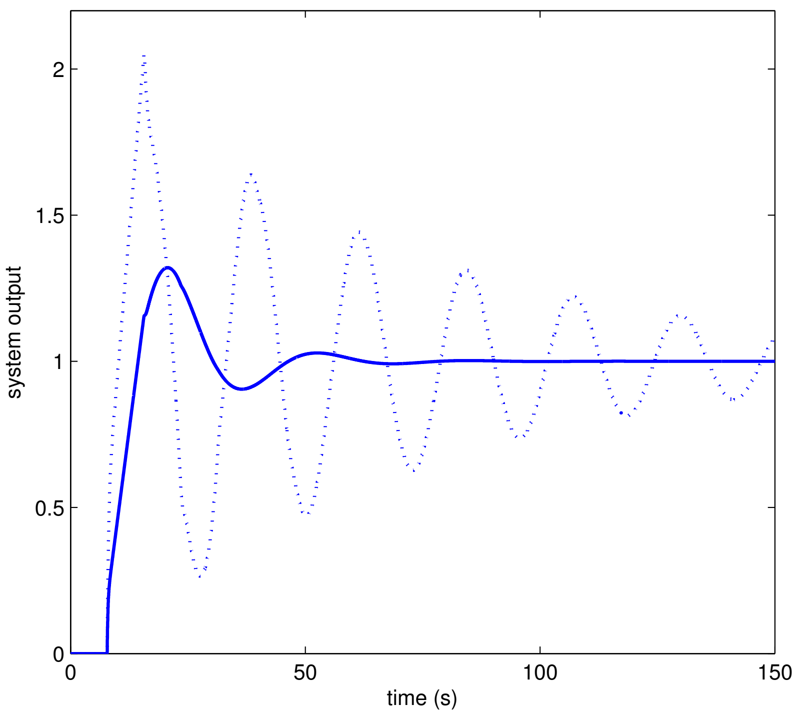


Figure 5.12: Robust responses for Example 3: (—) proposed and (...) Visioli's method [13]

Fig. 5.12 shows the step responses assuming a 30% perturbation in both the steady state gain and the plant delay. Visioli's PD parameters result in unstable responses whereas the proposed method is robust towards assumed parameter perturbations.

5.4 Controller Design in IMC Framework

The PI/PID settings for FOPTD and SOPTD process models in the IMC framework [14] are given in Table 5.9, where $\lambda_n = \lambda/\theta$ and the PI and PID controllers are of the form:

$$G_c(s) = K_p(1 + \frac{1}{sT_i}) \tag{5.17}$$

$$G_c(s) = K_p(1 + \frac{1}{sT_i} + sT_d) \tag{5.18}$$

Table 5.9: Controller parameters for various process models

$G_p(s)$	K_p	T_i	T_d
$\frac{Ke^{-\theta s}}{\tau s+1}$	$\frac{1}{K}\frac{\tau}{\lambda+\theta} = \frac{1}{\theta K}\frac{\tau}{\lambda_n+1}$	τ	
$\frac{Ke^{-\theta s}}{\tau^2 s^2+2\xi\tau s+1}$	$\frac{1}{K}\frac{2\xi\tau}{\lambda+\theta} = \frac{1}{\theta K}\frac{2\xi\tau}{\lambda_n+1}$	$2\xi\tau$	$\frac{\tau}{2\xi}$

Working on similar lines as in subsection 5.2.1 gives

$$\omega^4\theta^2 z^2 c^2 + \omega^2(4z^2c^2 + 8\theta z - \theta^2) - 4 = 0 \tag{5.19}$$

where $z = \lambda + \theta$. The condition that ω^2 should have repeated roots gives

$$4c^2\lambda^2 + \lambda(8c^2\theta + 4\theta c - 8\theta) + 4c\theta^2(1 + c) - 7\theta^2 = 0 \tag{5.20}$$

Dividing the above equation by θ^2, (5.20) becomes

$$4c^2\lambda_n^2 + \lambda_n(8c^2 + 4c - 8) - 7 + 4c + 4c^2 = 0 \tag{5.21}$$

The solution of (5.21) gives

$$\lambda_n = \frac{2 + 2\sqrt{1-c} - c - 2c^2}{2c^2} \tag{5.22}$$

The λ_n obtained by solving (5.21) for various values of c and the corresponding percentage overshoots are given in Table 5.10. Analytical expressions correlating M_s and λ_n to $\%os$ obtained using the curve fitting toolbox are

$$M_s = 32.79 \times 10^{-5}(\%os)^3 - 89.04 \times 10^{-4}(\%os)^2 + 88.75 \times 10^{-3}(\%os) + 1.385 \quad (5.23)$$

$$\lambda_n = -47.89 \times 10^{-5}(\%os)^3 + 0.0144(\%os)^2 - 0.1698(\%os) + 1.474 \quad (5.24)$$

Table 5.10: M_s^*, M_s, percentage overshoots and λ_n

M_s^*	M_s	$\%os$	λ_n
1.2	1.2	0	3.5568
1.3	1.3	0	2.1384
1.4	1.4	0.17	1.4458
1.5	1.6	3.48	1.0375
1.6	1.7	8.79	0.7691
1.7	1.8	14.51	0.5796

The normalized IMC filter parameter and the corresponding maximum sensitivity for smooth and tight control are obtained by calculating the respective values for 0 and 10 percent overshoot and are given in Table 5.11. The proposed tuning rules are given in Tables 5.12 and 5.13, respectively.

Table 5.11: M_s and λ_n for smooth and tight control

	M_s	λ_n
Smooth control	1.38	1.47
Tight control	1.71	0.74

Table 5.12: Proposed PI parameters

	K_p	T_i
Smooth control	$\frac{0.40\tau}{K\theta}$	τ
Tight control	$\frac{0.57\tau}{K\theta}$	τ

5.4.1 Limitations of IMC based Controller Design

As the IMC approach is based on pole zero cancelation, methods which incorporate IMC design principles result in good setpoint responses. However, the IMC settings result in a long settling time for the load disturbances for lag dominant processes. A good trade off between disturbance response and robustness is achieved by selecting $T_i = 8\theta$ for such class of processes [5]. The integral time constant is therefore taken as $T_i = min(T, 8\theta)$ and $T_i = min(2\xi\tau, 8\theta)$ for PI and PID

Table 5.13: Proposed PID parameters

	K_p	T_i	T_d
Smooth control	$\frac{0.81\xi\tau}{K\theta}$	$2\xi\tau$	$\frac{\tau}{2\xi}$
Tight control	$\frac{1.15\xi\tau}{K\theta}$	$2\xi\tau$	$\frac{\tau}{2\xi}$

controllers, respectively. Even though the modified integral settings degrades the setpoint performance, it is observed that the proposed controller settings give satisfactory performance for both setpoint tracking and load disturbance rejection for lag dominant processes also.

Extensive simulation results have shown that the closed loop performance of second order processes with $\theta\xi/\tau$ less than 0.6 can be improved by the following settings of the proportional gain of the PID controller.

$$\begin{aligned} K_p &= \frac{0.2}{K\theta}\frac{\xi\tau}{\lambda_n+1} && \text{if} \quad 0.01 < \frac{\theta\xi}{\tau} < 0.1 \\ K_p &= \frac{1.2}{K\theta}\frac{\xi\tau}{\lambda_n+1} && \text{if} \quad 0.1 \leq \frac{\theta\xi}{\tau} < 0.6 \end{aligned} \tag{5.25}$$

5.4.2 Simulation Study

The proposed tuning scheme is applied to FOPTD process models with various ratios of dead time to time constant and SOPTD processes to demonstrate the effectiveness of the proposed tuning method. The performance of the proposed controller is compared with the IMC based approach reported in [5]. The results are also compared with Wang et al.'s [1] method for SOPTD process models. The FOPTD process models considered in [84] are:

$$\begin{aligned} G_1(s) &= \frac{1}{s+1}e^{-s} \\ G_2(s) &= \frac{1}{s+1}e^{-5s} \\ G_3(s) &= \frac{100}{100s+1}e^{-s} \\ G_4(s) &= \frac{3}{100s+1}e^{-10s} \end{aligned} \tag{5.26}$$

The time constant and the plant delay are the same for $G_1(s)$ whereas $G_2(s)$ represents a delay dominated plant. $G_3(s)$ is a lag dominant process with θ/τ ratio of 0.01 and $G_4(s)$ represents the transfer function of an important viscosity loop in a polymerization process. Figs. 5.13 and

5.14 show the plant outputs for a unit change in the step input and a unit step change in the load disturbance. The controller tuned for smooth control gives plant output with no overshoot for $G_1(s)$ and $G_2(s)$ whereas settings for tight control yield faster and oscillatory output. The SIMC gives a response which is in between these two outputs as is evident from the step response plots. Unlike the SIMC, the proposed method can be used to obtain a family of responses starting from the smooth response to tight control. As $G_3(s)$ and $G_4(s)$ represent lag dominant processes, the IMC settings result in poor load disturbance rejection performance and hence, the integral time constant suggested in [5] is considered. It can be observed from Fig. 5.14 that this degrades the system setpoint performance and the overshoot is not zero for $G_3(s)$ and $G_4(s)$. However, the proposed controller gives satisfactory performance for both setpoint tracking and disturbance rejection even for these lag dominant processes. It can be thus concluded that the proposed tuning method can be used for a wide range of stable FOPTD process models. The robustness of the proposed tuning rules is evident from the step response plots shown in Fig. 5.15 assuming a 30% perturbation in θ and K.

The plant models considered for PID controller design are given by the following transfer functions:

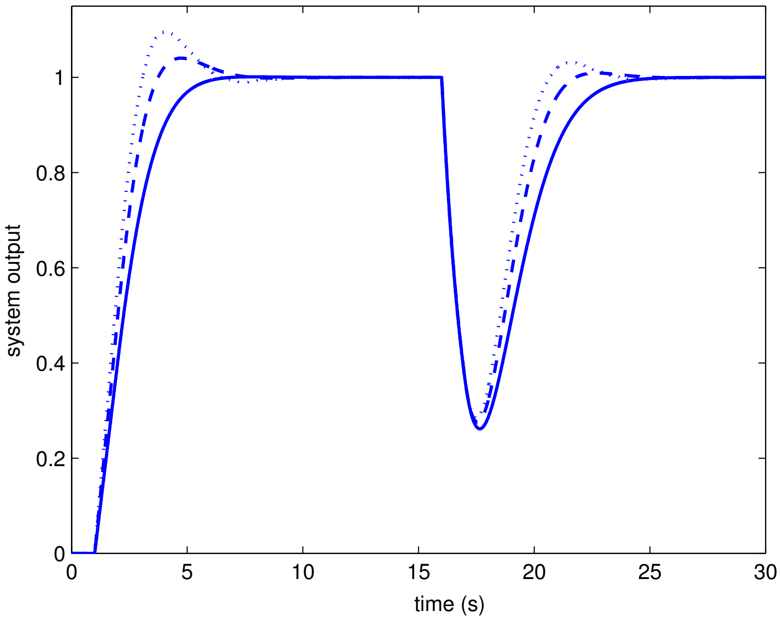
$$G_5(s) = \frac{1}{(s+1)(s+5)^2}e^{-0.5s} \tag{5.27}$$

$$G_6(s) = \frac{1}{(s^2+0.4s+1)}e^{-0.1s} \tag{5.28}$$

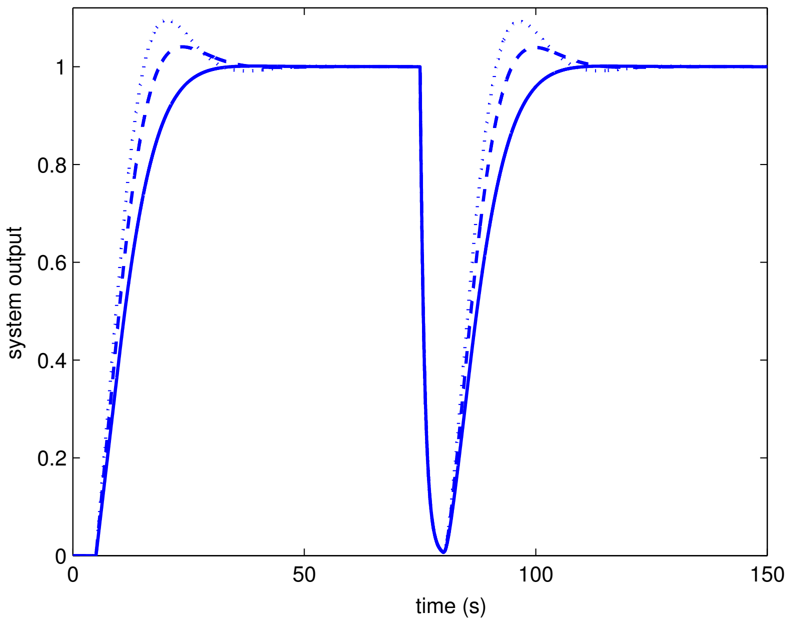
$$\hat{G}_5(s) = \frac{1}{7.724s^2+32.317s+25.220}e^{-0.606s} \tag{5.29}$$

$G_5(s)$ represents a higher order non oscillatory process and the equivalent second order plus time delay model is given by (5.29). The second example is an oscillatory plant with $\xi = 0.2$. The plant outputs are shown in Figs. 5.16 and 5.17, respectively. The proposed smooth controller gives no overshoot and satisfactory load disturbance rejection for $G_5(s)$. The controller tuned for tight control gives a faster response with less overshoot as compared to Wang et al.'s PID as is evident from Fig. 5.16. The load disturbance rejection performance is almost same for all the three methods. As the $\theta\xi/\tau$ ratio for $G_6(s)$ is 0.02, the modified proportional gain suggested in section 5.4 is used. Satisfactory performance is achieved by the proposed method whereas Wang et al.'s PID gives unacceptable closed loop step responses with large overshoot and long settling time.

The plant model $G_5(s)$ is considered again and the results are compared with Chen et al.'s [14] method to illustrate the robustness of the proposed tuning method. The controller settings

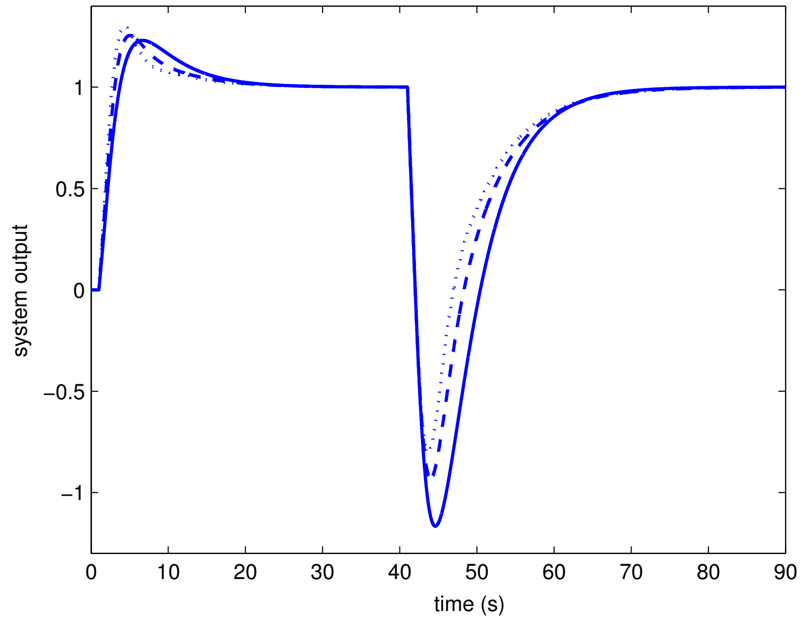


(a) $G_1(s)$

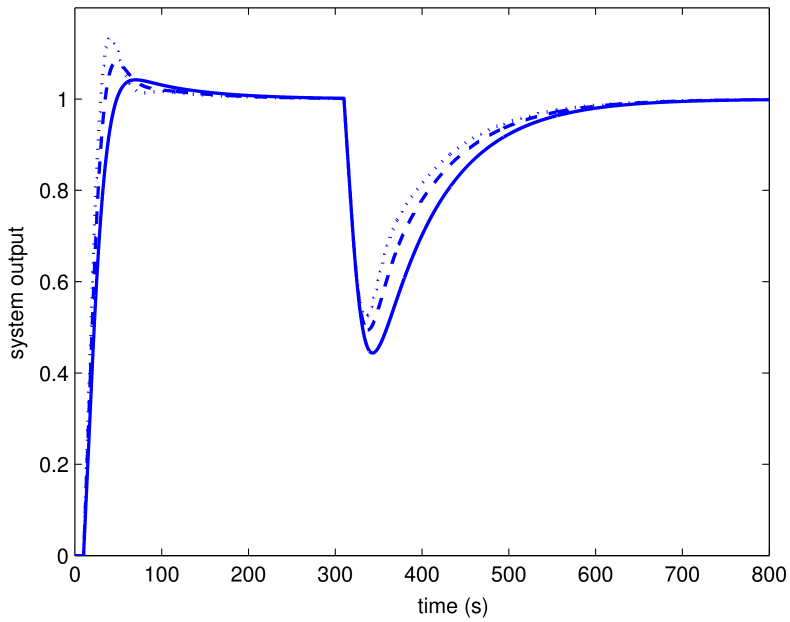


(b) $G_2(s)$

Figure 5.13: Step responses for $G_1(s)$ and $G_2(s)$: $(-)$ smooth control, (...) tight control, $(--)$ SIMC [5]



(a) $G_3(s)$



(b) $G_4(s)$

Figure 5.14: Step responses for $G_3(s)$ and $G_4(s)$: (−) smooth control, (...) tight control, (−−) SIMC [5]

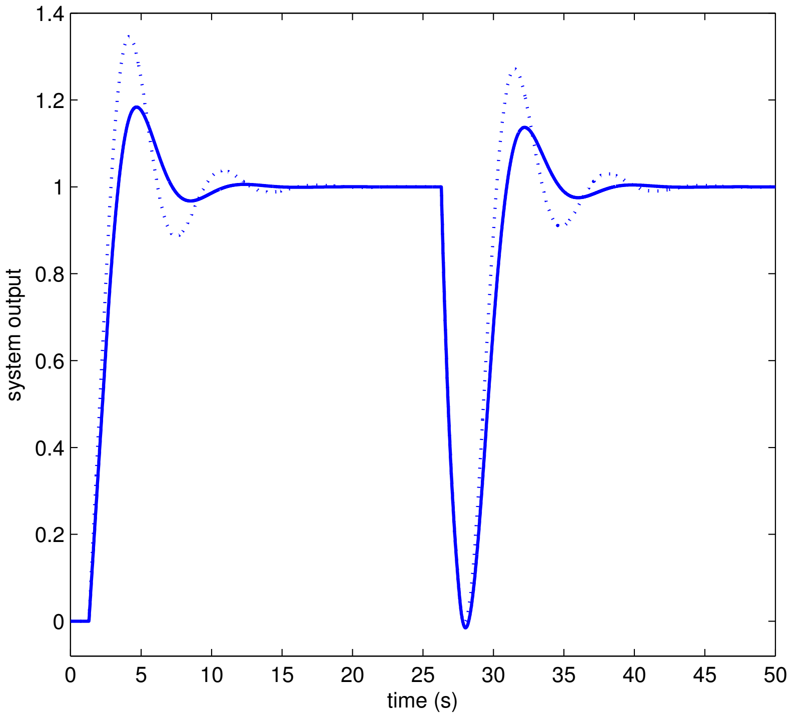


Figure 5.15: Robust responses for $G_1(s)$: (−) smooth control and (...) SIMC [5]

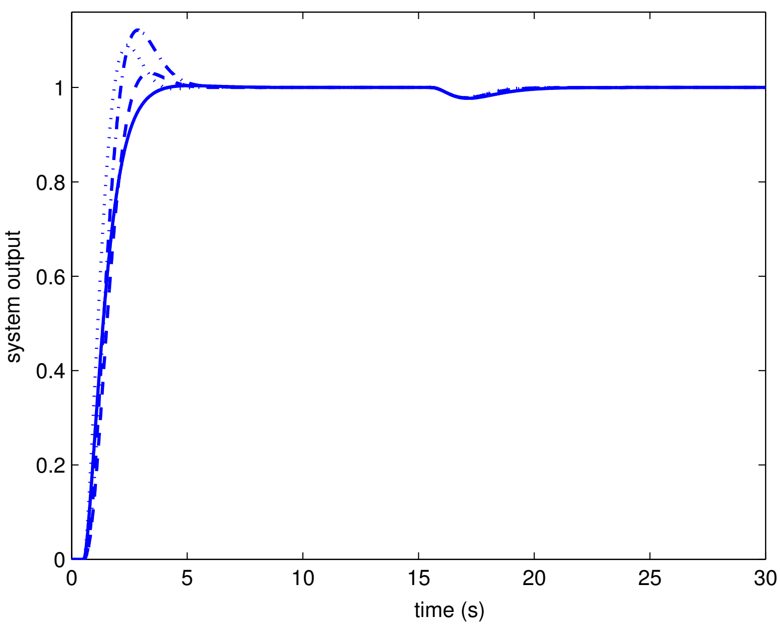


Figure 5.16: Step responses for $G_5(s)$: (−) smooth control, (...) tight control, (−−) SIMC [5], (−.−) Wang et al. [1]

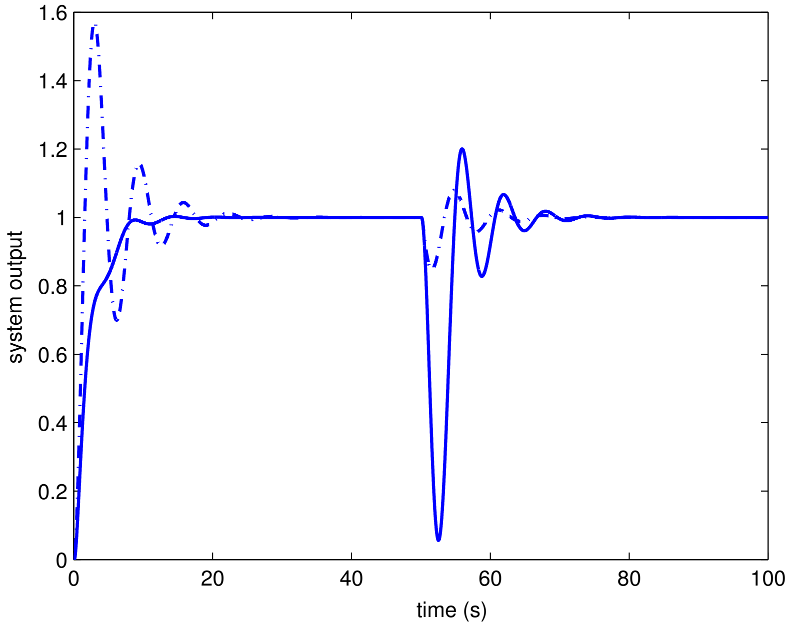


Figure 5.17: Step responses for $G_6(s)$: $(-)$ smooth control, $(-.-)$ Wang et al. [1]

corresponding to gain and phase margins of 3.14 and 61.4 °(as these are the best settings reported in [14] from robustness point of view) are considered in this work. The step responses for the nominal plant model and an assumed 20% uncertainty in both the steady state gain and the plant delay are shown in Fig. 5.18. It can be observed that the robustness of the proposed method towards assumed parameter perturbations are quite satisfactory.

Finally, the thermoelectric device TB-127-1.4-1.2 (Kyrotherm) which is used in biomedical thermocycler for PCR analysis and does not tolerates any overshoot in the output response is considered. A simplified model of the device is given by

$$G_7(s) = \frac{7}{100s+1}e^{-16s} \tag{5.30}$$

The controller parameters proposed in [15] are $K_p = 0.31$ and $T_i = 100$ whereas the proposed method gives $K_p = 0.36$ and $T_i = 100$. The performance of the closed loop system is evaluated using these controller settings by giving a unit step input in the setpoint and a negative step input of 0.50 at t = 200 seconds. It can be observed from the system outputs shown in Fig. 5.19 that the proposed method results in superior performance for both the setpoint tracking and load disturbance rejection. Also, exactly 5 percent overshoot is obtained by the proposed method whereas Mnif's method gives 7% overshoot, thereby proving the accuracy of the proposed tuning formulas.

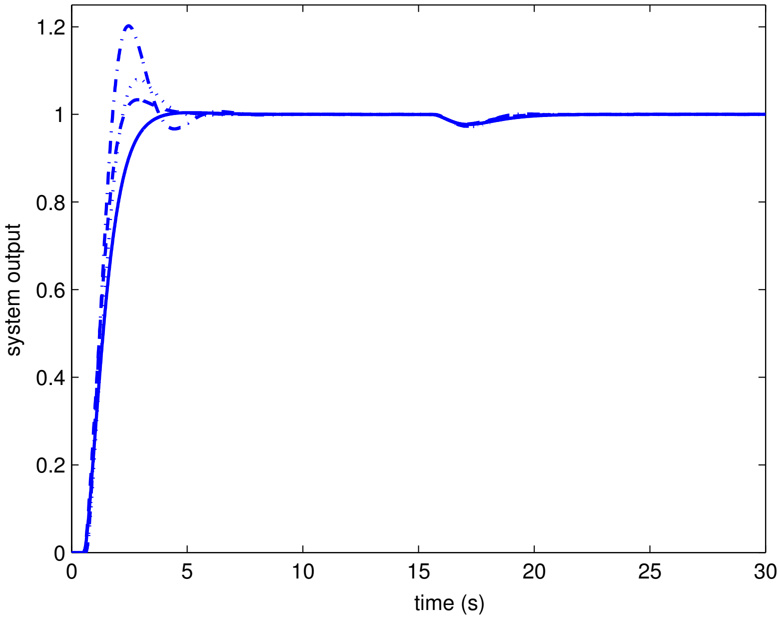


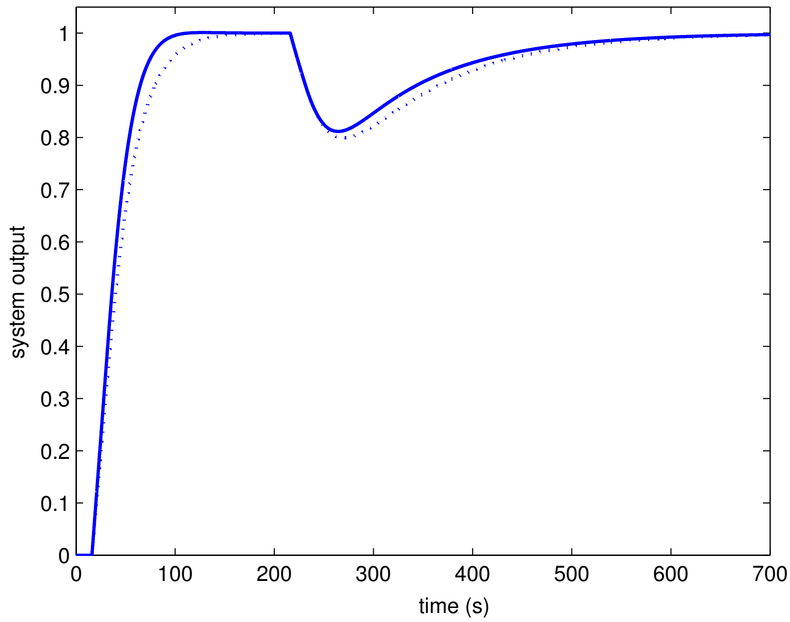
Figure 5.18: Step responses for $G_5(s)$: smooth control for nominal ($-$) and perturbed (...) process model, Chen et al. [14] for nominal ($--$) and perturbed ($-.-$) process model

5.5 Comparison of PI and PID Tuning Rules

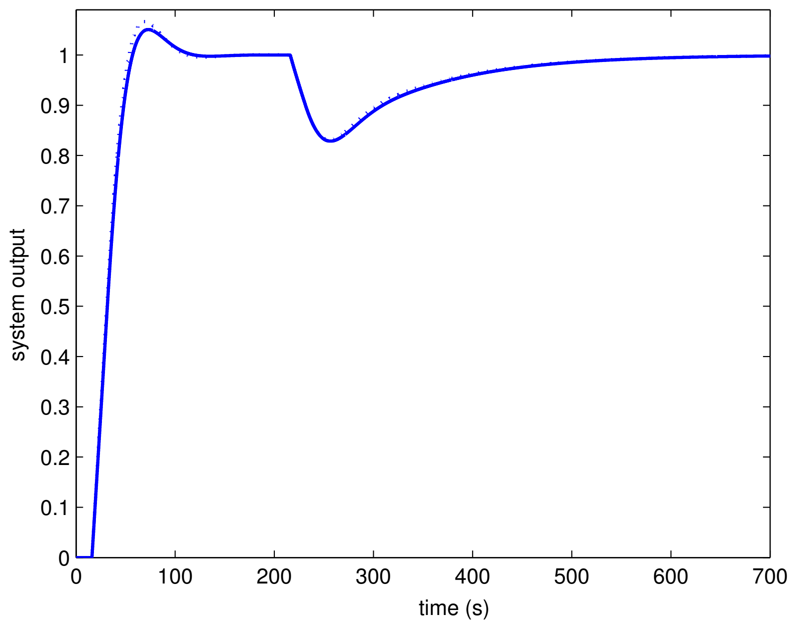
As tuning rules for PI/PID controller for stable FOPTD models have been proposed in this work, a comparative study of the achievable closed performances is done in this section. The system outputs for various values of θ for no overshoot are shown in Fig. 5.20. As is evident from the step response plots that the PID gives faster setpoint response and hence, a PID should be used instead of PI controller for improved closed loop performance.

5.6 Conclusions

The controller design problem for stable FOPTD and SOPTD plant models such that the response of the compensated system has overshoot below a prescribed value is widely addressed in the literature. An attempt has been made on similar lines in this work to propose analytical expression correlating the proportional gain of the PID controller to percentage overshoot for stable FOPTD plant models. The results are compared with Abbas's method to demonstrate the superiority of the proposed tuning rules. IPTD process models are considered next, and it is shown that the PD controller gives better servo control performances for such class of process models. However, the PD controller fails to reject the load disturbances and hence, PID settings for satis-



(a) no overshoot



(b) five percent overshoot

Figure 5.19: Step responses for $G_7(s)$: ($-$) proposed, (...) Mnif [15]

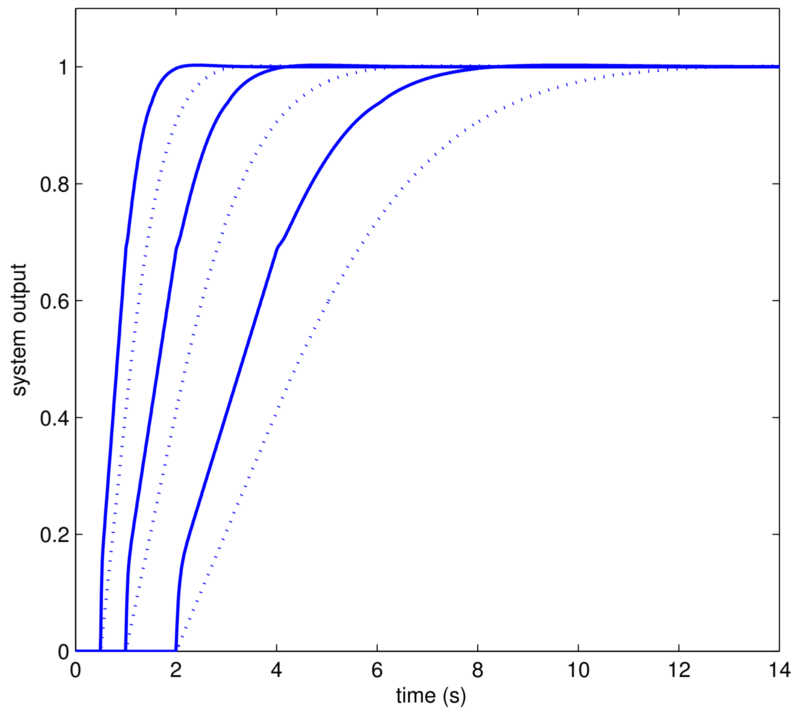


Figure 5.20: Step responses for FOPTD process models for $\theta = 0.5, 1, 2$: $(-)$ PID and (...) PI

factory load disturbance rejection performances are also proposed. Finally, the proposed approach is extended to the IMC framework to derive an analytical expression correlating the normalized IMC filter time constant to percentage overshoot. The proposed tuning rules are very simple and only a user-defined overshoot is required to design a PI/PID controller. Simulation examples are given to demonstrate the value of the proposed tuning method.

Chapter 6

Conclusions and Future Work

6.1 Concluding Remarks

The main purpose of this thesis has been to investigate some aspects of linear controller design methods for a class of processes in particular integrating plant models. The newly proposed evolutionary algorithm namely bacterial foraging is used to search the optimal PID parameters and the results are compared with those obtained by particle swarm optimization to illustrate the better convergence of the new algorithm. The integral of time $\dot{u}(t)$ squared is added to the ISTE criterion to obtain closed loop step responses with desired percent overshoot. A strategy to obtain the run length vector that ensures convergence and two modifications in the original BFA are proposed for better results. A new objective function that ensures smooth step responses with no overshoot for transfer function models commonly encountered in the process industry is also proposed. Tuning formulas for PID and its variants are obtained by minimizing the commonly used ISTE criterion and its higher versions for a class of integrating processes. The controller parameters are also obtained by minimizing the ISE criterion with the constraint that the slope of the Nyquist curve has a user specified value at gain crossover frequency. A comparative study of the closed loop performances that can be achieved by the proposed tuning methods is done to help the user in selecting tuning rules and the results are compared with some of the latest reported approaches to prove the value of the proposed design methods. Tuning formulas for PI-PD controller for IPTD and IFOPTD plant models by minimizing the proposed new objective function are also presented. As PID and its variants fail to give satisfactory closed loop responses for integrating plant models with large time delay, it is shown that the three controllers of the modified Smith predictor if tuned properly gives improved closed loop performances as compared to some of the recently reported approaches in the literature. The controller for setpoint tracking is designed so that the delay free component of the system output follows the response of a second order plant, assuming a perfect matching between the actual plant and model in both the dynamics and time delay. The PD controller considered for load disturbance rejection is tuned to achieve a user specified slope of the Nyquist curve at gain crossover frequency and a proportional controller with unity gain is considered for stabilizing the integrating plant models. The thesis finally addresses the controller design problem for stable FOPTD and SOPTD plant models such that the response of the compensated system has overshoot below a specified value. Analytical expressions correlating the controller parameters to percentage overshoot and plant model parameters are provided and it is shown that the PD controller gives better servo control performances for integrating processes. The approach is also extended to the IMC framework to obtain the normalized IMC filter time constant as a function of percentage overshoot. The major contributions of the work have been presented in Chapters 2-5 and conclusions of the investigation have been given at the end of each chapter.

6.2 Thesis Contributions

The contributions of the thesis are highlighted below:

1. An adaptive strategy for run length vector that ensures convergence of bacterial foraging algorithm
2. A modified ISTE criterion to obtain closed loop step responses with desired percent overshoot
3. A new objective function which results in robust closed loop step responses
4. Tuning formulas for PI and PID controllers for integrating plant models by
   - Minimizing ISTE, IST^2E and IST^3E integral criteria
   - Adjusting the slope of the Nyquist curve at gain crossover frequency
5. Tuning rules for PI-PD controller for IPTD and IFOPTD plant models
6. Analytical design method for the modified Smith predictor
   - A P controller with unity gain for stabilization

- Tuning formulas for the setpoint tracking controller as a function of the damping ratio and the natural frequency
- PD controller for load disturbance rejection based on loop slope adjustment

7. Tuning rules for stable FOPTD/SOPTD plant models to obtain specified percent overshoot

8. The normalized IMC filter parameter as a function of percent overshoot

6.3 Directions for Future Work

A few tracks for future research out of the present work are outlined below.

- Mathematical analysis of the chemotaxis step of BFA to study convergence properties will be an interesting and useful area of research.

- A PD controller gives better servo performance for integrating processes whereas a PID is needed for load disturbance rejection. Hence, proposing a two degree of freedom control architecture for the satisfactory closed loop operation of integrating processes warrants further investigation.

- Tuning formulas for unstable plant models by minimizing higher versions of ISTE criterion and a comparison with the latest reported approaches to find its suitability may be carried out.

- Controller tuning based on maximum sensitivity can be extended to non-minimum phase plants in the IMC framework.

- Analytical expressions correlating PD parameters to plant model parameters and percentage overshoot can also be obtained for IFOPTD and DIPTD process models.

Appendix A

Selection of Performance Criterion

A comparative study of the closed loop performances that can be achieved by the various integral criteria is done in this appendix. Assuming the steady state gain and the time constant of the plant model as unity, two values of θ $(0.5, 2)$ are considered for IPTD, IFOPTD and DIPTD plant models, respectively. Evaluating the results for the load disturbance rejection, a trade-off between the peak value of the system output and settling time appears; namely, employing the ISTE criterion means that a smaller value of the system output but a larger settling time occurs, with respect to the IST^{2}E and, more significantly, the IST^{3}E criterion. The IST^{3}E criterion is therefore recommended apart from the following two cases

1. PID control for IPTD plant model

2. PID control for IFOPTD plant model

where it is observed that the IST^{2}E and IST^{3}E criteria give almost similar closed loop step responses. The results are summarized in the following table.

Table A.1: Recommended Integral Criterions

Process model	PI	PID
IPTD	IST^{3}E	IST^{2}E
IFOPTD	IST^{3}E	IST^{2}E
DIPTD	–	IST^{3}E

A.1 IPTD Process Model

A.1.1 PI Controller

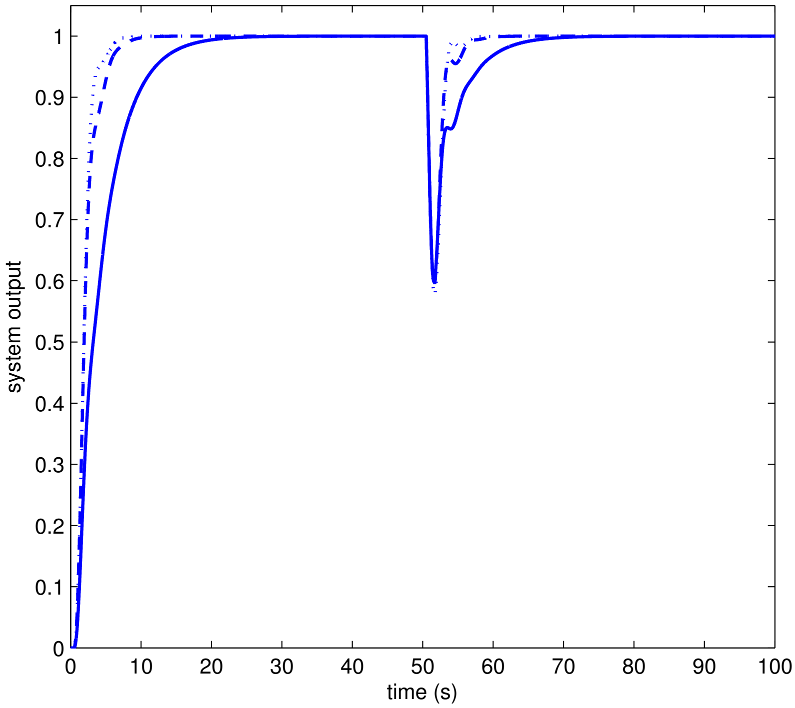


Figure A.1: Step responses for $\frac{e^{-.5s}}{s}$ $(L = -0.50)$: (–) ISTE, (– –) IST^{2}E and (...) IST^{3}E

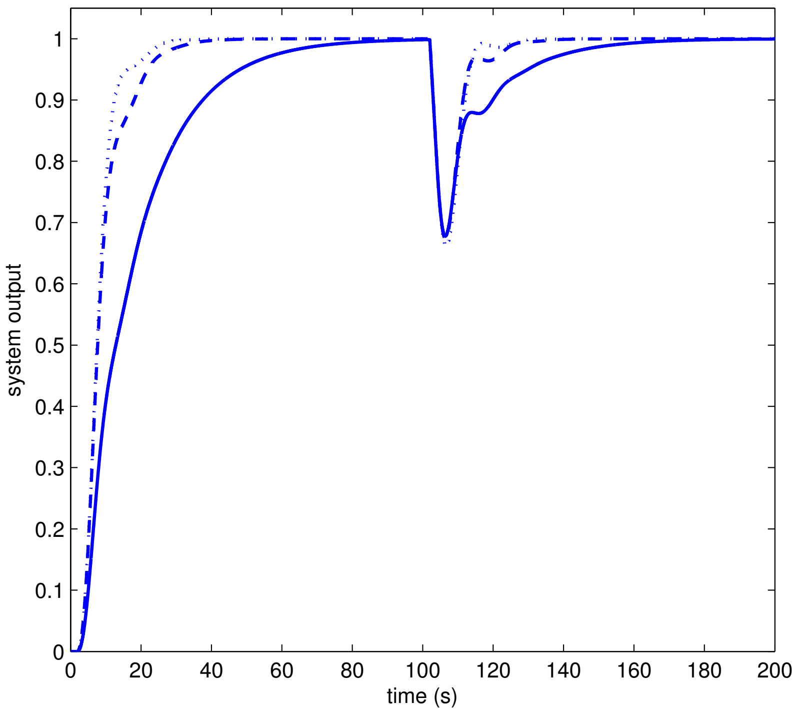


Figure A.2: Step responses for $\frac{e^{-2s}}{s}$ $(L = -0.10)$: (–) ISTE, (– –) IST^{2}E and (...) IST^{3}E

A.1.2 PID Controller

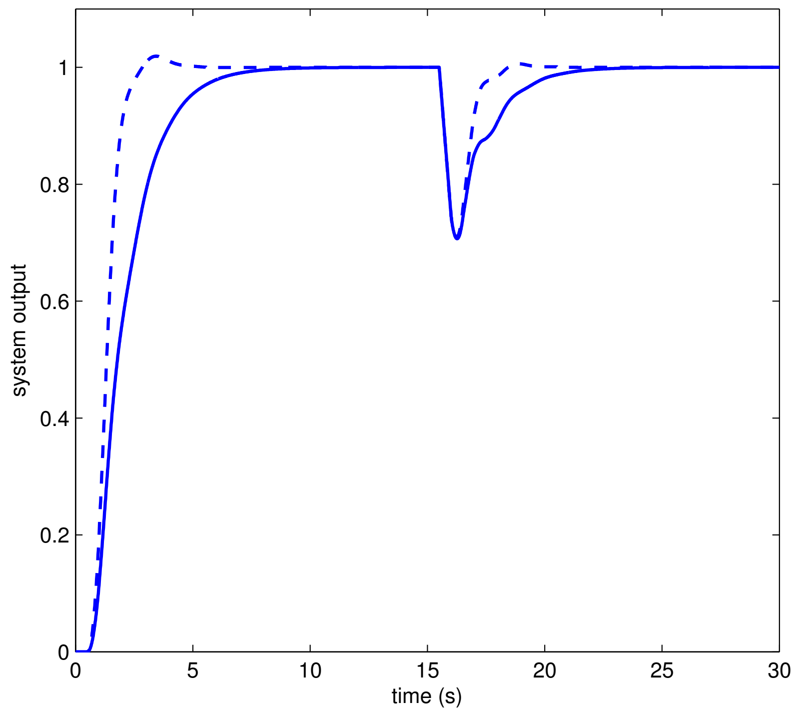


Figure A.3: Step responses for $\frac{e^{-.5s}}{s}$ ($L = -0.50$): (–) ISTE and (– –) IST^{2}E

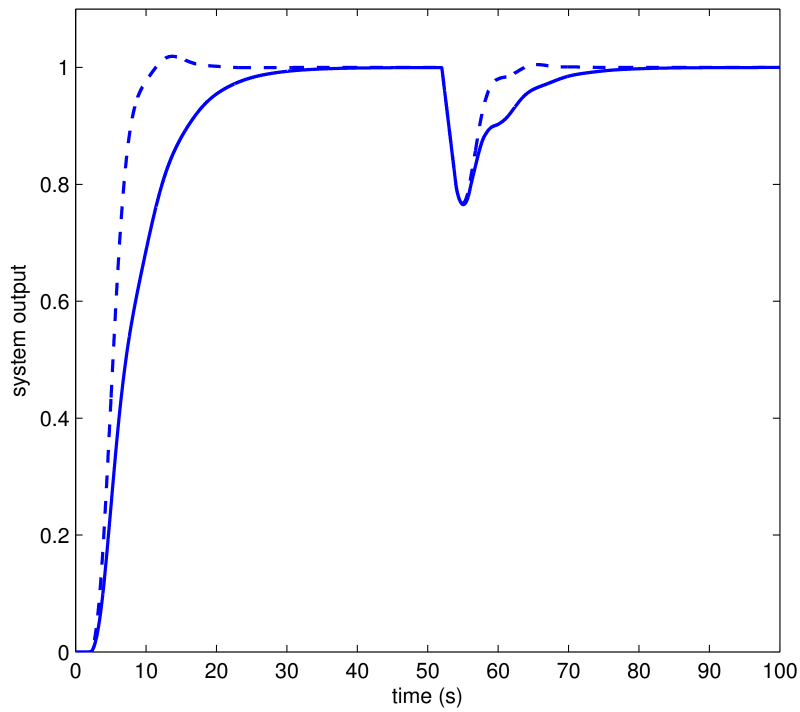


Figure A.4: Step responses for $\frac{e^{-2s}}{s}$ ($L = -0.10$): (–) ISTE and (– –) IST^{2}E

A.2 IFOPTD Process Model

A.2.1 PI Controller

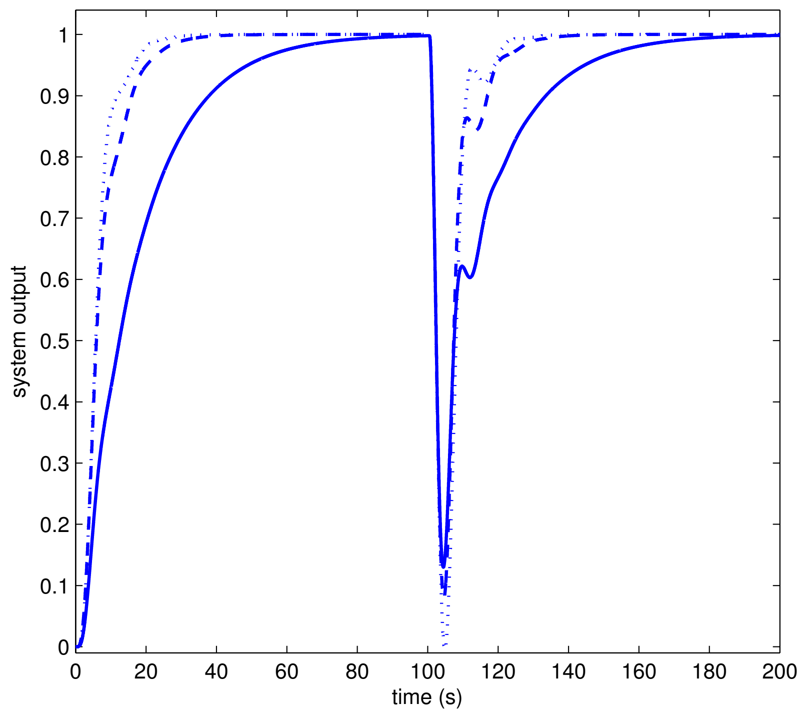


Figure A.5: Step responses for $\frac{e^{-.5s}}{s(s+1)}$ $(L = -0.50)$: (–) ISTE, (– –) IST^{2}E and (...) IST^{3}E

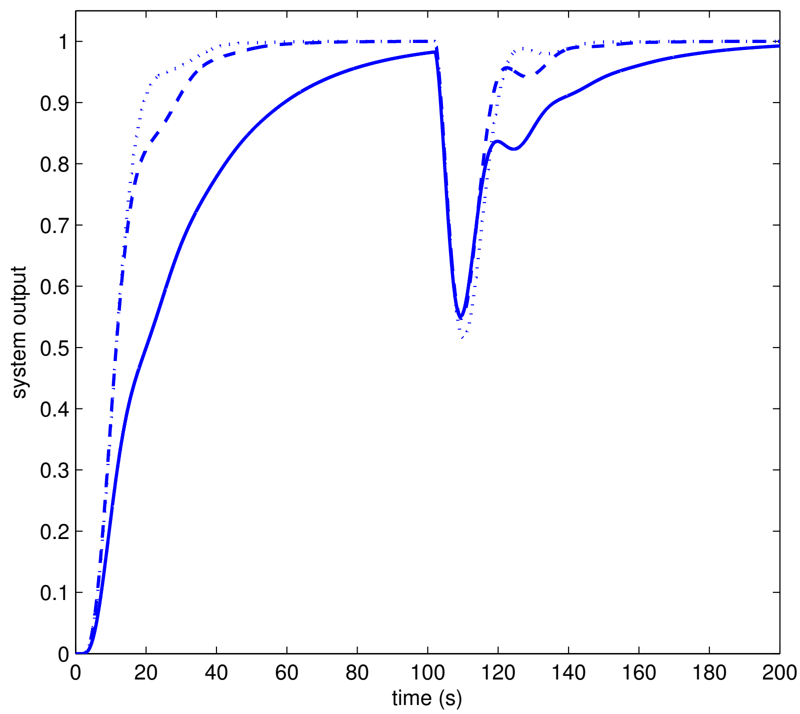


Figure A.6: Step responses for $\frac{e^{-2s}}{s(s+1)}$ $(L = -0.10)$: (–) ISTE, (– –) IST^{2}E and (...) IST^{3}E

A.2.2 PID Controller

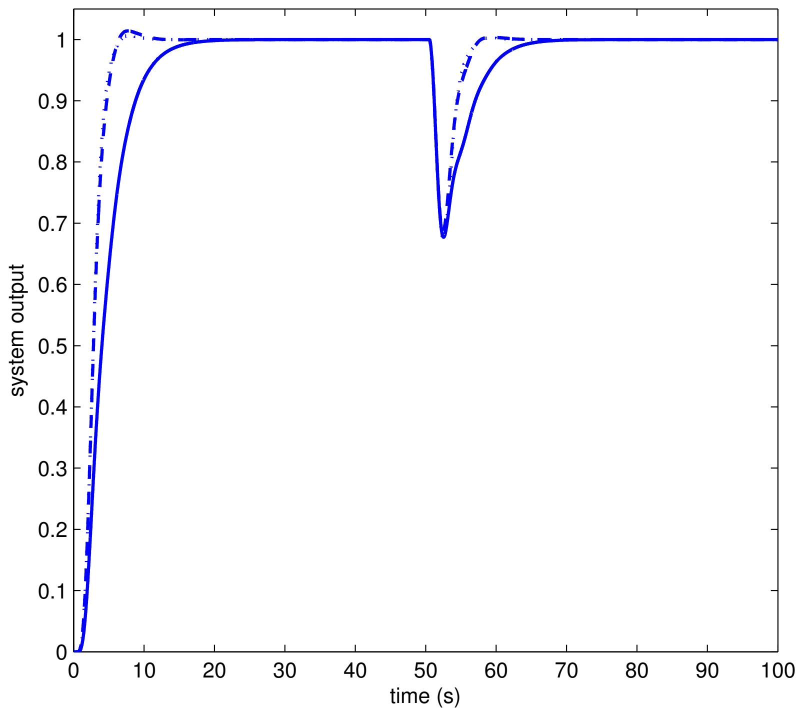


Figure A.7: Step responses for $\frac{e^{-.5s}}{s(s+1)}$ $(L = -0.50)$: (–) ISTE, (– –) IST^{2}E and (...) IST^{3}E

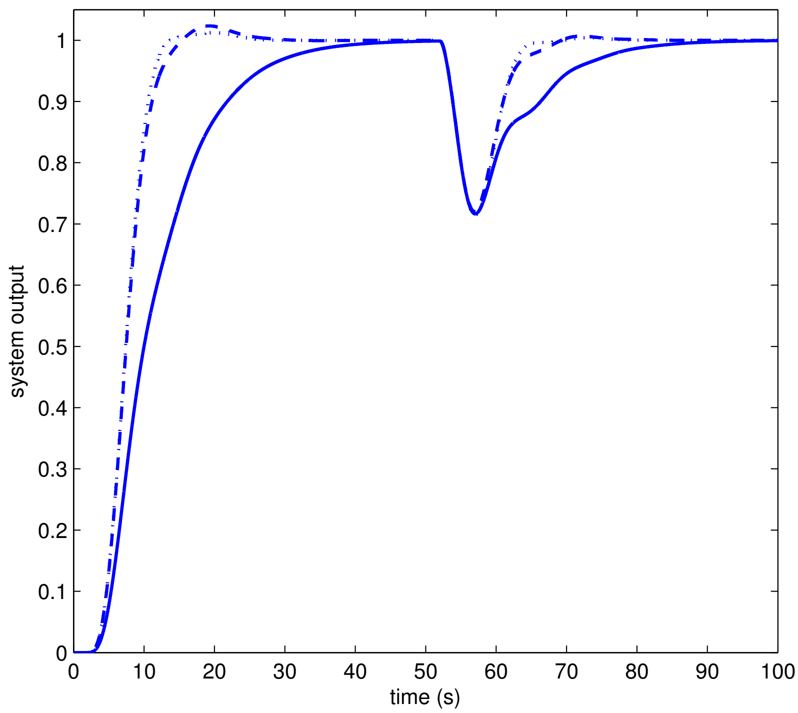


Figure A.8: Step responses for $\frac{e^{-2s}}{s(s+1)}$ $(L = -0.10)$: (–) ISTE, (– –) IST^{2}E and (...) IST^{3}E

A.3 DIPTD Process Model

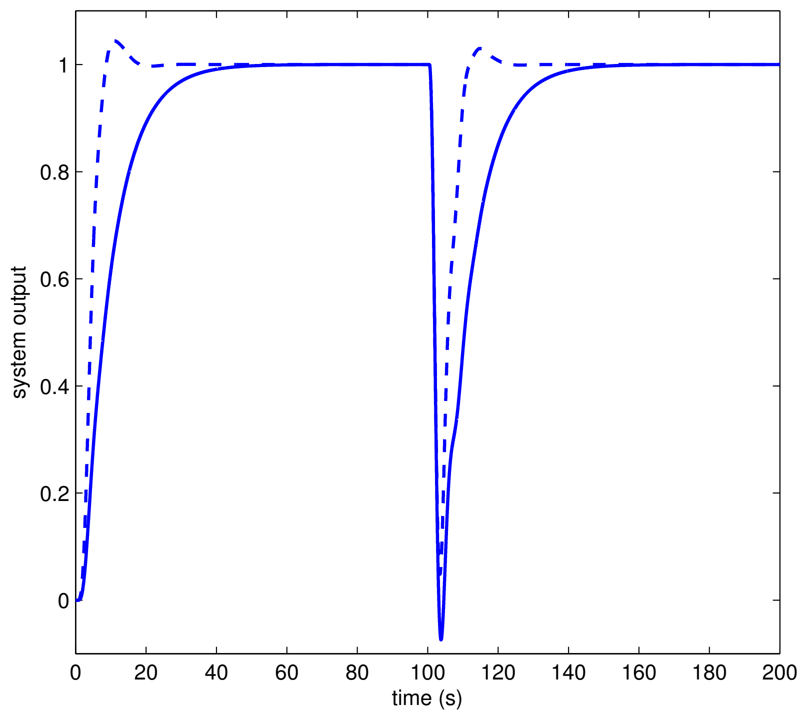


Figure A.9: Step responses for $\frac{e^{-.5s}}{s^2}$ $(L = -0.50)$: (–) IST^{2}E and (– –) IST^{3}E

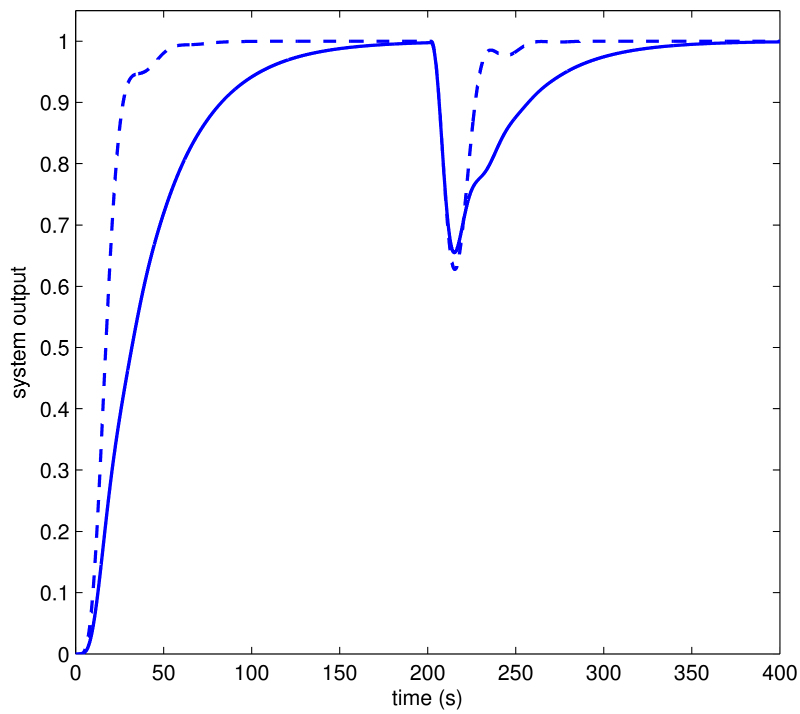


Figure A.10: Step responses for $\frac{e^{-2s}}{s^2}$ $(L = -0.01)$: (–) IST^{2}E and (– –) IST^{3}E

Appendix B

A Brief Review of Related Controller Design Schemes

- A method for tuning a PID controller for integrating plant models has been proposed by Sree and Chidambaram [7] so as to maintain a specified ratio between the coefficients of corresponding powers of s, s^2, s^3 in the numerator and the denominator, respectively. The suggested PID settings are: $K_p = \frac{0.896}{K\theta}$, $T_i = 2.5\theta$ and $T_d = 0.5\theta$.

- Rao et al. [11] have addressed the problem of controller design for a modified form of the Smith predictor for integrating and double integrating processes with time delay. The modified Smith predictor has two controllers, namely, a setpoint tracking controller and a load disturbance rejection controller for obtaining good servo and regulatory performances, respectively. The setpoint tracking controller is designed using the classical direct synthesis method based on the process model without considering the time delay whereas the PD controller considered for disturbance rejection is designed using optimal gain and phase margin approaches. The undesirable overshoot is taken care of by setpoint weighting. Guidelines have been provided by the authors for selecting the desired closed loop tuning parameter in the direct synthesis method and the setpoint weighting parameter.

- Kookos et al. [4] have proposed the following PI settings for an IPTD plant model based on the gain and phase margin specifications of 3.3 and 45°, respectively: $K_p = \frac{\pi}{7K\theta}$ and $T_i = 6.67\theta$.

- Poulin and Pomerleau [8] have presented a unified approach for the design of PI and PID controllers, based on the constant-M circles of the Nichols chart. The controller parameters are tuned such that the open loop frequency curve follows the corresponding constant-M circle using a predetermined maximum peak resonance. This approach gives the possibility of having the desirable maximum overshoot, the phase and gain margins and the bandwidth of the closed-loop system simultaneously.

- Visioli [13] has proposed PD/PID controllers for setpoint tracking and load disturbance rejection for IPTD plant models by minimizing ISE, ITSE and ISTE performance criteria. The recommended controller settings for servo and regulatory performances are $K_p = 0.90/(K\theta)$, $T_d = 0.45\theta$ and $K_p = 1.34/(K\theta)$, $T_i = 1.83\theta$, $T_d = 0.49\theta$, respectively.

- Table B.1 shows the PI/PID settings proposed by Skogestad [5] for a class of processes by equating the IMC filter time constant to the process model's equivalent time delay.

Table B.1: SIMC settings for various plant models

Process Model	K_p	T_i	T_d
FOPTD	$\frac{T}{K(\lambda+\theta)}$	$min\{T, 4(\lambda+\theta)\}$	$-$
IPTD	$\frac{1}{K(\lambda+\theta)}$	$4(\lambda+\theta)$	$-$
IFOPTD	$\frac{1}{K(\lambda+\theta)}$	$4(\lambda+\theta)$	T
DIPTD	$\frac{1}{4K(\lambda+\theta)^2}$	$4(\lambda+\theta)$	$4(\lambda+\theta)$

- Kaya [6] has proposed a model based PI-PD controller design for integrating and unstable plant models, where the PD feedback is used to change the poles of the plant transfer function to more desirable locations for control by a PI controller. The ISTE standard forms are used to obtain the parameters of the PI-PD controller as this enables the design to be completed by simple algebra.

- PID tuning rules based on loop shaping H infinity control have been proposed by Tan et al. [9] for stable and integrating plant models. The controller parameters are obtained using one design parameter which reflects the trade-off between robustness and time domain performance of the closed loop system, thus making it convenient to choose the controller parameters in the face of uncertainties in the plant parameters.

- The settings for P, PI and PID controllers which explicitly consider the characteristics of the closed loop response in particular, the percentage overshoot are proposed in [12]. The controller parameters are obtained in terms of plant model parameters by specifying the desired closed loop transfer function to follow an underdamped trajectory and by a suitable approximation of the plant delay. The suggested tuning relations are $K_p = \frac{a+b(\theta/T)^c}{d+e(\%os)^f}$, $T_i = T + \frac{\theta}{2}$, $T_d = \frac{T\theta}{2T+\theta}$ where a, b, c, d, e, f are the tuning constants and their values are given in the table below:

Table B.2: Parameters of the K_p equation

Controller	a	b	c	d	e	f
P	0.127	0.247	−1.050	0.918	−0.838	0.295
PI	0.148	0.186	−1.045	0.497	−0.464	0.590
PID	0.177	0.348	−1.002	0.531	−0.359	0.713

- The following tuning rules have been proposed by Chen et al. [14] to achieve a $+2$ dB closed loop amplitude ratio in the frequency domain.
  - FOPTD plant model: $K_p = \frac{0.7T}{K\theta}$ and $T_i = T$
  - SOPTD plant model: $K_p = \frac{1.4\xi\tau}{K\theta}$, $T_i = 2\tau\xi$ and $T_d = \frac{\tau}{2\xi}$

Appendix C

Why a PI controller fails to stabilize a DIPTD plant model?

The inability of the PI controller to stabilize a DIPTD plant model is illustrated using the concept of gain and phase margins. The open loop transfer function $G_{ol}(s)$ is given by

$$G_{ol}(s) = \frac{Ke^{-\theta s}}{s^2} \frac{K_p s + K_i}{s} \tag{C.1}$$

where $K_i = \frac{K_p}{T_i}$. Using equation (1.11), we get

$$\phi_m = -270^\circ - \omega_g \theta + \arctan \frac{K_p \omega_g}{K_i} + \pi \tag{C.2}$$

The maximum possible value of $\frac{K_p \omega_g}{K_i}$ is ∞ and hence, the above equation reduces to

$$\phi_m = -\omega_g \theta \tag{C.3}$$

From (C.3), it is observed that the phase margin is always negative irrespective of the gain crossover frequency thereby, indicating that a DIPTD plant model cannot be stabilized by a PI controller. Furthermore, the PID parameters obtained by minimizing ISTE criterion resulted in a very large integral time constant with poor convergence properties. The integral time constant was therefore, taken as infinity which resulted in a PD controller with excellent convergence of bacterial foraging algorithm.

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