Design of a Non-linear Analog PID Controller

by

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Abstract

In this study we propose an analogue nonlinear PID controller with antiwindup and dead-time compensation to optimise the control of loops experiencing degradation in the control performance as a result of dead-time and saturation nonlinearity. Loops containing a significant dead-time are notoriously difficult to control. The proposed controller optimises the control of loops experiencing the negative effects of saturation and dead-time.

Quality of control with respect to input and load disturbances are analysed using typical process models. The control performance of the proposed scheme shall be compared to widely used control algorithms to yield an improved control performance over standard PID control.

Combinations of linear and nonlinear control work well in improving the control performance of loops experiencing instability as a result of the negative effects of nonlinearities. Two nonlinear control schemes are proposed in this study. The first controller shall have variable nonlinear gains which are subject to specific nonlinear laws. For the second controller a PD modifier shall be proposed to overcome any degradation occurring in the performance of the control loop. Both controllers shall take into account the different needs of control with respect to the magnitude of the error signal and the rate of change of the error signal.

Improperly tuned control loops using standard PID control experience a poor control performance. The proposed PID and nonlinear modifier, with antiwindup and dead-time compensation, shall yield a good control performance even when it is not properly tuned. It will be easy to tune and shall be insensitive to a wide range of parameter settings.

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Chapter 1

Introduction to Proportional - Integral - Derivative Control

1.1 Background

The control of industrial processes is a well established engineering discipline (see Albert & Coogan, 1992, Åstrom & Hägglund, 1988, Martins de Carvalho, 1993, Shinskey, 1979). Its main aim is to modify the process behavior in order to drive processes, and in addition to safely achieve a set of prespecified desired objectives. These crucial tasks are accomplished by means of different types of controllers (see Albert & Coogan, 1992, Åstrom & Hägglund, 1988, Martins de Carvalho, 1993, Shinskey, 1979). Among these, the proportional-integral-derivative (PID) controllers are mostly used. These PID controllers are sometimes called three-term controllers.

It is estimated that over 70% of all installed industrial controllers are of the PID type (Brown, 1994). We will restrict our study only to those controllers that are used for single control loops. The reason for this is that the basic control problems appear normally in the single loop configuration, and if they are solved successfully on this level they can be extended to the more complex control topology. From an engineering point of view two aspects of control are important: disturbance rejection and setpoint tracking.

The former has much greater significance since usually the control loop most of the time operates under the constant reference signal and in such conditions the main task of the controller is to minimize the effects of disturbances.

Various factors within the control loop prevent the conventional PID controller from simultaneously achieving both:

- good disturbance rejection and
- good setpoint tracking.

Because of this, many alterations of these controllers are proposed (see for example Åstrom & Hägglund, 1988, Bajić, 1995a, Bajić & Rybalov, 1995, Glattfelder & Schaufelberger, 1986, Govender & Bajić, 1996) that utilize several different techniques like feedforward control, antiwindup mechanisms, nonlinear gains, and nonlinear proportional-derivative (PD) modification. In this study we will propose some alterations of the conventional PID control structure that will provide specific improvements of the control performance of a single loop structure. Particularly, we will propose new controllers suitable for control in single loops containing processes with long dead-time and significant actuator nonlinearity. These proposed solutions are useful and constructive combinations of several known techniques. Also an original contribution in this direction is made. The validity of these proposed techniques are demonstrated by using simulation experiments. Finally, based on simulation results, an electronic controller is developed and tested to confirm in practice the validity of the proposed control concepts.

The structure of this report is as follows. In Chapter 1 a review of the basic concepts related to linear PID control is given. Chapter 2 is devoted to the description and analysis of integral windup and measures to reduce its negative effects. In Chapter 3 a PID controller with nonlinear gains is proposed to provide improved setpoint tracking for loops containing processes with long dead time and significant actuator nonlinearity. In Chapter 4 the results of comuter simulation experiments are given for the controller from Chapter 3. A different and more efficient control structure is proposed in Chapter

5 and adapted to cater for systems with long dead-time and actuator nonlinearities in Chapter 6. Simulation results achieved with such a controller are given in Chapter 7. In Chapter 8 the electronic controller developed on the basis of results from Chapter 5 is presented with the experimental results. Finally, concluding remarks are given in Chapter 9, followed by a list of references in Chapter 10.

1.2 Basic Control Loop

This section describes the basic structure of the control loop, its most important components, and the signals involved. An explanation of the linear constituents of the PID controller is also given.

1.2.1 Extracts from Theory

Proportional (P), integral (I) and derivative (D) control action can be simultaneously utilized in a device called the PID controller. Today PID controllers are the standard building blocks for industrial automation (Åström & Hägglund, 1988) within the process industry. The popularity of these controllers comes from their functional simplicity and their ability to function in a wide variety of control applications with reasonable success (see Åström & Hägglund, 1988, Shinskey, 1979, 1994). Like for all other controllers, the main objective of this, so-called three-term, controller is to alter the behavior of a process with which it forms a control loop, in order to satisfy a number of particular requirements. One of these requirements may involve maintaining the process output as close as possible to a predetermined value which we usually call the setpoint. The other requirements can also be prespecified like the settling time, the rise time, the limit for overshoot, gain and phase margin (for an explanation of these concepts see ,for example Macdonald & Lowe, 1981) and so on. The controller achieves these requirements by adjusting its output in accordance with specific control laws, which in the case of a PID controller are the proportional, integral and derivative control laws. As a consequence the input signal to the process is modified and this directly reflects to the process output. A simplified control loop (having all linear components) with the PID controller is shown in Fig.1-1. An explanation of each of the components and signals involved with the control loop or which contribute to its functioning is discussed in following section.

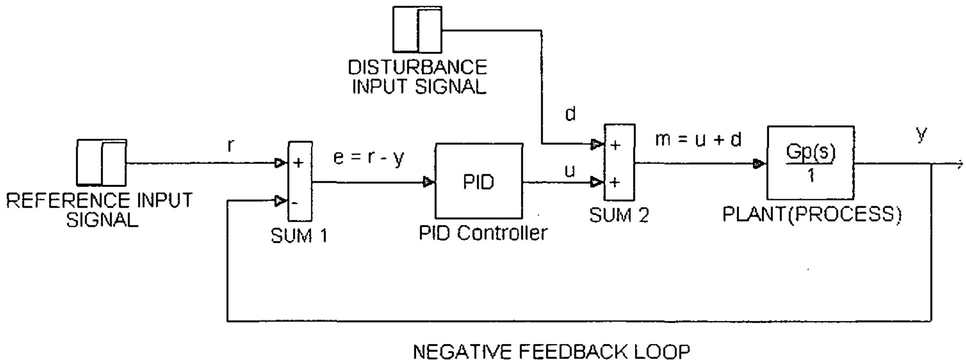

Figure 1-1: The position of a PID controller within a typical closed-loop feedback control system

1.2.2 Basic Control Loop Constituents

There are five main signals that appear in the single control loop. They are:

a/ the process output signal

b/ reference input signal

c/ error signal

d/ disturbance signal

e/ control signal.

An explanation of each of these signals, and the components of the control loop will be given.

- **Controlled output (process output) signal (y)**

  The *controlled output signal (y)* is that variable of the process which is aimed to be controlled. Usually it is required that it follows the reference input signal as close as possible. Together with the reference input signal it is used to form the error signal. In Fig.1-1 only one controlled output signal is considered, although in more complicated cases the plant may have several outputs to be controlled.

- **Reference input (setpoint) signal (r)**

  The *reference input signal (r)* represents a signal that needs to be followed (in most cases) as close as possible by the process output signal.

- **Error signal (e)**

  The *error signal (e)* is obtained as the difference between the reference input signal r and the process output signal y, i.e. $e = r - y$. This signal is used by the controller to generate a control signal u. The main reason for introducing feedback in control problems is the requirement to have an error signal that is in principle the main source of information for the controller.

- **Disturbance signal (d)**

  A *disturbance signal (d)* is an unpredictable, undesirable and usually unmeasurable input signal which could occur anywhere within the control loop. The most significant disturbances usually act on the system at the input of the process. This may influence the process output drastically. The main purpose of the controller in process applications is to minimize the effects of these disturbances on the process output.

- **Control signal or manipulated variable (u)**

  In principle, the controller uses the error signal to generate a *control signal (u)*. This control signal is a function of the error signal, controller parameters and controller dynamics. It is applied to the input of the process.

- **Process (plant)**

  The process is the system whose output(s) has to be controlled.

- **Controller**

  The main function of the controller is to drive the process output according to the requirements or, expressed more realistically, to ensure that the deviation between the process output signal and the reference input signal is within the acceptable bounds. The most primitive variant of the PID control fulfils this function by manipulating the error signal in accordance with the relevant control laws for the purpose of generating a control signal u. This control signal drives the process to ensure that the error is within the required bounds and that the process response is otherwise acceptable. The extent to which the controller influences the process will depend on the controller adjustments. Several well known techniques can be used to adjust (*tune*) the controller, and the most widely used in industrial applications are the Ziegler-Nichols (Ziegler & Nichols, 1942, 1943, Chiang *et al.*, 1952, Shinskey, 1979) tuning methods.

1.3 Linear PID Controller

1.3.1 Proportional Control

Proportional control is the most straightforward of all three previously mentioned types of control. The output of a linear proportional controller varies proportionally to the system error according to

$$u_p(t) = K_p e(t) + b$$

where t is the current moment, u_p is the controller output, b is the controller bias and K_p is the controller gain, usually called the proportional gain. K_p can be adjusted, within some limits, to make the controller output change bigger or smaller when the error e, i.e. deviation between the setpoint and the controlled variable, changes. The bias b can also be adjusted. The change of bias allows (in some cases) for the controller output to be driven to the appropriate value when the reference input signal is constant. This is frequently utilized in the so-called manual reset.

The gain of a proportional controller is usually described in terms of its *proportional band (PB)*. The concept of the proportional band is inherited from pneumatic controllers and is defined as

$$PB = \frac{1}{K_p} \times 100\%.$$

Note that a large proportional gain K_p corresponds to a small proportional band PB, while a large PB implies a small gain K_p. In most cases, as PB decreases (i.e. when K_p increases), the proportional action will cause a smaller error in the steady state. The steady-state error (e_{ss}) is the difference that exists between the reference input signal r and the process output signal y in the steady state. Sometimes it is also called offset. Consider now the system shown in Fig.1-2.

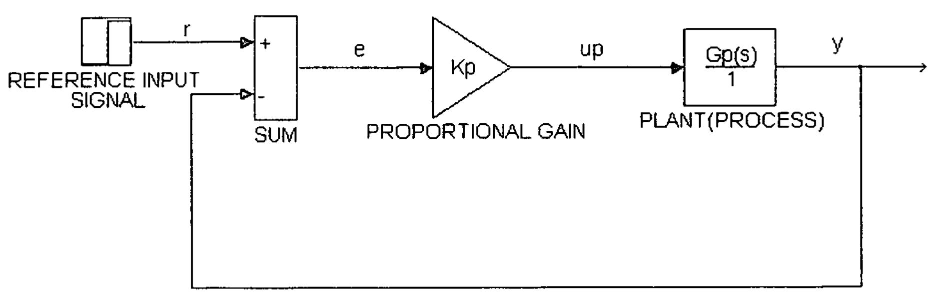

Figure 1-2: Position of the proportional controller within a closed-loop feedback control system

Let us assume that $e = 0$. The closed-loop transfer function of this control system is

$$\frac{C(s)}{R(s)} = \frac{K_pG_p(s)}{1 + K_pG_p(s)}$$

where G_p is the transfer function of the process. The Laplace transform of the error is given by

$$E(s) = \frac{R(s)}{1 + K_pG_p(s)}$$

The action of the proportional controller leads to offset in the output y in the steady-state for type-0 processes (McDonald & Lowe, 1981). This steady state error is influenced by the input signal and the type number of the control system. A type-0 system is characterized by the absence of an error integrating effect and by the existence of a steady-state error when a step type input signal is applied. The type-0 systems have $G_p(0) \neq \pm\infty$. Under certain conditions the steady-state error for the control system shown in Fig.1-2 can be calculated by using the so-called *final value theorem* (see MacDonald & Lowe, 1981) as

$$e_{ss}(+\infty) = \lim_{s\to 0}[sE(s)]$$

If a unit step is applied to the input of this control system, then

$$\begin{aligned} e_{ss}(+\infty) &= \lim_{s\to 0}\left(s\frac{1}{s}\frac{1}{1 + K_pG_p(s)}\right) \\ &= \lim_{s\to 0}\frac{1}{1 + K_pG_p(s)} \\ &= \frac{1}{1 + K_pG_p(0)} \end{aligned}$$

This indicates the presence of an offset if $G_p(0) \neq \pm\infty$, which is the case for type-0 systems. Also it can be seen that the absolute value of the steady state error can be reduced by sufficiently increasing K_p. Since K_p affects the system stability and dynamics, it will be limited by the stability constraints of the overall control system. A high value of K_p may lead to increased oscillations and large overshoots which could result in instability. It is for this reason that proportional control is combined with integral control in order to eliminate offset, while applying the smaller values of the gain K_p.

1.3.2 Integral Control

When an operator notices the existence of a steady-state error due to changes in the reference input signal and/or disturbances, he can correct for this 'manually' by altering either the bias b of the proportional-controller or the setpoint signal r until the error disappears. This is called *manual reset* (Palm, 1986). Integral control provides a means of *automatically* removing this steady-state error. In industrial terminology it is often referred to as *reset control* (Palm, 1986) or *automatic reset.*

Integral control is used in systems where proportional control alone is not capable of reducing the steady-state error within the acceptable bounds. Its primary effect on a process control system is to permanently attempt to gradually eliminate the error. The action of the integral controller is based on the principle that the control action should exist as long as the error is different from zero, and should have a tendency to gradually reduce it to zero. The integral control signal is proportional to both the magnitude and the duration of the error and is given by

$$u_i(t) = K_i \int_0^t e(s)ds$$

where K_i is called the *integral gain.* Frequently K_i is expressed as $K_i = K_p/T_i$, where T_i is the *integral action time* (or shortly *integral time*).

Integral Action as Automatic Reset

Integral action may be performed as a kind of automatic reset as shown in Fig.1-3. In fact, it is equivalent to permanently changing the bias of the proportional controller in an appropriate manner. To see this consider the scheme in Fig.1-3 and the following derivation using operator equations.

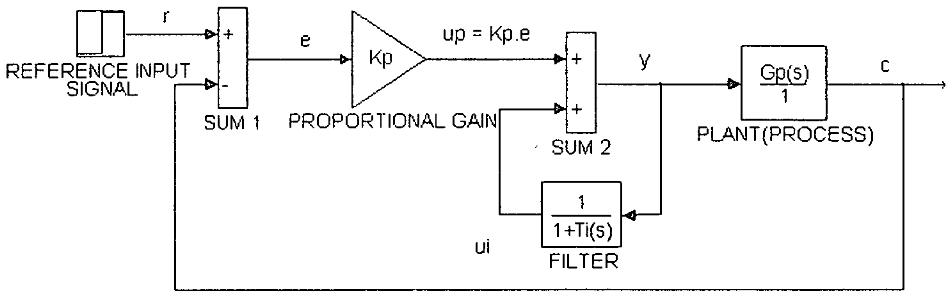

Figure 1-3: Proportional controller with an integrator as automatic reset

Let

$$y = K_p e + u_i$$

and

$$u_i = \frac{y}{1 + T_i s}$$

Then

$$\begin{aligned}
y &= K_p e + \frac{y}{1 + T_i s} \\
y - \frac{y}{1 + T_i s} &= K_p e \\
y\left(1 - \frac{1}{1 + T_i s}\right) &= K_p e \\
y\left(\frac{1 + T_i s}{1 + T_i s} - \frac{1}{1 + T_i s}\right) &= K_p e
\end{aligned}$$

$$
\begin{aligned}
\frac{y}{e} &= K_p \frac{1 + T_i s}{T_i s} \\
&= \frac{K_p}{T_i s} + \frac{K_p T_i s}{T_i s} \\
&= \frac{K_p}{T_i s} + K_p \\
&= \frac{K_p}{T_i} \times \frac{1}{s} + K_p \\
&= K_i \frac{1}{s} + K_p
\end{aligned}
$$

Solving for the controller output y one gets

$$
\begin{aligned}
y &= (K_i \times \frac{1}{s} + K_p) e \\
&= \frac{K_i}{s} e + K_p e
\end{aligned}
$$

This shows that the controller in Fig.1-3 behaves like a PI-controller, where the integral action takes the form of permanent adjustment of the bias of the proportional controller. There are also realizations of PI control different from the scheme shown in Fig.1-3. In the next subsection we will show that the PI-control structure has the ability to eliminate offset in some cases.

Offset Elimination Property of PI-controller structures

Proportional action comes into effect immediately as the error different from zero occurs. If the proportional gain is sufficiently high it will drive the error closer to zero. Integral control accomplishes essentially the same effect as the proportional controller having an infinitely high gain. This results in an *offset eliminating* property of the integral action for type-0 systems and step type input signals, and can be illustrated by applying the final value theorem to the PI controller structure shown in Fig.1-4.

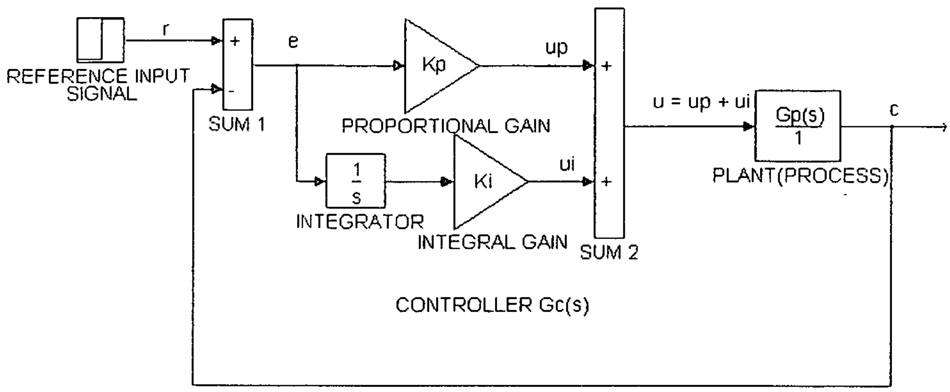

Figure 1-4: PI controller in a typical control loop

Consider in this case the Laplace transform of the error signal

$$E(s) = \frac{R(s)}{1 + G_p(s)G_c(s)}$$

where $G_c(s) = K_p + K_i/s$ and $R(s) = 1/s$. Based on the assumption that the final value theorem can be applied to calculate the steady state error (see MacDonald & Lowe, 1981), it follows that

$$\begin{aligned} e_{ss}(+\infty) &= \lim_{s\to 0}[sE(s)] \\ &= \lim_{s\to 0}\left[s \times \frac{1}{s} \times \frac{1}{1 + G_P(s)G_c(s)}\right] \\ &= \lim_{s\to 0}\left[\frac{s}{s + (K_p s + K_i)G_p(s)}\right] \\ &= 0 \end{aligned}$$

The value $e_{ss}(+\infty) = 0$ indicates that the offset is zero, which is what we intended to show.

Although integral control is very useful for removing steady-state errors, it is also responsible for sometimes introducing undesirable effects into the control loop in the form of an increased settling time, reduced stability and integral windup. A short explanation of each of these undesirable effects follows:

- **Increase of the settling time**

  An increase of the closed-loop system settling time is usually caused by the increased oscillations as a consequence of the present integral action (Matley *et al.*, 1986).

- **Reduction of stability**

  The presence of the integral action may lead to an increase of oscillations in the loop. These oscillations generally have a tendency to move the system more towards the boundary of stability. In some cases these oscillations will result in the loop becoming instable (Matley *et al.*, 1986).

- **Integral windup**

  Nonlinearities present within the loop, such as actuator saturation, may lead to different sensitivity of the process output to changes in the control signal. Whenever saturation occurs at the actuator output, the integral action will have to cease. If the integral-action does not stop, the integral part of the controller could command an action to which the actuator cannot react properly since it is saturated, and thus the aimed control effect will be violated. The result of this *windup* state could be severe overshoots in the controlled variable (Walgama & Sternby, 1989, 1993, Walgama *et al.*, 1992).

1.3.3 Derivative Control

In critical processes where 'tight' control is required, PI control alone may not suffice because:

- Both proportional control and integral control are incapable of responding to the rate of change of the error signal. This shortcoming could have a significant negative effect on stability due to possible increased oscillations in the system response, especially in applications where the value of the controlled variable is changing rapidly.

- The action of the proportional part of the controller becomes assertive only when a significant error has occurred.

- The gradual ramp type response of the integral part of the controller means that a significant time interval has to pass for the controller to produce a sizeable response. This time delay is due to the integral action of the integral controller.

It is for these reasons that derivative control is useful to respond to the rate of change of the error signal, even if the error is small. The control law of the derivative controller is given by

$$u_d(t) = K_p T_d \frac{de(t)}{dt} = K_d \frac{de(t)}{dt}$$

where K_d is called the derivative gain and T_d is called the *derivative action time* (or shortly *derivative time)*.

From the derivative control law it is evident that an important shortcoming of derivative action is that it will be absent when there is no change in the error signal. Even if there is a large constant error present within the system, the contribution of the derivative part of the controller will be zero. Derivative action on its own will therefore allow uncontrolled steady-state errors. It is for this reason that derivative control is combined with either proportional-control or proportional-integral control.

Another disadvantage of derivative-action is that it amplifies noise when it is contained in the error signal. Noise is the term used to describe random or otherwise irregular signals that can occur within the control loop. To prevent the possibility of any unstable behavior of the controller output occurring as a result of the noise, derivative action is often realized using the filtered process output signal, i.e. the process output is fed to the low-pass filter, the output of which serves as the input to the derivative action block. The control law for the derivative-controller then becomes

$$u_d(t) = K_d \frac{dy_f(t)}{dt}$$

where y_f is the output signal fed back through a low-pass filter. This signal is smoother than the process output signal y.

One way of including derivative action within the structure of a PD controller is illustrated in Fig.1-5. This configuration represents a textbook version of a PD controller. It is unsuitable for most industrial applications where step type changes in the reference input signal are present and/or where the noise is present in the process output signal.

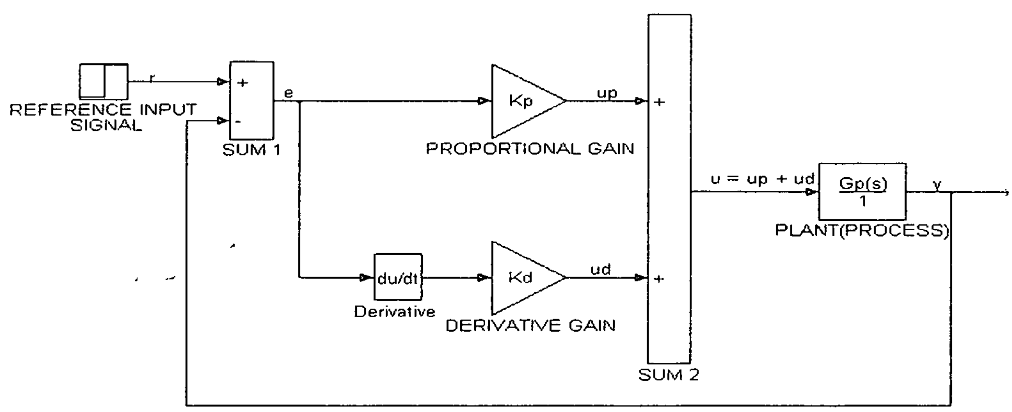

Figure 1-5: Text-book implementation of a PD controller in the closed loop

1.3.4 PID Controllers

All the control laws considered in the previous sections were linear. These control laws may be expressed in terms of the *degree-of-freedom* of the controller. If we call the controller an $n-$degree-of-freedom controller when there are $n-$input signals to the controller which can be independently manipulated within the controller, then for a one-degree-of-freedom linear PID controller shown in Fig.1-6 the control law will be

$$u_{c1}(t) = K_p e(t) + K_i \int_0^t e(s)ds + K_d De(t)$$

where $u_c(t)$ represents the output of the controller and $D \equiv d/dt$ is the derivative operator. Note that the term 'degree-of-freedom' of the controller can be used also with a different meaning.

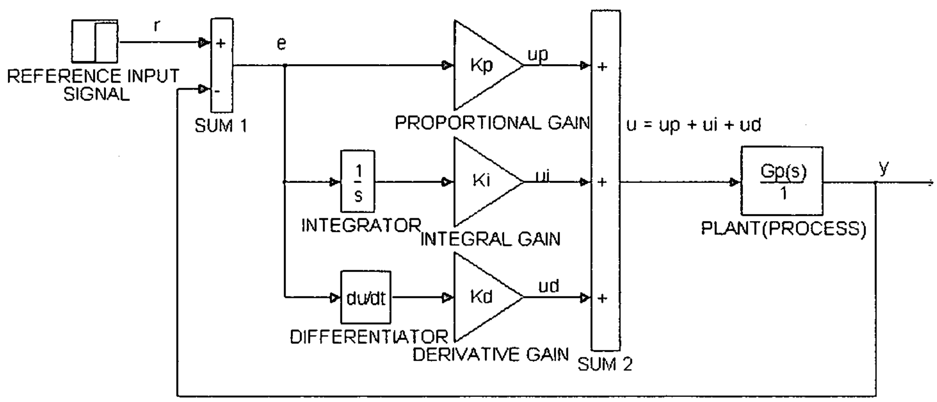

Figure 1-6: Implementing u_{c1} in the control loop

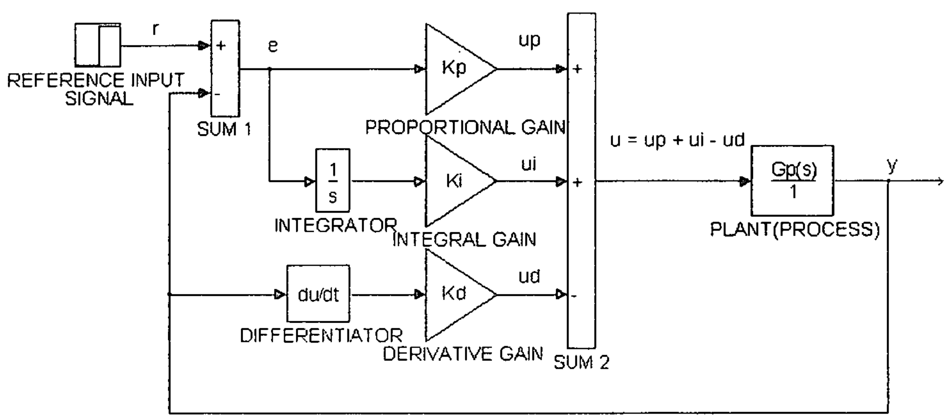

Figure 1-7: Implementing u_{c2} in the control loop

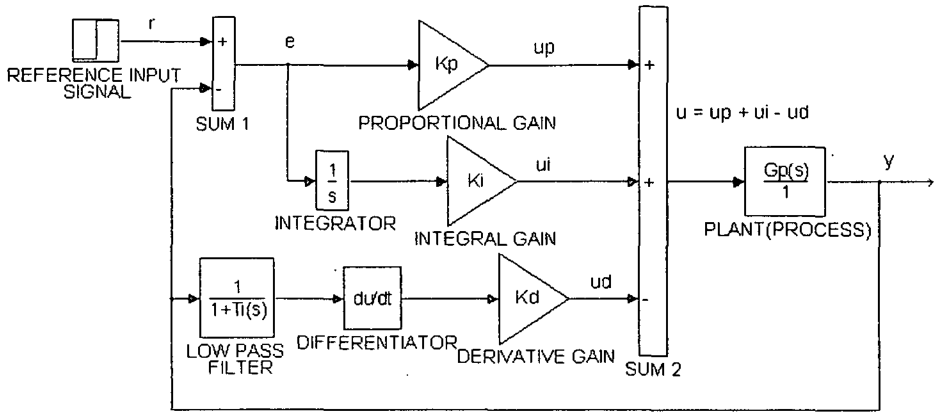

Figure 1-8: Implementing u_{c3} in the control loop

In the case of a two-degree-of-freedom linear PID controller shown in Fig.1-7 and Fig.1-8, the control laws are given by

$$u_{c2}(t) = K_p e(t) + K_i \int_0^t e(s)ds - K_d Dy(t)$$

or

$$u_{c3}(t) = K_p e(t) + K_i \int_0^t e(s)ds - K_d Dy_f(t)$$

respectively, where y_f is the filtered process output signal. These controllers can be represented in the Laplace transform domain as

$$U_{c1}(s) = \left(K_p + \frac{K_i}{s} + K_d s\right) E(s),$$

$$U_{c2}(s) = \left(K_p + \frac{K_i}{s}\right) E(s) - K_d s Y(s)$$

and

$$U_{c3}(s) = \left(K_p + \frac{K_i}{s}\right) E(s) - K_d s Y_f(s)$$

respectively. These laws form the *basis* of process control upon which control schemes are tailored in order to take into consideration the various factors which contribute to the dynamics of a control loop. The above-mentioned control laws are implemented in Figs.1-6, 1-7, 1-8.

The main disadvantage of the PID controller structure shown in Fig.1-6 is that it is sensitive to step changes in the reference input signal r and noise components in the process output signal y. The first problem is removed by the PID configuration in Fig.1-7, while the second problem is considerably reduced by the PID structure in Fig.1-8. The PID configuration shown in Fig.1-8 is also insensitive to step type changes in the reference input signal.

Other PID control strategies may also be used to improve the control performance of the controller. These strategies utilize set-point weighting (see Åstrom & Hägglund, 1988) or other feedforward techniques (see Astrom & Hägglund, 1988, Seborg *et al,* 1989, Bajić, 1994, 1995) to improve the control performance of the PID controller.

1.4 Conclusion

The PID controller is a synthesis of the three control modes discussed. If properly tuned it may provide a good control effect for a number of practical systems. All three modes of the controller interact with each other. Exactly how much of proportional, integral and derivative action is allowed for a good control will depend on the dynamics of the system which the controller aims to control, on the topology of the whole control system and the position of the controller in the loop.

Chapter 2

Actuator Nonlinearity and Antiwindup Measures

2.1 Introduction

The discussion thus far has been limited to systems consisting only of linear components. However, realistic models of physical systems normally contain nonlinearities. 'Nonlinearity' in control engineering is a term used to describe the functional relationship that exists between the input and output variables of a block. Nonlinearities appear naturally in any physical system. Ignoring these nonlinearities could yield models that do not reflect the true behavior of the system. An accurate analysis of any control system must take into consideration the presence of all significant nonlinearities which might influence the system performance.

Nonlinearities inherently present within a control system may exert a negative effect on the behavior of the control loop. If these nonlinearities are mild, there might not be too much of a problem. Where the nonlinearities are strong, such as in actuators, special consideration must be given to them. If a controller is used which neglects such nonlinearities, it may give rise to a deterioration in control performance (Walgama & Sternby, 1990). This deterioration in performance may have different manifestations and sometimes leads to increased oscillations, overshoots in response or otherwise unacceptable loop behavior.

The presence of nonlinearities such as saturation and others, as well as dead-time (which is considered a linear component), place severe constraints on the magnitude of the controllers tuning parameters (Albert & Coogan, 1992). For example, we make limitations on the magnitude of K_p in order to prevent a too large control signal that may lead to saturation of system components such as a valve. This reduction in the control signal magnitude has positive effect on control power savings and in ensuring system stability. But this limitation on the magnitude of K_p has the undesirable effect of reducing the speed of the system's reaction to any changes in external signals which may occur within the control loop, thus compromising the performance of the system.

Nonlinearities such as actuator saturation may also govern the behavior of the loop during start up, resulting in oscillations and overshoots. One way to reduce this negative effect of saturation would be to manually adjust the control signal sent to the valve until the system variables fall within the linear range of the actuator. Then the controller may be switched to automatic mode. Another more effective method to improve the start-up performance takes the nonlinearity into consideration within the design of the controller. The control law obtained improves the start-up performance of the controller by reducing overshoots and oscillations. This technique is elaborated upon in the following discussions.

2.2 Classification of Nonlinearities

The classification of nonlinearities is determined by the way how they appear in the control system or by their nature and structure. The former ones, the so-called system nonlinearities may be classified as inherent or intentional. From the latter ones we will mention only the saturation type nonlinearity due its most wide presence in a typical control loop.

2.2.1 Inherent nonlinearities

These nonlinearities exist within the *'presumably linear'* components selected to perform a function other than a nonlinear one in a system. Quite often the control system designer would be much happier if inherent nonlinearities do not exist at all. Typical examples of these nonlinearities are saturation and dead zone in a valve.

2.2.2 Intentional nonlinearities

Intentional nonlinearities are deliberately introduced into a system to improve its performance. For instance, the inclusion of a nonlinearity into a control loop may have some of the following positive effects on system performance like:

- compensation for the effects of nonlinearities occurring in the valve, usually referred to as nonlinear compensation (Chesmond, 1990, Atherton, 1982);
- a decrease in loop sensitivity (Chesmond, 1990, Elgerd, 1967);
- a minimization of loop settling time following a disturbance or setpoint change (Chesmond, 1990);
- a reduction in the effect of interactions with other loops (Chesmond, 1990).

2.2.3 The Saturation Nonlinearity

The inherent effects of nonlinearities such as saturation could be removed from the operating range of a plant by intentionally utilizing this nonlinearity as part of the control structure (Glattfelder & Schaufelberger, 1986). Generally speaking, including this nonlinearity within the control loop in the proper way results in a reduction in the magnitude of the overshoot. A short descriptive analysis of the saturation nonlinearity shown in Fig.2-1 and its gain characteristic as shown in Fig.2-2 follows.

The output of the actuator is bounded by its upper limit ($U_{\max}$) and lower limit ($U_{\min}$) to which correspond the input control signal values $u_{c\max}$ and $u_{c\min}$, respectively. The

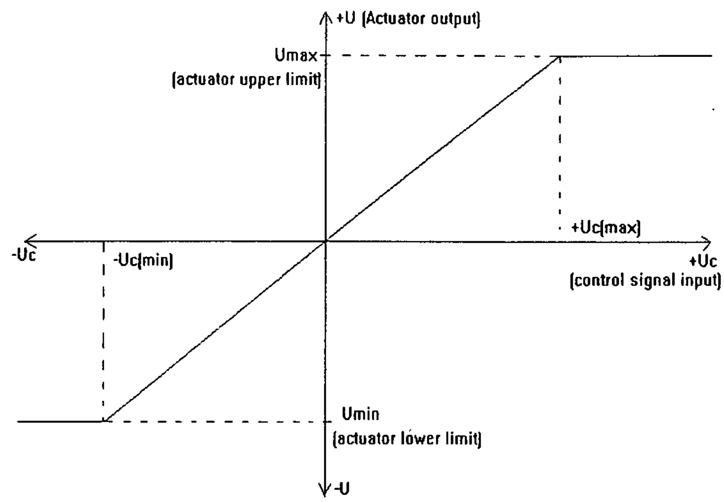

Figure 2-1: Saturation nonlinearity

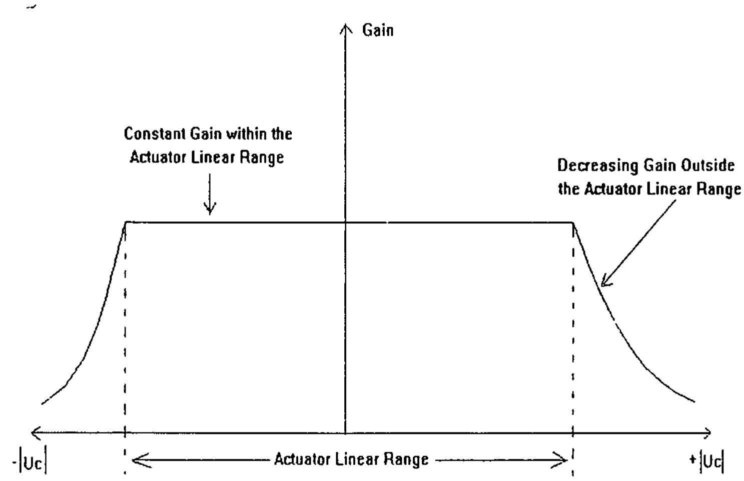

Figure 2-2: Gain characteristics for saturation nonlinearity

input control signal u_c may assume values beyond these limits. These limits fix the linear range of the control signal within the bounds of $u_{c\,\min} \leq u_c \leq u_{c\,\max}$. Within this linear range the actuator output is proportional to the control signal input and the gain is given by $U_{\max}/U_{c\,\max}$. Outside this linear range the ratio of output to input (and hence the gain) starts to decrease.

2.3 Integral Windup

The dynamic range of an actuator is bounded by its upper and lower limits, i.e. a valve will saturate if it is fully open or fully closed. In the controller shown in Fig.2-3 integral action will have to cease upon actuator saturation, otherwise the integration part of the controller will continue integrating the error signal. Any integral control signal beyond valve saturation is useless since it calls for a motion which the valve cannot produce. The continuous integration of the error signal, after the valve has saturated, results in an accumulation of the integral component ('winding up'), which must be removed later ('unwinding'). This 'winding' action of the integral part of the controller is called *integral windup* and leads to substantial overshoot and an increase in the loop settling and recovery times.

2.4 Antiwindup Schemes

There are many different *antiwindup schemes* that are common in use. In what follows we will mention some of them:

- **Disabling the integrator** (Franklin *et al.*, 1994): The actuator output is fed back positively via the integrator to generate the integral control signal u_i. The integral action is disabled by breaking the positive feedback loop before the actuator limits $U_{\min}$ and $U_{\max}$ are reached.

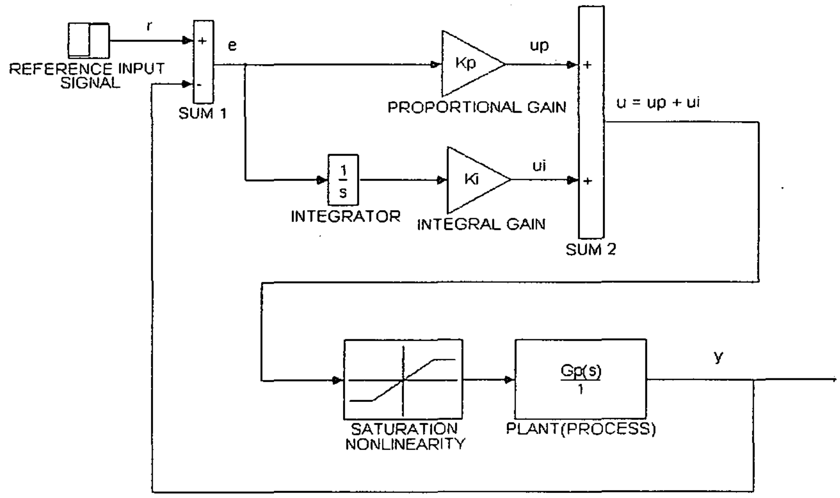

Figure 2-3: PI-controller with saturation nonlinearity and no windup protection

- **Clamping the integrator upon actuator saturation** (Walgama & Sternby, 1990): This technique involves clamping (stopping) the integrator before the control signal goes beyond the upper or lower saturation limits of the actuator. A major drawback of this method is that the control signal may be held in saturation as soon as the integral control action stops. This leads to severe overshoots in the output of the process. A solution to these overshoots is to use an *'intelligent integrator'* (Krikelis, 1980). An 'intelligent integrator' consists of a dead-zone nonlinearity in the feedback loop to 'turn off' the reset action as soon as integration exceeds a user specified limit (Krikelis, 1980).

- **Conditioning** (Walgama *et al.*, 1992): In this method the setpoint signal is modified to ensure consistency in the internal states of the controller when the valve saturates. The modified setpoint signal generates a new control error signal which adjusts the controller output until the effect of the nonlinearity is removed.

- **Resetting the integrator at saturation** (Walgama & Sternby, 1990): Resetting the integrator is achieved by connecting the feedback of the antiwindup scheme in a manner that makes it possible for feedback to act on the integrator input when the actuator saturates. A detailed discussion of this technique follows.

2.4.1 Resetting the Integrator at Saturation

Antiwindup measures have been analyzed by Hanus (1980) and Glattfelder & Schaufelberger (1986). Controllers implementing these measures correspond to one of the 'standard industrial analogue regulator structures' (cf. Glattfelder & Schaufelberger, 1986) in which the antiwindup circuit feedback acts on the integrator input when actuator saturation occurs. Early reference to this method is the work of Vandenbussche (1975). Good discussions are given by Åstrom (1981), Åstrom & Wittenmark (1984) and Åstrom & Hagglund (1988). The control action of a controller using this method of reset windup protection is governed by the following algorithm

$$u_c(t) = K_p e(t) + K_i \int_0^t e(s)ds - K_w \int_0^t e'(s)ds$$

where $e' = u_c(t) - u(t)$, with $u_c(t)$ and $u(t)$ representing the outputs of the controller and actuator, respectively. K_w is the nonlinear gain of the antiwindup module. This control law is implemented in the PI controller shown in Fig.2-4.

The operation of this controller may be explained as follows:

Roughly speaking, any mismatch between the controller output $u_c(t)$ and the actuator output $u(t)$ can be interpreted as windup. To reduce the occurrence of windup, an additional feedback path is provided by measuring the actuator output and forming a *feedback error signal* (e'). The purpose of e' is to modify the reset action till the controller output signal falls within the linear operating region of the actuator. As soon as the control signal is within the linear operating range of the valve, the feedback error signal will drop to zero and windup is eliminated. A simple time analysis of this antiwindup measure implemented in the controller shown in Fig.2-4 is as follows.

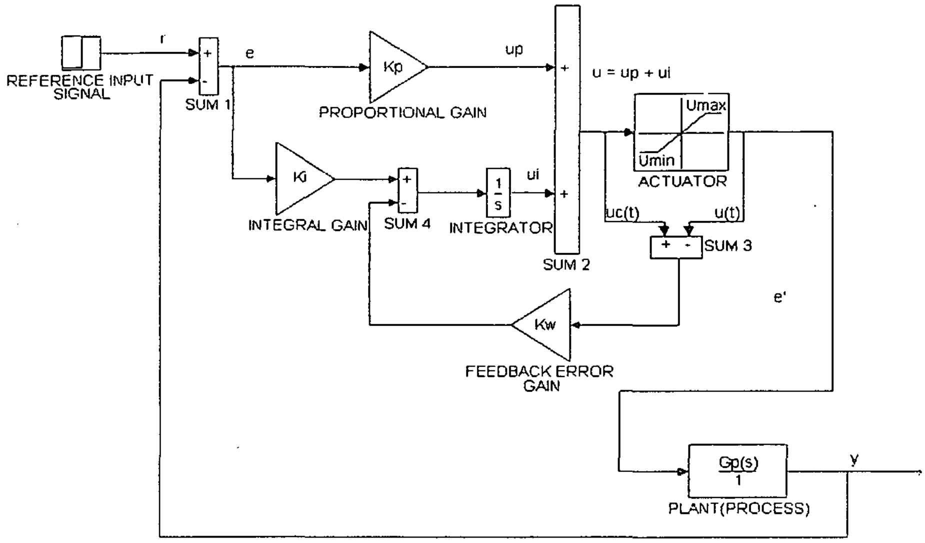

Figure 2-4: A PI-controller with saturation nonlinearity of the process and reset windup protection

When the output of the valve is within its linear range of operation, i.e. $U_{\min} < u(t) < U_{\max}$, the signal generated by the antiwindup module will be 0 since $u_c(t) - u(t) = 0$.

Let $U_{\min} = -U_{\max}$. If the output of the controller (u_c) is driven beyond the actuator's linear range of operation, say to $u(t) > U_{\max}$, then

$$u_c(t) = K_p e(t) + K_i \int_0^t e(s)ds - K_w \int_0^t [u_c(s) - U_{\max}]ds$$

$$u_c(t) = K_p e(t) + K_i \int_0^t e(s)ds - K_w \int_0^t u_c(s)ds + K_w \int_0^t U_{\max}ds$$

$$u_c(t) + K_w \int_0^t u_c(s)ds = K_p e(t) + K_i \int_0^t e(s)ds + K_w \int_0^t U_{\max}ds$$

Differentiating both sides one gets

$$Du_c(t) + K_w u_c(t) = K_p De(t) + K_i e(t) + K_w U_{\max}$$

and dividing throughout by K_w we get

$$\begin{aligned}
\frac{1}{K_w} \times Du_c(t) + u_c(t) &= \frac{K_p}{K_w} \cdot De(t) + \frac{K_i}{K_w} \cdot e(t) + U_{\max} \\
&= \frac{K_i}{K_w} \left[\frac{K_p}{K_i} \cdot De(t) + e(t) \right] + U_{\max}
\end{aligned}$$

So, if $|K_w| \to \infty$ then $u_c(t) = U_{\max}$. This corresponds to an ideal antiwindup model because the output of the controller will be within the linear operating region of the actuator for very large values of feedback error gain K_w.

The antiwindup scheme proposed has one shortcoming in that it is impractical to measure the output of a valve when it functions within its operating environment. This makes it difficult to implement the proposed antiwindup scheme. To overcome this problem, commercial controllers are now designed with an antiwindup mechanism that incorporates a mathematical model of the actuator nonlinearity where the user can define the assumed controller saturation limits $U_{\min}$ and $U_{\max}$. The choice of $U_{\min} \geq U'_{\min}$ and $U_{\max} \leq U'_{\max}$ in this scheme shown in Fig.2-5, where $U'_{\min}$ and $U'_{\max}$ denote the lower and upper saturation limits of the actual valve respectively, ensures that it functions in a similar manner to the one that excludes the actuator model mentioned previously in Fig.2-4.

2.5 Experiment to Illustrate the Effect of Integrator Windup

2.5.1 Preliminaries

Simulation studies were conducted to test the response of a PI controller under conditions where:

- a saturation nonlinearity was present in the loop and the controller used had no integral windup protection;
- an antiwindup scheme was fitted to the controller and it was then tested in a loop having a saturation nonlinearity.

The control algorithm for the linear controller is taken in the form of a simple PI controller

$$u_c(t) = K_p e(t) + K_i \int_0^t e(s)ds$$

while the law governing the nonlinear PI controller having an antiwindup scheme remains the same as previously mentioned, i.e.

$$u_c(t) = K_p e(t) + K_i \int_0^t e(s)ds - K_w \int_0^t e'(s)ds$$

All tests were conducted on an one-degree-of-freedom controller in a unity feedback system having the controller and the process series-connected in the feedforward branch. The reference input signal applied to the system is $r(t) = M_r h(t-10)$, where h is a step function. The actuator saturation limits are set to ± 1.

2.5.2 Purpose of the experiment

The purpose of this simulation experiment was to show that anti-windup measures may have a positive effect on setpoint tracking in loops containing a saturation nonlinearity.

2.5.3 Method used

Consider a process having a transfer function of

$$G_p(s) = \frac{1}{s}$$

The rest of the control system is as previously described. The structure of the PI controller with antiwindup mechanism is shown in Fig.2-5.

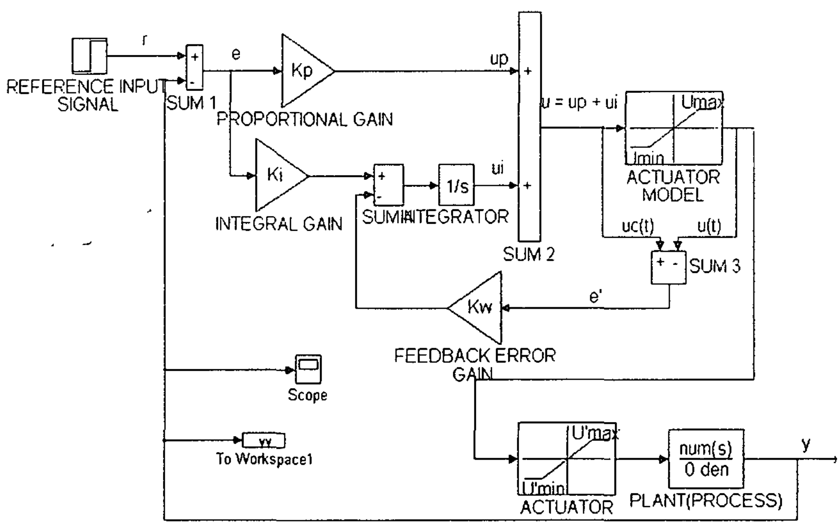

Figure 2-5: A PI-controller with reset windup protection scheme using an actuator model

The values of controller parameters used in these experiments are given in Table 2.1.

Parameters	K_p	K_i	K_w
with saturation nonlinearity and no windup protection	1.8	1	0
with saturation nonlinearity and windup protection	1.8	1	5.75

Table 2.1: Simulation parameters used in the experiment

2.5.4 Observation

From the responses shown in Fig.2-6, it is clear that control loops having an actuator and no windup protection experience large overshoots as a result of integral windup upon actuator saturation. Antiwindup improves setpoint tracking by reducing the magnitude of the maximum overshoot.

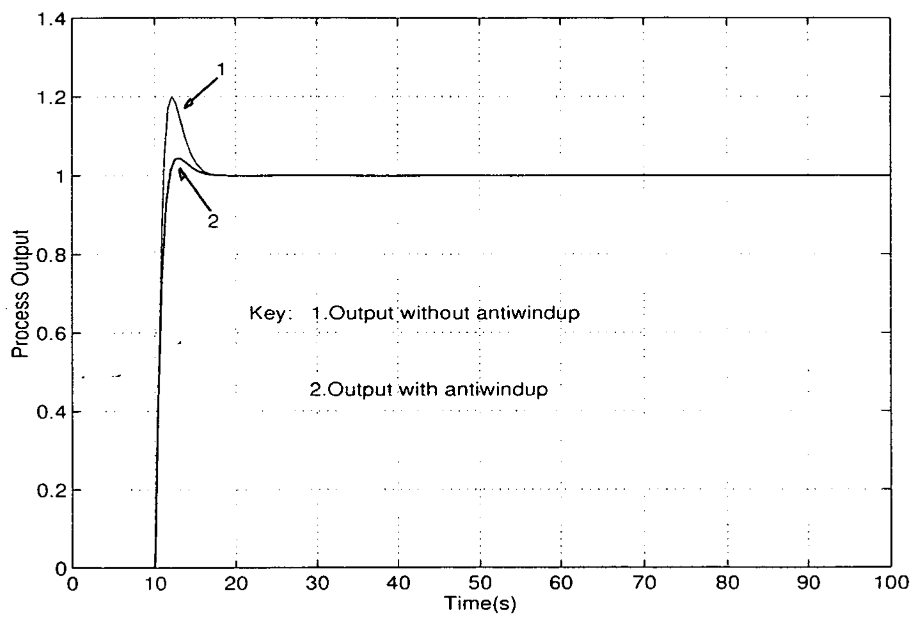

Figure 2-6: Effects of linear control with and without windup

2.6 Conclusion

A number of antiwindup schemes have been commented upon. Inclusion of any of these schemes within the structure of a conventional controller will alter the control algorithm, leading to a reduction in overshoot within loops having a saturation type nonlinearity.

Since one of the most common nonlinearities present in nearly all control loops is actuator saturation, including antiwindup as an option for the linear controller will ensure a versatility in the controller applications (i.e. the antiwindup module will make it possible to use the same controller in applications where actuator saturation is present). The implementation of the antiwindup mechanism to linear PID controllers is useful. However, even in the case of nonlinear controllers, as will be shown further in the text, their positive effects remain.

Chapter 3

Nonlinear Gain Scheduling PID Controller

3.1 Introduction and general background

Conventional PID controllers are simple but cannot always effectively control systems with loops prone to changing parameters (Åstrom & Hägglund, 1988). Practical problems inherent to process loops are usually the source of control performance deterioration.. The source of these problems could be attributed to the simultaneous need for good setpoint tracking and disturbance rejection, the presence of nonlinearities such as actuator saturation and static process gain, as well as process dead-time.

Nonlinearities and disturbances within a loop may lead to oscillations, overshoots or otherwise unacceptable behavior of the loop. Even with the optimal tuning of the linear conventional PID controller, the dynamic response of the plant with the controller in the loop will be adversely affected by the presence of nonlinearities and disturbances.

The dynamics of a control loop impose severe constraints on the controller gain parameters. For instance, limitations on the magnitude of the proportional gain (K_p) are necessary to prevent oscillations and the saturation of nonlinear components (such as actuators). These constraints on the magnitude of K_p leads to a smaller range in acceptable values of K_p, the consequence of which is that system reaction speed cannot be increased as may be required, i.e. the rise time, the loop settling time and the loop recovery time may not be reduced as might be necessary. Control loops that include an actuator nonlinearity may frequently become unstable if no limits are placed on the controller's integral gain. Introducing derivative action into the control loop contributes to the controller output with a portion of the signal proportional to the derivative of the error. The consequence of this derivative action is an overall reduction in the system reaction speed. This has a damping effect on changes in the controlled variable (cf. Åström & Hägglund, 1988, Martins de Carvalho, 1993, Shinskey, 1979). The presence of derivative action relaxes the constraints on K_p. When derivative action is present in the controller, K_p can be larger and still ensure stability. This increased proportional action improves system response speed. A conflicting situation of simultaneously satisfying loop reaction speed (by adjusting K_p) whilst maintaining minimal overshoot (by adjusting K_d) in the face of nonlinearities reduces the possibility of an easy optimal tuning of the loop.

In the discussion that follows, a possible alternative method is proposed to reduce the negative effects of compromised tuning, and to take into account the antiwindup measures and dead-time compensation via a Smith predictor (Smith, 1958) mechanism. This method is based on a nonlinear controller utilizing a nonlinear gain scheduling mechanism that has adjustable nonlinear gains (Bajić, 1994, Bajić *et al.,* 1995) and represents further enhancements of these results, extending them to cope with integral windup and long dead-times of the process.

3.2 Nonlinear Gain Scheduling of PID Control

A possible conventional nonlinear PID controller (see Bajić, 1994) that use a gain scheduling mechanism consists of three independent P, I and D actions with adjustable nonlinear gains. This nonlinear controller changes its gains according to nonlinear laws which are discussed later on. The scheme is derived originally from fuzzy logic presented in Zhao

Linguistic terms	**Symbols**
Big	B
Medium	M
Small	S
Positive Big	P
Positive Medium	PM
Positive Small	PS
Negative Big	NB
Negative Medium	NM
Negative Small	NS
Zero	Z

Table 3.1: Table of linguistic terms

Positions	**1**	**2,4,6**	**3**	**5**
error(e)	PB	NS, Z, PS	NB	PM
K_p	B	S	B	M
K_i	B	S	B	M
K_d	S	B	S	M

Table 3.2: Tuning rules for the nonlinear PID controller

et al.(1993) that captured a possible human perception and thinking processes in order to determine suitable the actions for the PID controller. To explain the background for the derivation of the nonlinear gains proposed in Bajić (1994) and Bajić *et al.* (1995), we will follow the logic of Zhao *et al.* (1993). In what follows, linguistic terms are used to indicate the magnitude of the error and the nonlinear gains. These terms are shown in Table 3.1:

The set of possible fuzzy rules associated with the PID controller utilizing fuzzy gain scheduling is given in Table 3.2 (see Zhao *et al.*, 1993).

The values given in Table 3.2 are used to denote the magnitude of the three nonlinear gains of the controller for different excursions of the error signal. This fuzzy-logic setting is converted to the ordinary nonlinear functions framework by Bajić (1994) and Bajić *et al.* (1995) that give the following possible explanation of Table 3.2 based on the step response in Fig.3-1 and the graphs shown in Figures 3-2 to 3-4

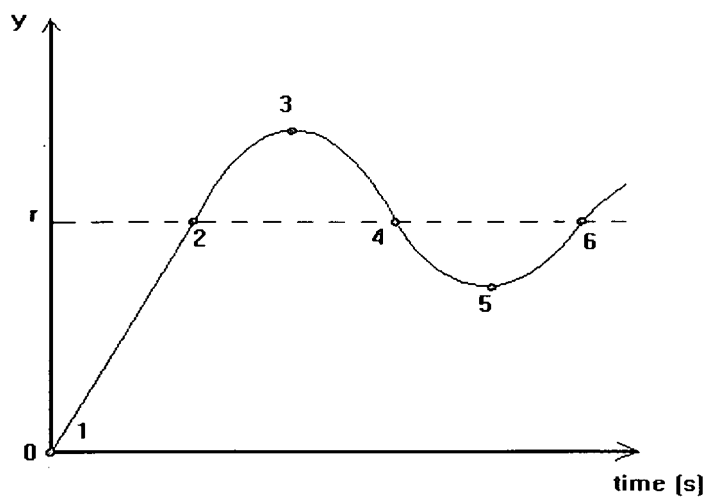

Figure 3-1: Step response used for derivation of control laws

- *Position 1:*: For a positive-big error, the nonlinear proportional and integral gains are big and the nonlinear derivative gain is small;
- *Positions 2, 4, 6:* When the error is close to zero, the nonlinear proportional and integral gains are small while the nonlinear derivative gain is big;
- *Position3:* For a negative-big error, the nonlinear proportional and integral gains are big and the nonlinear derivative gain is small.
- *Position 5:* When the magnitude of the error is positive-medium, all controller gains are set to medium.

These nonlinear characteristics for proportional, integral and derivative control are captured in the following control laws as proposed by Bajić (1994) and Bajić *et al.* (1995):

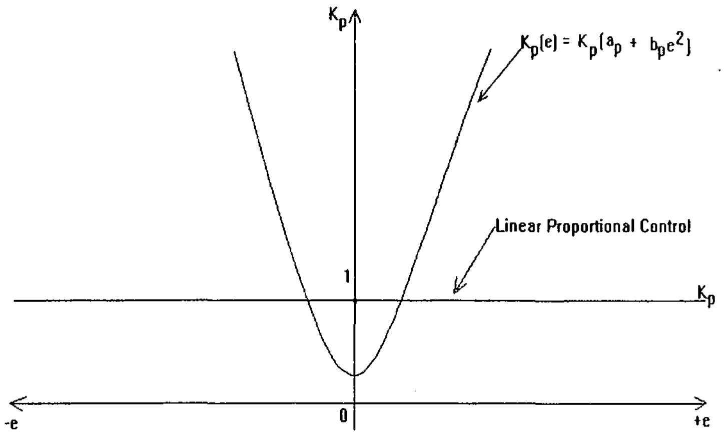

Figure 3-2: Nonlinear proportional gain

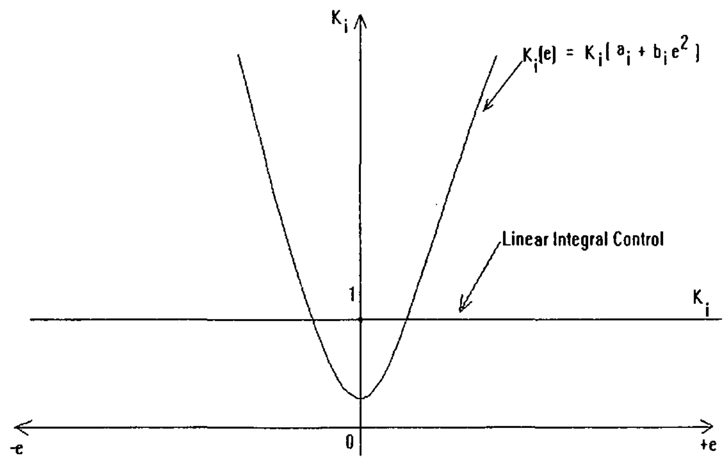

Figure 3-3: Nonlinear integral gain

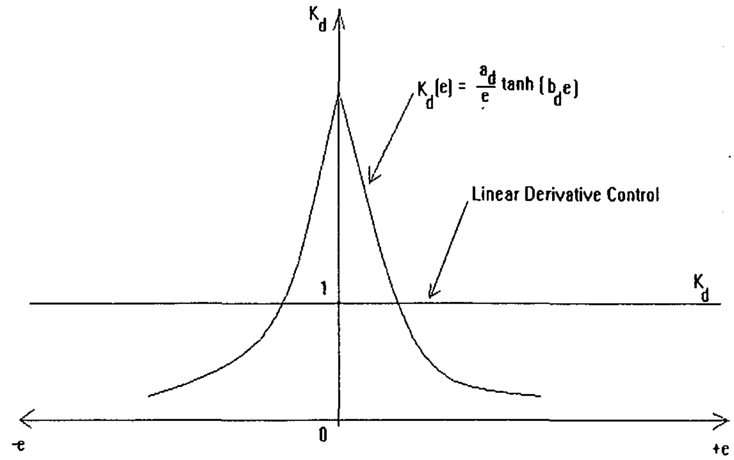

Figure 3-4: Nonlinear Derivative Law

$$u_p(e) = K_p(a_p e + b_p e^3) = K_p(a_p + b_p e^2)e = K_p f_p(e)e \tag{3.1}$$

$$u_i(e) = K_i \int_0^t (a_i e + b_i e^3)dt = K_i \int_0^t (a_i + b_i e^2)edt = K_i \int_0^t f_i(e)edt \tag{3.2}$$

$$u_d(e) = K_d \frac{d}{dt}[a_d \tanh(b_d e)] = K_d \frac{d}{dt} f_d(e) \tag{3.3}$$

where the integral gain $K_i = \dfrac{K_p}{T_i}$, derivative gain $K_d = K_p T_d$, and u_p, u_i and u_d represent the contributions of the proportional, integral and derivative components respectively. If the bracketed terms contained in u_p and u_i approach unity and the bracketed expression for u_d approaches the error e, the controller will behave as a linear controller. The adjustable parameters of nonlinear gains for each mode of control are denoted by $a_{p,i,d}$ and $b_{p,i,d}$. The behavior of this nonlinear controller will be governed by

$$u(t) = K_p(a_p + b_p e^2)e + K_i \int_0^t [(a_i + b_i e^2)e] ds + K_d \frac{d}{dt}[a_d \tanh(b_d e)] \tag{3.4}$$

These nonlinear laws are implemented in the controller structure shown in Figure 3-5

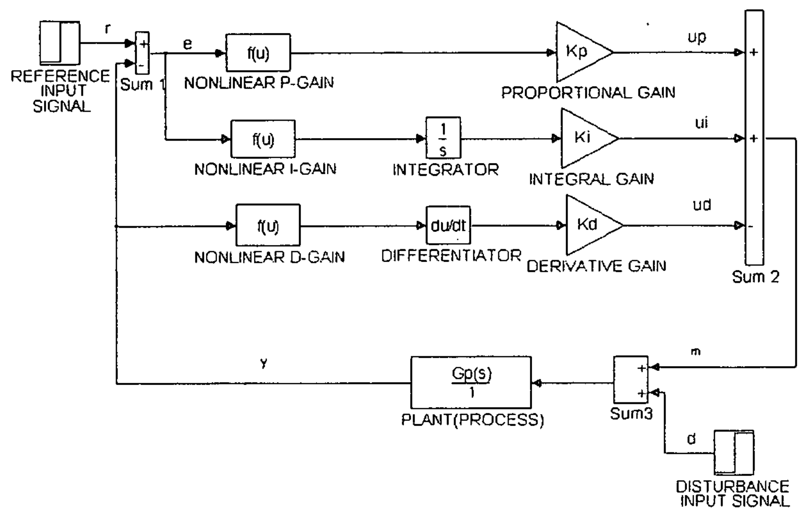

Figure 3-5: Loop with nonlinear PID controller having gain scheduling

3.2.1 Control of Loops that contain Saturation Nonlinearity

The presence of a saturation element such as an actuator in a control loop can lead to overshoot. This overshoot is a result of integrator windup when the actuator saturates. The causes and effects of integrator windup, and different anti-windup schemes to reduce its negative effects, were discussed in Chapter 2. The nonlinear controller with gain scheduling and with antiwindup mechanism implemented will behave according to

$$u(t) = K_p(a_p + b_p e^2)e + K_i \int_0^t [(a_i + b_i e^2)e] ds - K_w \int_0^t e'(s) ds + K_d \frac{d}{dt}[a_d \tanh(b_d e)] \tag{3.5}$$

where K_w is the gain of the feedback error e' contained in the antiwindup scheme. The block diagram of the controller implementing algorithm (3.5) is shown in Fig.3-6.

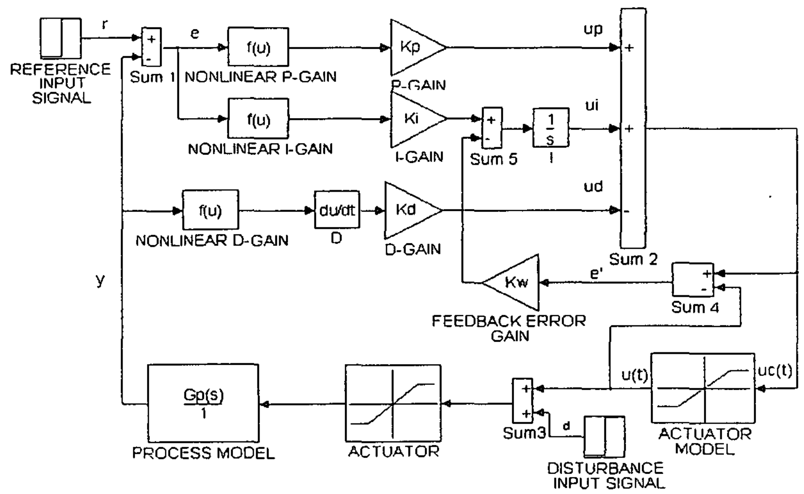

Figure 3-6: Loop with nonlinear PID controller with gain scheduling and integral windup protection

3.2.2 Control of Processes having Long Dead-time

Almost all industrial loops experience dead-time when a signal is transmitted from the process input to the process output. For example, if liquid flows through a 100 meter long pipe at a velocity of 5 meters per second, the transport delay will be 20 seconds, i.e. it will take 20 seconds for the liquid to flow through the pipe. If a control system is to maintain a certain composition of this liquid at the outlet of the pipe, there will be a 20 second delay between a control action being taken and the result being measured. This delay is referred to as dead-time and may lead to loop instability.

Dead-time may be especially troublesome in nonlinear controllers. The nonlinear characteristics of these controllers could increase their sensitivity to changes occurring within the loop, which when combined with process dead-time can easily produce instability. A Smith predictor control scheme (Smith, 1958) can be used to compensate for the negative effects of dead-time. Systems with Smith predictor control make use of an additional feedback loop containing approximated mathematical models of the process with dead-time. The compensatory effect of this feedback makes the controller to act on a model of the process when process dead-time is eliminated. If Smith predictor control is included within the nonlinear controller scheme governed by control law (3.4), then the algorithm for this nonlinear controller will contain the following operator equations:

$$e = r - (y - y_{m1} + y_{m2}) \tag{3.6}$$

$$y_{m1} = G_{m1}(s)\exp(-L_m s)(m) \tag{3.7}$$

$$y_{m2} = G_{m2}(s)(m) \tag{3.8}$$

$$y = G_p(s)\exp(-L_p s)(m + d) \tag{3.9}$$

where $G_p(s)$ denotes the process transfer function, $G_{m1}(s)$ and $G_{m2}(s)$ are the models used in the Smith predictor scheme, the output y_{m1} cancels the actual process output y, feedback signal y_{m2} is the output of the process model excluding the negative effects of dead-time, L_p is the loop transport delay and L_m is a model of the dead-time within the Smith predictor compensator. The nonlinear controller using a Smith predictor compensator is shown in Figure.3-7.

3.2.3 Control of Loops having Saturation Nonlinearity and Process Dead-time

The step response of many control loops experiencing the negative effects of actuator saturation and process dead-time may be improved by using a nonlinear controller with integral windup protection and Smith predictor compensation. The behavior of this nonlinear controller will be determined by (3.5) and operator equations (3.6)-(3.9). A block diagram of the nonlinear controller with anti-windup and Smith predictor compensation is shown in Fig.3-8.

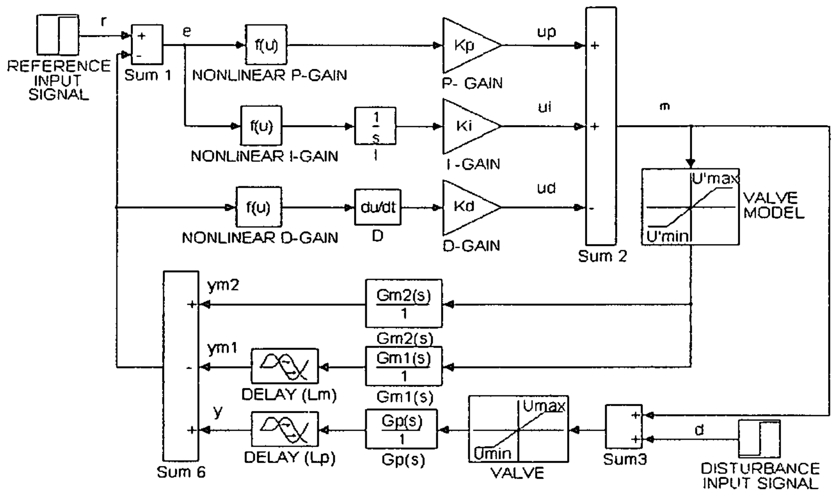

Figure 3-7: Nonlinear PID controller with gain scheduling and Smith predictor compensator

The control algorithms for the conventional linear PID controller used in the subsequent experiments in Chapter 4, were mentioned in Chapter 2 and are repeated here

$$u(t) = K_p e(t) + K_i \int_0^t e(s)d(s) + K_d De(t) \tag{3.10}$$

With antiwindup included the controller will be governed by

$$u(t) = K_p e(t) + K_i \int_0^t e(s)d(s) - K_w \int_0^t e'(s)d(s) + K_d De(t) \tag{3.11}$$

If windup protection and Smith predictor control is used in the conventional PID controller, then its operation will be determined by a combination of algorithm (3.11) and the one given by equations (3.6)-(3.9).

3.3 Conclusions

Inherent undesirable loop nonlinearities and dead-time of the process are the main factors responsible for contributing towards the deterioration in the performance of a control system. The control performance of the classical PID controller is sensitive to the presence

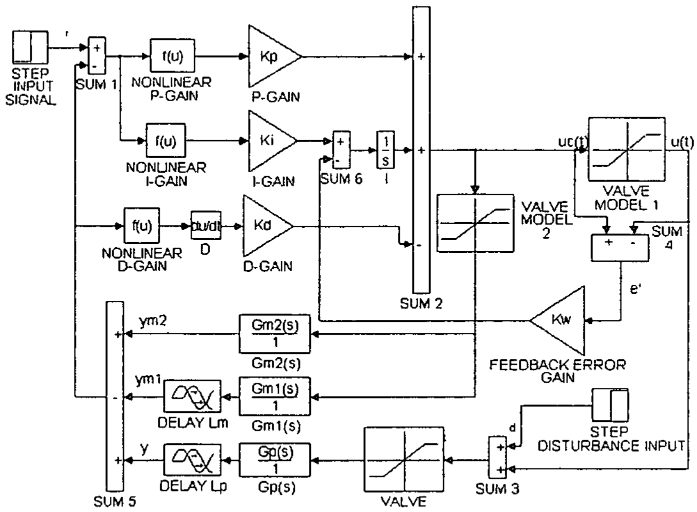

Figure 3-8: Nonlinear PID controller with gain scheduling, Smith predictor compensation and antiwindup

of such components within the loop. This sensitivity usually results in an unsatisfactory behavior of the process loop. The nonlinear gain scheduling in a PID controller combined with an antiwindup mechanism and Smith predictor schemes is proposed in this chapter for the purpose of improving the setpoint tracking properties of the loop. The exact stability analysis of the control loop containing the proposed nonlinear controller requires complicated qualitative analysis of nonlinear time-dealyed differential equations, and for this reason will not be discussed in this text. However, although such an analysis will not be given, the next chapter will use experiments to show that the proposed controller contributes strongly to the stability of control loops experiencing the undesirable effects of saturation and dead-time.

Chapter 4

Experiments with PID Controller that Has Nonlinear Gain Scheduling

4.1 Introduction

In this chapter we present results of simulation experiments conducted to compare the control performance of different variants of the PID controller having nonlinear gain scheduling developed in Chapter 3 and the performance of the PID controller without gain scheduling. The first type of PID controller we will call nonlinear, having in mind that its gains are given as nonlinear functions of the error signal, while the other one we will call linear (although it may contain antiwindup mechanism). This comparison analysis shall be made with respect to setpoint tracking and disturbance rejection, although the design developed in Chapter 3 is aimed at improving setpoint tracking and not disturbance rejection.

In all experiments a SISO negative unity feedback system, with the controller and process connected in series in the forward path of the control loop, will be considered. The performance of each controller will be tested on a hypothetical fourth order process. The process considered is self-saturating and is representative of numerous similar processes that can be found in practice. The reference input signal used in simulation is $r(t) =$

$M_r h(t - 10)$, where h represents a unit step function. A disturbance signal, $d(t) = M_d h(t - 40)$, combines with the controller output $u(t)$, so that the process input is governed by $u_{proc}(t) = u(t) + d(t)$. The process output y is fed back to the input of the controller to form the error signal e. Our comparison between the linear controller and the nonlinear controller will be done through the following experiments:

- **Experiments 4.1:** Process without dead-time and with the actuator in the loop.
- **Experiments 4.2:** Process with dead-time and without an actuator in the loop.
- **Experiments 4.3:** Process with both, dead-time and actuator in the loop.

Discussions of all these experiments will be given at the end of the chapter.

4.2 Experiments 4.1: Control of Loops Containing a Saturation Nonlinearity

4.2.1 Purpose

These experiments are chosen to demonstrate that a well tuned nonlinear controller without windup protection will yield a better control performance with respect to setpoint tracking and disturbance rejection than a well tuned linear controller. Another goal is to show that the performance of the nonlinear controller with antiwindup mechanism implemented is better that of linear one with the antiwindup structure.

4.2.2 Method

Consider the following process model

$$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$$

Parameters	M_r	M_d	K_p	K_i	K_d	$K_w = 0$
1. L(PID)	0.5	0.5	1	1.49	0.25	Fig.4-1
1. N(PID)	0.5	0.5	1	1.49	0.25	Fig.4-1
2. L(PID)	1	1	1	1.49	0.25	Fig.4-2
2. N(PID)	1	1	1	1.49	0.25	Fig.4-2
3. L(PID)	2	2	1	1.49	0.25	Fig.4-3
3. N(PID)	2	2	1	1.49	0.25	Fig.4-3

Table 4.1: Simulation parameters for Experiment 4.1 when no windup protection is used

Parameters	a_p	b_p	a_i	b_i	a_d	b_d
1.L(PID)	–	–	–	–	–	–
1. N(PID)	1.79	0.0049	1.1959	0.05	5	0.4277
2. L(PID)	–	–	–	–	–	–
2. N(PID)	1.79	0.0049	1.1959	0.05	5	0.4277
3. L(PID)	–	–	–	–	–	–
3. N(PID)	1.79	0.0049	1.1959	0.05	5	0.4277

Table 4.2: Parameters of controller gain scheduling for Experiment 4.1 when no windup protection is used

of the fourth order. This model is frequently used in simulation experiments for verification of controller performance (cf. Hagglund, 1992). The structure of the control system is as described in the previous Chapter in Section 3.2.1. The actuator limits are set at ± 3. Both controllers, linear and nonlinear, are tuned as follows. The linear PID controller is tuned according to Ziegler-Nichols open loop method. The same settings for is used for the controller with nonlinear gain scheduling and the parameters for the gain scheduling $a_{p,i,d}$ and $b_{p,i,d}$ are adjusted according to the tuning procedure as suggested by Bajić (1994) and Bajić *et al.* (1995). This tuning appears to be good for rejection of disturbances that are of the form of step-type functions. Gain K_w is adjusted to reduce overshoot caused by integrator windup. Table 4.1 and Table 4.2 list the utilized simulations parameters for the process considered. N(PID) represents the PID controller with nonlinear gain scheduling, and L(PID) stands for the linear controller. In the case when $K_w = 0$ no windup mechanism is implemented.

Parameters	M_r	M_d	K_p	K_i	K_d	$K_w = 2$
4. L(PID)	0.5	0.5	1	1.49	0.25	Fig.4-4
4. N(PID)	0.5	0.5	1	1.49	0.25	Fig.4-4
5. L(PID)	1	1	1	1.49	0.25	Fig.4-5
5. N(PID)	1	1	1	1.49	0.25	Fig.4-5
6. L(PID)	2	2	1	1.49	0.25	Fig.4-6
6. N(PID)	2	2	1	1.49	0.25	Fig.4-6

Table 4.3: Simulation parameters for Experiment 4.1 when antiwindup is used

Parameters	a_p	b_p	a_i	b_i	a_d	b_d
4.L(PID)	–	–	–	–	–	–
4. N(PID)	1.79	0.0049	1.1959	0.05	5	0.4277
5. L(PID)	–	–	–	–	–	–
5. N(PID)	1.79	0.0049	1.1959	0.05	5	0.4277
6. L(PID)	–	–	–	–	–	–
6. N(PID)	1.79	0.0049	1.1959	0.05	5	0.4277

Table 4.4: Parameters of controller gain scheduling for Experiment 4.1 when antiwindup is used

4.2.3 Observation

The results of simulations are presented for comparison in Fig.4-1 to Fig.4-6. It can be noted that the gain scheduling controller has better control effect to set point tracking and to disturbance rejection than the linear controller. Only for the large values of step-change in the reference input signal the linear controller has slightly smaller overshoot, but otherwise poorer setpoint tracking. However, with the antiwindup mechanism implemented the controller with nonlinear gain scheduling exhibits much better control performance in both cases, i.e. in setpoint tracking and in disturbance rejection.

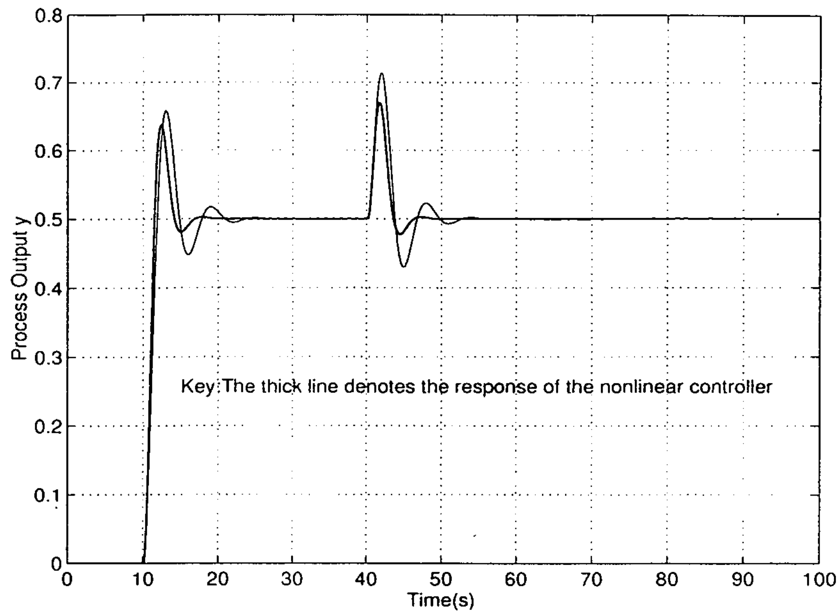

Figure 4-1: Setpoint tracking and disturbance rejection with no windup protection (Case 1)

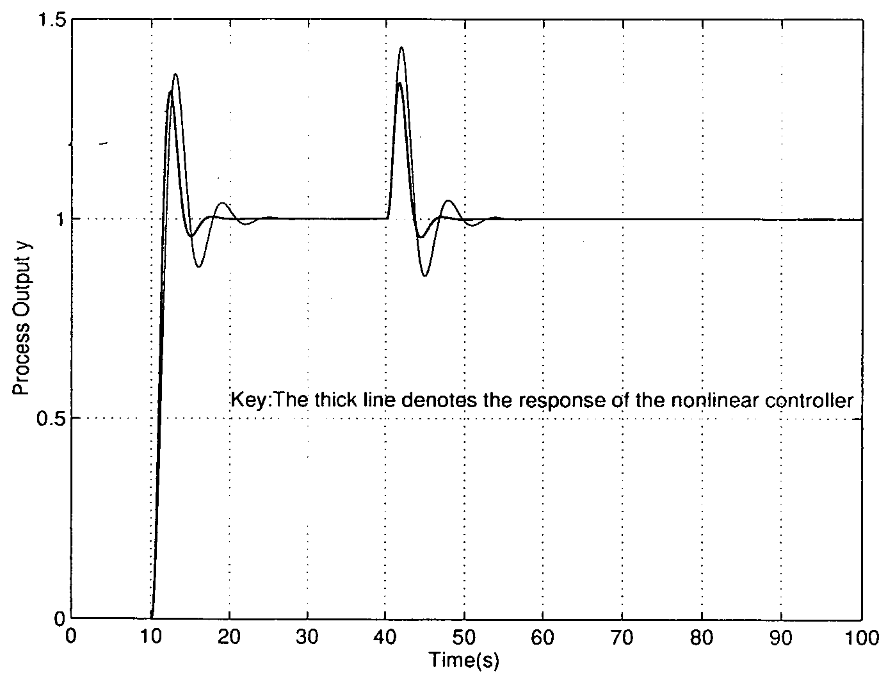

Figure 4-2: Setpoint tracking and disturbance rejection with no windup protection. (Case 2)

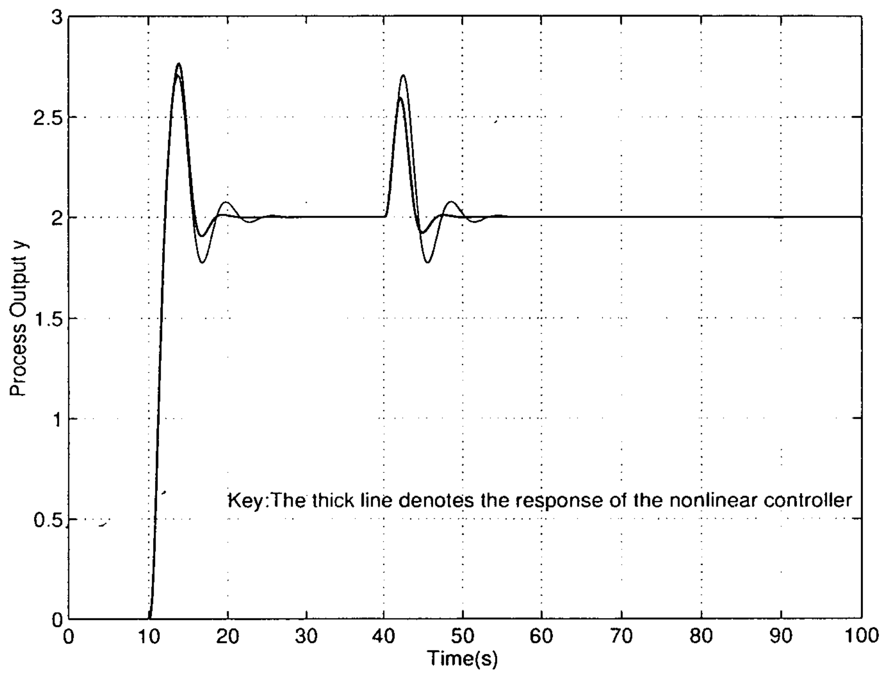

Figure 4-3: Setpoint tracking and disturbance rejection with no windup protection. (Case 3)

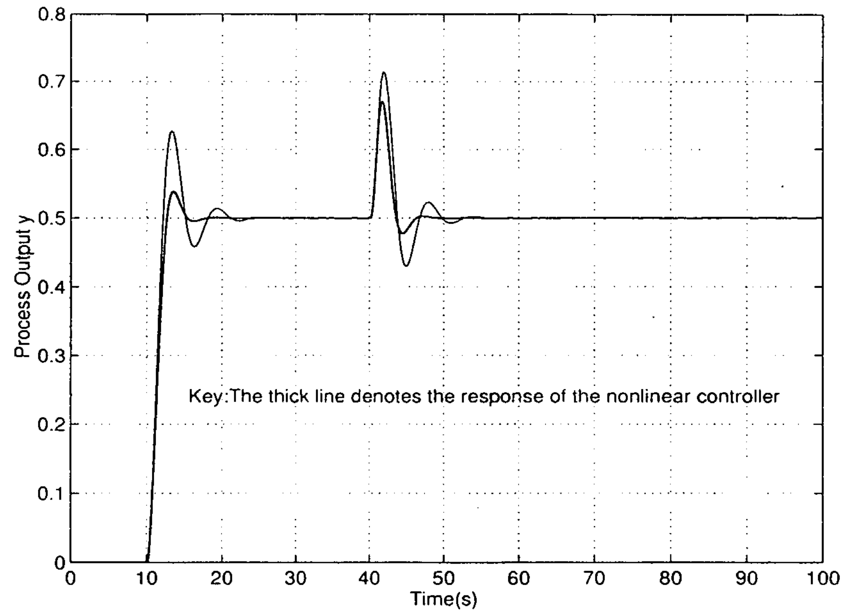

Figure 4-4: Setpoint tracking and disturbance rejection with windup protection (Case 4)

4.3 Experiments 4.2: Control of Loops Having Long Dead-Time

4.3.1 Purpose

These experiments should demonstrate that the PID controller with nonlinear gain scheduling with the Smith predictor schema implemented yields better control performance in loops having long dead-time than the linear controller with Smith predictor.

4.3.2 Method

Consider the system model $G_p(s)\exp(-L_p s)$, where G_p is the rational part of the transfer function of the process and L_p is the process dead-time. The Smith predictor control is used in the feedback loop to enable process dead-time compensation. Let us consider the process governed by

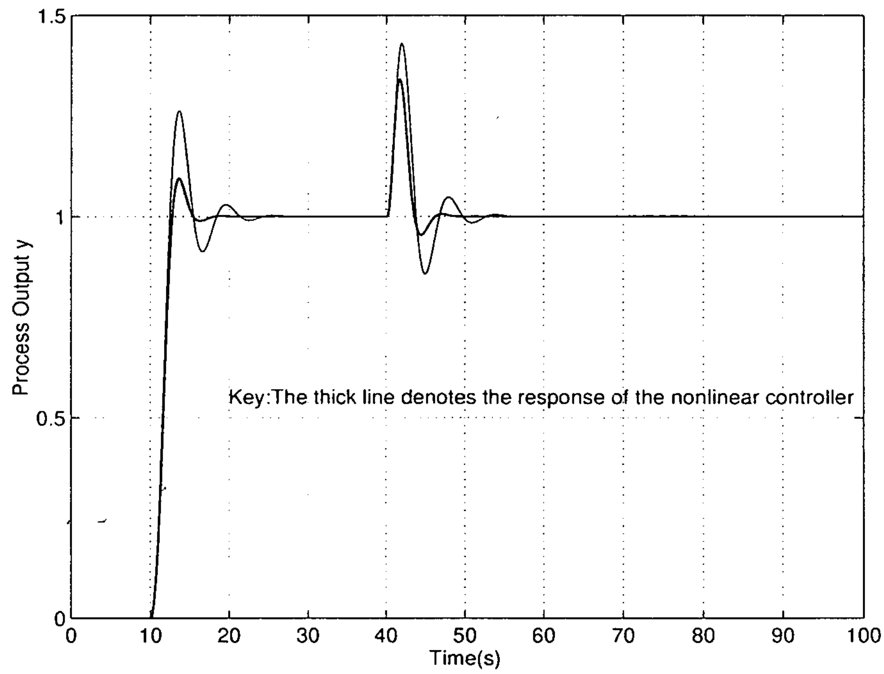

Figure 4-5: Setpoint tracking and disturbance rejection with windup protection. (Case 5)

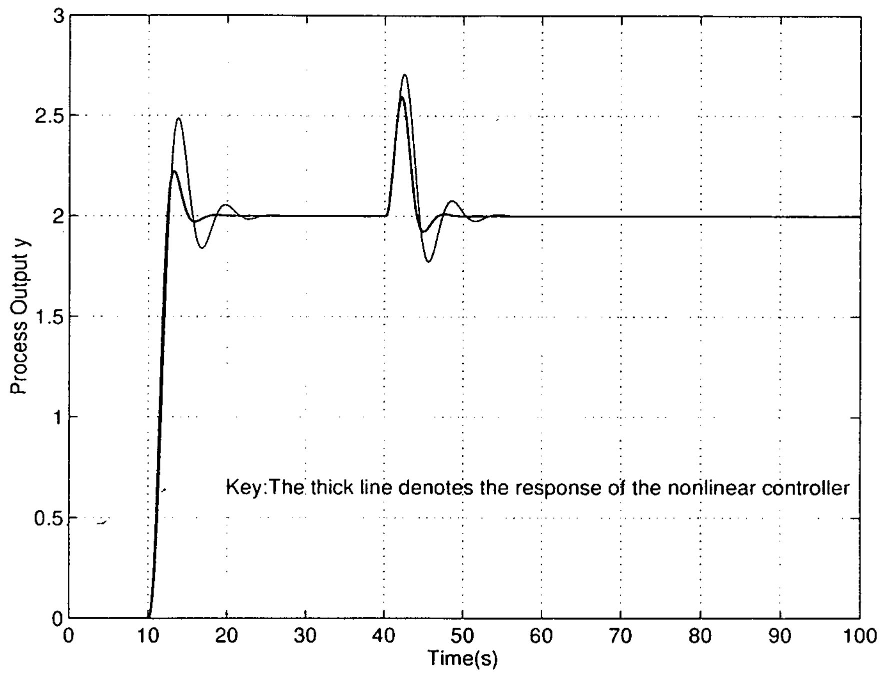

Figure 4-6: Setpoint tracking and disturbance rejection with windup protection. (Case 6)

$$G_p(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)} \exp\left(-L_p s\right)$$

The model chosen for good cancellation of the actual process output y in one parallel feedback branch of the Smith predictor is

$$\begin{aligned}
\frac{Y_1(s)}{U(s)} &= G_{m1}(s) \exp\left(-L_{m1} s\right) \\
G_{m1}(s) &= \frac{0.0365s^2 + 1.3087s + 2.8008}{s^3 + 3.972s^2 + 5.8549s + 2.8007} \\
L_{m1} &= 0.2575
\end{aligned}$$

The lower order model in the second parallel feedback branch for actual process control is determined as

$$\begin{aligned}
\frac{Y_2(s)}{U(s)} &= G_{m2}(s) \exp\left(-L_{m2} s\right) \\
G_{m2}(s) &= \frac{1.5523}{s^2 + 2.4884s + 1.5524} \\
L_{m2} &= 0.3381
\end{aligned}$$

The behaviour of the control system is governed by (3.4) and operator equations (3.6) - (3.9). The complete control scheme with the nonlinear gain scheduling controller operating in the forward path of the loop is shown in Fig.3-7. The response of both linear and nonlinear controllers will be tested using the process model defined by $G_p(s)$ in control loops containing dead-times of $L_p = 10$ [s] and $L_p = 20$ [s], respectively. Table 4.5 contains the parameter values used in the simulations, while Table 4.6 and Table 4.7 list the figures corresponding to the results of the experiment. The values of linear controller parameters are obtained optimizing disturbance rejection under the constraints

Parameters	M_r	M_d	K_p	K_i	K_d	a_p	b_p	a_i	b_i	a_d	b_d
1. L(PID)	0.2	0.2	20	20	10	-	-	-	-	-	-
1. N(PID)	0.2	0.2	20	20	10	2.45	0.0049	1.1959	0.05	4.75	0.3956
2. L(PID)	1	1	20	20	10	-	-	-	-	-	-
2. N(PID)	1	1	20	20	10	2.45	0.0049	1.1959	0.05	4.75	0.3956
3. L(PID)	2	2	20	20	10	-	-	-	-	-	-
3. N(PID)	2	2	20	20	10	2.45	0.0049	1.1959	0.05	4.75	0.3956

Table 4.5: Simulation parameters for Experiments 4.2

Parameters	L_p	Fig. no.
1. L(PID)	10	4-7
1. N(PID)	10	4-7
2. L(PID)	10	4-8
2. N(PID)	10	4-8
3. L(PID)	10	4-9
3. N(PID)	10	4-9

Table 4.6: Figures corresponding to Experiments 4.2 with Lp=10 [s]

of maximum of 20% overshoot in setpoint tracking, with the simultaneous constraints of $K_p \in [0, 20]$, $K_d \in [0, 20]$ and $K_d \in [0, 10]$. The parameters of the nonlinear gain scheduling are obtained keeping the values of the K_p, K_i and K_d as obtained for the linear controller and optimizing disturbance rejection under the constraints of maximum of 20% overshoot in setpoint tracking.

Parameters	L_p	Fig. no.
1. L(PID)	20	4-10
1. N(PID)	20	4-10
2. L(PID)	20	4-11
2. N(PID)	20	4-11
3. L(PID)	20	4-12
3. N(PID)	20	4-12

Table 4.7: Figures corresponding to Experiments 4.2 with Lp=20 [s]

Observations

The representative simulation results are shown in Fig.4-7 to Fig.4-12. From the above simulations it is clear that the system response speed for both controllers is comparable. The linear PID controller displays a poorer setpoint tracking results. In all the tests conducted, the response of the nonlinear controller in setpoint tracking with respect to maximum overshoot is better, but with a longer settling time. Disturbance rejection obtained by both controllers appear to be virtually the same.

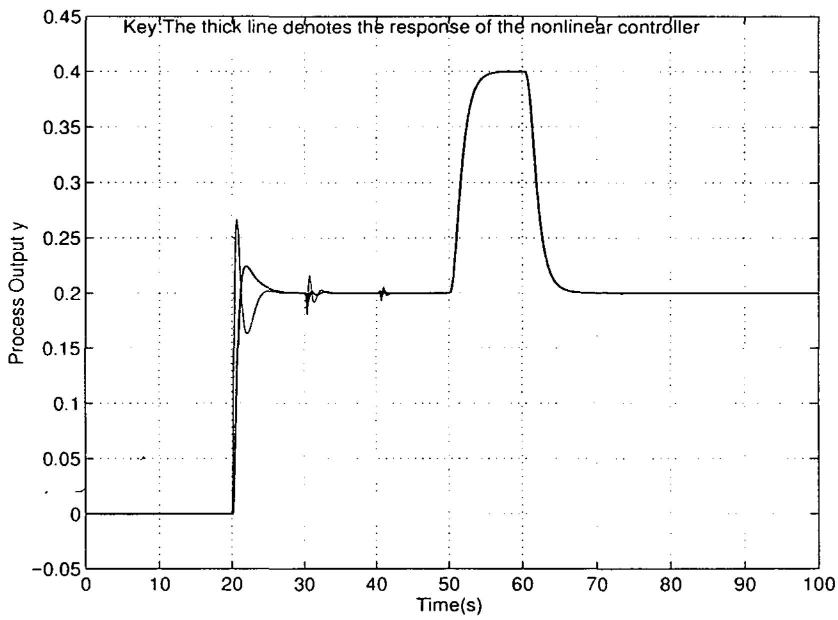

Figure 4-7: Setpoint tracking and disturbance rejection when $L_p = 10$ [s] (Case 1)

4.4 Experiments 4.3: Control of Loops Having Long Dead-Times and Saturation Nonlinearity.

4.4.1 Purpose

The purpose of these experiments is to compare the control performance of the linear PID controller and the controller with nonlinear gain scheduling in the situation when due to

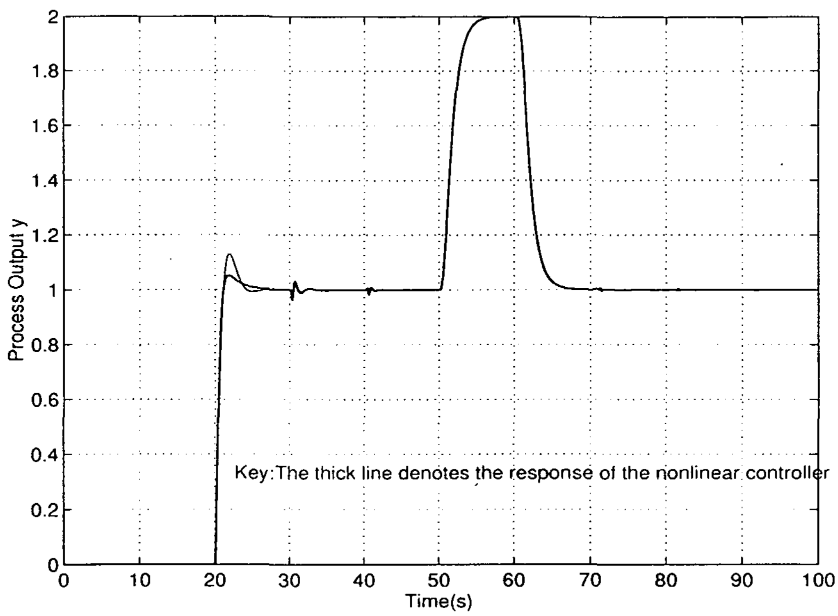

Figure 4-8: Setpoint tracking and disturbance rejection when $L_p = 10$ [s] (Case 2)

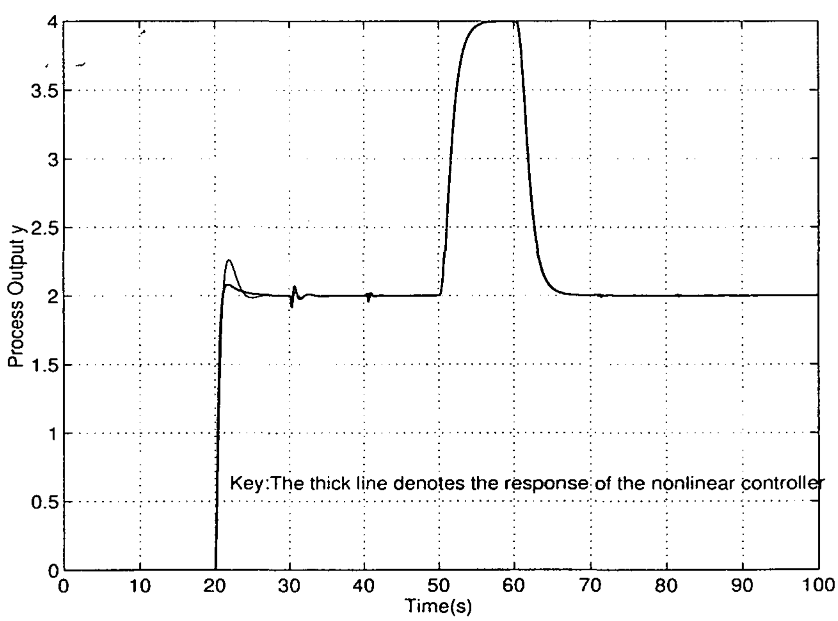

Figure 4-9: Setpoint tracking and disturbance rejection when $L_p = 10$ [s] (Case 3)

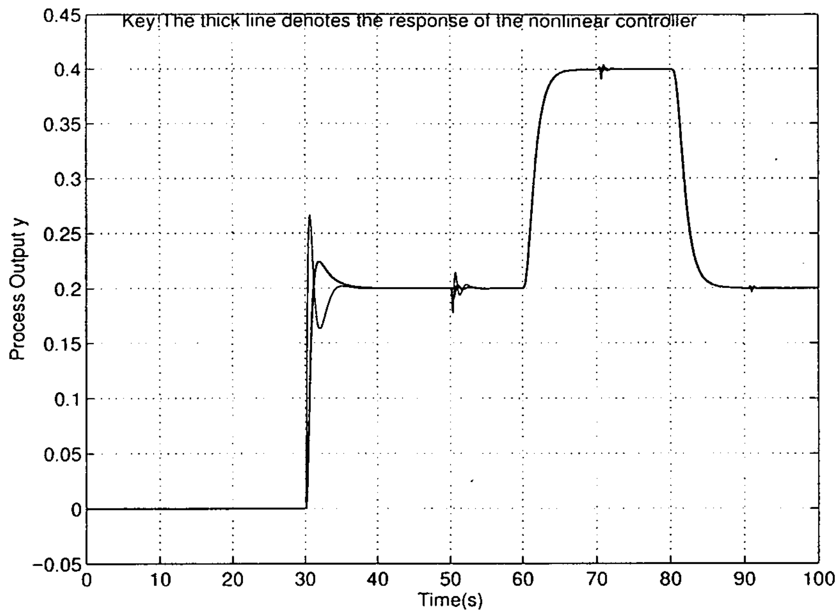

Figure 4-10: Setpoint tracking and disturbance rejection when $L_p = 20$ [s] (Case 1)

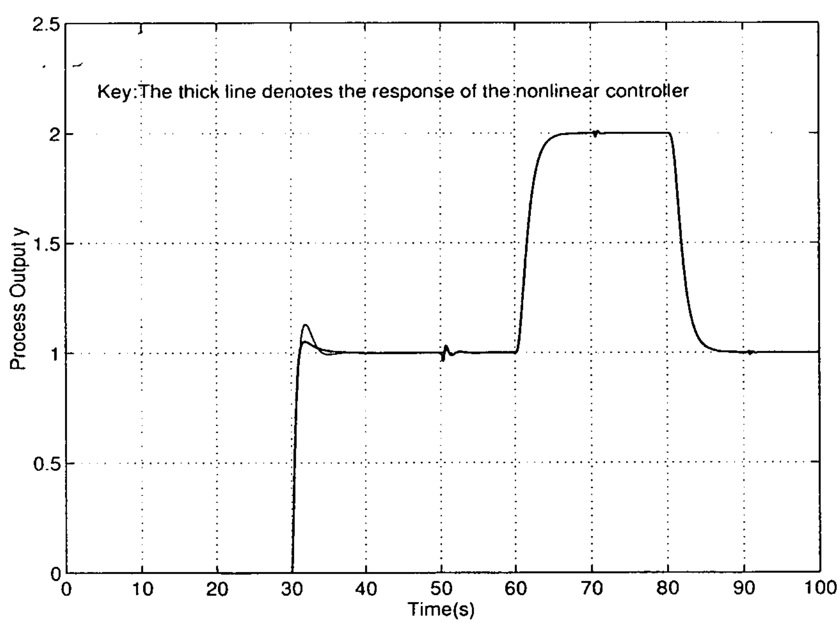

Figure 4-11: Setpoint tracking and disturbance rejection when $L_p = 20$ [s] (Case 2)

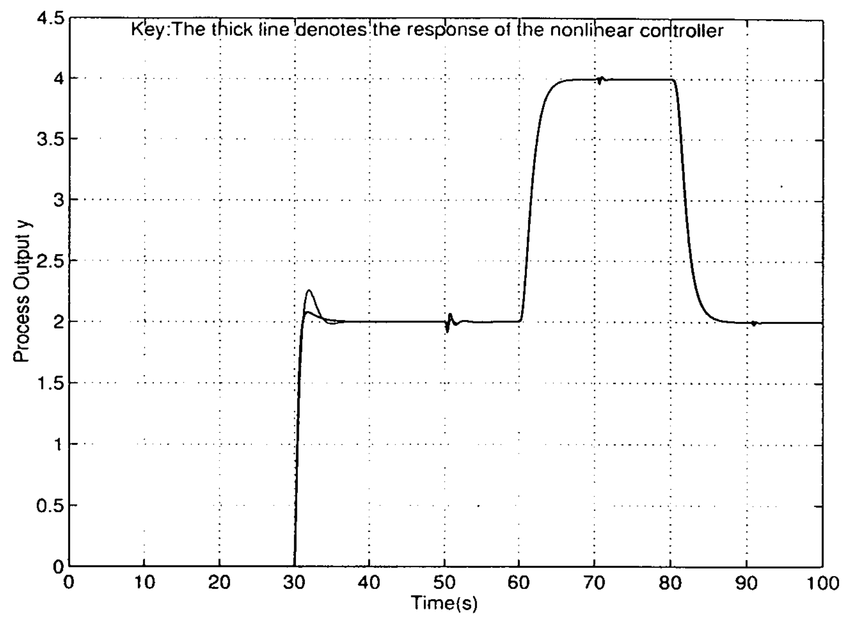

Figure 4-12: Setpoint tracking and disturbance rejection when $L_p = 20$ [s] (Case 3)

the long process dead-time the Smith predictor control is implemented and when due to the actuator saturation type nonlinearity an antiwindup mechanism is implemented.

4.4.2 Method

System models $G_p(s)\exp(-L_p s)$ from the previous experiments will be considered, with the Smith predictor control and the antiwindup mechanism. The nonlinear controller used in the simulations is governed by (3.5) - (3.9). The rest of the system is as discussed previously in Section 3.2.3. The complete control scheme is implemented in Figure 3-8. Simulation parameters used in the experiment are given in Table 4.8. The controllers tunings are the same as in Experiments 4.2. The saturation limits of the actuator nonlinearity is set to ± 10.

$K_w = 0.678$					
Parameters	M_r	M_d	K_p	K_i	K_d
1. N(PID)	1	1	20	20	10
1. L(PID)	1	1	20	20	10
2. N(PID)	2	3	20	20	10
2. L(PID)	2	3	20	20	10
3. N(PID)	4	5	20	20	10
3. L(PID)	4	5	20	20	10

Table 4.8: Parameters for Experients 4.3

Parameters	a_p	b_p	a_i	b_i	a_d	b_d
1. N(PID)	2.45	0.0049	1.1959	0.05	4.75	0.3956
1. L(PID)	-	-	-	-	-	-
2. N(PID)	2.45	0.0049	1.1959	0.05	4.75	0.3956
2. L(PID)	-	-	-	-	-	-
3. N(PID)	2.45	0.0049	1.1959	0.05	4.75	0.3956
3. L(PID)	-	-	-	-	-	-

Table 4.9: Parameters of nonlinear gain scheduling for Experiments 4.3

Parameters	L_p	Fig. no.
1. L(PID)	10	4-13
1. N(PID)	10	4-13
2. L(PID)	10	4-14
2. N(PID)	10	4-14
3. L(PID)	10	4-15
3. N(PID)	10	4-15

Table 4.10: Figures corresponding to Experiments 4.3 with Lp=10 [s]

Parameters	L_p	Fig. no.
1. L(PID)	20	4-16
1. N(PID)	20	4-16
2. L(PID)	20	4-17
2. N(PID)	20	4-17
3. L(PID)	20	4-18
3. N(PID)	20	4-18

Table 4.11: Figures corresponding to Experiments 4.3 with Lp=20 [s]

4.4.3 Observations

The results of the simulations are shown in Fig.4-13 to Fig.4-18. The simulations show the performance of the controller with nonlinear gain scheduling to be better with respect to setpoint tracking, than that of the controller without gain scheduling. The disturbance rejection is virtually the same for both of the controllers.

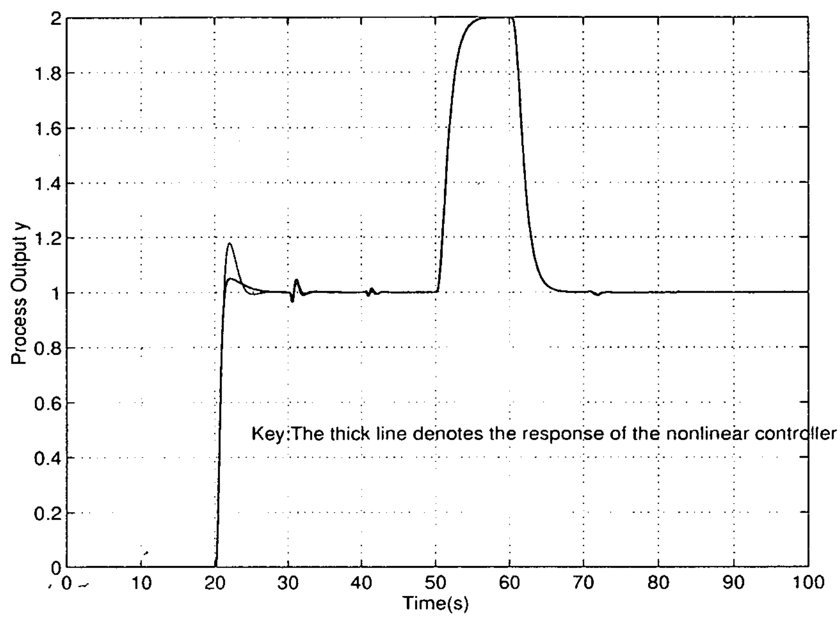

Figure 4-13: Setpoint tracking and disturbance rejection when $L_p = 10$ [s]. Smith predictor compensation and windup protection are present. Case 1.

4.5 Conclusions

The previous simulations indicate that a controller with nonlinear gain scheduling may be used to provide good control. In most instances, the gain scheduling controller yields a better control performance than the controller without gain scheduling that is properly tuned. This is something that should be expected. This is confirmed by comparing performance indices like overshoot and overall loop settling time. Overall, the controller

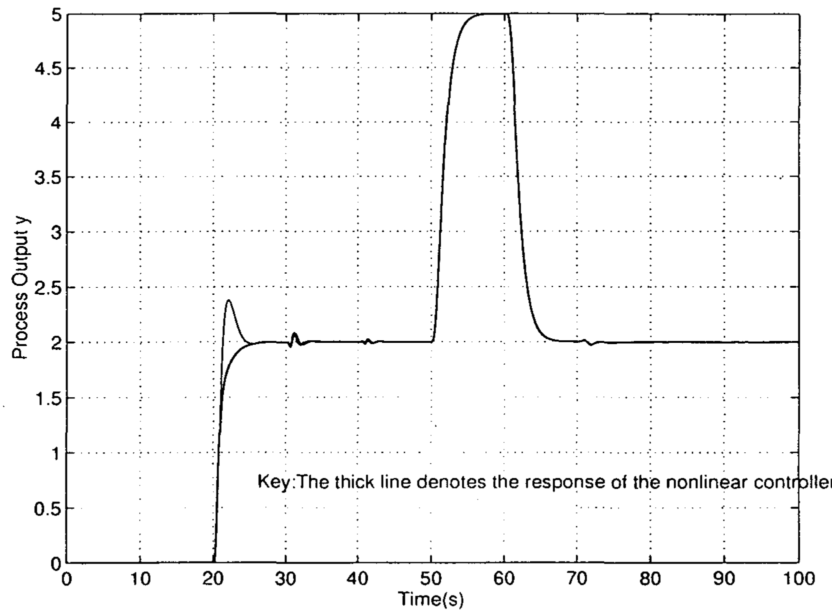

Figure 4-14: Setpoint tracking and disturbance rejection when $L_p = 10$ [s]. Smith predictor compensation and windup protection are present. (Case 2).

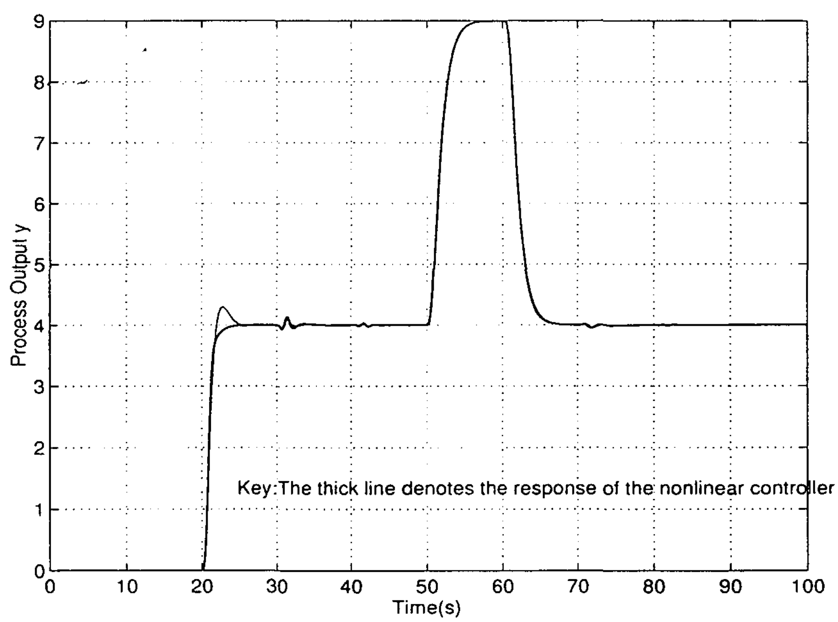

Figure 4-15: Setpoint tracking and disturbance rejection when $L_p = 10$ [s]. Smith predictor compensation and windup protection are present. (Case 3).

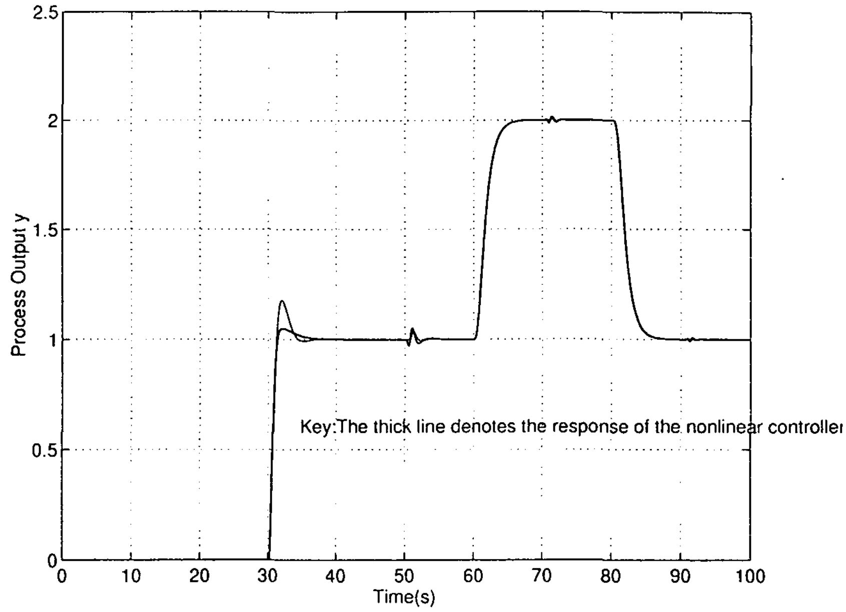

Figure 4-16: Setpoint tracking and disturbance rejection when $L_p = 20$ [s]. Smith predictor compensation and windup protection are present. (Case 1).

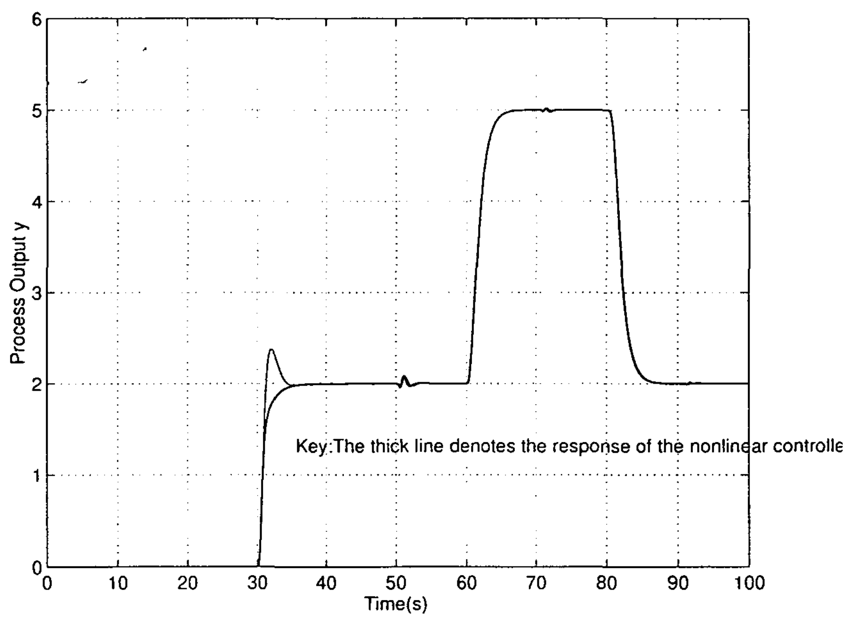

Figure 4-17: Setpoint tracking and disturbance rejection when $L_p = 20$ [s]. Smith predictor compensation and windup protection are present.(Case 2).

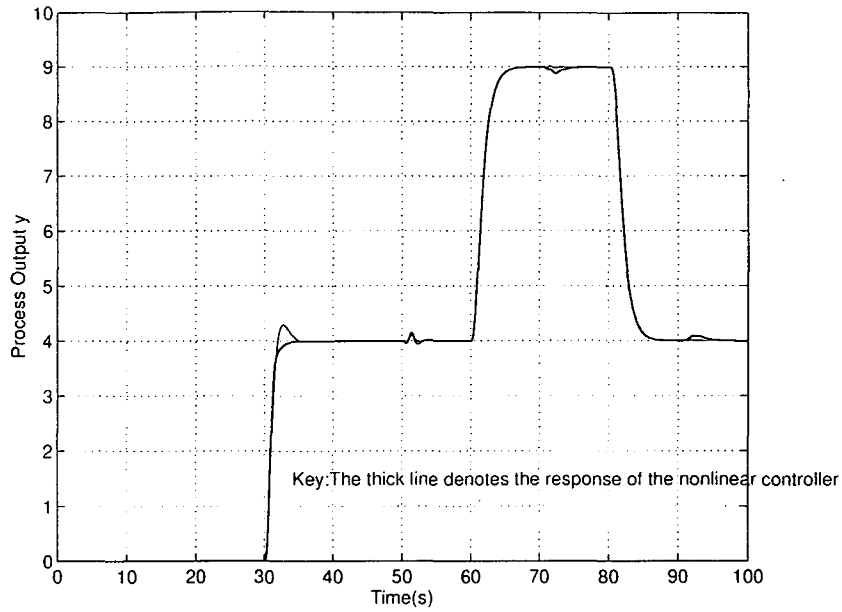

Figure 4-18: Setpoint tracking and disturbance rejection when $L_p = 20$ [s]. Smith predictor compensation and windup protection are present.(Case 3).

with gain scheduling performed better with respect to setpoint tracking keeping the same or better disturbance rejection performance.

'Hybrid' control schemes which combine the desirable characteristics of controllers with the fixed gains and with nonlinear gain scheduling can be used effectively for good control. Nonlinear gain scheduling may be used when a process is switched on for the first time, as well as during setpoint changes, for good setpoint tracking. Changing to fixed gain controller will retain the robust stability of the closed loop system, and still maintain the control function.

The main difficulty encountered with the nonlinear gain scheduling controller appears to be in its tuning. This problem is reduced by the tuning method proposed by Bajić (1994), but it is not as universal as what we would like. Proper tuning of the nonlinear controller is based on optimization methods and requires computer support for the cases not covered by the technique of Bajić (1994). To overcome the difficulties associated with the proper tuning and to improve the disturbance rejection properties of the PID controller, a variant of PID controller utilizing nonlinear proportional-derivative action is proposed by Bajić (1995), Bajić & Rybalov(1996) and Govender & Bajić(1996). This concept is extended in the next chapter.

Chapter 5

Nonlinear PD Modifier for Improved Setpoint Tracking and Disturbance Rejection

5.1 Introduction and General Backgound

Derivative (D) control is not too popular for the control of process loops in practice because it complicates the tuning process, a task which few can properly execute (Shinskey 1994). Popular perception is that D-action reduces the robustnes of a loop. Contrary to this, linear PID controllers, all tuned for minimum IAE and used in lag-dominant loops, exhibit far better disturbance rejection properties compared to a conventional PI controller (Shinskey 1994). One of the main reasons for D-action being omitted from a control loop stems from the fact that it is sensitive to noise components within the controlled variable. This shortcoming can be overcome (to a certain extent) by filtering the process output before introducing it into the derivative part of the controller.

Derivative action can be used to improve the stability of a control system. The problem with D-action is that it has the negative effect of slowing down a system's reaction speed to signal changes. This negative effect can be compensated for by increasing the proportional gain, but with certain constraints. There is therefore a need to improve the system's reaction speed without compromising its stability. This can be acheived, within limits, by utilising combinations of linear and nonlinear control which we shall designate as *'hybrid control'*. These hybrid controllers may be used for a high speed control as in certain applications where the speed of response may be critical (eg. servo control problems, missile control, etc.). The PID portion of this hybrid controller can be tuned in to a fast rise time and at setpoint the nonlinear component may be primed to activate and flatten the response (Shaw 1994).

5.2 The Hybrid Controller

The idea of a PID controller having a nonlinear proportional-derivative component constructed from the fuzzy-logic solution is borrowed from Bajić (1995) and Bajić & Rybalov (1996). Empirical evidence has shown that combinations of nonlinear and conventional control generally exhibit an improvement in system response (Zhao *et al.* 1993, Bajić & Rybalov 1996, Govender & Bajić 1996). The discussion hereafter will be restricted to instances within the control session when P-action and D-action have conflicting requirements with respect to setpoint tracking and disturbance rejection. The linear PID /L(PID)/ controller with a nonlinear modifier (NLM) may be a solution of this problem. This solution will be discussed in this chapter.

The step response shown in Fig.5-1 and the rule based logic are used to develop the algorithm of the nonlinear PD modifier. An analysis of this response is summarized in Table 5.1.

The following analysis is based on the step response shown in Fig.5-1:

Position A: $e > 0$, $De \leq 0$

For a positive large error, the derivative action should be as negative as possible in order to speed up the system response towards the setpoint.

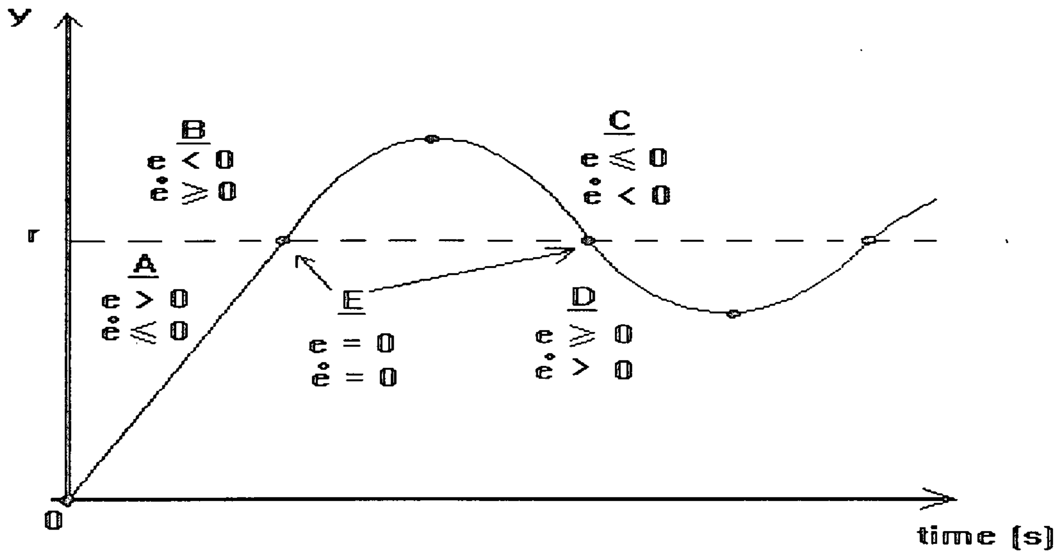

Figure 5-1: System response to a step input

Case	e	De
A	> 0	≤ 0
B	< 0	≥ 0
C	≤ 0	< 0
D	≥ 0	> 0
E	0	0

Table 5.1: Table of analysis for the step response

Position B: $e < 0$, $De \geq 0$

A negative error occurs for excursions of the output signal beyond the setpoint. The derivative action must be increased to damp the system response in order to reduce oscillations.

Position C: $e \leq 0$, $De < 0$

Derivative action must be as negative as possible in order to speed up the system response towards the desired output.

Position D: $e \geq 0$, $De > 0$

A positive error is present for excursions of the output signal below the setpoint. The effect of the increased derivative action is to damp system response and reduce the oscillations appearing in the process output.

Position E: $e = 0$, $De = 0$

Derivative action ceases when the output signal reaches setpoint.

The above-mentioned rules capture the fact that when the response is moving towards setpoint, De must decrease for the system's response speed to increase. Conversely, when the output moves away from the desired value, De must increase in order to damp the response and move it back to setpoint. Derivative action ceases when the system response is at setpoint. This is the logic used for the basic design of a nonlinear modifier (Bajić 1995) and later on is used for the construction of the fuzzy-logic modifier of the PID controller (Bajić & Rybalov 1996). Also, the non-fuzzy construction used for the control of systems with very long dead-time is given in Govender & Bajić (1996). In the discussions which follow in Chapter 6, we shall extend this solution to the case utilising an anti-windup mechanism and Smith predictor control.

5.2.1 Developement of the Nonlinear Modifier Algorithm

The theory promulgated for developement of the algorithm for the NLM is based on a connection with Lypanov's direct method with respect to the stability analysis of nonlinear systems having a very small dead time. The control algorithm proposed (Bajić

1995) is:

$$
\begin{aligned}
u_c &= K_p e + K_i \int_0^t e(s)ds - K_d(Dy - \alpha) \\
&= K_p e + K_i \int_0^t e(s)ds - K_d Dy + K_d \alpha \\
&= u_l + u_n
\end{aligned}
$$

where $K_p e + K_i \int_0^t e(s)ds - K_d D_y$ is the linear part of the controller denoted by u_l and $K_d\alpha$ is the nonlinear part denoted by u_n. The action of the simplified nonlinear modifier is governed by

$$
\begin{aligned}
u_n \quad &= \quad K_d\alpha \\
\theta_2 \quad &> \theta_1 \quad > 0 \\
g(e, De) \quad &= \quad |e| + 0.01\,|De|
\end{aligned}
$$

where

$$
\alpha = \begin{cases}
+\theta_1 g(e, De) \text{ for } (e > 0), (De \leq 0) \text{..........Rule A} \\
-\theta_1 g(e, De) \text{ for } (e < 0), (De \geq 0) \text{..........Rule B} \\
-\theta_2 g(e, De) \text{ for } (e \leq 0), (De < 0) \text{..........Rule C} \\
+\theta_2 g(e, De) \text{ for } (e \geq 0), (De > 0) \text{..........Rule D} \\
0 \text{ for } (e = 0), (De = 0) \text{.................. .. ...Rule E}
\end{cases}
$$

Rules A, B, C, D and E are based on the system response to a step input as shown in Fig.5-1. The multiplication factor 0.01 for De is used in this instance because it was found, through experimentation, to yield the most reasonable results. Following Bajić (1995), F_n is defined by

$$
\begin{aligned}
F_n(e, De)K_f &= K_d\alpha \text{ if } K_d.K_f \neq 0 \\
F_n &= 0 \text{ if } K_d.K_f = 0
\end{aligned}
$$

with

$$K_d = 0 \Leftrightarrow K_f = 0$$

so that the nonlinear part of the controller can be written as

$$u_n = K_d\alpha = F_n(e, De)K_f$$

The values of $\theta_1 = 0.3$ and $\theta_2 = 0.6$ have been obtained experimentally. They are used in this analysis and are not optimal in any specific sense. Gain K_f appears to be the only adjustable parameter of the nonlinear modifier. The complete control algorithm (u_c) for this hybrid controller can now be expressed in the following manner

$$u_c = K_pe + K_i \int_0^t e(s)ds + K_dDe + F_n(e, De)K_f$$

A block diagram implementing this control law is shown in Fig.6-1. We will extend this solution in Chapter 6 to include instances where windup and long dead-time within the process have a negative effect on the controllers performance.

5.3 Conclusion

The structure of a nonlinear modifier to improve the performance of the conventional PID controller has been presented. The proposed nonlinear modifier is connected in parallel with the linear PID controller. The control signal from the PID + NLM combination will enhance the performance of closed-loop processes. An original control scheme for processes having a saturation nonlinearity and long dead-times will be developed in Chapter 6. The proposed algorithm will be developed on the basis of the theory presented in Chapter 2 and in this chapter.

Chapter 6

Control of Processes Having Long Dead-time and Integrator Windup

6.1 Introduction

Following from the discussions in Chapter 2 and Chapter 3, we concluded that the presence of integral windup and dead-time within the process loop resulted in a deterioation in the performance of the loop. A detailed description and analysis of integral windup, and the various techniques used to reduce its negative effects, was given in Chapter 2. Chapter 3 introduced a controller scheme which consisted of a PID controller that has nonlinear gain scheduling for improved setpoint tracking, and an anti-windup mechanism together with Smith predictor control for the compensation of loops experiencing the negative effects of actuator saturation and long process dead-time, respectively. This chapter will discuss the control structures of the nonlinear controller on the basis of the solution proposed in Chapter 5. Combinations of the linear PID controller and the nonlinear modifier will be used to control:

- loops having no saturation nonlinearity or process dead-time;

- loops utilising an anti-windup scheme to reduce overshoots that may occur as a result of integral windup caused by the actuator saturating;
- loops with Smith predictor control to compensate for the negative effects of dead-time, and
- loops with antiwindup and Smith predictor control.

The emphasis shall be on the control of loops having long dead-time and significant actuator nonlinearity, and an *original control scheme* will be proposed to compensate for any negative effects which these nonlinearities may have on the behavior of the control loop. The solutions proposed in this chapter are illustrated in the next chapter via experiments to show the improved performance of the combination of the linear PID controller and the nonlinear modifier with antiwindup and Smith predictor control.

6.2 Control of Loops Having no Saturation Nonlinearity or Dead-Time

Chapter 5 discussed the developement of a nonlinear modifier to improve setpoint tracking and disturbance rejection. From Chapter 5 we deduced that the nonlinear modifier is designed so that, compared to the linear controller, it will increase the speed of the output y when y moves towards r, and strongly oppose any change in y when y moves away from r. This was the logic used in the design of the nonlinear modifier for systems experiencing no dead-time (Bajić 1995, Bajić & Rybalov 1996). The behaviour of the conventional linear PID controller is determined by

$$\begin{aligned} u_c(t) &= u_p(t) + u_i(t) + u_d(t) \\ &= K_p e(t) + K_i \int_0^t e(s)d(s) + K_d De(t) \end{aligned} \tag{6.1}$$

while the linear controller combined with the nonliner modifier is governed by

$$\begin{aligned} u_c(t) &= u_p(t) + u_i(t) + u_d(t) + u_n \\ &= K_p e(t) + K_i \int_0^t e(s)d(s) + K_d De(t) + F_n(e, De)K_f \end{aligned} \tag{6.2}$$

where $F_n(e, De)$ is determined in Chapter 5. The controller output signal u_c enters the process. Output signal y from the process is fed back to the input to form the error signal e. Fig.6-1 shows how this controller is implemented.

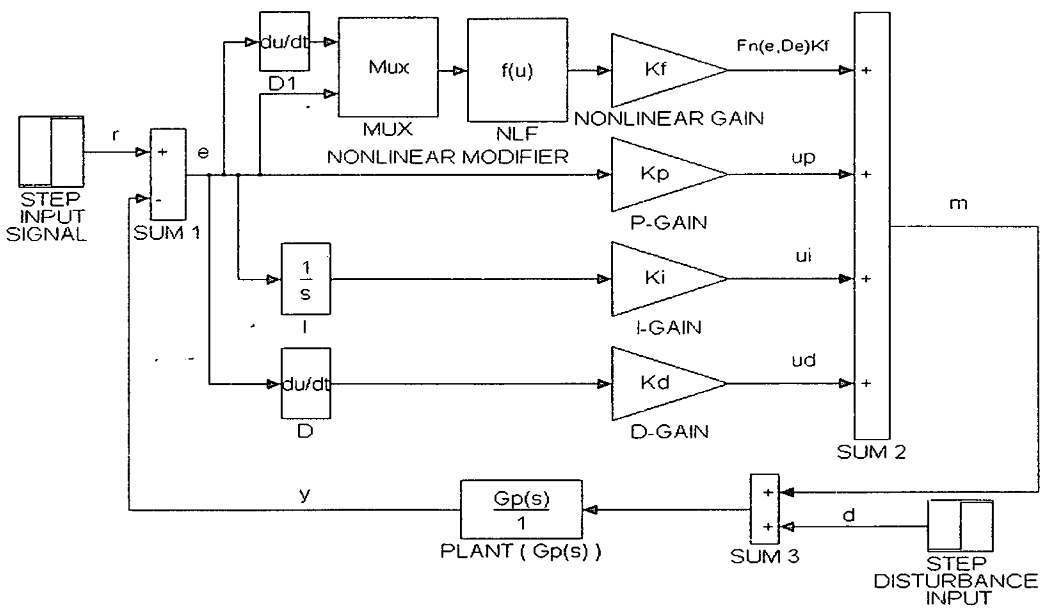

Figure 6-1: PID controller and nonlinear modifier for the control of loops having no saturation nonlinearity or dead-time

6.3 Control of Loops Experiencing Integral Windup

The saturation of an actuator leads to the integral controller calling for an action which the actuator cannot produce. This is known as integral or reset windup. Integral windup is especially troublesome during the initial startup of a plant, and it could lead to severe overshoots. Windup, and various techniques to reduce its negative effects by using a linear controller with an antiwindup scheme, was discussed in Chapter 2. The control law governing the behaviour of the linear PID controller with one possible antiwindup scheme is

$$\begin{aligned} u_c(t) &= u_l - u_w \\ &= K_p e(t) + K_i \int_0^t e(s)d(s) + K_d De(t) - K_w \int_0^t e'(s)d(s) \end{aligned} \tag{6.3}$$

where $e' = u_c(t) - u(t)$ is the windup protection feedback error and K_w is the gain of the antiwindup scheme. The action of the PID controller with antiwindup and a nonlinear modifier is governed by

$$\begin{aligned} u_c(t) &= u_l - u_w + u_n \\ &= K_p e(t) + K_i \int_0^t e(s)d(s) + K_d De(t) - K_w \int_0^t e'(s)d(s) + F_n(e, De)K_f \end{aligned} \tag{6.4}$$

The actuator output u is fed to the process. The process output y is fed back to the input to form the error signal e. This control scheme is implemented in Fig. 6-2.

6.4 Control of Loops Having Long Dead-time

The nonlinear nature of the proposed nonlinear modifier makes it very sensitive to dead-time. Any dead-time present in the control loop can result in control system instability. As mentioned in Chapter 3, one popular method to overcome the negative effects of dead-time utilises Smith predictor control (Smith 1958) to compensate for the process dead-time. The process models used in the Smith predictor compensator are approximated

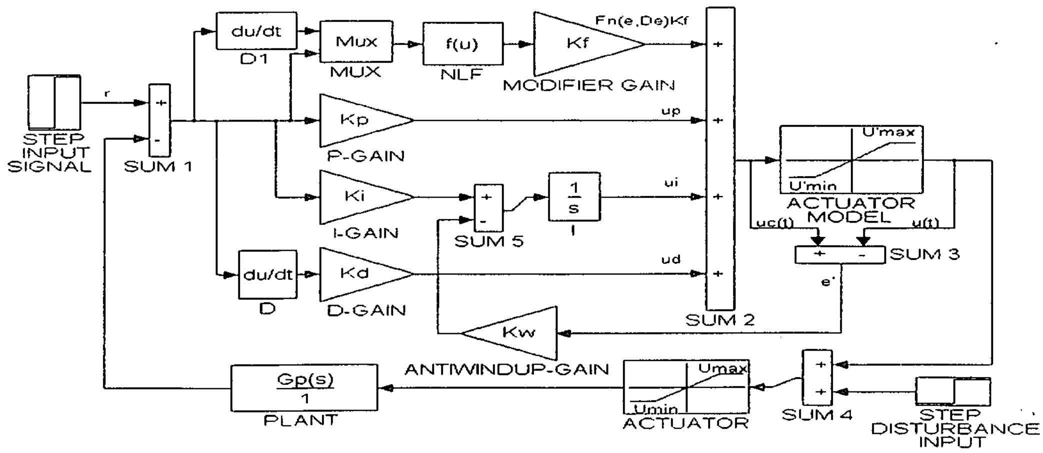

Figure 6-2: PID controller with nonlinear modifier, and antiwindup for controlling loops having a saturation nonlinearity

in order to reduce the sensitivity of the nonlinear modifier to dead-time. The structure of the Smith predictor is such that the signal fed back into the loop is responsible for generating an output that excludes any negative effects of dead-time. Let the process model be $G_p(s)\exp(-L_p s)$, where G_p is the rational part of the process transfer function and L_p is the process dead-time. Then the model chosen for good cancellation of the actual process output will be $G_{m1}(s)\exp(-L_{m1}\mathrm{s})$ and can be of relatively high order. The additional loop contains a lower order model, namely $G_{m2}(s)\exp(-L_{m2}s)$, for the proper control of the process. The combined PID controller and nonlinear modifier, with Smith predictor control, is governed by (6.2) and the following operator equations:

$$e = r - (y - y_{m2} + y_{m1}) \tag{6.5}$$

$$y_{m1} = [G_{m1}(s)\exp(-L_{m1}s)]\, m \tag{6.6}$$

$$y_{m2} = [G_{m2}(s) \exp(-L_{m2}s)] \, (m + d) \tag{6.7}$$

where m is the control signal and d denotes any disturbance that the loop may experience. The control signal is fed into the Smith predictor compensator and the process. The feedback signal is $y - y_{m2} + y_{m1}$. This signal represents the output of the process and Smith predictor scheme, excluding the negative effects of dead-time. The structure of this controller using a combination of (6.2), (6.5), (6.6) and (6.7) is shown in Fig.6-3.

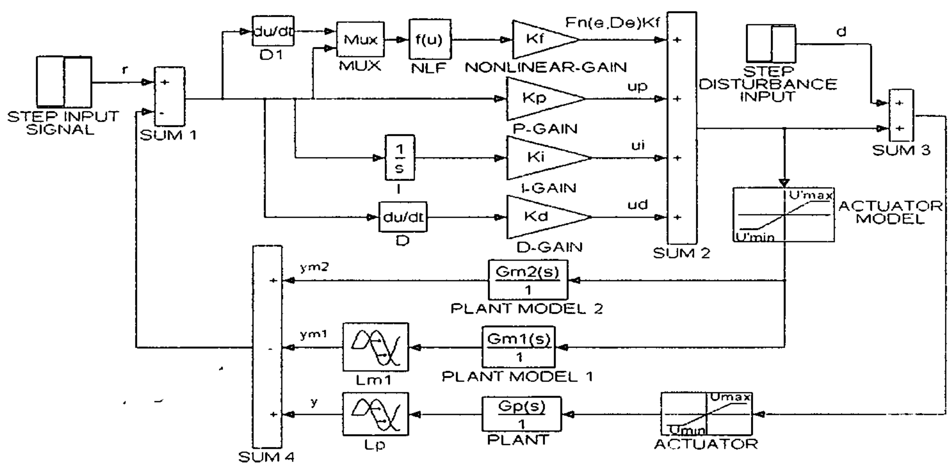

Figure 6-3: Linear PID controller with Nonlinear PD Modifier and Smith predictor compensator

6.5 Control of Loops Experiencing Integral Windup and Dead-time

Control loops having nonlinearities like actuator saturation, as well as those having processes with long dead-time, may experience instability as a result of the combined negative effects of these factors. The objective of this report is to develop a controller to that will improve the performance of such loops. A combination of the linear PID controller connected in parallel with the NLM, together with an antiwindup scheme and Smith predictor control, succeeds in improving considerably the control performance of systems experiencing the saturation nonlinearity and long process dead-time. The control algorithm governing the behavior of this controller with antiwindup and dead-time compensation is

$$\begin{aligned} u_c(t) &= K_p e(t) + K_i \int_0^t e(s)d(s) + K_d De(t) \\ &\quad - K_w \int_0^t e'(s)d(s) + F_n(e, De)K_f \end{aligned} \tag{6.8}$$

where the error e is a function of the Smith predictor compensator feedback scheme as determined by operator equations (6.5), (6.6) and (6.7). The controller output signal m goes to the Smith predictor and actuator, which feeds the process. The windup protection scheme generates an error signal e' when $(u_c(t) - u(t)) > 0$. This signal is used to reset the integral controller when the actuator saturates. The structure of this controller is shown in Fig.6-4.

6.6 Conclusion

An original control scheme to improve control of processes experiencing the negative effects of integrator windup and long dead-times has been proposed. The PID controller + NLM, with antiwindup control and Smith predictor compensation will positively enhance the control performance of closed-loop processes experiencing any negative effects of these factors. The next chapter will test the control performance of the proposed scheme, in loops having a saturation nonlinearity and long dead-times.

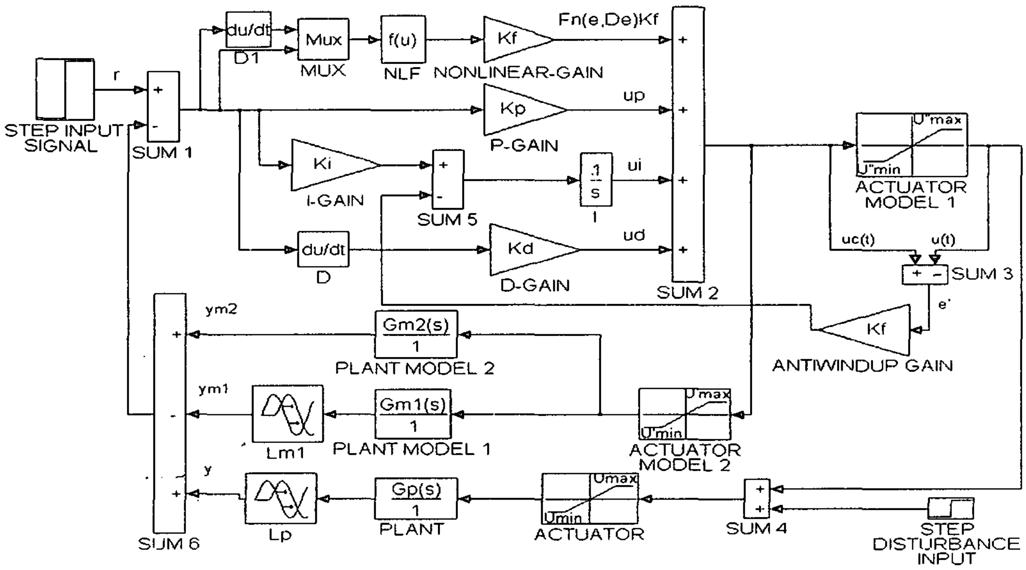

Figure 6-4: Linear PID controller with antiwindup, Smith predictor compensator and Nonlinear Modifier

Chapter 7

Experiments Using the Nonlinear Modifier

7.1 Preliminaries

The following experiments will illustrate the control performance of the controller schemes mentioned in the previous discussions. A SISO negative unity feedback system, with the process and controller connected in series in the feedforward path, will be considered in the subsequent experiments. The reference input signal is $r(t) = M_r h(t-10)$, where h is a unit step function. The disturbance input signal $d(t) = M_d h(t-50)$ is of the step type and acts as an additive to the controller output $u(t)$, so that the process input is governed by $u_{proc}(t) = u(t) + d(t)$. All subsequent experiments utilise a conventional linear three term PID controller governed by (6.1). A comparison shall be made between different combinations of the linear PID controller, and the linear PID controller + NLM governed by (6.2). The comparison will be made with respect to setpoint tracking and disturbance rejection, for different step inputs and step disturbances. This is done for an optimally tuned controller as well as for an improperly tuned one.The linear controller is optimally tuned for disturbance rejection and the criterion selected is IAE. The gains of the NLM and the antiwindup scheme (K_f and K_w respectively) are adjusted till a good response is obtained. Two different processes models, the first $G_{p1}(s)$ having a fourth order transfer function

$$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$$

and the second $G_{p2}(s)$ having a sixth order transfer function

$$G_{p2}(s) = \frac{64}{[(s+1)(s+2)(s+4)]^2}$$

will be considered in the subsequent experiments. These two processes will each be controlled (as illustrated in the following experiments) by combinations of:

Experiment 7.1: The linear PID controller, and the linear controller + nonlinear modifier (NLM) for loops having no dead-time or saturation.

Experiment 7.2: The linear controller + antiwindup, and the linear controller + NLM + antiwindup for loops containing saturation nonlinearity.

Experiment 7.3: The linear controller with Smith predictor control, and the linear controller with Smith predictor control + NLM, for loops having long dead-time.

Experiment 7.4: The linear controller with Smith predictor control + antiwindup, and the linear controller +Smith predictor control + antiwindup + NLM, for loops having long dead-time and saturation nonlinearity.

$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
1. linear controller	1	1	1	1	0	0
1. linear controller + NLM	1	1	1	1	0	0.65
2. linear controller	2	3	1	1	0	0
2. linear controller + NLM	2	3	1	1	0	0.65
3. linear controller	4	5	1	1	0	0
3. linear controller + NLM	4	5	1	1	0	0.65

Table 7.1: Optimal tuning parameters for the fourth order system

7.2 Experiment 7.1: Loops Having no Dead-time or Saturation Nonlinearity.

7.2.1 Purpose

The purpose of this experiment is to compare the performance of the linear PID controller to that of the linear PID controller having a nonlinear modifier. This comparison will be made for loops having no saturation nonlinearity or dead-time.

7.2.2 Method

The behaviour of $G_{p1}(s)$ and $G_{p2}(s)$ will be tested under the conditions described previously in Section 6.2. The linear PID controller is optimally tuned for disturbance rejection. Gain K_f of the nonlinear modifier is adjusted to improve setpoint tracking and disturbance rejection. The control scheme for this experiment is shown in Fig.6-1. The optimized values for the linear PID controller governed by (6.1) and the PID controller + nonlinear modifier governed by (6.2) are shown in Table 7.1 and Table 7.3. Table 7.2 and Table 7.4 list a set of possible parameters for an improperly tuned PID controller.

$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
4. linear PID controller	1	1	3.042	1.49	0	0
4. linear PID controller + NLM	1	1	3.042	1.49	0	0.65

Table 7.2: Improper tuning parameters for the fourth order system)

$G_{p2}(s) = \frac{64}{[(s+1)(s+2)(s+4)]^2}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
5 linear PID controller	1	1	1.75	1	1.15	0
5. linear PID controller + NLM	1	1	1.75	1	1.15	3
6. linear PID controller	2	3	1.75	1	1.15	0
6. linear PID controller + NLM	2	3	1.75	1	1.15	3
7. linear PID controller	4	5	1.75	1	1.15	0
7. linear PID controller + NLM1	4	5	1.75	1	1.15	3

Table 7.3: Optimal tuning parameters for the sixth order system

$G_{p2}(s) = \frac{64}{[(s+1)(s+2)(s+4)]^2}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
8. linear PID controller	1	1	1	1	1	0
8. linear PID controller + NLM	1	1	1	1	1	3

Table 7.4: Improper tuning parameters for the sixth order system

Case No.	**Figure Ref. No.**
1	7.1 & 7.2
2	7.2
3	7.2
4	7.3
5	7.4 & 7.5
6	7.5
7	7.5
8	7.6

Table 7.5: Table of references for Experiment 7.1

7.2.3 Observations

The results of the simulation for the linear PID controller and the linear PID controller with NLM are shown in Fig. 7-1 and Fig. 7-2 for the fourth order process model, and in Fig. 7-4 and Fig. 7-5 for the sixth order process model. Fig. 7-3 and Fig. 7-6 show the responses when the controller is not optimally tuned. From these results we observe that the quality of control with respect to setpoint tracking and disturbance rejection, for different inputs and disturbances, is retained by the combined linear controller and nonlinear modifier.

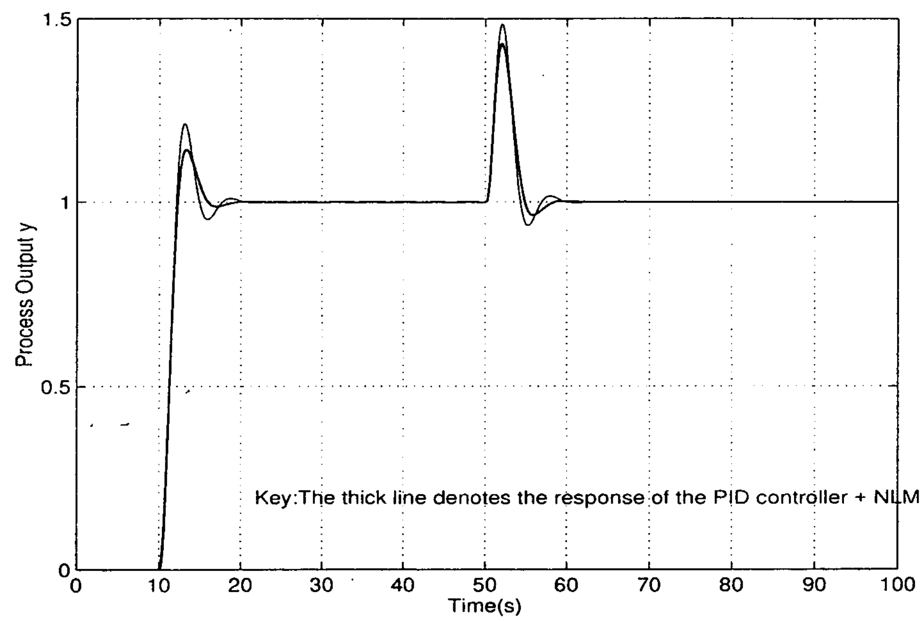

Figure 7-1: Step response for the fourth order system when the controller is optimally tuned

7.2.4 Conclusions

Combinations of the linear PID controller and the nonlinear modifier work well in optimising the performance of the loop, even in instances when the PID controller is improperly tuned.

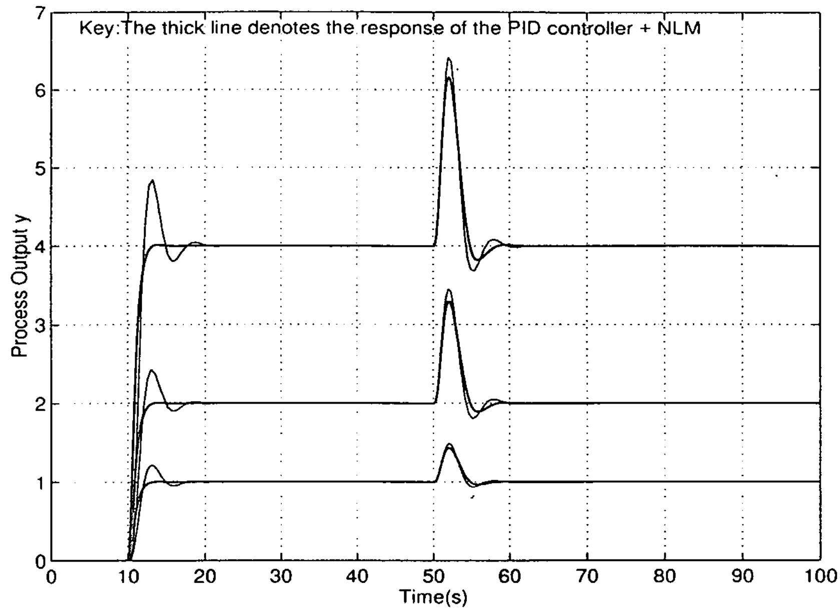

Figure 7-2: Combined responses of the fourth order system when the controller is optimally tuned

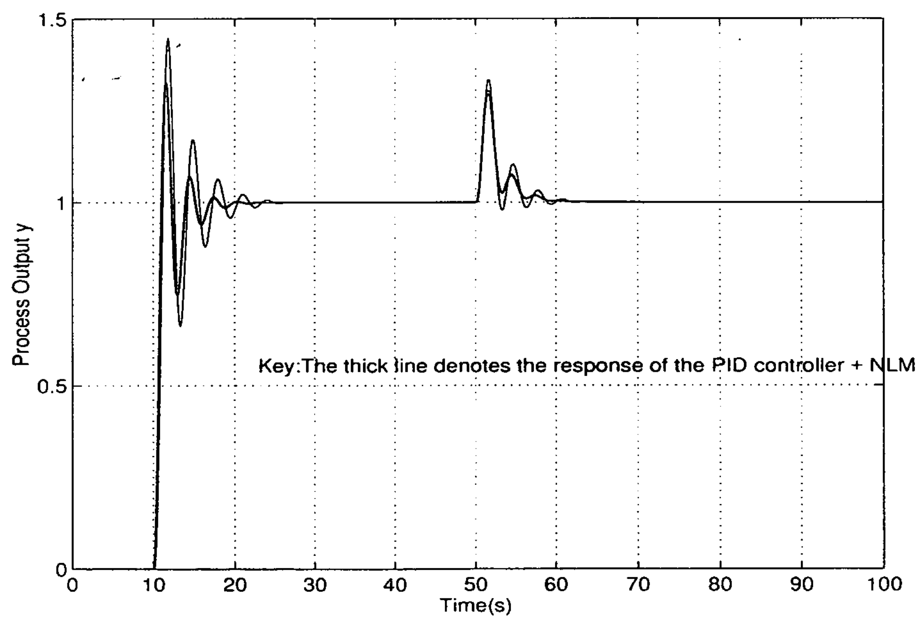

Figure 7-3: Step response of the fourth order system when the controller is not properly tuned

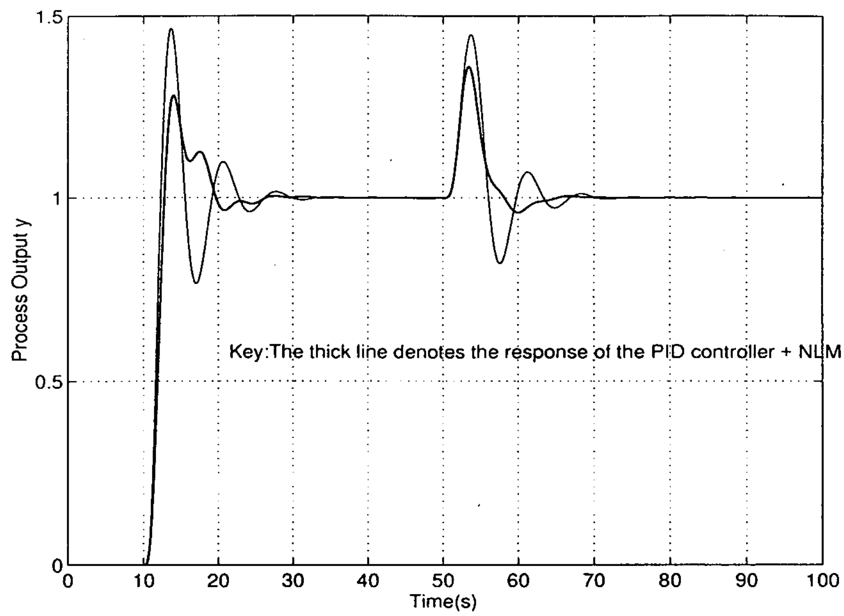

Figure 7-4: Step response for the sixth order system when the controller is tuned optimally

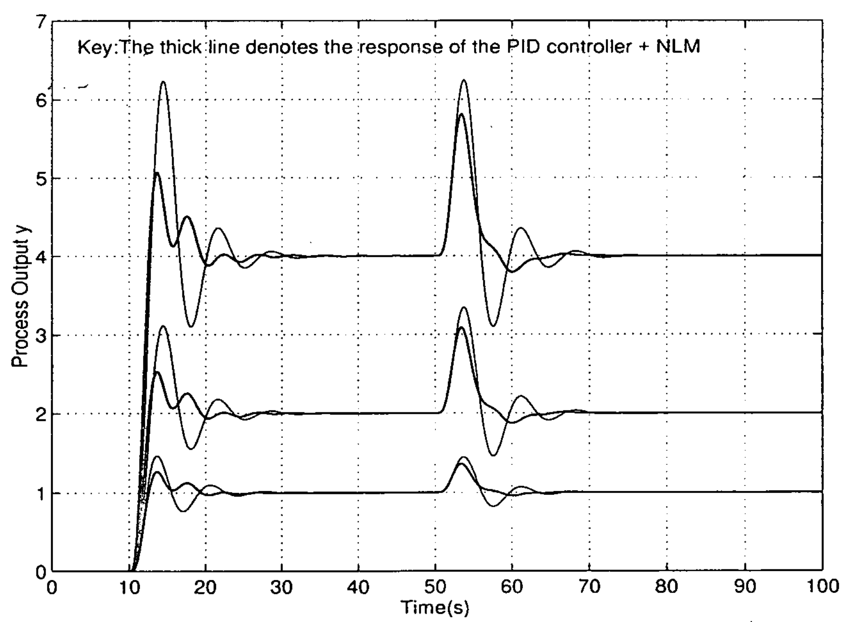

Figure 7-5: Combined responses for the sixth order system when the controller is optimally tuned

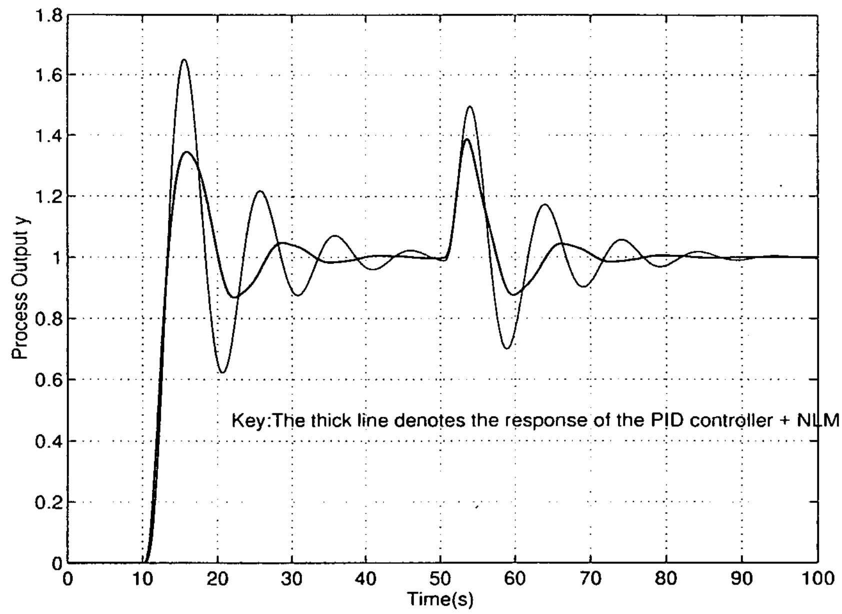

Figure 7-6: Step response of the sixth order system when the controller is not tuned properly

7.3 Experiment 7.2: Loops Having Saturation Non-linearity but without Dead-time.

7.3.1 Purpose

The purpose of this experiment is to compare the control performance of a linear conventional PID controller with anti-windup to that of the linear PID controller with anti-windup and the nonlinear modifier. This comparison will be made for loops that contain a saturation nonlinearity.

7.3.2 Method

The responses of process models $G_{p1}(s)$ and $G_{p2}(s)$ are tested under the conditions described previously in Section 6.3. The behaviour of the two controllers used in the experiment is governed by (6.3) and (6.4). Both these controllers are optimally tuned

$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$							
Parameters	M_r	M_d	K_p	K_i	K_d	K_f	K_w
1. linear PID + antiwindup	1	1	1	1.49	0.25	0	2
1. linear PID + NLM	1	1	1	1.49	0.25	3.5	0
1. linear PID + antiwindup + NLM	1	1	1	1.49	0.25	3.5	2
2. linear PID + antiwindup	2	3	1	1.49	0.25	0	2
2. linear PID + NLM	2	3	1	1.49	0.25	3.5	0
2. linear PID + antiwindup + NLM	2	3	1	1.49	0.25	3.5	2
3. linear PID + antiwindup	4	5	1	1.49	0.25	0	2
3. linear PID + NLM	4	5	1	1.49	0.25	3.5	0
3. linear PID + antiwindup + NLM	4	5	1	1.49	0.25	3.5	2

Table 7.6: Optimal tuning parameters for the fourth order system

$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$							
Parameters	M_r	M_d	K_p	K_i	K_d	K_f	K_w
4. linear PID + antiwindup	1	1	0	1	0	0	2
4. linear PID + NLM	1	1	0	1	0	3.5	0
4. linear PID + antiwindup + NLM	1	1	0	1	0	3.5	2

Table 7.7: Improper tuning parameters for the fourth order system

for disturbance rejection. Gain K_f of the nonlinear modifier is adjusted to improve the performance of the controller. K_w controls the feedback of the anti-windup scheme. K_w is adjusted in order to reduce overshoot, as a result of integrator windup, upon initial startup of the plant. The controller output u is fed into the input of the actuator whose saturation limits are set at -10 and $+10$. Fig.6-2 shows the structure of the controller used in this experiment. The optimized values for the linear PID controller + anti-windup, the linear PID controller + NLM and the linear PID controller + antiwindup + NLM are shown in Table 7.6 and Table 7.8. Table 7.7 and Table 7.9 list a possible set of tuning parameters when the linear controller is improperly tuned.

$G_{p2}(s) = \frac{64}{[(s+1)(s+2)(s+4)]^2}$							
Parameters	M_r	M_d	K_p	K_i	K_d	K_f	K_w
5. linear PID + antiwindup	1	1	1.75	1	1.25	0	1
5. linear PID + NLM	1	1	1.75	1	1.25	0.7	0
5. linear PID +antiwindup+NLM	1	1	1.75	1	1.25	0.7	1
6. linear PID + antiwindup	2	3	1.75	1	1.25	0	1
6. linear PID + NLM	2	3	1.75	1	1.25	0.7	0
6. linear PID + antiwindup + NLM	2	3	1.75	1	1.25	0.7	1
7. linear PID + antiwindup	4	5	1.75	1	1.25	0	1
7. linear PID + NLM	4	5	1.25	1.25	1.25	0.7	0
7. linear PID + antiwindup +NLM	4	5	1.75	1	1.25	0.7	1

Table 7.8: Optimal tuning parameters for the sixth order system

$G_{p2}(s) = \frac{64}{[(s+1)(s+2)(s+4)]^2}$							
Parameters	M_r	M_d	K_p	K_i	K_d	K_f	K_w
8. linear PID + antiwindup	1	1	1	1	1	0	1
8.linear PID + NLM	1	1	1	1	1	0.7	0
8.linear PID + antiwindup + NLM	1	1	1	1	1	0.7	1

Table 7.9: Tuning parameters when the sixth order system is improperly tuned

Case No.	**Figure Ref. No.**
1	7.7 & 7.8
2	7.8
3	7.8
4	7.9
5	7.10 & 7.11
6	7.10
7	7.10
8	7.12

Table 7.10: Table of references for the simulations

7.3.3 Observation

The results of the simulations are presented for comparison in Fig. 7-7 to Fig. 7-9 for the fourth order plant model, and Fig. 7-10 to Fig. 7-12 for the sixth order plant model. The results of the simulations clearly shows that overshoot as a result of actuator saturation is reduced with the inclusion of the anti-windup scheme. The nonlinear modifier contributes positively towards optimising setpoint tracking and disturbance rejection. Even in instances when the PID controller is improperly tuned, the nonlinear modifier stabilises the loop and improves setpoint tracking and disturbance rejection.

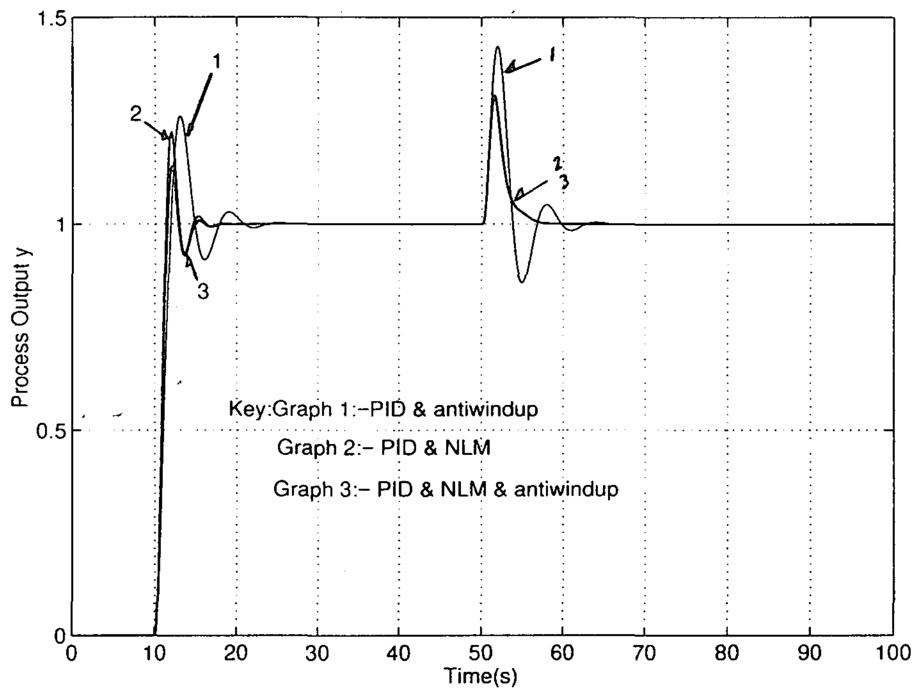

Figure 7-7: Step response for the fourth order system when the controller is optimally tuned

7.3.4 Conclusion

Combining the anti-windup scheme and the nonlinear modifier within the general structure of the conventional PID controller contributes towards the stability of the loop

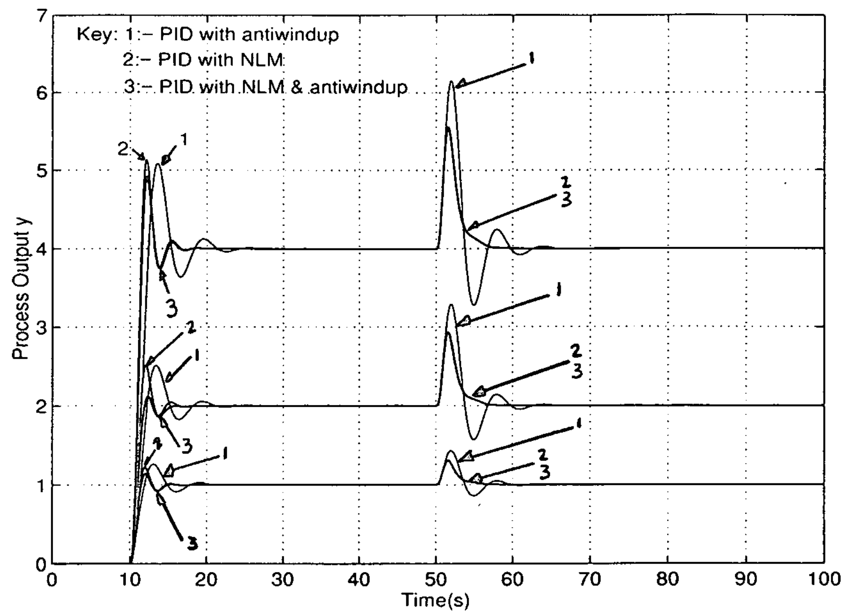

Figure 7-8: Combined responses for the fourth order system when the controller is optimally tuned

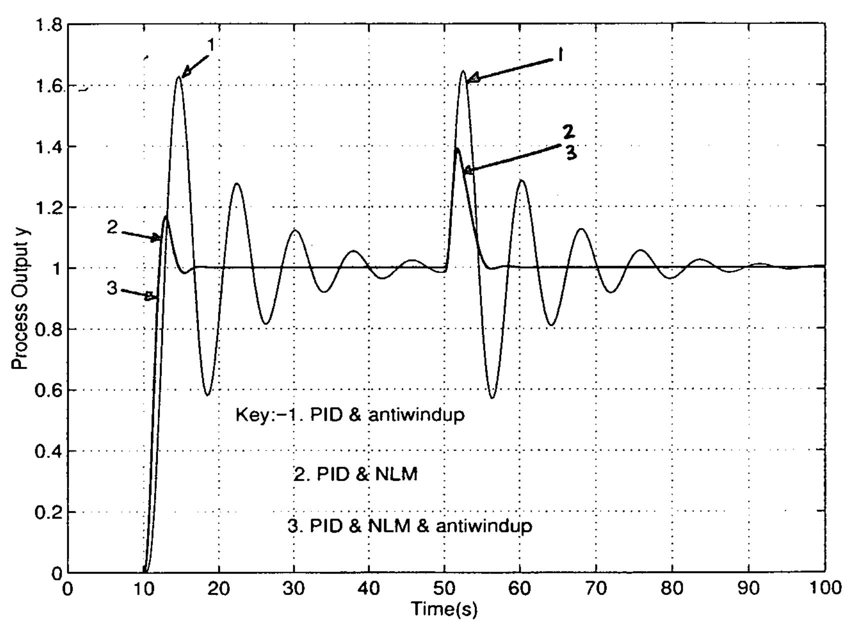

Figure 7-9: Step response for the fourth order system when the controller is not tuned properly

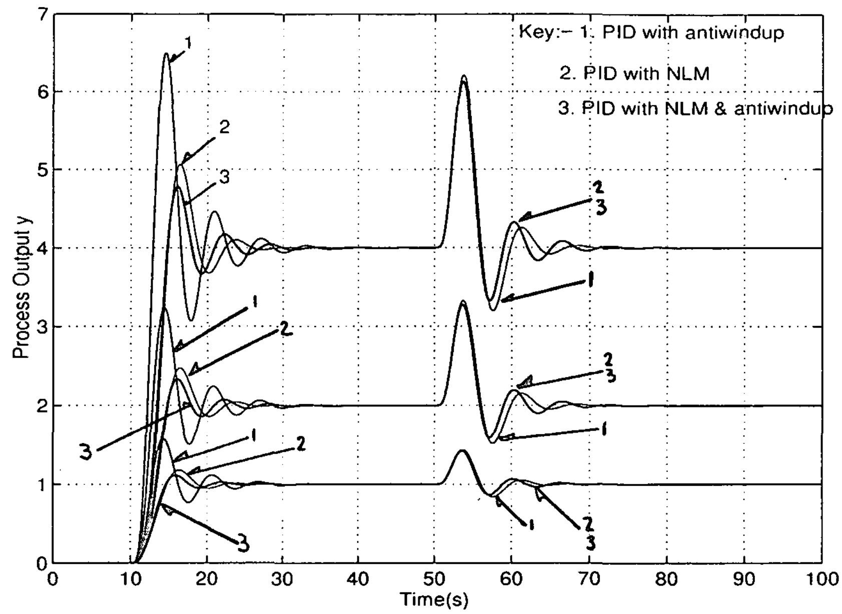

Figure 7-10: Combined responses for the sixth order system when the controller is tuned optimally

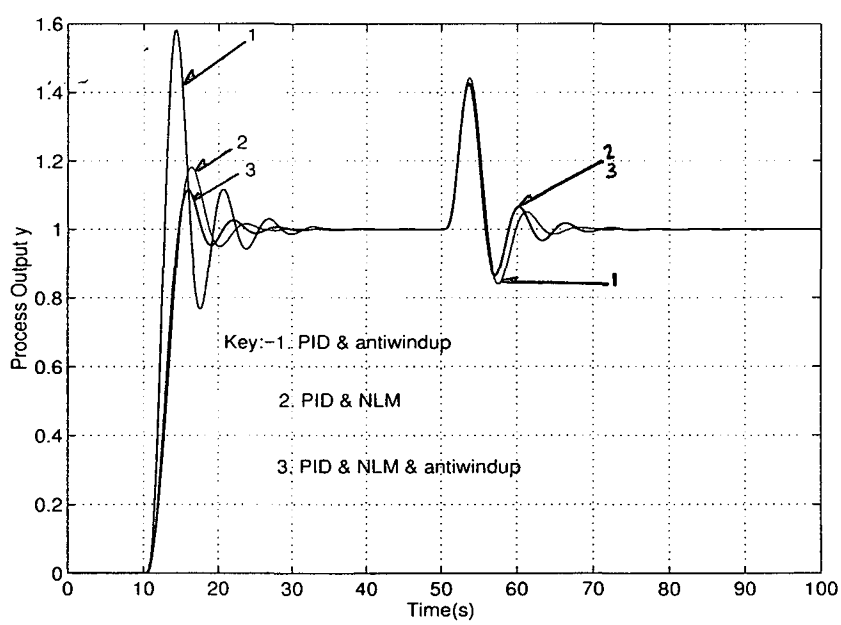

Figure 7-11: Step response for the sixth order system when the controler is optimally tuned

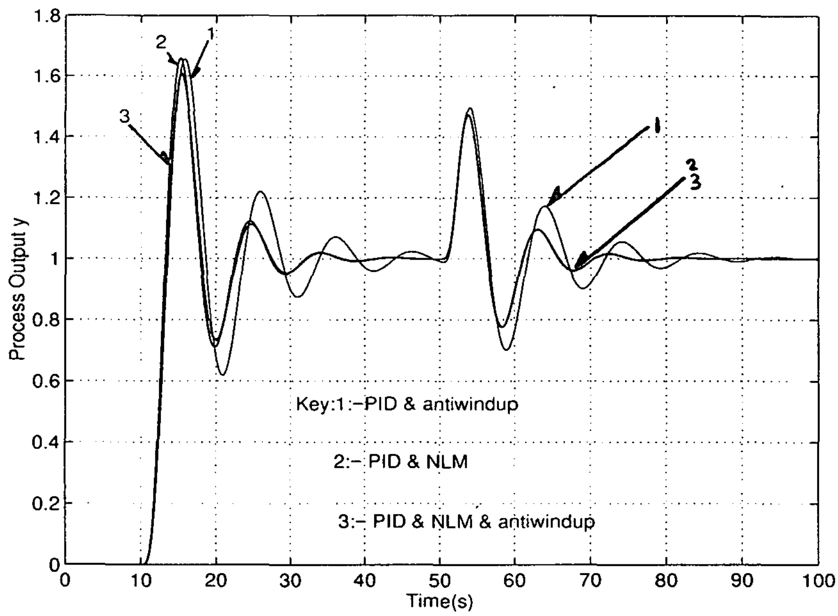

Figure 7-12: Step response for the sixth order system when the controller is not tuned properly

and enhances its control performance with respect to setpoint tracking and disturbance rejection.

7.4 Experiment 7.3: Loops Having Only Dead-time

7.4.1 Purpose

This experiment will compare the control performance of the conventional PID controller to that of a PID controller combined with the nonlinear modifier. The comparison is made for loops experiencing instability as a result of dead-time.

7.4.2 Method

Consider the process $G_p(s)\exp(-L_p s)$, where G_p represents the rational part of the transfer function of the process and L_p is the process dead-time:

For the first process model ($G_{p1}(s)$) of the fourth order given by

$$\frac{64}{(s+8)(s+4)(s+2)(s+1)}\exp(-L_{p1}s)$$

where $L_{p1} = 10$[s], the model chosen for good cancellation of the actual process output y in the first parallel feedback branch of the Smith predictor scheme is

$$\begin{aligned} G_{p1}^{m1}(s) &= \frac{0.0365s^2 + 1.3087s + 2.8008}{s^3 + 3.972s^2 + 5.8549s + 2.8007} \\ L_{p1}^{m1} &= 0.2575. \end{aligned}$$

The second order model in the second parallel feedback branch is

$$\begin{aligned} G_{p1}^{m2}(s) &= \frac{1.5523}{s^2 + 2.4884s + 1.5524} \\ L_{p1}^{m2} &= 0.3381. \end{aligned}$$

For a second process model of the sixth order

$$\frac{64}{[(s+1)(s+2)(s+4)]^2}\exp(-L_{p2}s)$$

where $L_{p2} = 10$[s], the third order model chosen for cancelling the process output y is

$$G_{p2}^{m1}(s) = \frac{0.0639s^3 + 0.0348s^2 + 1.4656s + 0.7597}{s^4 + 4.6507s^3 + 5.8618s^2 + 3.6052s + 0.7597},$$

$$L_{p2}^{m1} = 10.7175$$

In the second parallel feedback loop of the Smith predictor scheme, the second model used to identify the process is

$G_{p1}(s) = \dfrac{64}{(s+8)(s+4)(s+2)(s+1)}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
1. PID controller+Smith predictor	1	1	20	20	10	0
1. PID controller + Smith predictor + NLM	1	1	20	20	10	13
2. PID controller+Smith predictor	2	3	20	20	10	0
2. PID controller + Smith predictor + NLM	2	3	20	20	10	13
3. PID controller+Smith predictor	4	5	20	20	10	0
3. PID controller + Smith predictor + NLM	4	5	20	20	10	13

Table 7.11: Optimal tuning parameters for the fourth order system

$$\begin{aligned} G_{p2}^{m2}(s) &= \frac{0.5135}{s^2 + 1.2608s + 0.5135}, \\ L_{p2}^{m2} &= 11.0217. \end{aligned}$$

The complete control system used in this experiment is as described previously in Section 6.4. The response of the fourth and sixth order systems is tested in control loops having a dead-time of 10[s] and 20[s] respectively, using the controller given in Fig.6-3. The PID controller will first be optimally tuned, and then detuned to test the effectiveness of the nonlinear modifier at enhancing the perfromance of the control loop. The gain of the nonlinear modifier and the anti-windup scheme (K_f and K_w respectively) is adjusted till a good response is obtained with respect to disurbance rejection. The optimal tuning parameters for the linear PID controller + Smith predictor, and the linear PID controller + Smith predictor + nonlinear modifier are shown in Table 7.11 and Table 7.13. Table 7.12 and Table 7.14 list a possible set of tuning parameters when the controller is poorly tuned.

$G_{p1}(s) = \dfrac{64}{(s+8)(s+4)(s+2)(s+1)}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
4.PID controller+Smith predictor	1	1	0	2	1	0
4. PID controller + Smith predictor + NLM	1	1	0	2	1	13

Table 7.12: Improper tuning parameters for the fourth order system

$G_{p2}(s) = \dfrac{64}{[(s+1)(s+2)(s+4)]^2}$				
Parameters	M_r	M_d	K_p	K_i
5. PID controller+Smith predictor	1	1	6.8411	10
5. PID controller + Smith predictor + NLM	1	1	6.8411	10
6. PID controller+Smith predictor	2	3	6.8411	10
6. PID controller + Smith predictor + NLM	2	3	6.8411	10
7. PID controller+Smith predictor	4	5	6.8411	10
7. PID controller + Smith predictor + NLM	4	5	6.8411	10

Table 7.13: Optimal tuning parameters for the sixth order process

$G_{p2}(s) = \dfrac{64}{[(s+1)(s+2)(s+4)]^2}$						
Parameters	M_r	M_d	K_p	K_i	K_d	K_f
8. PID controller + Smith predictor	1	1	2	3	6.87	0
8. PID controller + Smith predictor + NLM	1	1	2	3	6.87	15

Table 7.14: Improper tuning parameters for the sixth order system

Case No.	**Figure Ref. No.**
1	7.13 & 7.17
2	7.14 & 7.18
3	7.15 & 7.19
4	7.16 & 7.20
5	7.21 & 7.25
6	7.22 & 7.26
7	7.23 & 7.27
8	7.24 & 7.28

Table 7.15: Table of references for the simulations in Experiment 7.3

7.4.3 Observations

The responses of the simulations are given in Fig. 7-13 to Fig. 7-28. We can observe that the tracking properties for an optimally tuned PID + NLM are better than that of the conventional linear PID controller. If the linear controller is not tuned properly, the nonlinear modifier stabilises the system response and contributes positively towards the tracking and rejection properties of the controller.

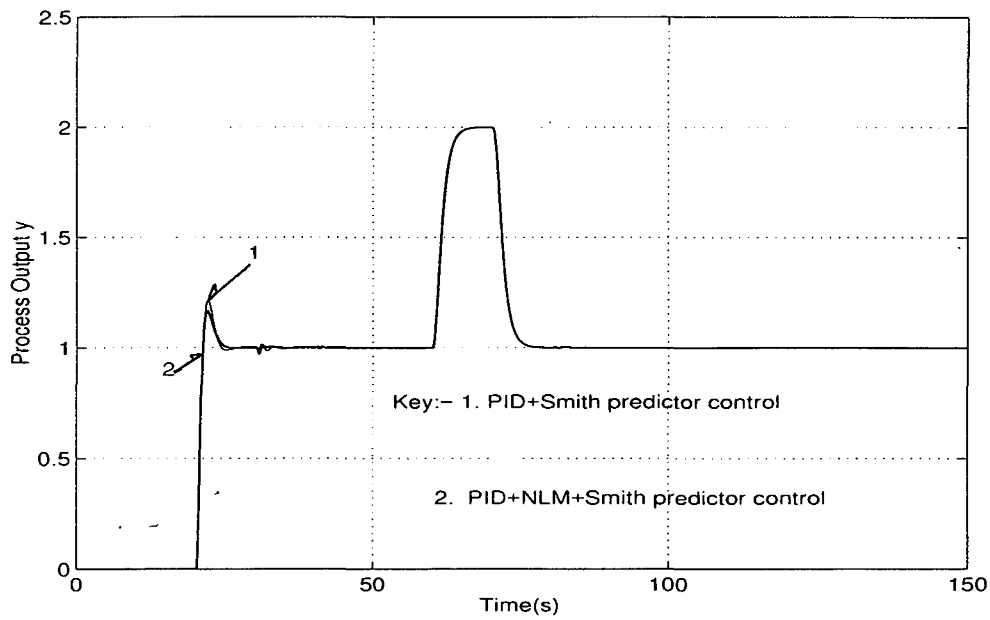

Figure 7-13: Optimal step response of the fourth order model when L_{proc} is 10[s]. (Case 1)

7.4.4 Conclusions

Combining the Smith predictor and nonlinear modifier within the structure of a conventional PID controller significantly improves the control of processes having very long dead-time. The Smith predictor compensates for dead-time while the nonlinear modifier positively enhances the performance of the loop with respect to setpoint tracking, keeping disturbance rejection good.

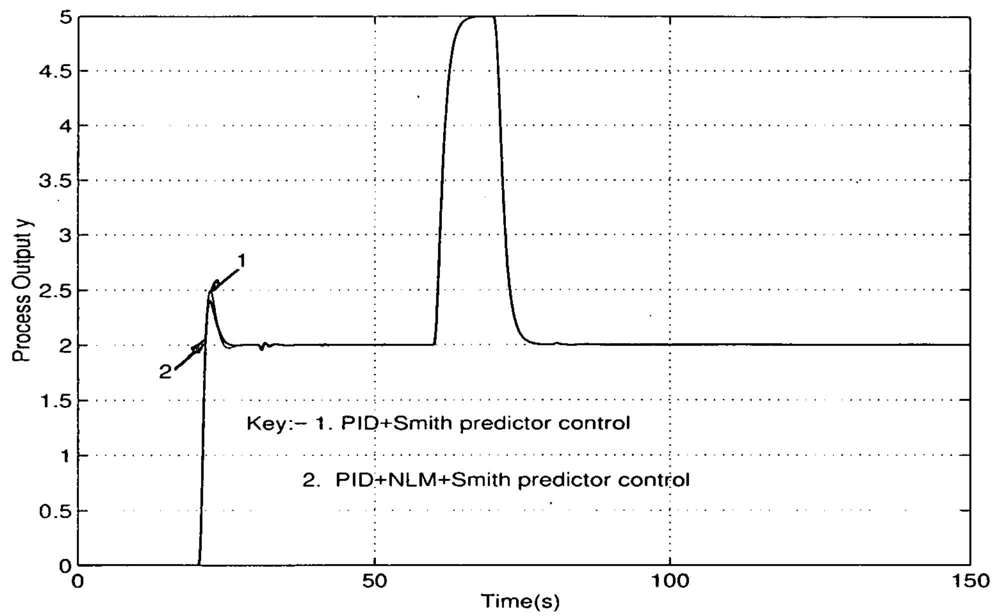

Figure 7-14: Optimal response of the fourth order system to two unit steps when L_{proc} is 10[s]. (Case 2)

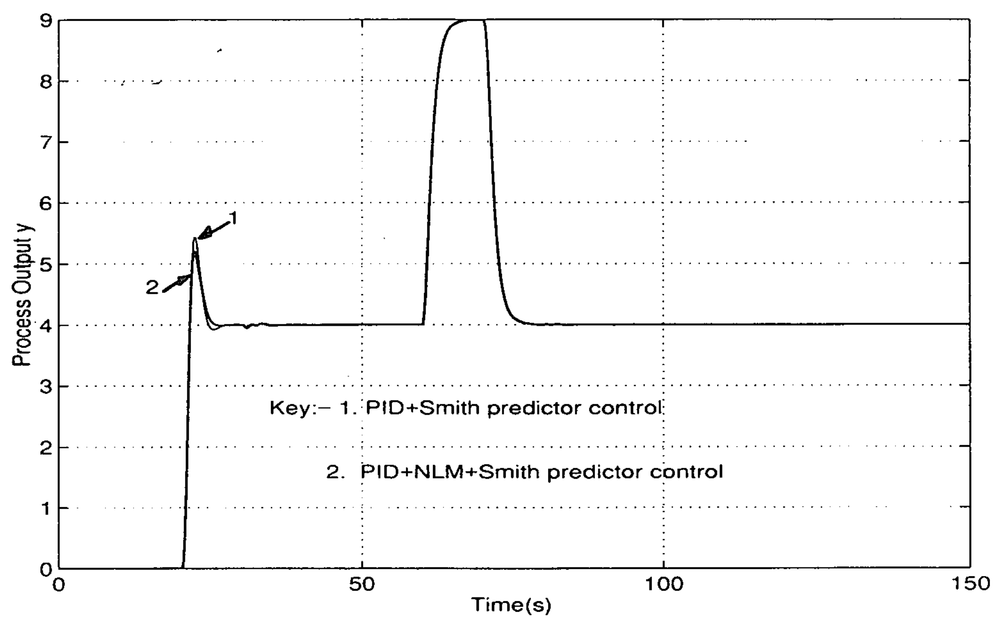

Figure 7-15: Optimal responses of the fourth order system to four unit steps when L_{proc} is 10[s]. (Case 3)

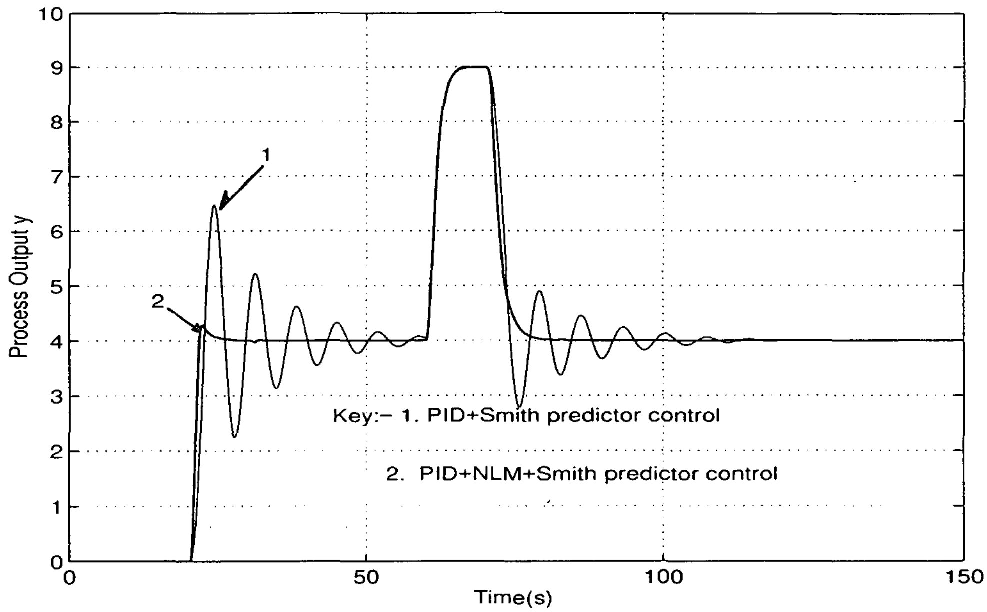

Figure 7-16: Step response of the fourth order system when the controller is not optimally tuned and L_{proc} is 10[s]. (Case 4)

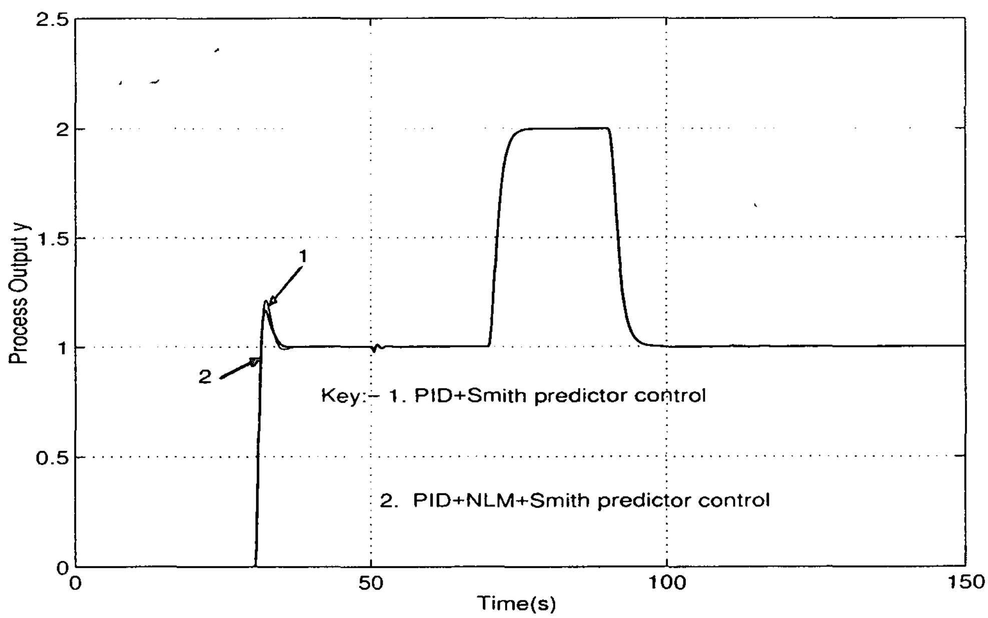

Figure 7-17: Optimal step response of the fourth order system when L_{proc} is 20[s]. (Case 1)

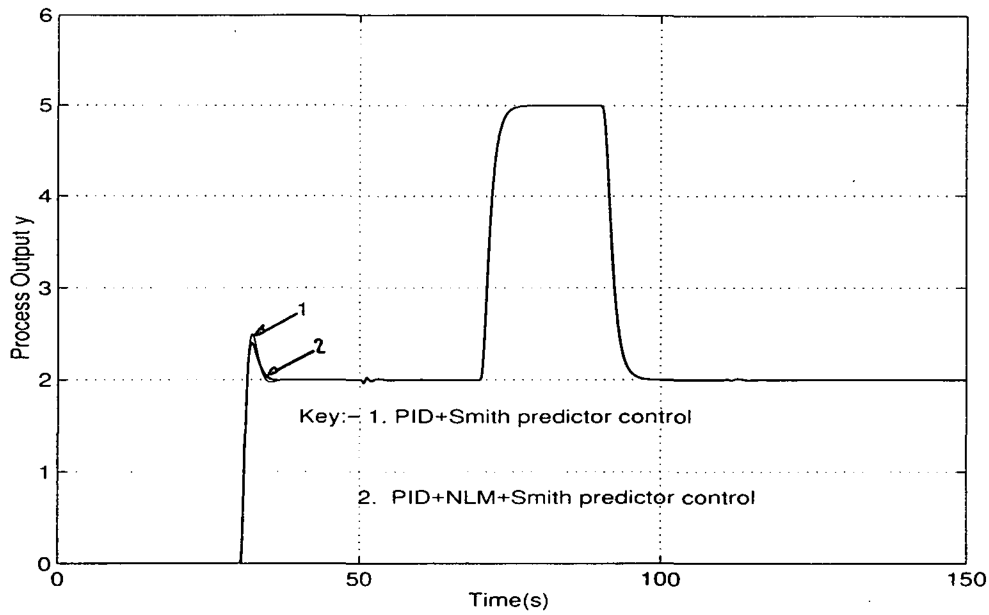

Figure 7-18: Optimal response of the fourth order system to two unit steps when L_{proc} is 20[s]. (Case 2)

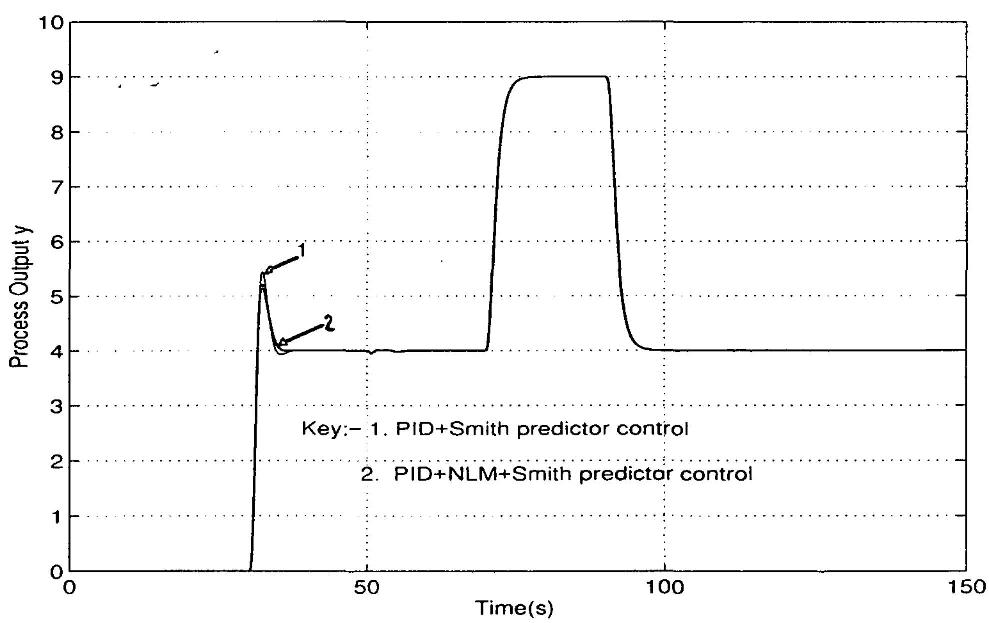

Figure 7-19: Optimal response of the fourth order system to four unit steps when L_{proc} is 20[s]. (Case 3)

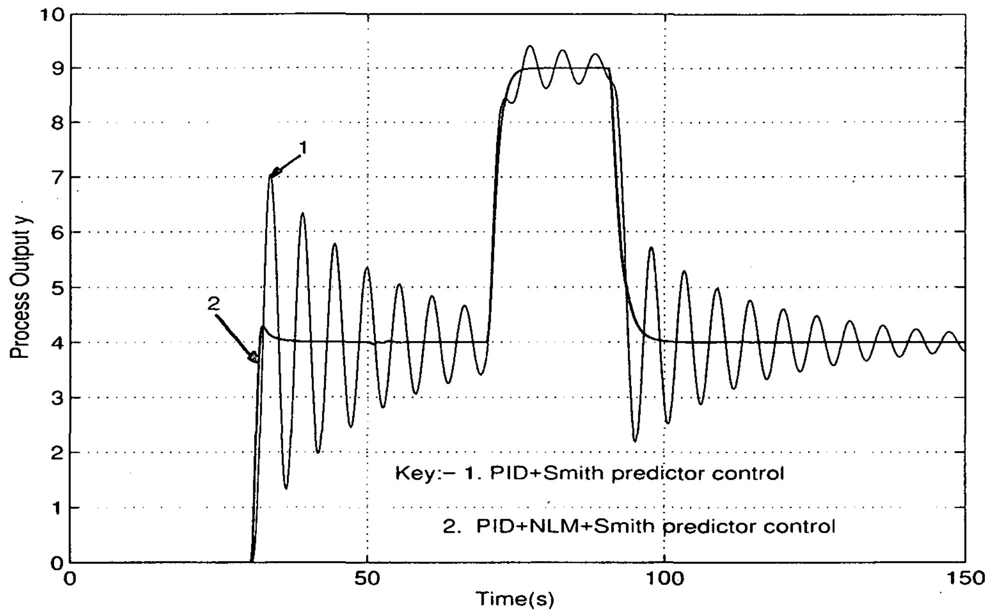

Figure 7-20: Step response of the fourth order system when the controller is not tuned properly and for L_{proc} of 20[s]. (Case 4)

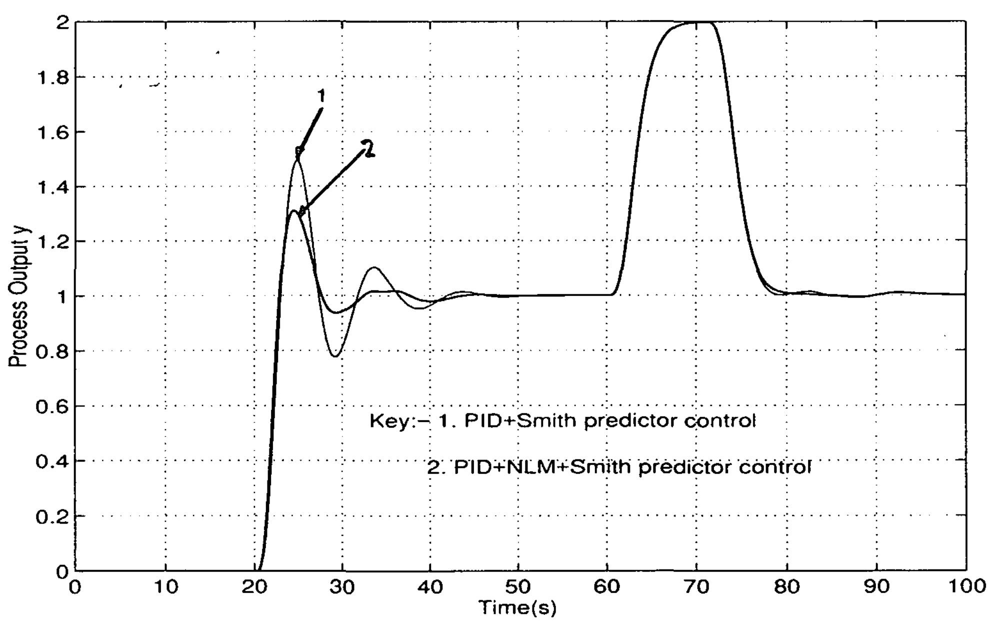

Figure 7-21: Optimal step response for the sixth order system when L_{proc} is 10[s]. (Case 5)

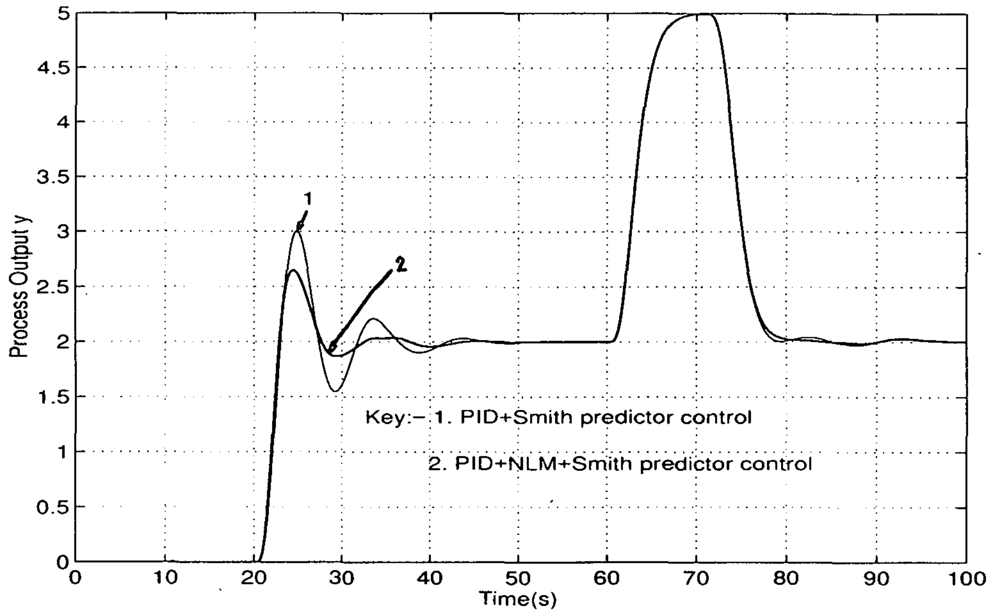

Figure 7-22: Optimal response of the sixth order system to two unit steps when L_{proc} is 10[s]. (Case 6)

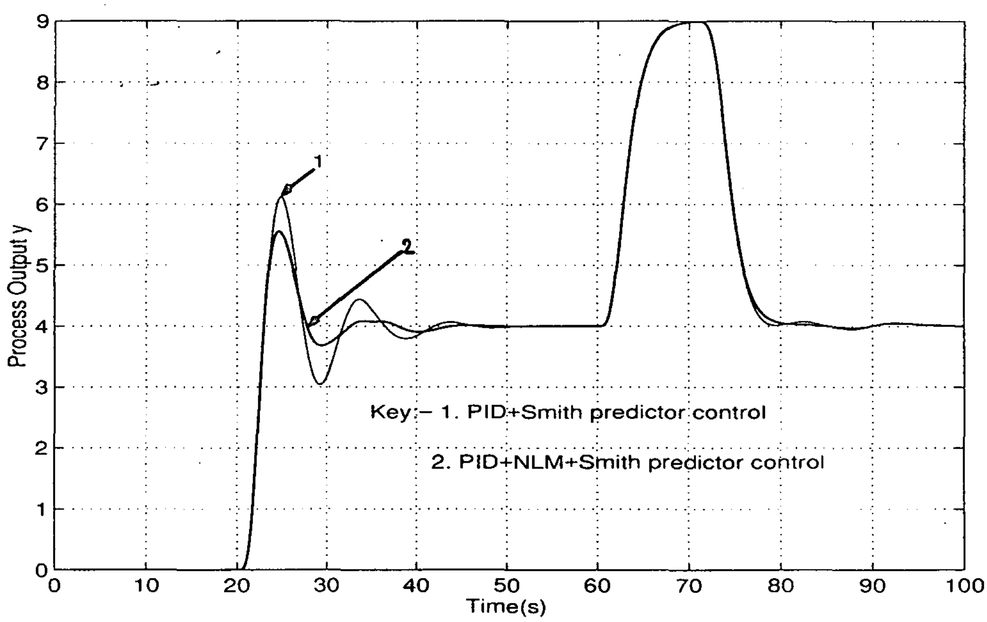

Figure 7-23: Optimal response of the sixth order system to four unit steps when L_{proc} is 10[s]. (Case 7)

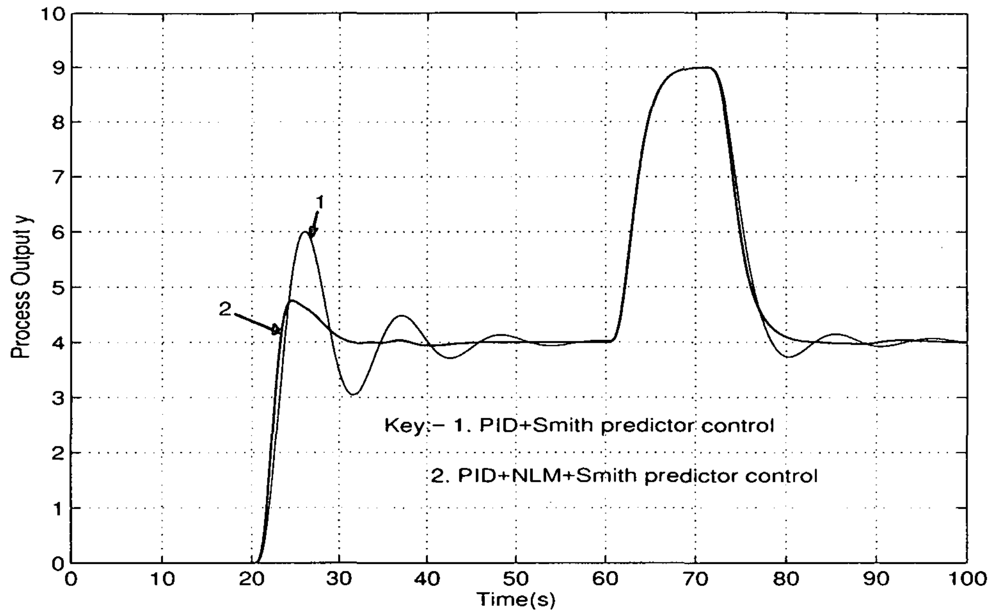

Figure 7-24: Response of the fourth order system to four unit steps when the controller is not tuned properly and L_{proc} is 10[s]. (Case 8)

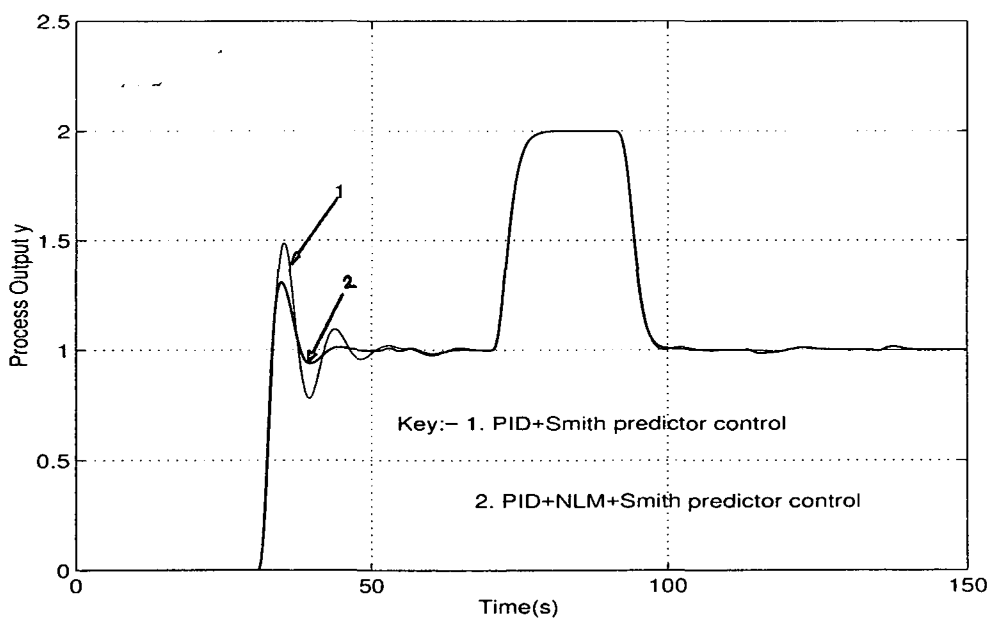

Figure 7-25: Optimal step response of the sixth order system when L_{proc} is 20[s]. (Case 5)

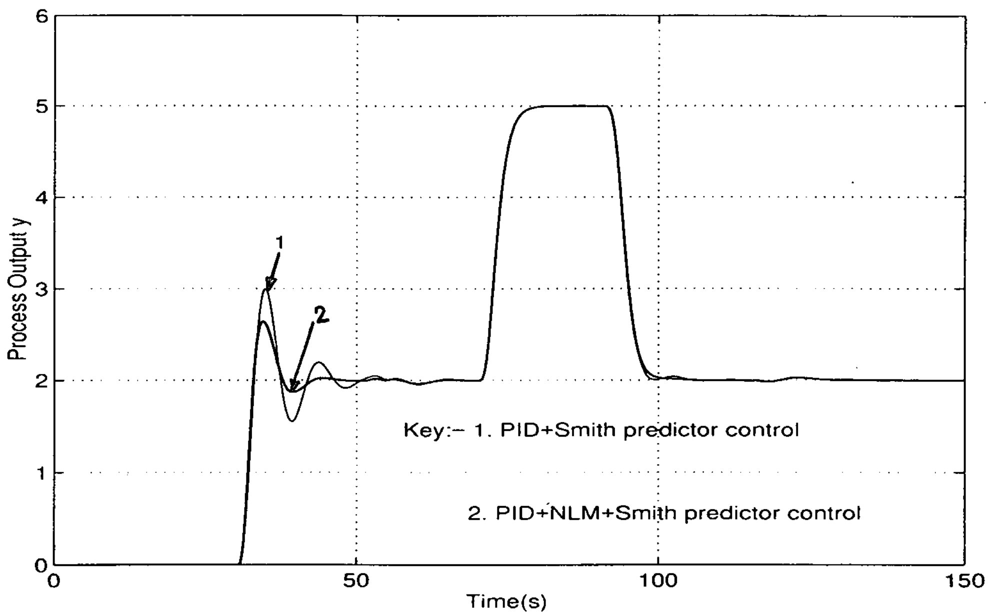

Figure 7-26: Optimal response of the sixth order system to two unit steps when L_{proc} is 20[s]. (Case 6)

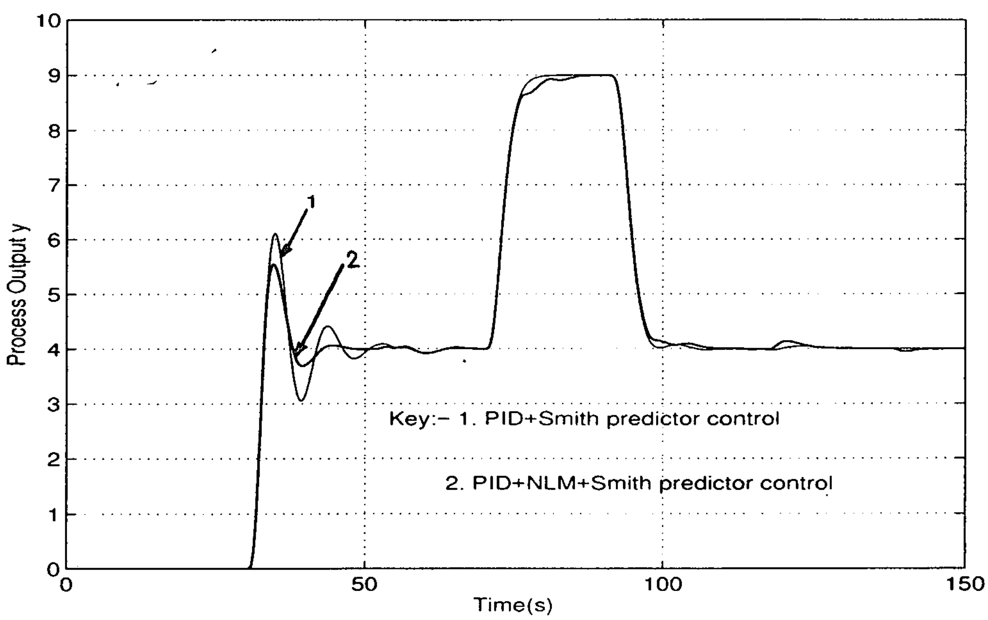

Figure 7-27: Optimal response of the sixth order system to four unit steps when L_{proc} is 20[s]. (Case 7)

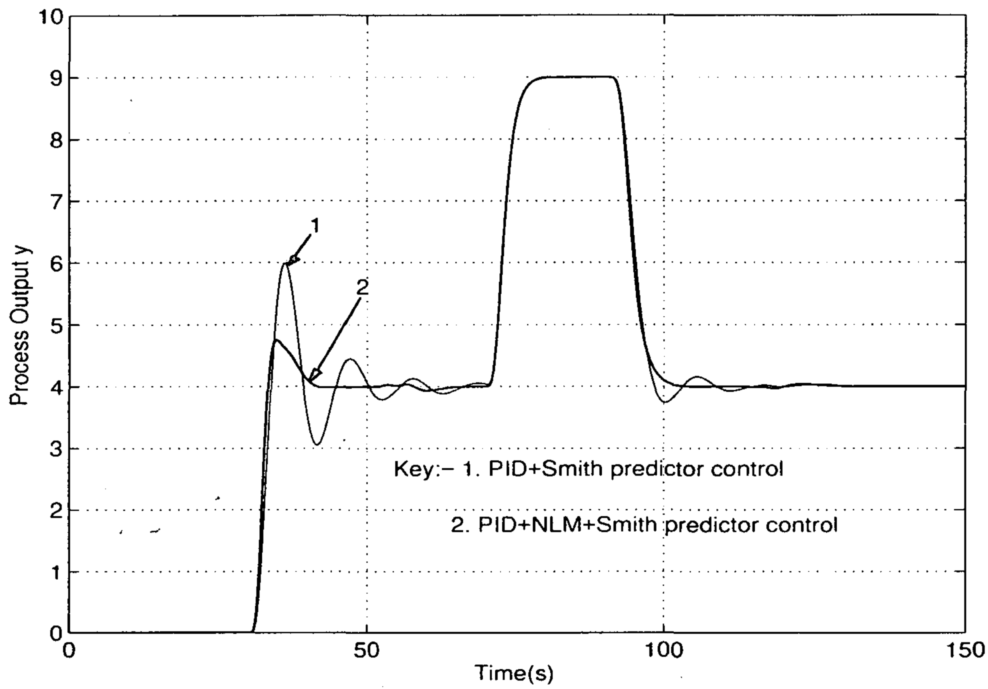

Figure 7-28: Response of the sixth order system to four unit steps when the controller is not tuned properly and for L_{proc} of 20[s]. (Case 8)

7.5 Experiment 7.4: Control of Processes Having Long Dead-Time and Integral Windup

7.5.1 Purpose

The objective of this experiment is to demonstrate the control performance of an *original control scheme*, using a combination of linear and nonlinear control, to positively enhance the control performance of loops experiencing the negative effects of long dead-time and integrator windup.

7.5.2 Method

Consider two process models, of the fourth order and sixth order, as defined by

$$\frac{64}{(s+8)(s+4)(s+2)(s+1)}\exp(-L_{p1}s)$$

and

$$\frac{64}{[(s+1)(s+2)(s+4)]^2}\exp(-L_{p2}s)$$

with $L_{p1} = 10$[s]. Smith predictor control is used to reduce the sensitivity of the controller to long dead-time. For the fourth order process model, the model chosen for cancellation of the actual process output y, in the first parallel feedback loop of the Smith predictor scheme is

$$\begin{aligned} G_{p1}^{m1}(s) &= \frac{0.0365s^2 + 1.3087s + 2.8008}{s^3 + 3.972s^2 + 5.8549s + 2.8007} \\ L_{p1}^{m1} &= 0.2575. \end{aligned}$$

An additional parallel feedback loop will contain a lower order model for proper identification of the process, namely

$$\begin{aligned} G_{p1}^{m2}(s) &= \frac{1.5523}{s^2 + 2.4884s + 1.5524} \\ L_{p1}^{m2} &= 0.3381. \end{aligned}$$

For the second process as defined by the sixth order transfer function with $L_{p2} = 10$[s], the corresponding models used in the Smith predictor scheme are

$$\begin{aligned} G_{p2}^{m1}(s) &= \frac{0.0639s^3 + 0.0348s^2 + 1.4656s + 0.7597}{s^4 + 4.6507s^3 + 5.8618s^2 + 3.6052s + 0.7597} \\ L_{p2}^{m1} &= 10.7175 \end{aligned}$$

and

$$\begin{aligned} G_{p2}^{m2}(s) &= \frac{0.5135}{s^2 + 1.2608s + 0.5135}, \\ L_{p2}^{m2} &= 11.0217. \end{aligned}$$

The output of the controller is fed into the actuator which has its saturation limits set at -10 and $+10$. The rest of the control system is as described previously in Section 6.5. The control algorithm of our controller, with Smith predictor compensation, anti-windup and a nonlinear modifier is defined by (6.8), and by (6.5) - (6.7. The control performance of our controller, as shown in Fig.6-4, will be tested in loops having a dead-time of 10[s] and 20[s], respectively. The PID controller is tuned to optimise disturbance rejection. Gain K_w of the windup protection scheme is adjusted to reduce the magnitude of any overshoot which may occur as a result of integrator saturation. The gain of the nonlinear modifier (K_f) is tuned to enhance the control performance of the loop. Our comparison will be done for systems using Smith predictor control and :

- **Case 1:** PID control with no antiwindup and no nonlinear modifier;

$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$							
Parameters	M_r	M_d	K_p	K_i	K_d	K_f	K_w
A. case 1	1	1	20	20	10	0	0
A. case 2	1	1	20	20	10	0	1
A. case 3	1	1	20	20	10	13	0
A. case 4	1	1	20	20	10	13	1
B. case 1	2	3	20	20	10	0	0
B. case 2	2	3	20	20	10	0	1
B. case 3	2	3	20	20	10	13	0
B. case 4	2	3	20	20	10	13	1
C. case 1	4	5	20	20	10	0	0
C. case 2	4	5	20	20	10	0	1
C. case 3	4	5	20	20	10	13	0
C. case 4	4	5	20	20	10	13	1

Table 7.16: Optimal tuning parameters for the fourth order system

$G_{p1}(s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}$							
Parameters	M_r	M_d	K_p	K_p	K_i	K_f	K_w
D. case 2	1	1	5	5	16	1	0
D. case 4	1	1	5	5	16	1	13

Table 7.17: Improper tuning parameters for the fourth order system

- **Case 2:** PID control with reset windup protection;
- **Case 3:** PID control in conjunction with the nonlinear modifier and
- **Case 4:** PID control with windup protection and a nonlinear modifier.

The optimal tuning parameters for the fourth and sixth order process models are shown in Table 7.16 and Table 7.20. Table 7.17 lists a possible set of parameters (for the fourth order process model) when the controller is not tuned properly.

Case	L_{proc}[s]	**Figure Ref. No.**
A	10	7.29
B	10	7.30
C	10	7.31
D	10	7.35

Table 7.18: Corresponding figures for the fourth process model when Lp = 10[s]

Case	L_{proc}[s]	**Figure Ref. No.**
A	20	7.32
B	20	7.33
C	20	7.34
D	20	7.35

Table 7.19: Corresponding figures for the fourth order process model when Lp = 20[s]

$G_{p2}(s) = \frac{64}{[(s+1)(s+2)(s+4)]^2}$							
Parameters	M_r	M_d	K_p	K_i	K_d	K_f	K_w
E. case 1	1	1	6.8411	10	4.9995	0	0
E. case 2	1	1	6.8411	10	4.9995	0	0.4
E. case 3	1	1	6.8411	10	4.9995	15	0
E. case 4	1	1	6.8411	10	4.9995	15	0.4
F. case 1	2	3	6.8411	10	4.9995	0	0
F. case 2	2	3	6.8411	10	4.9995	0	0.4
F. case 3	2	3	6.8411	10	4.9995	15	0
F. case 4	2	3	6.8411	10	4.9995	15	0.4
G. case 1	4	5	6.8411	10	4.9995	0	0
G. case 2	4	5	6.8411	10	4.9995	0	0.4
G. case 3	4	5	6.8411	10	4.9995	15	0
G. case 4	4	5	6.8411	10	4.9995	15	0.4

Table 7.20: Optimal tuning parameters for the sixth order process

Case	L_p[s]	**Figure Ref. No.**
E	10	7.36
F	10	7.37
G	10	7.38

Table 7.21: Corresponding figures for the sixth order process model when Lp = 10[s]

Case	L_p[s]	**Figure Ref No.**
E	20	7.39
F	20	7.40
G	20	7.41

Table 7.22: Corresponding figures for the sixth order process model when Lp = 20[s]

7.5.3 Observation

The results of the simulations are presented in Fig. 7-29 to Fig. 7-41. One of the observations from these results is that antiwindup schemes yield a better degree of control by reducing overshoot caused by integral windup when an actuator saturates. The nonlinear modifier improves the tracking and rejection properties of the linear controller. Combining the nonlinear modifier with antiwindup and Smith predictor control positively enhances the performance of the control loop with respect to setpoint tracking, keeping disturbance rejection good. Even in instances when the controller is not properly tuned (which is quite common in practice), the contributions of the anti-windup scheme and the nonlinear modifier have a strong stabilising effect on loop performance when compared to the conventional controller.

7.5.4 Conclusion

The antiwindup scheme contributes positively towards improving setpoint tracking by reducing the magnitude of the initial overshoot. This overshoot is caused by integral windup as a result of actuator saturation. The improvement in the setpoint tracking properties of the controller optimises control of the loop and reduces the role of the plant operator, who is usually in control when the plant starts up at the beginning of a production run.

The simple nonlinear modifier block connected to operate in parallel with the conventional linear controller ensures that the loop remains stable over a wide range of inputs and tuning parameters, and with very good performance. This is not true for the conventional linear PID controller.

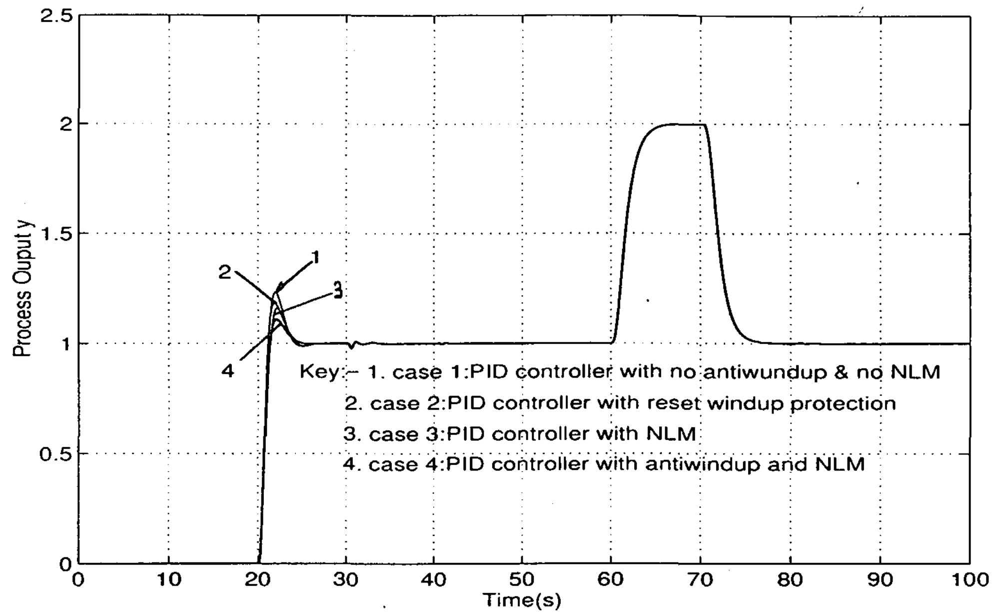

Figure 7-29: Optimal step response of the fourth order system. (Case A)

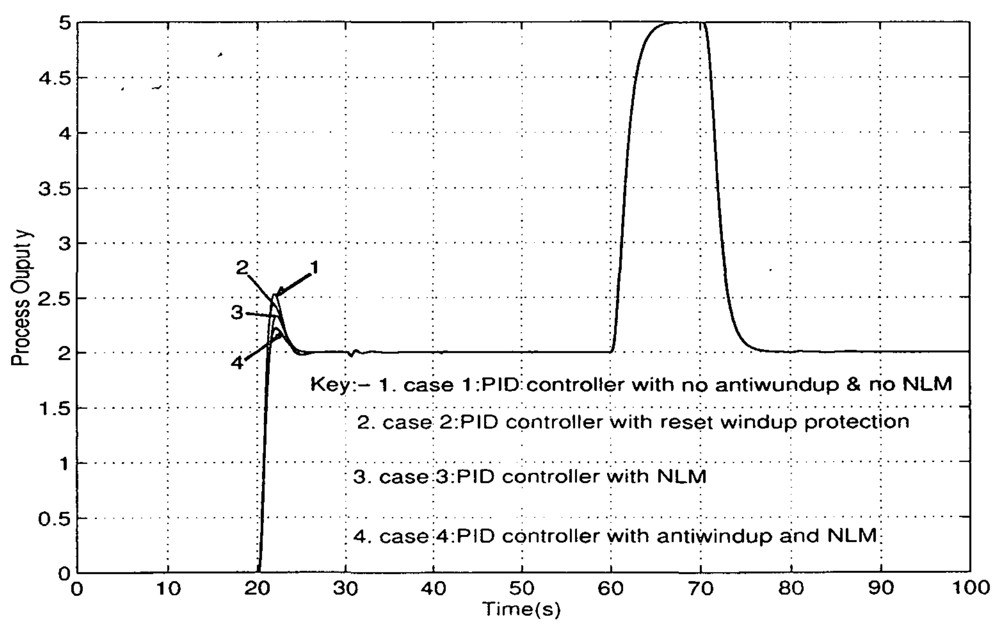

Figure 7-30: Optimal response of the fourth order system to two unit steps. (Case B)

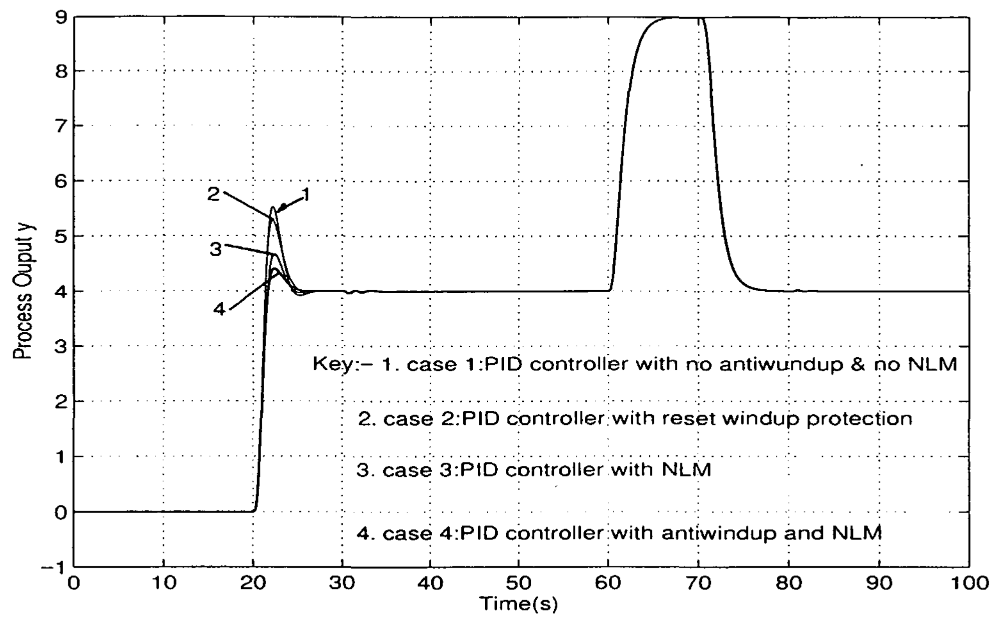

Figure 7-31: Optimal response of the fourth order system to four unit steps. (Case C)

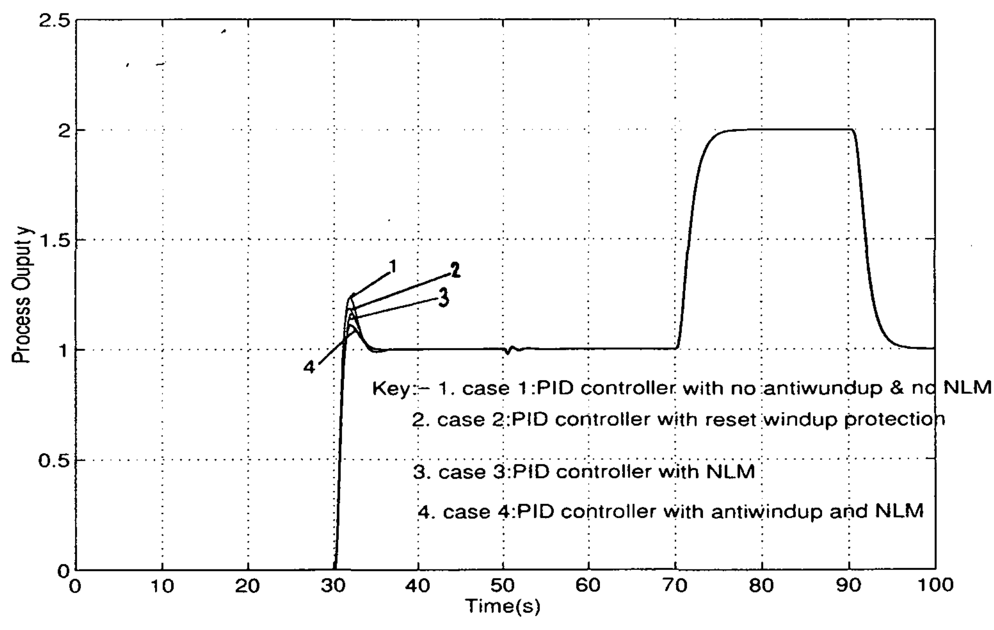

Figure 7-32: Step response of the fourth order system. (Case A)

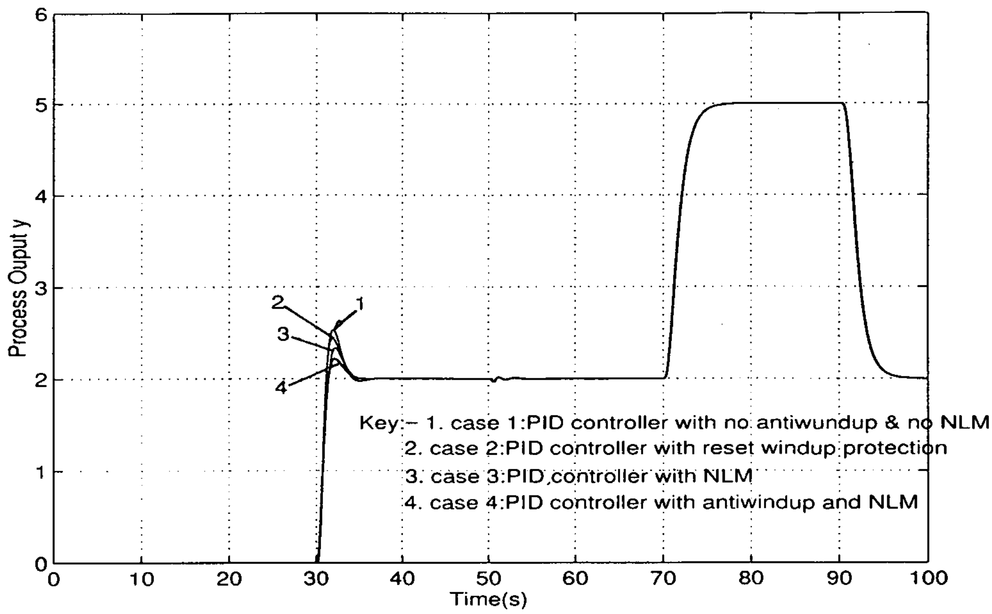

Figure 7-33: Optimal response of the fourth order system to two unit steps. (Case B)

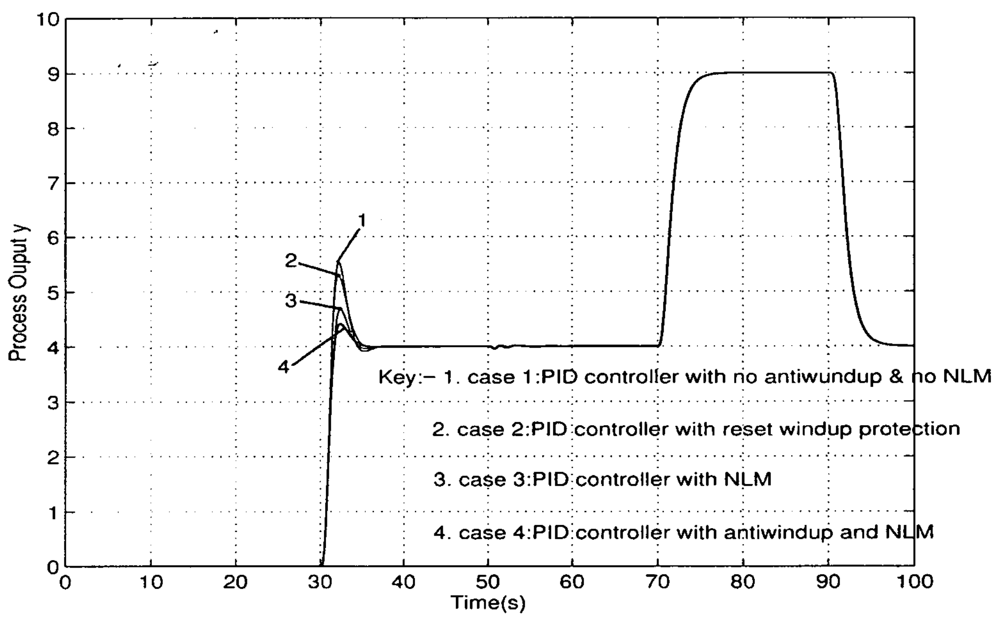

Figure 7-34: Optimal response of the fourth order system to four unit steps. (Case C)

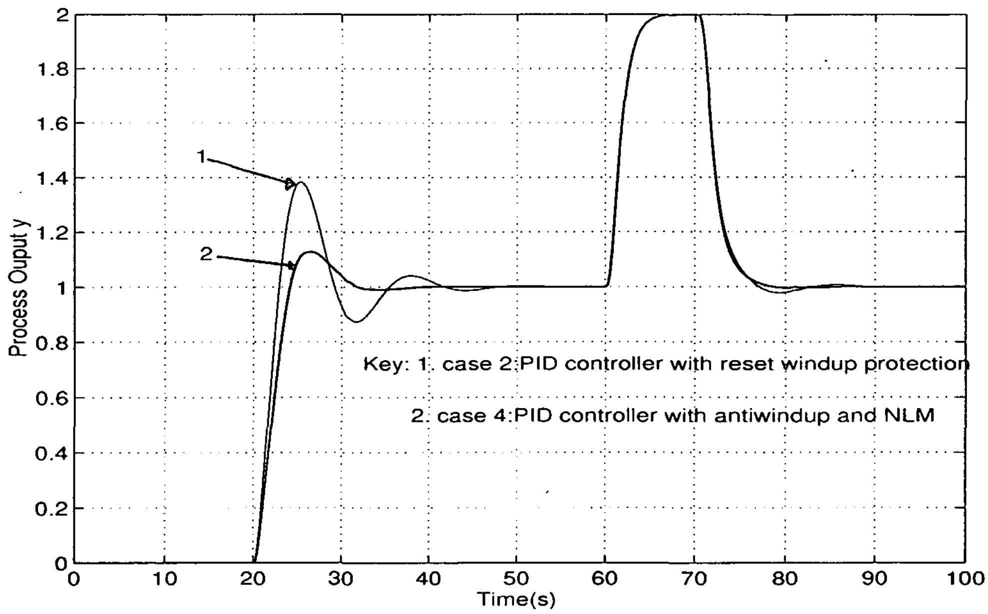

Figure 7-35: Step response of the fourth order process when the controller is not optimally tuned. (Case D)

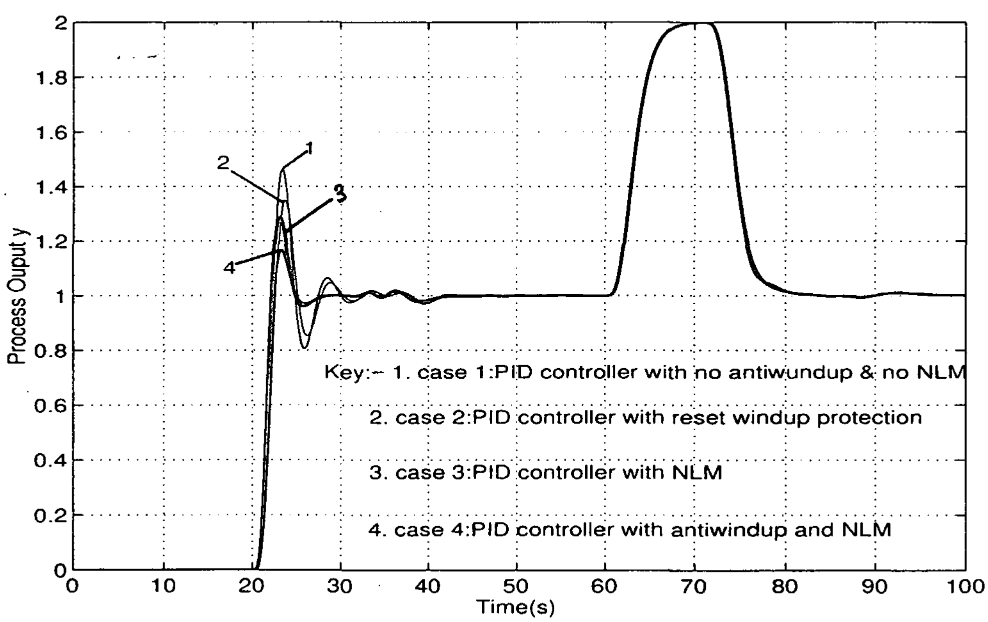

Figure 7-36: Optimal step response for the sixth order system. (Case E)

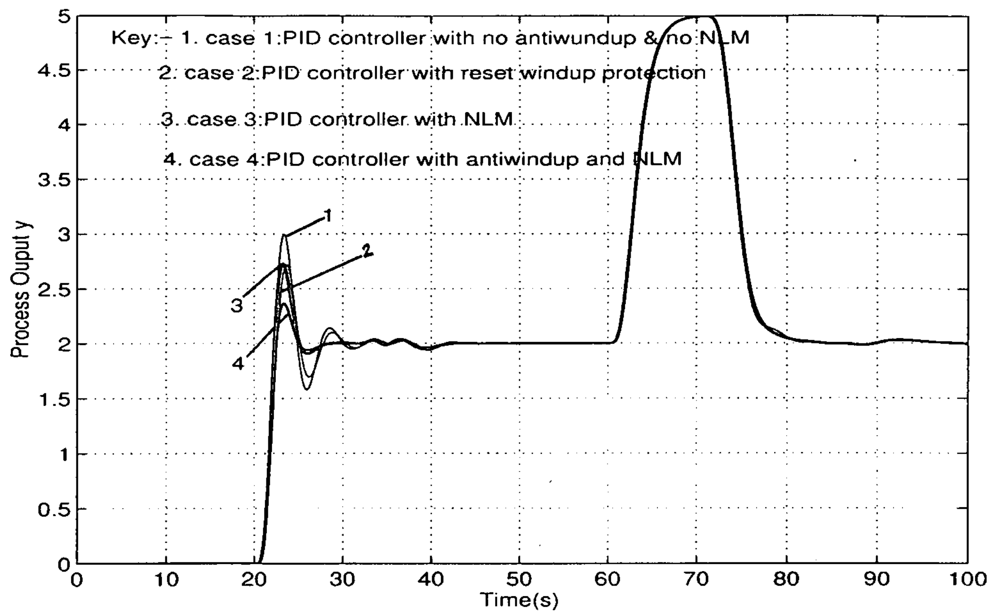

Figure 7-37: Optimal response of the sixth order system to two unit steps. (Case F)

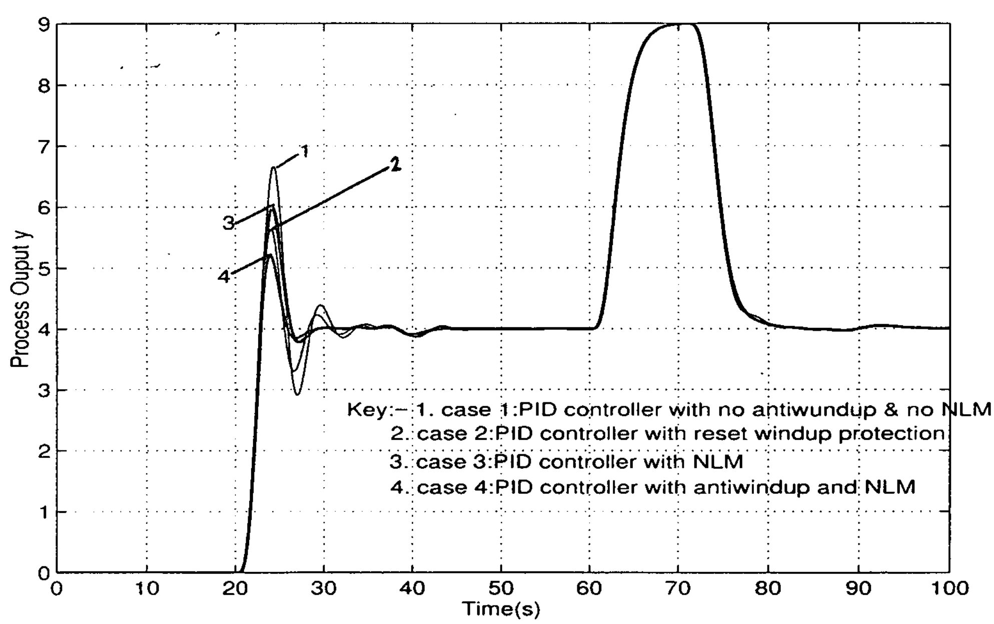

Figure 7-38: Optimal response of the sixth order system to four unit steps. (Case G)

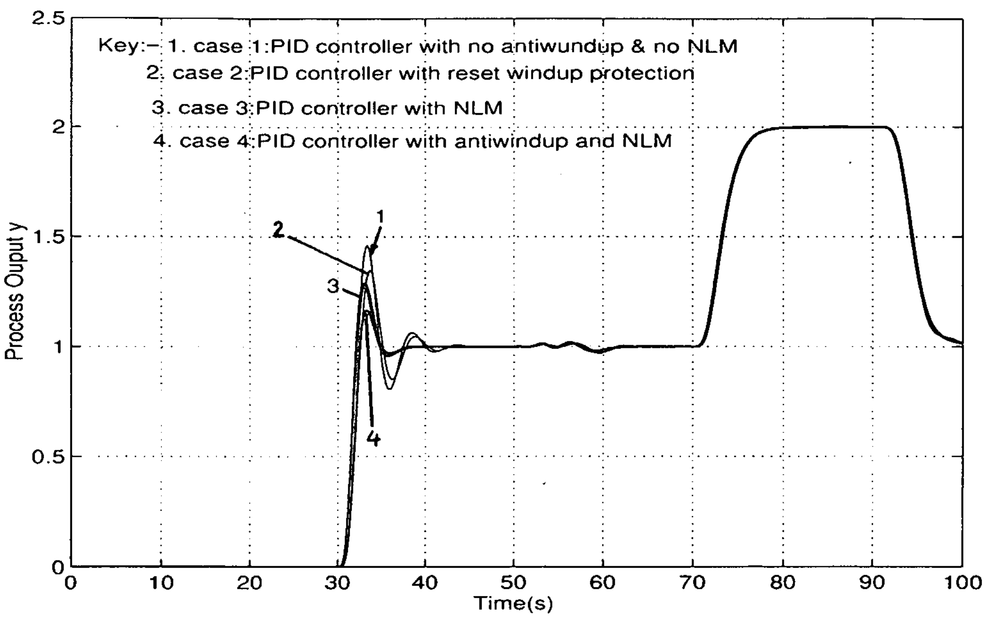

Figure 7-39: Optimal step response of the sixth order system . (Case E)

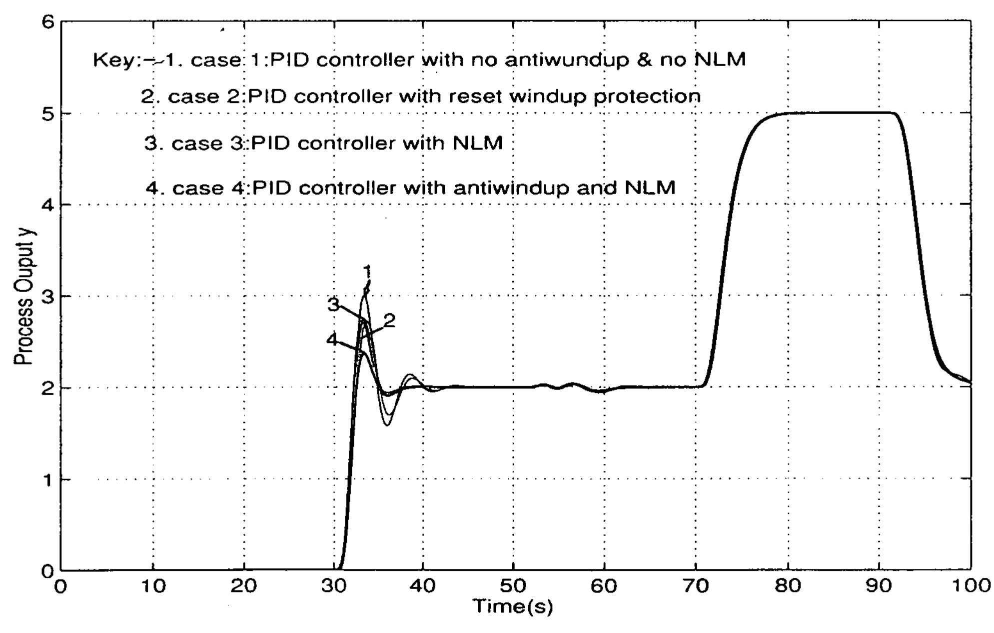

Figure 7-40: Optimal response of the sixth order system to two unit steps. (Case F)

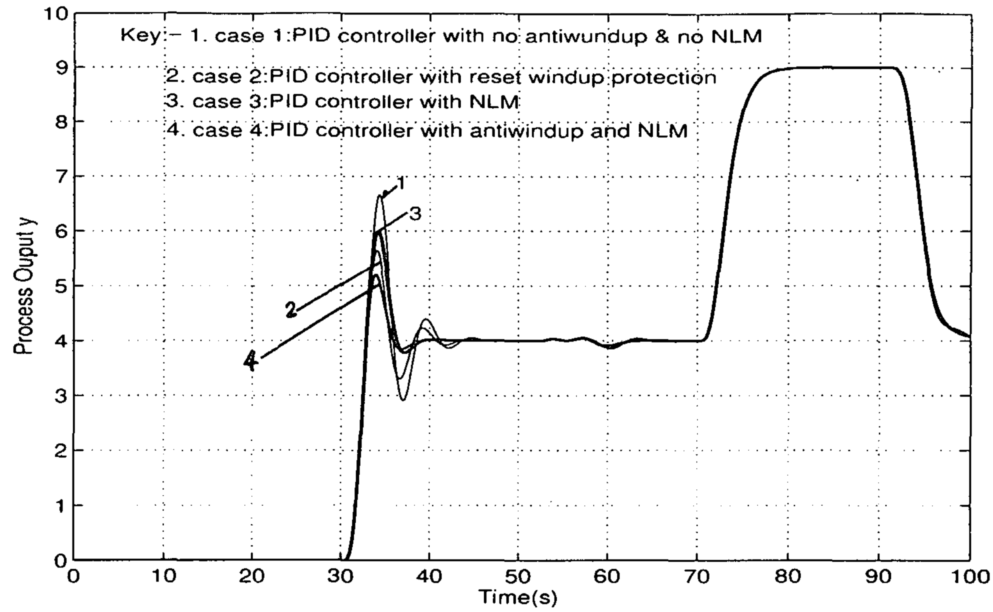

Figure 7-41: Optimal response of the sixth order system to four unit steps. (Case G)

The nonlinear modifier increases the robustness of the loop to a wide range of disturbances, even in instances when the linear controller is not tuned properly. This robustness is exhibited as a strong stabilising effect on system response. The need to optimally tune the controller (which is often difficult in practice) is replaced by the simple adjustment of a single gain, K_f.

Finally, we can conclude that combinations of linear and nonlinear control, together with Smith predictor compensation and windup protection, work well in providing good control for loops prone to instability caused by inherent nonlinearities and dead-time. A high degree of quality control has been acheived through the use of a relatively unsophisicated and inexpensive controller.

Chapter 8

The Analogue Electronic Controller

From the results of the simulation as shown in Chapter 4 and Chapter 7, we have observed that the control performance of the linear PID controller with the nonlinear PD modifier is better than that of the gain scheduling PID controller. The study involved rigourous testing of both control schemes within loops which were compensated for saturation nonlinearity and dead-time. During the study it was observed that the PID + NLM controller outperformed the gain scheduling controller for the following reasons:

- Good setpoint tracking and improved robust stability over a wide range of input and noise signals.
- Standard tuning can be used to tune the PID portion of the controller, and the nonlinear modifier can be easily tuned to optimise the control performance through the adjustment of a single gain K_f.
- The simple control scheme makes it possible to use a relatively unsophisicated controller to acheive a good control perfromance.
- The nonlinear modifier can be easily connected to a conventional PID controller.

For the reasons stated above, it was decided to construct the analogue version of the PID + NLM controller. A one degree of freedom, parallel type PID controller was utilised in the experiments. This PID controller with antiwindup, and the nonlinear modifier, was implemented on a breadboard using standard analogue components. The circuit diagram for this analog controller is given in Fig.8-1 to Fig.8-4. The controller was interfaced to an IBM compatible, 166 MHz Pentium personal computer. The interface circuit used was the PC-74 high performance analog I/O board for IBM PC's and compatibles. The PC-74 was accessed via I/O operations performed by the host computer. Analogue voltage input signals from the controller, occupying the 1 to 5 volt range, were converted into 12 bit digital codes at a sampling rate of 10 Hz. This digital code was then transferred to the host processor for processing, using polled I/O. The output of the simulated process is received from the host computer by the interface module and converted to analogue, within the 1 to 5 volt range. This analogue signal forms the feedback signal for the controller. The combination of the electronic controller and the computer was chosen because of the ease with which the process models could be changed within the Matlab/Simulink environment. The following experiments will demonstrate the performance of the PID controller with antiwindup + nonlinear modifier, on processes having long dead-time.

8.1 Experiment

8.1.1 Purpose

The purpose of this experiment is

- to verify the step response obtained in the simulation studies and
- to compare the control performance of the conventional analogue PID controller to that of the PID + NLM controller (in analogue version).

This is done by comparing the responses of the control loop consisting of the process (modelled on the PC in the Matlab/Simulink environment) and the electronic controller.

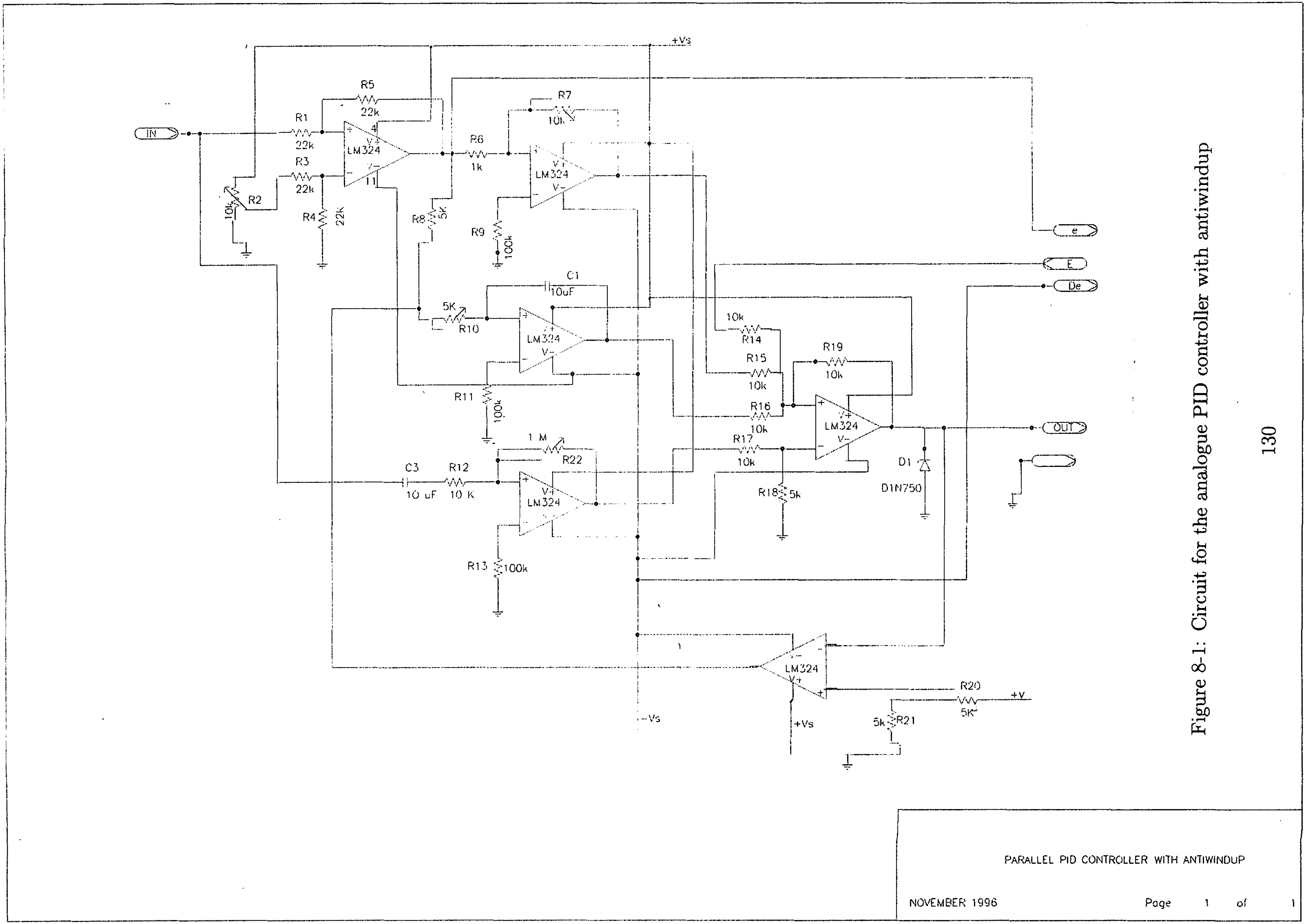

Figure 8-1: Circuit for the analogue PID controller with antiwindup

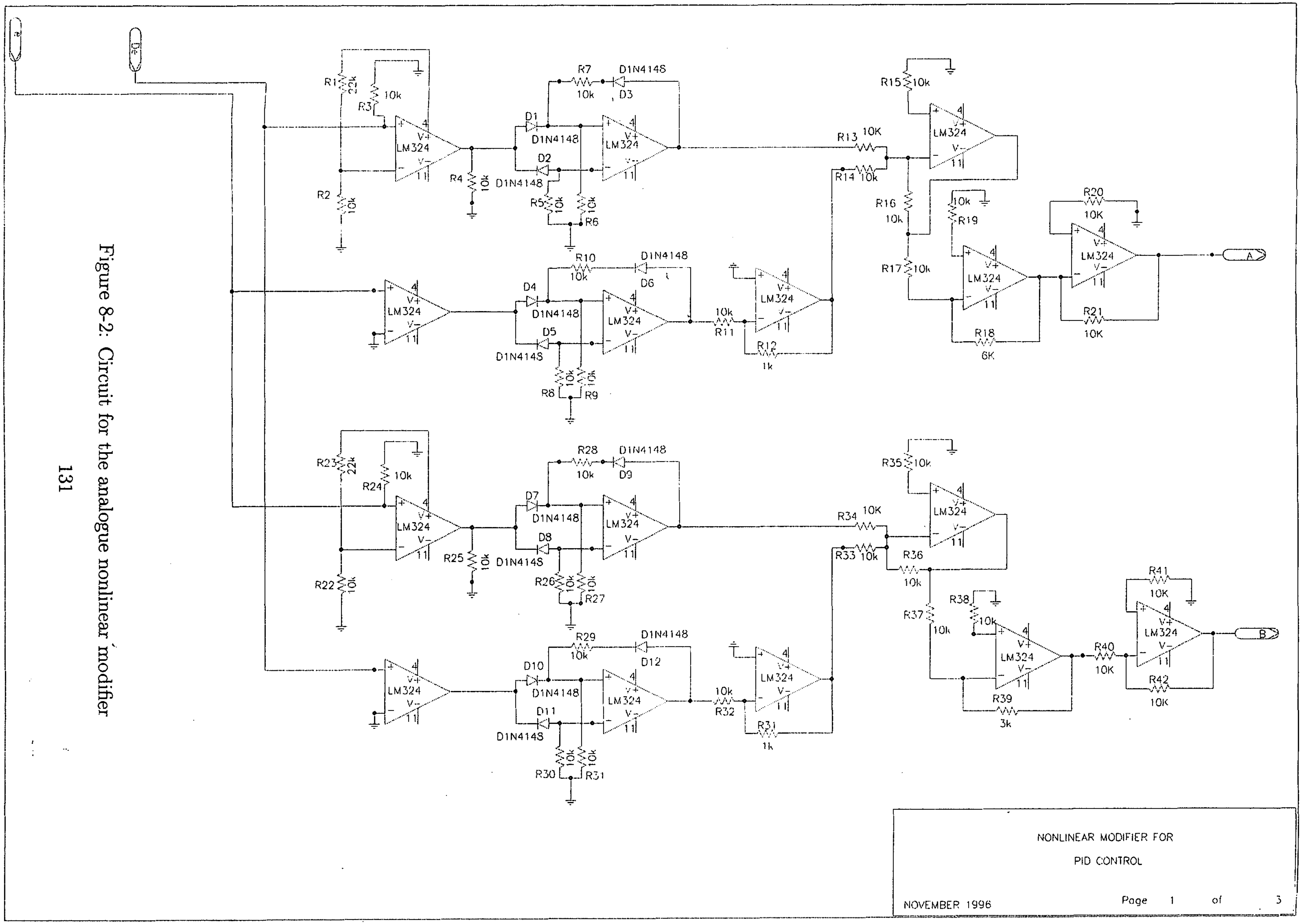

Figure 8-2: Circuit for the analogue nonlinear modifier

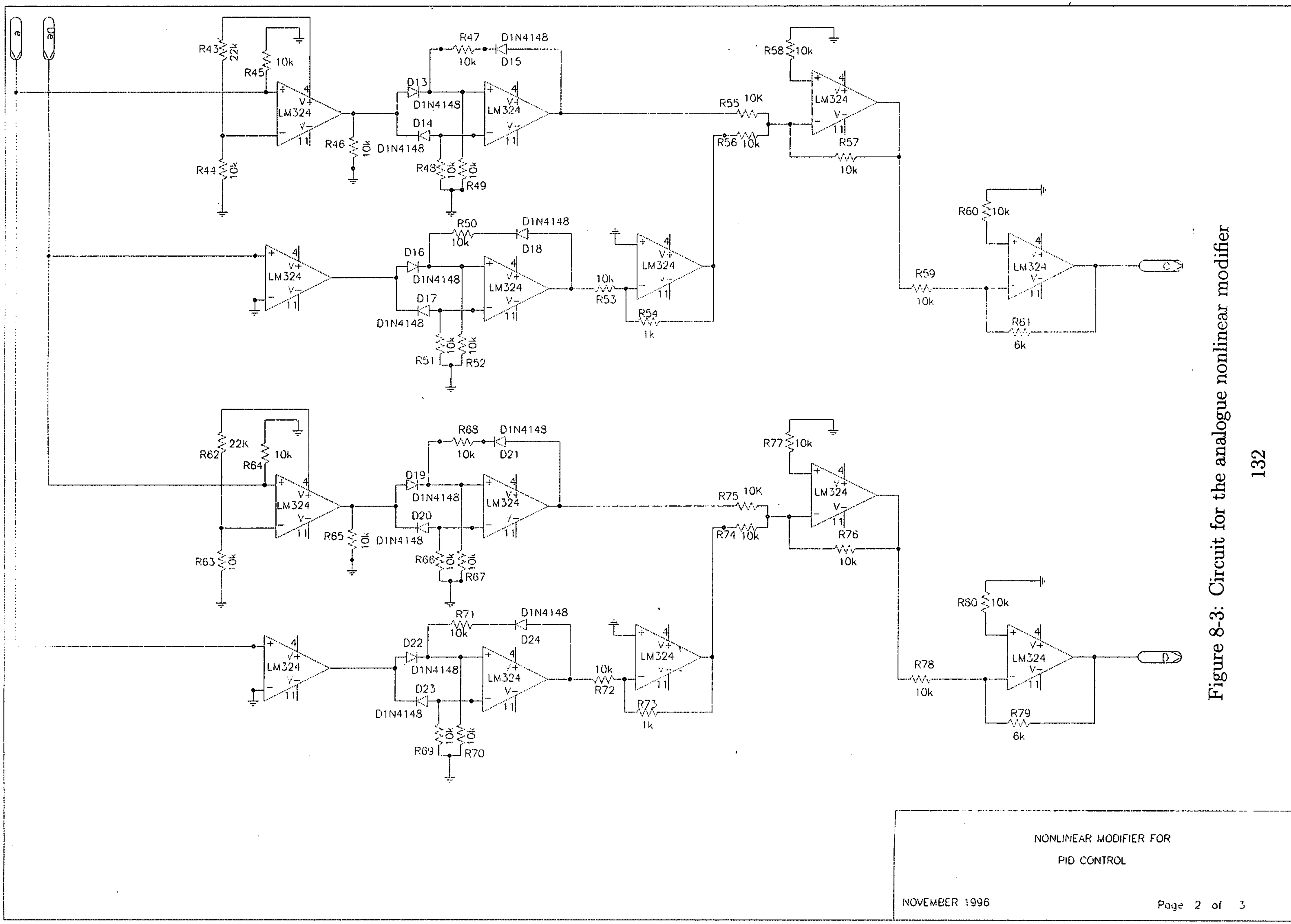

Figure 8-3: Circuit for the analogue nonlinear modifier

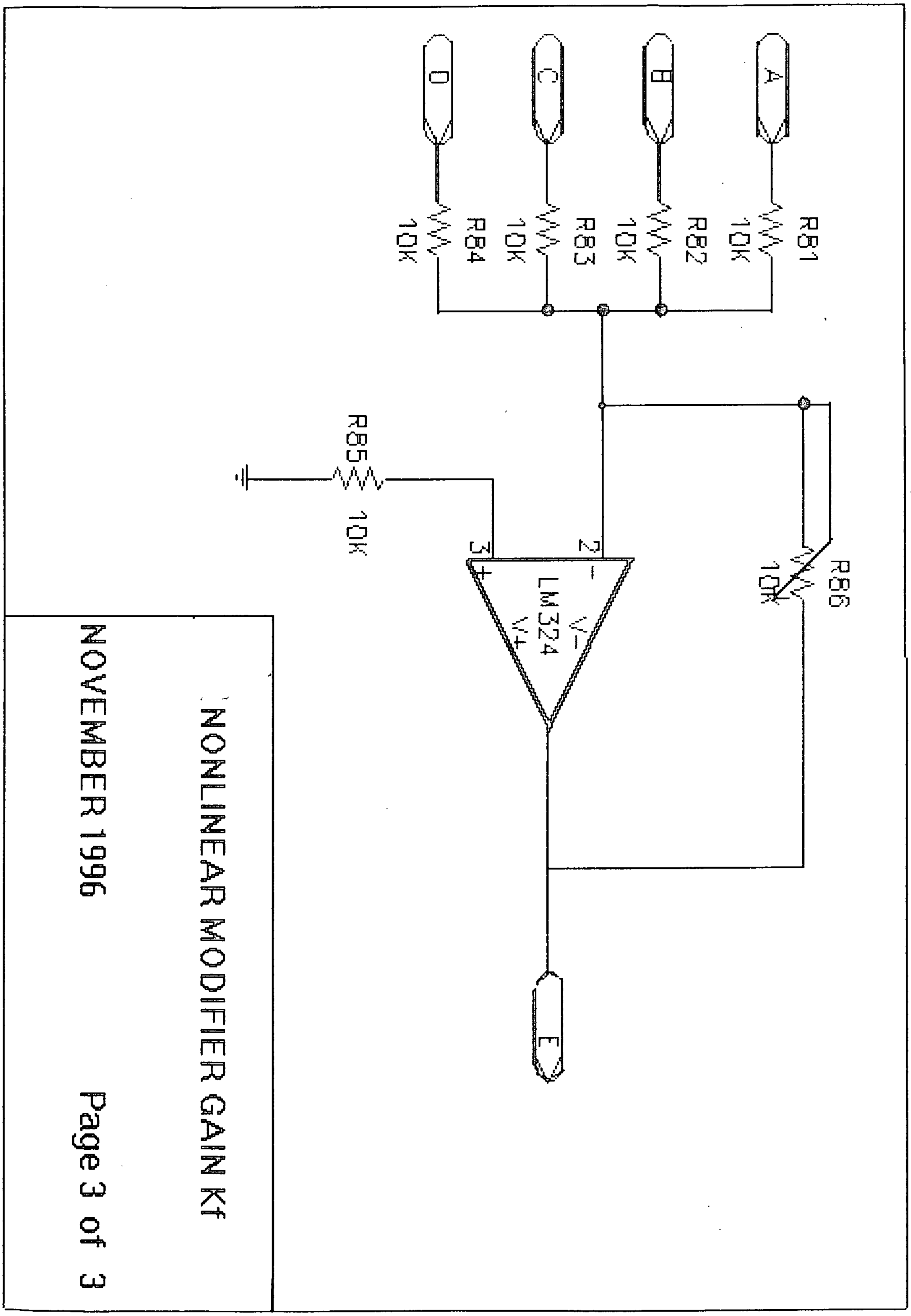

Figure 8-4: Circuit for analogue nonlinear modifier gain K_f

case	M_r	M_d	K_p	K_i	K_d	K_f	K_w	Figure Ref. No.
1	1	1	20	20	10	0	0	8-5
2	1	1	20	20	10	0	1	8-6
3	1	1	20	20	10	13	0	8-7
4	1	1	20	20	10	13	1	8-8

Table 8.1: Simulation parameters used in the test

8.1.2 Method

A control system having dead-time was simulated on the computer using Matlab/Simulink. The simulated fourth order process model is used to test the step response of the PID + NLM and is given by

$$\frac{Y(s)}{U(s)} = G_p(s)\exp(-L_p s) = \frac{64}{(s+8)(s+4)(s+2)(s+1)}\exp(-L_p s)$$

where $L_p = 10$[s]. The models chosen for the Smith predictor scheme are the same as in Experiment 7.4, i.e.

$$G_{m1}(s) = \frac{0.0365s^2 + 1.3087s + 2.8008}{s^3 + 3.972s^2 + 5.8549s + 2.8007}, L_{m1} = 0.2575$$

and

$$G_{m2}(s) = \frac{1.5523}{s^2 + 2.4884s + 1.5524}, \quad L_{m2} = 0.3381$$

These Smith predictor models were simulated on the PC within the Matlab/Simulink environment. The output of the controller u_c is fed into the input of an actuator. The actuator feeds the process. The controller was first tuned for a satisfactory response and then detuned. Table 8.1 lists the simulation parameters for a well tuned controller and Table 8.2 lists the parameters when the controller was detuned.

M_r	M_d	K_p	K_i	K_d	K_w	K_f	Figure Ref. No.
1	1	1	5	5	1	0	8-9
1	1	1	5	5	1	13	8-9
1	1	5	30	20	1	0	8-10
1	1	5	30	20	1	13	8-10
1	1	7	40	2	1	0	8-11
1	1	7	40	2	1	13	8-11

Table 8.2: Simulation parameters when the controllers are detuned.

8.1.3 Observations

The step responses of the closed-loop system with the analogue controller are shown in Fig.8-5 to Fig.8-11. A comparison of these responses shows the following:

- When the PID controller without the NLM is properly tuned (Table 8.1), the improvements obtained with the addition of the NLM is minimal - compare Fig.8-5 and Fig.8-7, as well as Fig. 8-6 and Fig.8-8.

- If the PID controller is not tuned well (situations shown in Table 8.2 and Fig.8-9 to Fig.8-11), then the presence of the NLM shows an extremely positive effect on the performance of the closed loop system, i.e. the nonlinear modifier provides considerably better setpoint tracking and disturbance rejection.

8.1.4 Conclusion

The responses obtained during the tests verify the simulation results given in Chapter 7. PID control with windup and dead-time compensation yields good control, but initial overshoot upon plant startup exists, albeit at a much reduced level. Combining the linear PID controller with antiwindup and Smith predictor control, together with a nonlinear modifier, provides a good degree of improvement in the performance of the control loop in the cases of the detuned controller. This is the most significant result from the viewpoint of practical implementation.

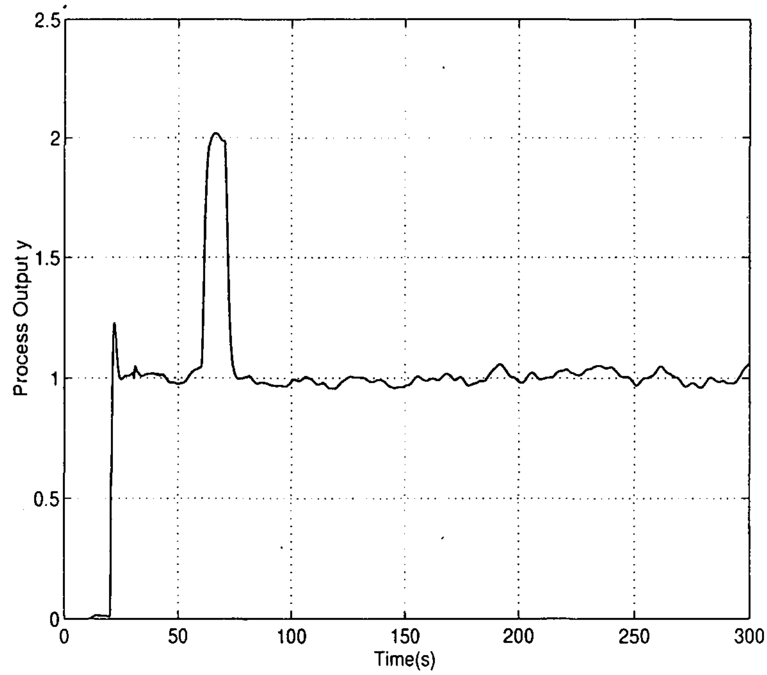

Figure 8-5: Step response of the analogue controller without the nonlinear modifier and without antiwindup. (Case 1)

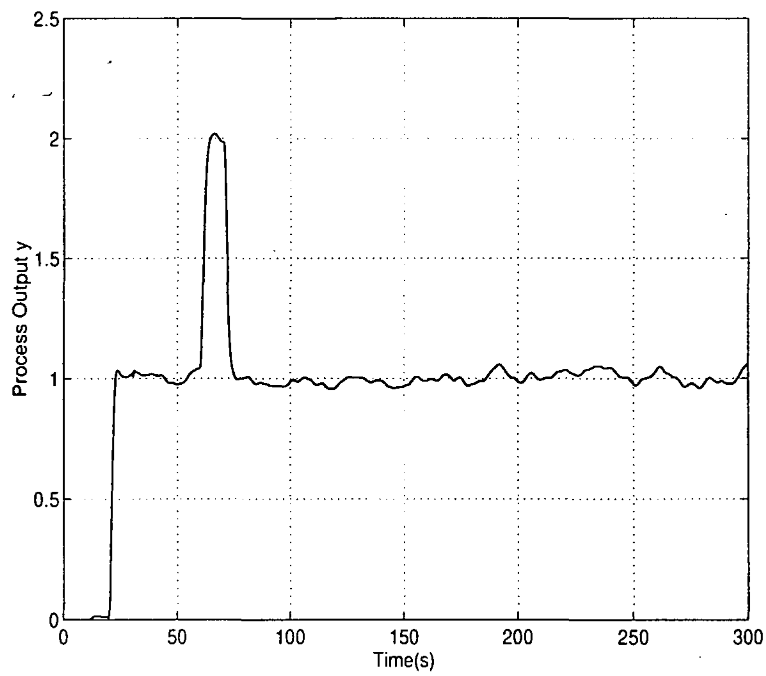

Figure 8-6: Step response of the analogue controller with antiwindup. (Case 2)

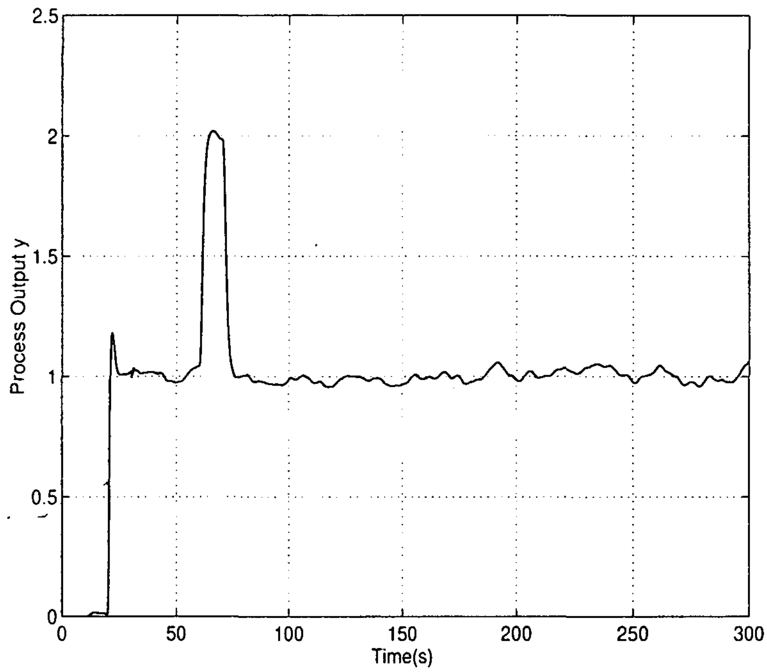

Figure 8-7: Step response using the analogue controller and the nonlinear modifier. (Case 3)

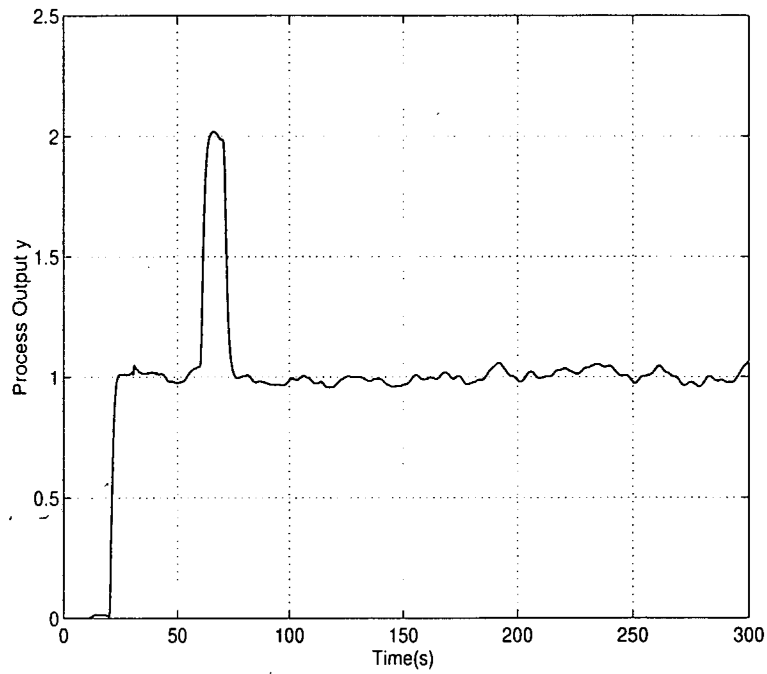

Figure 8-8: Step response of the analogue controller using the nonlinear modifier and antiwindup. (Case 4)

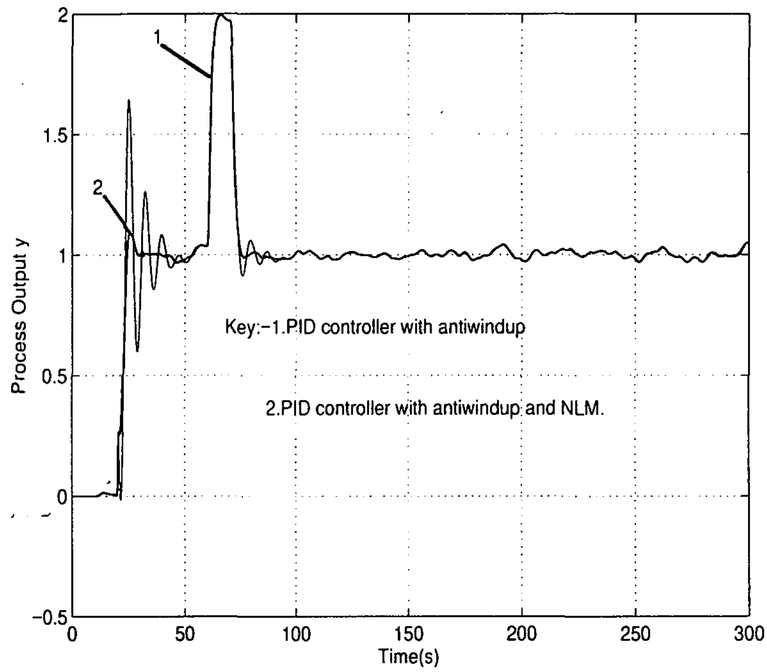

Figure 8-9: Comparing the step response of the analogue PID controller to that of the PID controller + NLM, when the controller is detuned.

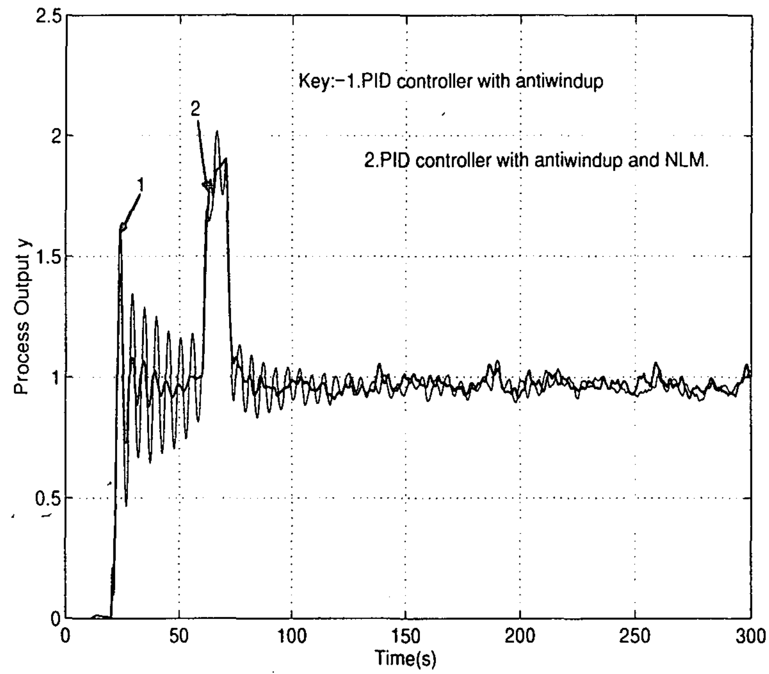

Figure 8-10: Comparing the step response of the analogue PID controller to that of the PID controller + NLM, when the controller is detuned.

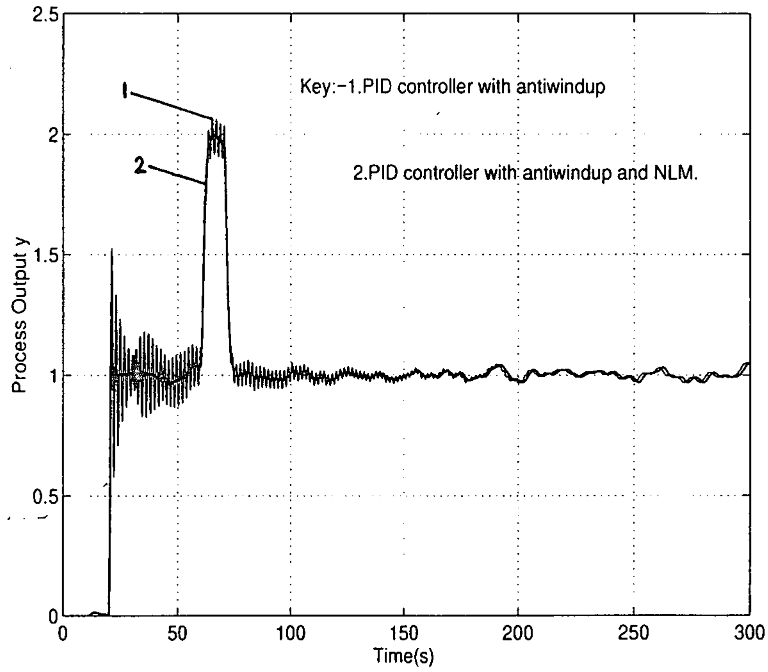

Figure 8-11: Comparing the step response of the PID controller to that of the PID + NLM, when both controllers are detuned.

Chapter 9

Conclusions

This study has presented one possible technological solution to common problems encountered by the classical PID controller. The control performance of the proposed controller was compared to popular widely present PID controller schemes.

A fair account of the antiwindup compensation schemes available in the literature and which are used in practice has been presented. A scheme in which the windup compensator feedback acts on the integrator input when the actuator saturates was used in the study. This scheme was combined with the proposed controller structures to yield a better degree of control the closed-loop process.

First, a nonlinear PID controller with gain scheduling, windup protection and Smith predictor compensation was proposed to reduce degradation in the control performance of processes experiencing the negative effects of long dead-times and integrator windup. The performance of the proposed controller with variable nonlinear proportional, integral and derivative gains, subject to specific nonlinear laws, was compared to that of the classical model having static gains. From the outcome of the tests, it was found that in most situations the proposed controller yielded better control with respect to setpoint tracking and disturbance rejection. An important shortcoming of the aforementioned controller is the fact that no specific technique could be used to optimise its tuning within the control loop.

Next, a controller scheme consisting of a nonlinear PD modifier was proposed to function in parallel with the standard PID controller. The proposed PID + NLM scheme, with antiwindup and Smith predictor compensation, was used to reduce the occurrence of control performance degradation in loops having significantly large dead-times and saturation nonlinearities. The outcome of the tests indicated that the proposed new control structure yields a better degree of control for loops encountering any undesirable effects of saturation nonlinearity and dead-time.

When compared to conventional PID control, the proposed PID + NLM scheme is less sensitive to any possible parameter variations within the closed-loop, and is less demanding with respect to the controller parameter settings. System robust stability is maintained by the proposed 'hybrid' control scheme for a wide range of parameter variations. Two important features of the proposed controller are that

- Optimization procedures were used to tune the linear portion of the controller, with the modifier requiring simple adjustment of a single gain to optimise control of the loop.
- The NLM + PID controller is very robust to a wide range of improper tuning of the linear PID controller and has a strong stabilising effect in a wide range of parameters.

The results obtained in this study are conclusive to indicate that the analogue implementation of the PID + NLM combination, with Smith predictor control and antiwindup can be used to acheive optimal control of loops experiencing instability as a result of long dead-time and actuator saturation.

Even though the responses of the above-mentioned analogue controller have been favourable, digital controllers are generally a better option for most control applications for the following reasons:

- Complicated algorithms are difficult to implement using analogue devices such as operational amplifiers, especially if the number of active devices (for example opamps) has to be limited for reliability purposes. These difficulties vanish when the same laws are implemented digitally, using software.

- Analogue controllers are inflexible. Changes to the control law governing the behaviour of the analogue controller may require modifications of the controllers hardware. Digital controllers on the other hand are flexible since any changes in the control algorithm are implemented quite easily by revising the software controlling the process.

The initial results obtained in this study are promising and it is hoped that further investigation will be conducted to exploit the possiblity for direct digital implementation of this controller.

Chapter 10

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