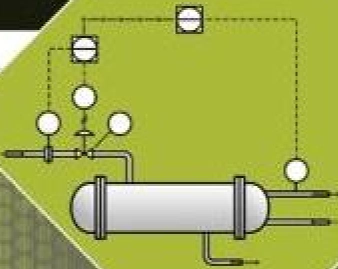
Basic and Advanced Regulatory Control: System Design and Application

2nd Edition

by Harold L. Wade



Setting the Standard for Automation™

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ISA–The Instrumentation, Systems,
and Automation Society

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DEDICATION

*To my mother, who provided an impetus for life-long learning,
and to Mary, who has provided love, support,
encouragement, and criticism when needed.*

ABOUT THE AUTHOR

Harold L. Wade is president of Wade Associates, Inc., a Houston, Texas, firm specializing in process control systems consulting and training. He has 45 years' experience in applying and installing process control systems in such industries as petroleum refining, chemical production, textiles, and water and waste treatment systems among others. He has held technical positions with Honeywell, Foxboro, and Biles & Associates.

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Dr. Wade has taught courses in process control systems design for ISA since 1986. He has presented process control and controller tuning seminars for many companies worldwide. He is also the developer of the process control training program, *PC-ControLAB*. Dr. Wade was a 2002 inductee into *Control* magazine's Process Automation Hall of Fame.

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PREFACE

This book presents a practical approach to process control for the chemical, refining, pulp and paper, utilities, and similar industries. It is the result of seminars in process control that I have presented both in the United States and abroad. A typical participant in my seminars is an engineer, currently employed by a processing company, who may have had formal training in an undergraduate process control course but who may not be able to fully relate the material from that course to his or her work experiences. This book aims to meet this need by explaining concepts in a practical way with only a minimal amount of theoretical background.

The book serves both the beginning and the experienced control systems engineer. For the beginning engineer, it initially presents very simple concepts. For the experienced engineer, it develops these initial concepts so as to provide deeper understanding or new insights into familiar concepts. The purpose is to provide everyone, beginner or experienced engineer, with something they can put to beneficial use in their plant.

This edition also develops a unique method of controller tuning and a novel form of decoupling control, both of which were only introduced briefly in the first edition. The impact on control strategy configuration of advances in the standardization of fieldbus communication systems for process control is discussed. The coverage of model predictive control has been expanded to reflect the wider acceptance of this technology, the development of more efficient systems, and falling prices for the supporting hardware platform. This edition also includes a new set of process control strategy application examples.

Although this is intended to be a practical "how-to" book, readers should not infer that this means it is devoid of mathematical concepts. Where such concepts are utilized, however, it is their *application* to practical situations, rather than the theory behind the concepts, that is emphasized. A theme of the first edition—that wherever I had to choose between providing mathematical rigor or promoting intuitive understanding, I always gave preference to understandability—has been carried forward into the present edition. This practicality distinguishes this book from many academic texts.

The book is organized generally into three parts. The first three chapters present background information, including a brief nonrigorous mathematical review, a discussion of symbols and terminology, and a description of general characteristics of processes and of selected types of control loops.

The second part—chapters 4 through 7—deals with feedback control. The objective is to provide the reader with a thorough intuitive grasp of feedback control behavior and all its nuances. In the chapter on feedback controller tuning (chapter 6), the discussion on improving as-found tuning (also called "intelligent trial-and-error tuning") has been considerably expanded, and supplemented by the presentation of a tuning flow chart that embodies this technique. This new tuning technique has been proven in practical applications and has been well accepted in training classes where it has been presented. In this same chapter, new material is included on tuning liquid-level control loops. The tuning of these loops, which have a completely different characteristic from most other process control loops, has in general received very little specific attention in the process control literature.

The last portion of the book—chapters 8 through 16—begins by defining the "feedback penalty" that must be paid if feedback control alone is used. This leads into a discussion of advanced regulatory control techniques (chapter 9), including chapters on cascade (chapter 10), ratio (chapter 11), feedforward (chapter 12), override (chapter 13), decoupling (chapter 13), model-based (chapter 14), and model predictive control (chapter 15). The chapter on feedforward control offers expanded coverage on the application of multiplicative feedforward control. The chapter on override (selector) control includes additional application examples for this technique, as well as an assessment of the performance of several alternative techniques. The chapter on the control of multiple-input, multiple-output (MIMO) processes (chapter 15) contains additional coverage of inverted decoupling. This MIMO technique was introduced in the first edition; new material previously available only in technical journals is presented here.

The chapter on model-based control in the first edition has been split into two chapters. Chapter 14, devoted primarily to dead-time compensation, covers Smith predictor control, internal model control, and Dahlin's algorithm. The other chapter, chapter 15, contains very significantly expanded coverage of model predictive control.

The concluding chapter, which is almost entirely new, covers process control application topics that do not readily fit into any of the other chapters. In addition to cross-limiting control for fired heaters, which was covered in the first edition, these new topics include floating control, techniques for increasing valve rangeability, and time proportioning control.

One of the themes of this book is to emphasize control strategies that are platform independent. However, since the appearance of the first edition, FOUNDATION™ Fieldbus (FF), which permits the control strategy to be distributed directly into field devices, has grown in acceptance. The network architecture, communication, and implementation aspects of FF are briefly summarized in chapter 5. In this edition, the process control aspects of FF receive greater coverage. Moreover, the chapters on modifications to feedback control, cascade, ratio, feedforward, and override (chapters 5, 9, 10, 11, and 12) all conclude with an example in which that chapter's strategy is implemented using FF function blocks.

I would like to express gratitude to the many students who, by asking probing questions, have enabled me to revise and sharpen my presentation and come up with more meaningful examples. In particular, I would like to thank the engineers at BASF–Freeport for encouraging me to develop the controller tuning flow chart, to the staff of the ISA Training Institute for their support during my seminars, and to Adrian and Ivan Susanto in Indonesia and Michael Wang in Taiwan for sponsoring courses and providing me with an opportunity for travel abroad.

I would also like to express my thanks to Dr. R. Russell Rhinehart for many helpful comments and suggestions, to my longtime friend and mentor (and reviewer of this book) Greg Shinskey, as well as to John Shaw, Jonas Berge, and Bryan Griffen who have reviewed all or parts of this book. And I have special thanks for Susan Colwell, who, through humor and patience, has helped me endure the arduous task of writing.

1 INTRODUCTION

The term *process control* implies that there is a *process* for which there is a desired behavior and that there is some *controlling function* that acts to elicit that desired behavior. This broad concept can embrace everything from societal processes governed by some regulatory control authority to automated manufacturing processes. In practically all cases, however, a common thread is that some measure of the actual process behavior is compared with the desired process behavior. This feedback action then generates a control policy that acts to minimize or eliminate the deviation between desired and actual behavior.

We are concerned in this book with a particular segment of automated process control—that which is applied to chemical, refining, pulp and paper, power generation, and similar types of processes. Even within this limited scope of applications, we will limit the discussion primarily to processes that are operated continuously for long periods of time and within a narrow region of the operating variables. In other words, we exclude such important operating modes as batch processing, start-ups, and grade changes. Many of the control techniques to be presented here, however, can be adapted to these other modes of operation.

For the processes we focus on in this book, the process's behavior is often characterized by measured values of such process variables as temperatures, flow rates, pressures, and the like. The desired behavior, then, is stated to be the set points of those process variables. Until fairly recent times, most applications of industrial process control used simple feedback controllers that regulated the flows, temperatures, and pressures. These controllers required a form of adjustment called *tuning* to match their controlling action to the unique requirements of individual processes. Occasionally, more advanced forms of control, such as ratio and cascade, could be found; even more rarely one might find a feedforward control loop. As long as most of the control systems were implemented with analog hardware, applications were limited to simple regulatory control. This was due to the cost of additional components, the additional interconnections more advanced control required, the burden of maintenance, and the vulnerability to failure of many devices in the control loop. With the advent of digital control systems, however, more sophisticated loops became feasible. Advanced regulatory control—which includes the previously mentioned ratio, cascade, and feedforward control as well as additional forms such as constraint (selector) control and decoupling—could readily be implemented simply by configuring software function blocks.

With this additional capability, however, a need developed for a systematic approach toward using it. This is called *control strategy design.* In order to design a technically successful and economically viable control strategy, the control system engineer must be well grounded in the techniques of feedback control as well as the tools of advanced regulatory control. The requisite knowledge includes both how to implement and how to tune. Even before that, however, the control system engineer must be adept at recognizing when to use (and conversely, when not to use) certain control methods as well as in projecting the expected benefits.

Using advanced regulatory control provides many benefits. One of the most important is simply closer control of the process. It will become very clear later in this book that with basic regulatory (i.e., feedback) control, there must be a deviation from set point before control action can occur. We will call this the "feedback penalty." The objective of advanced regulatory control is for the control action to be taken by incurring only a minimal feedback penalty. The reduction in feedback penalty may be stated in a variety of ways, such as a reduction of the maximum deviation from set point, as a reduction of the standard deviation, or simply as a reduction in the amount of off-spec product produced. This reduction in feedback penalty can provide several forms of economic benefit, such as improvement in product quality, energy savings, increased throughput, or longer equipment life.

Process control is but one part of an overall control hierarchy. It extends downward to safety controls and other directly connected process devices and upward to encompass optimization and even higher levels of business management, such as scheduling, inventory, and asset management (see Figure 1-1). Indeed, corporate profitability may be enhanced more significantly as a result of these higher-level activities than from improved process control *per se*. However, since each layer of the hierarchy depends upon the proper functioning of the layers beneath it, one of the primary benefits of advanced regulatory control is that it enables the higher levels, such as optimization and enterprise management and control.

❖ SYMBOLS

Chapter 2 discusses the graphical symbols used in control system documentation. Listed below are the mathematical symbols that are used generally throughout the book. Some symbols used in this book are used only for the discussion of a particular topic; these symbols are therefore defined in that discussion and are not listed here. Chapter 15 uses a unique set of symbols that are defined at the beginning of that chapter. The following are the symbols found throughout this book:

b	bias value (manual reset) on proportional-only controller output
e	error (deviation between set point and process variable)
E	when capitalized, refers to (Laplace) transform of error
K	steady-state gain of first-order lag
K_C	controller gain (noninteractive and interactive control algorithms)
K_D	derivative gain (independent gains control algorithm)
K_I	integral gain (independent gains control algorithm)
K_P	proportional gain (independent gains control algorithm)
K_p	process gain (change in process variable / change in controller output)
m	manipulated variable, controller output

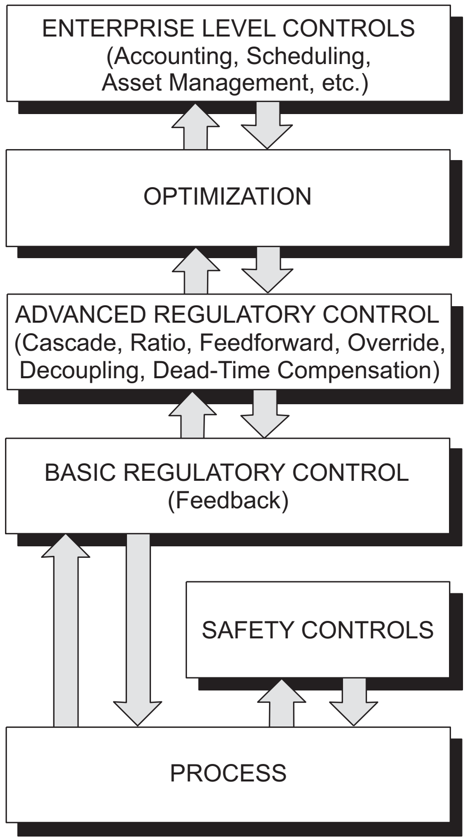

Figure 1-1. Overall Process Control and Information System Hierarchy

M	when capitalized, refers to (Laplace) transform of manipulated variable
PB	proportional band
PI	control algorithm with proportional and integral modes
PID	control algorithm with proportional, integral, and derivative modes
PV	process variable (see also symbol x)
SP	set point (see also symbol x_{SP})
T_D	derivative time (noninteractive and interactive control algorithms)
T_I	integral time (minutes/repeat) (noninteractive and interactive control algorithms)
x	process variable (see also symbol PV)
x_{SP}	set point (see also symbol SP)
u	disturbance variable
α	derivative gain when a derivative filter is used with noninteractive or interactive control algorithm)
θ	dead time
τ	first-order lag-time constant

2

MATHEMATICAL BACKGROUND, DIAGRAMS, AND TERMINOLOGY

Ask any engineer who is more than one year (one month?) out of college if he or she ever uses calculus on the job, and the answer will probably be "Never!" Ask that same engineer if he or she could still work a calculus (or differential equations) problem, and the response will likely be a horrified stare, followed by "Are you kidding?"

It is all too true that many engineers regard their math courses only as a necessary evil required to obtain their degree, but never to be used thereafter. "After all, I work in the real world!" they say. This negative notion of mathematics probably stems from their memory of late nights spent working tedious homework problems as well as their (or their professors') failure to associate the concepts of mathematics to the real world. It also may relate to the fact that these engineering students' subsequent success on the job probably does not depend upon their ability to produce an analytical solution to a calculus problem or a differential equation.

Outside the realm of the professional mathematician, the need to actually be able to solve a differential equation, on paper, has been all but eliminated by the availability of computer-generated solutions. Yet, in many engineering disciplines, the ability to conceptualize a problem in mathematical terms is still an invaluable asset. That ability is what distinguishes the engineer from the technician.

Nowhere is this truer than in the field of process control. Here we are concerned with dynamic phenomena, processes undergoing a transient change, control loops that are oscillating, and the like. Even the very name of the workhorse PID controller contains the words *integral* and *derivative*, two terms readily associated with everyone's freshman calculus class.

This book is not intended to be used as a college-level textbook. If it were, readers could be safely assumed to be "fresh up" on the topics this chapter covers. Rather, this book is intended to be used by the practicing engineer or the real-time computer programmer who is one or more years out of college and in a job environment that requires him or her to understand and improve real process control systems. Such a person may be a bit "rusty" in the mathematical concepts needed to understand the process control principles presented in the remainder of the book. In this chapter we will therefore focus on strengthening the reader's ability to understand and apply the concepts in real situations, not simply to work problems. Wherever confronted by the choice between advancing the reader's intuitive understanding and providing mathematical rigor, we will always favor intuitive understanding.

Let us begin by comparing how a mathematician and a process control engineer might view certain key mathematical topics. (For a more complete review of mathematics for the process control engineer, see *Basic Math for Process Control* by Bob Connell (Ref. 2-1).

❖ MATHEMATICAL FOUNDATION

♦ Functions of a Variable

The mathematician will often speak of the "functions of an independent variable." The functional relationship might be expressed in the form of either an equation or a graphical representation. The control engineer is also interested in the functions of an independent variable, but for him or her the specific independent variable is "time," and the functions are often called signals. The graphical relationship is presented on a strip chart recorder or a trend display at the process operator's console. Typical signals of interest are the process variable, the controller output, and the error signal.

♦ Derivatives

The mathematician is also often interested in the derivative of the function. The control engineer is as well, although the term *rate of change* may be the term he or she uses. Walk into the control room and ask the operator, "Hey, what is the derivative of the temperature?" The response will likely be "Huh?" (or worse!).

Instead, ask "How fast is the temperature changing right now?" The operator will examine the strip chart or trend display and respond with something like, "About two degrees every five minutes." Without being aware of it, the operator is using a concept from calculus: the derivative.

The mathematician uses analytical forms for expressing the derivative of certain functions. The following table presents a list of these functions and their derivatives:

$y = k$	$\dfrac{dy}{dx} = 0$
$y = kx$	$\dfrac{dy}{dx} = k$
$y = x^n$	$\dfrac{dy}{dx} = n\,x^{n-1}$
$y = k\,e^{ax}$	$\dfrac{dy}{dx} = a\,k\,e^{ax}$

The control engineer may have a component ("black box") inside a controller or a line of code inside a computer program that determines the derivative of an incoming signal. Suppose you have such a black box. Suppose also that the output of the black box is connected to a meter, specifically a center-zero, bi-deflectional meter as shown in Figure 2-1. To test your intuitive understanding of the idea of a derivative, suppose that you know the form of the input signal but cannot see the meter response. Cover up the right-hand side of Figure 2-1, observe the input responses shown on the left-hand side, and predict what the meter response should be. Then check yourself against the response shown on the right-hand side.

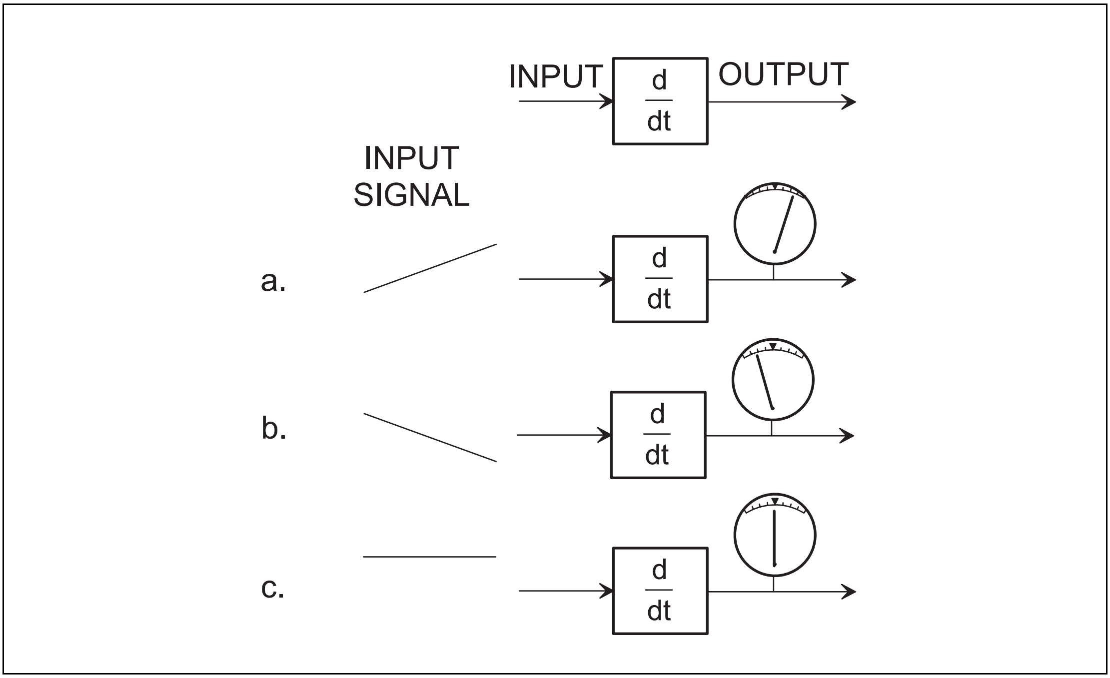

Figure 2-1. Practical Concepts of Differential Calculus

Remember, the output signal from the derivative unit "black box" will be zero if and only if the input signal is stationary. The actual value of the input signal does not matter. If the input signal is increasing, the output signal will be positive; if the input signal is decreasing, the output signal will be negative.

If you are thoroughly comfortable with these intuitive concepts, then you are as familiar with derivatives as you need to be for this book.

♦ Integrals

The mathematician is often interested in the integral of a function. Quite simply, if the function is represented in graphical form, the integral is simply the "area under the curve." The control engineer is also interested in the integral. For example, if the function in question is the error signal of a control loop, the control engineer might be interested in the area under the curve since the loop was last switched from manual to automatic or since the last load upset.

The mathematician uses analytical forms to express the integral of certain functions. The following table gives a list of these functions and their integrals:

$y(x) = 0$	$\int y(x)dx = k$
$y(x) = k$	$\int y(x)dx = kx$
$y = x^n$	$\int y(x)dx = \frac{1}{n+1} k\,x^{n+1}$
$y = k\,e^{ax}$	$\int y(x)dx = \frac{k}{a} e^{ax}$

As with derivatives, the control engineer may have a component ("black box") inside a controller or a line of code inside a computer program that will determine the integral of an incoming signal. As before, suppose you have such a black box. Suppose as well that the output of the black box is connected to a meter, specifically a center-zero, bi-deflectional meter as shown in Figure 2-2.

To test your intuitive understanding of integrals, suppose you can see the output meter reading but you do not know the form of the input signal. Cover up the left-hand side of Figure 2-2, observe the meter responses described on the right, and predict what the input signal must be. Then check yourself against the response shown on the left-hand side.

Remember, the output signal from the integral unit "black box" will be stationary if and only if the input signal is zero. The actual value of the output signal does not matter. If the input signal is positive, the output signal will be increasing; if the input signal is negative, the output signal will be decreasing. More specifically, if the input signal is both positive (negative) and constant, then the output signal will be increasing (decreasing) at a uniform rate.

If the only thing we know about the output signal is that it is stationary at the moment, then all we can conclude about the input is that it is zero at the moment. If we know that the output signal is currently both stationary and positive, as shown in Figure 2-2c, then we can conclude that, although the input is currently stationary, in the past it must have been positive more of the time, or to a greater magnitude, than it was negative.

If you are thoroughly comfortable with these intuitive concepts, then you know as much about integrals as you'll need to understand this book.

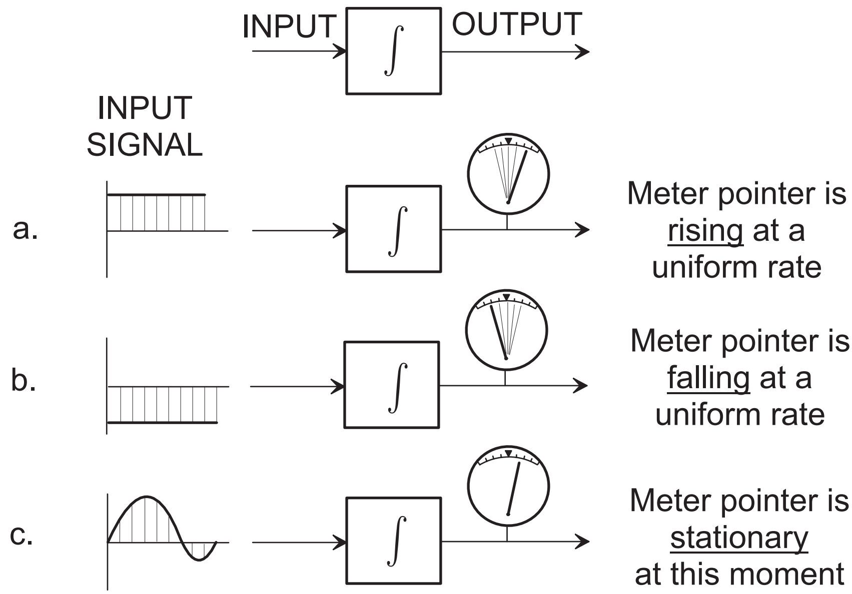

Figure 2-2. Practical Concepts of Integral Calculus

♦ Differential Equations

A differential equation expresses how a variable changes as a function of other variables, and perhaps as a function of itself as well. The study of differential equations and the means for obtaining solutions are both stocks in trade for the mathematician. The control engineer, however, is rarely if ever called upon to actually solve a differential equation. However, he or she often finds it very beneficial to know how to set up a differential equation that describes a system. But merely describing the system's dynamic behavior with a differential equation is not sufficient. The engineer needs to be able to predict what the dynamic response will be for certain forms of input. For this purpose, control engineers may use process dynamic simulation systems. In other cases, the control engineer can transform the differential equation into a simpler form that provides much insight into the behavior of the system. Let's illustrate this with an example.

We will use a simple process example of a dynamic system. A control engineer observes that the liquid level in a vessel changes more slowly as the difference between the inflow and outflow rate decreases. The dynamic behavior of such a process can be described by a simple differential equation. Likewise, the dynamic behavior of more complex processes can be described by more complex differential equations.

Suppose the tank shown in Figure 2-3 is initially in equilibrium (that is, the input and output flow rates are equal, consequently the level is stationary), and subsequently there is a step-change input flow rate. What will be the time response of the level?

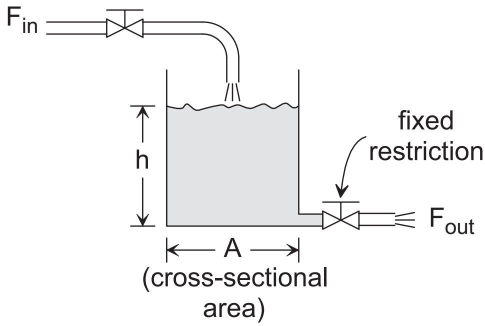

Figure 2-3. A Simple Dynamic Process

The answer depends upon what the output flow rate does. If the output flow rate is maintained constant, say by a flow controller and valve, then the level will continue to rise until the vessel overflows. On the other hand, if there is a fixed restriction on the output, the output flow rate will be a function of the hydrostatic head at the bottom of the vessel, hence a function of the level in the vessel itself. On a step change in inflow, the level will rise rapidly at first. This will increase the hydrostatic head and consequently increase the outflow. As the outflow increases, the level rises slower and slower until the tank is again in equilibrium with the input and output flow rates equal and the level stationary (assuming that the vessel does not overflow first).

Most control engineers will recognize the form of the response shown in Figure 2-4. Actually, this response could be determined by solving a differential equation. This author believes, however, that it is more important to have a good intuitive understanding of the physical response than to be able to predict the solution by solving the differential equation. (Of course, if you can do both, that is even better!)

Let us not leave the subject of differential equations too rapidly, however. For the vessel shown in Figure 2-3, the following differential Equation 2-1 describes the behavior of the level, *h*, in terms of volumetric input and output flows and the cross-sectional area:

$$A\frac{dh}{dt} = f_{in} - f_{out} \tag{2-1}$$

Now, suppose the outflow rate increases as the level increases. If we can make the simplified assumption that the outflow is proportional to the level[1]; that is,

1. Those familiar with hydraulic principles will recognize that a more accurate relationship for turbulent flow is that the outflow is proportional to the square root of the level, or:

$$f_{out} = b\sqrt{h}$$.

For the purpose of this discussion, such rigor would overly complicate the problem and obscure the concepts being presented.

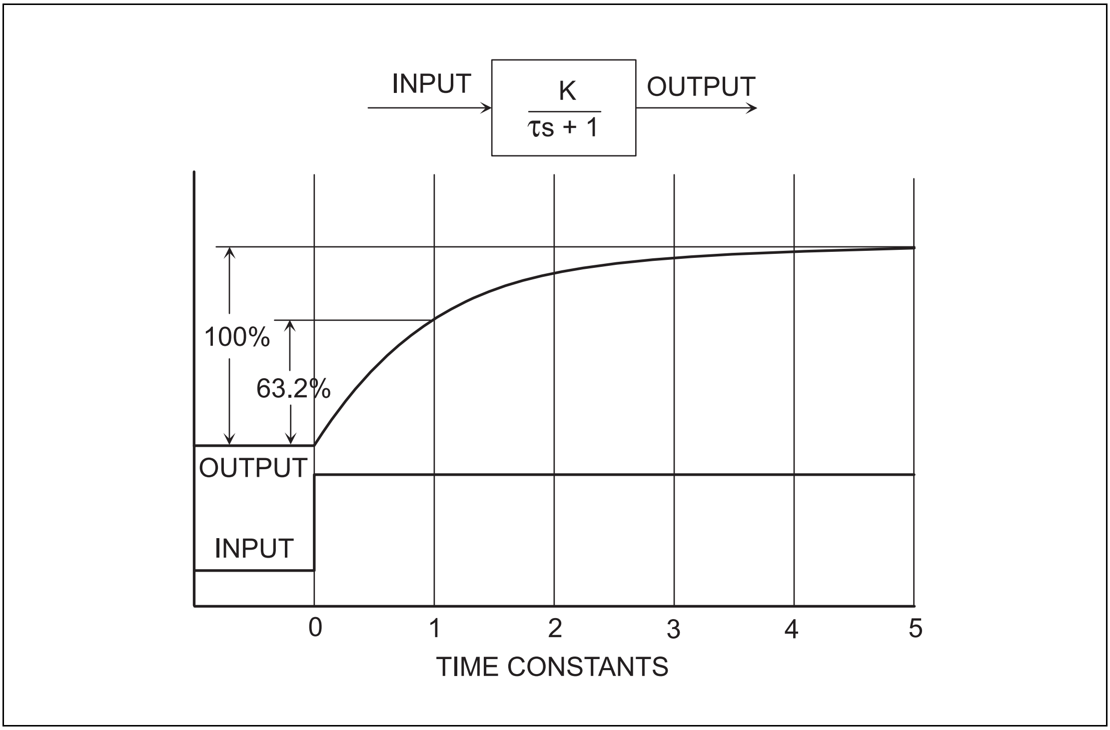

Figure 2-4. Response of Simple Dynamic Process to Step Change in Input

$$f_{out} = bh \tag{2-2}$$

then we can relate the rate of change of level to the inflow rate and the level itself. Thus, we have the following descriptive differential equation:

$$\frac{A}{b}\frac{dh}{dt} = \frac{1}{b}f_{in} - h \tag{2-3}$$

This equation can be solved using fairly elementary mathematical techniques, but doing so would not reinforce the concepts being presented here. Instead, in the following section we are going to take another approach: transforming the differential equation into a simpler algebraic equation. This will allow us to solve the equation by algebra. It will also provide us with a powerful tool for describing the dynamic behavior of this process (the tank) and many similar processes.

♦ Transfer Functions

Recent graduates of several engineering disciplines will have studied transfer functions (more specifically, the one called "Laplace transforms") in one or more of their college courses. These explorations probably entailed some degree of mathematical rigor (remember the initial value theorem, final value theorem, etc.?), so much so that the student often lost sight of the forest for the trees. That is, he or she failed to become sufficiently comfortable with transfer functions to use them as a way to think about real problems encountered on the job.

One of the first persons to use transfer functions as an aid to solving real problems was Oliver Heaviside, an English electrical engineer who practiced over a century ago. At that time, electrical engineers had pretty well mastered Ohm's law, but new types of problems involving circuit transients were emerging. For the most part, electrical engineers lacked the appropriate mathematical background to solve these types of problems. Oliver Heaviside developed a technique that enabled him to convert the differential equations that described the circuit transient response into an algebraic equation he could solve. For some time, the technique was known as "Heaviside transforms" until it was realized that what Heaviside was doing had already been done by the French mathematician Pierre Laplace many years before. Heaviside's fellow mathematicians and engineers were quite critical of him for using a mathematical technique when he could not say how or why it worked. Heaviside's response was, "I do not know how my digestive tract works, either, but I still enjoy eating a good meal."

Our approach will be similar to that of Heaviside. We will skip the mathematical rigor and immediately put transfer functions to work.

It is instructive to see how a transfer function is derived from a differential equation. Provided that a differential equation meets certain conditions, it can be transformed into an algebraic equation by performing the following three steps[2]:

(1) Replace a derivative symbol, $\frac{d}{dt}$, with the symbol s ;

(2) Replace an integral symbol, $\int \cdots dt$, with the symbol $\frac{1}{s}$.

Replace the symbols representing time-dependent variables, which should be written in lower-case letters in the differential equation, with their corresponding upper-case letter in the transformed equation.[3]

For readers who have never encountered transfer functions before, the reason for step 3 is undoubtedly obscure. We will deliberately leave it that way since a deeper understanding of the mathematical theory is not necessary for the use we will make of transfer functions.

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2. These conditions are linearity, piecewise regularity, and exponential order. The physical processes we deal with in this book meet the second and third of these requirements. We will use the following simple test for linearity: "Does the process respond the same way at any operating point?" Most real processes do not meet this criteria over a wide operating range. However, they do roughly meet the criteria within a reasonable range of a chosen operating point.
3. To be technically correct, the variables (signals) in the differential equation are functions of time. They therefore should be written in the form $x(t)$. According to mathematical convention, lower-case symbols (e.g., x) are used in the differential equation. In the transformed equation, the variables are transformed into functions of s. To note this transformation, $x(t)$ should be replaced by $X(s)$. By mathematical convention, upper-case symbols (e.g., X) are used in the transformed equation.

When these three steps are completed, we can solve the resulting algebraic equation for the dependent variable, H, in terms of the independent variable, F_{in}. An example will clarify this. The differential Equation 2-3 describing the tank level contains only a derivative symbol; there is no integral symbol. Therefore, it can be transformed by using only steps 1 and 3. When so transformed, it becomes:

$$\frac{A}{b}sH = \frac{1}{b}F_{in} - H \qquad (2\text{-}4)$$

Then, solving for level in terms of input flow:

$$H = \left[\frac{\frac{1}{b}}{\frac{A}{b}s + 1}\right]F_{in} \qquad (2\text{-}5)$$

The term within the brackets in Equation 2-5 is the *transfer function*. It tells how the dependent variable, H, responds dynamically (that is, its transient response) to a change in the independent variable, F_{in}. There are many forms of transfer functions. The form shown in Equation 2-5 is called a *first-order lag*. The physical significance of a first-order lag is that it usually represents the action at a place of mass or energy storage.

The general form of transfer function for a first-order lag is as follows:

$$\frac{K}{\tau s + 1} \qquad (2\text{-}6)$$

The reader should recognize three elements of a first-order lag transfer function:

First, there is the format itself:

$$\frac{\square}{\square s + 1}.$$

This format indicates that if the input (independent variable) changes in a step, then the output response (dependent variable) is as shown in Figure 2-4. The small squares represent placeholders for parameter values that describe the essential characteristics of the response.

Second, there is the ratio of the (eventual) steady-state change in output to the amount of input change. This is given by the parameter K; it is called the *steady-state gain*.

Finally, there is a parameter called the *time constant* and designated by the Greek letter τ (tau). This is an indication of how quickly the output comes to an equilibrium following a step change in the input. (Note that we said "an indication of," not "the time it takes to come to equilibrium." We shall see the significance of this distinction later.) The time constant can be stated in any convenient time units, such as seconds, minutes, and the like. For industrial process control work, it is usually most convenient to express τ in minutes.

Theoretically, the output will never come to equilibrium. Mathematicians have provided us with a convenient metric for comparing the speeds of response of various dynamic systems: the time required for the output to achieve 63.2 percent of equilibrium. This time is called the "time constant." Thus, τ represents the time it takes for the output to make 63.2 percent of its eventual change.

If we know that a dynamic system or process can be described by a first-order lag transfer function (Equation 2-6), then we immediately know the shape of the response to a step input change. If we know the steady-state gain, *K*, we know how much the output will change for a given change in input, and if we know the time constant, we know how fast the output will change.

Sometimes one hears that a dynamic system will come to equilibrium in five time constants. Actually, the system will reach 63.2 percent of equilibrium in one time constant, 63.2 percent of the remaining amount in one more time constant, and so on. From this, we can calculate the values in the following table:

Time since Step Input Change	Percentage of Steady-State Change
1 Time Constant	63.2
2 Time Constants	86.5
3 Time Constants	95.0
4 Time Constants	98.2
5 Time Constants	99.6

In five time constants the system (or process) is within ½ percent of equilibrium; the change thereafter will be practically imperceptible. If you already know the time constant, and being within ½ percent of equilibrium is sufficiently close for you to say "we're in equilibrium," then the statement that the system will reach equilibrium in five time constants is valid. What is *not* valid is to use a graph or chart of the change, make a guess at the time required to reach equilibrium, and then divide by five. This is because a great discrepancy may exist between several estimates of the point of equilibrium.

You can quickly approximate the time constant by determining how much time the system or process needs to change by two-thirds of its final amount in response to a step input change. This time will be slightly longer than the true time constant, but it will be accurate enough for most of our purposes. If desired, you can refine the estimate of the time constant by taking 90 percent of the time required for two-thirds of the change.

Another form of process response often encountered in actual applications is a time delay, often called "dead time." Without providing formal justification, we present the transfer function as follows:

$$e^{-\theta s} \tag{2-7}$$

This function describes a system or process that has the input-output response shown in Figure 2-5. The parameter θ represents the amount of dead time. For industrial process control, it is usually most convenient to express the dead time in minutes.

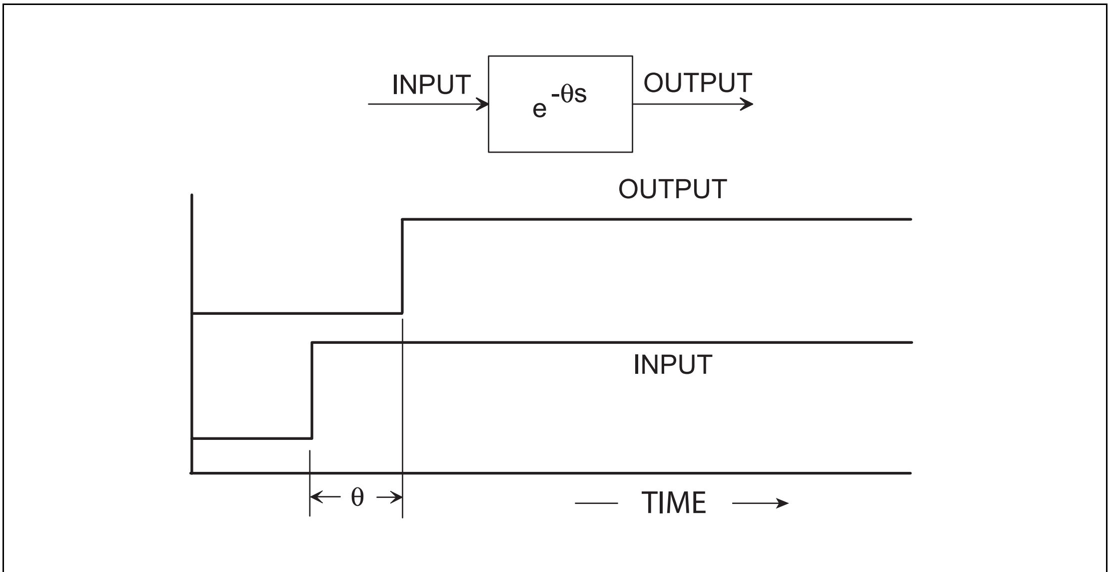


Figure 2-5. Laplace Transform and Graphical Diagram of Dead Time

The significance of dead time is that it usually represents the time for physical movement of mass or energy. Hence, it is often called the "transport lag."

A very common process response to a step input change is often approximated by a combination of first-order lag and dead time. This response, depicted by Figure 2-6, is typical of a process that contains both mass or energy storage as well as a transport time. It is described by the following transfer function:

$$\frac{Ke^{-\theta s}}{\tau s + 1} \tag{2-8}$$

Note that this form of response is always characterized by the following three parameter values:

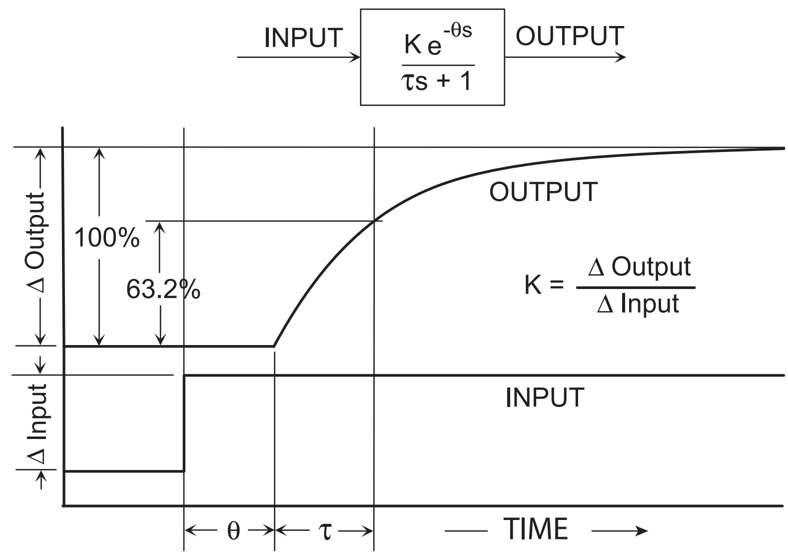


Figure 2-6. Laplace Transform and Graphical Diagram of First-order Lag Plus Dead Time

K	process gain;
τ	time constant;
θ	dead time.

In chapter 6, we discuss ways to determine values for these parameters by using process tests as well as how to use these parameter values to calculate controller tuning parameters. We can note here that an important property of feedback control loops is the ratio of dead time to time constant, θ/τ. This ratio gives a measure of how easy a loop is to control. If this ratio is small (meaning the dead time is much shorter than the time constant), the loop is fairly easy to control. As this ratio increases (say, to values reflecting dead time that is much longer than the time constant), the loop becomes increasingly difficult to control. The application of dead-time compensation techniques, discussed in chapter 14, should be considered for these cases.

♦ Frequency Response

We have described the step input response for two common transfer functions. In reality the transfer functions convey much more information than that. For example, if the input is a sinusoidal wave of a given magnitude and frequency, the output will also be a sinusoidal wave of the same frequency, although the magnitude will probably be changed and there will likely be a phase shift. The transfer function also conveys information about the magnitude ratio and the phase shift. Very elegant techniques—for example, Bode plots, root locus, and Nyquist diagrams—have been developed to utilize frequency-response information to support the analysis and design of control systems. While these techniques can provide great insight into the nature of control systems, they require a process model (transfer function) that is much more precise than is generally available in industrial process control. Hence, we will not discuss these techniques in this book.

❖ DIAGRAMS AND TERMINOLOGY

Every field has a terminology that is unique to it. In technical fields where information is often conveyed in the form of diagrams, there will be special symbols in addition to terminology. Both specialized terminology and symbols are used in the field of instrumentation and process control.

Figure 2-7 shows a pictorial diagram of a small portion of a process. Even a person with only moderate technical literacy might recognize this as a heat exchanger within a temperature control loop that consists of a temperature sensor, controller, and valve. This figure does not convey a lot of technical information, however.

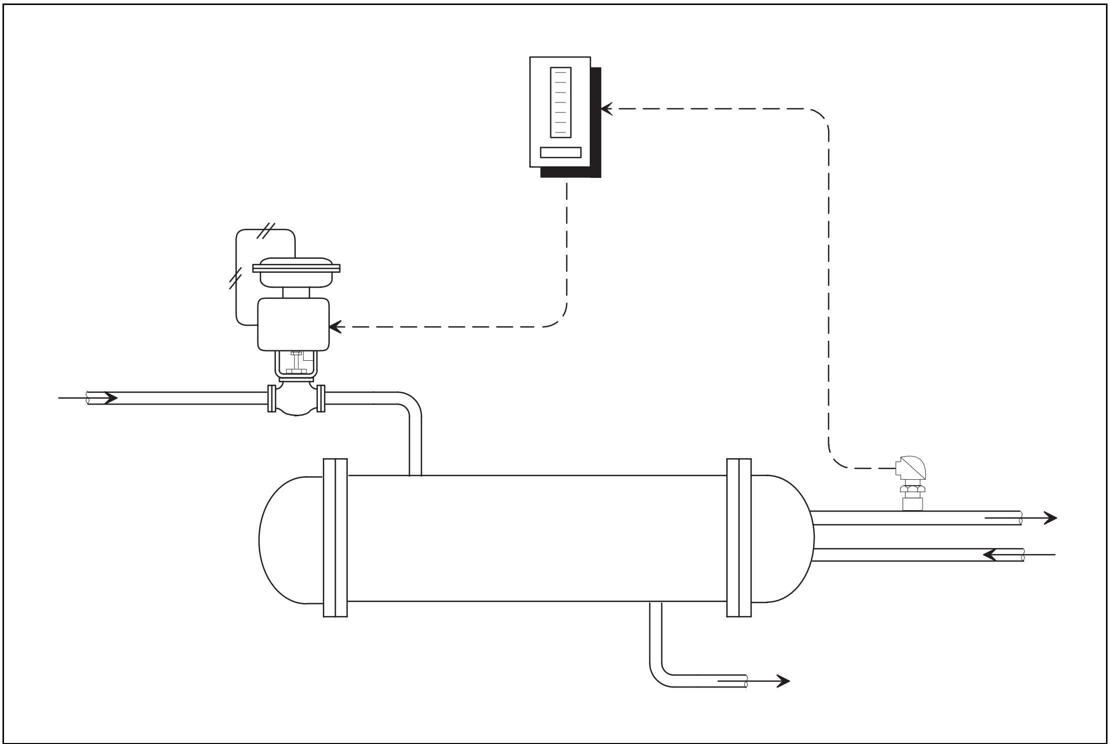


Figure 2-7. Pictorial Representation of a Process Control Loop

Instrumentation, control, and process engineers abstract this pictorial information into some form of iconographic diagram. The most familiar form is called a "Piping and Instrumentation Diagram," usually abbreviated as "P&ID" (not to be confused with PID controllers, to be discussed later). Figure 2-8 is an example of a P&ID. In practice, P&IDs are quite large (say 3 by

5 feet or more) and are quite detailed. Recently, there has been a trend toward machine-drawn P&IDs.

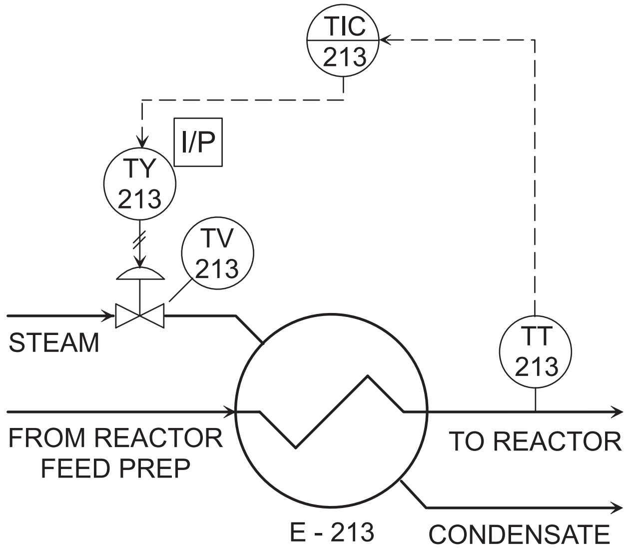


Figure 2-8. P&ID Control Loop Representation

A typical P&ID shows the outline of the process units and the piping that connects them as well as a symbolic representation of the instrumentation and control system. P&IDs are used almost universally in the process industries, although some large corporations have developed their own set of standard symbols for use with them.

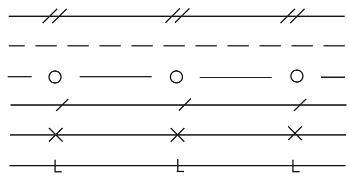
P&IDs typically use symbols to represent various types of instrumentation devices. These symbols, normally called "ISA symbols," are defined by ISA-5.1-1984 (R 1992), *Instrumentation Symbols and Identification* (Ref. 2-2). (A revision to this ISA standard is currently nearing completion.) Figure 2-9 presents a small subset of the symbols and device designations defined in ISA-5.1-1984. See Ref. 2-2, or the latest edition, for a complete set of designations.

Note that the ISA symbols indicate by graphical means where a device is located or accessed (on the control panel, in the field, etc.) as well as the type of signal (pneumatic, electric, etc.) that interconnects devices or functions. They also provide for a device tag that consists of a mnemonic designation for the overall function of a device (e.g., "TRC" for temperature recorder controller) plus additional letters or numbers that provide for unique loop identification. ISA symbols usually do not indicate detailed functionality of a device, such as whether it is a PI or PID controller, or whether or not it has a Manual-Automatic switch.

Another set of instrumentation and control symbols that are in use, particularly in the power industry and commonly referred to as SAMA (Scientific Apparatus Makers Association) sym-

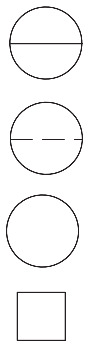
SIGNALS

Pneumatic (Usually 3 - 15 psi)
Electric (Usually 4 - 20 mA)
Internal Software Link
Undefined Signal Type
Capillary Tube
Hydraulic



INSTRUMENTS AND FUNCTIONS

DISCRETE DEVICES



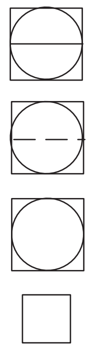
Located on Main Panel Board
Operator accessible

Located behind Main Panel Board
Not operator accessible

Field mounted
Not operator accessible

Mathematical function
(Symbol located upper
left or right of main symbol)

SHARED DISPLAY



Operator accessible
at Main Console

Not normally
accessed by
operator

Field function

Mathematical function
(Symbol located upper
left or right of main symbol)

LETTER IDENTIFICATIONS

LETTER	FIRST LETTER (Measured Variable)	MODIFIER	SUCCEEDING LETTERS
A	ANALYSIS		ALARM
C			CONTROLLER
D	DIFFERENTIAL		
E	VOLTAGE		ELEMENT
F	FLOW	RATIO	
H	HAND		HIGH
I	CURRENT		INDICATOR
K	TIME		CONTROL STATION
L	LEVEL		LOW
P	PRESSURE		
Q	QUANTITY	TOTALIZER	
R			RECORDER
S	SPEED	SAFETY	SWITCH
T	TEMPERATURE		TRANSMITTER
V	VIBRATION		VALVE
X	UNCLASSIFIED		
Y	EVENT/STATE		RELAY, COMPUTE, CONVERT
Z	POSITION		

Figure 2-9. Representative ISA Symbols (from ISA-5.1-1984)

bols, are defined by SAMA Standard PMC 22.1-1981, *Functional Diagramming of Instrument and Control Systems* (Ref. 2-3). While a table of Function Block Designations from PMC 22.1 was included in the 1984 revision of ISA-5.1 as Table 3, many of the primary SAMA symbols are being incorporated into the current ISA-5.1 revision work.

SAMA symbols are rarely used on P&IDs. Rather, they are typically used to diagram control systems at a detailed functional level. They provide no information about the device's location or function (Is it in the control system software or is it a PID controller in a field hardware device?) nor about the technology used to implement it (Is it pneumatic, electronic, or microprocessor-based?). Often, SAMA symbols show the control system without reference to the processing equipment or piping details. Nevertheless, for complex control systems, because they can show greater functionality compared with other documentation formats, they are often preferred for presenting control strategy details in some industries.

Because of the complexity of their control strategies, SAMA symbols are widely used in the power generation industry. Many manufacturers of control equipment, particularly distributed control systems, document their library of control algorithms using something similar to SAMA symbology.

Figure 2-10 shows an elementary control loop that is illustrated using SAMA symbols. Figure 2-11 is an abbreviated list of some of the more commonly used symbols.

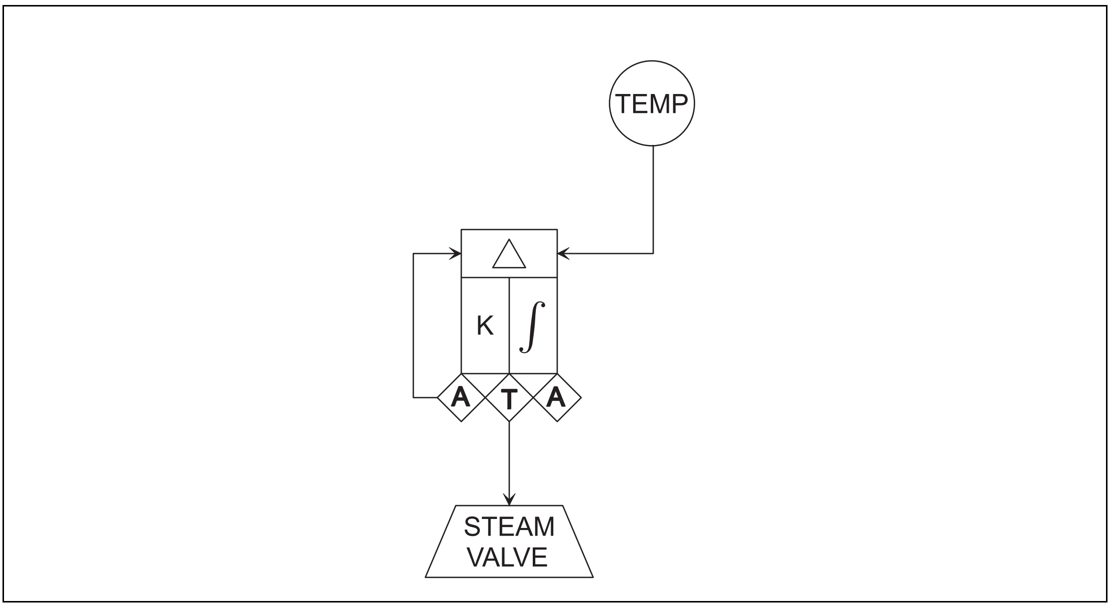


Figure 2-10. Control Loop Representation Using SAMA Symbols (From SAMA Standard PMC 22.1-1981)

This book uses both ISA symbols and SAMA symbols. It sometimes uses a mixture of the two when particular clarity is needed for some functions of a control scheme.

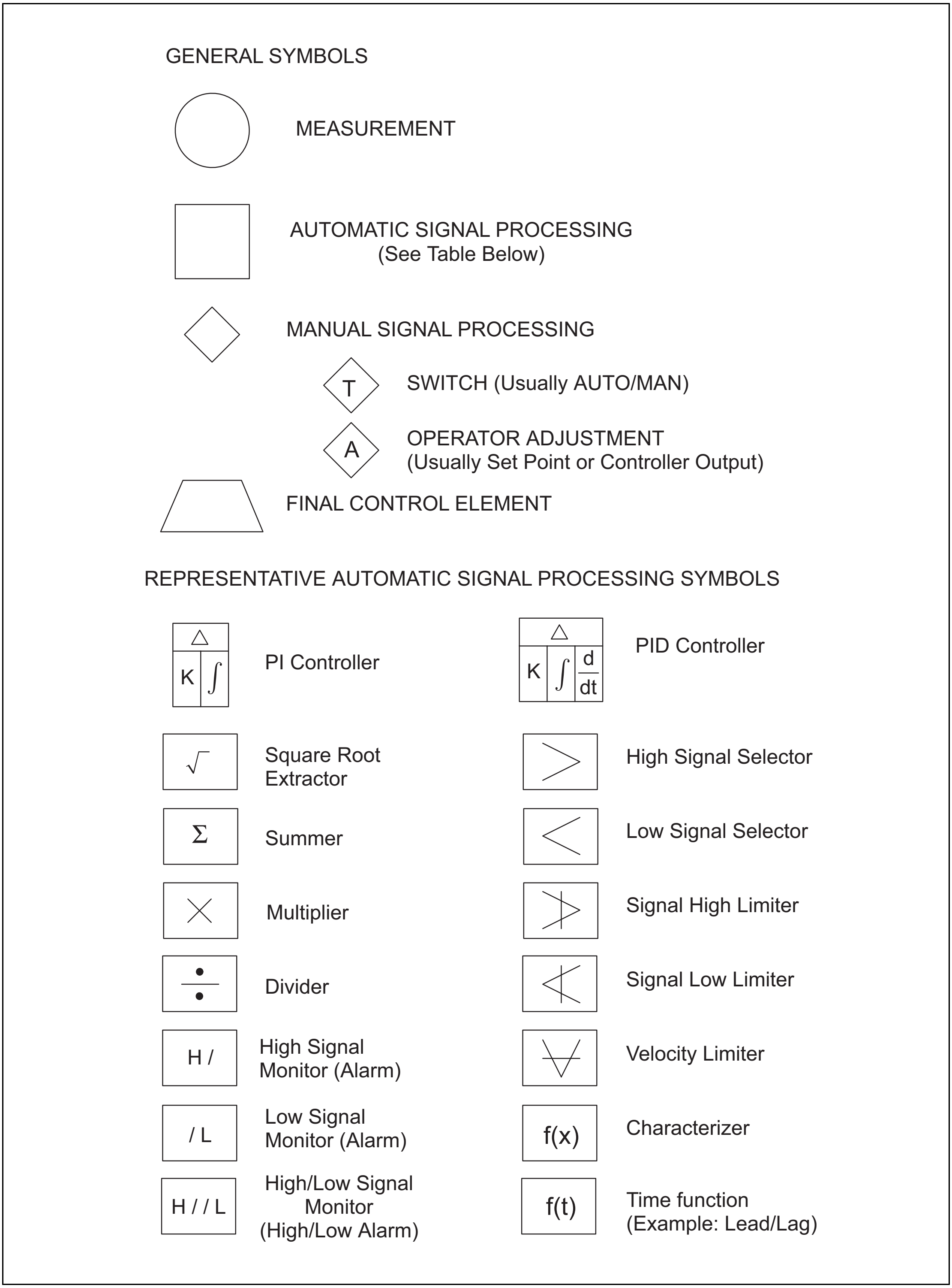


Figure 2-11. Representative SAMA Symbols (From SAMA Standard PMC 22.1-1981)

Other forms of documentation used by instrumentation and control engineers include process flow diagrams (PFDs) and loop diagrams. PFDs depict the process flow streams as well as all processing equipment. They designate the design flow rates, temperature, pressure, and composition for each of the process streams, as well as the size, capacity, and physical dimensions, as appropriate, for each piece of control equipment. PFDs provide only a basic representation of the control scheme, however.

Loop diagrams provide detailed installation information for a particular control loop. This information may include model numbers for hardware devices, calibration ranges, parameters used in calculation devices, cabinet and terminal numbers, and the like. These are invaluable in helping the installer or troubleshooter locate exactly which terminal numbers and which junction box are relevant to a particular loop. Since we are concerned here with the details of control strategy, not installation details, we will not discuss loop diagrams further.

For the purposes of describing and analyzing a control loop, when it isn't necessary to know whether it is implemented with analog or digital hardware, a block diagram is beneficial (see Figure 2-12). Block diagrams clearly show the closed-loop nature of a control system and focus on the information communicated between elements in the control loop.

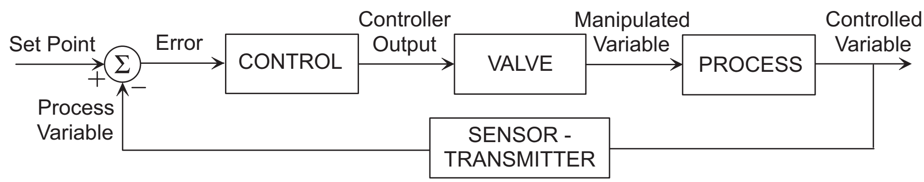


Figure 2-12. Representation of Process Control Loop Block Diagram

The most important element in the control loop is the process itself. The process can be a heat exchanger, a chemical reactor, a distillation tower, a food processing unit, and so on. The objective of a control loop is to regulate an attribute of the process that is variously called the *process variable*, *controlled variable*, or simply the measurement. (Not all process variables are measured; some are inferred or calculated from other variables.) In terms of information flow, we may speak of the process variable as the process output.

Going backward around the loop, we manipulate some physical quantity, usually a flow stream of mass or energy, to regulate the process variable. This is called the *manipulated variable*. The physical device by which we manipulate this stream is called the *final control element*. Quite often this is a valve, and for simplicity we will often use the term *valve* to refer to any type of final control element.

The intelligence within our control loop is contained within a controller. The means by which it takes control action is called the *control law* or *control algorithm*. In this book, we will give quite a bit of attention to various forms of control algorithms. The control algorithm is driven by an error, or the difference between a signal that represents the process variable and our desired value for that variable, which is called the *set point*. Finally, a very important component of the control loop is the actual measuring device for converting a physical attribute of the process (temperature, pressure, etc.) into a usable signal. On Figure 2-12 this is called the sensor and transmitter, although in common usage the device is simply called the *transmitter*.

For extensive definitions of the terms used by instrumentation and control engineers, see *The Automation, Systems, and Instrumentation Dictionary* (Ref. 2-4).

Figure 2-13 shows an even simpler block diagram for a control loop. This diagram segregates equipment outside the control room from the equipment inside. The aggregate of all equipment outside the control room is called the "process." (Some authors refer the aggregate as the "plant," to distinguish this from the process unit itself.) This includes the actual processing device (heat exchanger, etc.), the valve or other final control element, and the measuring/transmission equipment. From the controller's viewpoint, the aggregate of everything outside the control room receives the controller output signal, acts upon it, and returns a measurement signal. Note that in terms of information flow, the controller output is the process input. Likewise, the process output is the controller input.

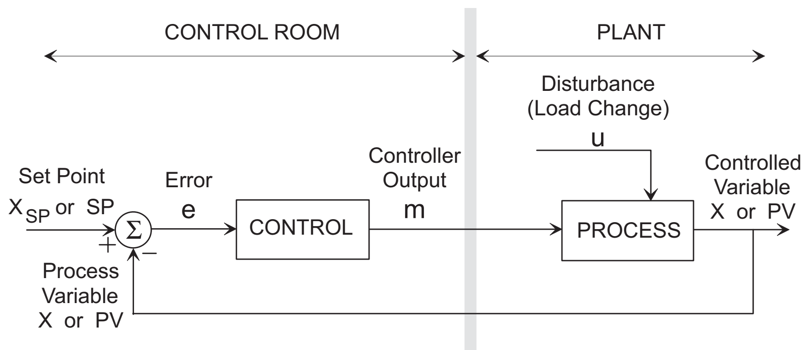


Figure 2-13. Simplified Block Diagram Representation of a Process Control Loop

In addition to designating the symbols for various signals within a control loop, Figure 2-13 also shows an aspect not previously shown: the disturbance or load change on the process. We will use the terms *disturbance*, *load change*, or simply *load* interchangeably to refer to a random phenomenon over which we have no control but which has a direct effect on our process. The phenomenon may be external, such as a feed rate or environmental effect, or it may be internal, such as catalyst decay or equipment malfunction.

Every control loop is subject to disturbances. If this were not so, we could run the process on manual. The purpose of the control system is to counteract the effect of the disturbances. Note that by using the terms *disturbance* and *load* interchangeably, we are departing from the meaning given the term *load* by the power generation industry where one may hear the instruction "set a certain load on that generating unit." In the power generation context, the term *load* is closer to the meaning we intend for the term *set point*. However, other segments of the process industry use the term *load* as we use it here.

❖ DIRECT- OR REVERSE-ACTING?

The block diagrams shown in Figures 2-12 and 2-13 indicated symbolically that the error was computed as being the difference between the set point and the process variable, specifically

$$e = SP - PV. \tag{2-9}$$

However, we could have computed error in the following way:

$$e = PV - SP. \tag{2-10}$$

Either method is correct under some circumstances, but for any given loop only one method is correct.

A controller, whether it is implemented in hardware or software, has an attribute that is either *direct-acting* or *reverse-acting*. These terms refer to the relative direction of movement of the process variable and controller output. If on an increase in the process variable the controller's response is to increase its output, then the controller is said to be direct-acting because the controller output directly follows the measurement. If on an increase in the process variable the controller's response is to decrease its output, the controller is said to be reverse-acting. Thus, computing the error as $SP - PV$ implies that the controller is reverse-acting, whereas $PV - SP$ implies that the controller is direct-acting.

This attribute, direct- or reverse-acting, must be set properly by an instrumentation or control system engineer in order for the control loop to function correctly. The proper setting depends upon the process as well as the failure mode final actuator. Within the loop, there must be an attribute called negative feedback, so named because if a measurement moves away from set point, the control action will be in the direction that returns it to set point. The opposite of negative feedback is obviously positive feedback. As an example of positive feedback, suppose we have a temperature control loop in which the controller adjusts the position of a steam valve. If, on a rise in temperature, the control action is such that the steam valve is opened, this will cause a further rise in temperature, which will cause a further opening of the steam valve, which will cause ... and so on. This is positive feedback; it must be avoided.

Engineers may use two thought processes to determine the proper setting. One way is to consider the control problem by thinking of the process action; the other is to consider the controller action. Either way will arrive at the same results.

To consider the process (plant) itself, as shown in Figure 2-12, suppose there is an increase in signal to the valve. (This could possibly occur when the controller is in manual, or it may be due to the action of a manual regulator installed in lieu of a controller.) Does this cause the signal that represents the process variable to increase or decrease? To answer this question, we must know the failure mode of the valve (fail-closed, often called "air-to-open," or fail-open, often called "air-to-close"). We also need to know the effect of the manipulated variable on the process variable (is it steam or cooling water?). We need to know whether the valve positioner is direct- or reverse-acting; we may also encounter a transmitter whose output signal increases when the physical variable decreases.

If an increasing signal to the process (valve signal) causes the measurement signal to increase, we say that the process is direct-acting. If an increasing signal causes the measurement signal to decrease, the process is said to be reverse-acting. Then, to have negative feedback, the controller must be the opposite of the process. A majority of processes are direct-acting (due to the fail-closed action of the valve); hence a majority of controllers are set reverse-acting. Figures 2-12 and 2-13, which show error computed as *SP – PV*, depict reverse-acting controllers.

The other thought process is to consider the required controller action. We ask ourselves the question, "If the measurement signal increases, do we want the controller output to increase or decrease?" If the desired action of the controller output is to increase, then set the controller direct-acting. Otherwise, set the controller reverse-acting. This thought process requires us to implicitly consider all of the factors mentioned above: Is the valve fail-open or fail-closed? Is the manipulated variable steam or cooling water? And so on. In contrast, the first thought process explicitly considers all of these factors.

Some manufacturers of digital control systems separate the consideration of the controller's direct or reverse action from the failure mode of the valve. The controller output signal, ranging from 0 to 100 percent, always represents the "percent open" of the valve. Therefore, the direct or reverse action of the controller represents the relative direction of the process variable and valve movement, regardless of whether the valve is fail-open or fail-closed. Then, a separate configuration question, which is applicable to the analog output function block, asks whether the signal should be reversed or not. If the signal is not reversed, 0 to 100 percent of the signal from the controller is converted into a 4–20 mA signal to the valve. This would normally be the choice for fail-closed valves. If the signal is reversed, 0 to 100 percent of the signal from the controller is converted into a 20–4 mA, typically for fail-open valves. This application is depicted in Figure 2-14.

❖ REFERENCES

2-1. Bob Connell. *Basic Math for Process Control*. ISA – The Instrumentation, Systems, and Automation Society, 2003.

2-2. ISA-5.1-1984 (R 1992), *Instrumentation Symbols and Identification*. ISA – The Instrumentation, Systems, and Automation Society, 1992.

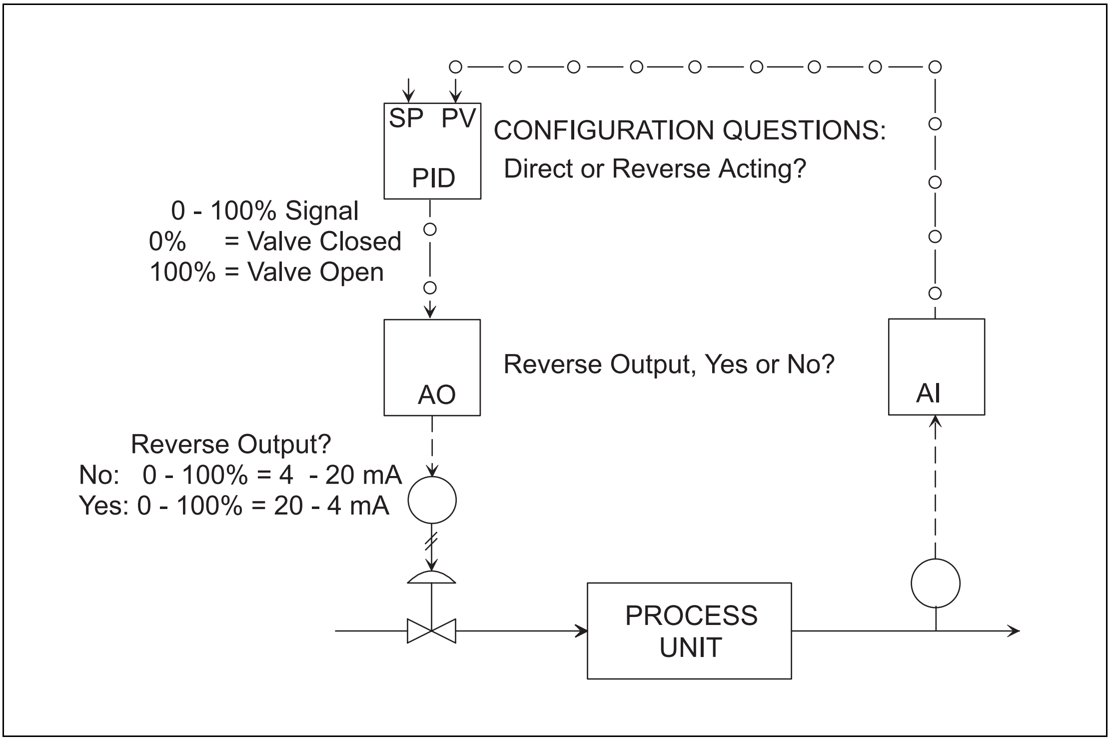


Figure 2-14. Direct- and Reverse-Acting Configuration Used in Some Digital Systems

2-3. SAMA Standard PMC 22.1-1981, *Functional Diagramming of Instrument and Control Systems*. Scientific Apparatus Makers Association, 1981 (MCAA web site: http://www.measure.org).

2-4. *The Automation, Systems, and Instrumentation Dictionary, 4th Edition.* ISA – The Instrumentation, Systems, and Automation Society, 2003.

3

PROCESS AND CONTROL LOOP CHARACTERISTICS

In order to design, analyze, or commission a process control system, one must be familiar with the characteristics of the process itself. Although it is highly beneficial for the control engineer to have a good understanding of the physical and chemical phenomena that govern the process, his or her view of the process will usually differ from that of the process design engineer. The discussion in this chapter is meant to develop the thought processes of a control engineer. Though some of the following points may seem to overstate the case, they will enable us to highlight the differences in the ways the control engineer and process design engineer think:

- The process design engineer is concerned with meeting production rate and quality specifications, which are often called the "design conditions." The control engineer is concerned with operating an existing process at other than design conditions, often with reduced throughput, variations in feedstock, or other abnormal conditions.

- The process design engineer's objective is often to minimize the initial cost (or the life-cycle cost) of the processing equipment. The control engineer's objective is to make the most efficient use of the equipment that is already installed.

- The process engineer considers those design parameters that can be specified as independent variables. Other parameter values that are derived from these are dependent variables. For example, the pressure of a saturated steam system might be an independent variable that can be specified during the design process; the temperature then becomes a dependent variable. The control engineer considers as independent variables the control points (for instance, valves or flow rates) that can be manipulated to affect the process. The steam pressure then becomes a dependent variable that results from those valve positions or flow rates.

- The process engineer is usually concerned with the steady-state conditions of the process. The control engineer must necessarily take into consideration both the steady-state and the dynamic, or transient, behavior of the process.

The characteristics of each process will be different. Even so, from the process control engineer's viewpoint, certain characteristics are similar from process to process. It is these characteristics that will be emphasized here.

❖ STEADY-STATE CHARACTERISTICS

When all inputs and external influences are held constant, most, but not all, processes come to a steady state. (Liquid level is different. Unless the inflow and outflow to a liquid-level process are equal, the process will not come to a steady state, even though inflow and outflow themselves are constant.) We will use the heat exchanger depicted in Figure 2-7 to illustrate the nature of the steady state. This is redrawn as Figure 3-1 here, but in this case the independent and dependent variables are identified. Let us assume that we have a liquid phase process that must be heated to a specified temperature. We also assume that we have a liquid phase heating medium, such as hot water or hot oil.

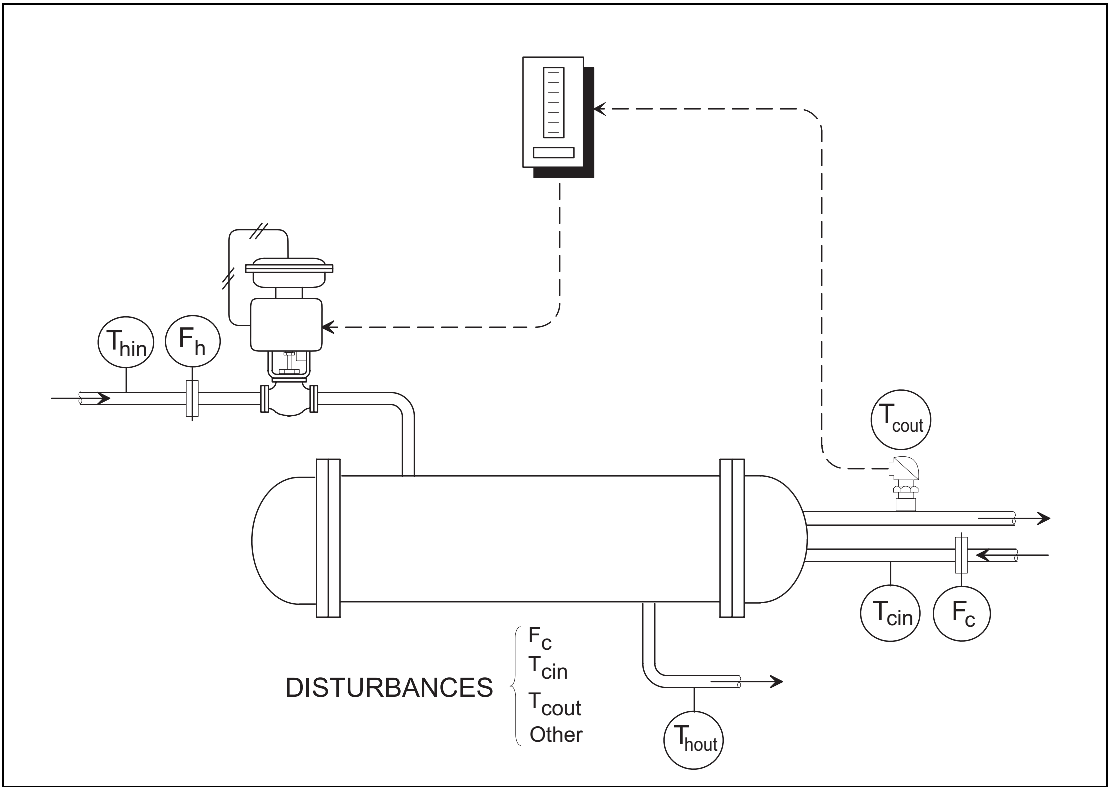


Figure 3-1. Disturbances to a Control Loop

From a process control viewpoint, the independent variable is the valve position, or equivalently, the controller output. The dependent variable of interest is the process outlet temperature, T_{out}. Other dependent variables are the flow rate of the heating fluid, F_h, and the outlet temperature of the heating medium, T_{hout}. In a typical operating plant, these may be monitored to detect abnormal operation, but from a control viewpoint they are inconsequential.

Other important variables to be considered are the disturbances to the process. These are sometimes called “load changes.” They can be considered as external, random influences on the process. It is the purpose of the control system to counteract the effect of these disturbances. Some of the disturbances that could affect the heat exchanger are the following:

- Changes in the process flow rate, F_c;
- Changes in the process inlet temperature, T_{cin};
- Changes in the source temperature of the heating medium, T_{hin};
- Changes in the upstream or downstream pressure of the heating medium. (This would change the hot stream flow rate, F_h, even though the valve position did not change.)
- Scaling of the heat exchanger tubes—thus affecting the heat transfer coefficient; and
- Environmental effects, if the heat exchanger is not perfectly insulated.

For the purposes of illustration, we will disregard the latter three of these disturbances (that is, we will assume that they are constant) and concern ourselves only with F_c, T_{cin} and T_{hin}. For the time being, we will also consider that these three variables are also being held constant. In other words, the only independent variable is the valve position, which uniquely sets the heating medium flow rate. With this consideration, we state a very important principle:

If all external influences on a process are held constant, then each value of the control signal (independent variable) produces a specific and unique value of the measurement value (dependent variable). (There are rare cases, such as the discharge pressure of a centrifugal compressor versus suction flow, or index of refraction versus composition, where this unique relation may not be true.)

This one-to-one relationship can be depicted in graphical form, as shown in Figure 3-2. We call this relationship the *process graph*. Keep in mind that the process graph depicts the steady-state relationship between the controller output (valve position) and the measurement for a particular combination of the disturbance variables. If any of the disturbance variables change in value, then we have a new process graph. Figure 3-3 shows the process graph for three combinations of disturbance variables. The graph for the original values of F_c, T_{cin} and T_{hin} is shown by the dotted line. The upper line shows the process graph for an increase in T_{cin}. The lower line shows the process graph for an increase in F_c.

The process graph—the steady-state relationship between the controller output and measured variables for a particular combination of disturbance variables—is an important concept for our understanding of control loop behavior. However, it is *not* something that we need to determine in actual practice. Indeed, it would be impractical to determine the infinite number of process graphs that would result from all combinations of the disturbance variables.

Nevertheless, we can deduce that if we wish to control the measurement to a particular value (set point), the process graph determines the required value of the controller output. If there are

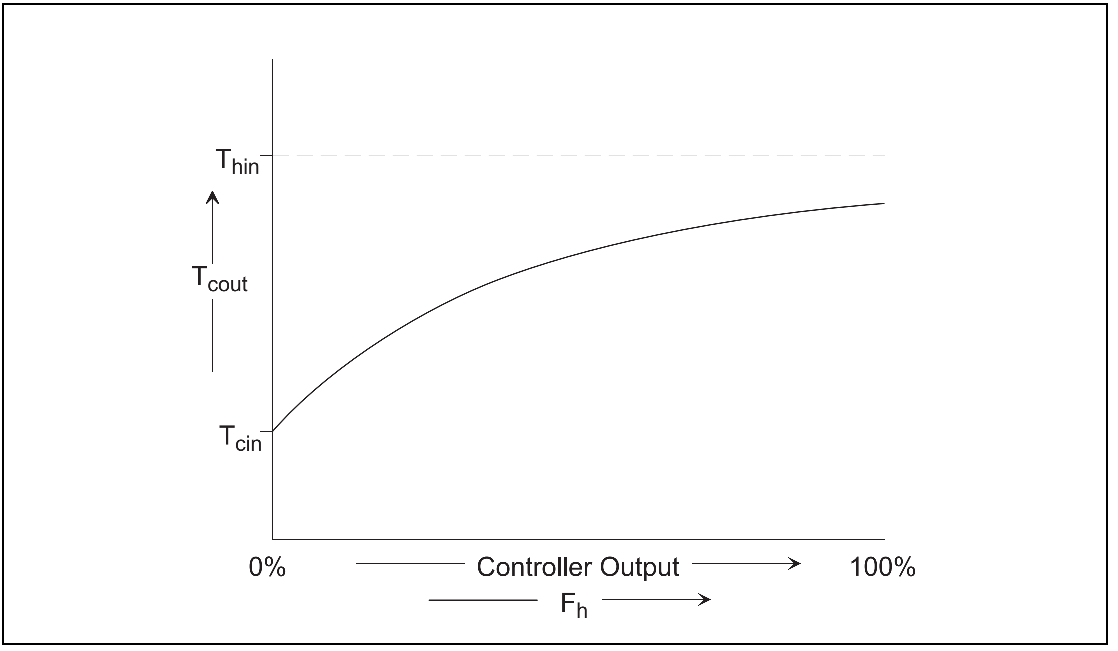


Figure 3-2. The Process Graph

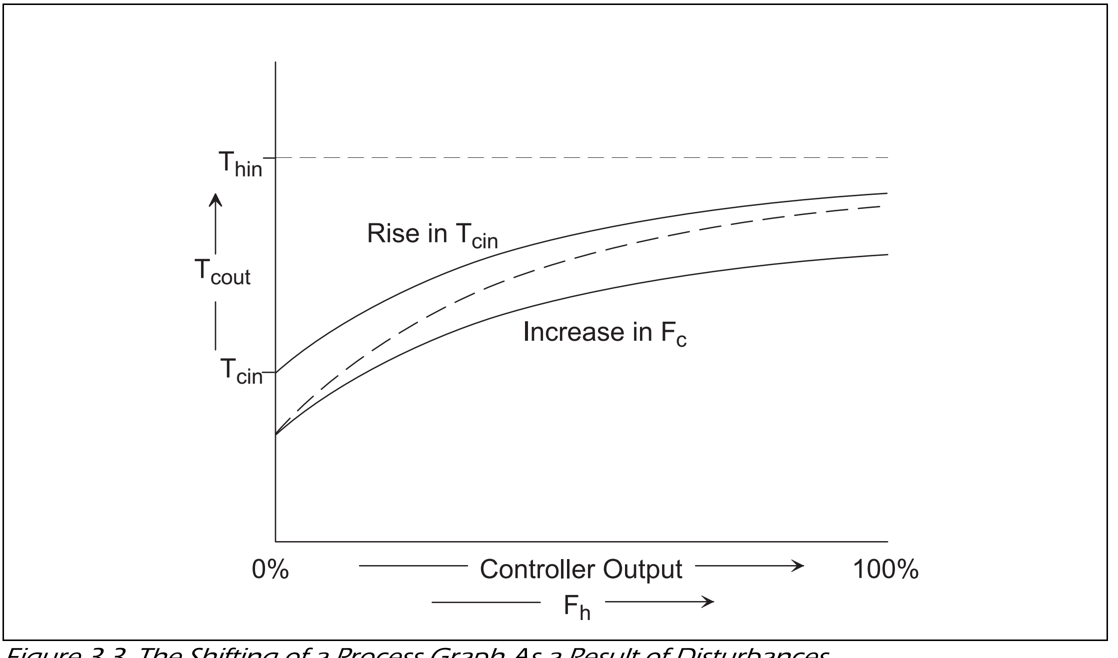


Figure 3-3. The Shifting of a Process Graph As a Result of Disturbances

load changes on the process that cause the process graph to shift, we will need a new value of the controller output. It is the duty of the controller to find the precise point on the process graph that brings the measurement to the desired value, as shown in Figure 3-4.

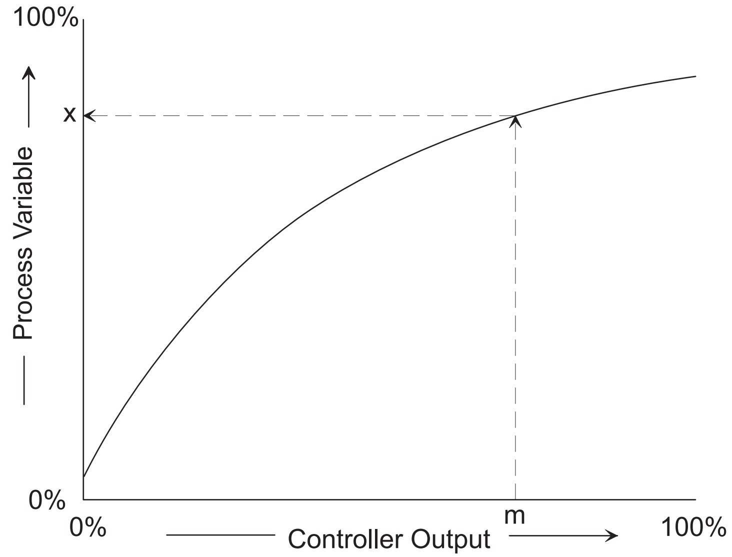


Figure 3-4. The Process Graph Determines the Controller Output Required to Bring the Measurement to a Desired Value

Although we will normally not have a precise process graph available for even one combination of disturbance variables, there are certain attributes that we should know. We must know whether the process graph slopes upward or downward. This is equivalent to saying that we must know whether the process is direct-acting or reverse-acting. An upward slope represents a direct-acting process (an increase in controller output causes an increase in measurement); a downward slope signifies reverse-acting. Recall from chapter 2 that to avoid positive feedback, the controller must be of opposite action—reverse-acting for a direct-acting process and vice versa.

We should also know, either explicitly or implicitly, the slope of the process graph, at least in the vicinity of the most probable operating point. The slope can be stated as the change in measurement divided by the change in controller output. This is called the *process gain*. Specifically, process gain, K_p, is defined by the following equation:

$$K_p = \frac{\text{Change in measurement}}{\text{Change in valve signal}} = \frac{\Delta x}{\Delta m} \tag{3-1}$$

The process gain often varies with operating point. This is equivalent to stating that the process, and hence the process graph itself, is often nonlinear. Except for some rare misbehaved processes, the process graph is monotonic. That is, it does not change the direction of the slope, so there is a unique relationship between each value of the controller output and the measurement.

Process nonlinearities can be caused by a number of conditions, including physical or chemical factors inherent in the process itself. In one frequently encountered situation the process variable responds linearly to changes in the *ratio* between two variables, such as the manipulated variable and a disturbance variable. To illustrate this, suppose that a process heater can be modeled by a simple heat-balance relationship:

$$F_p\, C_p\, (T_{out} - T_{in}) \;=\; F_g\, H_V\, E_{ff} \tag{3-2}$$

where: F_p = Heater feed rate (the disturbance variable)
C_p = Specific heat
T_{out} = Outlet temperature (the process variable)
T_{in} = Inlet temperature
F_g = Fuel rate (the manipulated variable)
H_V = Heating value of the fuel
E_{ff} = Heater efficiency

Equation 3-2 can be rearranged to express outlet temperature on the left-hand side of the equation and all other terms on the right-hand side:

$$T_{out} \;=\; T_{in} \;+\; \frac{H_V\, E_{ff}}{C_p}\frac{F_g}{F_p} \tag{3-3}$$

This demonstrates that the outlet temperature responds more or less linearly to the fuel-to-feed ratio, F_g/F_p. If a temperature controller directly manipulates the fuel rate, then the process gain seen by the controller is the sensitivity of the outlet temperature to changes in fuel rate. Specifically:

$$\frac{\Delta T_{out}}{\Delta F_g} \;=\; \frac{H_V\, E_{ff}}{C_p}\frac{1}{F_p} \tag{3-4}$$

In other words, the process gain of the control loop is inversely proportional to process flow rate. At low process flow rate (such as during start-ups), the process gain will be high; at higher flow rates, the process gain will be lower.

If the ratio itself were the manipulated variable, rather than simply the fuel rate, then the process gain seen by the control loop would be the following:

$$\frac{\Delta T_{out}}{\Delta\left(F_g / F_p\right)} \;=\; \frac{H_V\, E_{ff}}{C_p} \tag{3-5}$$

As long as the fuel heating value, heater efficiency, and specific heat of the process fluid remain fairly constant, then the control loop's process gain will remain constant. This strategy will be utilized in relation to ratio control in chapter 10 and in multiplicative feedforward control in chapter 11.

Nonlinearity in a control loop may also be caused by nonlinear characteristics of the valve.

Three common valve characteristics are linear, equal-percentage, and quick-opening. These terms describe various relationships between the area of opening for flow and the valve position. The basic relationship between area of opening and valve position is determined by the manufacturer of the valve, hence the terms *inherent* or *manufactured characteristics* are often used. If there is a constant differential pressure across the valve, the relationship between flow and valve position will be the same as the manufactured characteristics (see Figure 3-5).

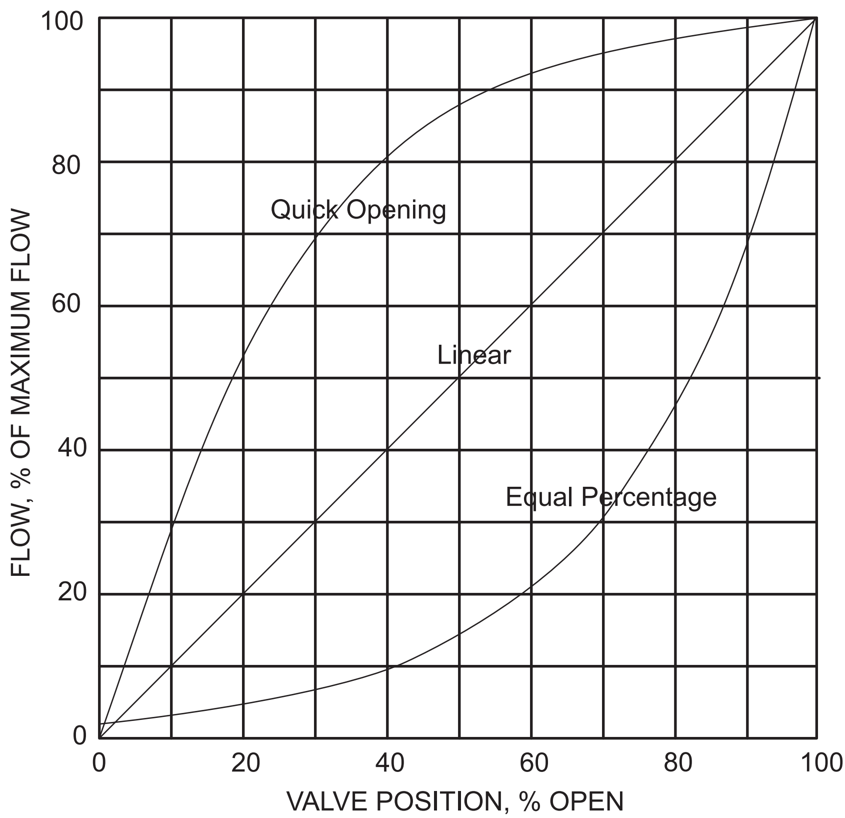


Figure 3-5. With Constant ΔP across the Valve, Flow versus Valve Position Follows the Manufactured Characteristics

Normally, the pressure differential across a valve is variable, being at a maximum when the valve is closed, and decreasing with increasing flow rate. The amount of decrease depends upon how much the other restrictions in the flow line tend to dominate at the higher flow rates. The actual relation between flow and valve position is determined by both the manufactured characteristics of the valve and the decrease in pressure at higher flow rates. This is termed the *installed characteristics*. Figures 3-6 and 3-7 show the installed characteristics for linear and equal-percentage valves, for a range of values of the ratio[1]:

$$\frac{\text{Minimum } \Delta P \text{ (when valve is wide open)}}{\text{Maximum } \Delta P \text{ (when valve is closed)}}$$

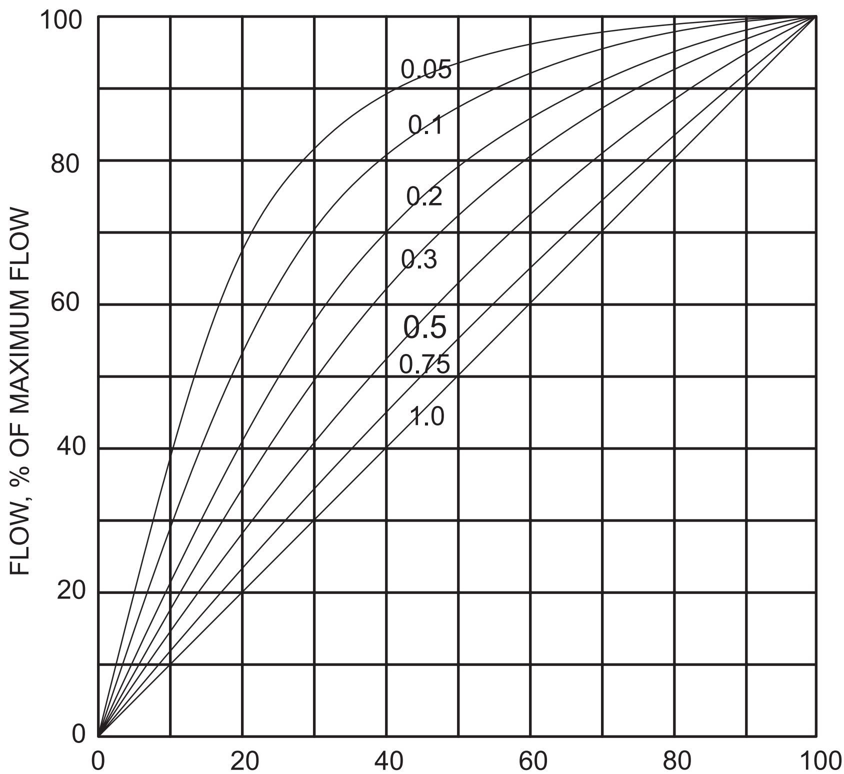


Figure 3-6. Installed Characteristics for a Linear Valve for a Range of Pressure Drop Ratios: $\frac{\text{Minimum } \Delta P}{\text{Maximum } \Delta P}$

1. Ref. 3-1 refers to this ratio as a “distortion coefficient” since its effect is to distort the manufactured characteristics into installed characteristics, as shown in Figures 3-6 and 3-7.

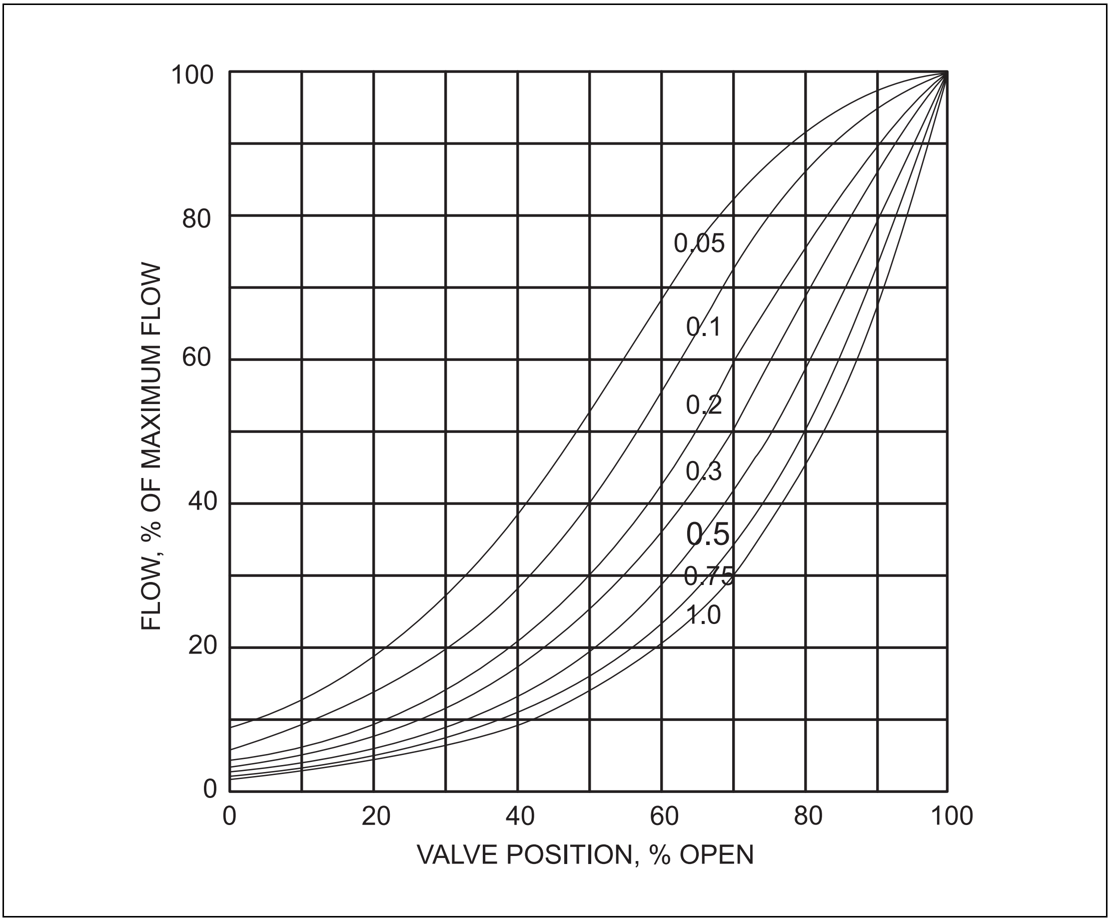


Figure 3-7. Installed Characteristics for a Linear Valve for a Range of Pressure Drop Ratios: $\frac{\text{Minimum } \Delta P}{\text{Maximum } \Delta P}$

Appendix B derives the relationships used to develop the graphs in Figures 3-6 and 3-7.

❖ DYNAMIC CHARACTERISTICS

Process dynamic characteristics can be classified into three broad categories:

- Self-regulating
- Non-self-regulating
- Open-loop unstable

The first two are depicted by the hydraulic analogies in Figure 3-8. The response of these three characteristics to a step change in input is shown in Figure 3-9.

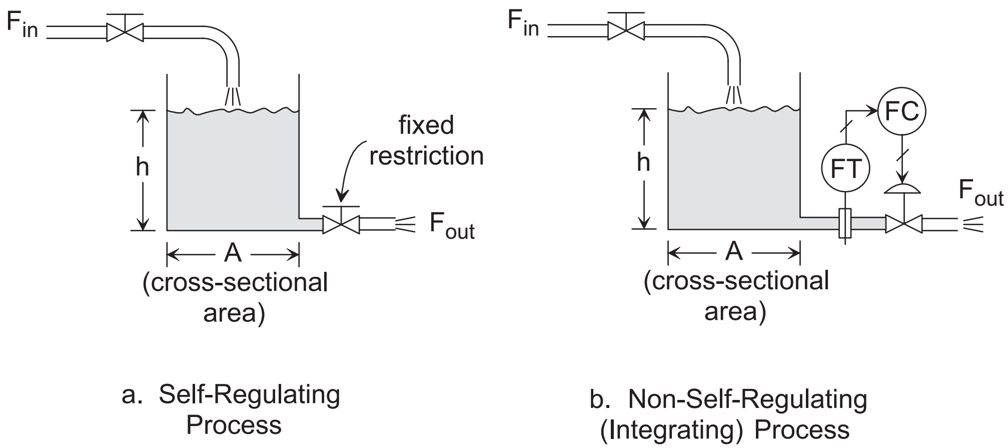


Figure 3-8. Hydraulic Analogies of Common Types of Process Characteristics

Self-regulating processes are those that, if all inputs are fixed, will seek their own equilibrium. Figure 3-8a shows a tank that has a fixed restriction in the outflow line. If the flow into the vessel is fixed, the level will reach equilibrium when the hydrostatic pressure at the base of the vessel causes the outflow to exactly equal the inflow. This is the analogy of many physical processes. For instance, in a thermal process if there is a certain rate of heat input, the temperature will rise or fall until the heat lost, either to the environment or when carried away by effluent streams, is exactly equal to the rate of heat input. Most processes fall into this category.

Non-self-regulating processes can be depicted by the hydraulic analogy of Figure 3-8b. Here, there is a fixed flow rate out of the tank that is controlled by the flow controller. The outflow does not depend upon the level in the tank. Unless the inflow is precisely the same as the outflow, the level will continue to rise or fall until either the tank overflows or becomes empty.

A mathematical expression for a process of this type is given by the following integral equation:

$$Ah = \int (f_{in} - f_{out})\, dt \tag{3-6}$$

where: h = height of fluid in tank (i.e., level)
A = cross-sectional area
f_{in}, f_{out} = volumetric flow rates

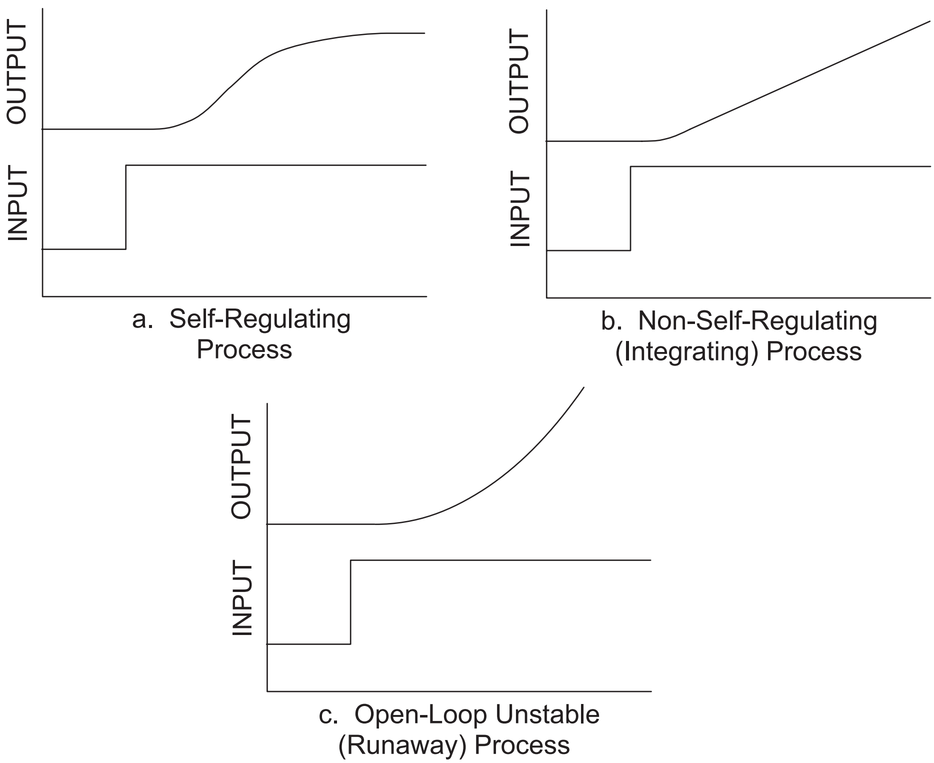


Figure 3-9. Step Input Response of Common Types of Process Characteristics

For this reason, non-self-regulating processes are often called *integrating* processes. In actual practice, liquid-level control can usually be represented as an integrating process. Another example of an integrating process is pressure control for a liquid-vapor system. If the heat input exceeds the heat removal rate, the pressure will continue to rise.

The number of processes that are self-regulating is much greater than the number that are non-self-regulating.

A few processes are unstable in the open loop (i.e., a loop without feedback control). These are called “runaway” processes. A typical example is a jacketed exothermic reactor. As the reactor temperature is increased, the reaction rate increases. But at the higher reaction rate, more heat is generated; also more heat is removed due to a higher differential temperature between the reactants and the jacket. If there is insufficient heat transfer, then excess heat will be generated, causing the temperature to rise even further.

Figure 3-10 illustrates heat flow versus temperature curves for an exothermic reactor. The curved line represents heat generation; the straight lines represent heat removal. The solid straight line represents one heat removal relation, determined by the heat transfer surface area.

The dashed straight line represents increased heat removal, due to, say, increased surface area. The intersection of the heat generation and heat removal curves represents an equilibrium point; at that point as much heat is being removed as is being generated. If the temperature is above (below) that point, and if the heat generation curve lies above (below) the heat removal curve, then excess heat is being generated (removed) which will cause the temperature to move away from the equilibrium point. On the other hand, if the heat generation curve is below (above) the heat removal curve, then the temperature will move toward the equilibrium point. The solid line for heat removal represents an open-loop unstable process, whereas the dashed line represents an open-loop stable process.

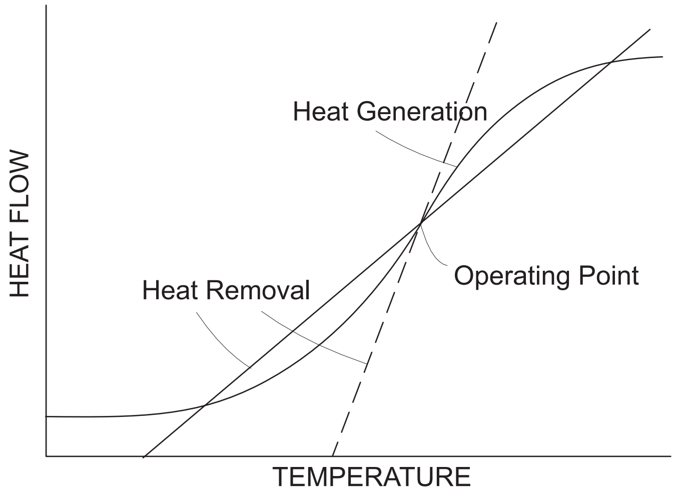


Figure 3-10. Heat Flow versus Temperature Curves for Exothermic Reactor

In chapter 4, we will mention the use of the derivative mode in a controller to provide a stabilizing effect in control loops containing open-loop unstable processes.

♦ Types of Dynamic Response

The simplest type of self-regulating process is one in which there is a single location for mass or energy storage. An example of such a process is shown in Figure 3-11. Here, a constantly flowing stream is heated in a well-mixed vessel. If the heat storage is negligible in both the walls of the vessel and in the heating coil itself, then the only point of heat storage is in the fluid itself. On a step increase in the heat input, the temperature will respond as shown in Figure 2-4. This is the typical response of a first-order lag; hence the process can be described by a two-parameter process model that consists of a process gain and a time constant.

Another elementary type of response is pure dead time. Usually associated with the physical movement of mass or energy, this is often called the “transport lag.” An example would be a

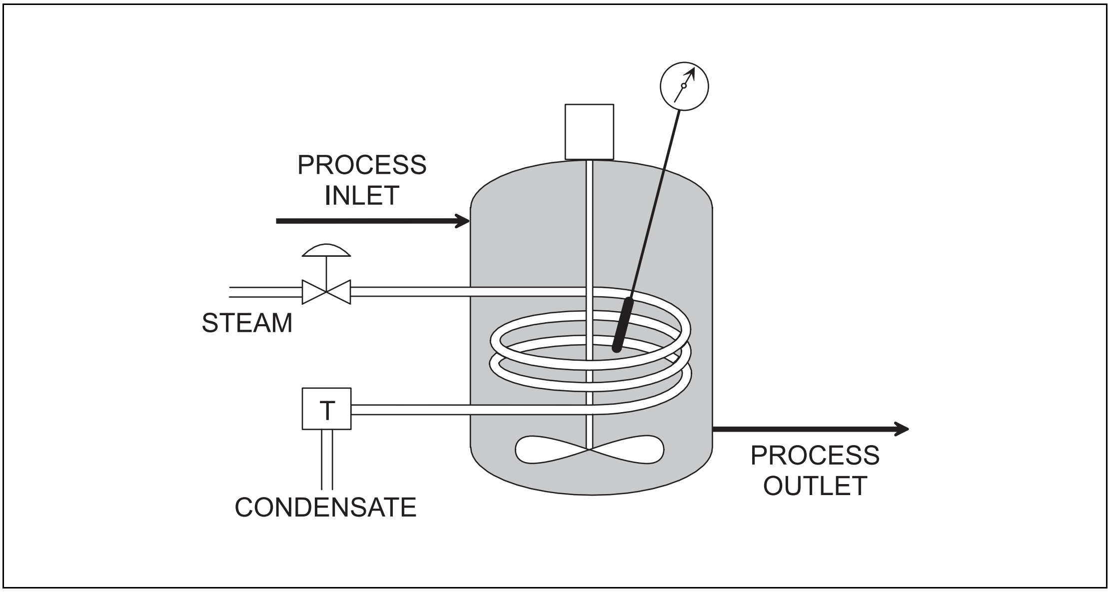


Figure 3-11. Process with Single Point of Energy Storage

well-insulated flowing pipeline, where the temperature is measured at two points separated by a considerable distance. The temperatures recorded from the two measuring points would be identical except they would be separated by the time required for the fluid to move from the upstream to the downstream point of measurement. This process can be described by a single parameter model that represents the dead time, as shown in Figure 2-5.

First-order lags and dead times can be considered as "elements" that, combined in a myriad ways, comprise the dynamic characteristics of real processes. Most self-regulating processes are not as simple as those just described; rather, they consist of multiple locations for mass or energy storage plus, perhaps, the time for the movement of the material. We will describe several elementary combinations, giving a physical example and a hydraulic analogy of each. Our purpose is to give the reader an intuitive understanding of why processes behave as they do.

Two different hydraulic analogies can be given for two locations for mass storage. In Figure 3-12a, the level in tank 2 has no influence on the flow of fluid between the tanks, whereas in Figure 3-12b, the driving force for flow is the difference in level between the tanks. Both of these situations represent two first-order lags, but in Figure 3-12a the lags are said to be *uncoupled*, whereas in Figure 3-12b, the lags are said to be *coupled*.

In both cases, the response to a step change in input will be an "S-shaped curve" rather than the idealized first-order lag response of Figure 2-4. A more detailed discussion of the shape of the S-shaped curve follows later in this section.

We note, however, that coupled lags are representative of many physical processes. For example, heat transfer from the hot to the cold side of a heat exchanger is dependent upon the differential temperature between the two sides.

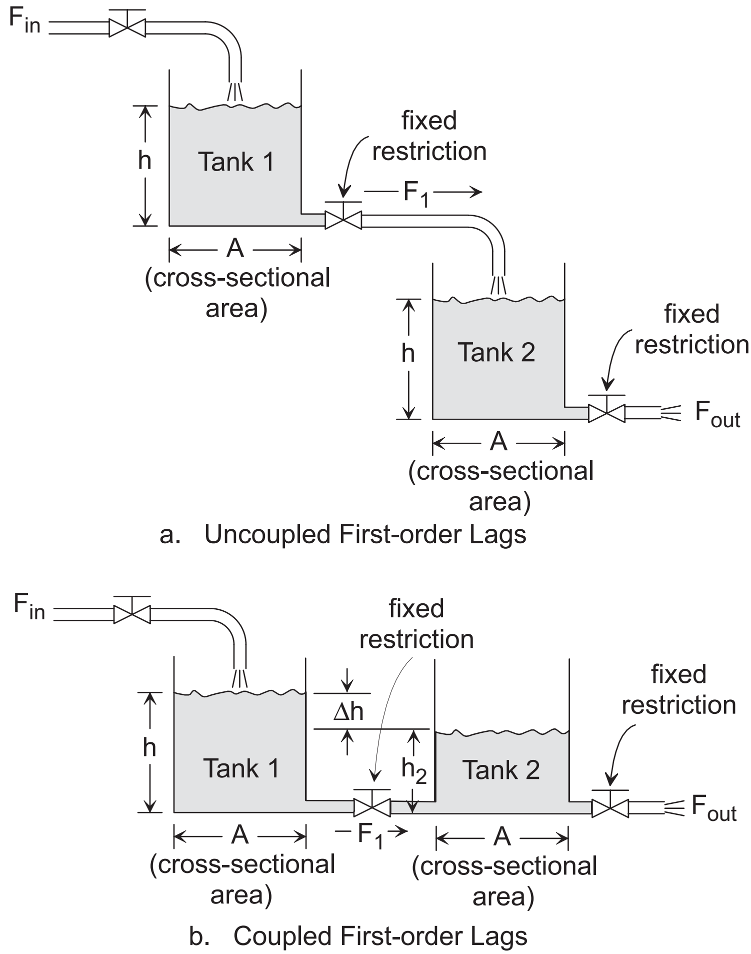


Figure 3-12. Hydraulic Analogies for Two Points of Mass Storage

It is interesting to consider the effect of many lags in series, both coupled and uncoupled lags. Continuing with our hydraulic analogies, assume that we start with one tank whose time constant is some value τ. The response to a step change in inflow is the familiar first-order lag graph, which is designated by “1” in both Figures 3-13 and 3-14.

Now suppose we replace that tank with two smaller tanks, each of which is half the size of the original tank. It is reasonable to assume that the time constant of each of the smaller tanks is $\tau / 2$. These tanks can either be configured as “uncoupled” lags or as “coupled” lags, as shown in Figures 3-13 and 3-14. Note that the mass is now distributed between the two tanks. The response to a step input change is shown by the graphs labeled “2” in Figures 3-13 and 3-14.

Now replace the two tanks with three smaller tanks, each of which is a third the size of the original tanks and has a time constant of $\tau / 3$. The step response is shown by the graphs labeled “3” in Figures 3-13 and 3-14. Continue in this manner, using *N* tanks, each holding 1/*N* volumetric units and having a time constant of τ / N.

For larger and larger values of N, the responses will continue to be S-shaped curves. With the uncoupled first-order lags, the graphs will be more severely curved. In the limit, as *N* approaches ∞, the S-shaped response will be converted into a pure dead time, with a time delay of τ minutes. (Chemical engineers will recognize this thought process as the passage from a stirred-tank reactor to a plug flow reactor.)

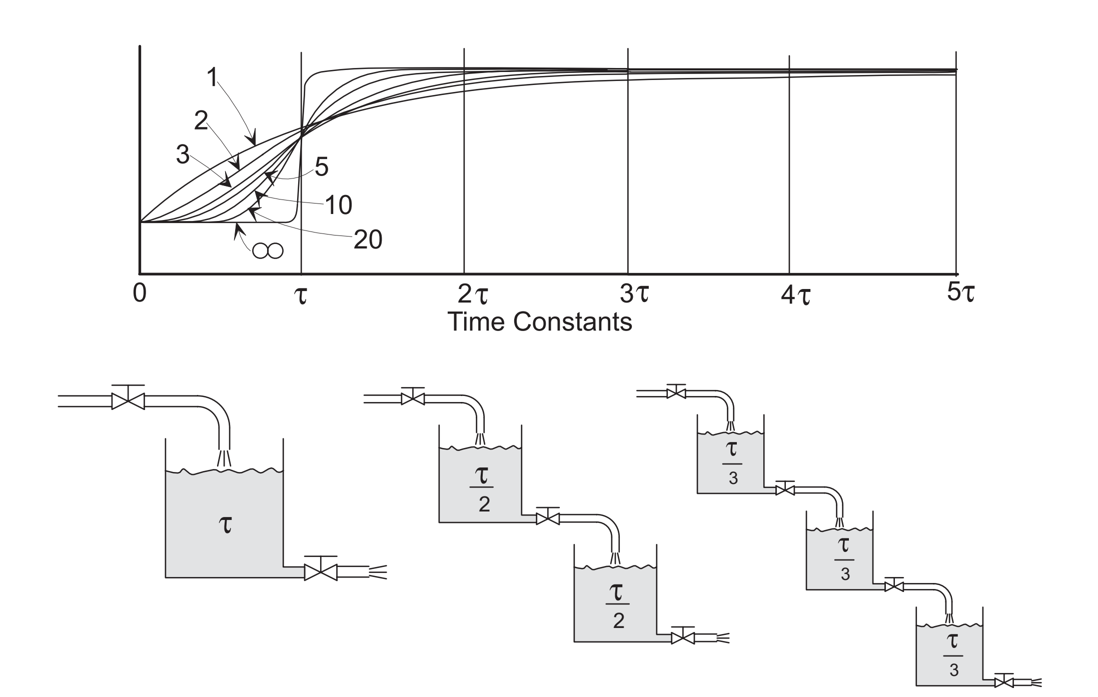


Figure 3-13. The Response of Multiple Uncoupled First-order Lags in Series

With coupled first-order lags, the apparent dead time does not increase as drastically, but the system becomes increasingly sluggish. The process appears to have a very long dominant time constant, much longer than if all the mass were concentrated at one location (a single first-order lag).

The physical import of this discussion is to demonstrate that the more widely the mass or energy is distributed, the more the process assumes the characteristics of dead time, particularly with uncoupled first-order lags. With coupled first-order lags, the process takes on the characteristics of a moderate amount of dead time plus a long dominant time constant.

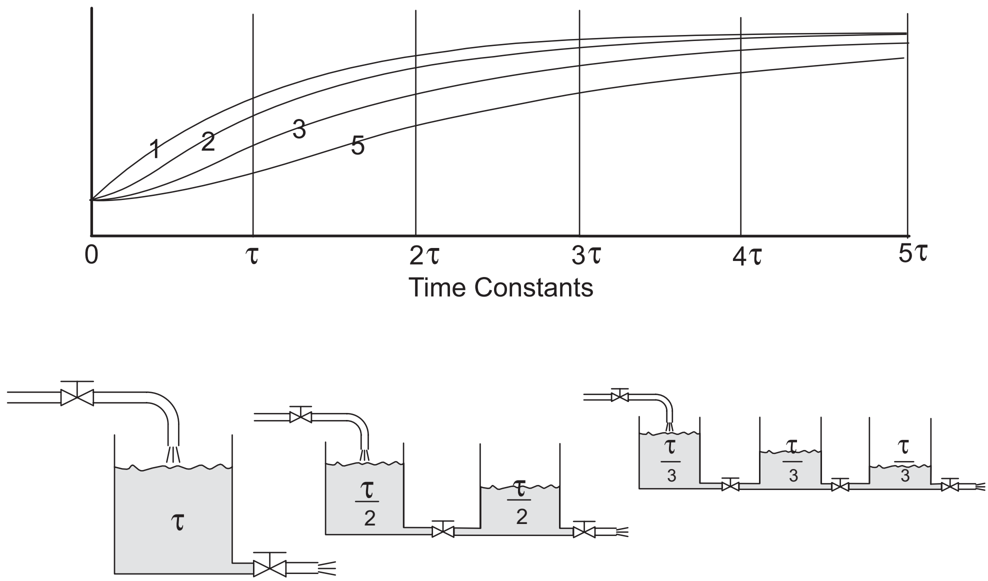


Figure 3-14. The Response of Multiple Coupled First-order Lags in Series

Let us consider a real example, rather than a hydraulic analogy. Suppose we have a liquid-filled, well-mixed tank that is heated by an internal steam coil, as shown in Figure 3-15. This is the same process shown in Figure 3-11, except that now we will consider the heat stored in the heating coil, the walls of the coil, and in the temperature-sensing bulb itself.

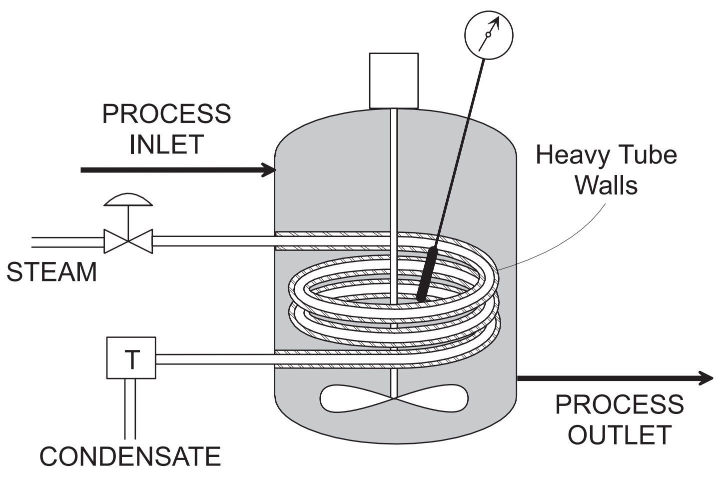


Figure 3-15. Example of Coupled First-order Lags

Suppose that initially the process is stable, with the steam valve in an intermediate position. There will be some amount of pressure drop across the steam valve. Assuming that there is a constant steam supply pressure, this will determine the pressure inside the coil. Since the steam is condensing within the coil, it will be in a saturated condition. Hence, its temperature is indirectly determined by the position of the valve.

From this stable condition, let us make a step increase in the valve position. This will increase the steam flow, causing the pressure in the coil to rise. The steam within the coil represents a point of (thermal) energy storage. The steam pressure, and hence the temperature in the coil, will not rise instantaneously. It will rise only as the mass of steam within the coil increases. This will be similar to the response of a first-order lag, with a very short time constant.

As the temperature of the steam within the coil rises, the temperature difference between the steam and the metal in the coil walls increases. This causes increased heat flow into the metal walls of the coil, which represents another point of energy storage. The temperature of the coil walls increases, which increases the temperature difference between the coil and the fluid itself. This in turn causes an increase in heat flow into the fluid, which causes the fluid temperature to rise. Finally, as the fluid temperature rises, the temperature difference between the fluid and the thermal bulb increases. This causes an increase in heat flow, and consequently an increase in the temperature sensed by the thermal bulb.

This somewhat tedious discussion shows that this process can be represented by the hydraulic analogy, much as in Figure 3-14 but with four tanks in series, representing a series of four coupled first-order lags. Hence, the expected response to a step change in valve position would be an S-shaped curve with a long response, something like those shown in Figure 3-14.

Let us consider another type of response. Start with the process represented by Figure 3-11, but now suppose that the vessel has a metal tank wall of significant thickness, as shown in Figure 3-16. This represents a point of energy storage. This is a first-order lag that is not in series with the main signal flow, but is at the side. A step change in heat into the tank must furnish heat not only to raise the fluid temperature, but also to raise the temperature of the metal wall of the vessel. A hydraulic analogy of a side lag is given in Figure 3-17.

If the ratio of the side lag time constant to the basic time constant is small (e.g., a very thin vessel wall and a high heat-transfer coefficient from the fluid to the wall), then the response will be essentially that of a first-order lag. On the other hand, suppose that the ratio of the side lag time constant to the basic time constant is large (e.g., a very thick vessel wall and a low heat-transfer coefficient from the fluid to the wall). In this case, the output response to a step input will be a relatively rapid rise to partial equilibrium, followed by a slow change the rest of the way to equilibrium. Figure 3-18 shows the output response for a range of ratios of the side lag to basic time constants. This phenomenon can explain such observed physical behavior as an initial rapid change in the variable followed by a lengthy period of slowly drifting change.

These examples have illustrated the types of response that may be expected if the points of mass or energy storage (first-order lags) are in parallel rather than in series, or at the side. In

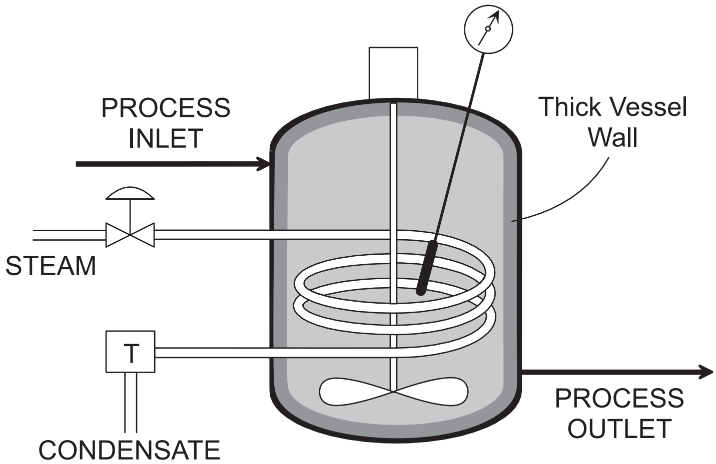


Figure 3-16. Example of Side Lag

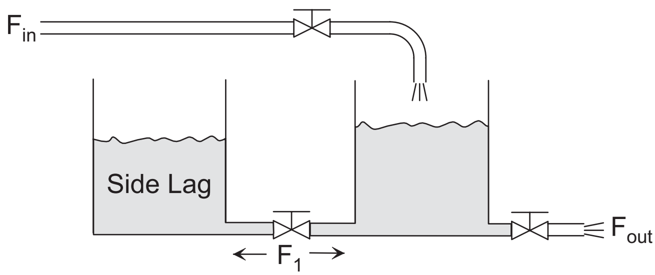


Figure 3-17. Hydraulic Analogy of Side Lag

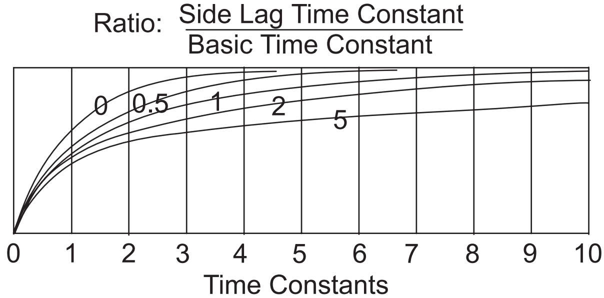


Figure 3-18. Response of Basic Lag Plus Side Lag

real processes, infinite combinations of these elementary examples are possible. We can have points of energy storage, some in parallel, some in series, and some at the side. Thus, depending upon the physical parameter values, we might expect an infinite number of responses for a step input change. But the responses will usually have the form of an S-shaped curve. Other forms of response that we occasionally encounter will be covered later.

We have purposely avoided giving rigorous forms of models, such as explicit transfer functions, for these examples. Doing so would not be meaningful. In a real industrial process application, we probably will not know the precise structure of the system—how many points of mass or energy storage there are in series, parallel, and so on. It would therefore be futile to try to construct an exact mathematical model. This situation differs from that of electro-mechanical systems (flight control, robotics, etc.), which are often comprised of discrete components and hence are amenable to precise mathematical modeling. For industrial process control, we have done well to recognize the potential forms of response and some of the factors that give rise to them.

In summary, we have concluded that the step response of most (but not all) self-regulating processes will be an S-shaped curve. We cannot formulate a precise process model, but we can approximate the response with a simplified form. The form we will choose is first-order plus dead time (FOPDT). This requires three parameters: process gain, pseudo-dead time, and pseudo-time constant. We will leave for chapter 6 a discussion of the task of obtaining actual numerical values for these parameters.

A self-regulating process can exhibit other forms of response, for example, the underdamped (decaying oscillation) form shown in Figure 3-19. While this type of response is often seen in closed-loop systems (when the controller is in the automatic mode), it is very unusual to see this as an open-loop response (when the controller is in the manual mode). The reason for this is that feedback is present in the closed loop.

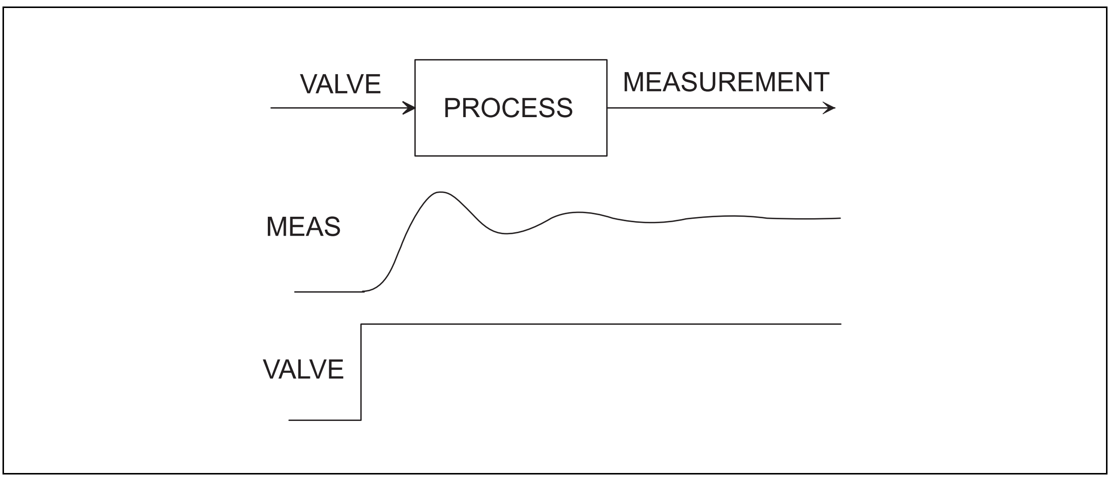


Figure 3-19. Open-Loop Response: Damped Oscillatory Behavior

If the presence of feedback is the cause for oscillation in the closed loop, then we have a clue as to what might cause oscillation in the open loop. Suppose that in a chemical process there is a point of control action (i.e., a control valve) and that downstream from this there is a sensor, perhaps a composition analyzer. Suppose further that between the point of control and the sensor there is a chemical recycle loop. This is a form of feedback that could result in oscillation.

As another example, suppose we have a cascade control system. (See chapter 9 for a full discussion of cascade control.) Normally, the outer, or primary, loop should be significantly slower than the inner, or secondary, loop. But if this is not the case, then when the secondary controller is in the automatic mode and the primary is in manual, a step change in the output of the primary controller may cause the primary measurement to oscillate because feedback is present in the secondary loop.

These examples are hypotheses of how oscillation could occur in the open loop. While actual situations resembling these may occur, the situation is sufficiently unusual in industrial process control systems that it need not be considered further here.[2]

A more important consideration, although still somewhat rare, is that of an inverse response. With an inverse response, when the valve undergoes a step change, the measurement tends to initially go the “wrong” way. It then reverses and comes to an equilibrium in the predicted direction, as shown in Figure 3-20a. This situation is also applicable to some integrating processes, whose response is shown in Figure 3-20b.

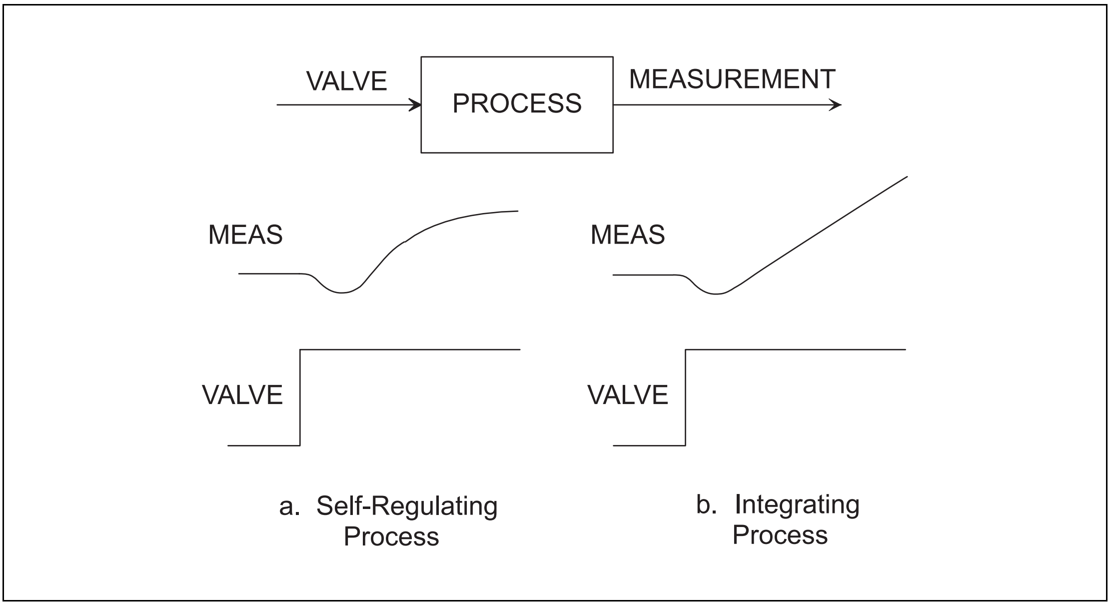


Figure 3-20. Inverse Open-loop Response

2. Underdamped behavior in the open loop is probably more frequent in mechanical and electro-mechanical systems. The typical behavior of the shock absorber of an automobile is a good example.

An oft-cited example of inverse response is the shrink-and-swell phenomena that occurs in the feedwater drum level of large steam generators. The drum level will remain constant if there is a constant feedwater rate, if there are no water losses and no blow-down within the boiler, and if the steam take-off rate, in mass units per hour, exactly matches the feedwater rate.

Suppose, however, that from this equilibrium situation the steam demand valve is opened slightly, increasing the steam flow. If the feedwater rate is unchanged, what will the drum level do? The first and obvious thought is that the drum level will go down, since more steam is being withdrawn than is being replaced with feedwater. However, the boiler drum does not respond that way because of a secondary phenomenon. Within the drum, as well as within the water tubes inside the boiler, there is water at its flash-point temperature. There are compressed bubbles of vapor, and when the steam demand is increased, the steam pressure falls slightly. This causes some of the water to flash into steam; it also causes the bubbles to expand. The expansion of vapor bubbles in the drum, together with the water forced from the tubes into the drum as a result of the expansion of vapor in the tubes themselves, causes the water level in the drum to rise momentarily.

If the feedwater is not adjusted, eventually the drum level will fall. This is because of the obvious first-principles effect of more steam being withdrawn than is replenished by the feedwater rate. This is an example of an inverse response on an integrating process.

If a simple level-control system were installed for this application, with the level controller setting the feedwater rate, then when steam demand increased, the level controller would, in the short-term, sense a rising drum level and decrease the feedwater flow rate—exactly the wrong control action to take.

For this reason, practically all large steam-generating systems are equipped with a three-element drum level-control system in which a measure of the steam rate immediately affects the feedwater rate. If this effect does not maintain long-term drum level control, then the drum level controller trims the feedwater rate. The drum level controller is tuned for a slow response so it ignores the initial inverse response of the drum level.

The process characteristics discussed so far have involved processes that have a single input and a single output or process variable. In industrial applications, however, a process often has multiple inputs (control valves) and multiple measurements. Each valve affects several measurements, and in turn each measurement is affected by several valves, as depicted by Figure 3-21. Such processes are said to be interacting. Between any single valve and a related measurement any of the forms of response we have depicted in this chapter may be exhibited. But the control system is complicated by the interaction that is present. For now, it will suffice to call the reader's attention to this type of process behavior. Chapter 13 will discuss control techniques for multiple-input, multiple-output processes.

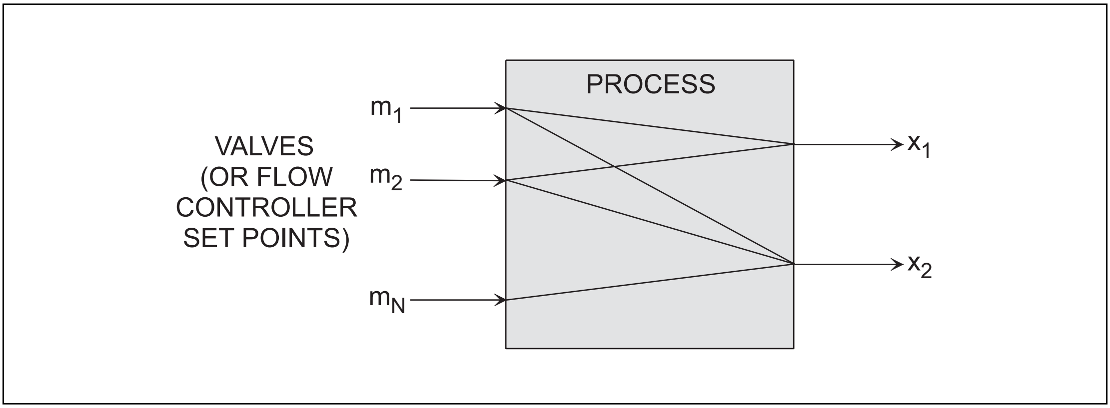


Figure 3-21. Multiple-Input, Multiple-Output Processes (Interacting Control Loops)

❖ CONTROL LOOP CHARACTERISTICS

The previous sections of this chapter have described both steady-state and dynamic characteristics of generic processes. In this section we will become more specific and describe the characteristics of common types of control loops. It is always a risky business to generalize about the characteristics of control loops (or, for that matter, of people, automobiles, stereo systems, etc.), since there are certainly exceptions to the generalization. In fact, a generalization is only what the writer or speaker believes to be the most prevalent characteristic based upon his or her experience and background. The risk is that a reader or listener may only have experienced the exception, and therefore take umbrage at the stated characteristic. Nevertheless, in this section we will present general characteristics, as well as point out exceptions, to common types of control loops, including flow, temperature, pressure, and liquid level. These types cover the largest portion of the control loops found at any plant (see Ref. 3-2).

Of necessity, we will use terms such as *controller gain*, *proportional band*, *integral*, *derivative*, *cascade control*, and the like, though we will not introduce these terms until later chapters, in particular, chapters 4 and 9. Most readers will probably have sufficient familiarity with these terms to understand their use here. We recommend that those readers for whom these terms are completely new reread this section after reading the chapters in which these terms are defined.

♦ Flow Control Loops

The characteristics of a flow control loop are influenced by several factors. These include what the flow stream is (liquid, gas, or two-phase liquid and vapor), how the flow is measured, how the flow is manipulated (i.e., the final control element), the relationship between the final control element and its piping environment, and the form of the controller itself. As a quick summary, a flow loop can be characterized as relatively fast, nonlinear, and often noisy. Let us explore each of these attributes.

The dynamic character of a flow loop is most often dominated by the dynamics of the final control element. If this element is a traditional control valve, then the speed of the actuator is the dominating dynamic element. A spring-opposed air-operated actuator will act as a first-order lag for small signal changes. For large signal changes, the changing volume within the actuator acts as a velocity limit on changes in stem position. In addition, the friction of the stem packing introduces a hysteresis effect between the signal-to-valve and the actual stem position. A valve positioner will decrease the effect of hysteresis, thereby improving the response of the stem positioning. While valve positioners are highly recommended for most control loops, in a flow loop the valve positioner and the flow controller may respond on approximately the same time scale, therefore causing the two to interact. (See chapter 9 for further discussion of the interaction of cascaded controllers.) Hence, it is often recommended that valve positioners not be used on flow loops. Many practitioners disagree with this recommendation, however, preferring to use a positioner on all control valves, even flow. Any tendency toward interaction between the positioner and the flow controller is then compensated for by reduced controller tuning. In general, it can be said that most flow loops, if they oscillate, will have a period of around one to three seconds.

We are interested in linearity because our preference would be that the loop have the same response with the set point at, say, 70 percent as it does when the set point is at, say, 30 percent. Flow loops are frequently nonlinear and have the maddening aspect that the type of nonlinearity found in one loop may not be the same as that found in another loop.

The nonlinearity is determined by the characteristics of the final control element, the type of flow measurement used, and the effect of other restrictions in the flow line. We discussed installed valve characteristics earlier in this chapter.

Another factor to consider when designing flow control loops is the flow sensor and its signal to the controller. If flow is measured by measuring differential pressure across an orifice plate, then the signal is proportional to the square of the flow, not to the flow itself. In older installations, the transmitted and displayed signal was proportional to the square of the flow. The display was on a nonlinear scale as shown in Figure 3-22. If this signal were also used as the process variable for a controller, then the squared relationship introduced an additional nonlinearity into the control loop. An equal-percentage valve would produce a great variation in process gain over the full range of the valve. A better choice of valve would be a quick-opening valve if there is no change in pressure drop across the valve, or a linear valve if there is a significant decrease in pressure drop across the valve as the flow increases.

With current technology, square root extraction is probably performed someplace within the loop, either in the sensor transmitter or in software in a digital-based controller. This removes one of the nonlinearities. As a result, an equal-percentage valve should be used if there is a significant decrease in pressure drop across the valve with increasing flow; a linear valve should be used if the pressure drop is relatively constant.

In addition to nonlinearities, another consideration in flow control loops is the measurement noise caused by turbulence in the line. Some types of sensors will produce more noise than

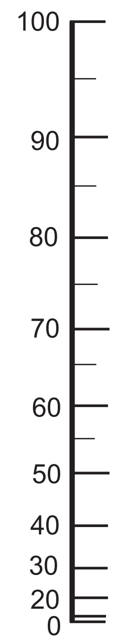


Figure 3-22. Nonlinear Display Scale When Flow Is Measured by a Differential Pressure Sensor without Square Root Extraction

others. Head (differential pressure)-producing devices and vortex meters will be the noisiest. Magnetic flow meters and Coriolis meters will be less noisy, and turbine meters will be the least noisy. To some extent, the noise can be filtered out, either in software or hardware, but doing so will retard the response of the control loop. There are two primary reasons for filtering out the measurement noise—first, to improve the appearance of the chart record or indication and second to preprocess the signal before it is sampled for logging and/or archiving purposes. In most analog controllers, the same signal is used for indication as for control. Hence, any filtering performed will affect both. In many digital-based systems, the control signal can be separated from the indication signal. Here, filtering can be applied to the indication signal without filtering and retarding the control signal.

Flow loops are usually tuned with a low gain (wide proportional band) and a relatively short integral time. (See chapter 4 for a description of tuning parameters and terminology.) This is especially true with digital flow controllers, in which the dead time caused by the scan time is significant compared to the rather short lag in the process. Analog controllers were often tuned with a high gain and long integral time. Because of the measurement noise present, derivative is never used in a flow loop.

♦ Temperature Control Loops

The almost diametric opposite of a flow control loop is the temperature loop. Usually, these are relatively slow and except in unusual circumstances are fairly noise-free. As for nonlinearity, in many temperature loops the process gain is inversely proportional to the process throughput.

The process dynamic characteristics are different for temperature control loops for heat exchangers, process heaters, and distillation columns. Heat exchangers are likely to exhibit significant dead time, whereas for process heaters, the loop will have the appearance of a single dominant location for energy storage. Most temperature loops are also self-regulating. Because of these factors, the most significant dynamics can be approximated as a first-order lag. Usually, there are several smaller points of energy storage (the valve actuator, piping walls, the thermocouple well, etc.). These combine to produce an apparent dead time. In addition, there may be a true dead time, due to the transport time of a flowing stream. (Transport time from the process unit to a sensing device counts as dead time. Transport time from a control valve to the process unit does not count as dead time.)

Thus, a FOPDT model is often a fairly good process approximation in a temperature loop. The time constant and the dead time will both increase, however, as the product flow rate (throughput) decreases.

Changes in throughput have a more significant effect on the process gain. As the throughput decreases, an incremental change in fuel flow will have an increasing effect on temperature, since there is less fluid to absorb the released heat. Hence, the process gain is inversely proportional to the throughput rate. If the temperature controller directly manipulates the fuel valve, an equal-percentage valve will provide a compensating nonlinearity. If cascade control is used, it is often in the form of ratio control, where the temperature controller sets a ratio of fuel (or heating source) to process load. Utilizing these control techniques tends to linearize an otherwise nonlinear loop.

Temperature control loops are often tuned with a relatively high gain or narrow proportional band and a moderately long integral time. Because of the absence of noise, they are ideal candidates for using derivative. The use of derivative in temperature loops is also helpful for overcoming the small secondary lag caused by the temperature measurement.

♦ Pressure Control Loops

Pressure control loops can be categorized according to the controlled fluid, which can be liquid, vapor or gas. For controlling pressure in a flowing line, pressure control loops can be further categorized by the arrangement of pressure sensor and control element. If the sensor is downstream of the control valve, the control loop is called a “pressure regulator” or “pressure reducing station.” If the sensor is upstream of the control valve, the control loop is called a “back pressure regulator.” The point of control may be far removed from the point of measurement. For instance, controlling the condenser capacity on the overhead of a distillation column may affect the pressure on a series of upstream columns.

Liquid-pressure loops are not common. In essence, liquid pressure is controlled by a flow balance into and out of the controlled volume. Hence, the characteristics are quite similar to that of flow loops.

If heat is added to a liquid that is at its boiling-point temperature, then a vapor is evolved that creates a pressure above the liquid. Suppose the vapor is removed through a fixed restriction, as shown in Figure 3-23a or 3-23b. Then the pressure can be controlled by manipulating the heat input. The pressure loop in this case controls a self-regulating process.

If, on the other hand, there is a fixed demand for the vapor, say by means of a flow controller as shown in Figure 3-23c or 3-23d, then the pressure can still be controlled by manipulating the heat input. However, the pressure is caused by the integral of the difference between energy input (heat) and energy removed in the compressed vapor. Therefore, this pressure loop controls an integrating process.

These elementary examples illustrate that pressure control of a boiling liquid may control either a self-regulating or an integrating process. Let us apply these examples to real-life situations. For example, at a steam generator there is normally an independent demand for steam, analogous to that shown in Figure 3-23c. At a distillation tower, the vapor may be controlled, either directly or inferentially, for composition control, analogous to that illustrated in Figure 3-23d. Hence, we can conclude that pressure control of a boiling liquid most often involves an integrating process.

In chapter 6, on feedback controller tuning, we will see that if a controller of an integrating process relies too heavily on the integral mode, an oscillation will result. Therefore, pressure control loops for boiling liquid vapors should be tuned with a relatively high gain (narrow proportional band) in order to minimize offset and relatively slow integral action. If the process is self-regulating, it is basically a thermal process, and hence has characteristics similar to temperature loops. This again leads to relatively high-gain, slow integral action tuning. Since these loops are almost noise-free, they, along with temperature loops, are good candidates for derivative.

The third type of pressure-control loop is that used for single-phase gas pressure. These loops are generally characterized as being fast and noise-free. If the closed piping network is relatively short (this excludes gas transmission pipelines), then the process appears as a single point of energy storage. Hence, it can be represented as a single first-order lag with minimal dead time. The controller tuning should be high gain (narrow proportional band), with perhaps no integral action, since the offset with load changes will be essentially negligible. Derivative action is not required.

Many gas-pressure loops, either reducing stations or back-pressure regulating stations, are equipped with self-contained pressure regulators that include the sensor, controller, and final control device. The controllers are simply high-gain amplifiers without integral action. Familiar examples of this include air sets that reduce instrument air from the header pressure down

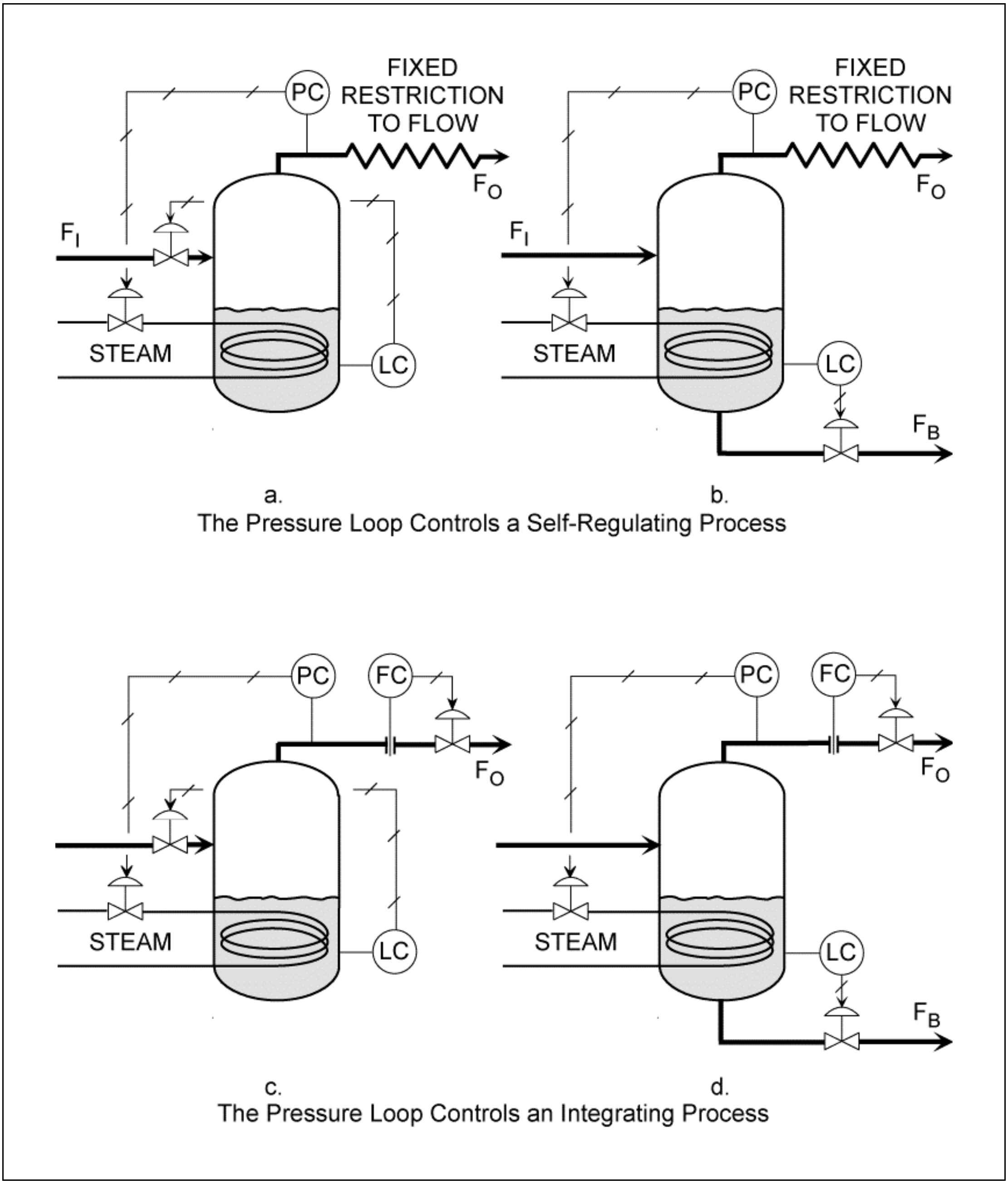


Figure 3-23. Elementary Examples of Pressure Control of a Boiling Liquid

to 20 psig to serve a particular instrument, and gas-pressure regulators for both domestic and industrial service.

Although the control of gas pressure has been described as (usually) being relatively easy, there are some *caveats*. If the upstream absolute pressure is greater than approximately 50 percent of the differential pressure across the valve, then the flow through the valve will resemble the "choked nozzle" effect. In this situation, the flow rate is a function of the upstream pressure and the valve position (area of opening) only. The process gain will be directly proportional to pressure. If the variation in pressure set point is wide, then the controller will have to be retuned at lower upstream pressures.

Another item of interest in the control of gas pressure relates to applications in which the controlled piping network is lengthy, such as in gas-transmission pipelines. Here, it definitely cannot be said that the process can be represented by a model in which all the energy is stored at one point. Because of the distributed nature of the energy storage, the controllers will be tuned with a lower gain than for in-plant pressure control loops. Consequently, integral action will definitely be required.

♦ Liquid-level Control Loops

Level controllers usually control an integrating process. This is because the accumulated liquid level is simply the integral of the difference between inflow and outflow. In actual practice the liquid level is not the driving force in determining either the outflow or inflow rate (despite the fact that we used hydraulic analogies to describe various types of process dynamics in the previous chapter). Thus, as with some types of pressure loops level-control loops must be controlled with a high gain (narrow proportional band), without relying excessively on integral action.

Level-control loops are usually noisy. There can be a random noise due to splashing, if a liquid stream is introduced above the liquid surface. There can also be an oscillatory noise caused by periodic sloshing across the surface of a vessel or by a "U-tube manometer effect" of the liquid oscillating between the interior of a vessel and an external cage level sensor. These oscillations will appear as an oscillation of the process variable itself, when as a matter of fact, the mass (or volumetric) holdup is unchanging. The level measurement signal should be filtered to minimize the appearance of these inherent oscillations.

Normally, liquid-level control is not critical. In fact, it may be more important to maintain an average liquid level over a longer period of time rather than a precise moment-by-moment liquid level. Special control algorithms for accomplishing this are mentioned in chapter 5 on feedback controller modifications. Because level controllers control integrating processes, tuning procedures differ from the procedures applicable to self-regulating processes. Chapter 6 contains a special section on the tuning of liquid-level control loops.

❖ SUMMARY

In this chapter we discussed generic forms of process characteristics, both steady state and dynamic. In discussing steady-state characteristics, we presented the concept of the *process graph* and noted that the position of the process graph changes with disturbances to the process.

In addition, we presented the concept of *process gain* and described nonlinearities arising both from physical and chemical properties of the process and from valve characteristics. We also described the three categories of process dynamic behavior: self-regulating, integrating, and "runaway" processes.

We used both hydraulic analogies and physical examples to illustrate basic forms of open-loop response, including first-order lag; dead time; multiple first-order lags; underdamped processes; inverse-response processes; and multiple-input, multiple-output processes.

We stated that precise models for industrial processes are often intractable. We described approximation techniques that can be used in the design of practical control systems.

Finally, we described general characteristics of flow, temperature, pressure, and liquid-level control loops.

❖ REFERENCES

3-1. H. W. Boger. "Flow Characteristics for Control Valve Installations," *ISA Journal*, November 1966.

3-2. F. G. Shinskey. *Process Control Systems,* 4th ed. McGraw-Hill Publishing Co., 1996.

4

PID CONTROL

❖ FEEDBACK CONTROL

The principle of feedback is one of the most intuitive concepts in process control. An action is taken, more than likely to correct a less-than-satisfactory situation. Then, the results of the action are evaluated. If the situation is not corrected, then further action may be warranted.

After inspecting our young son or daughter's report card, we give the instructions, "You had better improve your grades, else no more TV for you." At the next reporting period, we evaluate the results of our instruction; stronger action may be indicated. As we are driving our vehicle, if we detect that we are drifting out of the center of the lane, we make a slight adjustment of the steering wheel, then observe the effect. If we do not return to the center of our lane, then we make a further adjustment.

These examples, and others that could be provided *ad infinitum*, all involve actions that are carried out without conscious knowledge of the principle of feedback control. The corrective action, and the necessity for evaluating the effect for possibly additional corrective action, is intuitively obvious. Yet automatic feedback control, implemented with self-acting mechanisms, has only been widely utilized for a little more than 250 years. It first began when windmills and especially the early steam engines created the need for a speed-governing mechanism, which was implemented with the invention of the flyball governor (Ref. 4-1).

The study of feedback control as a science is younger than actual attempts at implementation. It was only in the century just past that the now-familiar PID (proportional-integral-derivative) form of feedback controller was developed. Although studies in stability analysis of systems described by linear differential equations were made in the 19th century by Routh and Hurwitz (Ref. 4-1), it was in the years prior to World War II that the theoretical underpinnings of feedback control began to be thoroughly explored (Ref. 4-2 and 4-3).

Feedback control can be classified by the form of the controller output. One of the simplest forms of output is a discrete form, also called on/off or two-position control. A familiar example of this is the household thermostat, which activates a heating unit if the temperature is below the setting or deactivates the unit if the temperature is above the setting. In actual practice, the physics of a thermostat is more complex than this, however. It embodies the principles of hysteresis so as to make the control action less susceptible to "chatter" when the temperature is near to the setting. It also embodies heat anticipation to minimize the swings in ambient temperature that are required to obtain the control action.

Although the field of thermostat construction contains many interesting topics for discussion, discrete control is not central to industrial process control and so will not be pursued further here. Conceptually, however, the idea of two-position control can be extended to multiposition or multistep control. (Commercial air-conditioning refrigeration equipment is operated by loading or unloading compressor cylinders, typically in four steps.)

The ultimate extension of multiposition control would be to have an infinite number of positions. This is called modulating control, an example of which is a process controller whose output can drive a valve to (theoretically) any position between 0 and 100 percent. Commercial process controllers provide an output signal that typically varies between 3 and 15 psig, or 4 and 20 milliamperes, representing 0 to 100 percent signal range.

A variation of modulating control is time-proportioning control. Here, the final control device can assume only two positions, off or on. However, the state is not determined merely by the relationship of the measured variable to the set point ("Are we above or below the temperature setting?"). Instead, the controller output changes state twice during each of a repetitive series of time periods. The fraction of the time in which the output remains in the "on" state is analogous to the output signal of a conventional modulating controller. For example, if a conventional modulating controller would produce a 57 percent output (say, 10.84 psi or 13.12 mA), then an equivalent time-proportioning controller's output would be in the "on" state for 57 percent of the time period and in the "off" state for 43 percent.

A typical application of time-proportioning control is one in which the final control element is an electric heating element, such as for plastics extrusion. The repetitive time period might be 10 seconds. Although the heating element is on for, say, 5.7 seconds and off for 4.3 seconds in every 10-second period, the mass of the extruder barrel acts to average out the heat input over time. As a result, the effect is approximately the same as if 57 percent of the power level were being applied continuously. (See chapter 16 for a further discussion of time-proportioning control.)

The study of feedback control centers on the use of modulating control devices, whether the physical output is 3–15 psi, 4–20 mA, time-proportioning, or some other equivalent form.

♦ MODES OF CONTROL

Conventional feedback controllers use one, two, or three methods to determine the value of the controller output. These methods, called the modes of control, are as follows:

Proportional	(P)
Integral	(I)
Derivative	(D)

In general, these modes can be used singly or in combination. The following lists all possible combinations, with a rough indication of their frequency of use in actual practice:

P	Sometimes used;
PI	By far the most often used;
PID	Sometimes used—probably should be used more than it is;
I	Used in special circumstances where the proportional mode must be avoided;
PD	Used only in rare applications;
ID	May be used in lieu of the I-only mode;
D	Never used alone.

This list shows that the combinations of P, PI, and PID cover almost all the actual feedback controller applications, with the PI combination being the most prevalent. Thus, although the P combination (proportional-only) is not important numerically, we will discuss it first, since a thorough understanding of the proportional mode will serve as the basis for understanding other modes and combinations.

Proportional Mode

With a controller containing only the proportional mode, the controller output, hence the position of the final control element, is proportional to the measurement value only. The proportional mode utilizes no history of the measurement value, nor considers its rate of change. Adjusting (tuning) the controller for the desired performance is simple since there is essentially only one (or at most, two) adjustment to be made. The proportional controller suffers from a serious deficiency, however: an offset exists between the set point and measurement value under most load conditions.

Figure 4-1 illustrates a proportional control system. The rate of fluid flow into the tank represents the load. To be in equilibrium, the outflow must be the same as the inflow. To achieve the required rate of outflow, the valve must be in a particular position. With the fixed mechanism consisting of a float, pivot, and linkage to the valve, this requires that the level must be in a particular position.

Suppose that the level is exactly where we want it to be, as indicated by the scale and pointer in Figure 4-1a. Then suppose the load (inflow rate into the tank) is increased, say, by opening the inflow valve a bit. With greater inflow, the tank level will rise. And rise it must, since to return to equilibrium the valve must open up to permit a greater rate of outflow. As the level rises, the linkage opens the valve. At a new equilibrium, the outflow and inflow will again be equal, the valve position will be increased, and the tank level will be higher, as shown in Figure 4-lb. The difference between the new and old tank level represents the offset caused by the increased load.

Let us now make this intuitively obvious behavior more precise. The behavior of a proportional controller is given by the following equation:

$$m = K_C e + b \tag{4-1}$$

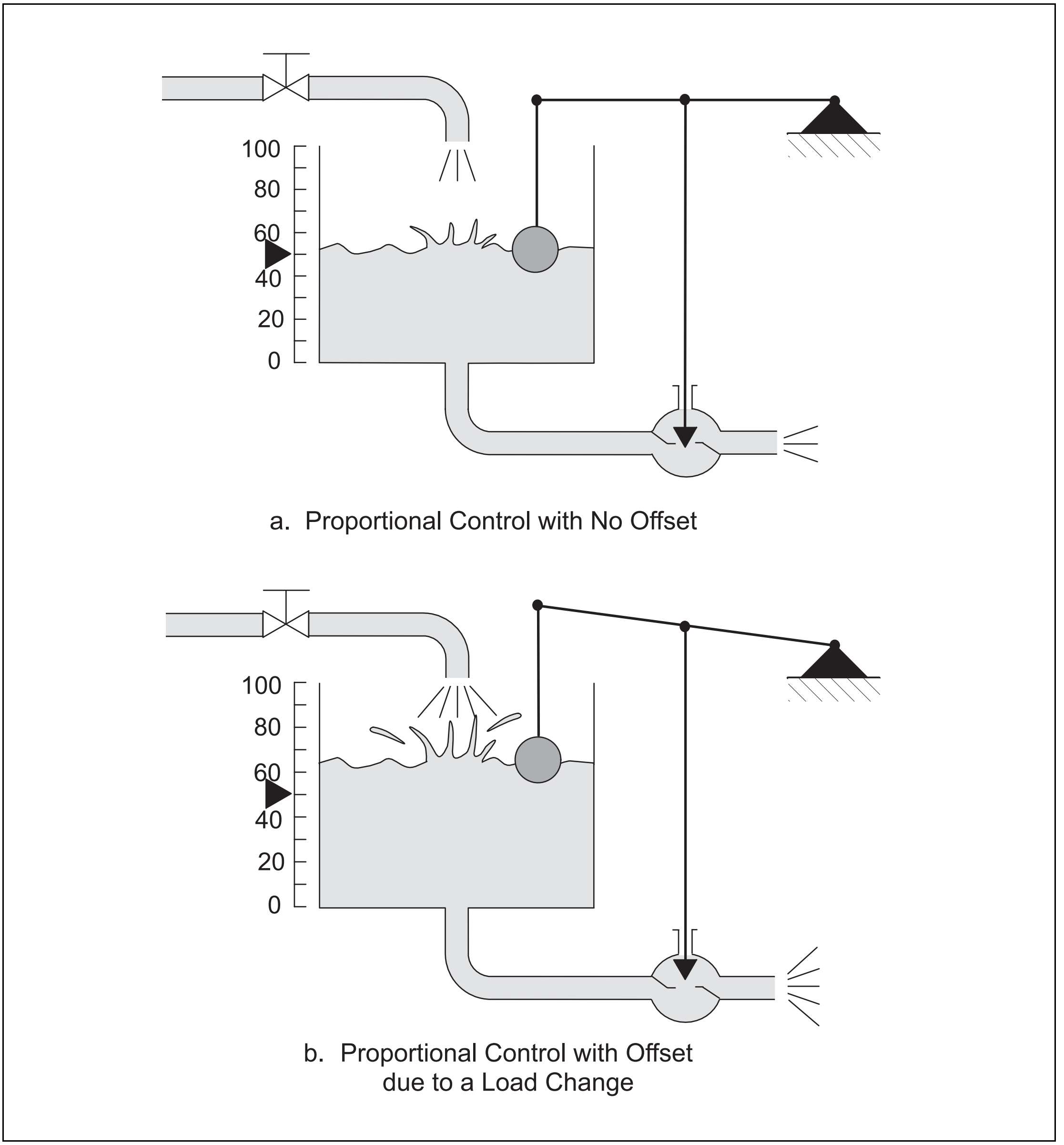


Figure 4-1. A Simple Illustration of Proportional Control

where: m = controller output;
K_C = controller gain;
e = error;
e = SP - PV (if the controller is reverse-acting);
e = PV - SP (if the controller is direct-acting);
b = output bias.

Figure 4-2 shows a block diagram representation of a proportional controller.

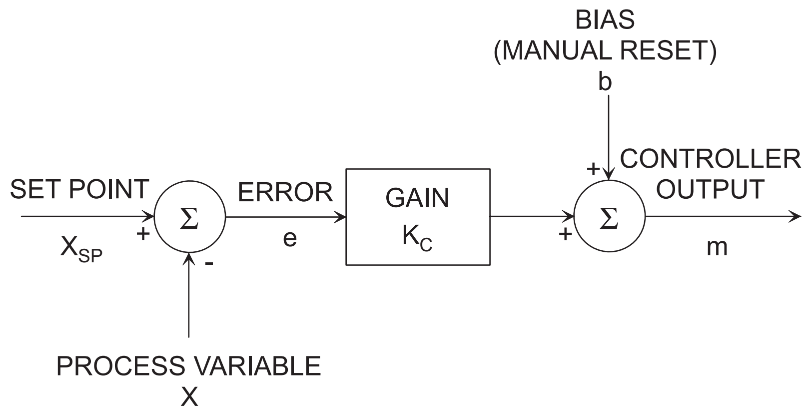


Figure 4-2. Block Diagram of a Proportional Controller

We assume that the values of the process measurement and the set point are converted into percentage of range. As a result, the output is also in percentage terms. In the next chapter, we will consider the case of digital control, where the set point, process variable, and controller output are expressed as real values in engineering units, rather than as a percentage. (The term *process variable* refers here to the input signal to the controller. Normally, this will come from a measurement transmitter. However, at this point, we are talking about the characteristics of the controller by itself, when it is not connected into a control loop. Equation 4-1 and the block diagram of Figure 4-2 are valid even if a dummy process-variable signal were being supplied to the controller from a manually adjusted signal source.)

The output bias term, *b*, provides the value of the controller output when the measurement and set point are equal. A very inexpensive controller might be constructed with a fixed value of *b*, say of 50 percent. Thus, the controller output would be 50 percent when there is no error in the loop. In most industrial-grade proportional-only controllers, the bias is adjustable; this is frequently labeled “manual reset.”

Equation 4-1 provides a linear relation between the process variable and controller output. This relation can be shown graphically, as in Figure 4-3. This figure shows the measurement on the vertical axis and the controller output on the horizontal axis. For the purposes of illustration, an assumed set point value has been plotted along the process-variable scale and an assumed value for output bias along the controller output scale.

Examine this figure and note the following:

- The figure depicts a reverse-acting controller, since an increase in the process variable will cause a decrease in the controller output.

- The position of the graph (line representing the controller input-output relationship) is established by three parameter values:

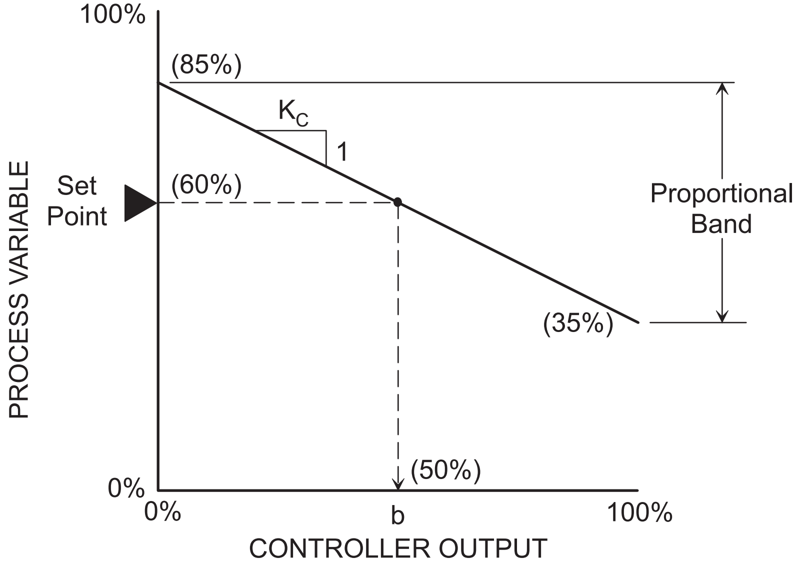


Figure 4-3. Graphical Relationship between Proportional Controller Input (Process Variable) and Controller Output

Set point,
Output bias,
Controller gain.

The set point and bias are coordinates of a point through which the graph must pass; the controller gain determines the slope of the graph.

- The graph in Figure 4-3 clearly indicates that when the process variable and set point are equal, the controller output is equal to the bias value.

- As Figure 4-3 illustrates, if the process variable is 85 percent, the controller output will be 0 percent; if the process variable falls to 35 percent, the controller output will be 100 percent. Thus, a change in input to the controller (change in process variable) of 50 percent will cause an output change of 100 percent. The graph depicts a controller with a gain of 2.0.

- If the gain is increased, say to 4.0, then only a 25 percent change in process variable is required to produce an output change of 100 percent. Graphically, a controller with a higher gain would be depicted by a line with a lesser slope. The reason a lesser slope represents a higher gain is that the independent variable—the input to the controller—is shown on the vertical grid, and the dependent variable—the controller output—is shown on the horizontal grid. This is the opposite of the usual way in which the dependent and independent variables are depicted. The reason for this depiction will become clear later in this chapter.

The amount by which the process variable must change in order to cause 100 percent change in controller output is called the *proportional band* (PB). (In Figure 4-3 this is 50%.) The relationship between controller gain and proportional band is given by the following:

$$PB = \frac{100}{K_C} \qquad K_C = \frac{100}{PB} \tag{4-2}$$

Among commercially available controllers, one may find either a proportional-band or a gain-adjustment knob as means for tuning the proportional mode of the controller. Some microprocessor-based systems permit the user to configure the system to display either proportional band or gain.

Now that we have established a graphical relationship between the controller input and output, let's review the relationship between the process input and output. In chapter 3, we presented the concept of the process graph as the steady-state relationship between the process input (signal to valve) and output (measurement) for a particular load condition. Figure 3-4 in that chapter provided an example of this. Note that when the loop is closed, the controller output is equivalent to the process input and the process output is equivalent to the controller input. Since the horizontal and vertical scales of Figures 3-4 and 4-3 are compatible, we can superimpose one graph upon the other, as shown in Figure 4-4.

This figure shows the graph of two entities, the process and the controller. The input of one is the output of the other. The only point at which both graphical relationships can be satisfied simultaneously is at the intersection of the two graphs. Thus, this is the equilibrium point of the control loop for the particular load condition used in plotting the process graph.

Figure 4-4 depicts the fortuitous circumstance in which the intersection of the two graphs coincides with the set point. Thus, at the load condition depicted, there will be no steady-state offset between measurement and set point.

Suppose there is a load change, as shown in Figure 4-5. This will cause a shift in the process graph but no change in the controller graph, since none of the three controller parameters (set point, output bias, and controller gain) have changed. (Note that we didn't say there would be no change in controller output.) The control loop will come to equilibrium at the new intersection between the controller and process graphs, but due to the load change, this new point of equilibrium is no longer at the set point.

This figure graphically depicts the same phenomena as Figures 4-la and 4-lb. With a proportional-only controller, at one particular load the set point and measurement will be equal. At any other load, there will be a steady-state offset.

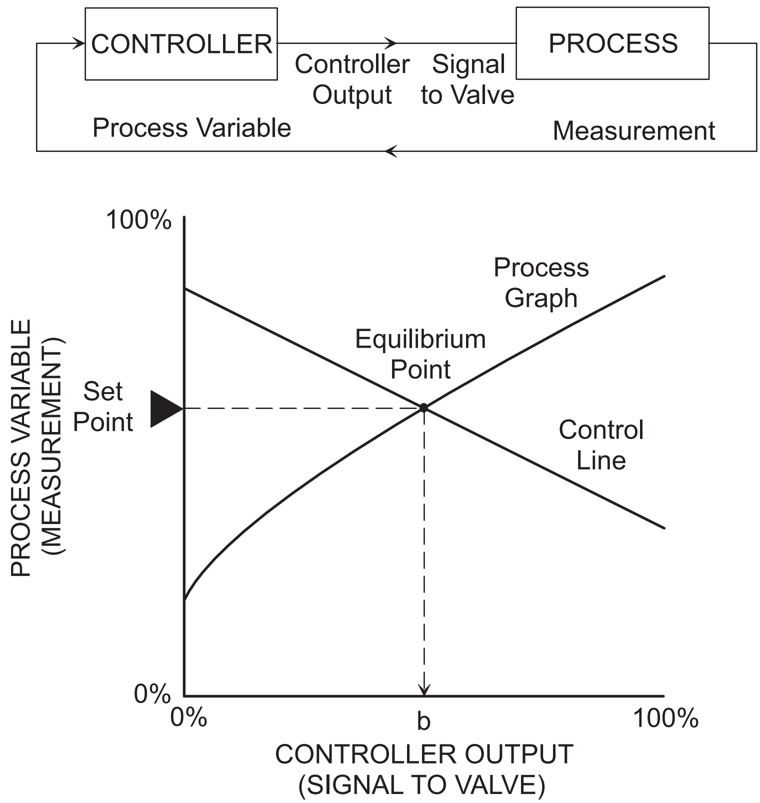


Figure 4-4. Proportional Control Graphical Relationship When Equilibrium Point Coincides with Set Point

Now that we have established the problem with the proportional-only controller, let's explore possible means of solving it. We stated earlier that the position of the controller graph was established by three parameters: controller gain, set point, and output bias. We will now explore, one at a time, the effect of adjusting each of these parameters.

Gain Adjustment

Figure 4-6 shows the effect of a gain adjustment, specifically a gain increase. Graphically, this causes the controller graph to pivot about the point that is determined by the coordinates of the other two parameters. Obviously, the steady-state offset is decreased.

Unfortunately, the amount by which the gain can be increased is limited. That limit is determined by the dynamics of the process. If the gain is increased excessively, then the loop will begin to oscillate, perhaps out of control. Thus, while increasing the gain is a step in the right direction, it is not a generally useful solution to the problem of eliminating the steady-state offset for any load condition.

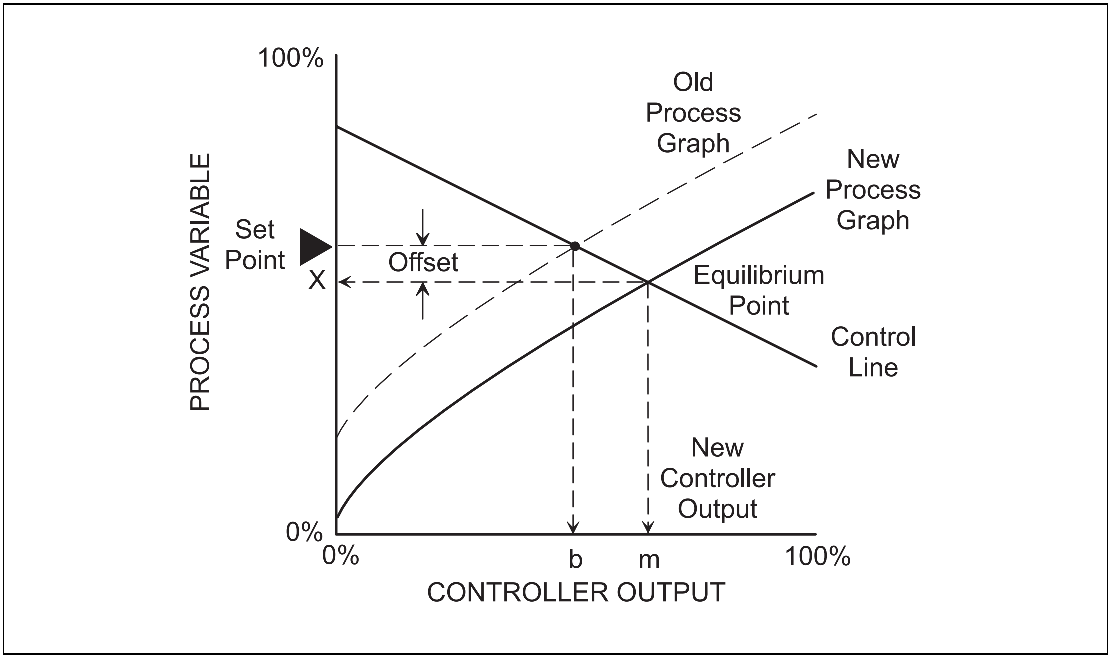


Figure 4-5. The Effect of a Load Change on a Proportional Control Loop

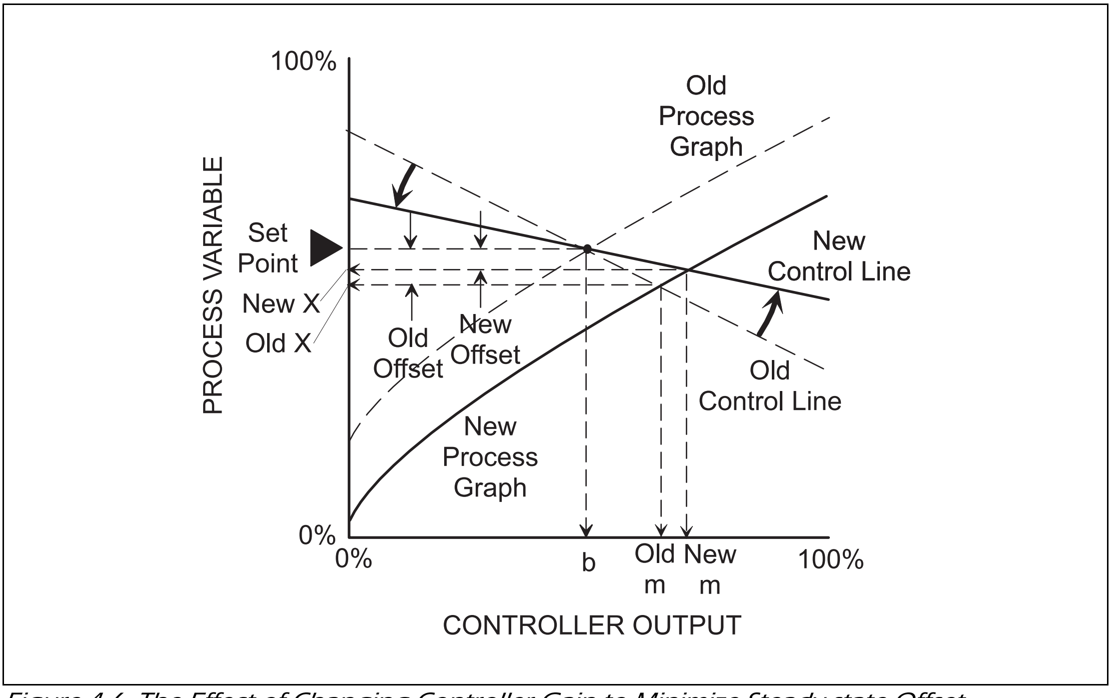


Figure 4-6. The Effect of Changing Controller Gain to Minimize Steady-state Offset

Set point Adjustment

We return to Figure 4-5, but this time we change the set point. (Analogy: If you set your thermostat for 75°F [23.9°C], but the thermostat maintains the room at a constant 70°F [21.1°C], you would probably raise the setting to 80°F [25.6°C], assuming that the thermostat is going to control the room to 5° lower than the setting.) Figure 4-7 shows that raising the set point raises the point determined by the coordinates of the set point and controller output bias; hence, the entire controller graph is raised. If it is raised by just the right amount, then the controller and process graphs will again intersect at the desired operating point (the original set point, not the "fake" set point just entered).

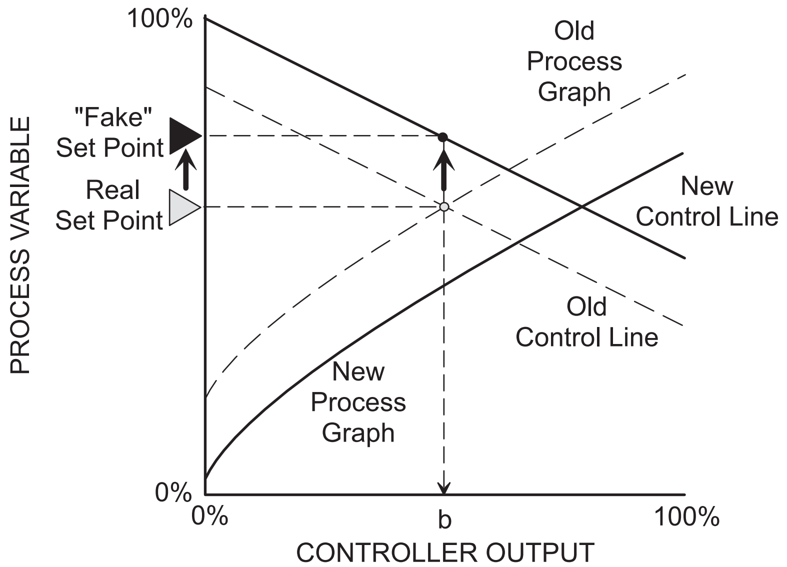


Figure 4-7. The Effect of Changing Set Point to Eliminate Steady-state Offset

This will eliminate the steady-state offset, until the next load change!

Output Bias Adjustment

We can explore the effect of one more parameter adjustment, the output bias. With the steady-state offset depicted in Figure 4-5, there is a positive error. (Since the depicted controller is reverse-acting, then *SP - PV* is greater than zero.) Increasing the output bias will move "point" to the right and then cause the entire controller graph to move to the right (see Figure 4-8). If it is moved by the correct amount, the intersection of the controller and process graphs will again coincide with the set point.

Again, this will eliminate the steady-state offset, until the next load change!

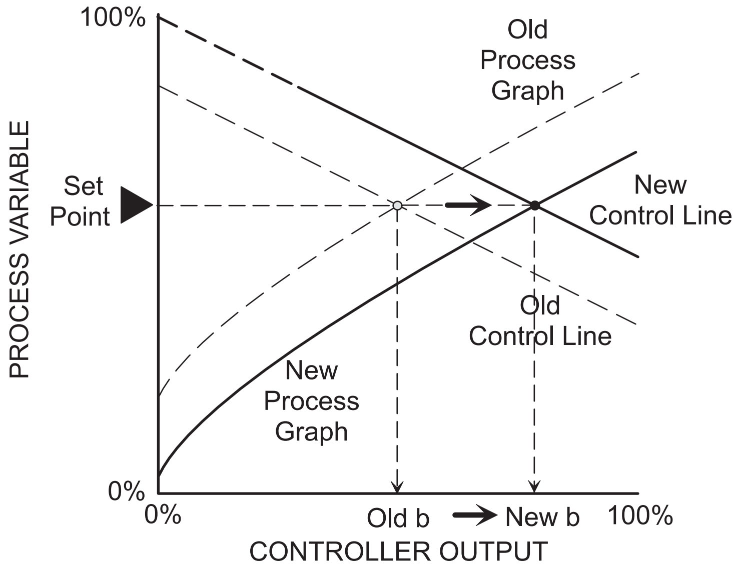


Figure 4-8. The Effect of Changing Output Bias to Eliminate Steady-state Offset

It appears that we are not progressing. The only advantage of the latter adjustment (changing the bias) over the former (changing the set point) appears to be that we maintain the set point at our true desired operating point. We have, however, developed a concept that we wish to pursue. If the error is positive, we can increase the bias. If the error is negative, we decrease the bias; if the error is zero, we leave the bias unchanged. A little reflection also tells us that we should change the bias at a rate that is proportional to the magnitude of the error.

Now, suppose we build a mechanism into our controller or into our software system that will automate the procedure just described—increase or decrease the bias at a rate that is proportional to the magnitude of the error. Increase it if the error is positive, decrease if the error is negative. The graphical effect is that as the process load changes—shifting the process graph—the controller graph is shifted laterally, always attempting to keep the intersection at the set point.

Integral Mode

An integrator is the ideal device for automating the procedure for adjusting the controller output bias. When we considered the output bias to be manually adjusted, we called it "manual reset." We will now set the bias automatically by the output of the integrator; hence, we will call it "automatic reset." Often this term is shortened to merely "reset."

Figure 4-9 shows this implementation. Obviously, if the input to the integrator, that is, the control loop error, is positive (negative), then the integrator output and the controller output will shift upward (downward). Only if the error is zero will the integrator output, as well as the controller output, be stationary.

Although we have established the concept of automatic reset, we will now make two modifications to Figure 4-9 to be more in agreement with commercially available process controllers. The first modification is to use the product of gain-times-error as the input to the integrator rather than the error itself. Since the controller gain is always a positive number, we have not violated our criterion: that when there is a positive (negative) error, the controller output should be increased (decreased); and if the error is zero, then the controller output should be maintained stationary. (One of the PID modifications presented in the next chapter presents an alternative formulation.)

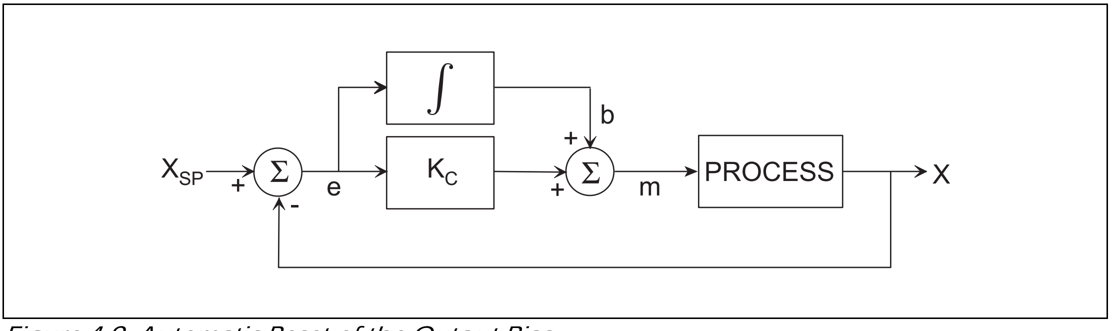


Figure 4-9. Automatic Reset of the Output Bias

The second modification is to place a parameter, in the form of $1/T_I$, in series with the integrator, so the effect of the integral mode can be adjusted. Both of these modifications are shown in the following equation and are incorporated into the PI (proportional-plus-integral) controller block diagram shown in Figure 4-10.

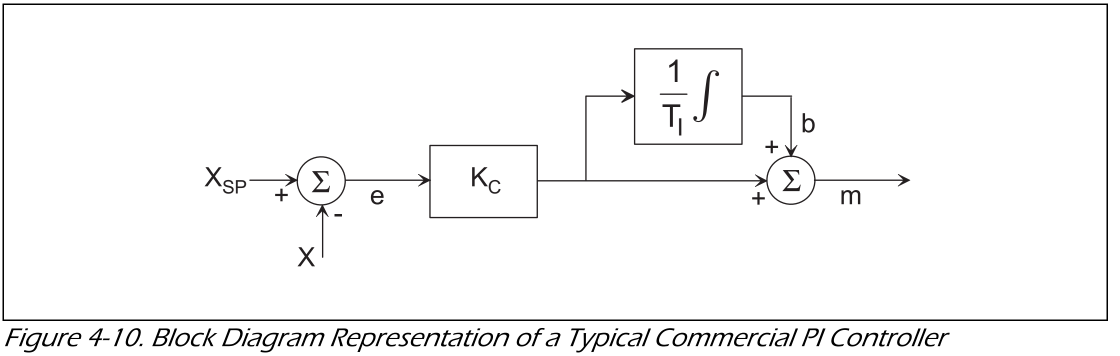


Figure 4-10. Block Diagram Representation of a Typical Commercial PI Controller

The following equation represents the PI controller mathematically:

$$m = K_C\left(e + \frac{1}{T_I}\int e\,dt\right), \tag{4-3}$$

where: T_I = integral time, minutes per repeat.

A corresponding equation in Laplace notation is as follows:

$$M = K_C\left(1 + \frac{1}{T_I s}\right)E. \tag{4-4}$$

Before continuing, let's assign a meaning to the parameter T_I. We will do this by means of a "thought" experiment. Suppose we have a PI controller sitting on a workbench. It receives its measurement from a signal generator while its output is merely recorded on a chart recorder. This is obviously an open-loop setup. We have described this explicitly to emphasize the fact that the definition that follows is based on an open-loop phenomena, not a closed loop.

Suppose we begin our experiment by adjusting the controller set point and the signal generator that supplies the measurement to be the same value. The controller will have zero error and the controller output will be stationary; let's assume that it is at some mid-range value on the chart. We also set the controller tuning parameters at some reasonable values. We assume that there are two controller tuning parameters, one for controller gain and the other labeled "reset" for adjusting T_I.

From this starting point, we make a step change in set point. This makes a step change in error, hence a step change to the input of the integrator.

At the instant of change, the proportional mode contributes a step change to the controller output. The magnitude of this change is the controller gain times the change in error. This is called the proportional response of the controller output.

After the change, the error persists at the same value; the controller doesn't correct for the error, since it is in an open loop. Also, immediately after the change occurs, the integrator has a nonzero input (and constant input, if we make no further changes in the set point or measurement). Hence, the integrator output begins ramping (changing gradually) at a rate that is determined by the amount of error signal and by the value of T_I. This integral response is added to the proportional response of the controller output, as shown in Figure 4-11.

Suppose we ask ourselves, "How long will it take for the integral response to cause an additional output change that is equal in magnitude to the proportional response?" If we time this, we see that the time it takes for the integral action to repeat the proportional response is equal to our setting for the parameter T_I. Hence, the usual units of T_I are *minutes per repeat,*[1] *meaning the minutes required for the integral response to repeat the proportional response (in the open loop).*

If we want a slow integral response, we should set a large number for T_I. If we want a fast response, we set T_I to a small number. Intuitively, however, if we want a fast response, we would like to set on a dial—or enter—a large number; if we want a slow response, we would

1. To be technically correct, the units for T_I are minutes. Common terminology, however, uses minutes per repeat.

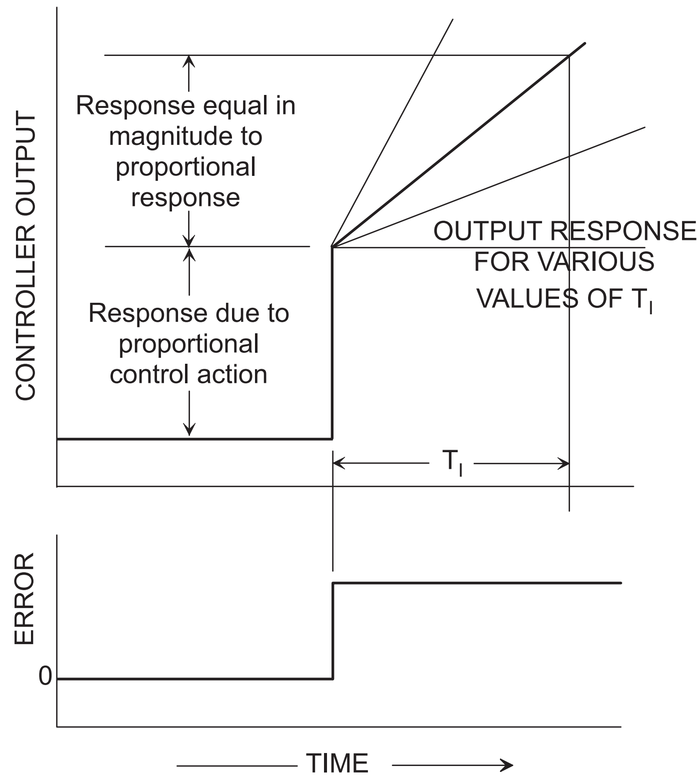


Figure 4-11. The Significance of T_I

like to enter a small number. For this reason, many manufacturers calibrate the integral-mode adjusting dial as $1/T_I$, not T_I. The units of $1/T_I$ are usually called *repeats per minute*, not minutes per repeat.

Among commercially available controllers, integral-mode adjustment knobs are found calibrated in both "minutes per repeat" and "repeats per minute." Some microprocessor-based systems permit the user to configure the system to display either minutes per repeat or repeats per minute. Also, some system use "seconds" as the time basis rather than minutes.

Let us make one other observation. Earlier in this chapter (in Figure 4-3), we noted the relationship of the controller input-output graph and the proportional band. We noted that the width of the proportional band is determined by the controller gain and that the position of the PB is determined by a combination of values of set point and output bias (see Figure 4-3). Since the integral contribution to the controller output is the output bias, then as the integrator output changes, the *position* (but not the width) of the proportional band must also change. This is illustrated by Figure 4-12, which shows a strip-chart recorder in which the process-variable scale is indicated on the left-hand edge. Overlaying this is a series of lines that indicate the output of a controller with a 50 percent PB. Initially, the process requires a 50 percent controller output to bring the PV to the set point. Then, there is evidently a decrease in load on the process because, later on, only a 30 percent valve position is required for the PV to equal the set point. Thus, the entire proportional band has been shifted downward by 10 percent of the PV scale. (A PB setting of 50 percent is equivalent to a controller gain of 2; hence, a shift in the position of the proportional band by 10 percent will cause a decrease in controller output of 20 percent.)

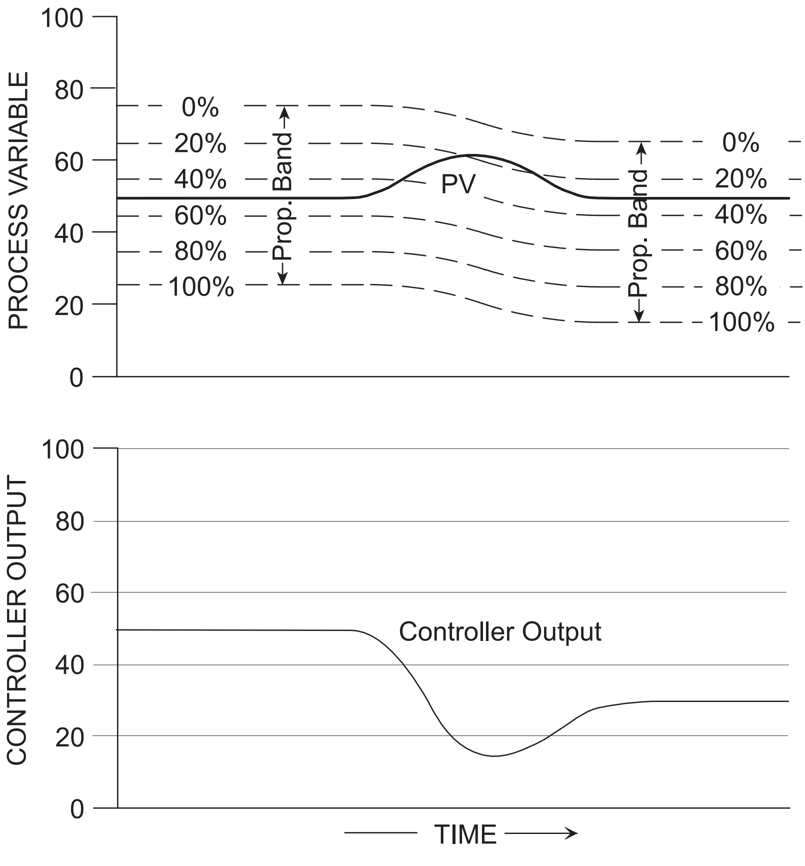


Figure 4-12. Shifting of the Proportional Band As a Result of Integral Action

We have now established the form for the proportional-plus-integral controller. Its primary virtue is that it automatically eliminates steady-state offset in the presence of load changes, set point changes or changes in tuning. For this reason, PI is the most widely used form of control in actual process control applications. To obtain an acceptable performance, however, we do have two parameters to adjust, rather than the one (gain) we had with the proportional-only controller.

Derivative Mode

Now that we have provided for the elimination of the steady-state offset, let's consider an enhancement for our control-loop performance. By adding a component to the controller output that is proportional to the *rate of change* of the measurement, we can anticipate the effect of load changes, thereby reducing the total amount of deviation.

Equation 4-5 and Figure 4-13 show the derivative mode added to the previously established proportional and integral modes. The contribution of the derivative mode to the controller output is based upon the rate of change (derivative) of the product of controller gain times error. The tuning parameter, T_D, allows us to adjust the relative effect of this mode of control. The mathematical representation of the PID is

$$m = K_C\left(e + \frac{1}{T_I}\int e\,dt + T_D\frac{de}{dt}\right), \tag{4-5}$$

where: T_D = derivative time, minutes.

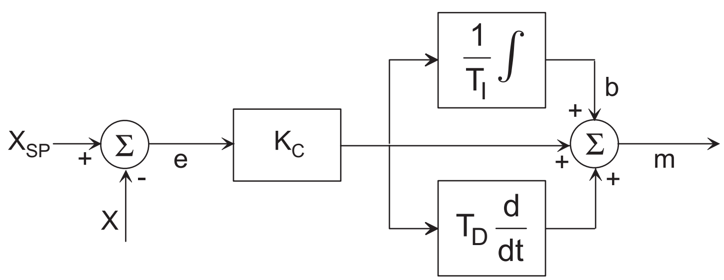


Figure 4-13. Block Diagram Representation of an "Ideal" PID Controller

An equation corresponding to Equation 4-5 but in Laplace notation is:

$$M = K_C\left(1 + \frac{1}{T_I\,s} + T_D\,s\right)E \tag{4-6}$$

Just as we performed a "thought" experiment to give a meaning to the parameter T_I, we will perform a similar experiment to give a meaning to the parameter T_D. In the previous experiment, our controller contained only the proportional and integral modes. In this experiment, our controller will contain only the proportional and derivative modes.

Our equipment setup is similar to that used before. The controller measurement is supplied by a ramp generator, which has a ramp-adjusting rate and an on/off switch. The controller output is connected to a chart recorder.

For our first experiment, we start with the ramp generator off, T_D set to zero (note that with that setting, we are utilizing only the proportional mode), and the controller output set at a mid-range position on the chart. It is not necessary to have zero error, since we have removed the integral mode from our controller.

When the ramp generator is turned on, a ramp form of error signal is produced. This, in turn, with proportional control only, generates a ramp form of controller output. (Since the controller is open loop, the controller output doesn't affect the error.) When the ramp generator is turned off, both the error and the output revert to a constant value (see Figure 4-14).

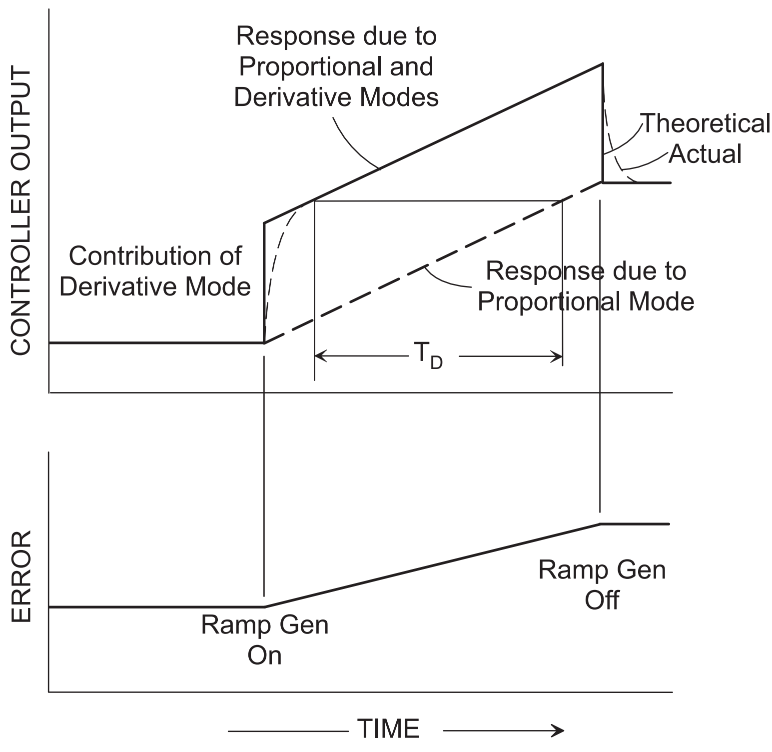


Figure 4-14. The Significance of T_D

Now, return to the same initial conditions and repeat the experiment, this time with a positive value set for T_D. When the ramp generator is turned on, the proportional mode contribution to the controller output is the same as before. Additionally, there is the derivative mode contribution; this is constant since the rate of change of the measurement is constant. When the ramp generator is turned off, the controller measurement stops changing, so the contribution of derivative mode to the output goes to zero. These responses are all shown on Figure 4-14.

The contribution of the derivative mode to the controller output is as follows:

$$K_C T_D \frac{de}{dt} \ .$$

If the set point is not changing (and in this thought experiment it is not), then this is equivalent to the following:

$$-K_C T_D \frac{dx}{dt} \qquad \text{for reverse-acting controllers,}$$

or

$$K_C T_D \frac{dx}{dt} \qquad \text{for direct-acting controllers.}$$

where x is the measurement value. Note that the derivative contribution is always in a direction that will arrest the direction and rate of movement of the measurement.

This statement does not provide as much insight, however, as the following fact: with proportional-plus-derivative control, the controller output (e.g., the valve position) leads the movement it would otherwise have had with proportional control alone by T_D minutes. This leading action provides a faster response to load upsets in loops where it can be used. This is one of the virtues of the derivative mode.

Another interpretation of the effect of the derivative mode is illustrated by Figure 4-15. This figure shows the chart record of a process variable that is changing at a constant rate. If proportional control only were used, then at any given time we would use the present value of the measurement to determine the present value of the controller output. However, if we used proportional-plus-derivative control, the controller projects ahead by T_D minutes and determines the predicted value of the measurement T_D minutes from now. This predicted value of measurement is used to compute the controller output, rather then the present measurement value.

Employing the present rate of change, the predicted value of the measurement T_D minutes from now is:

$$\hat{x}\left(t + T_D\right) = x\left(t\right) + T_D \frac{dx(t)}{dt} .$$

An equation for a PD controller (recall that this illustration does not use the integral mode) is:

$$m = K_C \left(e + T_D \frac{de}{dt} \right) + b \qquad (4\text{-}7)$$

(Compare Equation 4-7 with Equation 4-1 for a proportional-only controller.) Since, for a reverse-acting controller, e is given by $e = x_{SP} - x$, then

$$\begin{aligned} m &= K_C\left(x_{SP} - x - T_D\frac{dx}{dt}\right) + b \\ &= K_C\left(x_{SP} - \hat{x}(t+T_D)\right) + b. \end{aligned}$$

This illustration clearly demonstrates the predictive nature of the derivative mode. A bit of reflection will show that the two illustrations depicted by Figures 4-14 and 4-15 are equivalent.

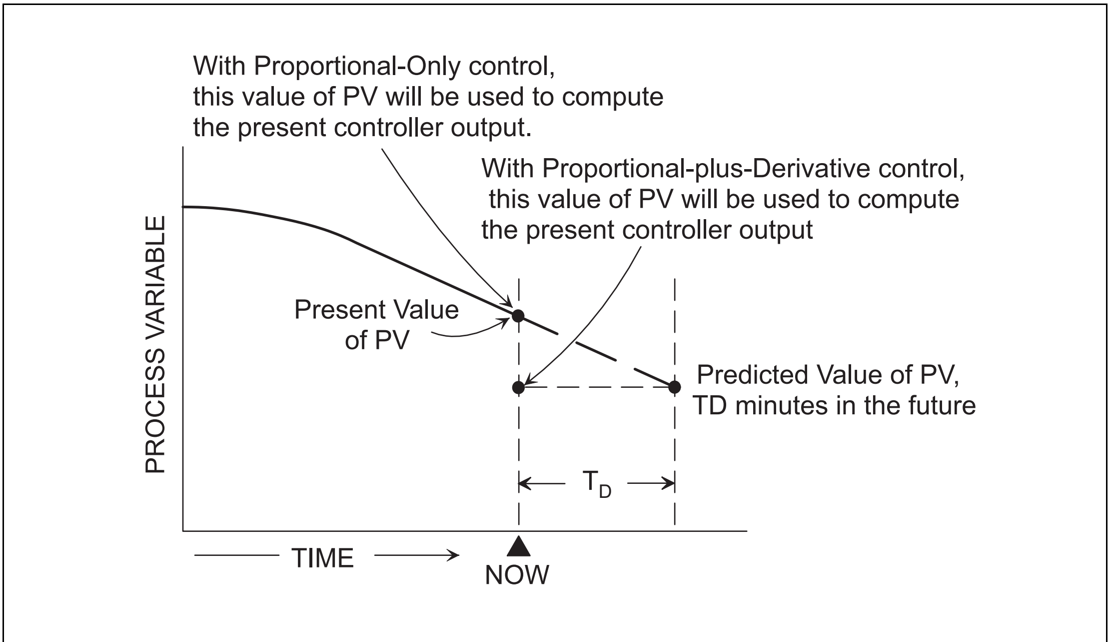


Figure 4-15. The Predictive Nature of Derivative

We first began our discussion of the derivative mode by stating that, in contrast with the integral mode—which is essential to eliminate steady-state offset—the derivative mode only enhances, or improves, the behavior of a control loop. A logical question to ask is, "Where would the derivative mode be used?" Rather than ask that question directly, let's ask, "Where should the derivative mode *not* be used?" We can then gain an answer to the first question by a process of elimination.

Derivative should not be used in control loops in which there is an excessive amount of noise on the measurement. This is because the predictive nature of derivative would amplify the noise and cause an excessive amount of noise on the controller output. This will eliminate most flow and level loops, and possibly others as well.

Plants probably do not need to use derivative to speed up the response of loops that already have a relatively fast response. This fact will eliminate flow loops (again), gas-pressure loops, and possibly others from consideration for derivative control.

By the process of elimination, we arrive at temperature loops, or loops that have relatively noise-free and relatively slow response, as the primary candidates for the use of derivative. These likely include temperature and composition loops, but may also include other types of loops with similar attributes. With open-loop unstable processes such as exothermic reactors, the temperature controller will practically always use the derivative mode to stabilize the otherwise unstable process.

When process control systems were implemented with only analog controllers, either pneumatic or electronic, the specification of the derivative mode, in addition to the proportional and integral modes, represented an increase in the initial cost of the instrument. Hence, one purchased only the modes that were likely to be used, and this often did not include the derivative mode. Once the instrument had been purchased and installed, the question no longer arose, “Should we use the derivative mode?” It was not there to be used.

With the current microprocessor-based controllers, the manufacturer always provides the control algorithm complete with all three modes. If derivative is not required, T_D is merely set to zero. But the present technology offers the opportunity to use derivative much more than in former times. Derivative control is probably an underutilized technology that could be used in many applications to improve the performance of many control loops.

❖ SUMMARY

In this chapter, we have explained the form of the primary control modes for feedback controllers and have provided an intuitive feel for their behavior. We have indicated where combinations of these modes may be used, and we have defined the terms used to describe the tuning parameters. We have not described procedures for tuning a feedback controller—that will come in chapter 6.

The culmination of this chapter are Equation 4-5 and its Laplace transform equivalent, Equation 4-6, which describes the full PID controller. This is the classical, or “textbook” (see Ref. 4-4), form of the PID equation. It could also be described as the “ideal” PID—not “ideal” in the sense of “perfect,” but meaning the standard against which other forms of PID are compared.[2]

If you were to shop for a commercial process controller with just this form of control action, or if you were to examine the control algorithm library of a typical distributed control system, you probably would not find exactly this form. The manufacturers offer many modifications.

2. Some manufacturers, as well as some authors, refer to the PID form represented by Equations 4-5 and 4-6 as the “ISA” form. This author believes this is erroneous usage, since ISA has never sanctioned any particular form for the PID algorithm.

Often, understanding and properly using such modifications is the key to the success of a control system application. The following chapter describes many of the PID modifications that are available in the commercial world.

Table 4-1. Summary of Feedback Control Modes

Mode	Common Name(s)	Tuning Parameter	Application
Proportional	Proportional	Gain, K_C or Prop. Band, PB	Used when: Simple form of control is desired, load does not change significantly, or offset is acceptable. Also used when the control loop dynamics permit a relatively high gain to be set without causing excessive oscillation. Then, even if significant load changes are present, offset is only minimal.
Integral	Reset Automatic Reset	Min./Repeat, T_I or Repeats/Min. $1/T_I$	Used almost always in conjunction with proportional mode to eliminate steady-state offset. Occasionally used alone; known as integral controller. For most applications, I-only controller provides inferior performance when compared with PI modes.
Derivative	Rate Action Pre-Act	Deriv. Time, T_D	Used usually in combination with P and I modes to improve loop performance by anticipating the effect of load changes. Used mainly on temperature loops or other loops that have similar characteristics (low noise level, fairly slow response).

❖ REFERENCES

4-1. S. Bennett. *A History of Control Engineering, 1800 ~ 1930.* Institution of Electrical Engineers and Peter Peregrinius Ltd., 1979.

4-2. Nyquist, H., "Regeneration Theory." *Bell System Tech J.*, 1932.

4-3. Bode, H. W., *Network Analysis and Feedback Amplifier Design*, D. Van Nostrand Col, New York, 1945.

4-4. K. J. Åström and T. Hägglund. *PID Controllers: Theory, Design and Tuning.* ISA – The Instrumentation, Systems, and Automation Society, 1995.

5

MODIFICATIONS TO STANDARD PID CONTROL

The last chapter concluded with the development of the standard (also called the "textbook," "classical," or "ideal") form of the PID controller. Equation 4-5 presented it as:

$$m = K_C\left(e + \frac{1}{T_I}\int e\,dt + T_D\frac{de}{dt}\right) \tag{5-1}$$

Although it is desirable to have a thorough understanding of both the rationale and the behavior of this standard form, in real applications one often finds the need for subtle deviations from the standard form. Indeed, if one attempted to purchase exactly that form, or choose it from a microprocessor-based system vendor's library of control algorithms, frustration would be the likely result. The reason: few manufacturers offer the standard form of the PID; most offer one or more variations of it, called "configuration options."

Some of the modifications are offered by only a few manufacturers, or even only one. After all, each manufacturer claims to have features that will be an advantage to users of that product. On the other hand, all manufacturers offer many of the modifications in forms that are at least similar to each other. In this chapter we describe some of the more common modifications available, as well as their application or purpose. Several manufacturers' software control forms are reviewed at the end of the chapter. These particular manufacturers were selected to illustrate a range of features; their inclusion here should not be interpreted as an endorsement of their products.

Because of the power and flexibility of the microprocessor, users will encounter most of the modifications, or optional features, to standard PID control in a digital-based system rather than in an analog system. Nevertheless, for clarity of presentation, we will refer to analog elements, gains, integrators, derivative units, as we did in the previous chapter, rather than rely on difference equations which would be a more exact representation of the digital form. Before the end of the chapter, however, we will consider the subject of the digital implementation of PID control.

❖ SET POINT "SOFTENING"

A step change in set point can be a rather harsh disturbance for a control loop. Several modifications are available that "soften" the effect of such a change: derivative mode on measurement, proportional mode on measurement, linear combination of inputs to modes, and set point ramping.

♦ Derivative Mode on Measurement

In control loops where the standard form of the PID (Equation 5-1 and Figure 4-13) is used, a set point change causes a large, but short-duration "spike" in the controller output. This is due to the derivative mode's response to the very rapid change in error. In addition to the derivative spike, there is the proportional response, followed by a gradual change caused by the integral response, until the measurement again achieves equilibrium with the new set point (see Figure 5-1a).

Consider the derivative spike in the controller output. Most likely, the final control element will not be fast enough to respond totally to this rapid change. Furthermore, the process itself will act as a filter and prevent much of this signal from reaching the measurement value. Even so, this spike on the output signal is unwanted because it is probably undesirable to move the valve this rapidly. For instance, we may not wish to cause a thermal shock to heat-exchanger tubes, or we may not wish to upset a reactor catalyst bed.

One of the reasons for adding the derivative mode to the controller was to improve the response of the control loop to a load upset. This can be achieved by making the derivative unit responsive to measurement only, rather than to the error. Doing so will avoid the derivative spike caused by a set point change. A block diagram of this configuration option is shown in Figure 5-2; the response to a set point change is shown in Figure 5-1b. This modification is represented by the following equation:

$$m = K_C\left(e + \frac{1}{T_I}\int e\,dt - T_D\frac{dx}{dt}\right) \qquad (5\text{-}2)$$

Note that the sign of the derivative contribution to the controller output has been changed from "+" to "–" in Figure 5-2. Since the derivative contribution must always act to oppose the direction of motion of the measurement, the derivative contribution must be negative on a load increase for a reverse-acting controller. If Figure 5-2 depicted a direct-acting controller, the sign of the derivative contribution would be changed to "+" and the signs at the summation point for set point and measurement would be reversed. Each figure in this chapter uses a reverse-acting controller as the basis of illustration.

Suppose two controllers are mounted side by side, one with the standard form of the PID, and the other identical except for the derivative mode on measurement rather than on error. If the controllers were controlling identical processes and they were identically tuned, the response to a load upset would be identical. The only difference in the behavior of the controllers would

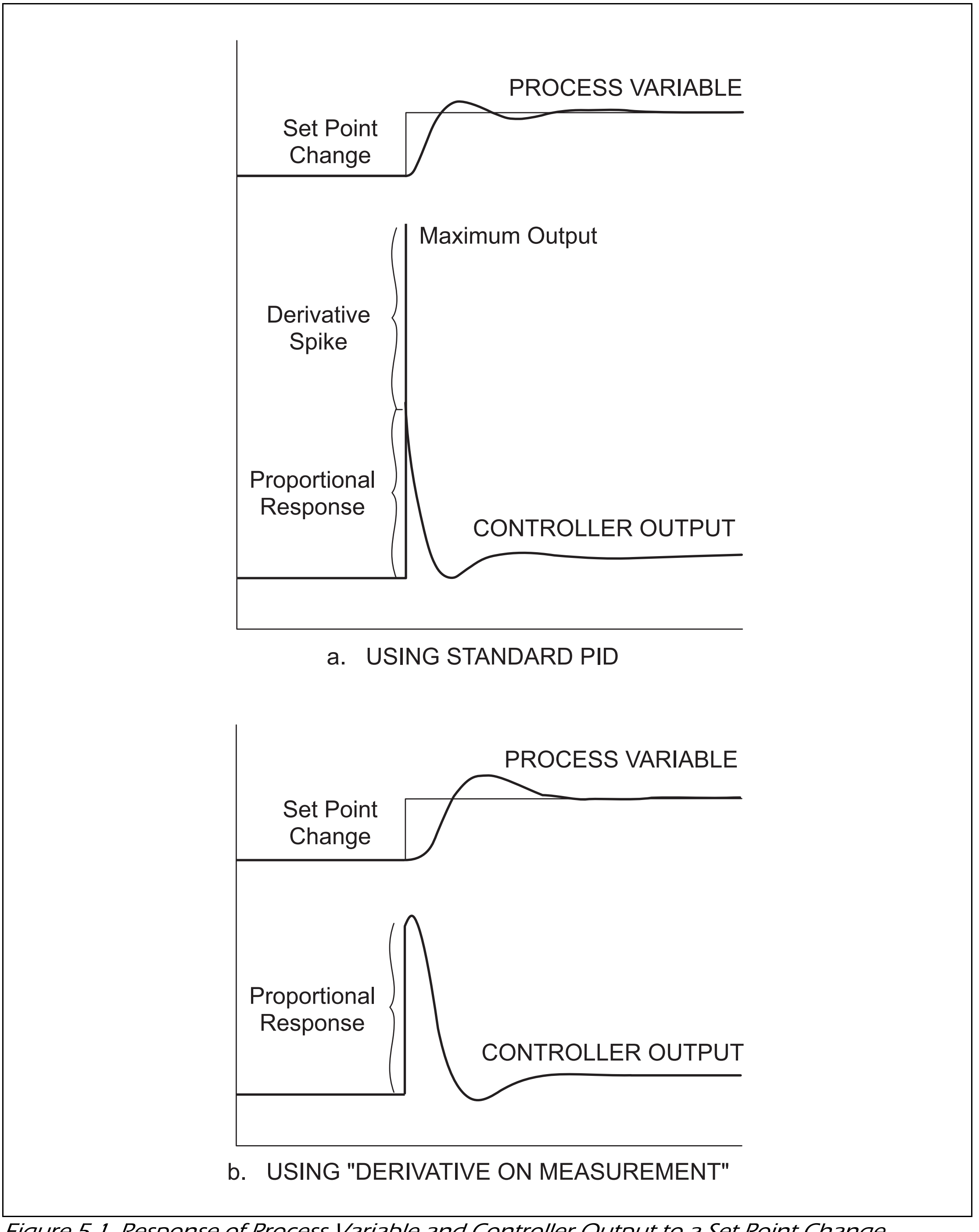


Figure 5-1. Response of Process Variable and Controller Output to a Set Point Change

be the action that follows a set point change: the controller with the standard PID would cause a spike in the controller output whereas the other controller would not.

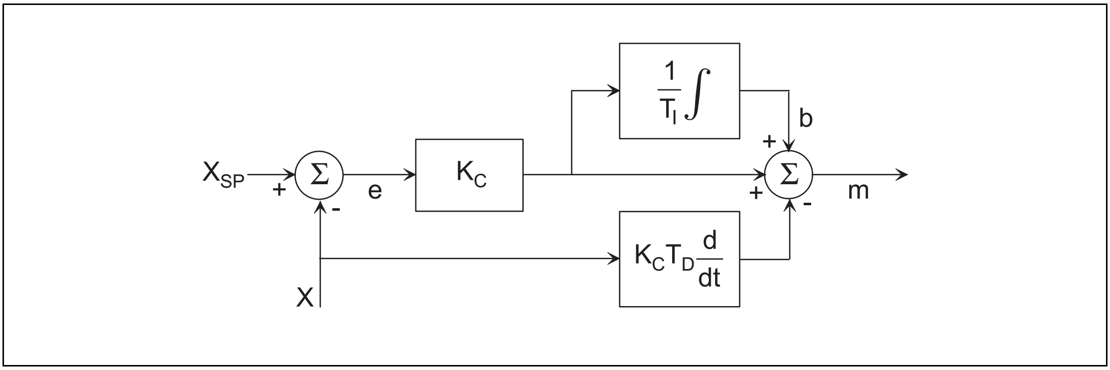


Figure 5-2. Derivative on Measurement

Equipment manufacturers often provide a user configuration option called "derivative on error" or "derivative on measurement," especially in microprocessor-based control systems. Almost always, the better choice from an application viewpoint is "derivative on measurement." Some manufacturers do not give the user this choice; rather, they provide derivative on measurement as the only option. This is not necessarily a deficiency in a controller, however, since "derivative on measurement" is the preferred choice for almost all applications.

A rare instance in which derivative on error would be preferred is when the derivative mode is used in the secondary controller of a cascade loop. (Cascade control is discussed in chapter 9.) For instance, in controlling an exothermic chemical reactor, the set point of a jacket water-temperature controller may be set by the output of the primary reactor temperature controller. Both the primary and the secondary may utilize the derivative mode to provide the fast response and stabilization that the exothermic process requires. In this case, however, the primary controller will not make abrupt changes in the secondary controller set point. For this reason, derivative response to changes in error, whether caused by set point change, measurement change, or both, is acceptable.

♦ Proportional Mode on Measurement

By placing the derivative mode on the measurement, we have eliminated the derivative spike caused by a step change in set point. We still have, however, a step change in controller output, as shown in Figure 5-lb. This is called the *proportional response* or "proportional kick." Even this may be a more abrupt change to the process than is desirable. The proportional response to a set point change can also be eliminated by making the proportional mode, as well as the derivative mode, responsive only to the measurement signal rather than to the error. This is shown in Figure 5-3 and mathematically by Equation 5-3. This leaves only the integral mode acting on the error. The loop response to a set point change is shown in Figure 5-4.

$$m = K_C\left(-x + \frac{1}{T_I}\int e\,dt - T_D\frac{dx}{dt}\right) \qquad (5\text{-}3)$$

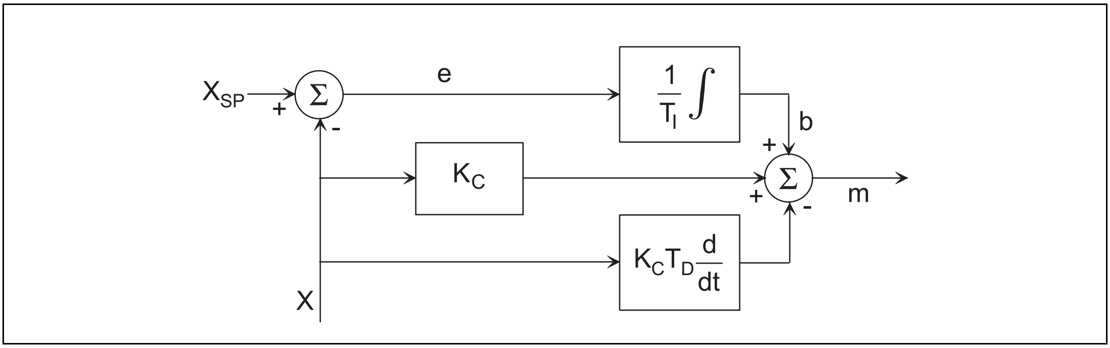


Figure 5-3. Proportional and Derivative Modes on Measurement

Note that in Figure 5-3 the sign of the proportional contribution to the controller output has been changed from "+" to "–" to be consistent with the reverse-acting controller depicted there. For a direct-acting controller, both the sign of the proportional contribution to the output and the signs at the error summation point would be reversed.

Suppose that two controllers are mounted side by side, one with the standard form of PID and the other identical except that the proportional and derivative modes are on measurement rather than error. If the controllers were controlling identical processes and if both controllers were identically tuned, the response to a load upset would be identical. On a set point change, the output of the controller with P and D on measurement will exhibit neither the derivative spike nor the proportional kick. Instead, the integral mode of the controller will cause the controller output to begin a gradual change until the measurement achieves equilibrium with the new set point.

Because of the absences of the initial rapid forcing of the process caused by the proportional response, this controller would probably not drive the measurement to the new set point as quickly as would a standard PID, or even a PID with the derivative on measurement modification. However, this is not necessarily a fault in this controller, since the purpose in using this modification is to be gentler on the process. Because of the slower response to a set point change, this modification should not be used in the secondary controller in a cascade loop. (Cascade control is discussed in chapter 9.)

♦ Linear Combination of Inputs to Modes

The configuration choices represented by the block diagrams in Figures 4-13, 5-2, and 5-3 can be expressed in a comprehensive equation as follows:

$$\begin{aligned} m &= K_C\left(\beta e - (1-\beta)x + \frac{1}{T_I}\int e\,dt + T_D\frac{d}{dt}\big(\gamma e - (1-\gamma)x\big)\right), \\ &= K_C\left((\beta x_{SP} - x) + \frac{1}{T_I}\int e\,dt + T_D\frac{d}{dt}(\gamma x_{SP} - x)\right), \end{aligned} \quad (5\text{-}4)$$

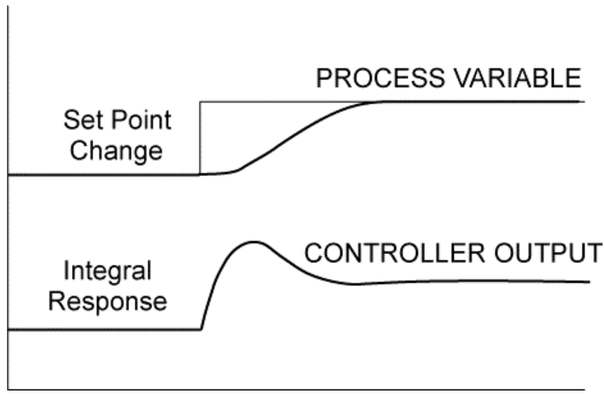


Figure 5-4. Process Variable and Controller Output Response to a Set Point Change with Both Proportional and Derivative Modes on Measurement

where: e = error $(x_{SP} - x)$, for reverse-acting controller,
x = measurement,
x_{SP} = set point,
β = proportional mode selection factor,
$\beta = 1$ for proportional on error,
$\beta = 0$ for proportional on measurement,
γ = derivative mode selection factor,
$\gamma = 1$ for derivative on error,
$\gamma = 0$ for derivative on measurement.

The definitions here for β and γ indicate that they can have values of only 0 or 1. This provides the option of having proportional and derivative modes that are wholly responsive to either the error or the measurement. Suppose that the restrictions on β and γ are relaxed so that either or both may take on any value between 0 and 1, that is:

$$0 \leq \beta \leq 1,$$
$$0 \leq \gamma \leq 1.$$

In this case, one could obtain a linear combination of proportional and derivative modes on error or measurement. (One manufacturer refers to this as a "two degree of freedom" controller.) This variation is probably most useful when only an intermediate value for β is chosen, since γ will usually be left at 0 (providing derivative on measurement). When the value for β is intermediate, the controller can be tuned with a higher value of gain, thus providing a better response to a disturbance. There will also be some proportional action taken on a set point change. As a result, the time available to achieve a new set point will be greater than it would be if all of the proportional action was on measurement.

Functionality that is equivalent to the combination of proportional mode on error or measurement can be obtained by placing a lead-lag filter on the set point. This filter should take the following form:

$$\frac{\beta T_I s + 1}{T_I s + 1},$$

where β is the proportional mode selection factor, as earlier, and T_I is the integral time. One manufacturer refers to this as "set point filtering."

For the same reason that the proportional-on-measurement configuration option should not be used as the secondary controller in cascade loops, this configuration option should also be avoided for that application.

♦ Set Point Ramping

Some controllers permit a step set point change to be made into a new target value. The actual set point used by the controller, however, is changed gradually ("ramped") from its present value to the target value at a specified rate of change. This will result in a more gradual change in the process variable, considerably reducing overshoot.

❖ INTEGRAL-ONLY MODE

Occasionally, it is desirable to eliminate the proportional and derivative modes entirely, utilizing only the integral mode. Even when used alone, the integral mode, sensing the error between set point and measurement, will manipulate the controller output until that error is reduced to zero. For most control loops, however, the proportional mode will reduce the phase lag through the controller caused by the integral mode. Hence, proportional-plus-integral will produce superior performance to integral-only control.

With the PID control algorithm formulated as shown by Figure 4-13 and Equation 5-1, one cannot achieve a true integral-only controller by setting the gain and derivative time to zero since the gain is a common multiplier for all three modes. In a manufacturer's library of software algorithms, a separate algorithm (integral-only) is normally provided. This will have the following form:

$$m = K \int e\,dt \qquad (5\text{-}5)$$

The gain term may be expressed in several different ways, such as K, K_I, or $1/T_I$. Also, some manufacturers may provide an integral-plus-derivative algorithm, in which case it may be converted into an integral-only algorithm simply by setting the derivative tuning parameter to zero.

❖ INTERACTIVE OR NONINTERACTIVE CONTROLLER

Which came first, the chicken or the egg? Commercially available analog controllers, using pneumatic mechanisms that achieved the general objectives of proportional, integral, and derivative control, were developed before a mathematical relationship for the "ideal" PID controller (Equations 4-5 and 5-1) had been formulated. When the working mechanisms were subsequently analyzed mathematically, they did not meet the "ideal" form. Instead, they could be described by the block diagram of Figure 5-5, and by the following equation, written in Laplace notation[1]:

$$M(s) \;=\; \hat{K}_C\left(\frac{(1+\hat{T}_I s)(1+\hat{T}_D s)}{\hat{T}_I s}\right)E(s) \tag{5-6}$$

(The "^" over the symbols indicates the entered value for the tuning parameters.)

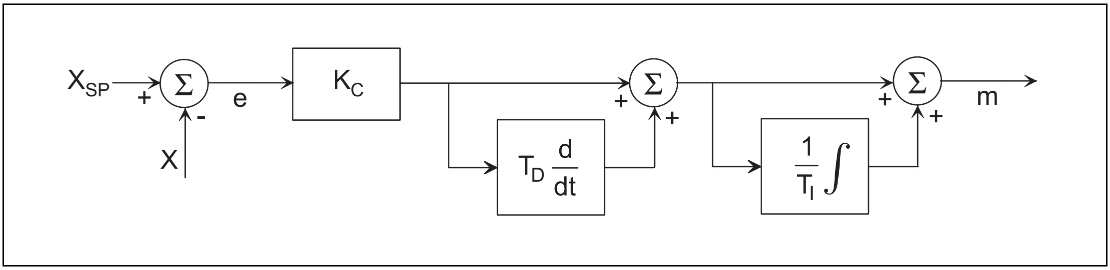


Figure 5-5. Block Diagram of Interactive Controller with All Modes on Error

We will show that, although the mathematical descriptions between the ideal PID and the traditional analog PID controllers differ, by setting the tuning parameters properly, they can be made to behave identically. By some simple manipulations, Equation 5-6 can be reformulated to take the same form as Equation 4-6:

$$\hat{K}_C\left(\frac{\hat{T}_I+\hat{T}_D}{\hat{T}_I}\right)\left(1 + \frac{1}{(\hat{T}_I+\hat{T}_D)s} + \left(\frac{\hat{T}_I\hat{T}_D}{\hat{T}_I+\hat{T}_D}\right)s\right) \tag{5-7}$$

When this is done, the parameters that represent controller gain, integral time, and derivative time no longer have the same meaning as they do with the "ideal" controller. With the controller that is represented by Equations 4-5 and 4-6, the effective controller gain, integral time, and derivative time are not directly set by the entered parameters $\hat{K}_C$, $\hat{T}_I$ and $\hat{T}_D$, but by a combination of these parameters. For example, if the parameter $\hat{T}_D$ is adjusted, it affects the effective value of controller gain, integral time, and time constant. Similarly for the other

1. A further refinement of the mathematical description of commercially available analog controllers adds the effect of a filter to suppress measurement noise. This is discussed later in this chapter.

parameters. For this reason, the controller form that represents the original working mechanisms is often called the *interactive* form, whereas the "ideal" form is often called the *noninteractive* form.[2]

Thus, if one has an interacting controller, the equivalent parameters for a noninteracting controller are given by:

$$K_C = \hat{K}_C\left(\frac{\hat{T}_I + \hat{T}_D}{\hat{T}_I}\right) \tag{5-8}$$

$$T_I = \hat{T}_I + \hat{T}_D \tag{5-9}$$

$$T_D = \frac{\hat{T}_I \hat{T}_D}{\hat{T}_I + \hat{T}_D} \tag{5-10}$$

Conversely, if one has a noninteracting controller, the equivalent parameters for an interacting controller are given by:

$$\hat{K}_C = \lambda K_C \tag{5-11}$$

$$\hat{T}_I = \lambda T_I \tag{5-12}$$

$$\hat{T}_D = \frac{T_D}{\lambda} \tag{5-13}$$

$$\text{where: } \lambda = \frac{1}{2} + \sqrt{\frac{1}{4} - \frac{T_D}{T_I}}\,. \tag{5-14}$$

Example: Suppose we have two controllers, one formulated as a noninteractive controller, the other formulated as an interactive controller. Suppose, further, that the following controller tuning values are entered for each controller:

K_C	2.0,
T_I	8 minutes per repeat,
T_D	2 minutes.

2. Manufacturers do not use the terms *interactive* and *noninteractive* consistently. Later in this chapter, we will present another meaning some manufacturers apply to the term *interactive* when applied to PID. The terms *series* and *parallel* are also sometimes used to describe what we have called interactive and noninteractive controllers. Finally, as we noted in chapter 4, some manufacturers refer to the noninteractive (standard) form as the "ISA" form, although ISA has never endorsed a particular form of the PID control algorithm. Caution is advised when using manufacturer's terminology.

The effective values for the noninteractive controller would be the same values as the entered values. For the interactive controller, however, the effective values would be the following:

Effective K_C 2.5,
Effective T_I 10 minutes/repeat,
Effective T_D 1.6 minutes.

On the other hand, suppose that we maintained the same dial settings for the noninteractive controller, but adjusted the dial settings for the interactive controller using Equations 5-11 – 5-14. The effective controller tuning parameters for both controllers are now the same, even though the entered values are different. In other words, if the controllers were controlling identical processes, we could not discern a difference in their behavior.

To conclude this discussion, note the following points:

(1) If T_D is set to zero, then there is no difference between the interactive and noninteractive forms.

(2) We can enter the effective controller tuning parameters directly if we have used a formal method (such as one of the Ziegler-Nichols methods described in chapter 6) to determine what these tuning parameters should be and if we have a noninteractive controller. If, however, we have an interactive controller, we should calculate the required entry values from the effective values, according to Equations 5-11 – 5-14. (See discussion in the sidebar on page 132 relative to applicability of Ziegler-Nichols tuning relations to noninteractive or interactive controller.)

(3) If we were unaware that we had an interactive controller and entered the required effective values for the dial settings, then the actual effective values produced would be given by Equations 5-8 – 5-10. The error would not be too serious since the actual effective values are within 25 percent of the desired effective values.

(4) As a practical matter, if the controller we are tuning is an analog controller, then the error in calibration of the dial settings is probably greater than the error introduced by our failure to take into account the difference in the controller forms.

(5) Shinskey (see Ref. 5-1) claims that the interactive form is safer since it is impossible to set a combination of dial settings that will produce an effective value of derivative time that is greater than the effective value of integral time. With a noninteractive form, nothing prevents the controller from being tuned, with the derivative time in excess of the integral time.

(6) There is no functional, technical advantage to either the interactive or noninteractive form. The noninteractive (standard) form has a wider choice of effective coefficients (i.e., T_D can exceed T_I for overshoot reduction on set point changes); therefore, it is more flexible than the interactive form. The word "interactive" often connotes benefits, as in interactive graphics. It has no such meaning here, however.

(7) With many microprocessor-based control systems, the control algorithm is formulated to mimic the analog controller (interactive form), not the ideal (noninteractive) form. Here, the conversion from required effective values into parameter entry values might be somewhat more important. Some manufacturers give the user the choice of the interactive or noninteractive form.

(See footnote on page 87 regarding inconsistency in nomenclature among various manufacturers.)

(8) Most computer-resident process control software systems use the noninteractive form.

❖ INDEPENDENT GAINS

We have pointed out that with the traditional formulation of the PID, the controller gain is a multiplier that affects all three modes. Some manufacturers formulate the PID with an independent gain on each mode, as shown by Equation 5-15 and in Figure 5-6. (They may refer to this as the "parallel" form of PID.)

$$m = K_C e + K_I \int e\,dt + K_D \frac{de}{dt} \tag{5-15}$$

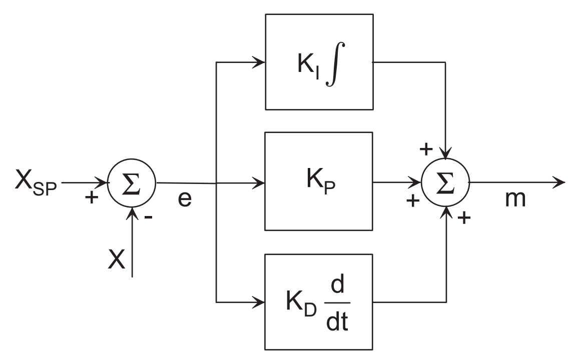


Figure 5-6. Independent Gain Adjustment on Each Mode

Earlier, we noted that equivalent behavior can be obtained between the noninteractive and interactive PID forms by adjusting the tuning parameters appropriately. So too, here, we can obtain equivalent behavior between the standard PID and that shown in Figure 5-6 by setting the following relationship between the tuning parameters:

Standard PID Figure 4-13	Independent Gains Figure 5-6
K_C	K_P
$\frac{K_C}{T_I}$	K_I
$K_C T_D$	K_D

Thus, a performance advantage cannot be claimed for either form. The form represented by Figure 5-6 is probably more likely to be found in control applications such as electric drives and power generation than in the more traditional applications in the process industries. Plants choose this form more as a matter of preference and tradition for those applications than for actual performance.

❖ INTERNAL FILTER

In chapter 4, we stated that if the derivative mode is used in the presence of measurement noise, then the derivative action will amplify the noise and produce an excessive noise component on the controller output. For this reason, we recommended that the derivative mode not be used on control loops that have excessive noise.

If a moderate amount of noise is present, however, it is still possible to use derivative, provided that a filter is used to attenuate the noise. The actual physical construction of a filter can take many different forms, such as a restriction and a bellows in a pneumatic controller, an R-C network (or an R-C network with an operational amplifier) in an electronic analog controller, or a few lines of code in a digital control algorithm. Mathematically, the filter is often represented as a first-order lag, although more complex forms of filters are available in some commercial controllers.

Structurally, the filter can be incorporated into the controller in several ways, as shown in Figure 5-7. The filter can either be placed on the incoming measurement signal to the controller, as shown in Figure 5-7a, or merely in series with the derivative unit, as shown in Figure 5-7b. When the controller uses the interactive form, the filter can be placed on the error itself, as shown in Figure 5-7c. This controller can be represented in Laplace notation by Equation 5-16. Many texts state that this form more nearly represents traditional analog controllers than does Equation 5-6:

$$M(s) = \frac{K_C(T_I s + 1)(T_D s + 1)}{T_I s\left(\frac{T_D}{\alpha} s + 1\right)} E(s) \cdot \tag{5-16}$$

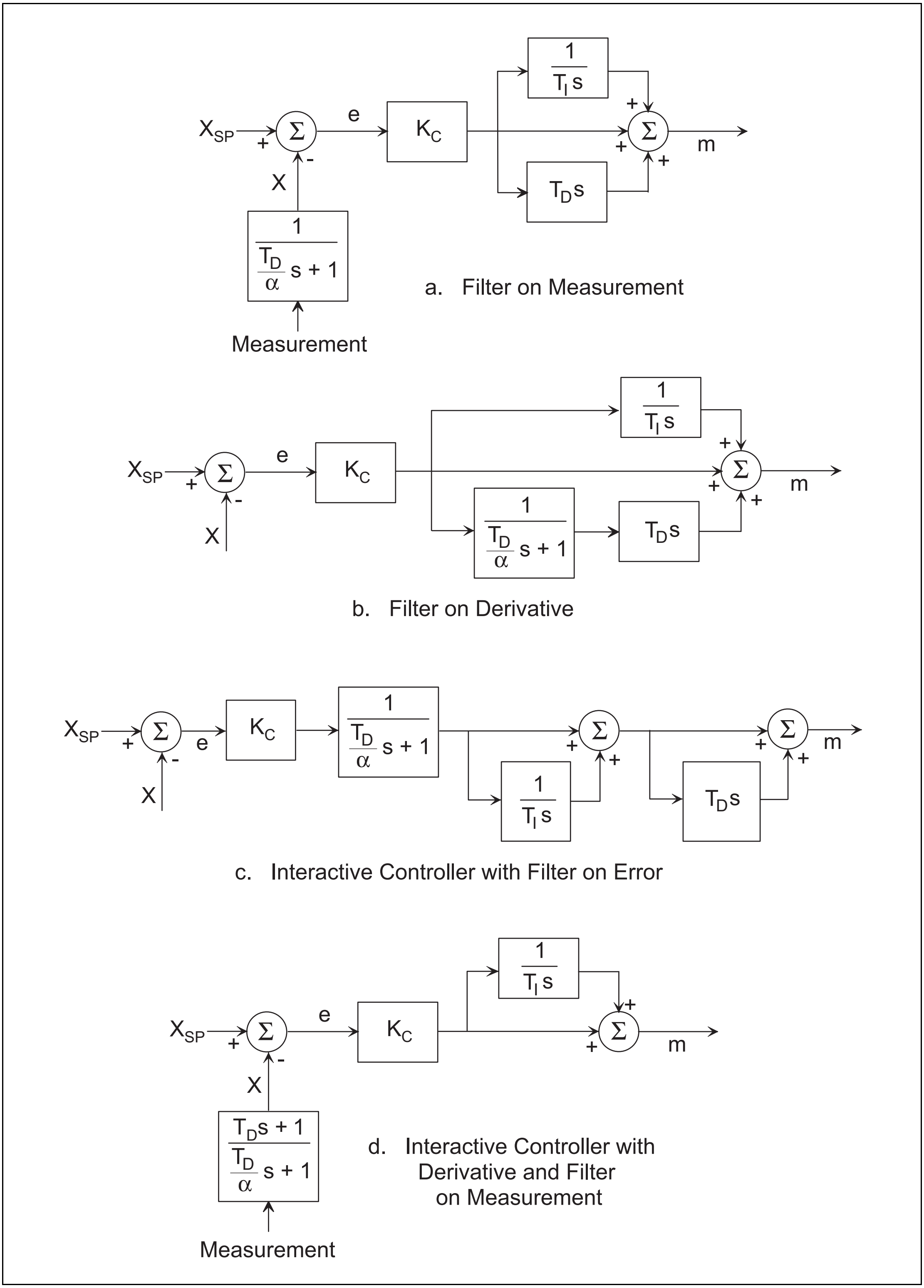


Figure 5-7. Internal Filter Configurations

Analog and microprocessor-based controllers that apply the derivative mode to the measurement can be represented by Figure 5-7d and Equation 5-17:

$$M(s) = K_C \left[\frac{(T_I s + 1)}{T_I s} E(s) - \frac{K_C (T_I s + 1)(T_D s + 1)}{T_I s \left(\frac{T_D}{\alpha} s + 1 \right)} X(s) \right] \tag{5-17}$$

If the filter is formulated as a first-order lag and the filter time constant is short relative to the process dynamics, then the filter placement will not have an appreciable effect on the behavior of the control loop. On the other hand, if there is heavy filtering (i.e., the filter time constant is the same order of magnitude as the dead time of the process), and if the filter is on the overall measurement signal, then the process's apparent dead time will be increased by approximately one-half of the filter time constant. This will significantly affect the controller tuning.

The filter time constant is often made proportional to the user-adjustable derivative time. For example, the filter time constant is given by T_D/α. The parameter α can have a value of from 6 to 20; it often has a built-in value of 10. Thus, the filter time constant is automatically set at 0.1 times the derivative time. This limits the amplitude of the derivative spike, when the step change in set point is made, to 10 times the proportional response. This is often called the "derivative gain." Some manufacturers make the derivative gain accessible so users can adjust it.

❖ NONLINEARIZATION

A PID algorithm in any of the forms described in this chapter is termed a "linear" controller. For a given amount of error, it will always respond in the same manner. There are circumstances where one might want the algorithm to perform in a different fashion. If a PID is used as a liquid-level controller, it may be desirable for the controller to have very conservative action when the error is small, to avoid excessive fluctuations of the flow rate. If, however, the error is large, more aggressive controller action may be desired to prevent the vessel from draining or overflowing. To achieve this, the error signal can be modified by a characterizer function. The modified (pseudo) error signal, designated $\hat{e}$, is then used by the PID modes, as shown in Figure 5-8.

One popular nonlinearization function is called "error-squared" or "absquare" (for "absolute value of the square of the error"). Neither of these terms is a precise mathematical description of the function; that is best provided by Equation 5-18 and by the graphical relation shown by curve A in Figure 5-9.

$$\hat{e} = \frac{e \times |e|}{100} \tag{5-18}$$

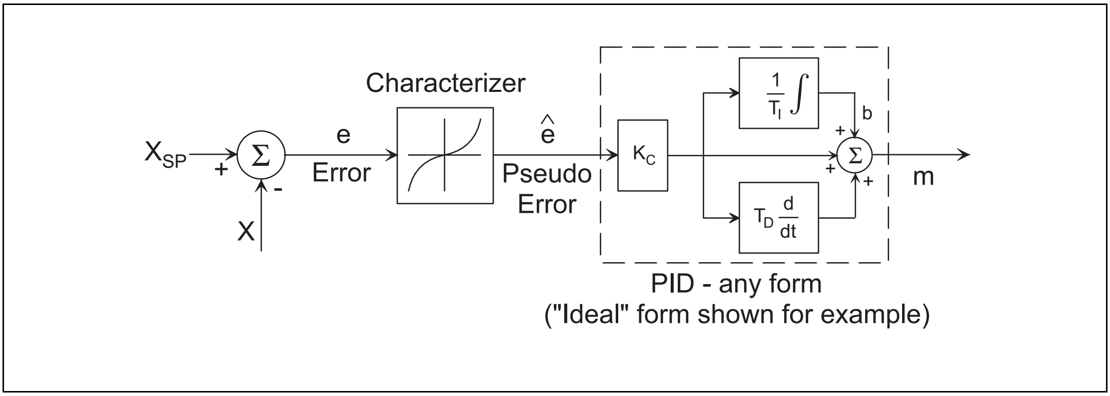


Figure 5-8. Insertion of a Characterizer in a PID Algorithm

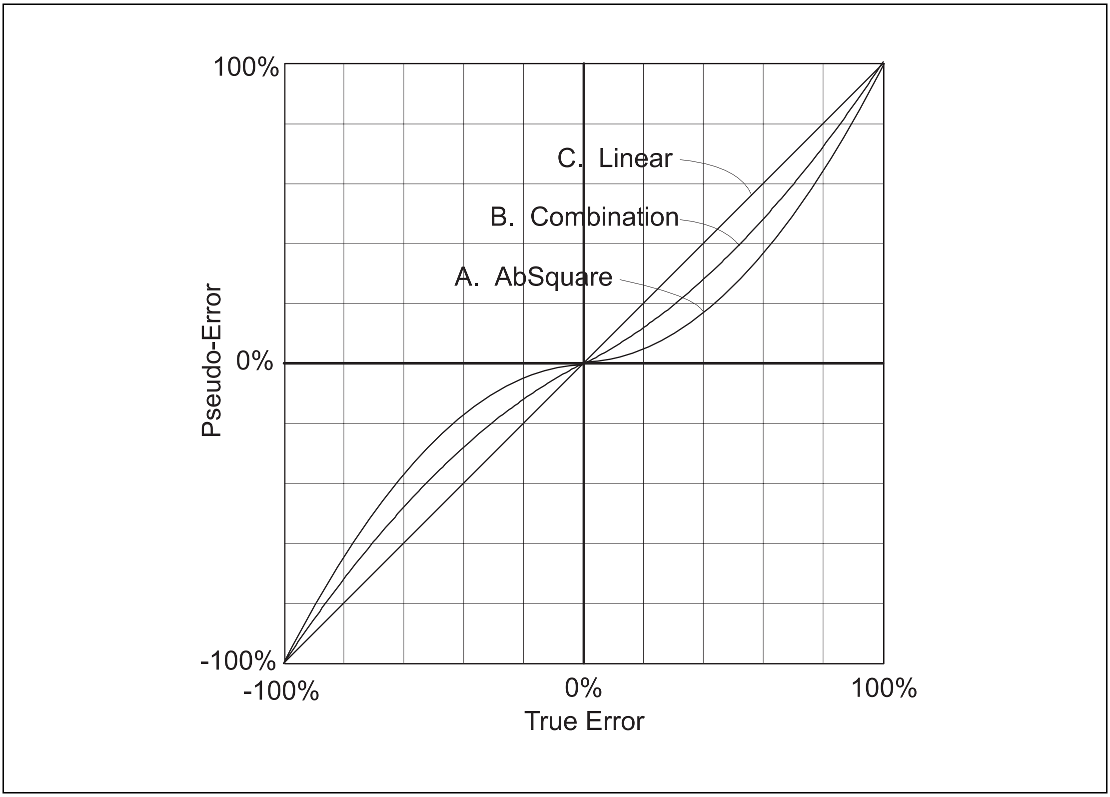


Figure 5-9. Error versus Pseudo Error Relationship

The effect of this nonlinearization function is both to lower the controller gain when the error is near zero and to gradually increase the controller gain as the deviation from set point increases. In fact, when the error is zero, the effective controller gain is zero. Possible variations of Figure 5-8 would be to place the non-linear characterization on only the proportional mode, or on only the integral mode, rather than on the common error signal. A slightly less drastic arrangement of Figure 5-8 would be to take a combination of the simple error and the absquare error, as shown by curve B in Figure 5-9 and described by the following equation:

$$\hat{e} = \theta e + (1-\theta)\frac{e \times |e|}{100} \tag{5-19}$$

where θ is an adjustable parameter that can take any value from 0 to 1. With this arrangement, if $\theta > 0$ there is a finite controller gain, even at zero error.

Page 168 contains further discussion related to the use of error-squared algorithms for level control.

In addition to its use in liquid-level control, the "error-squared" or "absquare" form of nonlinearization has also been used in pH control. Suppose you are neutralizing a waste stream by adding a reagent. The titration curve is very nonlinear and is essentially the opposite nonlinearity provided by Figure 5-9. Therefore, using this nonlinear control feature will tend to cancel out the nonlinearity of the process.

Another form of nonlinearization is to characterize the error with straight-line segments. For a custom application, this may be nonsymmetrical. Commercial systems would have symmetrical characterization, as shown in Figure 5-10, curve A. If the central segment has zero slope as shown in Figure 5-10, curve B, this is called a "gap action" or "dead zone" algorithm. An application for this algorithm would be for positioning a final actuator that has a reversible electric motorized valve. If the process variable is near the set point (deviation $< \pm b$), the controller acts as if there were zero error, hence it doesn't move the valve. The valve motor is only activated if the deviation exceeds the break-point limits. The objective of this form of control is to prevent "chatter" of the valve motor, albeit at the cost of control of somewhat reduced quality.

❖ SET POINT TRACKING AND BUMPLESS TRANSFER

Most feedback controllers have some form of auto-manual switch, either in the hardware or software, on the controller output. This allows the operator to intervene, if necessary, and manually set a value of controller output instead of having the controller set it automatically. It is usually desirable that a switch from manual to automatic, or from automatic to manual, be made without abruptly changing the position of the valve or other final control device. This is called "bumpless transfer."

At one time the responsibility for achieving bumpless transfer was entirely up to the operator. He or she had to preset, or balance, certain signals before switching. While many of these controllers are still in use, the procedure for switching bumplessly varies so much from one manufacturer to another that it is not appropriate to cover the subject here. Suffice it to say that with modern controllers, the procedure that achieves bumpless transfer has been automated. This places much less burden on the operator to manipulate the process correctly.

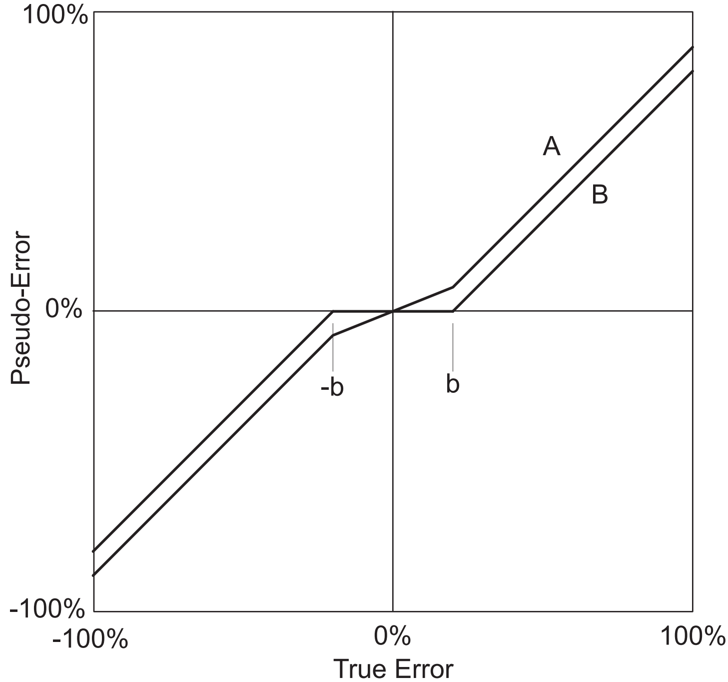


Figure 5-10. Error Characterization with Straight-line Segments

One bumpless transfer procedure involves set point tracking (also called "PV tracking" by some manufacturers). With set point tracking, whenever the controller is in manual, the set point follows, or "tracks," the measurement value, even though the measurement itself may be varying. Thus, when the controller is switched from manual to automatic, there is initially no error in the feedback loop. One other provision is required for bumpless transfer—the initial value of the integral mode contribution must also be made equal to the output value entered by the operator. These two provisions will achieve complete bumpless transfer from manual to automatic. The exact procedure for implementing these provisions will vary from manufacturer to manufacturer.

Although set point tracking is one way to achieve bumpless transfer, it is not the only one. Bumpless transfer is universally desirable—there are no conceivable situations where it would not be wanted. On the other hand, for some applications set point tracking has its drawbacks. For example, if the control loop in question is controlling a quality attribute of a final product, such as composition, then it would be desirable to be able to enter the required set point to meet product specifications rather than have this target value altered by manual/automatic switching. Here, bumpless transfer without set point tracking would definitely be advantageous.

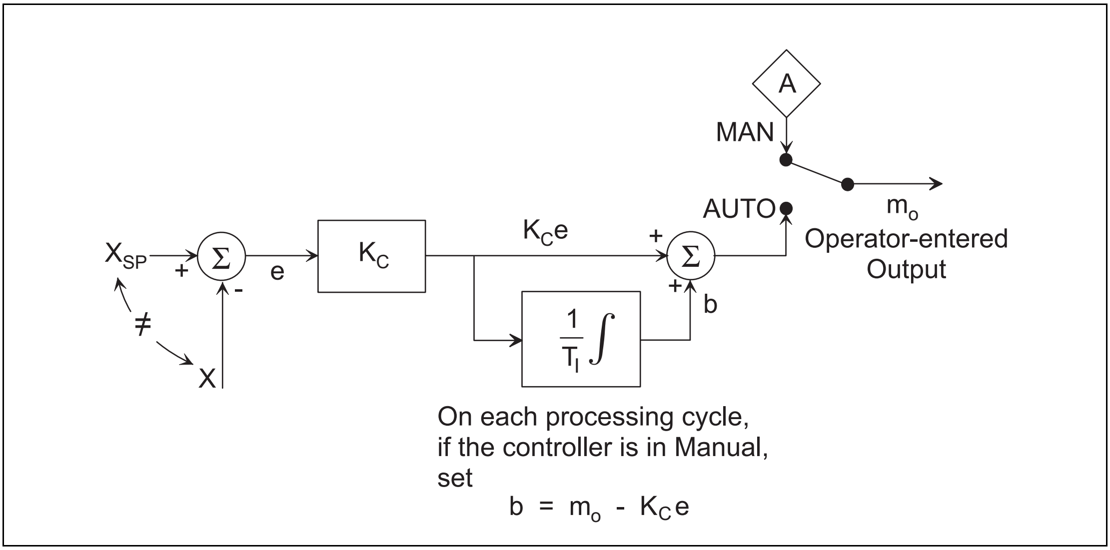


Figure 5-11. Bumpless Transfer without Set Point Tracking

It can be achieved by causing the initial value of the integral mode contribution ("*b*" in Figure 5-15) to track the proper value whenever the controller is in manual:

$$b = m_o - K_C e, \tag{5-20}$$

where m_o = operator-entered output value.

On a switch from manual to automatic, the position of the final control element will not change abruptly. However, if there is a deviation between set point and measurement, the integrator will react to this error by gradually moving the valve to a new position that will eliminate the error. Thus, we will have achieved bumpless transfer without set point tracking. Many contemporary microprocessor-based systems implement manual/automatic bumpless transfer in essentially this manner.

We have only discussed bumpless transfer when switching from manual to automatic. It is also desirable that the switch from automatic to manual be bumpless. This is a simple hardware design task for analog controllers. We will leave a discussion of control algorithms implemented in software for a later section of this chapter.

❖ "BUMPLESS" TUNING

"Bumpless tuning" is not a common term. Most users have not considered it as a configuration option or feature. But it is a feature that manufacturers must provide in their controllers to avoid "bumping" the controller output whenever controller tuning parameters are changed.

The integral tuning parameter can be changed anytime, even when the controller is in automatic mode, without causing an abrupt change in the controller output. This is not so with the gain and derivative tuning parameters, however. If the gain (or proportional band) is changed when the controller is in automatic and *the set point and measurement are not equal*, then there will be a "bump" in the controller output unless the manufacturer has provided for bumpless tuning. The provision that must be made is to readjust the contribution of the integral mode so that the output is unchanged by the tuning change.

Assume that we are speaking of a discrete proportional-plus-integral algorithm (which we cover in more detail in the next section) that is processed periodically. Also assume that the controller gain has been changed since the last sampling instant. The controller output before and after the gain change is given by the following:

$$m_{old} = K_{C,old}(e + b_{old}),$$
$$m_{new} = K_{C,new}(e + b_{new}),$$

where: e = error,
b = integral-mode contribution to the controller output.

Since we want m_{new} to be the same as m_{old}, we can equate the right-hand side of these two equations and determine a value for b_{new}, based on the error and the amount by which the gain was changed:

$$b_{new} = b_{old} + e \times (K_{C,old} - K_{C,new}) \tag{5-21}$$

Similarly, if we are using PID and the derivative is changed, then before the next sampling time the integral output should be adjusted by this equation:

$$b_{new} = b_{old} + \frac{de}{dt} \times (T_{D,old} - T_{D,new}) \tag{5-22}$$

❖ PREVENTING RESET WINDUP

If a feedback controller containing the integral mode is unable to bring the measurement to the set point, say, due to a valve being at a limit, there will be a sustained error in the loop. The integral mode will eventually drive the controller output to a saturation limit, such as 0 percent or 100 percent. Such a condition is called "reset windup," or merely "windup." Reset windup is especially bothersome in analog controllers where there are no hard limits on the output at 0 percent and 100 percent (3 and 15 psig, or 4 and 20 mA). For example, a pneumatic controller output can go beyond the 15 psig limit, all the way to the supply air pressure, which may be 18 to 20 psig. Similarly, an electronic analog controller output can go to about 24 mA. When the controller does regain control of the loop, there must be a significant error in the opposite direction either of sufficient magnitude or sufficient duration to cause the controller output to come back within sight, that is, within the 3–15 psig or 4–20 mA nominal range. With a controller that is implemented in a digital processor, if there is a sustained error, and in the absence of other provisions, the integrator output will continue to grow without limits. Clearly, a mechanism for avoiding reset windup in such situations is desirable.

There are several methods for preventing reset windup; some are only applicable in certain circumstances. All the techniques we will mention here assume that the controller is implemented in a digital processor.

One of the simplest methods for anti-reset windup protection is to simply stop the integration if the controller output reaches a limit. When the condition that causes the windup is removed, the controller output will recover back to a normal operating point, at a rate dependent primarily on the integral tuning.

Some manufacturers offer a considerable improvement over this technique. If the controller output reaches a limit and then begins to recover, the integral action is accelerated by a factor of sixteen until the process variable has returned to its normal operating point. This permits the system to recover much faster from a windup condition.

♦ External Reset Feedback

Override (or selector) control systems have a unique need for reset windup protection. (Override control is discussed in detail in chapter 12). In a typical override control application, one controller controls the valve in normal circumstances. In abnormal circumstances, however, this controller's output is "overridden" by another controller, which then takes control of the valve. If an ordinary PI (or PID) controller's output is overridden by another controller, it will be unable to achieve its set point; consequently, it will wind up. This is a problem posed by using ordinary PI controllers in override applications.

Some manufacturers provide a modified controller form to overcome this problem. This modification is said to use "external reset feedback," "external reset," or simply "reset feedback." Since the derivative mode does not contribute to the problem of reset windup, we will use only the P and I modes in our discussion of this modification.

For a PI controller, the modification that uses external reset feedback is indicated in transfer function form by the block diagram of Figure 5-12 and by Equation 5-23.

$$M(s) = K_C E(s) + \frac{1}{T_I s + 1} M(s) \tag{5-23}$$

This equation and figure indicate that the controller output is computed as the sum of the controller gain times error, plus a lagged value of its own output. The controller output is fed back into a first-order lag whose time constant is the desired integral time of the controller.

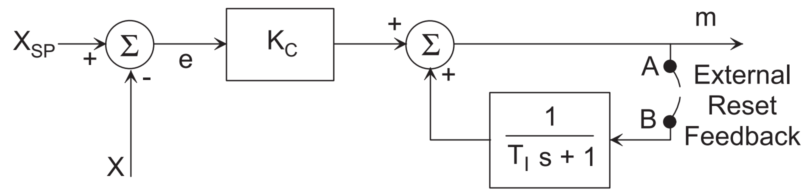


Figure 5-12. PI Controller Formulated with External Reset Feedback

Although it is not immediately obvious, a controller that is formulated in this way has exactly the same behavior as a standard PI. One way of demonstrating this is to solve Equation 5-23 for M(s). After a bit of algebraic manipulation, this yields

$$M(s) = K_C\left(1 + \frac{1}{T_I s}\right)E(s) \tag{5-24}$$

which is exactly the same transfer function as for the standard PI (see Equation 4-4).

The internal formulation of the controller would be of no interest to us if its only attribute were that it behaved the same as a standard PI. The real benefit occurs if the manufacturer has configured the controller, either in hardware or software form, so that the link from A to B in Figure 5-12 can be removed. Then we can input the signal to the external reset feedback port B from anyplace we choose. A common configuration in override control is to take the reset feedback from the output of the selector device that selects between the normal controller and an abnormal controller (see Figure 12-4). Since there is no explicit integrator in the controller, the nonselected controller will not wind up, even in the presence of a long-term error because of the overriding action of the alternative controller.

♦ Batch Switch

For batch process applications, as well as for similar applications such as process startups or significant changes in operating point (grade changes), the measurement value may be in transition between an old set point and a new one for a considerable duration. During this time, the action of the integral mode will cause windup. For example, suppose the set point of a temperature loop is raised significantly. The integral action of the controller may cause the valve to open fully long before the measured temperature reaches the new set point. Then, getting the valve back to its normal operating range will require a significant overshoot of the set point.

Recall from chapter 4 that the position of the proportional band is shifted by integral action (see Figure 4-12). Thus, the reason for the overshoot is that during the temperature rise the integral action shifts the proportional band so it lies entirely above the set point. To bring the valve back from its wide-open position, the measurement must be somewhere within the proportional band, and hence must exceed the set point. Only then does the reversal of the sign of the error cause the integral action to shift the proportional band back down to a normal operating region.

Windup can also occur if there is a limitation in the controlling medium itself. For instance, if there is a loss of steam supply, a temperature-control loop will wind up simply by attempting to achieve a constant set point. When the deficiency is corrected, there will be a significant overshoot of the set point before the valve gets back to its normal operating point.

Anti-reset windup techniques for this application generally consist of forced-shifting of the proportional band by some means other than the normal integral action. This can be accomplished either in hardware controllers or in software control algorithms. We will present the functional details here, which can be implemented in either technology. This anti-reset windup technique is often called a “batch controller” or a “controller with a batch switch” (Ref. 5-1). With this controller (as with every PI controller), in normal operation the controller output is determined by the following:

$$m = K_C e + b$$

where b is the integral-mode contribution.

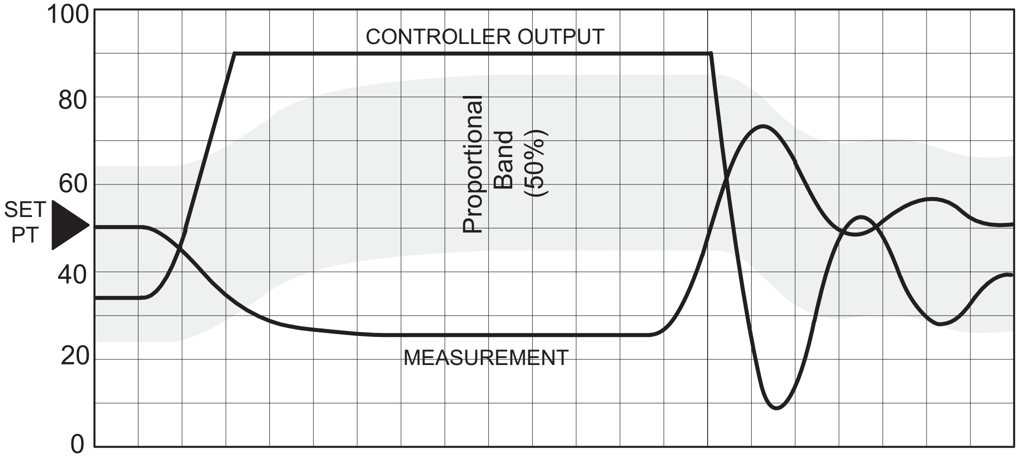
With the batch switch feature, however, if a sustained error causes the controller output to reach a maximum value, the contribution of the integral mode is back-calculated so as to hold the controller output at this maximum value.

$$b = m_{max} - K_C e \tag{5-25}$$

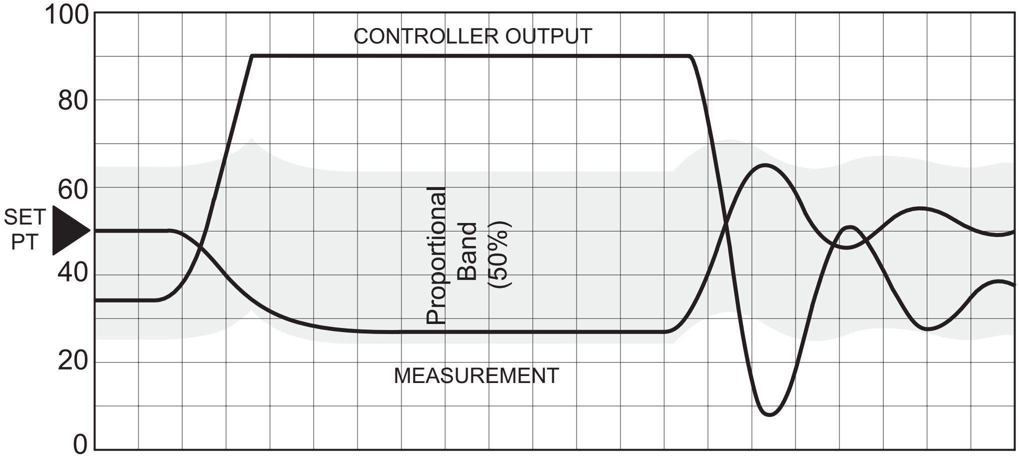
The effect is to shift the proportional band downward during the period of noncontrol. When control can be resumed, the measurement value will already be within the proportional band, so the controller output will start to cut back right away. The net result is a reduction in overshoot when control is resumed.

Figure 5-13 shows the results of a simulation-generated demonstration of this “batch controller” technique. The simulation scenario is that of a temperature controller controlling a steam valve. During a period of normal control, the steam supply is interrupted, causing the controller output to wind up to a maximum value (set at 90 percent). The characteristic of interest is the recovery of the control loop when the steam supply is resumed.

In Figure 5-13a, the controller is a conventional PI controller without anti-reset windup. Note that when control is lost, the integral action shifts the proportional band upward so that most of it lies above the set point. (Had the maximum output been set at 100 percent, the PB would have been entirely above the set point.) When the steam supply recovers, the measurement must rise to within the PB before the controller output starts decreasing. Note that there is a significant overshoot before normal control is reachieved.



a. Without Batch Switch



b. With Batch Switch

Figure 5-13. Overshoot Reduction Using the Batch Switch

In Figure 5-13b, the controller has been modified to a PI with anti-reset windup. Note that when control is lost, the PB is initially shifted upward. However, once the controller output reaches the maximum the PB is shifted downward, as a result of back calculation of the output bias (Equation 5-25). Upon recovery, the measurement must still rise to within the PB before the controller output is reduced, but this occurs much quicker; consequently, the overshoot is reduced.

In a situation like this, using derivative would also reduce the overshoot, and in fact using both derivative and anti-reset windup will produce very dramatic results. If the proportional band is shifted downward too far, the result will be a "negative overshoot" (i.e., the measurement will not initially reach the set point). Consequently, batch controllers of this type have a lower limit on b. This is called the "reset preload" value; it is user adjustable.

Other tuning parameters that must be adjusted on a batch controller are maximum controller output and preload settings. In a given situation, the batch controller's performance depends greatly on these settings as well as on the process load. The batch controller works best when it is adjusted for one repeatable situation, including set point and process load.

❖ DISCRETE CONTROL ALGORITHMS

Although we have referred to computer or microprocessor-based control, our examples of PID control have so far used equations and block diagrams to illustrate only continuous form (Equation 5-1 and Figure 4-13, for example). Let us now consider how the form represented by Equation 5-1 would be implemented in discrete form for a digital computing control system (Ref. 5-2). The formulation for a digital device is called a control algorithm.

First, let us assume the following:

- That the algorithm is processed on a repetitive basis, say, every ΔT seconds. (ΔT can be as short as 50 ms in some distributed control systems or as long as several minutes in some host computer-based systems. We assume that for a particular control loop in a particular system ΔT is constant.) We can speak of processing the algorithm at the n^{th} processing instant and designate values computed at that instant with a subscript "n".

- That the process variable is sensed and converted either into engineering units or into a normalized value by a combination of hardware and software. On each processing cycle of the control algorithm the most recent value of the process variable, designated x_n, is available to us.

- That, likewise, the set point is available to us; this is designated SP_n.

- That other variables, computed on previous processing cycles, have been saved in memory and are available to us. These are designated with subscripts "n-l", "n-2", etc.

In the automatic mode, a processing cycle begins by determining the error between set point and measurement:

$e_n = SP_n - x_n$ if the controller is reverse-acting, or
$e_n = x_n - SP_n$ if the controller is direct-acting.

The next step is to approximate the integration operation of the continuous algorithm. Since integration computes the area under the curve, the integration operation can be approximated by a technique called *box car integration*, in which each discrete value of error is multiplied by ΔT, thus computing the area of a rectangle. The area of all rectangles are summed. Alternatively, as a new value of error is determined, a new sum of rectangular areas can be computed by adding the new rectangular area to the previous sum:

$$S_n = S_{n-1} + \Delta T\, e_n \,. \tag{5-26}$$

The term S_n is equivalent to the integral of the error up until time n.

While this series of steps is conceptually sound, it is equally valid and more convenient to simply sum up all past errors, then multiply the resulting sum by ΔT in a later operation:

$$S_n = S_{n-1} + e_n \,. \tag{5-27}$$

The differentiation operation can be approximated by a discrete differencing, for example:

$$\frac{de}{dt} \approx \frac{e_n - e_{n-1}}{\Delta T}$$

It is equally valid and more convenient, however, to compute merely the difference between the present and previous error, then divide by ΔT in a later operation.

We are now ready to compute the controller output value at time n:

$$m_n = K_C\left(e_n + \frac{\Delta T}{T_I} S_n + \frac{T_D}{\Delta T}\left(e_n - e_{n-1}\right)\right) \tag{5-28}$$

Equation 5-25 is the discrete counterpart of the standard PID equation written in continuous form in Equation 5-1. Because this equation computes the position of the final control element, it is usually called the *position* form.

It is sometimes desirable to calculate the increment by which the controller output should change rather than the actual output itself. The increment of change is given by:

$$\Delta m_n = m_n - m_{n-1}$$

This expression is not a valid way to calculate Δm_n, however, since we know neither m_n nor m_{n-1}. We need an equation for calculating Δm_n directly. We first write an expression for m_{n-1}, similar to Equation 5-25, except all the subscripts are decreased by 1:

$$m_{n-1} = K_C\left(e_{n-1} + \frac{\Delta T}{T_I}S_{n-1} + \frac{T_D}{\Delta T}(e_{n-1} - e_{n-2})\right) \tag{5-29}$$

Now make a term-by-term subtraction of Equation 5-29 from Equation 5-28, using Equation 5-27 to compute $S_n - S_{n-1}$:

$$\Delta m_n = K_C\left(e_n - e_{n-1} + \frac{\Delta T}{T_I}e_n + \frac{T_D}{\Delta T}(e_n - 2e_{n-1} + e_{n-2})\right) \tag{5-30}$$

This equation is also a discrete counterpart to the continuous form of the standard PID, Equation 5-1. It calculates the incremental change in controller output over a time period ΔT, and is often called the *incremental* or *velocity mode* form.

All the modifications to the standard form of the PID mentioned earlier in this chapter are also applicable to both the position and the incremental forms of the discrete PID. For example, the incremental form of the PID, which has modifications that place both the proportional and derivative on measurement rather than error, is as follows:

$$\Delta m_n = K_C\left(x_{n-1} - x_n + \frac{\Delta T}{T_I}e_n - \frac{T_D}{\Delta T}(x_n - 2x_{n-1} + x_{n-2})\right) \tag{5-31}$$

The incremental form calculates the required change in controller output or valve position. It depends upon some other hardware or software provision to "remember" the last actual position of the controller output or valve position. In the following paragraphs, we will describe a very common way of achieving this.

For many microprocessor-based systems, a control strategy is configured by a series of software function blocks. Many times, these function blocks are the counterpart of hardware modules in an analog control system. Just as a set of hardware modules require interconnections (through wiring or pneumatic piping) to form a complete control system, so a set of software function blocks also requires interconnection. This is often called "softwiring."

Figure 5-14 shows a simple feedback loop in which the microprocessor-based control portion consists of three function blocks:

An analog input (AI) function block that causes an analog-to-digital (A/D) converter (hardware device) to convert the incoming sensor signal (say, 4–20 mA) into an analogous value that may be in engineering units or a normalized signal (say, 0 to 100% of transmitter range). This value is deposited in a memory register that can be considered as a part of the AI function block.

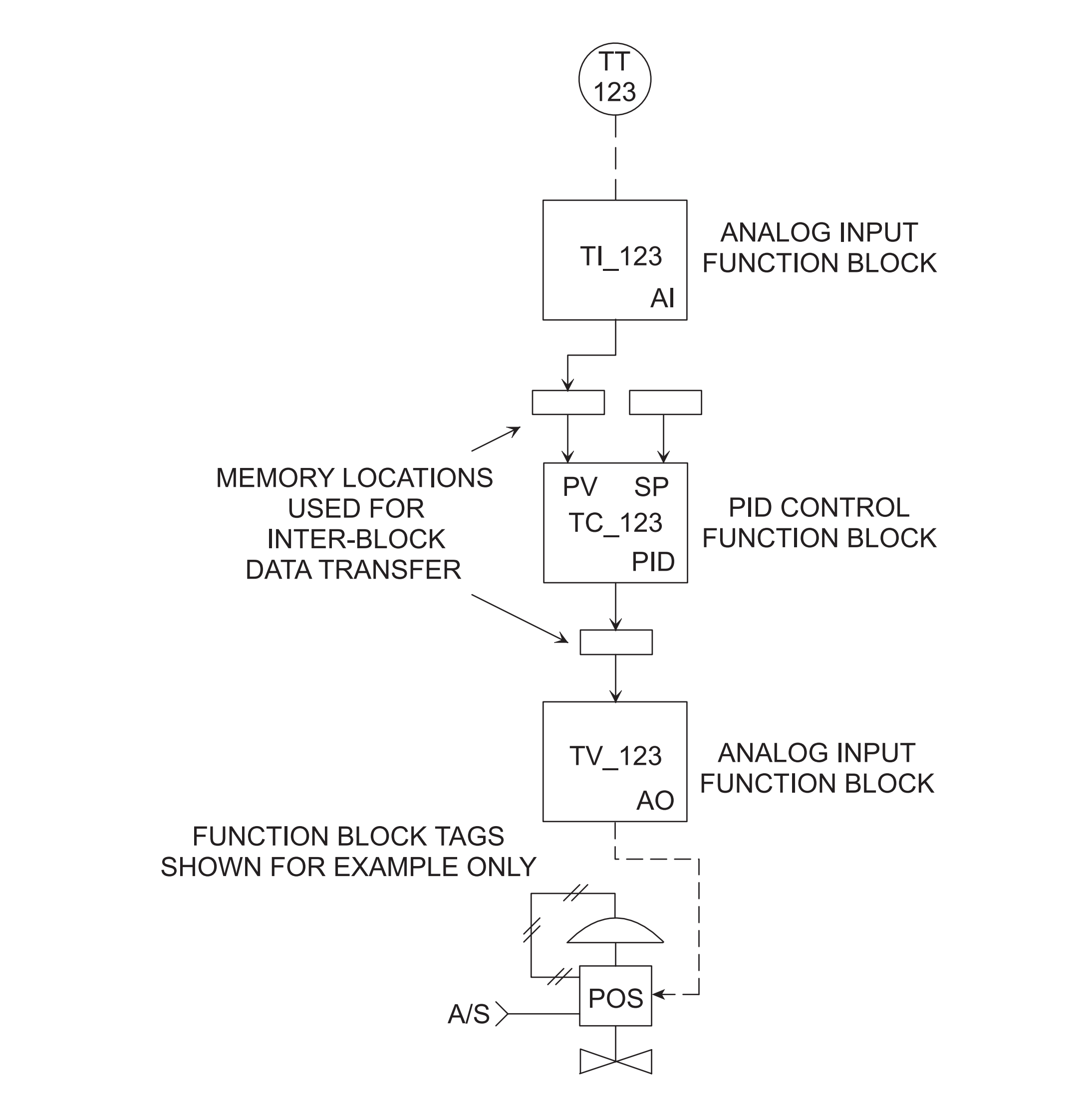


Figure 5-14. Typical Control Loop Hardware/Software Structure

A control function block (PID) that obtains the measurement value from the AI block, compares this with the set point, then executes a discrete form of PID algorithm (Equation 5-28, 5-30, or a form containing any of the modifications), and calculates either the required controller output or change in controller output.

An analog output (AO) function block that obtains from a memory register the required position of the valve or other final control element. This value (0 to 100%) is converted by a digital to-analog (D/A) converter (hardware device) into a signal (say, 4–20 mA) to the valve.

If the PID is a position type, it calculates a new value for the required valve position and overwrites the old value in the memory register referenced by the AO block. If the PID algorithm is an incremental type, it calculates the required change in valve position and adds this change to the current value in the memory register. In either case, the AO block has a new value at which to reposition the valve or other final control element (see Figure 5-15). In normal operation, both the position and incremental algorithm forms produce essentially identical behavior in the control loop. Hence there is no technical advantage to the user of one form over the other. However, from the viewpoint of the software implementer, it is easier to deal with certain situations such as manual/automatic switching and tuning parameter changes using the incremental form. Hence, many microprocessor-based control system vendors use this form.

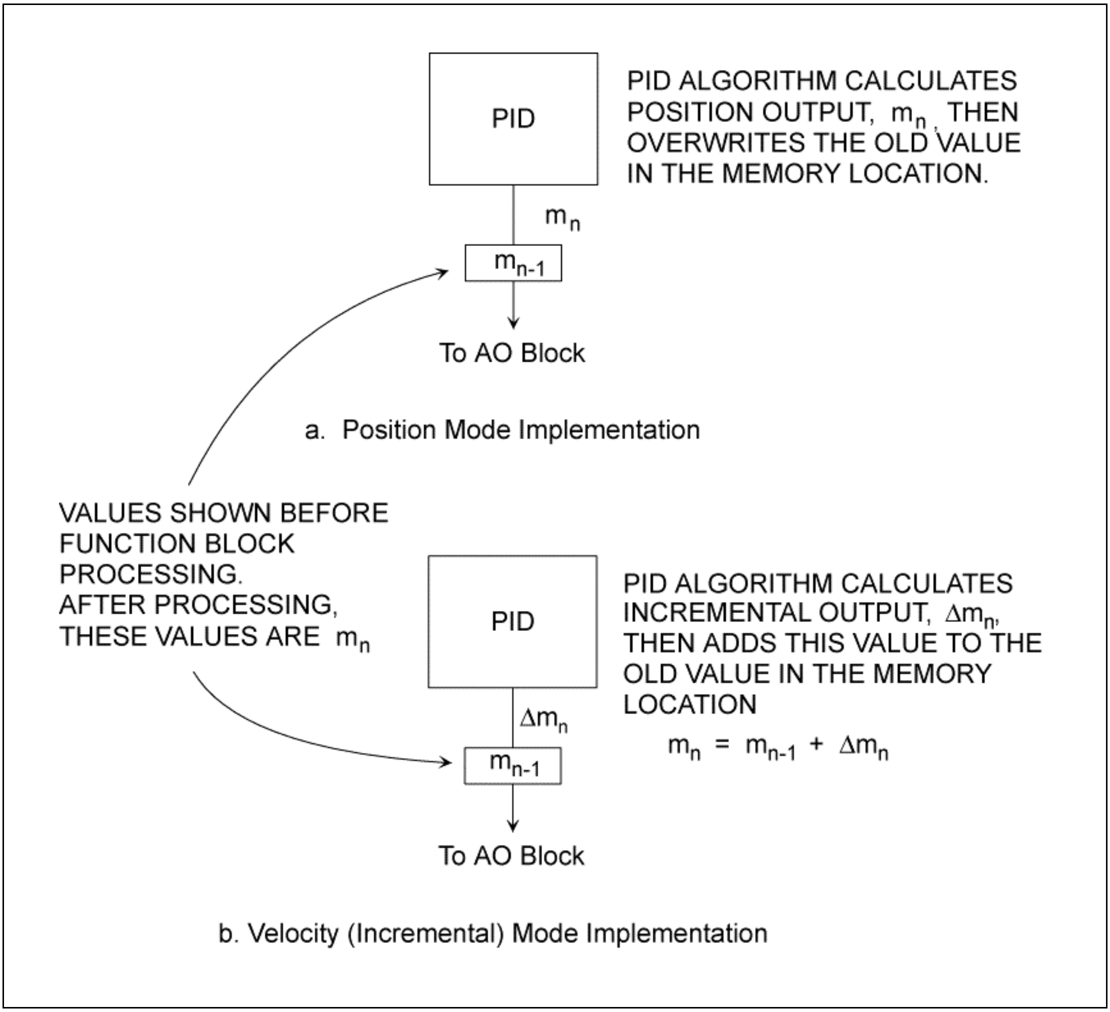


Figure 5-15. Position and Incremental (Velocity) Mode Algorithms

Dealing with abnormal situations involves additional communication between function blocks. This is often accomplished by "flag passing," wherein one function block sets status bits that other function blocks recognize and act upon. The exact features and details of implementation are specific to each vendor, but the following features are fairly common:

When the PID block is switched from automatic to manual, it simply stops calculating its incremental change. The position of the final control element remains at its last value. Thus, bumpless transfer from automatic to manual is inherently achieved.

The lower-level block (the AO block in the preceding example) will probably have limits beyond which the upper-level block cannot change the value of the required valve position. This is analogous to setting output limits of 4 mA (0%) and 20 mA (100%) on an analog controller output. These limits are user adjustable.

Because there is no memory register within the incremental algorithm that is acting like an integrator, there is no windup within the algorithm itself. However, if an external influence interrupts the control loop (say, a sticking valve), then the memory register that communicates with the AO block can wind up to its limiting value.

The PID block may impose a limit on the absolute value of the incremental change. Thus, the user may set with certainty the maximum rate of change of the final control element. (If the loop is well behaved, this limit will only be exercised infrequently. This limit should not be used to determine normal loop behavior in lieu of adequate tuning of the PID itself, as the author witnessed at one installation.)

Possibly the original reason the incremental form of the PID was developed was for control functions that resided in a central computer that directly determined valve positions. (This is known as direct digital control, or DDC.) To prevent the loss of many (perhaps several hundred) control loops during computer failure, the “memory” of the last valve position was contained in an external device, such as an electrical stepping motor. The PID controller in the computer calculated an incremental change in required output. This incremental change was transmitted to change the position of the external stepping motor; the stepping motor position was then converted into a signal that positioned the final control element. In the event of computer failure, there were no further changes to the stepping motor, hence the loop was said to fail to a “hold-as-is” condition.

❖ INCORPORATING ENGINEERING UNITS IN CONTROLLER GAIN

Scaling the controller gain is another aspect of discrete algorithms. When we presented the standard form of the PID, we did not characterize the gain term, K_C, as requiring engineering units; we implied that K_C was dimensionless. Indeed, it is if both the measurement signal and the controller output are both in a normalized form, such as 4–20 mA, 3–15 psi or 0 to 100 percent of scale. However, if a digital system (host computer or microprocessor-based system) converts the measurement into engineering units before making it available to the PID function block, and if the controller output is also calculated in engineering units, then, unless other provisions are made, the controller gain would also have to be in engineering units. (How would you like to tune a controller by setting the gain at 0.1 gpm per degree F?). To avoid this messy situation, the manufacturer will probably make one of the following provisions:

The process variable and set point will be scaled back to a normalized value, such as percent, before being used in the PID algorithm. The controller output, in percent, will be scaled to engineering units, if required by a downstream function block.

The controller gain is divided into two terms, as shown in the following equation:

$$m_n = K K_C \left(e_n + \frac{\Delta T}{T_I} S_n + \frac{T_D}{\Delta T} (e_n - e_{n-1}) \right) \qquad (5\text{-}32)$$

where K_C is the normal dimensionless tuning parameter and K is a scaling parameter that makes the equation dimensionally consistent. The engineering units for K are as follows:

$$\frac{\text{engineering units of controller output}}{\text{engineering units of measurement}}.$$

In some systems, K is transparent to the user and is calculated behind the scenes from the process variable and the output range data obtained from configuration questions. In other systems, K is explicitly set by the user during controller configuration, and will not be adjusted during tuning operations. A reasonable value for K is as follows:

$$K = \frac{\text{span of output}}{\text{span of measurement}} \qquad (5\text{-}33)$$

Example: Suppose a pressure-control loop is configured as a cascade system, with the primary controller being a pressure controller. Its output sets the set point of a secondary controller, which controls a cooling water flow. Suppose further that all signals are in engineering units rather normalized values. The signal ranges are these:

Pressure transmitter: 200–500 psig,
Flow transmitter: 0–150 gpm.

The set point of the flow controller (output of the pressure controller) must have the same range as the flow measurement. Thus, using Equation 5-33, the correct value for K is:

$$K = \frac{150 \text{ gpm - } 0 \text{ gpm}}{500 \text{ psig - } 200 \text{ psig}}$$

$$= 0.5 \frac{\text{gpm}}{\text{psig}}.$$

❖ COMMERCIAL EXAMPLES OF MODIFICATIONS

This chapter has presented some of the more commonly used PID controller modifications that may be found in commercially available controllers. Not all of these modifications will be found in every manufacturer's product, and indeed there are additional modifications (probably called "features" in the manufacturer's literature) which are not discussed here. In this section, we will describe two specific products, by manufacturer and product name, and illustrate how these modifications are used in these products. The choice of manufacturers was made merely for illustrative purposes, and should not be considered as an endorsement of those products over many other equally fine process control product lines currently available.

Essentially all manufacturers of digital control systems, whether distributed control systems, programmable controllers, single-station controllers, or computer-based software packages, will have a library of standard control algorithms. These enable the user to select the appropriate algorithms and configure his or her application-specific control strategy. This strategy may involve a feedback loop, as discussed so far, or a more complicated loop, such as ratio, cascade, feedforward, and so on, as we discuss later in this book.

There are two general approaches to formulating a library of control algorithms (see Ref. 5-3). One approach is to formulate comprehensive algorithms. For instance, a PID algorithm may also have the capacity to perform signal conditioning on the input, alarming, manual/automatic switching, and output limiting, as well as many other functions. With this approach, when a user designates a software function block to execute a PID algorithm, these additional features can also be performed with no further configuration.

The other approach is to formulate a library of elementary algorithms. With this approach, each algorithm performs only one (perhaps two) functions. To perform additional functions, the user will configure several elemental algorithms into a control strategy.

With the former approach, the manufacturer's library of control algorithms will be relatively shorter than with the latter. The advertised function-block processing rate ("blocks per second") will probably be less. This does not imply that either approach is superior to the other—the choice is simply one of vendor preference. In addition to other features, the two products described next illustrate these two approaches.

♦ Honeywell Distributed Control Systems

Most of the control algorithms in the Honeywell TDC 3000 and TPS distributed control system reside in a module called the Process Manager (PM), Advanced Process Manager (APM), or High-performance Process Manager (HPM). Control strategies are established by configuring function blocks, each of which executes a particular type of algorithm from a manufacturer-supplied library of comprehensive function blocks. For each function block, the user chooses the algorithm to be executed, makes selections from an extensive number of options for that algorithm, and interconnects ("softwires") the function block to other function blocks to form a complete control strategy (Ref. 5-4).

The PID algorithm executes the incremental form of the algorithm and, as usually applied, adds the incremental change in controller output to the reference value (set point) of a downstream function block.

During configuration, the user is given a choice of interactive or noninteractive controller, with these terms having the same meaning as earlier in this chapter. The interactive form has a preset filter on the derivative, as described by Equations 5-7 and 5-8, with a fixed value of 10 for α. The user is also asked to select the A, B, C, or D form of the algorithm. Table 5-3 can enable these designations to be correlated with the modifications discussed here:

Table 5-3. PID Algorithm Forms for the Honeywell PM, APM, and HPM

Algorithm Designation	Proportional mode	Integral mode	Derivative mode
A	Error	Error	Error
B	Error	Error	Measurement
C	Measurement	Error	Measurement
D	—	Error	—

There is an extensive number of additional options from which to choose, as well as a system for interblock communication that prevents an upper-level block from winding up if a lower-level block is in manual or if its output is saturated.

Data is passed between blocks in engineering units form rather than as normalized values. However, the PID controller gain is adjusted as a dimensionless value. It is automatically scaled by the engineering range of the process variable and controller output from user data that was entered when the function block was configured.

♦ Siemens Energy and Automation Loop Controllers

The Siemens family of loop controllers (formerly manufactured by Moore Products Co.) include the models 352P, 353, and 354N. These are microprocessor-based units with a large set of user-configurable elementary function blocks. This capability results in a highly flexible process control unit that is adaptable to a wide variety of applications (Ref. 5-5).

Figure 5-16 shows a block diagram that illustrates the essence of a typical feedback control loop configuration. This figure includes the following function blocks:

- Analog input (AIN), which converts the incoming measurement signal from 4–20 mA into a value in engineering units;
- PID controller (PID) with four essential inputs:

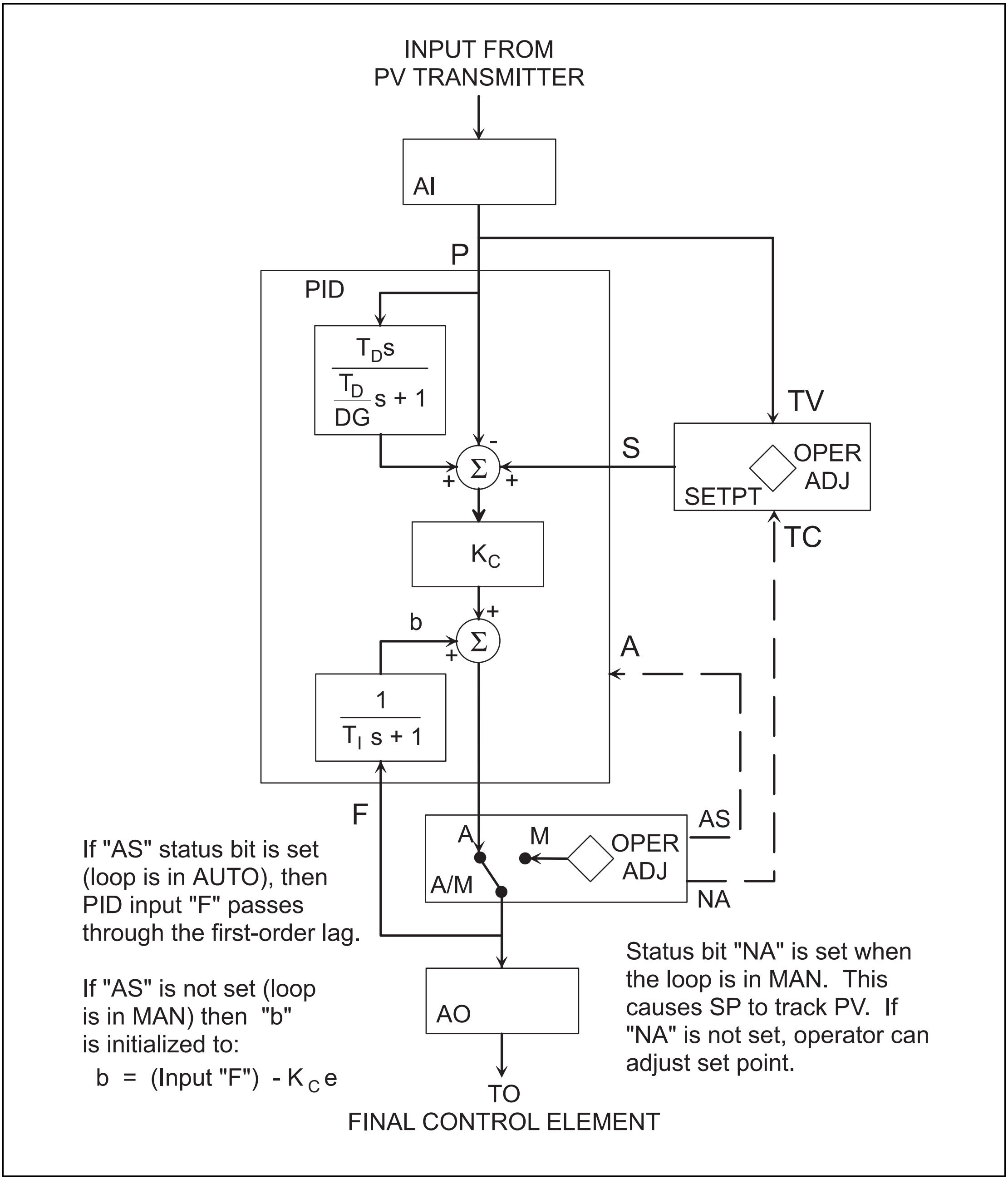


Figure 5-16. Feedback Control Loop Using Siemens Loop Controller Function Blocks

- P —Process variable,
- S —Set point,
- F —External reset feedback from output of auto-manual switch,

- A —A status signal with a value of either 0 or 1. Its role is described later.
- Automatic/manual (A/M) switch block;
- Analog output (AOUT), which converts the controller output, or operator-entered manual output value, into a 4–20 mA valve signal;
- A set point track/hold function block (SETPT) that either holds the set point value that the operator entered or tracks the measurement value (AIN block output) whenever the A/M switch is in manual.

In automatic operation, a software switch in the A/M function block passes the PID controller output (calculated in position form) directly to the AOUT block, where the signal is converted into 4–20 mA form. The output of the A/M switch is also connected to the external reset feedback port (F) of the PID controller, thus providing integral action.

Status signals are passed from the A/M block to both the PID block and to the set point track/hold block. When the status is Auto, the PID performs in its normal manner. The set point track/hold block maintains its last output as the set point to the PID. This value can be adjusted by the operator.

When the A/M is in manual, the external reset feedback signal is used to initialize the output of the first-order lag (see Equation 5-20). The output of the set point track/hold block follows the AIN block output (measurement value). Thus, in manual, the error is zero, so the other signal into the summing junction will be zero. The result is that the signal waiting on the “automatic” side of the A/M switch is the same as the output value the operator entered. Switching the control loop from automatic to manual does not cause a “bump” to the controller output. The set point tracking block maintains as the set point the measurement value at the time of switching.

Note from Figure 5-16 that this is an interactive controller with derivative on measurement. Also note the filter on the derivative component of the signal. The time constant of the derivative filter is the derivative time scaled by the value “DG” (derivative gain). The DG parameter can be adjusted by the user in this controller.

This family of controllers contains additional elementary function blocks that can be used to configure ratio, cascade, feedforward, and other control strategies, which we describe in subsequent chapters. Also, in this family the user can combine an extensive set of logic function blocks with the continuous control function blocks, making possible the configuration of sequential control strategies.

❖ PROCESS CONTROL USING FOUNDATION™ FIELDBUS

A fieldbus (uncapitalized) is a method for digital communication between control devices. Since the advent of digital communication technology, several fieldbuses have been developed. These have varying features and levels of capability, and are generally intended to serve particular segments of the automation industry. The most highly developed fieldbus for process control applications, the FOUNDATION™ Fieldbus,[3] was developed by a consortium of manufacturers known as Fieldbus Foundation. It is one of the fieldbuses approved by the IEC 61158 standard.

FOUNDATION™ Fieldbus (hereafter abbreviated as FF) is more than a digital communication technology. The standard also includes the definition of function blocks that make it possible to distribute the control strategy into field devices (Refs. 5-6, 5-7). For instance, a transmitter can send a signal to a valve positioner, which contains a PID algorithm. Both of these devices communicate over the same two-wire digital network to a host computer, which provides the human interface. Figure 5-17 shows one segment of an FF installation.

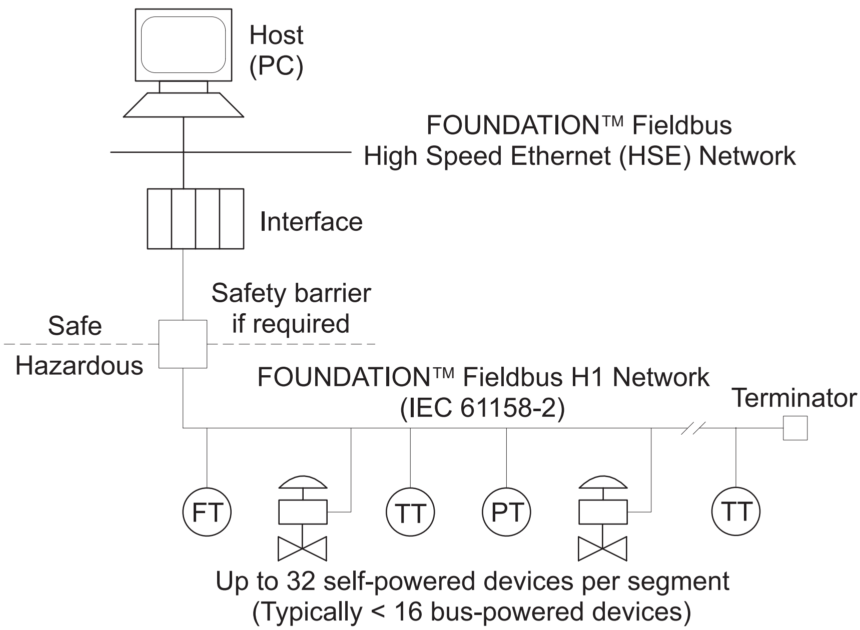


Figure 5-17. FOUNDATION™ Fieldbus Architecture

Among the cited benefits of this approach are these:

- Lower installation costs, since the amount of wiring, conduit, and marshaling panels required is reduced.
- Lower capital equipment cost, because the system provides single-loop integrity, thus providing high availability without the need to purchase redundant components.

3. FOUNDATION™ is a trademark of the Fieldbus Foundation, Austin, Texas.

- Less time spent in engineering, as a result of the standardized approach.

- Less time spent in commissioning, since there are fewer wiring terminations.

- Reduced maintenance costs, since the field devices provide extensive diagnostic information regarding their own health, combined with the fact that much of the maintenance can be done from the control room, thus avoiding (in many cases) the need to send a technician to the field.

- Interoperability. Since every manufacturer is required to adhere to strict standards, devices from various manufacturers can co-exist on the network.

- Limited interchangeability. Also a benefit of the standardization; the devices of one manufacturer can be replaced with the devices of another. (However, if a manufacturer has added enhanced features, another manufacturer's product may not provide those same enhancements. At the least, some reconfiguration effort may be required.)

- Fewer spare parts and devices required, due to the interchangeability feature.

- Improved control. Although it is primarily the control strategy itself and the tuning of the controllers that determine the quality of control, not the physical residence of the PID algorithm, there are control improvement benefits that can accrue due to a number of FF features. For example, FF can provide true reset windup protection on every loop, originating from the actual position of the valve. If a DCS were used, obtaining position feedback from the valve would require the positioner to have a 4–20 mA output card and the DCS to have an additional AI point, plus the wiring between them. Although possible in theory, this was rarely done in practice, due to the additional expense. Other features providing control improvement are status determination (e.g., good/bad/uncertain signal validity); increased accuracy due to the elimination of D/A and A/D converters in transmitters, controllers, and positioners; and the availability of secondary measurements from devices, such as process temperature, static pressure, and density.

There are numerous sources of information regarding the communication, engineering, configuration, and installation aspects of FF. We will briefly cover here only the process control aspects of FF and the basic regulatory control strategies available using FF. Subsequent chapters will introduce various types of advanced regulatory control strategies, and each will close with a discussion of the FF application considerations for that particular control strategy.

The Fieldbus Foundation document FF-891, Part 2 (Ref. 5-8) defines ten function blocks:

AI	analog input
DI	discrete input
ML	manual loader
B	bias/gain

CS	control selector
PD	proportional-derivative controller
PID	proportional-integral-derivative controller
RA	ratio
AO	analog output
DO	discrete output

FF-892, Parts 3, 4, and 5 (Ref. 5-9) define additional function blocks. These include the following function blocks, which can be used in the advanced regulatory control strategies discussed in subsequent chapters:

IS	input selector
LL	lead-lag
DT	dead time
CA	calculate
AR	arithmetic

The definitions are very explicit about the way function blocks interact with other function blocks, even with function blocks in devices provided by different manufacturers. This includes block linking, status passing, block initialization, and the like. For the PID function block, the standard defines sixty-five parameters, including the mnemonic, the exact data structure for each, and the type of access for each. As one example, the CONTROL_OPTS parameter (parameter #13) is a two-byte bit string, accessible only when the function block is in the out-of-service mode. Each of the bits has a defined purpose. For instance, one of the bits specifies whether the block is direct- or reverse-acting.

All manufacturers who provide a PID function block must use this exact parameter definition list. Manufacturers may, however, enhance their products by adding to the defined parameter list. Since the standard does not specify a mathematical formulation for the PID algorithm itself, manufacturers are free to choose a form or add to the defined parameter list to provide optional forms of the PID algorithm. In their standard PID function block, Smar (Ref. 5-10) implements the "ISA" algorithm (called earlier in this chapter the "noninteractive" form) with derivative on PV. In contrast, Fisher-Rosemount (Ref. 5-11) gives the user a configuration option of the "standard" (called "noninteractive" in this chapter) or "series ("interactive") form, with choices of PID action on error, PI action on error—D on PV or I action on error—PD action on PV. This is determined by the setting of parameter #73, MATH-FORM, which is not one of the sixty-five parameters in the standard definition. On the other hand, Smar also offers an advanced PID (APID) function block that provides algorithm configuration options similar to Fisher-Rosemount, determined by the setting of parameter #76, PID_TYPE.

These are merely examples of manufacturer-to-manufacturer differences in areas beyond the standard FF definition. They do, however, highlight the fact that if you are using enhancements to a function block beyond the FF standard definition, and you replace the device with another manufacturer's device, the enhanced features on the original device may not be present on the replacement device. Or you may have to make some changes to the configuration to utilize similar features on the replacement device. To this extent, the devices from the two manufacturers may not be truly interchangeable.

♦ FF Function Block Classes

There are four subclasses of function blocks (Ref. 5-6):

- Input class
- Control class
- Calculate class
- Output class

Input class blocks connect to sensor hardware via an input transducer block over a hardware channel.[4] Control class blocks perform closed-loop control and have back-calculation functionality to provide bumpless mode transfers and reset windup protection, among other features. Calculate class blocks perform auxiliary functions required for control or monitoring, but do not support the back-calculation mechanism. Output class blocks connect to actuator hardware via output transducer blocks over a hardware channel and support the back-calculation mechanism.

♦ FF Basic Control Strategy

A basic control strategy configured from standard FF function blocks is shown in Figure 5-18. This does not appear to be all that different from Figure 5-14, except that the AI function block must be in a transmitter and the AO function block must be in a valve positioner. The PID function block can be in the transmitter, the valve positioner, or in some other device. Its location will have an implication on loading of the communications network, however.)

The AI block receives its input from a transducer block. It filters and scales the input; performs any required calculations, such as square-root extraction; and passes the value to its OUT parameter in engineering units (e.g., 307ºF). Since every function block has a number of modes, including out-of-service (O/S), automatic (AUTO), manual (MAN), and others, the AI block can be placed in the MAN mode and a simulated measurement value entered at the host.

In the basic regulatory control strategy, the PID block receives an operator-entered set point. Its IN parameter, the process variable, is linked to the OUT parameter of the AI block. The input is filtered, scaled, and passed to the PID algorithm and also to alarm detection. The output of the PID algorithm is scaled, passed to output selection (manual/automatic switch), then to output limiting, and then to the block's OUT parameter. The BKCAL_OUT parameter will be used in cascade control schemes, which are presented in chapter 9.

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4. A transducer block is the device-specific interface between the physical measurement and the standard AI or AO block. It is not considered to be a part of the control strategy configuration.

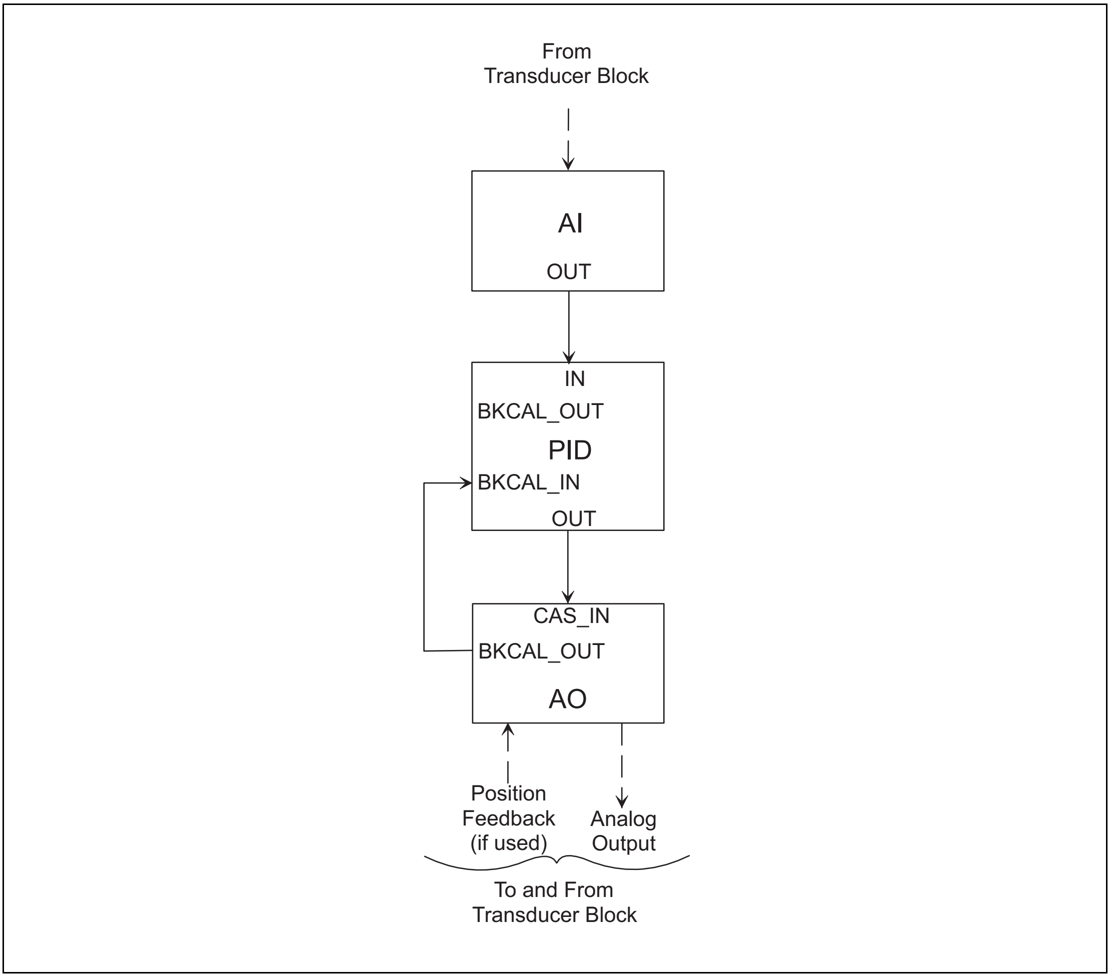

Figure 5-18. Feedback Control Strategy Using FOUNDATION™ Fieldbus Function Blocks

All parameters that are passed from one block to another are also passed with appended status bits. The configuration of the receiving block determines the action to be taken in the event of abnormal status. For example, the block may revert to an initialization mode (IMAN) or hold its output at the last value.

The AO block receives its cascaded set point (CAS_IN) from the OUT parameter of the PID. (Note that the input to the AO block is considered as its "set point"; since this is originating from another block, this is considered as "cascade." (This is slightly different from conventional instrumentation terminology. The AO block can be considered as a "servo positioning PID controller" where the desired valve position is the SP and the actual valve position is the PV.) This signal is scaled, rate-limited (if configured), and passed to the transducer block for valve actuation.

Normally, the scaled and limited set point is passed to the AO block BKCAL_OUT parameter. This is linked to the BKCAL-IN parameter of the PID block. Suppose the loop is opened and the valve positioned by hand, say, by someone using the local interface on the positioner itself to set the AO block to Auto, rather than its usual Cas mode. This is reported to the PID through the BKCAL_OUT – BKCAL_IN link. If the I/O option "Use PV for BKCAL_OUT" is set in the AO block, the process variable (actual stem position) is used for the BKCAL-OUT parameter, and the status bits of BKCAL_OUT force the PID block into an initialization mode (IMAN). Not only does this assure bumpless transfer when the valve is taken out of "hand" operation, but it also prevents windup in the event that the valve stem is limited, either physically or in software.

There are many additional features of the FF function blocks, but they are beyond the scope and space limitations of this book.

❖ REFERENCES

5-1. F. G. Shinskey. *Process Control System – Application, Design and Tuning*. 4th ed., McGraw-Hill. 1996.

5-2. C. L. Smith. *Digital Computer Process Control.* Intex Publishers, 1972.

5-3. Samuel M. Herb. *Understanding Distributed Processor Systems for Control.* ISA,1999.

5-4. Honeywell Automation and Control Solutions. *HPM Control Functions and Algorithms.*

5-5. Siemens Energy and Automation. *Model 354N Universal Loop Controller User's Manual.*

5-6. J. Berge. *Fieldbuses for Process Control: Engineering, Operation and Maintenance*. ISA – The Instrumentation, Systems, and Automation Society, 2002.

5-7. Fieldbus Foundation, *FOUNDATION Fieldbus System Engineering Guidelines*. AG-181, Rev. 1.0.

5-8. Fieldbus Foundation, *Foundation Specification: Function Block Application Process*, *Document FF-891, Part 2.*

5-9. Fieldbus Foundation, *Foundation Specification: Function Block Application Process*, *Document FF-892, Parts 3, 4, and 5.*

5-10. Smar, *FOUNDATION Fieldbus Function Block Manual*, February 2003.

5-11. Fisher-Rosemount, *Foundation Fieldbus Blocks, 00809-0100-4783*, Rev. BA.

6

TUNING FEEDBACK CONTROL LOOPS

The power of the PID controller is that it can be adjusted to provide the desired behavior on a wide variety of process applications through the judicious choice of one, two, or three parameter values, and with only modest knowledge about the process. Determining acceptable values for these parameters is called "tuning" the controller.

In the process industries, the person who tunes the controller often faces a number of adverse factors. The process dynamics are usually not well known; they probably change with operating conditions; there is often an unwanted signal component (called "noise") on the measurement; the loop may be subject to random load changes; and frequently the interaction between control loops makes it difficult to discern the tuning effects of a particular loop from the interactive response with other loops. Furthermore, the loop tuner often must work on an ongoing process, which allows for only minimal or no experimentation or testing. Given these adversities, the wonder is that so many PID loops provide more or less satisfactory performance. On the other hand, it is probably true that the tuning could be improved for a significant number of all control loops.

In this chapter we will explore both theoretical and practical concepts behind controller tuning. The primary techniques to be covered are these:

- trial-and-error tuning
- tuning from open-loop test data
- tuning from closed-loop test data
- improving "as found" tuning (also called "*intelligent* trial-and-error tuning")

If any one of these techniques were clearly superior to the others, there would be no reason to discuss the others. However, each of these techniques has both advantages and disadvantages. Gaining an understanding of each will therefore provide tremendous insight into the tuning task.

As a separate topic, we will also discuss the tuning of liquid-level control loops. The reason for this special coverage will be made clear in that section.

❖ PERFORMANCE CRITERIA

One decision that must be made very early in the loop-tuning procedure is a criterion for acceptable performance. Often this criterion specifies the decay ratio following a set point change. The traditional definition of decay ratio is the ratio of the deviation from set point at the second peak after a set point change to the deviation at the first peak. This is depicted in Figure 6.1a. Occasionally, the set point response is such that this definition is not useful. A better definition of decay ratio is the ratio of the difference between the second peak and its succeeding valley to the difference between the first peak and its succeeding valley.[1] This is depicted in Figure 6.1b. This definition, though more cumbersome, will work in all cases. Most of the time, however, the simpler and more widely used definition depicted by Figure 6.1a will suffice.

The decay ratio can also be defined for a disturbance or load upset. For a step change in load, the behavior depicted in Figure 6.1c is typical. Here, the decay ratio must be determined by the ratio of peak-to-valley differences. A load upset response like that depicted in Figure 6.1d is somewhat unusual for most processes, but is typical of the load upset response of level control loops.

One well-known criterion for controller tuning is a decay ratio of one-fourth following a set point change. This is also called "quarter-wave decay," "quarter-wave damping," and "quarter-amplitude decay." This criterion states that if a loop is oscillating, each peak deviation should be only one-fourth of the previous peak deviation on the same side of set point. This is equivalent to stating that on each half cycle, the amplitude of deviation should be decreased by approximately one-half, making the total decrease one-fourth for a full cycle.

A control loop that is tuned for quarter-amplitude decay (see below) following a set point change (Figure 6.1a) will respond somewhat sluggishly to a load change (Figure 6.1c). On the other hand, if a loop is tuned to give quarter-amplitude decay response to a load upset as shown in Figure 6.1d, then it may be too oscillatory for a set point change.

Intuitively, a loop that exhibits quarter-amplitude decay appears to be well tuned. The decay ratio alone, however, does not provide a complete performance specification. For a given process, two different combinations of tuning parameters can both produce quarter-amplitude decay, yet the period of oscillation can be considerably different between the two.

For many applications, the quarter-amplitude-decay criterion provides acceptable damping of the oscillations that follow a set point change. For other applications, quarter-amplitude decay may be too oscillatory. A plant's operations and engineering staff may be willing to accept a more sluggish response to a load upset in order to minimize the overshoot that follows a set point change. Some operators prefer to see loops that are critically damped; that is, they want the measurement to rise rapidly to the new set point value and yet avoid overshoot.

---

1. This second definition is more defensible theoretically, since the set point response is the composite of a filtered exponential rise and a damped sinusoidal signal. If the rise time of the exponential is sufficiently fast, then the two definitions are essentially the same. This is why the first definition is valid in most circumstances.

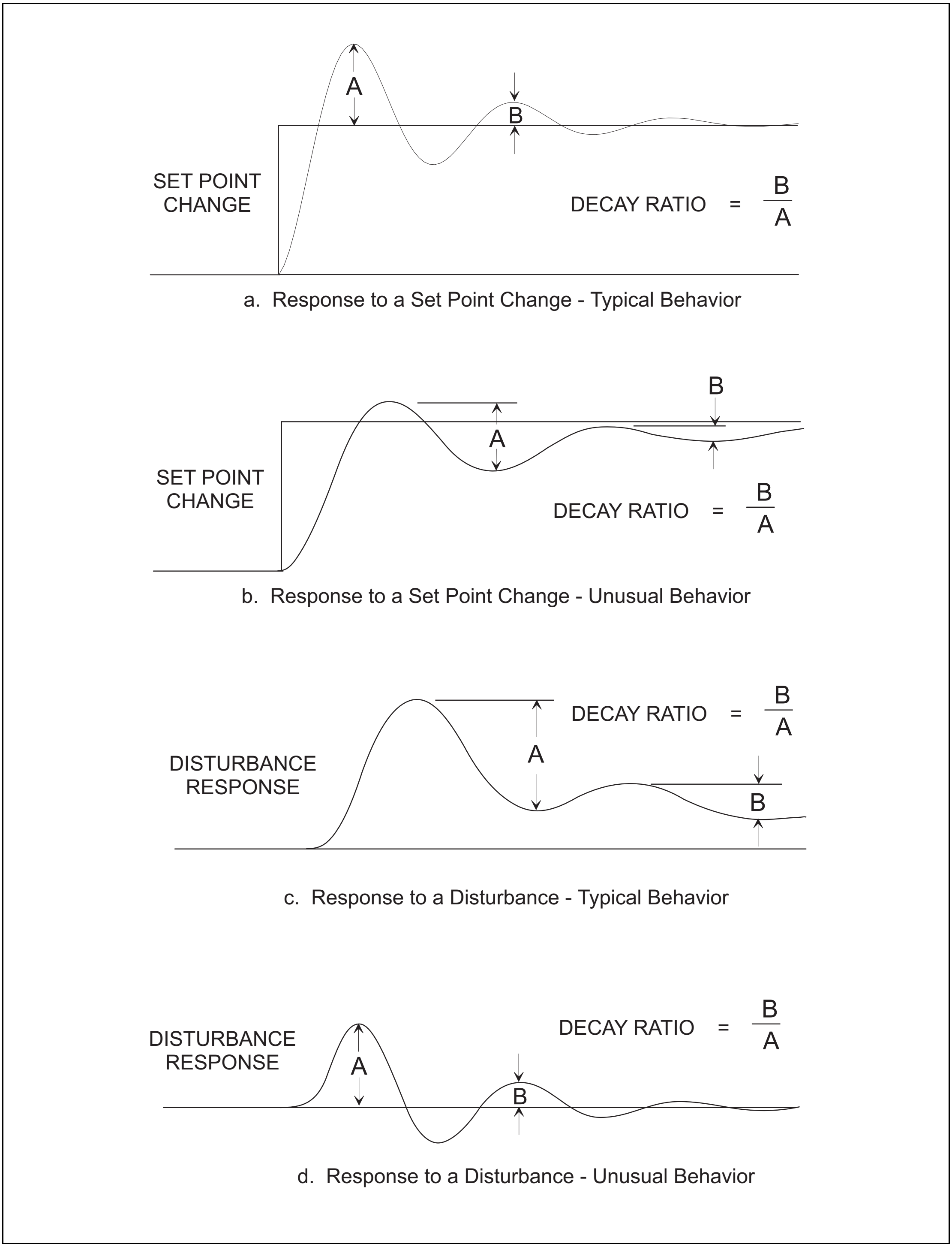

Figure 6-1. Quarter-Amplitude Decay Responses

Also, for many control loops, the set point is rarely changed. The purpose of these loops is to minimize the effect of disturbances. Even so, because set point changes are usually made more easily than load changes, in actual practice many loops are tuned for a suitable response to a set point change. Then, the resulting response for a load upset is accepted even though it may not be the best. A preferable tuning strategy would be to tune the controller for best response to a load change, then use one of the set point "softening" techniques described in chapter 5 to ameliorate the effect of occasional set point changes. If actual load changes cannot be made, the effect can be simulated by placing the controller in manual, changing the controller output, then quickly returning the loop to automatic.

Other criteria that are sometimes used to specify, or measure, control loop performance include *rise time*—the time between a set point change and the first crossing of the set point; *overshoot ratio*—the ratio of the magnitude of the first overshoot above set point to the magnitude of the set point change itself; *settling time*—the time, following a set point change or disturbance, that it takes for the oscillation to become so small that the deviation does not exceed some specified amount.

Still more criteria that can be used are based on minimizing the integral of some function of the error. The following four integral-error criteria can be considered:

Integral of the absolute value of the error (IAE):

$$IAE = \int |e| dt \tag{6-1}$$

Integral of the square of the error (ISE):

$$ISE = \int e^2 dt \tag{6-2}$$

Integral of time x absolute value of the error (ITAE):

$$ITAE = \int t |e| dt \tag{6-3}$$

Integral of time x square of the error (ITSE):

$$ITSE = \int t e^2 dt \tag{6-4}$$

Note that the simple integral of the error is not a valid criterion since the integration of a positive error would be canceled out a half-cycle later by integration of a negative error. The four criteria just listed avoid this, either by taking the absolute value of the error or by squaring the error.

Minimizing each of these integral-error criteria will produce a different response. For example, a plant that uses the ISE criterion pays an increasingly large penalty as the magnitude of the error increases. Hence, for a given loop the ISE criteria will result in a smaller maximum deviation value than the IAE criterion, but it may cause the oscillation to persist longer.

The rationale for the last two criteria listed (ITAE and ITSE) is that the longer an error persists after a set point or load change, the more heavily it should be penalized. Thus, the ITAE criterion will permit a greater initial deviation than an IAE criterion, but it will force the oscillation to die out sooner.

If there is noise on the process variable, any of these criterion will increase without bounds. To be valid as measures of performance, the same time span should be used for integration in any cases to be compared.

Integral-error criteria are useful for theoretical and control simulation studies as well as to provide insight into the tuning process. They may also be used in control loop audits. They are rarely used in actual control loop tuning, however.

In summary, what constitutes good tuning is often a subjective matter that can vary from application to application—as well as from person to person tuning the loop.

❖ TUNING FOR SELF-REGULATING PROCESSES

Because self-regulating and non-self-regulating processes have a different character, the tuning procedures that are applicable to one may not be applicable to the other. In this section, we will restrict our comments to self-regulating processes. The next section is devoted entirely to tuning controllers for non-self-regulating processes, primarily liquid-level control.

♦ Trial-and-Error Tuning

Most loops are tuned by an experimental technique. Even when a formal technique, such as open-loop or closed-loop testing, is used to determine the initial tuning, a final bit of fine tuning may be in order. Trial-and-error tuning requires the user to observe the response of the loop to a previous event, either a set point change or load change, then decide what tuning parameter (or parameters) should be changed, in which direction, and by how much. Experience helps in interpreting the response. The user must also have a thorough understanding of the effect of changing each of the tuning parameters.

Response to Various Tuning Parameter Combinations

For a self-regulating process controlled by a PI controller (interactive or noninteractive), the tuning map on Figure 6-2 depicts the response to a set point change for various combinations of proportional and integral tuning. The graph in the upper left-hand corner (graph Al) depicts the closed-loop response to a set point change when the controller is tuned with very low gain (wide proportional band) and no integral action. If integral action cannot be turned off, then the left-hand column represents minimum integral action, the largest possible value for minutes per repeat, or the smallest possible value for repeats per minute.[2] Graph A1 shows an overdamped response—no oscillation—resulting in a significant steady-state deviation from set point.

As the gain is increased (proportional band is decreased), move down the left-hand side of the tuning map. As you do, you'll note first that there is a slight tendency to overshoot, then a greater tendency to oscillate, with a corresponding reduction in steady-state offset. Finally, if the gain is increased too much, the loop is driven to a state of undesirable oscillation (graph A4 in Figure 6-2), and the question of steady-state offset is a moot point—the loop is never in a steady state.

Note that going from a low gain (wide proportional band) to a high gain (narrow proportional band) is a move in a *destabilizing* direction.

Now return to the graph in the upper left-hand corner. We will maintain the low gain. This time, we will increase the speed of the integral action from very slow (or none) to fast integral action. We do this by decreasing the minutes per repeat setting or by increasing the repeats per minute.

Initially, the integrator within the controller, sensing the steady-state error that would otherwise exist, gradually changes the controller output to reduce the steady-state error to zero. The process response is as if the "tail of the curve" were bent slowly upward, until the error is reduced to zero. This is depicted in graph B1 in Figure 6-2.

If the speed of the integral action is increased further, the tail of the curve is bent upward so rapidly that overshoot results (see graph C1). An even further increase in the integral action will create very undesirable oscillation of the loop, as shown in graph D1. Thus, we can state that an increase in the speed of response of the integral action is a move in a *destabilizing* direction.

At this point, a normal question might be, "Don't we want the loop to be as stable as possible at all times? Why don't we just set both the gain and the integral time to a minimum value?" If we truly wanted maximum stability for a control loop, we could put the controller in manual. There would then be no way to compensate for the effect of disturbances. Thus, we don't want maximum stability; we want reasonable stability plus the ability to compensate for disturbances.

We have shown that if the gain is too large or the integral action is too fast, the control loop will be driven into unstable oscillation. However, for a given process, the character of the oscillations will differ. Oscillation that is due primarily to excessive gain (e.g., graph A4 of Figure 6-2) will show a relatively high frequency and low amplitude when compared with oscillation caused primarily to the integral action being too fast (e.g., graph D1). We will utilize this fact later when we discuss a technique for improving existing tuning values.

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2. Minutes per repeat and repeats per minute are the most widely used measures of integral action. In general, we will speak of minutes per repeat in this book. Some manufacturers as well as the Fieldbus Foundation use "seconds per repeat" for integral action and "seconds" for derivative time.

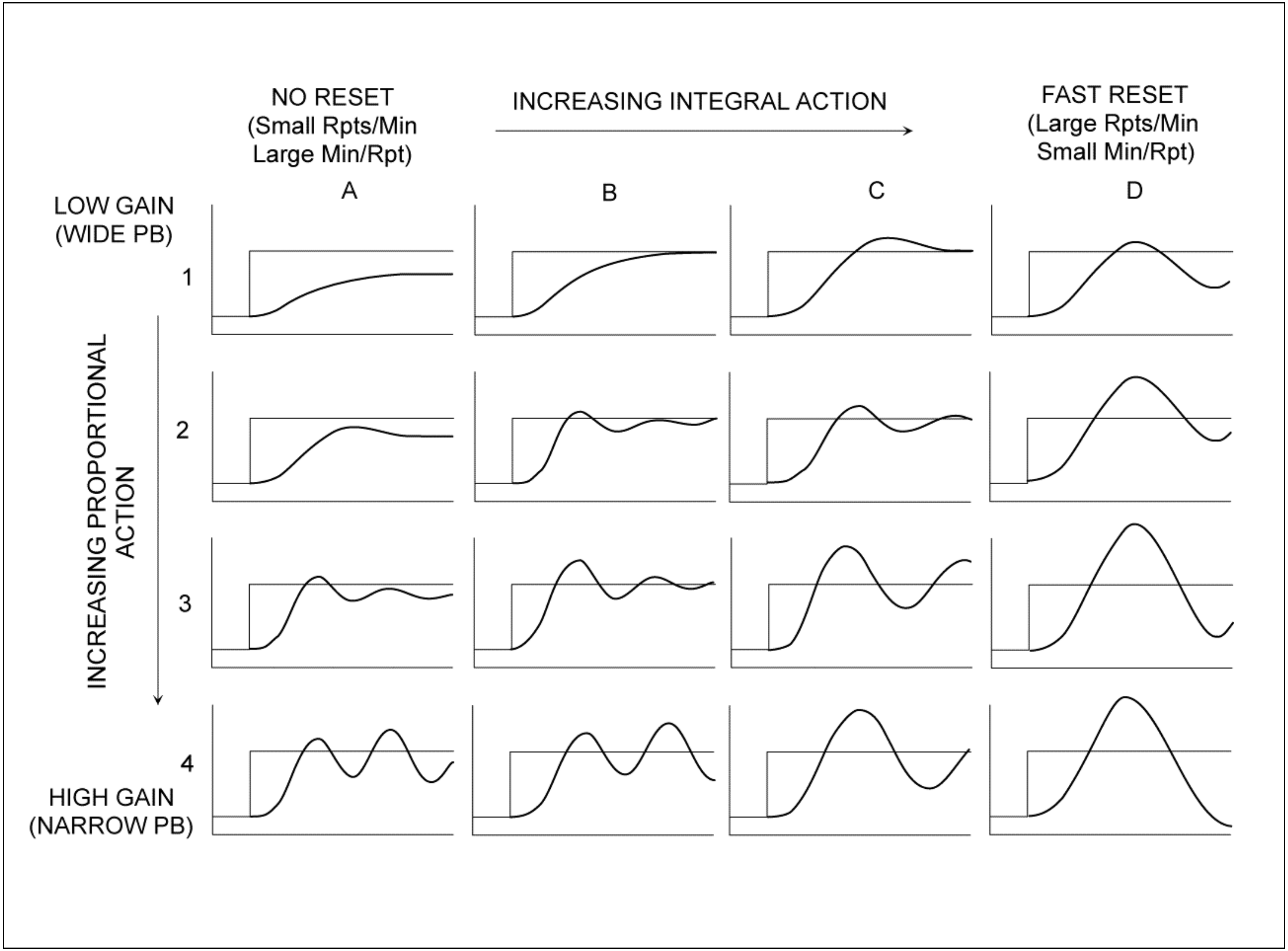

Figure 6-2. Tuning Map for Proportional and Integral Effects

The rest of the graphs on the tuning map in Figure 6-2 depict various combinations of gain and integral action. Note that the graphs in the center of the map—in particular, graphs B3 and C2—depict somewhat similar responses. This illustrates the fact that a range of combinations of gain and reset will produce approximately the same response. It also illustrates that a trade-off can often be made between gain and reset. If, for some reason or other, the controller gain is increased—a move in the destabilizing direction—compensation can be made by decreasing the integral response—a move in the stabilizing direction. For example, move from graph C2 to graph B3. The benefit of this move is an improved response to a disturbance in the loop but at the expense of slightly more oscillation.

The tuning map, Figure 6-2, does not show the effect of adding the derivative mode to the controller. If a graph could be drawn in three dimensions, we would add a third dimension showing the effect of derivative. It would start with no derivative on the tuning map and extend to an excessive use of derivative, for all combinations of gain and integral action.[3] Then we could illustrate the fact that, for a self-regulating process, incorporating derivative is a move in the *stabilizing direction.*

Since adding derivative has a stabilizing effect, we can increase the gain and/or increase the speed of the integral response, both of which have a destabilizing effect. However, larger gain and/or faster integral action will cause more control action in response to a load upset, and hence will reduce the maximum deviation from set point as well as reduce the settling time. This is the primary motivation for using the derivative mode in a controller. Deterrent factors for the use of derivative include noise on the process measurement, and the difficulty of tuning one additional parameter.

Consider the following scenario. Suppose we have adjusted gain and reset, with no derivative, to produce acceptable response, as shown in Figure 6-3, graph B1. Now add a small amount of derivative. This stabilizing move allows us to increase the gain and integral response. Suppose that we repeat this procedure until we have a good response using gain, integral, and derivative. We note that for the set point change, both the maximum deviation and the settling are reduced, as illustrated in graph B2.

The question occurs to us: Did we just “luck out” and find a better combination of gain and reset, or is it really the derivative that is helping us? To answer this question, suppose that we keep the same gain and reset as used in graph B2 of Figure 6-3 and take out the derivative. By removing it, we can see how much it was aiding us. The response will be as shown in graph A2.

Comparing graphs B1, B2, and A2 of Figure 6-3 will give us an intuitive feel for the effect of derivative. With well-tuned PI control, the proportional and integral modes move the final control element to bring the measurement to set point (graph B1). With derivative added, higher gain and faster integral action are permitted so control action is greater; that is, there is

3. In the text that follows, we assume that we are using a noninteractive controller with derivative mode on measurement. If we are using an interactive controller, then the terms *gain*, *reset*, and *derivative* will refer to *effective* values of these tuning parameters (see Equations 5-8 – 5-10).

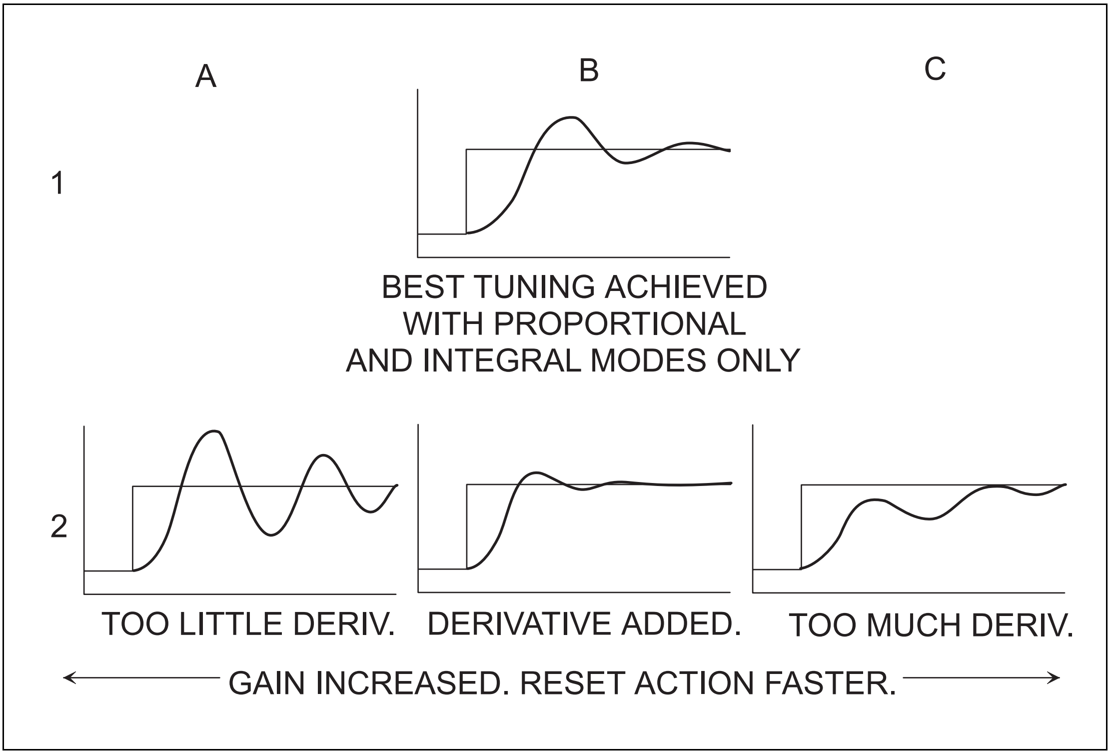

Figure 6-3. The Effect of Adding Derivative

more movement of the final control element. This tends to “push” the measurement to the set point faster (graph B2).

The measurement can be pushed to set point faster because something analogous to “brakes” are operating on our control system, namely, derivative action. Recall from chapter 4 that the contribution of the derivative mode is proportional to the rate of change of the measurement (or error) and that its direction is opposite that of the measurement’s movement.

If we “throw out our brakes” (graph A2 of Figure 6-3), then we have only the magnitude, sign, and integrated history of the error to arrest the measurement’s rate of movement. The result is a large deviation from set point and severe oscillation.

What if we retain the “good” setting for gain and integral that we had in graph B2, but increase the derivative even more? The response will be as shown in graph B3 of Figure 6-3. If we apply too much derivative action, the measurement oscillates without crossing the set point, and the controller output shows an excessive amount of activity.

Trial-and-Error Tuning Procedure

The purpose of the previous scenario was to give you an insight into the effect of making an adjustment to each tuning parameter. However, it was not intended as a recommended procedure that should be followed to tune an actual control loop. If a new control is to be tuned by trial and error, the procedure given below can be followed. Variations in the procedure are noted for different forms of the PID control algorithm.

- Put the controller in manual and adjust the controller output until the measurement is stable and near the normal operating value.
- Take out all reset and derivative action.
- Set the gain to a low value (proportional band to a wide value).
- Put the controller in automatic. Test the process by changing the set point or by making a process load change.
- Adjust the proportional gain until you observe a slight oscillation. Repeat this step until you observe the desired damping characteristics (for instance, quarter-amplitude decay). Do not be concerned at this point by an offset between set point and measurement, since without integral action the offset will most likely be steady state.

When you have achieved the desired damping characteristics by using proportional control alone, increase the integral action until the measurement lines out at set point.

If you are using either a noninteractive or interactive form of PID, you can determine a reasonable initial value for reset by measuring the period of oscillation (in minutes) using proportional control alone. Set the following:

Integral time, minutes per repeat = 0.67 times the period or
Reset rate, repeats per minutes = 1.5 / period

If you are using an “independent gains” form of PID, you can determine a reasonable initial value for the integral gain from the following calculation:

$$K_I = \frac{1.5\ K_P}{P} \tag{6-5}$$

where K_P represents the proportional gain and P is the period, the time between two successive peaks of the oscillation.

The proportional gain must then be decreased or the proportional band increased to compensate for the destabilizing effect of adding reset. Hence:

Decrease proportional gain to 90 percent of its former value, or
Increase proportional band by 110 percent of its former value.

Test the process and make fine-tuning adjustments if necessary. From here, it will probably be better to adjust the gain to achieve the desired damping characteristics.

Whether or not you can use derivative depends on the amount of noise on the measurement, as well as the degree to which it is filtered. If derivative is to be used, the procedure will vary significantly, depending upon the form of PID control that you use. In all cases, it is assumed that the derivative mode acts on measurement only.

If the controller is a noninteracting PID then:

- Increase the derivative setting. The initial value should be about one-tenth the integral time (minutes per repeat). Depending upon the amount of noise on the measurement, this may or may not produce undesirable chatter of the controller output. If it does, reduce the derivative. If it does not, you can increase the derivative up to a maximum of about 15 percent of the integral time determined for PI control alone.

  After adding derivative, increase the gain by 25 percent of its former value, and decrease the integral time to about two-thirds of its former value.

- If the controller is an interacting PID then:

- Increase the derivative setting. The initial value should be about one-tenth of the integral time (minutes per repeat). Depending on the amount of noise on the measurement, this may or may not produce undesirable chatter of the controller output. If it does, reduce the derivative; if not, you can increase the derivative, up to a maximum of about 30 percent of the integral time determined for PI control alone.

- After adding derivative, decrease the integral time from the value determined by PI control alone by twice the amount of the derivative time. (For example, if the original integral time was ten minutes per repeat, and you set the derivative to three minutes, then the new integral time should be 10 - 2 x 3 = 4 minutes.)

- Do not change the gain. Since the derivative mode is on measurement, the initial response to a set point change will be the same as for a PI controller. For a disturbance, however, the interactive algorithm will provide a higher effective gain than the actual value entered. This will provide an improved response to the disturbance.

If the controller is an "independent gains" form of PID then:

- Calculate the maximum allowable derivative gain for a noise-free process from this equation:

$$K_{Dmax} = \frac{0.2K_P^2}{K_I} \qquad (6\text{-}6)$$

where K_P and K_I are the values of proportional and integral gain used with P_I control. Start with a smaller value, say about one-half of the maximum. The derivative gain can then be either increased or decreased, depending on how much valve movement is being caused by measurement noise.

As derivative is added, the proportional gain, K_P, can be increased by up to 133 percent of its prederivative value, and the integral gain, K_I, can be increased to 220 percent of its prederivative value.

As a practical matter, you should never change a tuning parameter by more than 50 percent of its former value before performing a new process test to determine its effect. If the process response is already "in the ballpark" then adjustments of 25 percent, or less, between process tests are more reasonable. On the other hand, an adjustment of 5 percent or less will probably not have a big enough effect to justify the time spent making an additional process test.

The procedure just outlined can be followed if a loop is to be tuned by trial and error for the first time. Often, however, the challenge is not that of initial tuning, but of improving existing tuning. This is discussed later in this chapter.

♦ Tuning from Open-loop Test Data

Formal methods of tuning involve testing the process and extracting data from the tests that will permit tuning parameters to be calculated directly. Two types of tests are the open-loop and closed-loop tests.

In the open-loop test, the controller is placed in manual, and controller output is adjusted until the measurement is near the normal operating point. Then, the controller output is changed in a step fashion. Parameter values for a simple process model are then determined from the process response to this step change. The theoretical step response of this simple process model should approximate the response of the actual process.

For most self-regulating processes the response to a step change in process input (controller output) is an "S-shaped curve" that initially rises very gradually, perhaps following some delay, and then rises more rapidly, followed by a gradual rise to equilibrium. This type of process response can usually be approximated by a first-order-plus-dead-time (FOPDT) model, as shown in Figure 6-4.

Three parameter values are required:

Process gain	K_p
Process time constant	τ
Dead time (delay)	θ

Once these three parameter values are determined, one can use the equations in either column 2, 3, or 4 of Table 6-1 to determine controller tuning values for the modes of control, P, PI, or PID, that will be used. Table 6-1 and Table 6-2 both assume that a noninteractive controller with derivative mode on measurement, with other modes on error, will be used. If the interactive form of the PID controller is used, then the tables give *effective* values for the controller

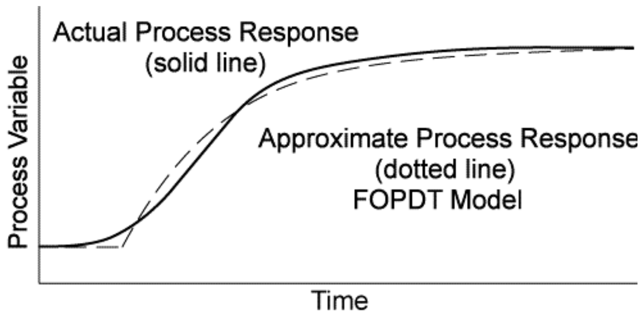

Figure 6-4. Actual and Approximate Process Responses to an Open-loop Test

tuning parameters. These must then be converted into *entry* values using Equations 5-11 through 5-14.

Table 6-1. Controller Tuning Equations Using Open-loop Data

Controller Type	Ziegler-Nichols (Reference 6-1)	Cohen-Coon (Reference 6-2)	IAE, Disturbance (Reference 6-3)
Prop Only (P) Gain (K_C)	$\frac{\tau}{K_p \theta}$	$\frac{\tau}{K_p \theta}\left(1+\frac{\theta}{3\tau}\right)$	$\frac{0.902}{K_p}\left(\frac{\tau}{\theta}\right)^{0.985}$
Prop + Integral (PI)			
Gain (K_C)	$\frac{0.9\tau}{K_p \theta}$	$\frac{0.9\tau}{K_p \theta}\left(1+\frac{\theta}{11\tau}\right)$	$\frac{0.984}{K_p}\left(\frac{\tau}{\theta}\right)^{0.986}$
Integral Time (T_I) (minutes/repeat)	3.33θ	$3.33\theta\left(\frac{\tau+0.1\theta}{\tau+2.22\theta}\right)$	$1.645\theta\left(\frac{\tau}{\theta}\right)^{0.293}$
Prop + Integral + Deriv (PID)			
Gain (K_C)	$\frac{1.2\tau}{K_p \theta}$	$\frac{1.33\tau}{K_p \theta}\left(1+\frac{3\theta}{16\tau}\right)$	$\frac{1.435}{K_p}\left(\frac{\tau}{\theta}\right)^{0.921}$
Integral Time (T_I) (minutes/repeat)	$2.0\ \theta$	$2.5\ \theta\left(\frac{\tau+0.2\theta}{\tau+0.6\theta}\right)$	$1.139\ \theta\left(\frac{\tau}{\theta}\right)^{0.251}$
Derivative Time (T_D) (minutes)	$0.5\ \theta$	$0.37\theta\left(\frac{\tau}{\tau+0.2\theta}\right)$	$0.482\ \theta\left(\frac{\tau}{\theta}\right)^{-0.137}$

The success of the open-loop test method depends upon several factors, including how well a first-order-plus-dead-time model actually matches the true process response and how accurately the model parameters are determined.

The size of the step change to the controller output will depend upon local conditions. A larger step change (say, 10%) is desirable to make it possible to clearly distinguish the response to valve change from measurement noise or other disturbances. On the other hand, a smaller change (5% or less) is usually required because of the necessity to avoid making a major upset to the process.

Many correlations for controller tuning parameters have been published. Table 6-1 presents three of these. The second column presents the best known and oldest of these correlations, the well-known Ziegler-Nichols (Z-N) open-loop controller tuning relationships (Ref. 6-1). The calculations are simple (remember, these equations were developed before hand calculators existed!), although they are not necessarily the best performing. The second column presents the Cohen-Coon correlations (Ref. 6-2), which give considerably better control-loop performance at higher values of the ratio θ / τ. The third column presents correlations developed by Lopez et. al (Ref. 6-3) to minimize the IAE criterion for a step disturbance applied at the process input. The process assumed by Lopez was a pure FOPDT.

See Tables 1.21i and 1.21j in Ref. 6-4 for other controller tuning correlations.

The question is often asked, "What form of PID, interactive or noninteractive, did Ziegler and Nichols use to develop their relations?" Since they had available the Fulscope pneumatic controller made by Taylor Instrument Co. (now part of ABB), this is generally believed to have been their target controller. This controller was highly interactive and can be described by a transfer function similar to Equation 5-3, except that the term $(T_I - T_D)s$ is in the denominator. Thus, if the integral and derivative times were set equal, the controller acted as if it had an infinite gain. The question of Ziegler and Nichols' PID is one of only historical interest, however. A more appropriate question is, "To what form of PID controller should the Z-N relations be applied today?" (Note that this question is only applicable if derivative is being used. Otherwise, there is no difference.) A survey of textbooks and articles that list the Z-N relations shows that most do not identify the intended form of PID. A simulation study by the author for a number of process models used the Z-N values directly as well as the converted values (see Equations 5-8–5-14) in both interactive and noninteractive forms of the PID with derivative on measurement. The performance indicated by various measures (IAE, overshoot, decay ratio, maximum deviation for load change) was surprisingly similar. However, one desirable trait was noted when the Z-N relations were assumed to apply to a noninteractive PID, then corrected for use with an interactive PID. This produced the smallest value of the controller output's proportional response to a set point change. On the other hand, if the Z-N relations were assumed to apply to an interactive PID, then corrected for a noninteractive PID, then the largest proportional response to a set point change was observed. On this basis, we recommend that the Z-N relations, given in Tables 6-1 and 6-2, be applied to a noninteracting PID, and converted (Equations 5-11–5-14) for use with an interactive PID.

Once the process achieves a new equilibrium, the process gain can be calculated as follows:

$$K_P = \frac{\text{change in measurement (percent of span)}}{\text{change in controller output (percent)}} = \frac{\Delta x}{\Delta m} \qquad (6\text{-}7)$$

The dead time and time constant can be determined by analyzing the measurement response record. Several different methods are described in the following sections.

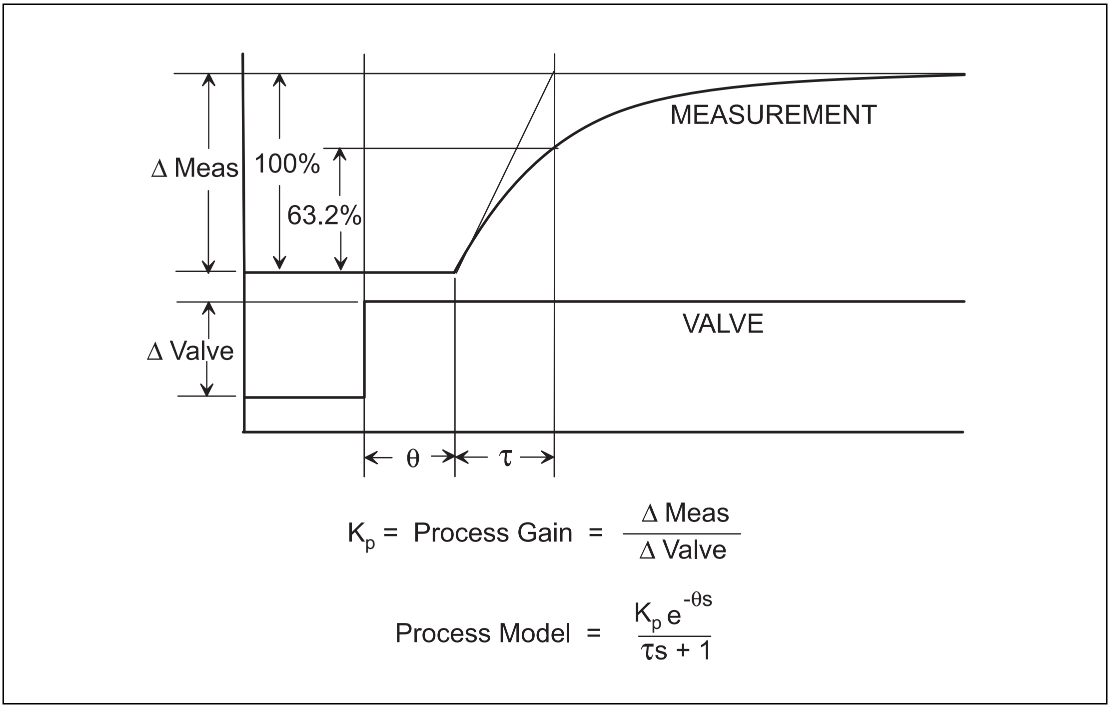

Figure 6-5. Determining Process Parameters for a True FOPDT Process

True First-order Plus Dead Time

If the process can truly be represented by a FOPDT model, the response will be as shown in Figure 6-5. The dead time is obvious. The time constant can be determined in one of the following two ways. For a true FOPDT process the time constants determined by these two methods will be identical.

(1) Draw a tangent at the point of steepest rise—this will be at the beginning of the rise. Continue the tangent until it intersects a horizontal line representing the ultimate equilibrium point of the measurement. The rise time of the tangent is the time constant.

(2) Draw a horizontal line at 63.2 percent of the ultimate change in measurement. The time constant is the time difference from the end of the dead time to the time the process has risen to 63.2 percent of its ultimate value.

Approximate First-order Plus Dead Time

Very few processes, however, will exhibit a true FOPDT response. For most processes, there are an unknown number of lags (places of mass or energy storage) in the system, which may be arranged in an infinite variety of ways. The smaller lags produce an apparent dead time, even if no true dead time is present.

We can proceed by drawing a tangent at the point of steepest rise, as we did previously. The apparent dead time, θ, is represented by the time difference between the time of controller output change and the intersection of the tangent with the lower equilibrium horizontal line. The rise time of the tangent to 100 percent of the measurement change is a value we will call τ_1.

We also draw a horizontal line at 63.2 percent of the ultimate change in the measurement. The rise time of the process, from the end of the apparent dead time to the 63.2 percent, is a value we will call τ_2.

One or the other of two situations may occur, as shown in Figure 6-6.

(1) The rise time of the tangent, τ_1, is greater than the rise time of the process to the 63.2 percent point, τ_2 (see Figure 6-6a). This is a very common situation. It indicates a process that has multiple lags in series, as shown in Figure 3-13 or 3-14.

(2) The rise time of the tangent, τ_1, is less than the rise time of the process to the 63.2 percent point, τ_2 (see Figure 6-6b). This is the situation in which the measurement quickly rises to a plateau, then slowly drifts upward for a long time. It indicates a significant lag in parallel with the other lags (see Figure 3-17 and 3-18). It is less common than the first situation just described.

There appears to be a dilemma regarding the method to be used to determine the apparent time constant, τ of the process; the rise time of the tangent; or the rise time of the process to the 63.2 percent point. However, if we examine Table 6-1, we see that the time constant is only used to calculate the controller gain. Furthermore, the shorter the time constant, the lower the controller gain. For pragmatic reasons, it is better to calculate controller tuning parameters that are too conservative (i.e., produce an overly sluggish closed-loop response) than parameters that are too aggressive (i.e., produce an overly oscillatory closed-loop response). Thus, we should use as the process time constant whichever is the shorter value: the rise time of the tangent or the rise time of the process to the 63.2 percent point.[4] That is:

4. It may come as a surprise that we do not always consider the longest time constant in the process as the dominant time constant. If we have two first-order lags in parallel, in which one has a significantly longer time constant than the other and there are additional lags that create apparent dead time, then we can tune our controller on the basis of the *shorter* time constant of the two lags in parallel. The signal through the longer lag will act as a slow disturbance that can be compensated for by additional feedback control action.

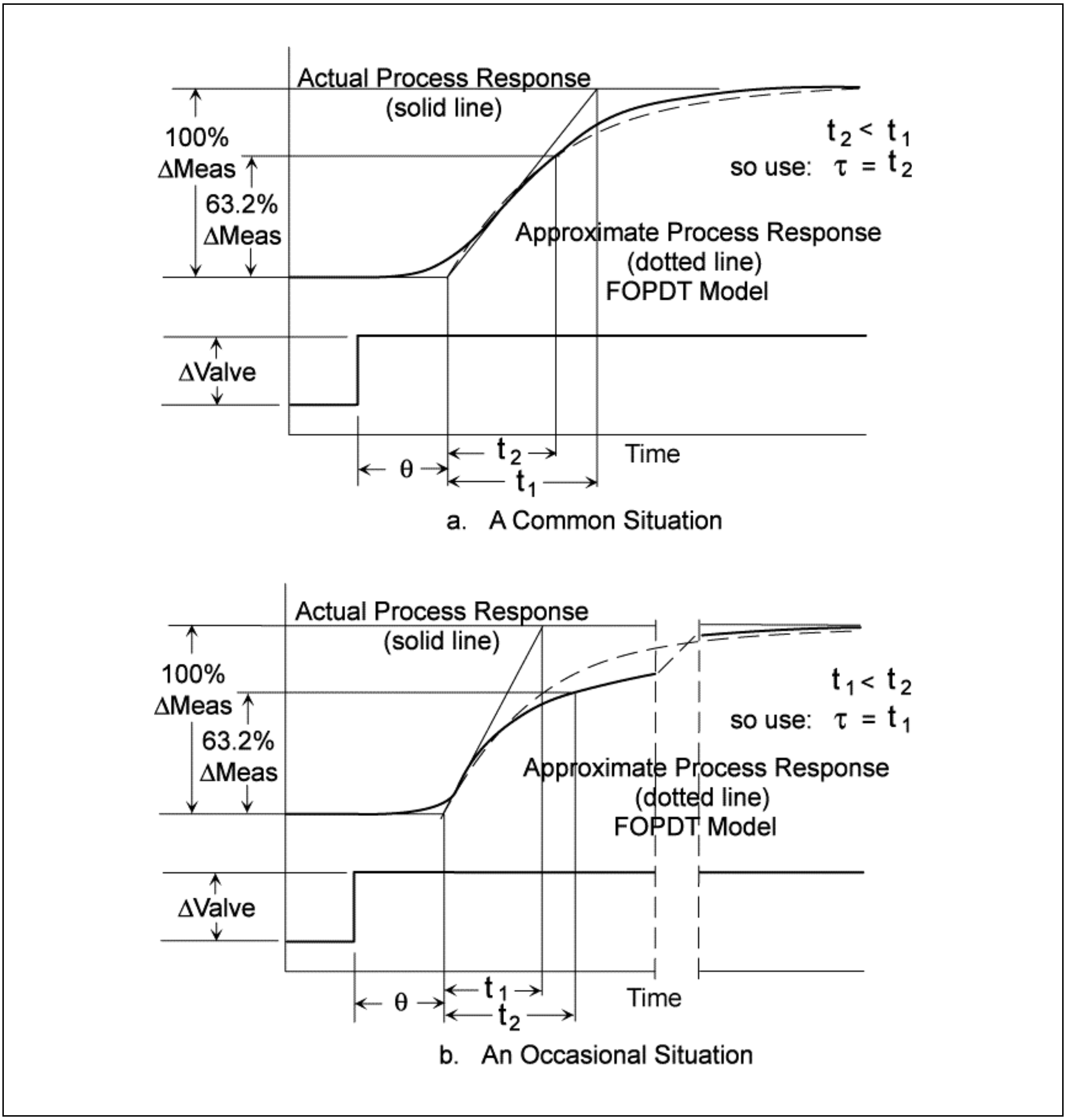

Figure 6-6. Approximating an Open-loop Test Response with a FOPDT Process Model

$$\tau = \text{minimum of } \tau_1,\ \tau_2$$

Indeterminate Process Gain

For open-loop process testing, we have assumed that the test starts with the process in an equilibrium condition and continues until a new equilibrium is reached. The difference in measurement values was used to calculate the process gain, K_p. The situation may occur, however, where the process does not reach a new equilibrium. An example of this is integrating processes found in level-control loops; these are discussed later in this chapter. Another example is where the process may not be a true integrating process, but the new equilibrium value is outside the feasible region of operation, as shown in Figure 6-7. In these two cases, the process gain cannot be determined by graphical analysis. We can, however, obtain sufficient information to allow us to determine controller tuning values.

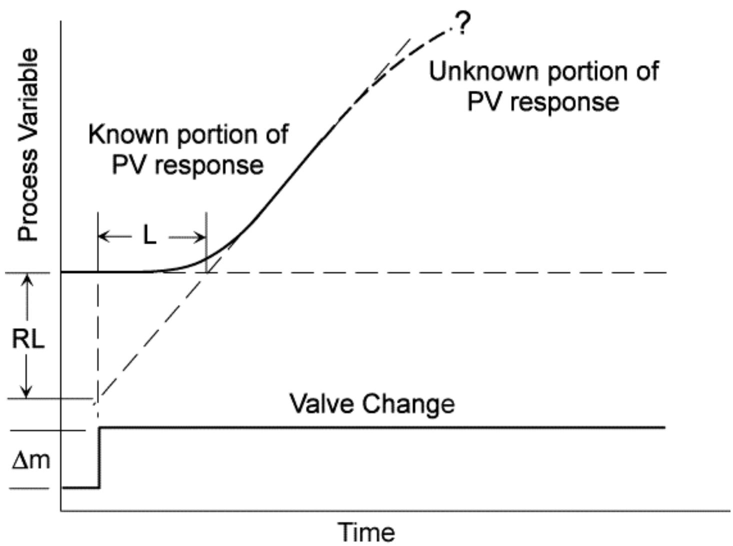

Figure 6-7. An Open-loop Process Response with Indeterminate Process Gain

If there is process response data for only a portion of the way to equilibrium, draw a tangent at the point of steepest rise, and extend this below the initial equilibrium value to the vertical line that represents the time of step valve change. Call the vertical distance "RL," as shown in Figure 6-7. Call the time from the changing of the valve to the intersection of the tangent with the base line "L". Given the values RL and L and the amount of valve change, Δm, calculate the following:

$$\frac{K_p}{\tau} = \frac{RL}{L \times \Delta m} \tag{6-8}$$

$$\theta = L \tag{6-9}$$

Knowing values for q and the parameter group K_p/τ will allow us to enter Table 6-1, column 2 (Ziegler-Nichols) and calculate controller gain, K_C. The process dead time, q, determined graphically, allows us to calculate the integral time, T_I, and also derivative, T_D, if desired.

Determining Process Parameters without Drawing a Tangent

A method for estimating the dead time and time constant without graphically drawing the tangent is described in Figure 6-8.

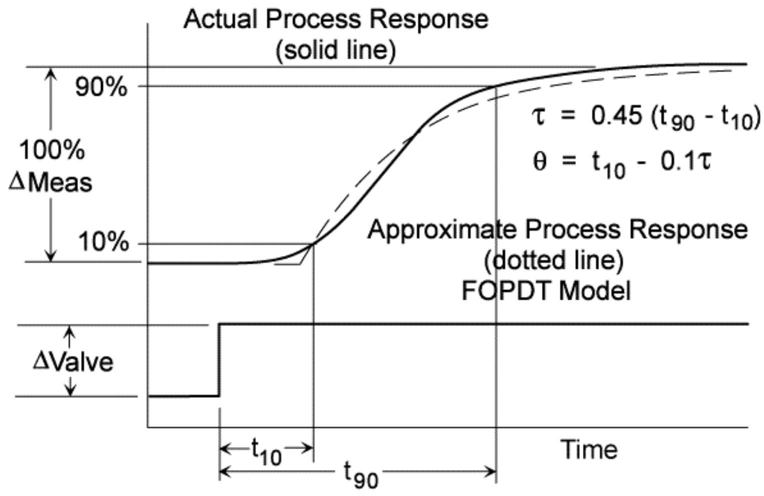

Figure 6-8. Nontangent Method for Approximating an Open-loop Test Response with a FOPDT Process Model

Practicalities

If we are using a graphical method, we may not be able to draw the tangent precisely. If we are using the nontangent method, process noise may prevent us from determining precise values for the 10 percent and 90 percent points. Thus, some uncertainty may exist in the values of the process parameters, particularly the apparent dead time. If so, we should use a longer value for dead time, rather than a shorter value. Examining the equations of Table 6-1 shows that using a shorter value of dead time will result in a higher controller gain and a shorter integral time, both of which are effects that tend to produce oscillation or instability in the control loop. On the other hand, using a longer dead time will result in a lower gain and a longer integral time; these will be more conservative tuning settings. In essence, when using the equations of Table 6.1, we do not want to represent the process as being easier to control than it really is. (This line of reasoning contradicts the frequently heard definition of dead time as "the time from when the valve is changed until movement of the measurement is first detected.")

Hence, some practical advice is appropriate regarding estimating FOPDT model parameters. *If there is any uncertainty:*

- Use a *larger value* for the estimate for process gain;
- Use a *short* estimate for process time constant; use a *long* estimate for process dead time.

Of course, each of the model parameters should be estimated as carefully as possible. Using all of these pragmatic uncertainty rules simultaneously may produce overly conservative controller tuning.

Some processes do not exhibit the same response when the valve position is increased as they do when it is decreased. Thus, it is probably desirable to test the process at least twice, by moving the valve in opposite directions. If you observe a considerable difference in the response, then use the parameter set that gives the more conservative controller tuning. For noisy processes, take the average of four responses.

If the process is controlled by a digital system that executes the control algorithm on a periodic basis, then the loop will have additional dead time equal to one-half the control period. Filtering the measurement adds more dead time, so the total contribution is dead time that is equal to a full-control sampling period. Hence, this amount should be added to the apparent dead time before using the equations of Table 6-1. However, unless the control period is 20 percent or more of the apparent dead time, this correction is not significant enough to be necessary.

The use of Table 6-1 should be limited to situations where the ratio of apparent dead time (corrected if necessary by the control period) to apparent time constant is between 0.1 and 1.0. This will cover most actual applications. If dead time that is less than one-tenth the time constant is used, the equations of Table 6-1 will produce an exceptionally high controller gain. Furthermore, it is unlikely, given the realities of the testing and approximation method, that a dead time-to-time constant ratio of less than 0.1 will be determined with sufficient accuracy to justify its use. Hence, a minimum dead time of 0.1 times the time constant should be used with Table 6-1. If the actual dead time happens to be less than this, the equations will produce conservative tuning values that can then be improved by trial-and-error fine-tuning.

If the dead time is longer than the time constant, then some other form of control, such as a dead-time compensation algorithm. Dead-time compensation and other model-based control techniques are discussed in chapter 14.

This discussion of tuning by open-loop testing has focused on self-regulating processes. The FOPDT model cannot approximate an integrating process, such as is found in some level-control applications, since it never reaches equilibrium. Considerations for tuning level-control loops are discussed later in this chapter.

Lambda Tuning

The Lambda tuning technique (Ref. 6-5) utilizes the same open-loop data, K_p, τ, and θ as the previous techniques. It requires an additional parameter, λ, however, which is the desired time constant of the closed-loop response to set point change. In other words, we are trying to force the set point response to follow a trajectory similar to that shown in Figure 6-9 where the process variable comes to set point gradually and with no overshoot. Since we have the freedom to choose λ, we can choose how aggressive the control action will be.

The tuning relations are as follows:

$$K_C = \frac{\tau}{K_p(\lambda + \theta)} \tag{6-10}$$

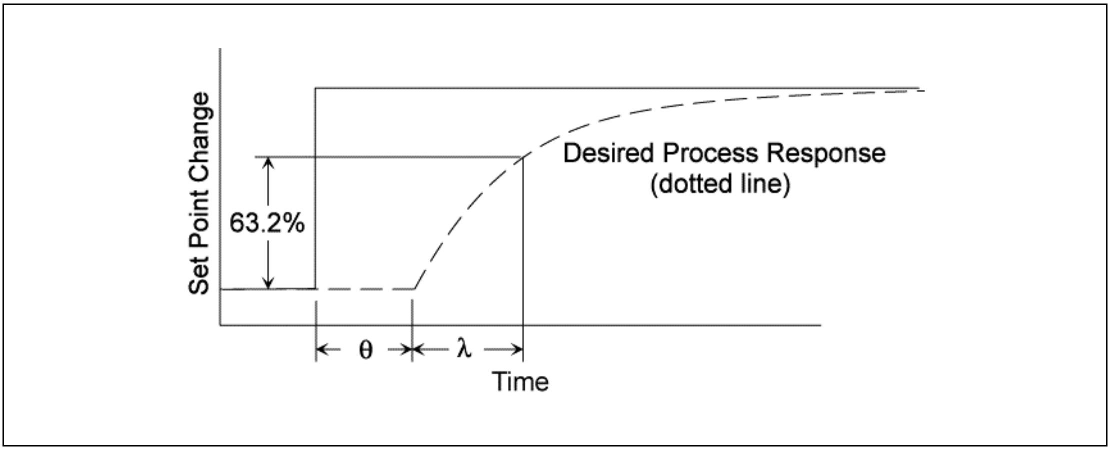

Figure 6-9. Desired Closed-loop Response to Set Point Change: The Basis of Lambda Tuning

$$T_I = \tau + \frac{\theta^2}{(\lambda + \theta)} \tag{6-11}$$

Lambda tuning is a form of internal model control (see chapter 14). Note that in the absence of dead time, the response is a true first-order lag with a time constant of λ. That is:

$$\frac{PV(s)}{SP(s)} = \frac{1}{\lambda s + 1}.$$

The choice of λ depends upon the degree of confidence in the process model parameters, K_p, τ, and θ. If there is high confidence, choose $\lambda = \tau$. This will provide a relatively fast response to set point changes. If there is low confidence, choose a value for λ somewhere between 2τ and 4τ. If λ is chosen to be less than τ, then the controller will be overly aggressive, and the output will initially overdrive its final value. One disadvantage that the lambda tuning technique shares with other internal model controllers is that it provides poorer response to disturbances.

Lambda tuning is favored by some industries, particularly the pulp and paper industry, for its ability to make gentle changes to the process on a set point change. One factor favoring its application in that industry is that dead time is often known exactly, since it is inversely proportional to paper-machine speed.

♦ Tuning from Closed-loop Test Data

The procedure for making a closed-loop test is as follows: set the integral action to a minimum, remove all derivative action, and set the gain to a low value.[5] Put the controller in automatic, with the measurement near the normal operating point. Then make a small change in set point. If the process does not oscillate, or if the oscillation quickly decays, then increase the gain, perhaps by as much as 50 percent, and repeat the test. The objective is ultimately to have the gain high enough so that a sustained oscillation (neither increasing nor decaying) will result. Normally, observing for three cycles or less will be sufficient to determine whether there is sustained oscillation. If the controller gain is high, it may be easier to observe the controller output rather than the process variable. Be sure that neither the measurement nor the controller output reaches an upper or lower limit. If the process is operating near a limit, it may not be possible to use this test.

Once sustained oscillation is attained, measure the period of oscillation. This is called the ultimate period, P_u. Note the gain that ultimately produced sustained oscillation; this is called the ultimate gain, K_{Cu} (see Figure 6-10). (The ultimate proportional band, PB_u, may be determined instead. If so, use the usual relationship,

$$K_{Cu} = \frac{100}{PB_u}$$

to find the ultimate gain.) Once the ultimate gain and ultimate period have been found, use equations from Table 6-2 to calculate tuning parameters for the chosen controller modes.

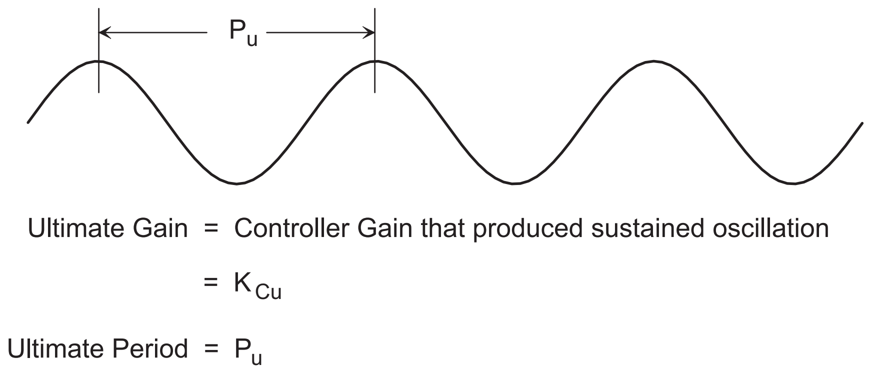

Figure 6-10. Determining Process Characteristics by a Closed-loop Test

5. Integral action can be set to a minimum by setting the value of repeats per minutes to the lowest dial setting (preferably 0.0) or the value of minutes per repeat to the largest dial setting. With some microprocessor-based systems, the integral adjustment is calibrated in terms of minutes per repeat, yet these systems accept a value of 0.0. This is interpreted by the system not as “zero minutes per repeat” but as “turn the integral action off.”

Table 6-2. Controller Tuning Equations Using Closed-loop Data (Ref. 6-1)

Controller Type	Parameter	Equation
Proportional Only (P)	Gain	$K_C = 0.5\ K_{Cu}$
Proportional + Integral (PI)	Gain	$K_C = 0.45\ K_{Cu}$
	Integral Time (minutes/repeat)	$T_I = 0.83\ Pu$
Prop + Integral + Deriv (PID)	Gain	$K_C = 0.6\ K_{Cu}$
	Integral Time (minutes/repeat)	$T_I = 0.5\ P_u$
	Derivative Time (minutes)	$T_D = 0.125\ P_u$

The mechanics of performing a closed-loop test are relatively easy to describe. It may not be so easy to implement the test, however, for several reasons:

(1) It is difficult to control the amplitude of oscillation. A large amplitude is not required; in fact it need be only large enough to distinguish control oscillation from the measurement noise band. Even so, a small change in set point may yield a larger than expected amplitude of oscillation.

(2) For many applications a sustained oscillation may not be tolerable.

(3) Many supervisory and operations personnel may object to a sustained oscillation, even though the purpose of the test is to obtain better controller tuning.

(4) Several tests that require a long testing period, and so a lengthy period of off-spec production, may be needed to obtain sustained oscillation.

Despite these disadvantages, the closed-loop test has the following advantages over the open-loop test:

(1) The closed-loop method makes no a priori assumption about the form of the process model. We do not try to force the process to look like a FOPDT model.

(2) The quality of data obtained from the closed-loop test, and therefore the quality of the tuning parameters produced, is much higher with the closed-loop method than with the open-loop method. This is because frequency (or period) can be measured very precisely, whereas dead time and time constants can only be approximated.

(3) The effects of a sticking valve are a part of the closed-loop test data. Therefore, the resulting tuning parameters inherently take this into consideration.

To gain the advantages of both of these methods, we offer the following compromise method:

(1) Start out as if you were going to make a closed-loop test, that is, no reset, no derivative, low gain, and with the controller on automatic.

(2) Continue making step tests, with higher and higher gain, until you attain quarter-amplitude decay. This should occur around the midpoint of the oscillations, not necessarily around the set point, since with no integral action there may be an offset. Note the period, P_q, of the response with quarter-amplitude decay. If we continued the test until we reached a sustained oscillation, the ultimate period would not be appreciably shorter than P_q. Hence, from the data we have, we can estimate P_u as follows:

$$P_u(est) = 0.9\, P_q\ . \qquad (6\text{-}12)$$

(3) Note the gain, K_{Cq}, which produces the quarter-amplitude decay. If we continued the test until we reached a sustained oscillation, the ultimate gain would be considerably larger than K_{Cq}, but by a predictable ratio. Usually, it would be between 1.6 to 1.75 times K_{Cq}. Hence, we can estimate K_{Cu} by the following rule of thumb:

$$K_{Cu}(est) = 1.67 K_{Cq}\ . \qquad (6\text{-}13)$$

(4) Once we have estimates for K_{pu} and P_u, use Table 6-2 to determine tuning parameters for the selected control modes. The error in the controller tuning parameters will probably not exceed 5 percent to 10 percent of the values that would have been determined had the tests continued to sustained oscillation. Furthermore, it will usually be much more expedient to drive the process into only quarter-amplitude decay oscillation rather than a sustained oscillation.

♦ Improving "As-Found" Tuning Parameters

We are often called on to improve the tuning of a loop that is already operating. Usually, the request is the result of the fact that the loop is oscillating unacceptably or that the loop response is sluggish in returning to the set point, particularly after a load upset.

The cause of the unsatisfactory performance may not lie in the controller. So, before considering controller tuning, we should first investigate several process, equipment, and operating conditions. For the purposes of this discussion, however, let us suppose that we have completed that investigation, have ruled out other causes, and have decided that we truly have a controller tuning problem. Let us also suppose that we are using a proportional-integral controller—no derivative. Thus, our method will apply in a vast majority of actual situations.

The tuning parameters and the characteristics of the process behavior represent the "as-found" conditions. In the absence of other control-loop disturbances, the existing tuning parameters give rise to the existing process behavior. Thus, some relationship must exist in the "as-found" data. In other words, the “as-found” data represents a quantity of knowledge about the process. If we were to make either an open- or a closed-loop test on the process, we would in effect be “starting over,” discarding any knowledge we already have about the process.

Many tuners will attempt trial-and-error tuning. Without an organized procedure, a trial-and-error technique is likely to resemble a random walk and may or may not result in improved tuning. Frequently, an inexperienced tuner will reduce the controller gain so as to minimize cycling, then make the reset action faster in an attempt to force the process variable to return to set point faster. These efforts are at cross purposes, since one is a stabilizing move, and the other a destabilizing move. At the very least, unorganized trial-and-error tuning may consume an excessive amount of the tuner’s time and/or result in a significant amount of poor production.

We will now present an organized procedure for controller tuning. The objective is to adjust the “as-found” parameters as quickly as possible to give acceptable response. That is, we want to minimize the number of tuning parameter changes. We also want to avoid making either of the traditional open-loop or closed-loop tests. We will first present the essence of the technique; then we will make it more formal by means of a flow chart.

If the loop is oscillating unacceptably, if we have determined that this is because of poor tuning, and if we are using a proportional-plus-integral controller, then the oscillations are caused either by the gain being too high or by the integral action being too fast, or both. Which is it?

It is almost always possible to decrease the gain (increase the proportional band) and diminish the oscillation. If the loop is going out of control (increasing amplitude of oscillation), decreasing the gain is probably the right thing to do immediately, but it may not be the best long-range solution. It could be that the basic cause of oscillation is that the integral action has been set too fast.

If the oscillation appears to be acceptable (i.e., oscillates with a quarter-amplitude decay response or less), how can we be sure that the loop is tuned as well as it can be? It is possible that the integral action has been set too fast (a destabilizing effect), only to be compensated for by setting a low gain (a stabilizing effect). In this situation, under a load upset, the sluggishness of the loop may result in a lengthy deviation from set point.

If a PI controller is acceptably tuned, there is a predictable relation between the integral time and the period of oscillation, P. As a general rule, this relationship is as follows:

$$1.5 < \frac{P}{T_I} < 2.0 \tag{6-14}$$

(More will be said about these limits later.) Hence, we can determine whether the reset is more or less set correctly by comparing the integral time, in minutes per repeat, with the period of oscillation. If the reset is approximately correct, then the following relation will hold:

$$1.5\, T_I < P < 2.0\, T_I\,. \tag{6-15}$$

This rule of thumb gives a window, based on the current integral time. The period of oscillation should lie within this window. If the period is longer than this window allows, indications are the existing integral time is too fast. Use the present period to choose a new integral time that satisfies the following relation:

$$\frac{1}{2}P < T_I < \frac{2}{3}P\,. \tag{6-16}$$

If the period is shorter than this window allows, indications are either that the integral action is too slow or the present controller gain is too high. If the decay ratio is greater than desired, then reduce the gain. Otherwise, readjust the integral time to some value between one-half to two-thirds of the present period.

If the period satisfies Equation 6-15, then adjust the gain to achieve a satisfactory decay ratio.

The previous paragraphs have presented the essence of an organized tuning procedure for improving the tuning of an existing loop. The procedure can be formalized by the flow chart of Figure 6-11 and the additional chart shown in Figure 6-12. The flow chart assumes that the target response is a quarter-amplitude decay following a set point change, and that the target period-to-reset time ratio can be expressed by Equation 6-15. When the flow chart requires the integral time to be changed, use Equation 6-16 as a guide for selecting a new value. When the flow chart requires that the gain (or proportional band) be changed, only the direction of change is given, not a numerical guide for selecting a new value. That guide is given, however, by Figure 6-12, which uses the present decay ratio to select a multiplier (divisor) for adjusting the present gain (proportional band). The flow chart provides a couple of "escapes" in the event less oscillation than a quarter-amplitude decay is desired.

The flow chart attempts to cover all contingencies. It does so only if we make a correct interpretation when anomalous control-loop behavior occurs. In Figure 6-13a, the loop is not oscillating so by exiting at the right end of decision block 2 and using Figure 12 with a decay ratio of 0, we are instructed to multiply the gain, or divide the proportional band, by a value of 1.35.[6] In Figure 6-13b, the loop is oscillating because we can measure a period. But either the first or second peak, or both, failed to rise above the set point, hence the traditional definition of decay ratio (see Figure 6-1a) cannot be applied. However, the alternative definition of decay ratio, using the ratio of peak-to-valley differences (see Figure 6-1b), can be applied. In Figure 6-13c, the loop is tending to oscillate, but the second peak is indeterminate. By measuring the time between the first peak and the valley, we can get a measure of half the period.

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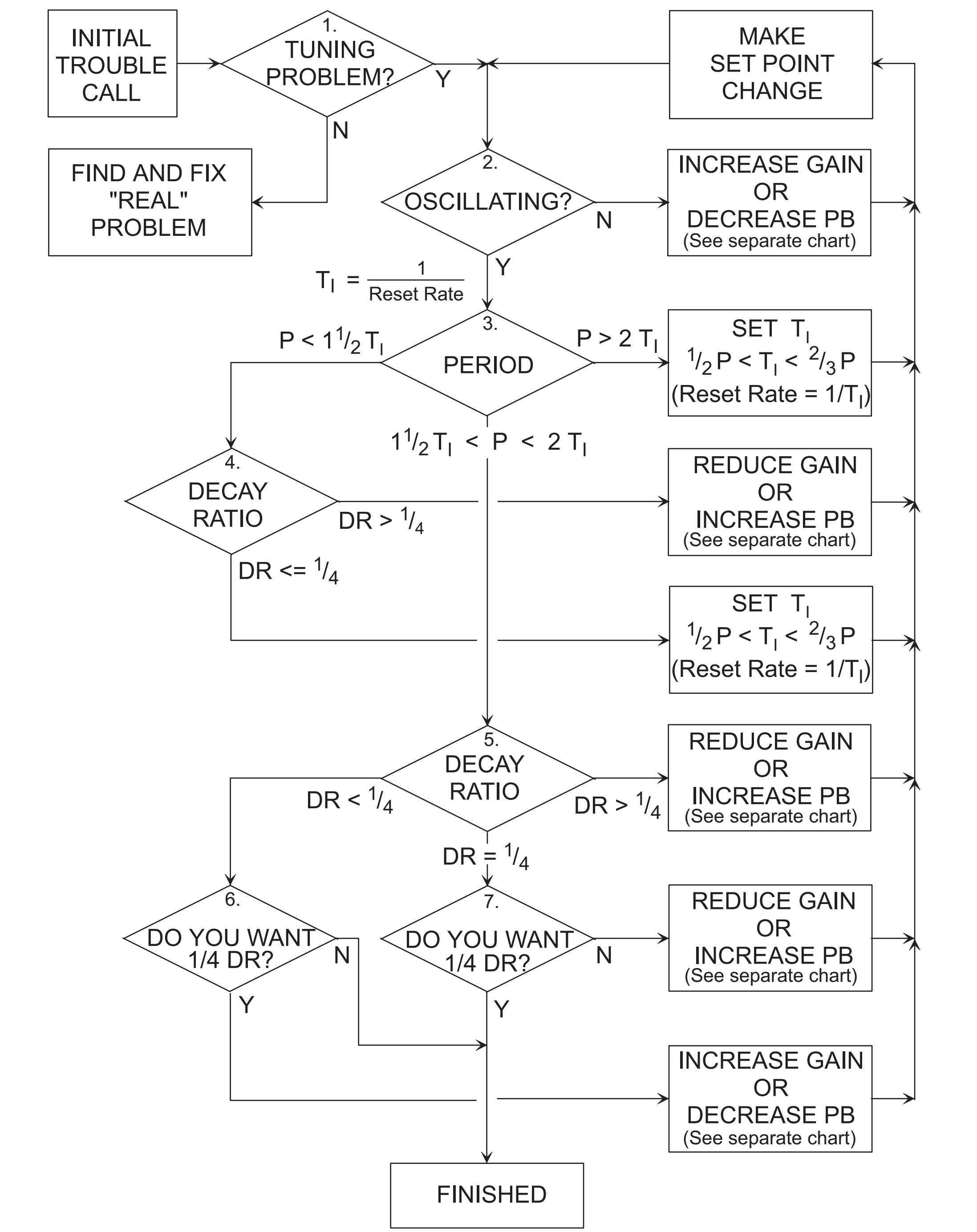
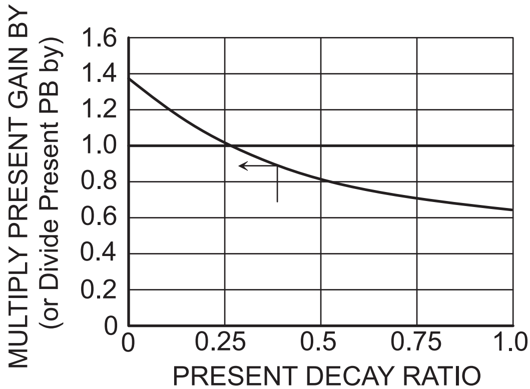
6. The term *oscillation* can be applied to three forms of oscillation: sustained, decaying, or expanding. Some persons incorrectly apply the term *oscillation* only to the behavior exhibited by Figure 6-10. When asked if the behavior depicted by Figure 6-1a is oscillating, they would reply, "No, it is dying out." In fact, both figures depict some form of oscillation.



(See Figure 6-12 for gain (or PB) adjustment chart.)

Figure 6-11. Flow Chart for PI Controller Tuning for a Self-Regulating Process



INSTRUCTIONS: When the Flow Chart recommendation is to
Change Gain or Change PB,
enter the graph on the horizontal axis at the present decay ratio.
Read the related factor on the vertical axis.
Multiply the present Gain, or Divide the present PB by this factor.

Figure 6-12. Gain (or PB) Adjustment Chart for Use with the Flow Chart

Double this value for the full period. Since there is no discernible second peak or valley, the decay ratio should be considered as zero.

Example 1:

Suppose you are called into the control room to look at an errant loop that is using a PI controller. The current response is as shown in Figure 6-14a. After exploring all other possibilities (Is it a process problem? Is it being influenced by some other loop?), you conclude that it is a tuning problem. You note the "as-found" tuning parameters, and also determine the current period of oscillation and decay ratio. (It is highly recommended that you keep a log of your tuning parameter changes and the resulting behavior as shown in the Example 1 Tuning Log.)

Following the flow chart, the loop is oscillating, so check the period. The period is greater than 2 times T_I, so exit the right end of decision box 3 to a box that instructs us to set T_I to somewhere between one-half and two-thirds of the present period, or between 7.5 and 10 minutes. Suppose we choose 8.0 minutes per repeat. Enter this and make a small set point change. (Note that we only change one parameter at a time.) Now our response appears as Figure 6-14b. We record the new tuning parameter and resulting behavior as shown in line 2 (Tuning Trial 1) on the Tuning Log.

In a plant situation, we would probably stop at this point. However, for instructional purposes, suppose we try to get closer to quarter-amplitude decay. The benefit would be a slightly improved response to a disturbance. The period-reset time criterion is met, so exit the bottom of block 3, the left end of block 5, and then exit the bottom of block 6. (If we really did not want quarter-amplitude decay, we would exit at the right end of block 6 and be finished.) Our

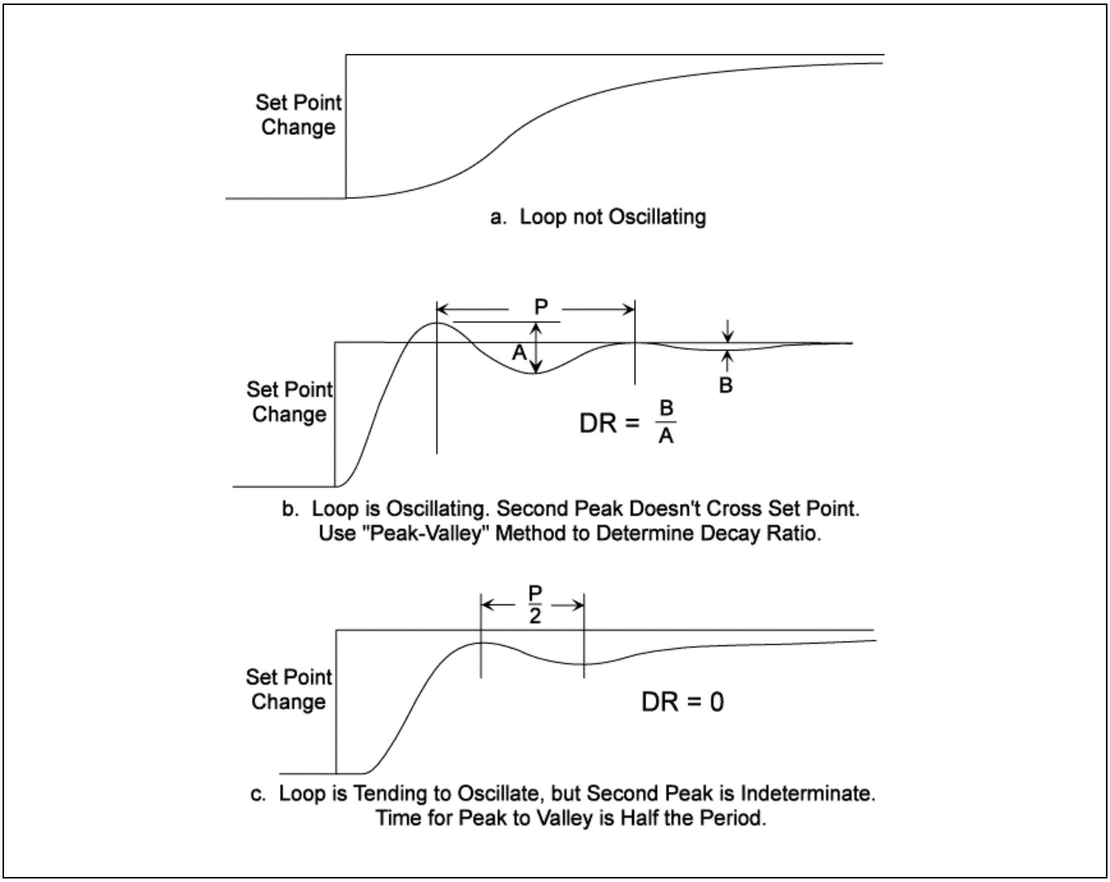


Figure 6-13. Anomalous Behavior for Set Point Response

Example 1 Tuning Log

Tuning Trial	K_C	T_I	Decay Ratio	Period
As Found	*2.0*	*3.0 min/rpt*	*0.35*	*15.0 min.*
1	*2.0*	*8.0 min/rpt*	*0.13*	*14.6 min.*
2	*2.3*	*8.0 min/rpt*	*0.18*	*13.6 min.*
3	*2.53*	*8.0 min/rpt*	*0.23*	*13.0 min.*

instructions are to increase the gain. At a decay ratio of 0.13, Figure 6-12 gives a gain adjustment factor of about 1.15. Multiply this by the current gain to get a new gain value of 2.3. Enter this and make another small set point change. We observe the resulting behavior as shown by Figure 6-14c. We record the new tuning parameter and the behavior on line 3 (Tuning Trial 2) on the Tuning Log.

Again, we are tempted to say “good enough,” but suppose we make one more attempt. Following the same path as before, then using Figure 6-12, we get a gain adjustment factor of 1.1,

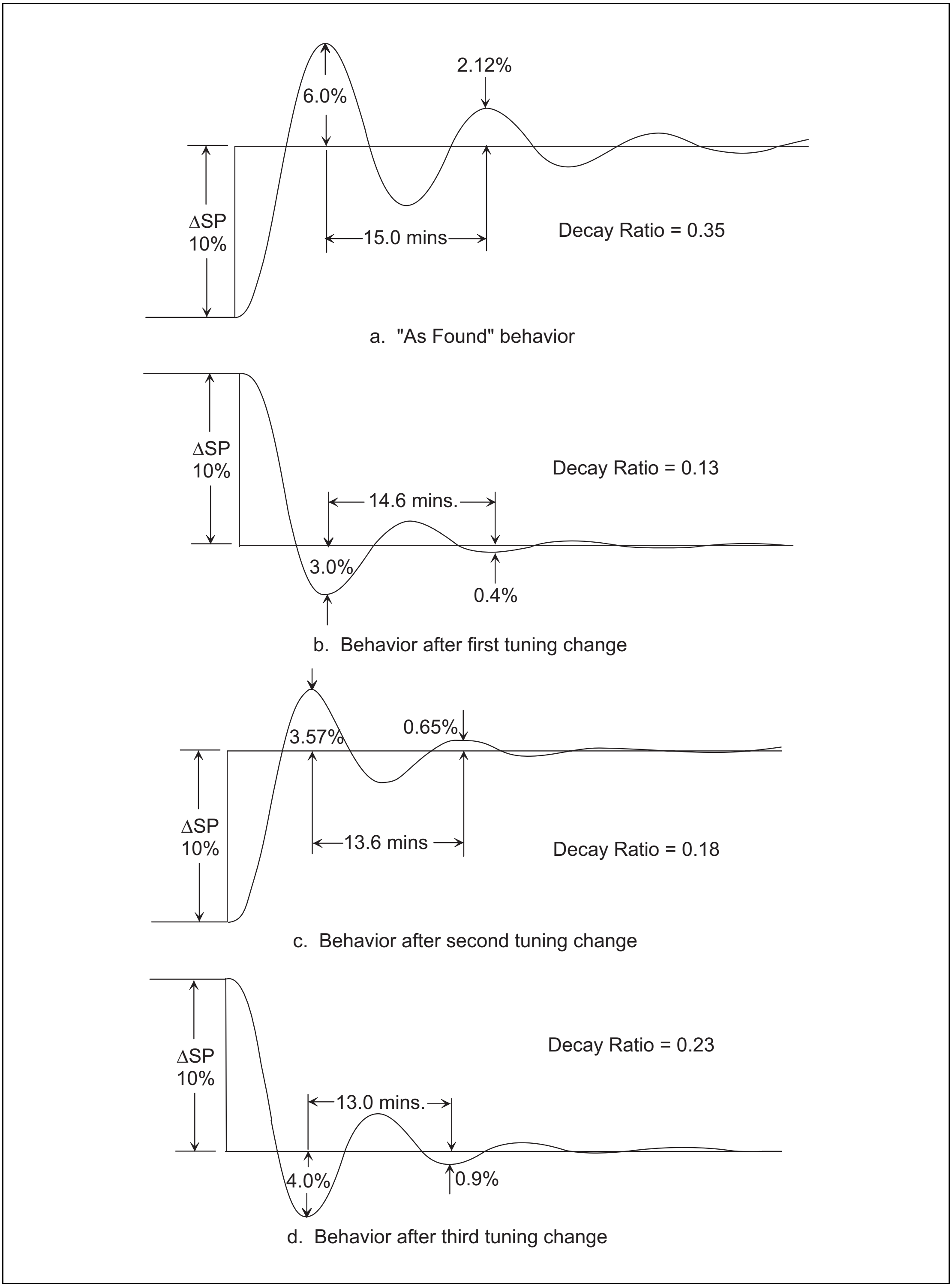


Figure 6-14. Example of Successive Response for Controller Tuning

which results in a new gain of 2.53. This produces the behavior shown in Figure 6-14d. Our tuning log appears as shown in line 4 (Tuning Trial 3) on the Tuning Log.

This time we are sufficiently close to quarter-amplitude decay. (Don't be a slave to precise numbers!) We accept this as our final tuning.

In this example, we have improved the "as-found" tuning, with at most three tuning moves. In a plant situation, you may not be quite this lucky. But by using the rules of thumb given previously as a guide, you should be able to tune the loop much more efficiently (i.e., with fewer tuning moves) than by a pure trial-and-error procedure.

End of Example 1.

Limits for the P/T_I Ratio

The limits stated by Equation 6-10 for the P/T_I ratio are "soft" limits. While they are reasonable target limits for most loops, they need not be adhered to slavishly. If the objective is to minimize the effect of a disturbance, then the limits should be raised as the ratio of dead time-to-dominant time constant increases. Lopez et al. (Ref. 6.3) present empirical equations for minimizing the IAE for a pure FOPDT process that is subjected to a step disturbance at the process input. A simulation study showed that their tuning rules resulted in an increase in the P/T_I ratio from 1.7 at a dead time-to-time constant (θ/τ) ratio of 0.1 to 2.6 at a θ/τ ratio of 1.0 Thus, if you know even an approximation to the θ/τ ratio, or if you have experience with similar types of processes to serve as a guide, you can increase the limits for P/T_I (Equation 6-12). Equations 6-13 and 6-14 should then be modified. The flow chart, with the modified limits, will still provide an organized procedure for improving as-found tuning.

If you are also using the derivative mode, then it is less clear what the limits for the rule-of-thumb window should be. With a low amount of derivative, the limits will not change appreciably from those stated previously. With a significant amount of derivative (up to one-quarter of the integral time), and with a corresponding increase in the controller gain and decrease in the integral time, the period will be decreased slightly. Therefore, the following limits appear to be reasonable when derivative is used.

$$2\,T_I < P < 3.33\,T_I\,. \qquad (6\text{-}17)$$

❖ TUNING LIQUID-LEVEL CONTROL LOOPS

The previous sections of this chapter have dealt with the tuning of controllers for self-regulating processes such as temperature and flow. Liquid-level control, however, has characteristics distinctly different from those of the previous loops. Some of the differences are these:

- Liquid level is usually a non-self-regulating (integrating) process.

- Intuitive rules of thumb used for tuning other loops ("If it's cycling too much, reduce the gain.") do not apply, and in fact will usually produce results that are the opposite of those expected.

- Liquid-level control loops, once properly tuned, do not usually go out of tune.

Most processes can be described at best by an approximate process model that must often be determined by process testing. On the other hand, most liquid-level control loops readily yield to an analytical approach. A simple process model can be formulated, desired performance parameters can be established, and from this controller tuning parameters can be calculated. Once this is done, other attributes of the control loop, such as the period of oscillation, can be predicted.

Determining tuning parameters for a liquid-level loop should probably be considered as an engineering activity, rather than being left for field trial-and-error tuning, for two reasons: the counterintuitive nature of liquid-level loops makes tuning by trial-and-error techniques difficult, and liquid-level loops are amenable to an analytical approach. Hence it can be said that *liquid-level control loops should be engineered, not tuned*.

Many engineering studies start with an ideal model, then incorporate subsequent considerations to account for differences found in real situations. Engineering design problems are usually based upon some assumed worst-case conditions, even though those conditions may never be experienced in reality. We will follow the same approach in engineering liquid-level control loops.

♦ Idealized Control Loop Model

An idealized liquid-level control system is shown in Figure 6-15.

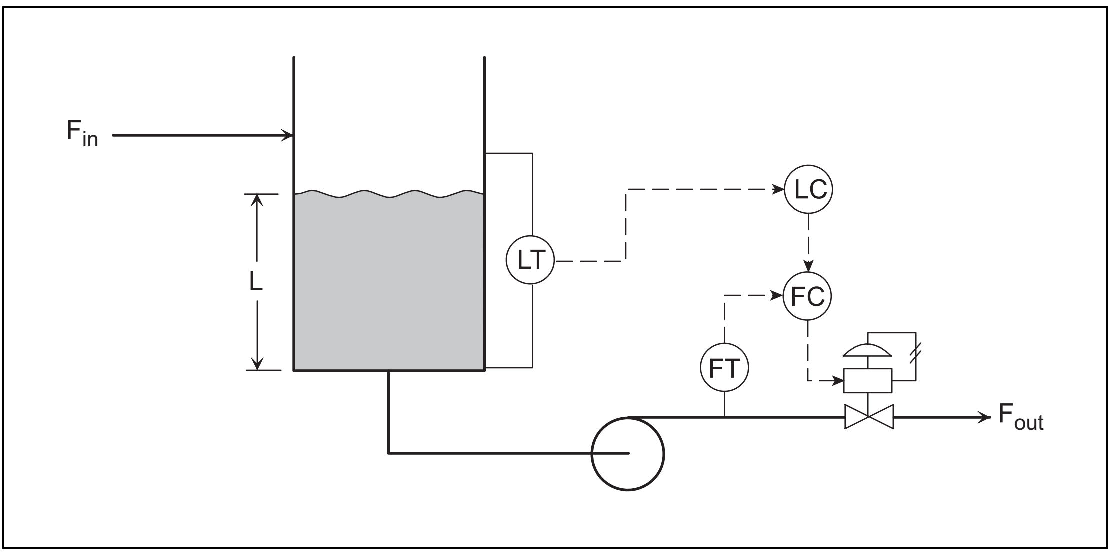


Figure 6-15. Ideal Process for Liquid-level Control

Attributes of this idealized system are these:

- The tank has constant cross-sectional area.
- The level controller is cascaded to a flow controller.
- A valve positioner is installed on the flow control valve.
- All inflow goes to outflow; the tank is merely a buffer storage tank.
- The maximum outflow is the same as the maximum inflow.
- The size of the tank is substantial.
- There is no thermal effect such as the boiling liquid found in boiler-drum level control.
- The level controller operates at constant set point.

The implications for these attributes are as follows:

- The level-control loop constitutes a linear system.
- There is no effect from upstream or downstream pressure, line loss, or pump curve.
- The system is not affected by the size of the valve.
- The dynamics of the flow loop are significantly faster than the dynamics of the vessel, and therefore can be ignored.
- There is no dead time in the loop.
- Response to set point change need not be considered since set point changes are rarely made; instead, the response to a disturbance is of critical importance.

Figure 6-15 shows a common situation in which the level controller manipulates the outflow in response to changes in inflow. There are also cases where the level controller manipulates the inflow in response to varying demands for the outflow. The technology developed here is applicable to both cases, generally by exchanging the words *inflow* and *outflow.*

A key parameter required for analysis is the tank holdup time, also called the *tank time constant.* If the tank geometry (diameter, distance between the level taps) and maximum outflow rate (flow rate corresponding to 100 percent output of level controller) are known, then the tank time constant can be calculated from the following:

$$T_L = \frac{f_{out}}{Q} \tag{6-18}$$

where: T_L = tank time constant
Q = holdup quantity, between upper- and lower-level sensor taps
f_{out} = maximum outflow

Q and f_{out} should be in compatible units, such as "gallons" and "gallons per minute." The time constant is then expressed in minutes.

A block diagram for the loop with PI control is shown in Figure 6-16.

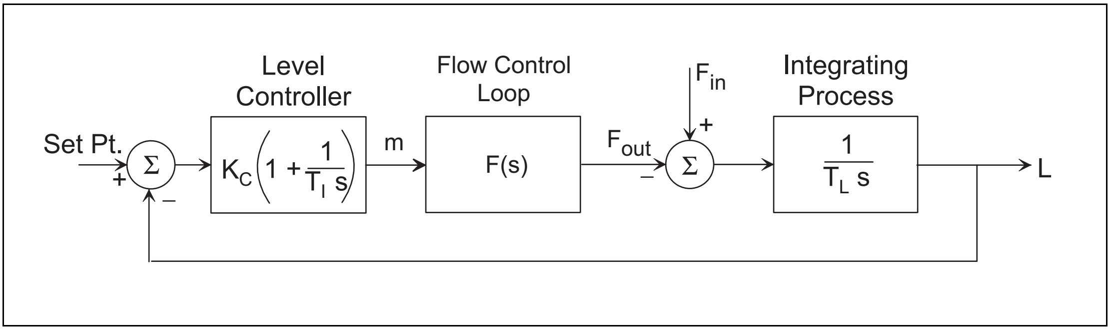


Figure 6-16. Block Diagram for Ideal Liquid-level Process Model

If the dynamics of the flow loop are negligible compared with the tank dynamics, then it can be assumed that *F(s),* the closed-loop transfer function of the flow loop, is equal to unity. If the loop operates at constant set point, then we are more interested in the response to a disturbance (i.e., to a change in f_{in}) than to the set point response. However, in addition to the response of the level to a change in f_{in}, we may also be interested in the response of the outflow, f_{out}, to a change in f_{in}. Transfer functions representing these two responses, derived from Figure 6-16, are given by the following two equations:

$$\frac{L(s)}{F_{in}(s)} = \frac{\frac{s}{T_L}}{s^2 + \frac{K_C}{T_L}s + \frac{K_C}{T_I T_L}} \tag{6-19}$$

$$\frac{F_{out}(s)}{F_{in}(s)} = \frac{\frac{K_C}{T_L}s + \frac{K_C}{T_I T_L}}{s^2 + \frac{K_C}{T_L}s + \frac{K_C}{T_I T_L}} \tag{6-20}$$

These equations display the fact that the loop acts as a second-order system. These transfer functions, written with traditional servomechanism terminology, are:

$$\frac{L(s)}{F_{in}(s)} = \frac{\frac{s}{T_L}}{s^2 + 2\zeta\omega_n s + \omega_n^2} \tag{6-21}$$

$$\frac{F_{out}(s)}{F_{in}(s)} = \frac{2\zeta\omega_n s + \omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \qquad (6\text{-}22)$$

where ζ, the damping factor, and ω_n, the natural frequency, are given by:

$$\zeta = \frac{1}{2}\sqrt{\frac{K_C T_I}{T_L}} \qquad (6\text{-}23)$$

$$\omega_n = \sqrt{\frac{K_C}{T_I T_L}} \qquad (6\text{-}24)$$

The damping factor is a dimensionless number. The natural frequency is in radians per minute if T_I is in minutes (minutes per repeat) and T_L is in minutes.

If $0 < \zeta < 1$, the control loop is said to be *underdamped.* A step change in disturbance or a set point change will result in a decaying, oscillatory response.

If $\zeta = 1$, the control-loop response is said to be *critically damped*. A step change in disturbance or a set point change will result in a relatively rapid return to set point without oscillation.

If $\zeta > 1$, the control-loop response is said to be *overdamped*. A step change in disturbance or a set point change will result in a relatively slow return to set point without oscillation.

Many practicing engineers may be more familiar with the concept of decay ratio, *DR*, rather than the damping factor, ζ. (Decay ratio is defined for both a set point change and a disturbance in Figure 6-1.) It can be shown that the damping factor and decay ratio are related by the following relations:

$$\zeta = \frac{-ln(DR)}{\sqrt{4\pi^2 + (ln(DR))^2}} \qquad (6\text{-}25)$$

$$DR = exp\left(\frac{-2\pi\zeta}{\sqrt{1-\zeta^2}}\right). \qquad (6\text{-}26)$$

For example, for the familiar quarter-amplitude decay, ζ has a value of 0.215.

♦ Determining Tuning Parameters

In addition to knowing the tank holdup time, T_L (see Equation 6-18), analytical determination of tuning parameters requires three additional decisions:

- ΔF_{in} – the maximum step change in disturbance (inflow) that can be expected, in percentage of full scale (this is the "worst-case" assumption);
- ΔL_{max} – the maximum allowable deviation from set point, in percentage of full scale (choosing this parameter lets us determine whether we want "tight" or "loose" tuning);
- DR – the desired decay ratio, in the event of a step inflow disturbance.

Using these values, the value for T_L , and Equation 6-25 to convert decay ratio into damping factor, the relations given in Table 6-3, 6-4 and 6-5 can be derived.[7] Table 6-3 presents relations for calculating tuning parameters, while Tables 6-4 and 6-5 present relations for predicting attributes of the level and outflow response to a step change in inflow.

Table 6-3. Liquid-level Tuning Parameter Relations for Ideal Model

Tuning Parameter	Underdamped $\zeta < 1$		Critically Damped $\zeta = 1$
	Rigorous	Simplified	
K_C	$2\zeta\, e^{-\zeta f(\zeta)}\left(\dfrac{\Delta F_{in}}{\Delta L_{max}}\right)$	$2\zeta\, e^{-\zeta f(\zeta)}\left(\dfrac{\Delta F_{in}}{\Delta L_{max}}\right)$	$\dfrac{2}{e}\left(\dfrac{\Delta F_{in}}{\Delta L_{max}}\right)$ $(e = 2.71828\cdots)$
T_I	$2\zeta\, e^{\zeta f(\zeta)}\left(\dfrac{T_L\,\Delta L_{max}}{\Delta F_{in}}\right)$	$4\zeta^2\left(\dfrac{T_L}{K_C}\right)$	$4\left(\dfrac{T_L}{K_C}\right)$
Note: In the table above, $f(\zeta) = \dfrac{1}{\sqrt{1-\zeta^2}}\tan^{-1}\dfrac{\sqrt{1-\zeta^2}}{\zeta}$			

The column labeled "Rigorous" demonstrates that the tuning parameters are entirely a function of the four fixed or chosen parameters, T_L , ζ (as determined from the chosen decay ratio), ΔF_{in} and ΔL_{max}. The column labeled "Simplified" produces the same results, calculated in terms of some previously calculated quantity.

7. Complete derivations are available from the author. Contact hlwade@aol.com.

Once the tuning parameters have been calculated, the predicted behavior for the level as well as for the outflow can be calculated from Tables 6-4 and 6-5.

Table 6-4. Predicted Behavior for Level – Ideal Model

Behavior Attribute	Underdamped $\zeta < 1$		Critically Damped $\zeta = 1$
	Rigorous	Simplified	
Arrest Time T_{aL}	$f(\zeta)e^{\zeta f(\zeta)}\left(\frac{T_L \Delta L_{max}}{\Delta F_{in}}\right)$	$\frac{f(\zeta)}{2\zeta}T_I$	$\frac{T_I}{2}$
Period P	$\frac{2\pi}{\sqrt{1-\zeta^2}}e^{\zeta f(\zeta)}\left(\frac{T_L \Delta L_{max}}{\Delta F_{in}}\right)$	$\frac{\pi}{\zeta\sqrt{1-\zeta^2}}T_I$	N/A
IAE	$\left(\frac{1+e^{\frac{-\zeta\pi}{\sqrt{1-\zeta^2}}}}{1-e^{\frac{-\zeta\pi}{\sqrt{1-\zeta^2}}}}\right)e^{2\zeta f(\zeta)}\left(\frac{T_L(\Delta L_{max})^2}{\Delta F_{in}}\right)$	Same as ←	$e^2\left(\frac{T_L(\Delta L_{max})^2}{\Delta F_{in}}\right)$
Note: In the above table, $f(\zeta) = \frac{1}{\sqrt{1-\zeta^2}}\tan^{-1}\frac{\sqrt{1-\zeta^2}}{\zeta}$			

The level arrest time, T_{aL}, represents the time from the disturbance until the maximum deviation from set point. The period, P, is the predicted period of oscillation of the level-control loop, if it is underdamped. See Equation 6-1 for the definition of *IAE*.

$\Delta F_{out-max}$ represents the maximum change in outflow. The ratio

$$\frac{\Delta F_{out-max}}{\Delta F_{in}}$$

will always be greater than 1. The outflow arrest time, T_{aF}, represents the time from the disturbance until the maximum change in outflow. The maximum rate of change of outflow is given since it is this quantity, rather than the size of the outflow change itself, that represents the maximum disturbance to a downstream process unit.

The relations given in Tables 6-3, 6-4, and 6-5 show the development of this tuning technique, but they are not very useful as working relations because of the amount of computation required. For three specific decay ratios, Tables 6-6 and 6-7 present working relations. The three decay ratios chosen are these:

Table 6-5. Predicted Behavior for Outflow – Ideal Model

Behavior Attribute	Underdamped $\zeta < 1$		Critically Damped $\zeta = 1$
	Rigorous	Simplified	
Maximum Change in Outflow $\Delta F_{out-max}$	$\left(1+e^{-2\zeta f(\zeta)}\right)\Delta F_{in}$	$\left(1+e^{-2\zeta f(\zeta)}\right)\Delta F_{in}$	$\left(1+e^{-2}\right)\Delta F_{in}$
Arrest Time T_{aF}	$2 f(\zeta) e^{\zeta f(\zeta)} \left(\frac{T_L \Delta L_{max}}{\Delta F_{in}}\right)$	$2T_{aL}$	$2T_{aL}$
Max Rate of Change of Outflow $\left(\frac{dF_{out}}{dt}\right)_{max}$	$\zeta \le ½$ $\left[exp\left(-\zeta f(\zeta) - \frac{1}{\sqrt{1-\zeta^2}} tan^{-1}\left(\frac{(1-4\zeta^2)\sqrt{1-\zeta^2}}{\zeta(3-4\zeta^2)}\right)\right)\right] \times \left(\frac{(\Delta F_{in})^2}{T_L \Delta L_{max}}\right)$	$\zeta \le ½$ Same as ←	$\frac{2}{e}\left(\frac{(\Delta F_{in})^2}{T_L \Delta L_{max}}\right)$
	$½ < \zeta < 1$ $2\zeta e^{-\zeta f(\zeta)} \left(\frac{(\Delta F_{in})^2}{T_L \Delta L_{max}}\right)$	$½ < \zeta < 1$ Same as ←	
Note: In the following expressions, $f(\zeta) = \frac{1}{\sqrt{1-\zeta^2}} tan^{-1} \frac{\sqrt{1-\zeta^2}}{\zeta}$			

- critically damped,
- one-quarter decay,
- one-twentieth decay.

Critically damped is chosen because it represents a recognized extreme form of tuning; one-quarter decay is chosen because of its familiarity. The last, although less familiar, is chosen because it provides both the minimum IAE and the lowest maximum rate of change of outflow. Figure 6-17 depicts these three forms of response with equal values of maximum deviation.

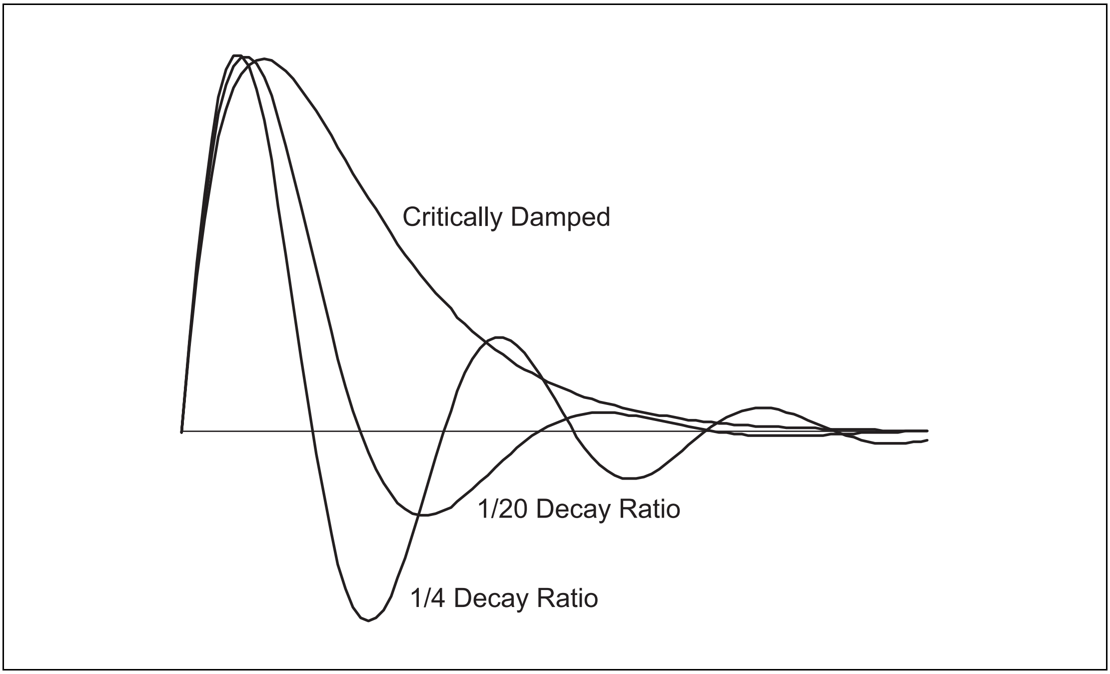


Figure 6-17. Three Forms of Disturbance Response; Equal Magnitude Maximum Deviation

Table 6-6. Working Equations Liquid-level Tuning Parameters – Ideal Model (If the loop has dead time or is not a cascade loop, use Tables 6-8 and 6-9 rather than 6-6 and 6-7.)

Decay Ratio DR	**Gain** K_C	**Integral Time** T_I
Critically Damped	$\frac{0.74\,\Delta F_{in}}{\Delta L_{max}}$	$\frac{4.0\,T_L}{K_C}$
0.05	$\frac{0.50\,\Delta F_{in}}{\Delta L_{max}}$	$\frac{0.74\,T_L}{K_C}$
0.25	$\frac{0.32\,\Delta F_{in}}{\Delta L_{max}}$	$\frac{0.19\,T_L}{K_C}$

Table 6-7. Working Equations for Predicted Behavior ~ Liquid-level Control – Ideal Model (If the loop has dead time or is not a cascade loop, use Tables 6-8 and 6-9 rather than 6-6 and 6-7.)

Decay Ratio DR	Level Response			Outflow Response		
	Level Arrest Time T_{aL}	Period P	IAE	Outflow $\Delta F_{out\text{-}max}$	Outflow Arrest Time T_{aF}	Outflow Max Rate of Change
Critically Damped	$0.5\ T_I$	Not Applicable	$7.40 \times \dfrac{T_L (\Delta L_{max})^2}{\Delta F_{in}}$	$1.14\ \Delta F_{in}$	T_I	$0.74 \times \dfrac{(\Delta F_{in})^2}{T_L\ \Delta L_{max}}$
0.05	$1.45\ T_I$	$8.09\ T_I$	$4.61 \times \dfrac{T_L (\Delta L_{max})^2}{\Delta F_{in}}$	$1.34\ \Delta F_{in}$	$2.90\ T_I$	$0.52 \times \dfrac{(\Delta F_{in})^2}{T_L\ \Delta L_{max}}$
0.25	$3.22\ T_I$	$14.93\ T_I$	$5.45 \times \dfrac{T_L (\Delta L_{max})^2}{\Delta F_{in}}$	$1.55\ \Delta F_{in}$	$6.43\ T_I$	$0.61 \times \dfrac{(\Delta F_{in})^2}{T_L\ \Delta L_{max}}$

Example 2:

Suppose you have a tank with the following specifications:

Tank diameter:	5.0 feet,
Distance between level transmitter taps:	8.0 feet,
Maximum outflow (upper end of outflow transmitter measuring span):	250 gpm.

Calculate the tank holdup time:

Surge volume $$= \frac{\pi}{4} \times 5^2 \times 8 = 157.3\ \text{ft}^3,$$

Surge quantity $$= Q = 157.3\ \text{ft}^3 \times 7.48\ \frac{\text{gal}}{\text{ft}^3} = 1176.6\ \text{gal},$$

Holdup time $$= T_L = \frac{1176.6}{250} = 4.7\ \text{min}.$$

Also, suppose that you anticipate that a worst-case disturbance would be a step inflow change of 10 percent. In the event of this disturbance, you want the level to deviate no more than 5 percent (about five inches). You would like for the system to settle out fairly rapidly, so you choose a 0.05 decay ratio.

$$\Delta F_{in} = 10\%,$$
$$\Delta L_{max} = 5\%.$$

With this data, use Table 6-6 to calculate tuning parameters:

$$K_C = \frac{0.50 \times 10}{5} = 1.0$$

$$T_I = \frac{0.74 \times 4.8}{1.0} = 3.55 \text{ min / repeat}.$$

We can use Table 6-7 to predict other properties of the response:

Level arrest time: $T_{aL} = 1.45 \times 3.55 = 5.15 \text{ minutes},$

Period: $P = 8.09 \times 3.55 = 28.7 \text{ minutes}.$

The period may seem to be excessive, but recall that because of the fast settling behavior selected, the maximum deviation during the second half-cycle will be about 1.1 inches, during the third half-cycle about 0.25 inches, and so on.

End of Example 2.

♦ "Real-world" Considerations

The results presented so far have been based on an idealized process model. Many real applications will fail to meet one or more of the criteria for the idealized model. The following are commonly encountered situations, along with suggested procedures for coping with them.

Irregularly Shaped Vessels

For irregularly shaped vessels, such as horizontal or spherical drums, using the level measurement directly as the process variable for the level controller may create highly nonlinear characteristics for the behavior of the control loop. In this case, the loop can be linearized using the volumetric holdup in the vessel rather than the level measurement. This may be computed from the vessel geometry and the actual level measurement. The process variable then should be scaled in terms of percentage of maximum volumetric holdup.

No Cascade Loop

In cases where there is no secondary flow controller as shown in Figure 6-15, the holdup time cannot be calculated from Equation 6-14 because the maximum outflow cannot be related to the maximum setting of a secondary flow controller. It would probably be futile to attempt to determine the maximum rate of outflow with a wide-open valve since there are unknown variables such as line loss, the effect of the pump curve, head effects in the tank, and so on. In addition, the process response is probably nonlinear so the maximum flow rate with a wide-open valve would probably vary with level in the tank. What is required is the apparent holdup time at the nominal operating point. An additional required parameter is the valve gain, K_V. (K_V replaces $F(s)$ in Figure 6-16.)

We now present a method by which we can determine the apparent holdup time and the valve gain at the nominal operating point from process tests. Both of these methods start with the supposition that the operation is stable with constant inflow, with constant level at the nominal set point, and with the controller in the automatic mode with its output somewhere between the extreme limits. Also, we assume that the inflow remains constant for the duration of the test. The valve should have a positioner, or at least be free of stiction and hysteresis.

Put the controller in manual and change the output by a small amount, say Δm percent. The outflow will change by an amount ΔF, and the level will begin changing (see Figure 6-18). After a certain period of time, say Δt, change the controller back to its original position. The level should stop changing. Now determine the change in level, ΔL, during the test. (Use the absolute values in percentage of full scale for ΔL, ΔF_{out}, and Δm.) The apparent holdup time can be estimated from:

$$T_L = \frac{\Delta F_{out}\,\Delta t}{\Delta L}\ , \qquad (6\text{-}27)$$

and the valve gain from:

$$K_V = \frac{\Delta F_{out}}{\Delta m}\ . \qquad (6\text{-}28)$$

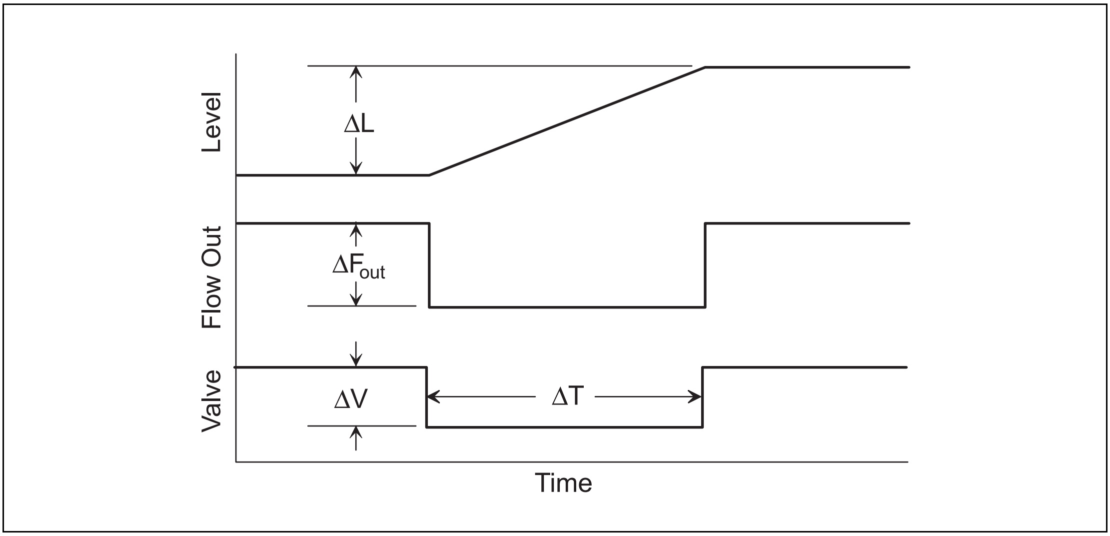


Figure 6-18. Testing for T_L and K_V

Dead Time

The ideal process model for a liquid-level loop contains neither dead time nor lag time. Real processes may contain either dead time or lag. For instance, a cascaded flow-control loop may have a finite response time. For specific forms of response, Tables 6-8 and 6-9, presented later in this section, contain correction factors for Φ, the ratio of dead time to holdup time. In addition, Table 6-8 contains the valve-gain factor, K_V. For cascade loops, K_V can be set to unity.

Unequal Inflow and Outflow

When considering step changes in feed rate, consider the feed rate as the liquid actually going to the reservoir, whether from the feed stream, from liquid falling from lower trays on a distillation tower, or from some other source. Hence, the change in feed rate, ΔF_{in}, can be the result of any cause, such as actual change in vessel feed rate, change in the percentage of liquid of the feed, change in reboiler heat, or change in liquid load on the trays in the distillation tower.

Flashing Liquid

In some cases, flashing liquid or foam and froth can produce a false indication of level. An example is the "shrink-and-swell" effect in a boiler drum. The shrink-and-swell effect approximates that of dead time; hence the dead-time correction factors of Tables 6-8 and 6-9 may apply.

♦ Modified Tables for Tuning Parameters

Tables 6-8 and 6-9 are modifications of Table 6-6 and 6-7 to account for two real-world phenomena:

- Noncascade control,
- Dead time in the level-control loop.

These correction factors were determined as a "best fit" to simulation results.

♦ Sinusoidal Disturbance

If an oscillating inflow is anticipated (for instance, as a result of the cycling of a control loop of an upstream process unit), then both the level and the outflow will oscillate with the same frequency. The maximum deviation in level (half the amplitude of its oscillation) may be more than would result from a step change in inflow equal to half the amplitude of inflow oscillation. Also, the level-control loop may act as an amplifier, causing the outflow's amplitude of oscillation to exceed that of the inflow. Hence, both the amplitude of oscillation for both the level and the outflow should be investigated, with the possibility that the tuning parameters will need to be modified.

If the frequency of oscillation of the inflow is known, then a key parameter is the ratio of this frequency to the undamped natural frequency (or simply "natural frequency") of the level-control loop, ω/ω_n. Table 6-7 or 6-9 gives relations for calculating the predicted period of oscillation, P, of the control loop. The natural frequency, in radians per minute, can be calculated from the following:

Table 6-8. Controller Tuning Parameters for Step Change in Inflow with Corrections for Dead Time and Noncascade Control

ΔF_{in} = max. step change in disturbance ΔL_{max} = max. allowable deviation of level from set point T_L = holdup time, minutes		θ = Dead Time $\Phi = \frac{\theta}{T_L} \quad (0 \le \Phi \le 0.5)$ K_V = Valve Gain (=1.0 if level is cascaded to flow)	
Decay Ratio DR	**Damping Factor** ζ	**Gain** K_C	**Integral Time** T_I
Critically Damped	1.0	$\frac{0.74\,\Delta F_{in}}{(1-\Phi)^{0.5}\,K_V\,\Delta L_{max}}$	$\frac{5.44\,T_L\,\Delta L_{max}}{\Delta F_{in}}$
0.05	0.430	$\frac{0.5\,\Delta F_{in}}{(1-\Phi)\,K_V\,\Delta L_{max}}$	$\frac{1.47\,T_L\,\Delta L_{max}}{(1-\Phi)^{0.25}\,\Delta F_{in}}$
0.25	0.215	$\frac{0.32\,\Delta F_{in}}{(1-\Phi)^{1.5}\,K_V\,\Delta L_{max}}$	$\frac{0.58\,T_L\,\Delta L_{max}}{(1-\Phi)^{0.9}\,\Delta F_{in}}$

Table 6-9. Predicted Performance for Step Change in Inflow with Corrections for Dead Time and Noncascade Control

Decay Ratio DR	**Level Response**			**Outflow Response**		
	Level Arrest Time T_{aL}	**Period** P	**IAE**	**Outflow** $\Delta F_{out\text{-}max}$	**Outflow Arrest Time** T_{aF}	**Outflow Max Rate of Change**
Critically Damped	$2.72\,(1-\Phi)^{0.96} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	Not Applicable	$7.39(1-\Phi)^{0.52} \times T_L \frac{\Delta L_{max}^2}{\Delta F_{in}}$	$\frac{1.14\,\Delta F_{in}}{(1-\Phi)^{0.1}}$	$5.44\,(1-\Phi)^{1.26} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$\frac{0.74}{(1-\Phi)^{0.70}} \times \frac{\Delta F_{in}^2}{T_L\,\Delta L_{max}}$
0.05	$2.14\,(1-\Phi)^{0.70} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$11.91\,(1-\Phi)^{0.87} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$4.61(1-\Phi) \times T_L \frac{\Delta L_{max}^2}{\Delta F_{in}}$	$\frac{1.34\,\Delta F_{in}}{(1-\Phi)^{0.22}}$	$4.27\,(1-\Phi)^{1.11} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$\frac{0.52}{(1-\Phi)^{1.37}} \times \frac{\Delta F_{in}^2}{T_L\,\Delta L_{max}}$
0.25	$1.87\,(1-\Phi)^{0.47} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$8.67\,(1-\Phi)^{0.66} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$5.45(1-\Phi)^{0.54} \times T_L \frac{\Delta L_{max}^2}{\Delta F_{in}}$	$\frac{1.55\,\Delta F_{in}}{(1-\Phi)^{0.25}}$	$3.74\,(1-\Phi)^{0.85} \times T_L \frac{\Delta L_{max}}{\Delta F_{in}}$	$\frac{0.61}{(1-\Phi)^{1.16}} \times \frac{\Delta F_{in}^2}{T_L\,\Delta L_{max}}$

$$\omega_n = \frac{2\pi}{P\sqrt{1-\zeta^2}} \tag{6-29}$$

Once the normalized frequency ratio is known, Figure 6-19 can be used to determine

$$K_C \frac{L(\omega)}{F_{in}(\omega)}.$$

If the frequency of oscillation of the inflow is not known, then a worst-case condition would be when

$$\frac{\omega}{\omega_n} = 1.$$

For an assumed amplitude of input oscillation, $F_{in}(\omega)$, if half the (peak-to-peak) amplitude of level variation, $L(\omega)$, exceeds ΔL_{max}, then recalculate a new value for K_C from

$$K_{C-new} = K_C \frac{L(\omega)/2}{\Delta L_{max}}. \tag{6-30}$$

(Note that we are adjusting K_C *upward* from its original value.) Then return to Table 6-7 and calculate a new value of T_I. (If you are using Table 6-19, use column 3 to back-calculate a new value for ΔL, resulting from the adjusted value K_C. Then use these new values for K_C and ΔL in column 4 to calculate T_I.)

Example 3.

In Example 2, for ΔF_{in} = 10% and ΔL_{max} = 5%, the following tuning parameters and predicted period of oscillation were calculated:

K_C = 1.0,
T_I = 3.55 minutes/repeat,
P = 28.7 minutes.

The natural frequency is given by:

$$\omega_n = \frac{2\pi}{28.7\sqrt{1-0.430^2}} = 0.24 \text{ radians/minute.}$$

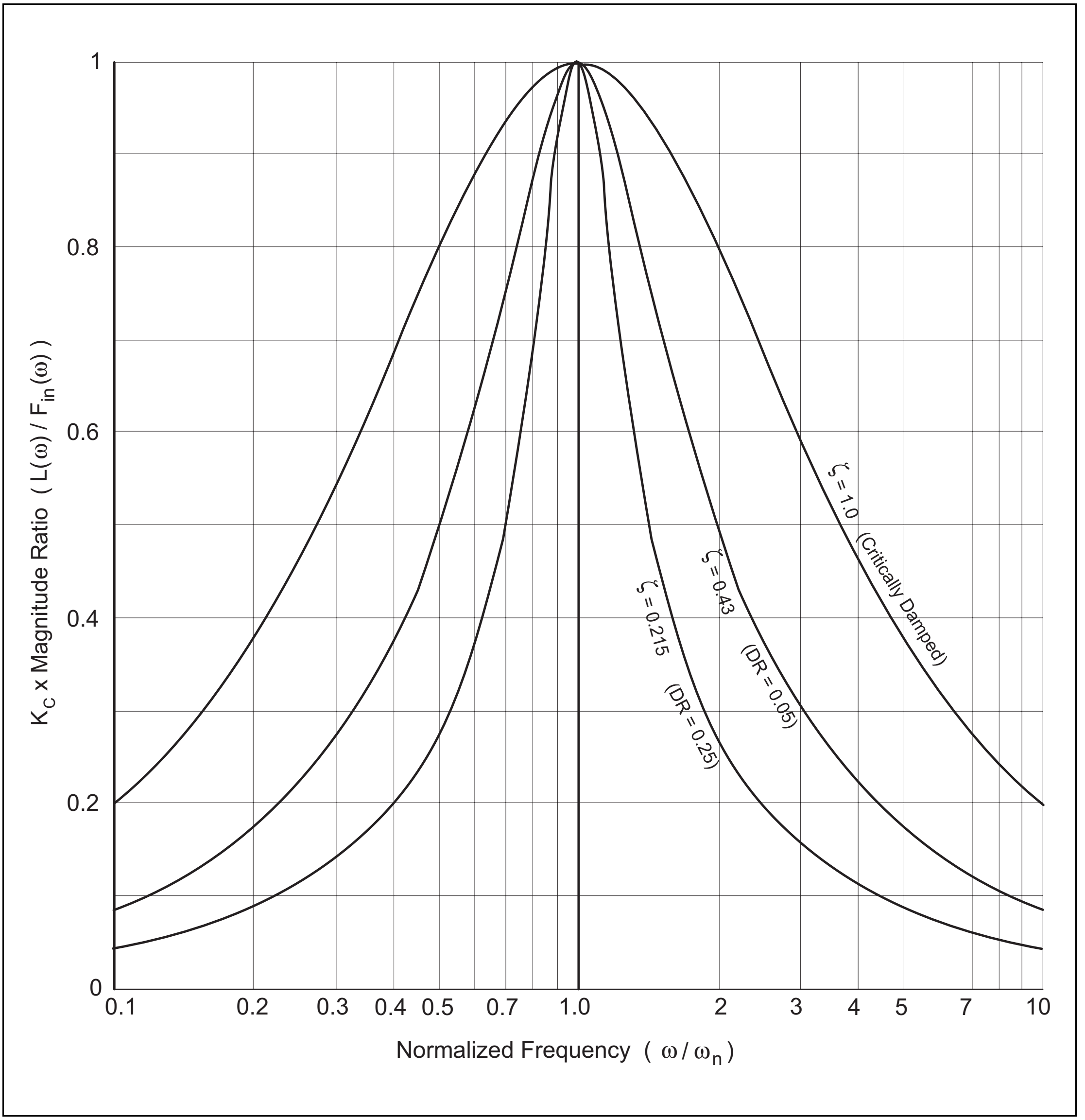


Figure 6-19. Magnitude Ratio of Changes in Level to Sinusoidal Changes in Inflow

If, rather than a step input change of 10 percent, we anticipate that, as a worst case, there will be an oscillating input of the same frequency whose peak-to-peak amplitude is twice this, then from Figure 6-9:

$$K_C \frac{L(\omega)}{F_{in}(\omega)} = 1.0 \times \frac{L(\omega)}{20} = 1.0 \, .$$

or $L(\omega) = 20$.

Since half the amplitude exceeds ΔL_{max}, then recompute K_C as

$$K_{C-new} = \frac{20}{2\times5}\times1.0 = 2.0$$

With this new value for K_C, reenter Table 6-7 and calculate T_I.

$$T_I = \frac{0.74\times4.7}{2.0} = 1.74 \text{ minutes/repeat.}$$

We had assumed that as a worst case, the frequency of oscillation was the same as the natural frequency, whatever value that is. If the frequency of input oscillation were fixed, say at 0.24 radians/minute (the same as the initial natural frequency of the loop), then we would proceed as we did before and calculate new values for K_C and T_I. With a new value for T_I, however, the predicted period of oscillation of the loop would be 14.08 minutes. Consequently, the value for ω_n would change to 0.49 radians/minute. At a normalized frequency ratio of 0.24/0.49, from Figure 6-19,

$$K_C \frac{L(\omega)}{F_{in}(\omega)} = 0.5$$

so that $L(\omega) = 0.5\times\frac{20}{2} = 10$.

Half of this amplitude is equal to ΔL_{max} so our adjusted tuning is satisfactory. If our new amplitude were still greater than $2 \times \Delta L_{max}$, we would have had to calculate a further adjustment for K_C and T_I.

End of Example 3.

For a range of values of the normalized frequency ratio, Figure 6-20 depicts the magnitude ratio for the amplitudes of oscillation of outflow and inflow. This table illustrates the fact that if the normalized frequency ratio, ω/ω_n, is near to unity, the level-control loop will act as an amplifier for the outflow. If the ratio is considerably less than 1, the outflow and inflow will oscillate at approximately the same amplitude. If the ratio is greater than unity, however, the amplitude of outflow oscillation will be considerably attenuated from that of the inflow. (By *inflow*, we refer to the net inflow into the liquid pool in the vessel. This may differ from the total liquid inflow into the vessel itself.) As a consequence of this, we deduce that if two level-controlled vessels (such as distillation towers) are in series, it is preferable that the level-control loops be tuned so their natural frequencies are separated. In particular, the natural frequency for the second vessel should be greater than the first, to take advantage of the attenuation of inflow oscillation provided by the first vessel. Since the vessel holdup times are fixed and the assumed magnitude of disturbances should not be varied to produce the desired tuning characteristics, the primary "handle" that we have for ultimately determining the natural frequencies is ΔL_{max} for each vessel. If we have already chosen ΔL_{max} as large as possible for the first vessel, then we should choose a smaller ΔL_{max} for the second vessel, if possible. If it is not possible to make the natural frequency of the second vessel greater than that of the first vessel, we can go in the other direction and deliberately make it slower than the first vessel. At the very least, the natural frequencies of vessels in series should be separated by a factor of two.

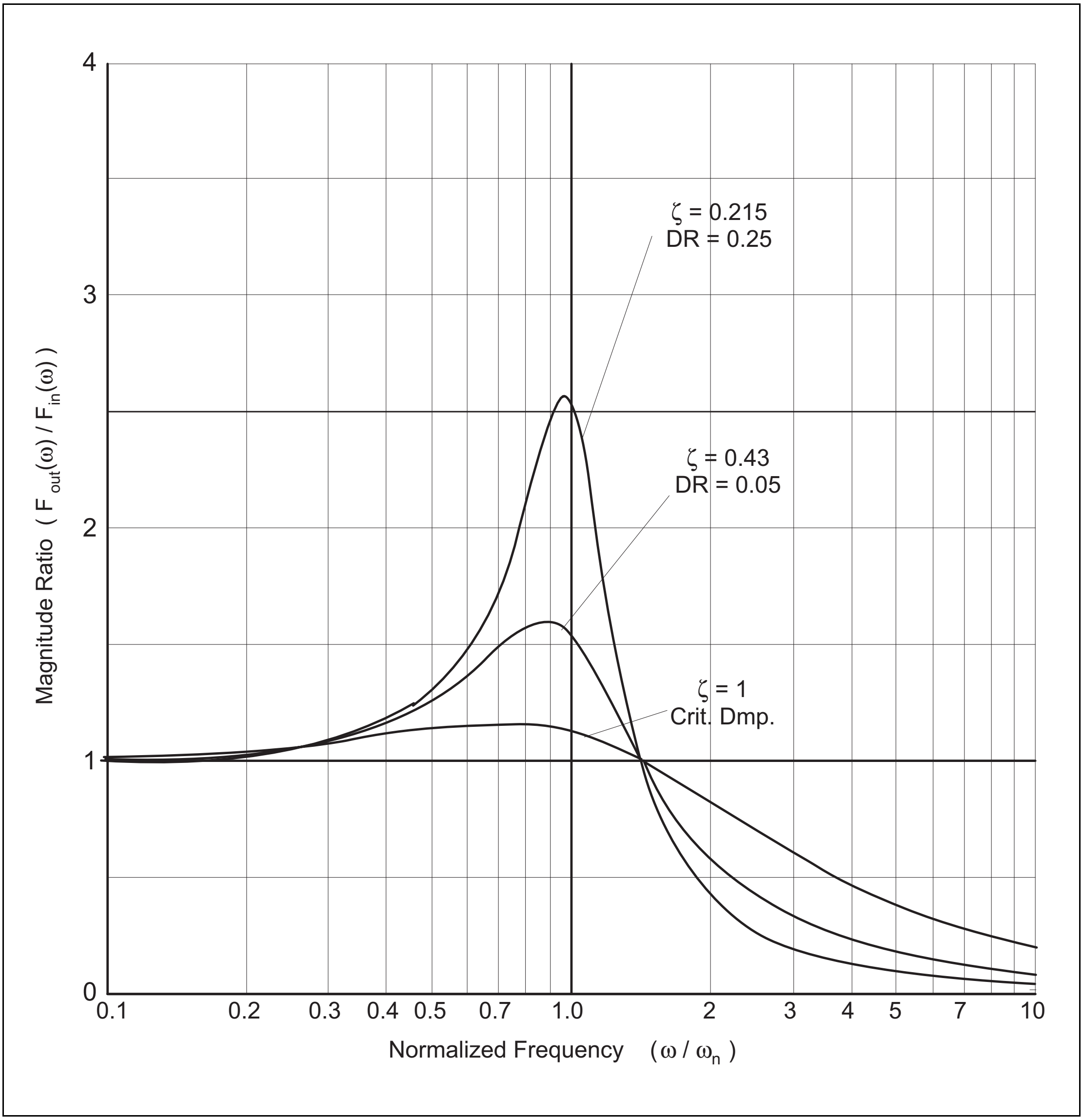


Figure 6-20. Magnitude Ratio of Changes in Outflow to Sinusoidal Changes in Inflow

♦ Other Approaches to Liquid Level Controller Tuning

Ziegler-Nichols Tuning

Some engineers have reportedly used a Ziegler-Nichols approach for tuning liquid-level loops. If the response of an open-loop process test shows an identifiable amount of dead time, then the parameter determination method depicted by Figure 6-7 and Equations 6-8 and 6-9 can be used. Table 6-1 can then be used to determine tuning parameters. The disadvantage of this technique is that it does not take into consideration either the worst-case disturbance nor the allowable maximum deviation of level.

Proportional-Only Control

If there is no reason to maintain the level at a fixed set point, but the aim is rather to prevent deviations beyond a particular limit, then you can use a proportional-only controller. Assuming that the allowable excursions above and below set point are equal, then the controller gain should be set according to this equation:

$$K_C = \frac{100}{2\,\Delta L_{max}} \qquad (6\text{-}31)$$

or the proportional band to $2 \times \Delta L_{max}$. The bias (manual reset) should be set to 50 percent. With this arrangement, if the disturbance is such that a controller output of 50 percent is required, the level will be at set point. Otherwise, there will be a steady-state offset between set point and level measurement. On a step change in inflow, the level will respond as a first-order lag with a time constant equal to the holdup time divided by the controller gain.

For a given value of ΔL_{max}, this technique provides the lowest rate of change of the outflow, hence the minimum amount of disturbance to a downstream process unit. This can be a real advantage because when the inflow is very low and a large increase can be expected, there is more room to absorb the inflow. When the inflow is very high the level is high, so there is a large volume to allow a slow decrease in the outflow. The disadvantage of this method is that the level is rarely at set point. This is more of a disadvantage from the standpoint of the operator's acceptance of the technique than a technological limitation, however.

If closer control about the set point is desired, the controller gain can be increased. Although there is no theoretical upper limit for the gain on a proportional-only controller for an integrating process, in practice, this will be limited by resonance that may occur within the loop. If the level sensor is an external cage type, there may be manometer effect between the liquid in the vessel and the liquid within the level-sensor cage. This will appear as an oscillation within the control loop, even though the total mass holdup may be unchanging. If the liquid has a large surface area, a resonant sloshing may occur, with a period that is proportional to the cross-sectional dimension. For a probe or other type of point-source measurement, this will also show up as an oscillation within the loop. Furthermore, splashing, such as from upper trays in a distillation tower, may cause noise to appear on the level measurement. Thus, there will be a practical limit to the controller gain. Even so, many level loops are successfully controlled using a controller gain as high as 10 or 20 (proportional band of 5% to 10%). With a high gain, any measurement noise present will cause excessive valve action. Therefore, the gain may be reduced, in favor of utilizing some integral action within the controller.

Averaging Liquid-level Control

We mentioned earlier that many liquid-level loops are not critical. It is possible to tolerate fluctuation, even offset, if it smooths out the flow to a downstream process unit. This can be accomplished by lowering the controller gain and using a small amount of integral action. Such a technique is called "averaging liquid-level control" since it maintains the long-term average of the level at the set point.

Nonlinear Control

Some manufacturers provide a nonlinear control algorithm that has the effect of increasing the controller gain as the measurement gets further away from set point. The "error-squared algorithm" was introduced in chapter 5. This algorithm has a very low gain at set point, with increasing gain as the measurement gets further away from set point. Some manufacturers accomplish a similar function by using linear characterization of the error signal. Sometimes the nonlinearization is applied only to one controller mode, such as proportional mode or integral mode, with the other controller modes seeing the normal error signal. This approach provides a form of averaging level control.

An interesting approach for surge tank and averaging level control is presented in reference [6-6]. Here the following form of error-squared algorithm for PI control is recommended:

$$m = K_C\left(\frac{e|e|}{100} + \left(\frac{|e|}{100}\right)^2 * \frac{1}{T_I}\int e\,dt\right) \quad (6\text{-}32)$$

When written in the form:

$$m = K_C\frac{|e|}{100}\left(e + \frac{|e|}{100 * T_I}\int e\,dt\right) \quad (6\text{-}33)$$

it is obvious that this is a form of scheduled tuning in which both the effective controller gain, K_C, and the effective integral time, T_I, are made proportional to the absolute value of error. The result is that the product, effective controller gain times effective integral time, is constant, at all values of error.

$$\hat{K}_C = K_C\frac{|e|}{100},$$

$$\hat{T}_I = T_I \frac{100}{|e|}.$$

Therefore $\hat{K}_C \hat{T}_I = K_C T_I$.

According to reference [6-6], other forms of implementation of error-squared PI control (including that depicted by Figure 5-8) may go unstable, since the product, effective gain times the effective integral time, varies with error.

There are certain analogies between this concept and the analytically developed technique for tuning liquid level loops presented earlier. From Equation 6-21 it is seen that:

$$K_C T_I = 4\zeta^2 T_L \tag{6-34}$$

Thus, for a selected decay ratio, which determines ζ through Equation 6-26, and a fixed holdup time, T_L, the product of gain and integral time is constant. While the analogy probably cannot be stretched too far, there are additional analogies related to controller tuning which can be made:

- A larger value for the product, $K_C T_I$, will cause a more stable control loop.
- For a fixed value of $K_C T_I$, a larger value of K_C, consequently smaller value for T_I, will reduce both the maximum deviation due to a step load change and the period of oscillation.

As far as is known, no manufacturer provides an error-squared algorithm which is the equivalent of Equation 6-32 or 6-33. This form could be custom implemented by the special programming capabilities of many manufacturers' systems, however.

❖ OTHER TUNING SITUATIONS: RUNAWAY PROCESSES

Runaway processes are unstable in the open loop. That is, if the controller is left on manual, the measurement will continue to climb or fall until some physical (possibly disastrous) limit is reached. An example is an exothermic reactor. As the temperature rises, the reaction rate increases, causing a greater rate of heat evolution, and consequently a more rapid rise in temperature. Such processes are often controlled by modulating the flow of cooling water to the jacket of the reactor.

An interesting phenomenon occurs with runaway processes. As with most processes, an excessive controller gain will cause the loop to oscillate. But, in contrast with most processes, a gain that is too low will also cause loss of process control by allowing the measurement to continue to climb or fall. Thus, there is both an upper limit and a lower limit for the controller gain. The difference between these limits is the permissible window for the gain. If there is a wide window, then the process should be relatively easy to control. On the other hand, much trouble may ensue if the window is narrow.

Exothermic chemical reactors are often developed by first constructing a laboratory or pilot-sized unit. Suppose such a reactor has been built and, because of the wide window of permissible controller gain, is relatively easy to control. Then, the reactor is scaled up to a production-sized unit. As the size is increased, the volume, which determines the rate of heat evolution, increases as the cube of the dimensions, whereas the jacket surface area, which determines the rate of heat removal, increases as the square of the dimensions. These two factors cause the acceptable window for controller gain to decrease as the size of the reactor is increased. In other words, a small, easy-to-control reactor may scale up to a large, very difficult-to-control reactor.

Often an exothermic reactor is controlled by measuring the reactor (or reactor effluent) temperature and cascading this temperature controller to a jacket water-temperature controller. Whether the cascade is present or not, a significant amount of derivative action is usually required in the reactor temperature controller to improve the stability of the loop. If a cascade is present, it can also have derivative action. This is one example of where derivative on error, rather than derivative on measurement, is preferred. The secondary controller may or may not have integral action.

Admittedly, we have not given any tuning rules for the runaway process, since each process will be decidedly different. However, we have discussed some of the factors influencing the choice of controller modes as well as some of the tuning problems involved.

❖ TYPICAL TUNING VALUES FOR PARTICULAR TYPES OF LOOPS

If the valve and sensor are properly sized and ranged, then most loops of a particular type will have similar tuning parameters. Thus, by using a table of typical values, one can start with tuning values that are at least “in the ballpark.” Table 6-10 presents typical values for the four most common types of control loops: flow, temperature, pressure, and level.

❖ PRACTICAL CONSIDERATIONS FOR LOOP TUNING

Suppose you’re called into the control room to correct an alleged loop tuning problem. Probably the worst thing you could do is to immediately begin changing a tuning parameter. There are several things you should do first.

Table 6-10. Rules of Thumb for Tuning Common Control Loop

Loop Type	Gain (Prop Band)	Reset, Mins/Repeat (Repeats/Min)	Derivative, Minutes
Flow	0.4 – 0.65 (150% - 250%)	– 0.25 (4 – 10)	None
Temperature	2-10 (10% - 50%)	2 – 10 (0.1 – 0.5)	to 2.0 (always less than reset)
Pressure, Gas	20 – 50 (2% - 5%)	May not be needed	None
Pressure, Liquid	0.5 – 2.0 (50% - 200%)	– 0.25 (4 – 10)	None
Pressure, Vapor	2 – 10 (10% - 50%)	2 – 10 (0.1 – 1.0)	0.1 to 2.0 (always less than reset)
Level	See text	See text	None
Composition	– 1.0 (100% - 1000%)	10 – 30 (0.03 – 0.1)	Varies

(1) Find out as much as you can about the loop. If this is not a new loop, then has this tuning problem just started? If so, what has changed? Is the process operating at a different condition? Is the set point different? Does this happen some of the time but not others?

(2) Put the loop on manual. If the oscillations persist, there is certainly not a tuning problem with this loop; rather, some other loop is oscillating and that is being reflected by the oscillations of this loop.

(3) Consider other equipment in the loop. Give particular consideration to the valve. Is it sticking? Is it operating very close to either end of its calibrated travel?

(4) Understand the process phenomena. Is there something unusual about the process or its behavior?

(5) If there is a controller tuning log available (it is a "must" for every control room!), then examine it. What has been the tuning history of the loop?

(6) Finally, if you are convinced that this truly is a tuning problem, then make note of the existing tuning parameters. Try to improve the loop performance by the technique "improving as-found tuning parameters" described earlier, rather than by making either open-loop or closed-loop process tests.

Once you have found new parameters, test the process by making a small set point change, and if possible, a small load upset. Before leaving, note these in the tuning log, along with any process data (throughput rate, feed composition, product specifications, etc.) that are pertinent to this particular situation.

❖ REFERENCES

6-1. J. G. Ziegler and N. B. Nichols. "Optimum Settings for Automatic Controllers," *Transactions of the ASME,* vol. 64, 1942, pp. 759-68.

6-2. G. H. Cohen and G. A.Coon. "Theoretical Considerations of Retarded Control," Taylor Instrument Companies Bulletin #TDS-10A102.

6-3. A. M. Lopez, J. A.Miller, C. L. Smith, and P. W. Murrill. "Controller Tuning Relationships Based on Integral Performance Criteria," *Instrumentation Technology,* November 1967, pp. 57-62.

6-4. B. G. Lipták, ed. *Instrument Engineers' Handbook – Process Control*, 3d ed., Chilton Book Co., 1995.

6-5. E. B. Dahlin. "Designing and Tuning Digital Controllers," *Instruments and Control Systems,* June 1968, p. 77.

6-6. *Product Application Manual*, Techmation, Inc., Tempe, AZ. Also available from http://www.protuner.com/app6.pdf.

7

SELF-TUNING

Self-tuning, or auto-tuning as it may also be called, is a technology that has made rapid strides in the past few years. Although the theoretical concepts underlying it have been around for some time, only recently has computing power become inexpensive enough for calculation-intensive self-tuning techniques to be economically viable in commercial products. Today, practically all distributed control systems and microprocessor-based single-loop controllers, even those aimed at the low end of the marketplace, include some type of self-tuning.

There are several reasons to use self-tuning. If the process is non-linear—that is, if it does not exhibit the same response at one operating point as another—and if the process is operated under widely varying conditions, then the controller's tuning parameters should be changed to match operating conditions. This change could be accomplished by using an operator's log that tabulates various combinations of tuning parameters for different operating conditions. This method assumes that each operating condition can a priori be associated with a particular combination of tuning parameters. It also depends on an operator or instrument technician's diligence in entering the appropriate tuning parameters as operating condition changes. However, if process characteristics change rapidly, or cannot be categorized from simple measured data, it probably is unreasonable to expect frequent manual changes to the tuning parameters to produce satisfactory results.

An entirely different motivation for using self-tuning is the desire to have a procedure that will determine an acceptable set of tuning parameters automatically during startups. This would minimize the task of a control engineer or instrumentation technician in manually tuning the loops.

A similar need for self-tuning exists if the end user cannot be presumed to have the knowledge or experience to successfully tune the control loops manually. This problem is frequently faced by manufacturers of lower-priced single-loop controllers (e.g., the "quarter-DIN" controllers. "DIN" refers to a standard instrument faceplace and panelboard cut-out size.) whose market often does not include sophisticated control equipment users.

The variety of motivations for using self-tuning naturally leads to the emergence of different techniques, and a variety of self-tuning techniques are now on the market. Generally, these can be grouped into the following categories: scheduled tuning, on-demand tuning, and adaptive tuning.

❖ SCHEDULED TUNING

With scheduled tuning, one or more tuning parameters are changed automatically as process conditions or the operating point change. "Adaptive gain" is one form of scheduled tuning, in which the control algorithm receives an external input that represents the controller gain. It is the user's responsibility to provide the correct value for the controller gain, either by means of a table lookup or user-written program. Many controllers have provisions for adaptive gain.

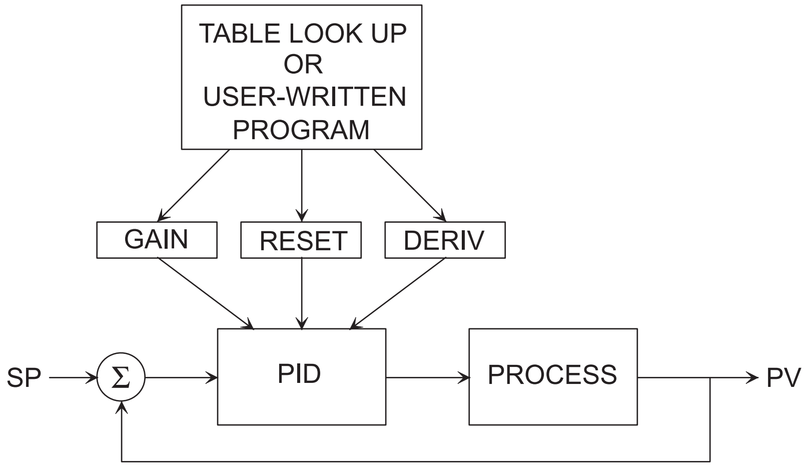
A simple extension of this adaptive gain concept permits all three tuning parameters—gain, reset, and derivative—to be set from an external input. For example, a controller may provide for a table lookup to determine which set of preset tuning parameters to use, as shown in Figure 7-1. Three sets of tuning parameters are stored in the table; this represents three different operating regions. Also stored with each set are limits that determine the boundary for that region, indexed by a key variable. The user can designate the variable that is the key to the table. This may be the set point, process variable, error, controller output, or some variable not directly related to the control loop. When this variable is within a region defined by specified boundary limits, then the related tuning parameters are entered to the controller. If the controller uses the position-mode algorithm, then some form of "bumpless tuning," as described in chapter 5, will be required.

Scheduled tuning is merely an automation of the "operator's log" concept, in which predetermined tuning parameters are manually entered according to the current operating point. It should be noted that whereas the controller itself operates in the closed loop, the tuning procedure is open loop. That is, if erroneous tuning values are specified for one or more regions, the controller uses these values just as if they had been directly entered through a tuning display. The controller makes no attempt to assess and modify its own performance by determining improved tuning parameters. Even so, this method is a considerable improvement over attempts to determine a single set of tuning parameters for all situations ("one size fits all"). Scheduled tuning is probably more reliable than depending on manual entry of predetermined tuning for various conditions. Furthermore, how the tuning parameters will be changed is completely obvious and within the control of the user. This feature may not be present in more automated tuning procedures.

❖ ON-DEMAND TUNING

One form of on-demand tuning is simply an automation of the open-loop testing method for controller tuning that was described in the previous chapter. In its simplest form, the controller uses existing tuning parameters until a command is given or a button labeled "tune" is pressed. Then the following procedure is carried out automatically:

- The control loop is switched to manual;
- The controller output is changed by a specified amount. The amount of change is usually established in advance by the user as a configuration parameter;



KEY VARIABLE FOR REGIONS: ___________
(Can be SP, PV, Error or Other Measured Variable)

REGION	BOUNDARIES	GAIN	RESET	DERIV
1	0 - 30%	G1	R1	D1
2	30% - 70%	G2	R2	D2
3	70% - 100%	G3	R3	D3

(Example Figures)

Figure 7-1. Scheduled Tuning

- The response to the step change in controller output is sampled and stored until the complete response is observed;

- Some type of numerical analysis is performed on the response data to determine an approximate mathematical model for the process; this may be as simple as first-order plus dead time;

- Once process model parameters have been determined, some type of correlating equations are utilized to determine tentative values for controller tuning parameters. These equations could be one of the equations sets presented in Table 6-1, or they could be proprietary equations provided by the manufacturer;

- The tentative tuning values are displayed to the user for confirmation. If confirmed, they are inserted into the control algorithm.

Several elaborations on this on-demand tuning procedure may be incorporated. For example, one manufacturer's self-tuning algorithm first makes a single step to determine the approximate time constant and dead time of the process. The step is removed and the process allowed to settle, then a second step is applied for a length of time equal to 2.5 times the sum of the preliminary estimates of dead time and time constant. This step permits the process gain to be estimated. When this step is removed, the algorithm makes a final estimate of dead time and process time constant. The objective of this sequence is to obtain better process response data than a single step test would provide.

Another elaboration of this on-demand tuning procedure could be to fit a more precise process model to the data than a simple first-order plus dead time.

In all of these cases, the essence of the technique is the same—make an open-loop step test, observe the process, and calculate tuning parameters.

The problems with this on-demand technique are the same as those presented in chapter 6 for manually initiated open-loop testing. Furthermore, a load upset may occur midway through the response that masks the controller output's response to the step change. Because of these problems, the requirement for confirmation permits plant personnel to apply discretion before the parameters are used on line by the control algorithm.

The advantages of this on-demand tuning technique are its simplicity and the fact that the user does not need expertise in controller tuning. For these reasons, this technique is widely used, especially by the so-called quarter-DIN controllers that are targeted for the lower end of the controller market.

The technique can also be combined with other techniques. For instance, it can be combined with scheduled tuning to determine tuning parameters for multiple operating regions. Within any operating region, the user can command a "tune" operation, which initiates the bump-test procedure described previously. If the resulting parameter set is confirmed, then the parameters are entered into the table for the appropriate operating region.

Another approach to on-demand tuning, based upon the work of Åström and Hägglund (Ref. 7-1), is called the "relay method." It is related to the closed-loop tuning technique described in chapter 6. When the command to tune is given, the controller is left in automatic at its current operating point. This approach utilizes preconfigured high and low limits on the controller output. These limits should be a selected amount above and below the current controller output. An on/off control strategy (relay controller) is used. This causes the controller output to vary in a square-wave fashion between the minimum and maximum output limits. Consequently, the process variable will oscillate in an approximate sine wave as shown in Figure 7-2.

If the time in which the controller output dwells on one limit exceeds the dwell time at the other limit, then both the high and low limits are moved an equal amount in the direction of the longer dwell time. When the output response is symmetrical (dwell times at each limit are equal), the period of oscillation is the same as the ultimate period, P_U, which is determined by the closed-loop test method described in chapter 6. (This is proven in Ref. 7-1.) Furthermore,

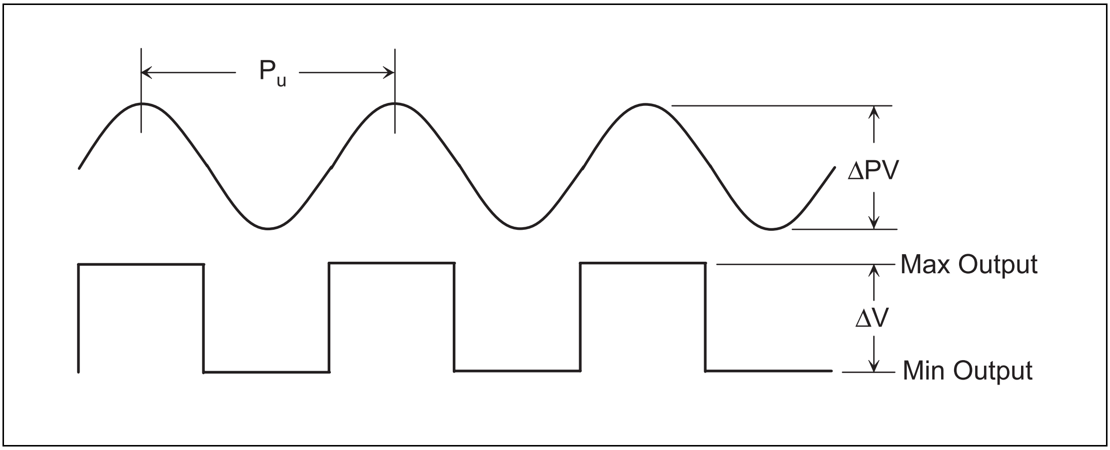


Figure 7-2. Controller Output and PV Response for Relay Tuning

the ultimate gain, K_{CU}, is a function of the ratio of amplitude of oscillation of the process variable and controller output. Specifically:

$$K_{CU} = \frac{4}{\pi}\frac{\Delta PV}{\Delta V} \tag{7-1}$$

where: ΔV = Amplitude of square-wave oscillation of controller output
ΔPV = Amplitude of oscillation of process variable.

Once K_{CU} and P_U have been determined, then an appropriate tuning parameter correlation, such as Table 6-2, can be used to determine tuning parameters. This procedure, or some modification of it, has been automated and is the basis of several vendors' automated loop-tuning procedure.

❖ ADAPTIVE TUNING

There are several techniques for on-line tuning in which the tuning parameters are determined by an auxiliary program that automatically evaluates the closed-loop behavior and calculates and modifies the tuning parameters whenever necessary. The Foxboro EXACT™ is an example.[1] This tuning procedure observes the pattern of the response and then invokes a set of rules for determining new tuning parameters that will drive the pattern closer to a desired response pattern (Ref. 6-2). In essence, the technique attempts to mimic what an experienced, competent human would do in tuning the loop. According to Foxboro, the EXACT technique offers the following features:

- It does not require artificial load upsets; instead, it utilizes the normal process disturbances that occur;

1. EXpert Adaptive Controller Tuner. EXACT is a trademark of The Foxboro Co.

- It does not attempt to impose an arbitrary mathematical model on the process.

As a part of the technique's setup procedure, the user establishes a band around the set point, as shown in Figure 7-3. This band should be wider than the normal noise band of the process. As long as the process variable remains within this band, the self-tuner is deactivated, although the loop may be under closed-loop control utilizing existing tuning parameters.

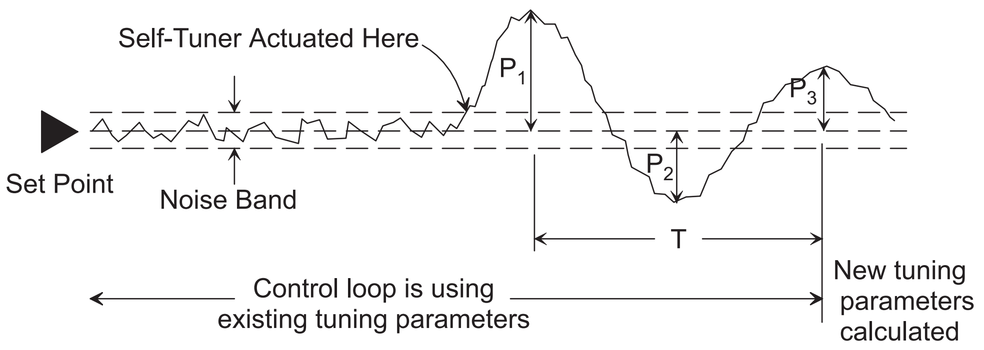


Figure 7-3. Pattern Recognition Used by Foxboro's EXACT™ Self-tuning Algorithm

Suppose that a load upset occurs that forces the process variable outside this band. The self-tuner is then activated and begins to observe the pattern of the response. The pattern is characterized by three successive peak deviations from set point P_1, P_2, and P_3, and the period of oscillation as measured by the time, T, between the first and third peaks. When the three peaks have been determined, the damping ratio, *DMP*, and overshoot ratio, *OVR*, are calculated from:

$$DMP = \frac{P_3 - P_2}{P_1 - P_2} \tag{7-2}$$

$$OVR = -\frac{P_2}{P_1} \tag{7-3}$$

The user, again as a part of the setup procedure, can specify target values for these ratios. In essence, this is equivalent to specifying the desired shape of the response. Usually, these ratios are not independent; hence, the technique does not attempt to drive both ratios to the specified targets. Instead, the ratio that is closer to its target is determined. The difference between this ratio and the target is used to calculate multiplying factors, which are applied to the present tuning values to calculate new values. The period of oscillation is also used to adjust the reset and derivative settings. Bear in mind that the actual steps are based on a series of heuristic rules, not on a formal mathematical procedure. The newly calculated parameters are used to update the working parameters in the PID algorithm.

The rules contain many statements that are intended to cope with anomalous conditions such as these:

- "What if there are false peaks?"
- "What if the second and/or third peak cannot be distinguished from the noise band?"
- "What if the control loop is overdamped, so that upon a load or set point change, only one peak occurs?"

There are several setup parameters to be entered. We have mentioned the noise band and target values for the damping and overshoot ratios. Other parameters include initial values for the three tuning parameters, a factor that determines the extent to which the algorithm will use derivative (an entry of 0.0 makes the algorithm use the proportional and integral modes only), the maximum allowable damping and overshoot, a factor that places an upper and lower limit on the calculated proportional band and integral time values, the maximum wait time for the third peak, and a high-frequency output cycling limit. Even these setup parameters can be determined in a semiautomated fashion, using the "pretune" procedure that is an integral part of the EXACT package. This is similar to the "on-demand" open-loop tuning previously described. The user places the controller in the pretune mode, specifies an amount of output change (amount of process "bump"), and then enters a command to begin the process test. This test allows the controller to learn enough about the process to calculate tentative values for the setup parameters. The user can either accept these or change any of them at will. Once the pretune test is completed, the EXACT procedure can be enabled to maintain control-loop tuning automatically thereafter.

Although the EXACT procedure can be left in the enabled state indefinitely, many users prefer to enable it only for a short time so as to determine acceptable tuning parameters, then disable it (thus retaining constant tuning parameters) until it is felt that the loop again needs retuning.

Reports from users of this technique have been mixed. Some installations have reported very good results—typical of these has been pH control, where the process is both highly nonlinear and time variable. Other reports have noted problems with processes that are too quiescent, processes in which load disturbances and measurement noise are essentially indistinguishable, and control loops that interact with other loops. The fact that problems have been cited should not be interpreted as an indictment of the technique, however. They merely force us to recognize that for this, as for any self-tuning technique, there are both more favorable and less favorable applications.

❖ TUNING AIDS

A number of tuning aids are available as software packages from third-party vendors. Typically, these are executed in a Windows®-based computer that can access process data in a variety of ways[2]:

- Using OPC or DDE servers;
- Through A/O converters connected to a particular control loop's I/O;
- Using custom-designed software that connects directly to a DCS or PLC data highway;
- Using an ASCII file from a data historian.

These packages vary in their features and capabilities, but in general all of them receive input process data, analyze this data, and calculate recommended tuning parameters. The better-known packages provide the following features:

- The loop can be tested in either the manual or automatic mode.
- The data can be edited to remove noise spikes.
- The user can choose from a number of performance criteria, including fast or slow response to a load upset, the form of set point response, or lambda tuning.
- The user can choose a safety factor, which in essence provides a trade-off between robustness of the loop and loop performance.
- The package provides a database of controllers, by manufacturer and control algorithm type, so the calculated tuning parameters are specific to the controller being used.
- After calculating tuning parameters, the user can ask "what-if"-type questions, in order to explore other values of the response selection or safety factors.
- A number of analysis tools can be utilized, making it possible to display robustness plots, standard deviation of the variable before and after tuning, and simulated responses to set point and load upsets using the recommended tuning parameters. (Robustness plots show how much variation in process gain or dead time can be tolerated while still maintaining loop stability.)
- If the package is connected to the loop in real time, then on command, the recommended tuning parameters can be downloaded to the controller.

2. "Windows" is a registered trademark of Microsoft Corporation.

❖ REFERENCES

7-1. K. J. Åström and T. Hägglund. *PID Controllers: Theory, Design and Tuning,* ISA – The Instrumentation, Systems, and Automation Society, 1995.

7-2. T. W. Kraus and T. J. Myron. "Self-Tuning PID Controller Uses Pattern Recognition Approach," *Control Engineering,* June 1984, pp 106-11.

8 ADVANCED REGULATORY CONTROL

The term *advanced control* is often used, abused, misused, and overused. Even among knowledgeable control engineers, the term does not have a consistent meaning. It is used to refer to everything from cascade control, which can be implemented with analog instrumentation, to optimization and model-based predictive control, which usually require a host computer interfaced to a lower-level distributed control or data acquisition system.

We use the term *advanced regulatory control*. By this we refer to a collection of control techniques, from ratio and cascade up through decoupling control and dead-time compensation. The unifying concept underlying the techniques that we include in advanced regulatory control is that they can all be implemented with hardware modules or, what is more likely, with software function blocks in a microprocessor-based control system. With technology such as FOUNDATION™ Fieldbus, it is also possible for these techniques to be implemented with function blocks distributed in fieldbus devices. We will not focus on the platform for implementation, however, other than to say that the techniques covered here are on a level below the control tasks that are traditionally reserved for a host computer, such as optimization, scheduling, and model predictive control.

The topics that we consider to be in the category of advanced regulatory control include:

- Ratio control,
- Cascade control,
- Feedforward control,
- Override control,
- Control of multiple input, multiple output processes (decoupling control),
- Dead-time compensation and elementary model-based control.

The last topic in this list will serve as a lead-in to model predictive control, for which an overview will be presented in chapter 15. Model predictive control can truly be called "advanced process control" rather than advanced regulatory control.

These topics can all be considered as tools in a control system engineer's toolkit. Just as a good workman is competent in the use of his tools, so the control engineer should know how and when to use each of these tools. But, like the workman, he or she should not feel obligated to use each tool on every project.

Before we begin the discussion of individual topics, we will first investigate the motivation for using advanced regulatory control (or any form of advanced control, for that matter). Suppose that at some time (to be designated as "time 1") we observe the record or trend display for a process variable, along with the controller output signal, shown in Figure 8-1. Both variables are steady, and the process variable is at set point.

At some time later (time 2), we again observe the record or trend of these variables. Again, both variables are steady, and the process variable is right on set point. But between time 1 and time 2, a significant load change must have occurred because the controller output at time 2 is considerably different from its value at time 1.

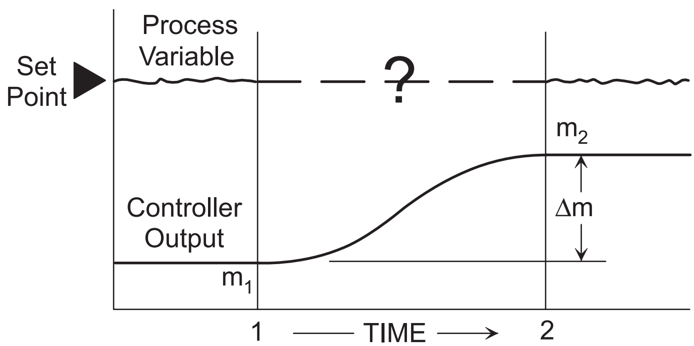


Figure 8-1. Two Separate Observations of the Behavior of a Control Loop

Let us suppose that we do not know what happened between time 1 and time 2, and we have no historical data that can be recalled (not a very realistic situation, but go along with me anyway). Do you think it is possible for the controller to have been tuned well enough for the process variable to have remained on set point between time 1 and time 2, that is, to have ridden through the load change without deviation?

Regardless of your intuitive answer, let's do a bit of formal analysis. Using Equation 4-3, we compute the controller output at time 1 as follows[1]:

1. We can use a proportional+integral controller here, rather than PID. The derivative mode would have no role in the discussion that follows.

$$m_1 = K_C\left(e_1 + \frac{1}{T_I}\int_0^1 e\,dt\right) + m_0 \qquad (8\text{-}1)$$

Note that we are talking about a particular time (time 1), so we use a definite integral here, with subscripts and limits on the integral sign, which represent a particular time. The lower limit of integration, time “0”, can refer to the last time that the controller was switched into automatic, and the term m_0 refers to its initial output value at that time.

Since the *PV* and *SP* were equal at time 1, the e_1 is zero. Equation 8-1 can be simplified as follows:

$$m_1 = \frac{K_C}{T_I}\int_0^1 e\,dt + m_0 \qquad (8\text{-}2)$$

Similarly, since the error at time 2 is also zero, the controller output at time 2 is given by:

$$m_2 = \frac{K_C}{T_I}\int_0^2 e\,dt + m_0 \qquad (8\text{-}3)$$

This integral can be broken into two regions, giving:

$$m_2 = \frac{K_C}{T_I}\int_0^1 e\,dt + \frac{K_C}{T_I}\int_1^2 e\,dt + m_0 \qquad (8\text{-}4)$$

We are interested in the *change* in controller output, which is given by:

$$\Delta m = m_2 - m_1.$$

Subtracting Equation 8-3 from Equation 8-4 yields:

$$\Delta m = \frac{K_C}{T_I}\int_1^2 e\,dt \qquad (8\text{-}5)$$

Equation 8-5 highlights the fact that in order to have feedback control action, there must be an error. In this case, the error is represented by the area under the curve between times 1 and times 2. There are an infinite number of possible trajectories of the process variable from time 1 to time 2. One possible trajectory is shown in Figure 8-2. All possible trajectories would have the same net area, given the required valve movement from m_1 to m_2 and the tuning parameters K_C and T_I.

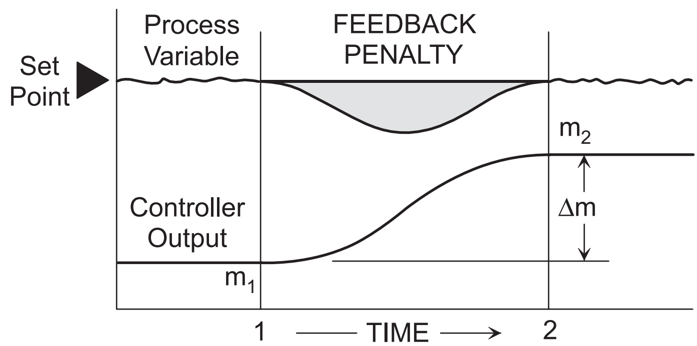


Figure 8-2. The Feedback Penalty

We can surmise (and this is confirmed by practical experience) that if the load change is very gradual, the deviation between set point and measurement would be very small. Our analysis is consistent with this. If times 1 and 2 are widely separated, the same net area under the curve is still required, but since it is distributed over a long time period, the maximum deviation is reduced. (Think of cutting a rubber sheet the shape of the shaded area in Figure 8-2, then stretching the ends of the cut-out sheet.) Thus, for a slow disturbance, feedback control by itself is adequate. For a faster disturbance, something more is needed.

This discussion points out the curse of feedback control—to have control action, a penalty must be paid in the form of an error (deviation) in the feedback loop. This fact presents both a challenge and an opportunity for advanced regulatory control. How can you obtain the required control action (valve movement) without paying the feedback penalty?

This challenge is the theme of the following chapters.

9

CASCADE CONTROL

❖ CASCADE CONTROL TECHNOLOGY

When one steps beyond basic feedback control, cascade and ratio control are probably the first of the so-called advanced regulatory control techniques one encounters. In a cascade control system, one feedback controller adjusts the set point of another feedback controller. The upper-level controller is called the "primary," while the lower level is called the "secondary." A typical application of cascade control is a temperature controller cascaded to a flow controller. Figure 9-1 depicts a cascade control system, using both ISA (Ref. 9-1) and SAMA symbols (Ref 9-2).

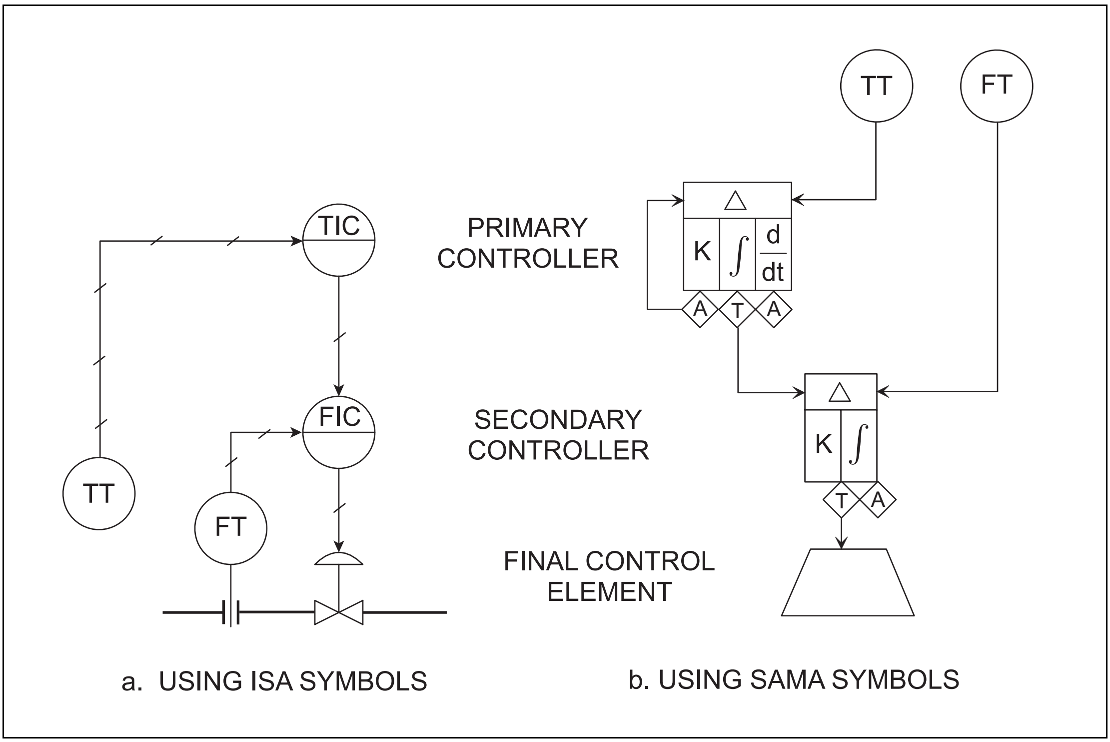


Figure 9-1. Symbolic Representations of a Cascade Control System

Before investigating the finer points of cascade control systems, let us first consider a situation that illustrates the motivation for using cascade control. In Figure 9-2a the process outlet temperature of a heat exchanger is sensed. The temperature controller then adjusts the set point of the steam-flow controller to maintain the outlet temperature at set point. An alternative control strategy would be for the temperature controller to directly manipulate the control valve, as shown in Figure 9-2b. To compare these two schemes, we need to consider the disturbances to the process. We will consider several sources of disturbance, both with and without the cascade control system.

First, suppose there is no cascade, as shown in Figure 9-2b, and suppose also that the process flow rate increases. This increased load on the heat exchanger will cause the outlet temperature to drop. The temperature controller will react by increasing the signal to the valve. The new valve position will in turn cause an increase in steam flow. The effect of this change in steam flow must then pass through the heat exchanger before its corrective effect is felt by the temperature sensor.

Now, suppose that the temperature controller cascades a steam-flow controller, as shown in Figure 9-2a. Suppose that there is the same increase in the process flow rate as before. Again, there will be a drop in outlet temperature. The temperature controller will react by increasing the set point of the steam-flow controller, which, in turn, will increase the signal to the valve. With the increased valve position, the flow will quickly respond to the increased demand from the temperature controller. The effect of this change in steam flow must then pass through the heat exchanger before its corrective effect is felt by the temperature sensor.

These two scenarios illustrate the fact that with this particular disturbance, the cascade control loop provides little or no difference in response. Consider two other scenarios. We begin as we did before, with no cascade—the temperature controller is connected directly to the valve. This time, the assumed disturbance is a drop in the steam header pressure. At the current valve position, this will cause a drop in steam flow; after the effect passes through the heat exchanger, this results in a drop in outlet temperature. From there on, the events are the same as in the first scenario. The temperature controller reacts by increasing the valve position. This restores the steam flow, but the effect must then pass through the heat exchanger before the corrective effect is felt by the temperature sensor. The response of the temperature loop to this disturbance is approximately the same as it was to the increase in process flow.

One last scenario. We return to the cascade control configuration and suppose there is a drop in steam header pressure. The steam-flow rate drops, but this is sensed by the flow transmitter. The flow controller quickly responds by increasing the valve position, restoring the flow to the set point demanded by the temperature controller. The net effect is that the steam flow is disturbed only momentarily. Given that this momentary disturbance must pass through the heat exchanger whose dynamics will tend to filter the effect, the process outlet temperature will show very little of the effect of this disturbance.

With this example in mind, look at the block diagram of a generic cascade control loop shown in Figure 9-3. This figure shows an inner loop and an outer loop. The inner loop is comprised

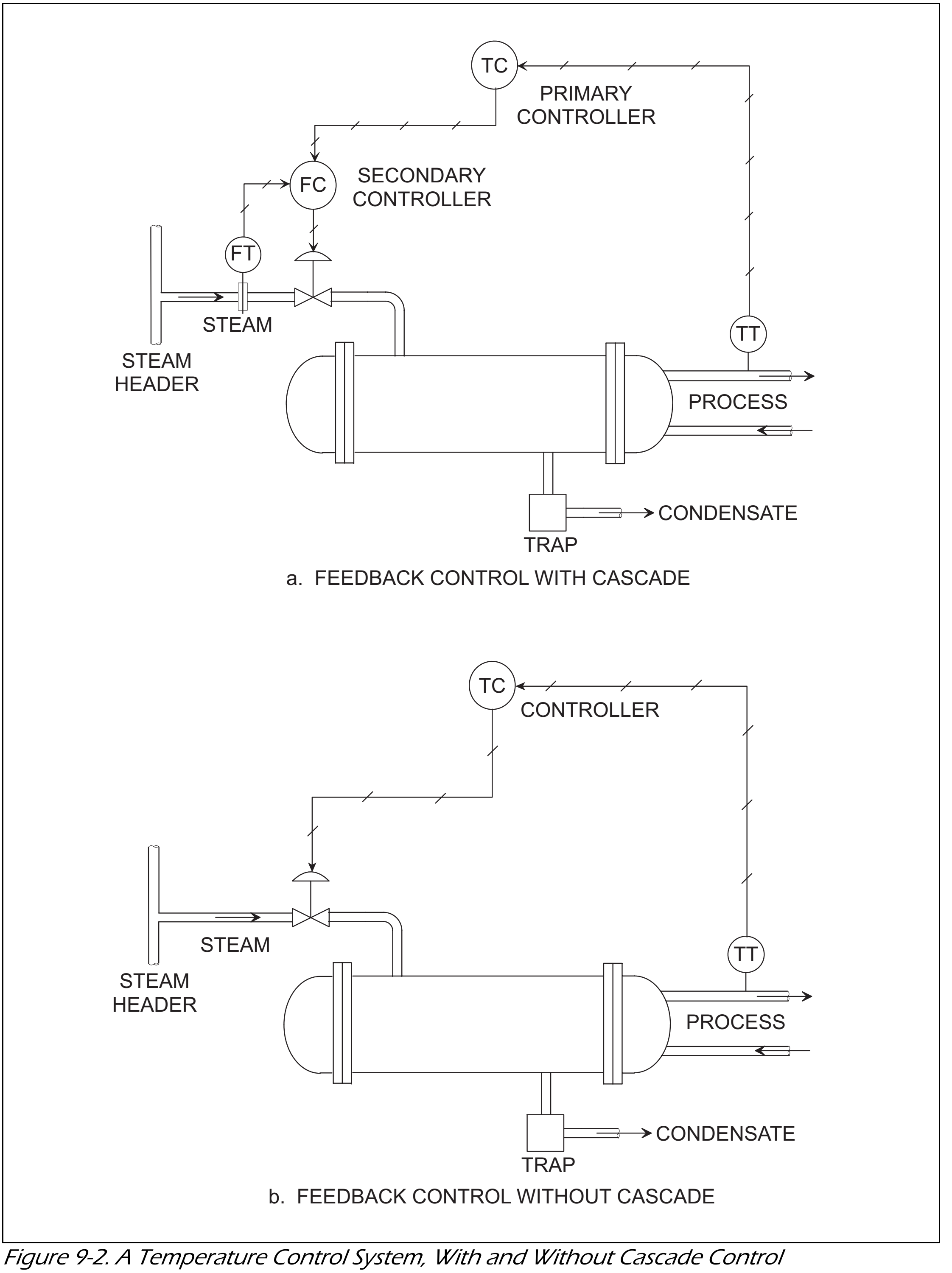


Figure 9-2. A Temperature Control System, With and Without Cascade Control

of a secondary controller and a secondary process. The primary controller and the primary process are components of the outer loop. The inner loop is also a component of the outer loop, since the primary controller sets the set point of the secondary controller. Also, the output of the secondary process is an input to the primary process.

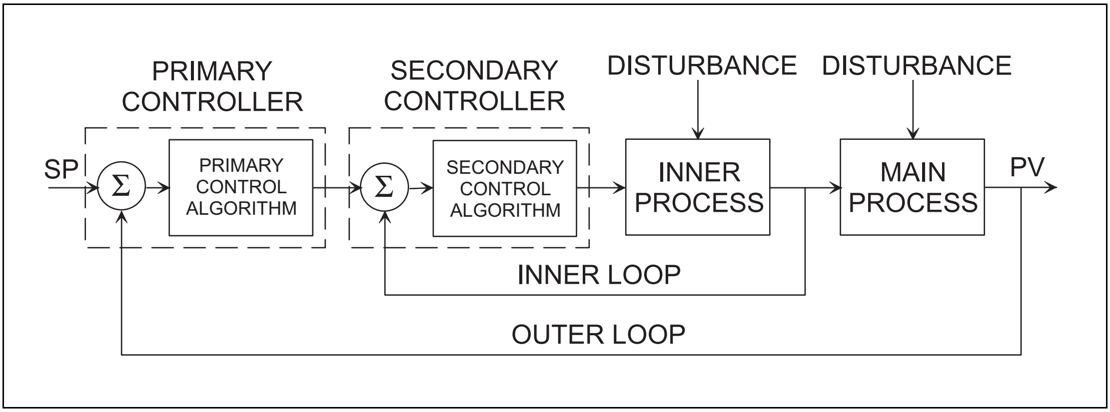


Figure 9-3. Block Diagram of a Cascade Control System

In the example shown in Figure 9-2, the primary controller is the temperature controller, the secondary controller is the flow controller, the primary process is the heat exchanger itself, and the secondary process is the flow-control valve. The output of the secondary process (steam flow) is the input to the primary process (heat exchanger). It is also measured and becomes the process variable for the flow controller.

The block diagram shows two disturbances, one into the primary process and one into the secondary process. The primary process disturbance is analogous to a change in process flow, whereas the secondary disturbance is analogous to changes in the steam header pressure.

The results of analyzing the example can be extended to the more general case. For disturbances that enter the primary loop but are outside the secondary loop, the cascade is of little benefit. For disturbances that enter the inner loop, the secondary controller is very beneficial in compensating for them and preventing the disturbances from affecting the outer loop. If the secondary controller were not present then only the outer loop would exist. A disturbance to the secondary process would inherently be a disturbance in the outer loop, and we would have to pay the feedback penalty at the primary controller to compensate for it. When the secondary loop is present, it provides the necessary control action without paying the feedback penalty in the primary loop.

One requirement for cascade control that we have alluded to but not explicitly stated is that the inner (secondary) loop should be significantly faster than the outer (primary) loop. There is no precise definition of "significantly faster." A good rule of thumb is that the frequency of oscillation of the secondary loop, when well tuned, should be at least three times that of the primary loop, also when well tuned. If the primary loop's speed of response is near that of the secondary loop, adverse interaction ("fighting") can occur between the primary and secondary controllers. Furthermore, if primary loop responds quickly, then cascade control may not be necessary in the first place, since a single feedback controller can cope with equal ease to disturbances to the secondary or primary process.

Normally, the loops' relative speed of response is not a problem, since with most common applications of cascade, such as temperature cascaded to flow, the difference in speed of response will be greater than the rule of thumb requires. If, for some reason, the primary and secondary loops have a similar speed of response, then the primary loop can be dampened by reducing the controller gain and lengthening the integral time.

❖ IDENTIFYING CANDIDATE APPLICATIONS

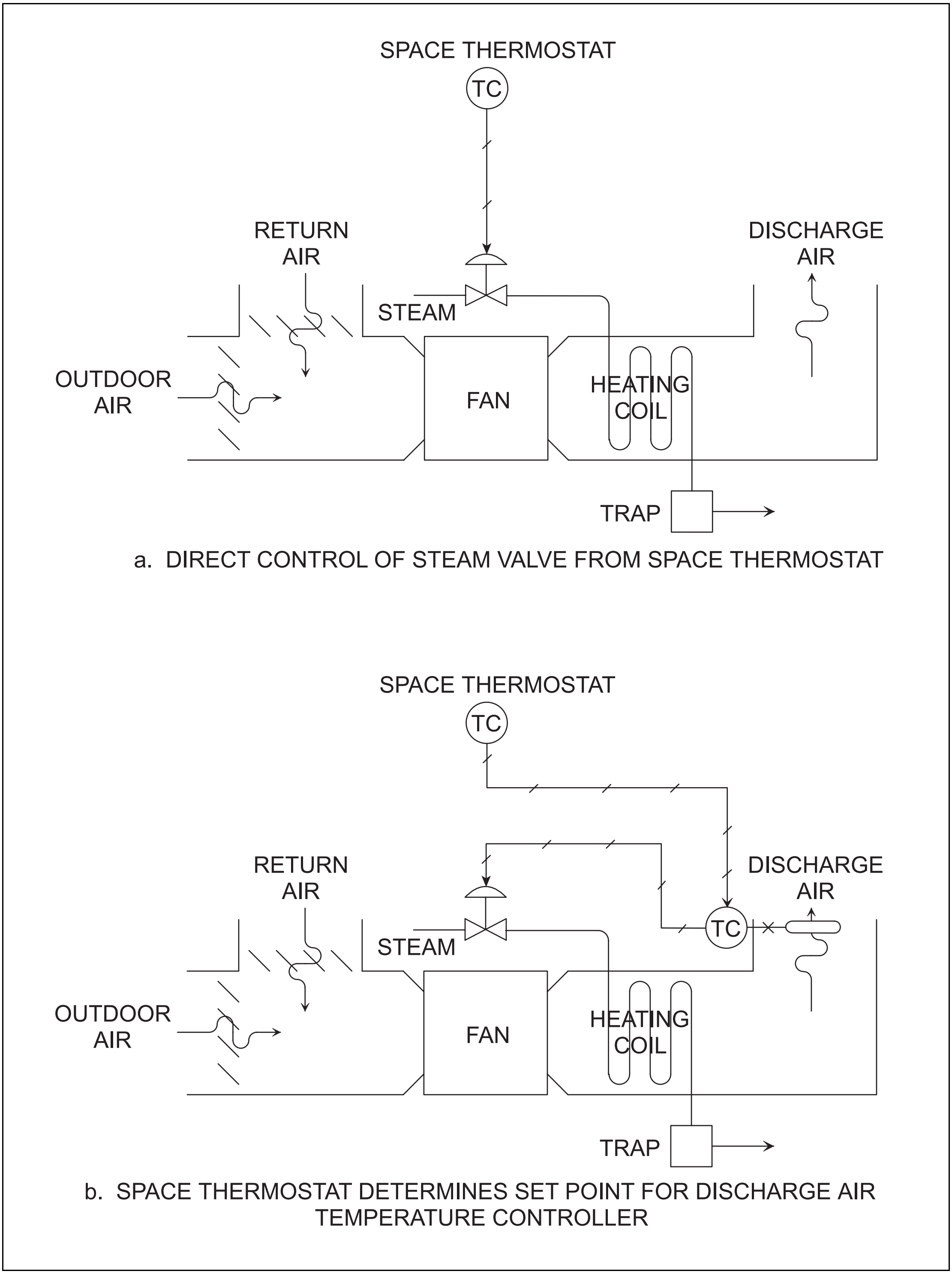
How should one go about spotting likely candidates for cascade control? Suppose that you have a process with a primary measurement (process variable) and a feedback controller and final control element (e.g., valve). First, analyze the source of disturbances. Then ask, "Is there an intermediate measurement that can be used to close an inner loop that will encompass the disturbance?" If so, this application is a candidate for cascade control. Consider the following example from the heating, ventilating, and air conditioning (HVAC) industry.

Example

An air-handling unit for a commercial building consists of a fan, heating coil, and discharge air duct to deliver heated air to the space. In the conditioned space, a throttling-type thermostat (not the on/off type common in residential applications) senses the temperature and controls a steam valve on the heating coil. Return air from the space is mixed with outdoor air at the inlet to the fan. The fixed ratio of outdoor air to return air is used to meet the requirements for ventilation (see Figure 9-4).

Suppose that we are presented with the following problem. In normal operation, the thermostat does an adequate job of maintaining a stable space temperature. However, in the particular climate of this application, the outside air temperature sometimes drops very rapidly. When it does, the mixed air temperature drops, as does the discharge air temperature. This eventually causes a drop in the space temperature. The thermostat senses and corrects for this, but because of the large volume of space, it takes an excessively long time to recover to the desired temperature. Can we improve the control system to alleviate this problem?

We begin by asking, "Is there an intermediate measurement that could be made and used in an inner feedback loop that would encompass the disturbance?" One possible solution is to measure the steam flow and cascade the output of the thermostat to a steam-flow controller. This is similar to the heat-exchanger example that was used at the beginning of this chapter. An intermediate measurement (steam flow) is used in an inner feedback control loop. The flaw in this solution is that the inner loop (steam flow) does not encompass the disturbance (change in outdoor air temperature). In other words, this is a solution to the wrong problem. Had we been presented with the problem of rapid fluctuations of the steam header pressure, then this would have been the appropriate solution.



a. DIRECT CONTROL OF STEAM VALVE FROM SPACE THERMOSTAT

b. SPACE THERMOSTAT DETERMINES SET POINT FOR DISCHARGE AIR TEMPERATURE CONTROLLER

Figure 9-4. A Cascade Control Application in the HVAC Industry

Another possible solution is to measure the discharge temperature and let that control the steam valve, as shown in Figure 9-4b. When the outside air temperature drops, and consequently the mixed air temperature, the discharge temperature controller will sense this. The discharge temperature control loop will be rapid, in comparison with the space temperature control loop. Hence, the discharge temperature will be maintained approximately constant at its set point.

The drop in outside air temperature will also affect the space temperature (due to heat loss through the walls, windows, etc.), albeit on a much slower time scale. The output of the thermostat will gradually increase the set point of the discharge temperature controller. The maximum deviation of space temperature from the thermostat setting will be minimal, however.

This is the correct solution to this problem. We have found an intermediate measurement, namely, discharge temperature, with which we can close an inner control loop. This inner loop encompasses the disturbance, that is, the drop in outdoor air temperature.

What if we had been presented with both problems—variations in steam header pressure and a drop in outdoor air temperature. It would be technologically feasible to employ a flow controller to encompass the first disturbance, a discharge air temperature controller to encompass the second disturbance, and finally a thermostat to control space temperature. This would be a three-level cascade control system. This solution would probably be overkill, however. If the discharge temperature controller alone were used as the inner loop, it would encompass both of the disturbances. True, a steam-flow controller would minimize the effect of steam pressure variations on the discharge temperature. However, the discharge temperature control loop is sufficiently fast, relative to the space temperature loop, that we can tolerate the short-term effect of steam-pressure variations on discharge temperature. Hence, the flow controller is unnecessary.

This does not mean that multi-level cascade control systems should never be used. The author has personally supervised the installation and commissioning of a four-level cascade system for a distillation tower, shown in Figure 9-5. The innermost loop was reflux flow; this was cascaded from an upper-tray temperature controller, which in turn was cascaded from an overhead composition controller. The outermost loop was a composition control loop from a downstream distillation tower.

Working from the innermost loop outward, each of the loops was slower than the one beneath, yet not so slow that it was not beneficial to apply the next lower level of cascade to encompass disturbances to that loop.

Though there may be no obvious disturbance to a secondary loop (such as a change in steam header pressure in the heat-exchanger example), benefits may still accrue from cascading a primary process variable controller to a secondary flow controller. Adverse conditions involving the valve, such as a sticking valve, mis-sized valve, or wrong valve characteristics, will be more or less confined to the inner loop and will have a much less detrimental effect on the outer loop. In essence, when a flow controller manipulates a valve with a positioner, this con-

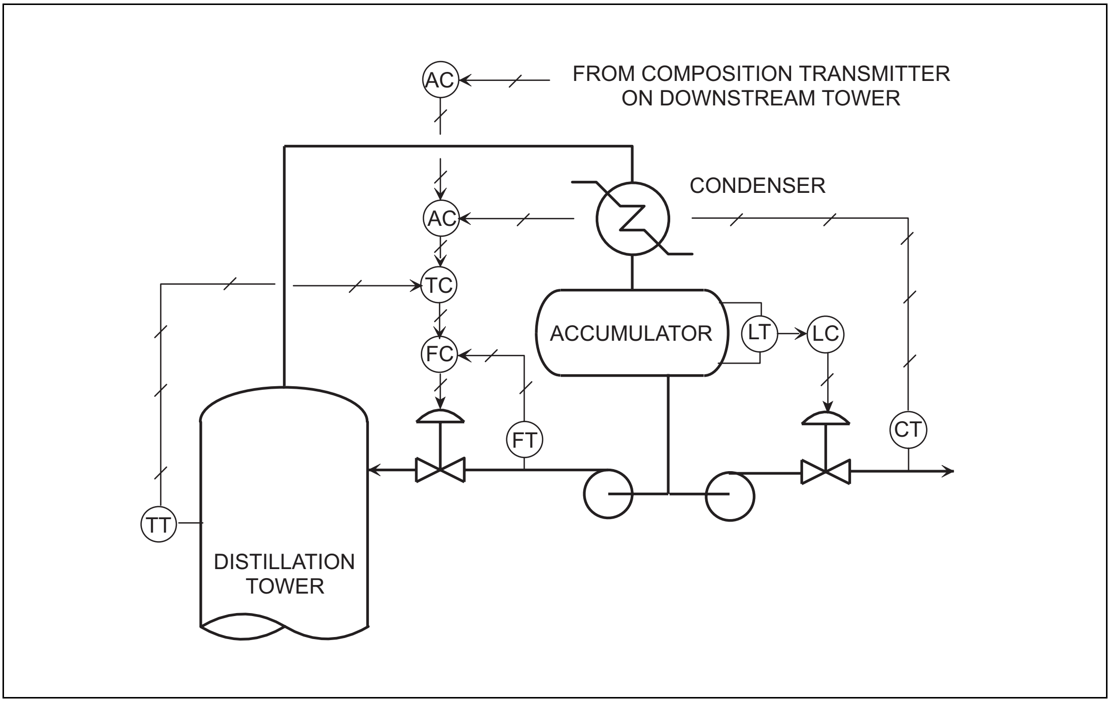


Figure 9-5. A Multi-level Cascade Control Application in the Petroleum Industry

stitutes a cascade control loop, with the flow controller serving as the primary controller and the valve positioner serving as the secondary controller. Although this configuration is not commonly called "cascade," it has the technological characteristics, as well as benefits, of cascade control.

In addition to situations where a cascade loop is used to prevent disturbances to the secondary loop from affecting the primary loop, feedforward control represents another application in which one should use a cascaded secondary (usually flow) controller. (Feedforward control is discussed in chapter 11.) With feedforward control, it is much easier to determine the change in flow rate (or mass or energy) that is required to compensate for a disturbance than it is to determine a change in valve position.

❖ IMPLEMENTATION, OPERATION, AND TUNING

♦ Implementation

Using traditional, analog-based control devices, the primary and secondary controller will be separate devices, with the output of the primary controller (4–20 mA signal) connected to the set point of the secondary controller. With this type of technology, no status information is passed between the devices, other than the set point for the secondary controller. Thus, manual/automatic switching and bumpless transfer are achieved by operator manipulation.

With current technology, however, cascade loops, as well as most other types of control functions, will be implemented using some type of microprocessor-based system. The details of implementation and operation will differ from one manufacturer to another. The following description will agree in its general underlying concept with all manufacturers' systems.

The cascade loop will be configured as a series of software function blocks, as shown in Figure 9-6. These function blocks are executed sequentially. A list of the blocks, with their function and order of execution is as follows:

- Primary analog input block (AI)—converts the primary measuring signal into a numerical value that is deposited in a specific memory location;
- Secondary analog input block (AI)—converts the secondary measuring signal into a numerical value that is deposited in a specific memory location;
- Primary PID controller (PID)—if in the automatic mode, fetches the set point from its memory location as well as the measurement value from the output of the primary AI block, calculates an output value, and leaves it in a specific memory location;
- Secondary PID controller (PID)—if in the automatic mode, fetches its set point from the output of the primary PID as well as the measurement value from the output of the secondary AI block, calculates an output value, and leaves it in a specific memory location;
- Analog output (AO)—fetches the output value from the secondary PID and converts it into an analogous electrical output signal (e.g., 4–20 mA) for transmission to the valve.

♦ Operation

The operation of a cascade loop depends upon the type of hardware employed. With conventional analog control instrumentation, both the primary and secondary controller will have a manual/automatic switch, plus a means for adjusting the controller output manually. In addition, the secondary controller will have a cascade-local switch for selecting the set point source. If the loop is in fully cascade-automatic, then when the secondary is switched to manual or to local set point, the primary must also be switched to manual to prevent it from winding up. To return the loop to fully cascade-automatic requires considerable manual manipulation of switches and settings, such as the following:

- Match the secondary set point to its process variable,
- Switch the secondary to automatic,
- Match the primary output to the secondary set point,
- Switch the secondary to cascade,

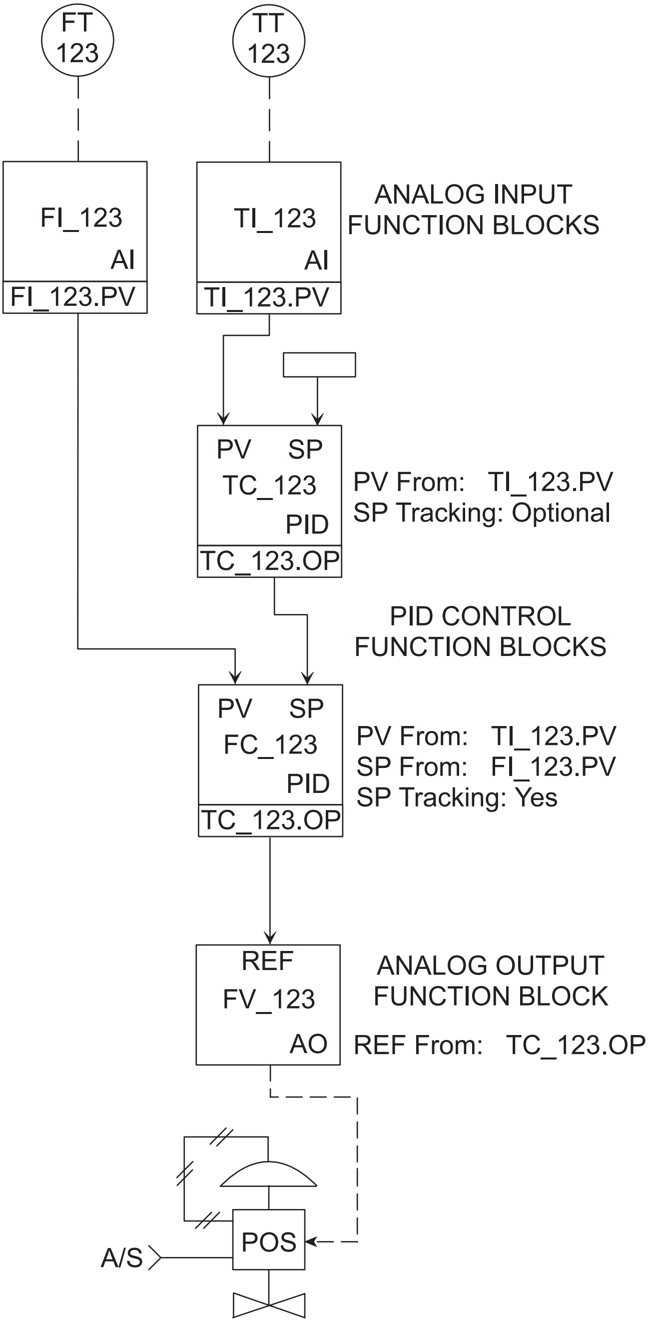


Figure 9-6. Implementation of Cascade Control Using Manufacturer's Function Blocks

- Match the primary set point to its process variable,
- Switch the primary to automatic.

With digital processor technology, however, for the operator to switch to manual control of the valve or to local secondary set point entry, he or she merely has to switch the secondary controller. If the secondary controller was switched to manual, its set point is forced to track the process variable. If it is switched to local, the secondary set point is a value entered by the operator. In either case, the primary controller recognizes the status change and goes into an initialization mode wherein its output is made equal to the secondary set point.

While the secondary controller is returned to fully cascaded-automatic, the primary controller returns to the automatic mode. Since its output has been tracking the secondary controller set point, no bump will occur in its output.

These details show that, from the operator's viewpoint, manual/automatic switching is much simpler using a microprocessor-based system than with a conventional analog system, since the manipulations required to achieve bumpless transfer are performed automatically.

Furthermore, windup protection can easily be built into systems implemented in software. If the output of the secondary is limited, say, with limits of 0 and 100 percent, then when the output reaches one of those limits, the primary controller could wind up if adequate provisions have not been made. For example, suppose the primary is a temperature controller, cascaded to a flow controller, which controls a fail-closed steam valve. Both the primary and the secondary will be set for reverse action. Should a deficiency in steam supply occur, then the secondary controller will drive its output to the maximum limit (100%) in an attempt to achieve its set point. Once it reaches the maximum limit, further increases in its set point will have no effect. The secondary function block will communicate this status to the primary block. This will inhibit the primary block from increasing the secondary set point. The primary will be allowed to decrease the set point of the secondary, however, once control is again achievable.

Should the secondary block drive its output to a minimum value, the reverse situation will occur. The primary block will be inhibited from decreasing the secondary controller's set point, but will be permitted to increase its set point. These mechanisms prevent windup in the primary, should the secondary become saturated. If there is a multiple-level cascade, the same type of mechanism will be passed to higher levels of the cascade, preventing windup at all levels.

If the controllers are provided with external feedback (see chapter 12 for a description of external feedback), then another method for preventing windup in the primary controller is to use the secondary measured variable as the external feedback to the primary controller.

♦ Commissioning and Tuning

When commissioning a cascade control system, one should start with the inner loop and work outward. Put the primary controller in manual and tune the secondary controller. Once the secondary controller is tuned and in automatic operation, it becomes merely a part of the process as seen by the primary controller. The primary can then be tuned using any of the tuning procedures discussed in chapter 6, then placed in automatic operation.

❖ CASCADE CONTROL USING FOUNDATION™ FIELDBUS

The basic architecture of FOUNDATION Fieldbus (FF) was described in chapter 5, together with a detailed description of the basic feedback control strategy. We described the use of the BKCAL_OUT signal, containing both a parameter value and status bits, from the AO block to the BKCAL_IN port of the PID block. When the mode in the AO block is set to Auto (instead of the usual Cascade), the PID is informed through the BKCAL_OUT to BKCAL_IN link and can thus initialize its output, thus assuring bumpless transfer when the block is switched from manual to automatic. This same link also forces the PID block into IMAN if the valve position is limited, either physically or in software. This prevents windup of the PID block (Refs. 9-3, 9-4, and 9-5).

In the cascade control strategy, this concept is extended from the secondary PID to the primary PID. A BKCAL_OUT signal (parameter value plus status bits) from the secondary is transmitted to the PID BKCAL_IN port of the primary. If the secondary PID is in an initialization mode or its set point source is not in cascade then this signal forces the primary PID into an initialization mode (IMAN). When the secondary is in IMAN, its set point is forced to track its process variable, if the set point tracking option is set. The parameter value contained in the BKCAL_OUT signal that is transmitted to the primary is the set point of the secondary. The primary output is forced to track this value. This prevents windup of the primary PID if its block output is broken in any way. In summary, the cascade initialization is back propagated (e.g., all the way from the valve, up through the secondary PID, to the primary PID initializing all controllers in its path).

Since the function blocks are resident in field devices, several different physical configurations are possible. Each analog input transmitter can contain its own PID, with the secondary PID's output transmitted to an AO block in the valve transmitter. This configuration, however, requires two forward and two backward links if the output is to be transmitted on the bus. A better alternative is for the primary PID to be in its analog input transmitter and the secondary PID and AO block to be in the valve actuator. This requires two forward and one backward links on the bus. However, if the valve actuator supports two PID function blocks, an even better configuration is for both the PIDs to reside in the valve actuator, each receiving a signal from its respective analog input transmitter. This requires only the two forward signals on the bus, which thereby minimizes bus loading.

❖ REFERENCES

9-1. ISA-5.1-1984 (R 1992), *Instrumentation Symbols and Identification*. ISA – The Instrumentation, Systems, and Automation Society, 1992.

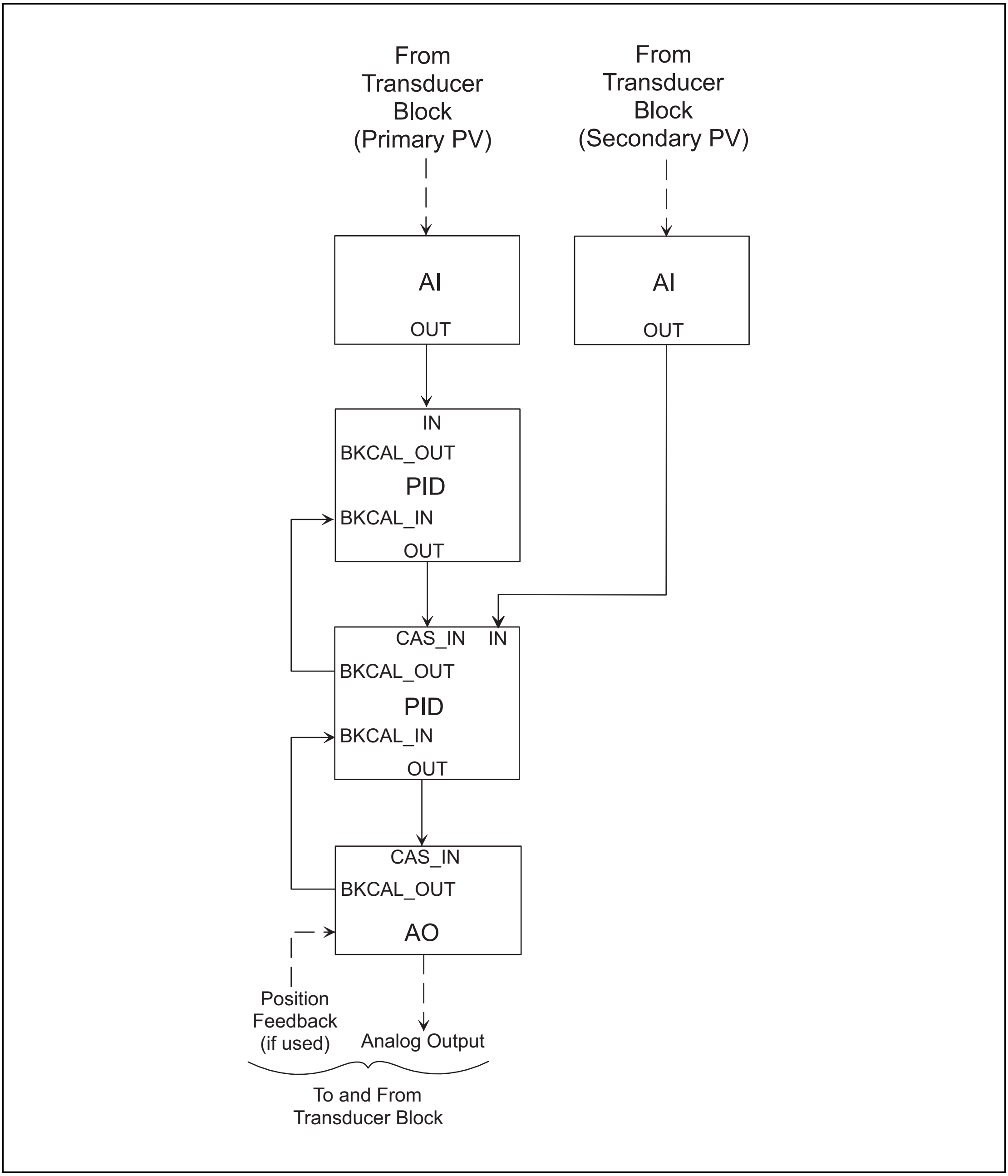


Figure 9-7. Cascade Control Strategy Using FF Function Blocks.

9-2. SAMA Standard PMC 22.1-1981, *Functional Diagramming of Instrument and Control Systems*. Scientific Apparatus Makers Association, 1981 (MCAA web site: http://www.measure.org).

9-3. J. Berge. *Fieldbuses for Process Control: Engineering, Operation and Maintenance,* ISA – The Instrumentation, Systems, and Automation Society, 2002.

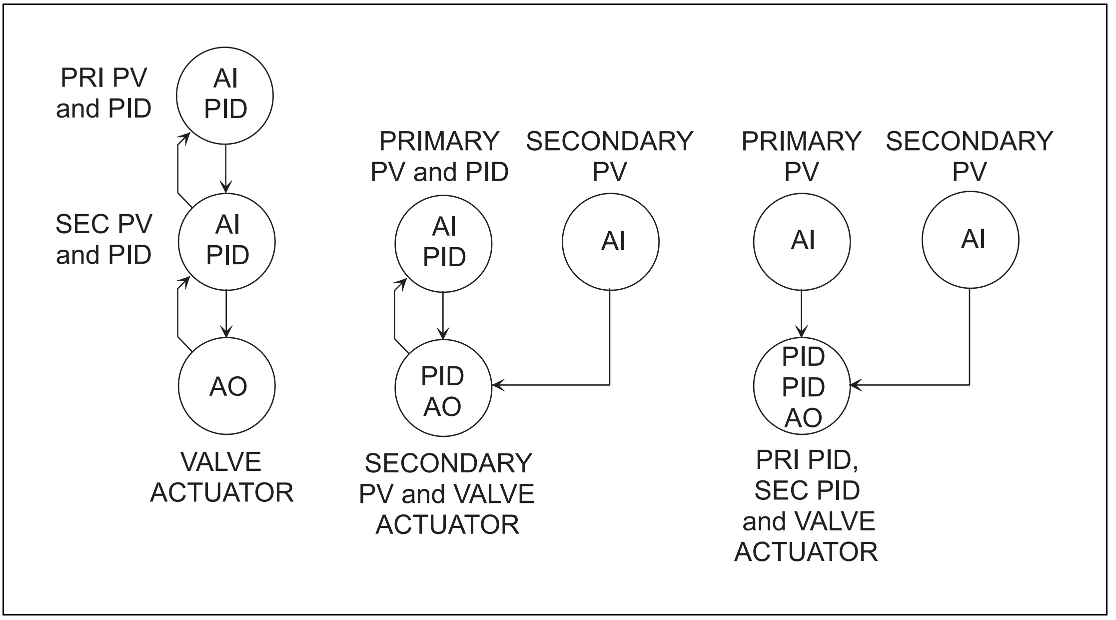


Figure 9-8. Alternative Locations for FF Function Blocks for Cascade Control

9-4. Fieldbus Foundation, *FOUNDATION Fieldbus System Engineering Guidelines*. AG-181, Rev. 1.0.

9-5. Fieldbus Foundation, *Foundation Specification: Function Block Application Process*, *Document FF-891, Part 2.*

10 RATIO CONTROL

❖ RATIO CONTROL TECHNOLOGY

Ratio control paces (controls) the flow rate of one stream so as to maintain a specified ratio between that stream and the measured flow rate of another stream. The measured flow rate is called the "wild" flow, since (at least within the jurisdiction of this control loop) it is uncontrolled. The controlled flow loop is often called the "secondary" loop. Figure 10-1 depicts a ratio control system using both ISA (Ref. 10-1) and SAMA (Ref. 10-2) symbols.

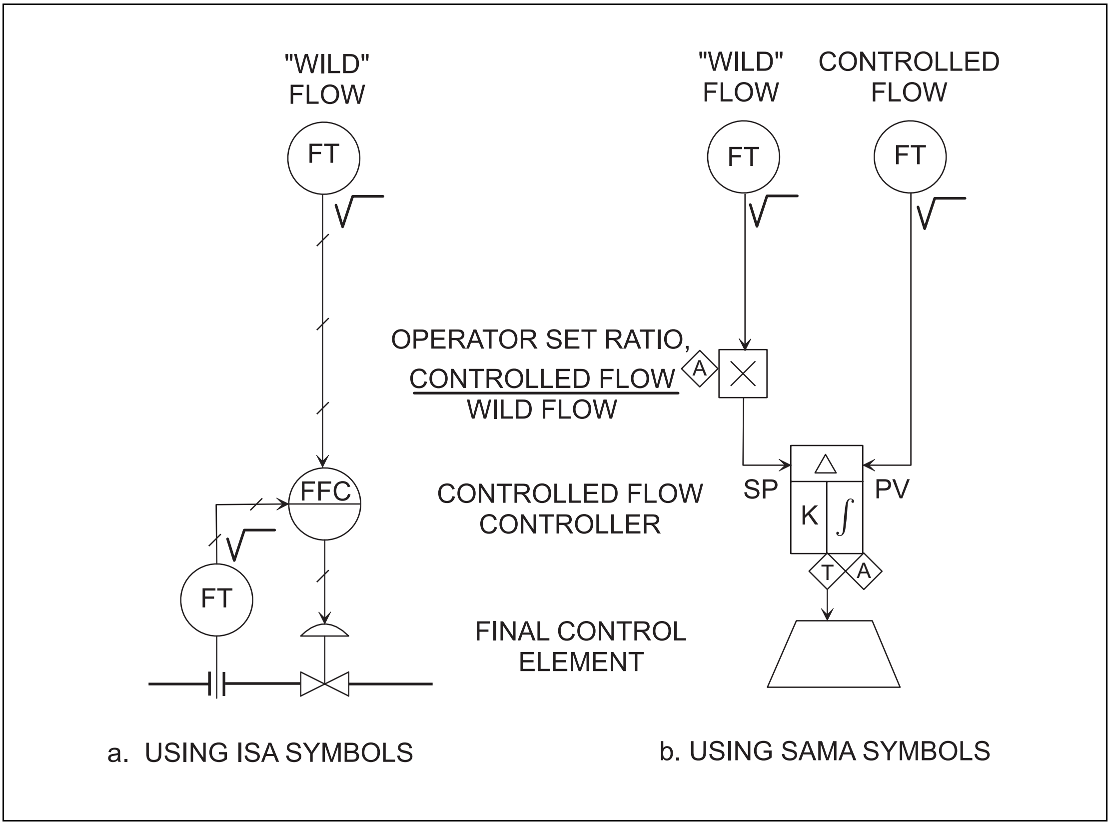


Figure 10-1. Symbolic Representations of a Ratio Control System

Although there may be some applications for ratio control that are not flow-based, most applications ratio flow-to-flow. The broad definition of flow used here is any form of mass or energy transfer. Typical applications for ratio control include:

- Blending two or more components. One ingredient may be set for the master production rate; other ingredients are then ratioed to this master ingredient.

- Air-fuel ratio control for a combustion process. In a simple air-fuel ratio control system, the fuel flow may be controlled by a temperature or pressure controller. A measure of the fuel flow is then used to determine the set point of the air-flow controller.[1]

- Controlling a product stream to feed rate, as a means of composition control. This is a common control technique for distillation towers.

In the most common configuration for ratio control, the measured value of the wild flow is multiplied by the specified ratio, controlled flow-to-wild flow. The product then becomes the set point for the controlled flow loop. For example, fuel flow could be multiplied by a required air-to-fuel ratio, which would result in the required air rate or set point for the air-flow controller.

When ratio control is implemented with conventional analog hardware, the ratio adjustment setting as well as the multiplication function are built into the controller itself. In addition, there is usually a ratio/local switch that, in the local position, permits the set point of the secondary loop to be directly adjusted.

When ratio control is implemented in a microprocessor-based system, the multiplication function may be a part of the PID software function block or it may be a separate function block, depending on the vendor's choice.

Several modifications can be made to the basic ratio control configuration:

- A bias value may be added to (or subtracted from) either the wild or the controlled flow, as shown in Figure 10-2a. With this arrangement, the relationship between the two is offset rather than being a ratio that extends all the way to zero.

- The wild flow rate may be used as the set point for the controlled flow controller, then the controlled flow PV is multiplied by the required ratio, wild-to-controlled flow, as shown in Figure 10-2b. (Notice that this is the inverse of the ratio shown in Figure 10-2a.) Thus, although the controlled-flow PV is scaled in units of the wild flow, it is sensitive to changes in controlled flow. This scheme was frequently used with analog control systems that used flow controllers for both the wild and controlled flows. If the controllers were mounted adjacent to each other on a panelboard, then the set point and PV indicators for both controllers would be in an identical position when the actual ratio was being maintained.

1. In practice, air-fuel ratio control systems are often more complex than described here. An enhanced form of air-fuel ratio control system, called "cross-limiting" or "lead-lag," is discussed in chapter 16.

Occasionally, ratio control is implemented by calculating the actual ratio from the two measured values, as shown in Figure 10-2c. The ratio itself then becomes the process variable of the PID controller. The required ratio is then directly entered as the set point of the PID. This has the advantage that the process variable of the controller as well as the set point are displayed in terms of the desired quantity, which is the ratio between the two flows. It also has the advantage that limits may be placed on the values for entering set points. The disadvantages of this scheme, however, include:

(a) If the flow that represents the denominator of the calculation falls to a low value, the calculated ratio will go to an extremely large value;

(b) The gain of the ratio controller is inversely proportional to the controlled flow rate, even if the flow-measurement signal is linear or has been linearized.

One distributed control system (DCS) manufacturer provides a ratio control algorithm in which the process variable and set point displayed on the CRT screen are the calculated and desired ratios, respectively. However, the actual PID control algorithm uses as its process variable the actual measurement of the secondary flow. The set point used by the PID is the product of the entered (and displayed) ratio as well as the measured value of the wild flow. This configuration is shown by Figure 10-2d. It provides the display advantages of the previous scheme without encountering the disadvantage of the nonlinearity. This scheme can probably be configured in most DCSs by using multiple function blocks.

❖ AUTOMATIC RATIO ADJUSTMENT

Ratio control using a constant value of the required ratio is no more technically challenging than an ordinary flow controller. However, a more interesting scheme with broader application is when the required ratio is adjusted automatically by the output of another feedback controller. We will use a combustion process, specifically an air-fuel ratio control system, to illustrate this control scheme.

Figure 10-3 shows a P&I diagram of a process heater in which the fuel flow is measured and multiplied by the required air-to-fuel ratio. This results in the required air-flow rate, which is introduced as the set point of a feedback controller. The new feature of this diagram is that the required air-to-fuel ratio is automatically adjusted by the output of a stack O_2 controller. For simplicity, the same control scheme, without the process equipment, is shown in Figure 10-4. Additional components of the heater control system are shown with lighter lines. This includes an outlet heater temperature measurement and controller that sets the fuel-flow rate. These components do not lie within the scope of our present discussion, however.

The ultimate purpose of air-to-fuel ratio control is energy efficiency. For a given amount of fuel, a theoretical amount of air is needed if there were complete combustion of all the fuel. This is called the stoichiometric ratio of air and fuel. In practice, however, a greater amount of air is required to provide good mixing of the air and fuel. If there is insufficient air, efficiency will drop because of the loss of unburned fuel. On the other hand, if there is too much air, effi-

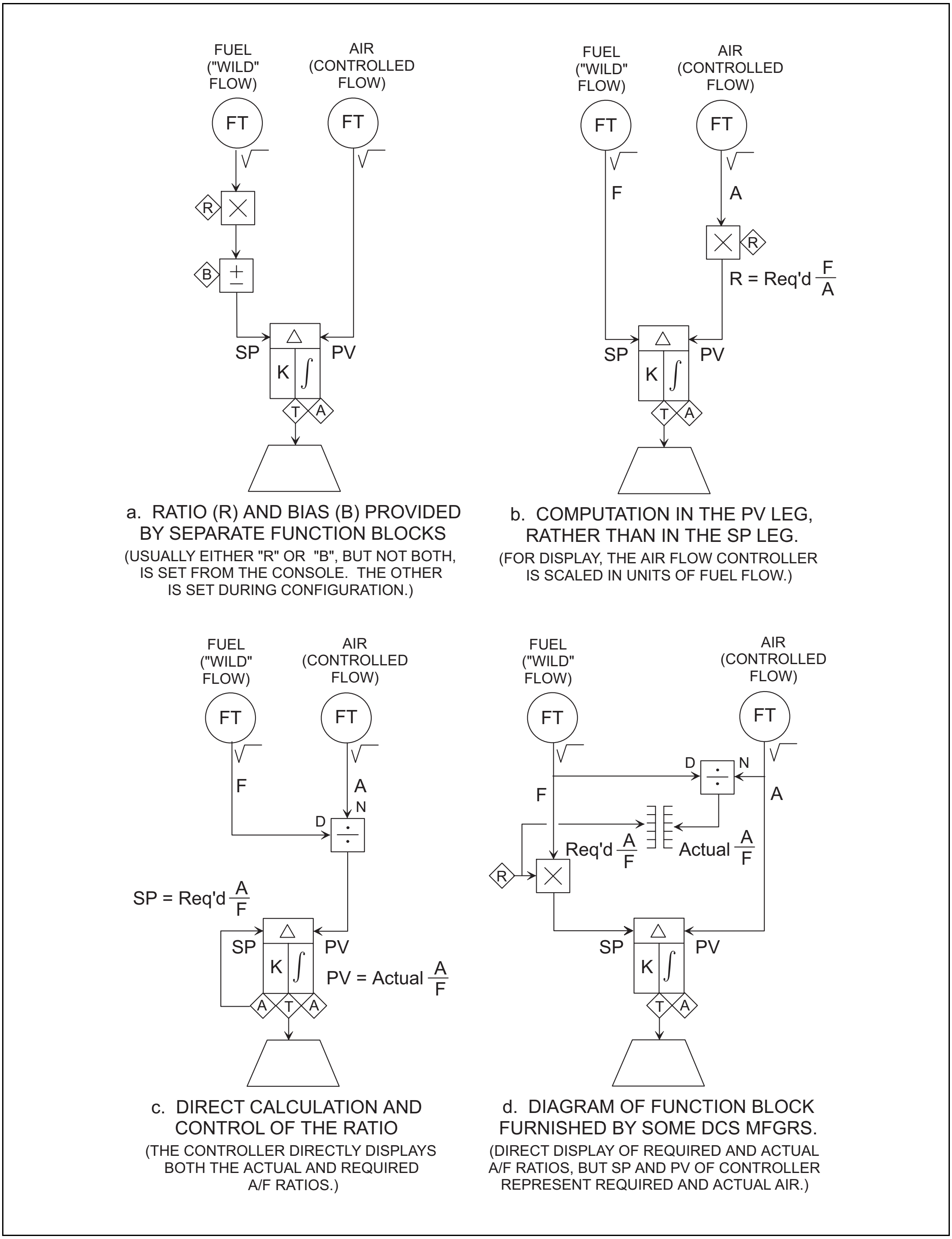


Figure 10-2. Alternative Ratio Control Configurations

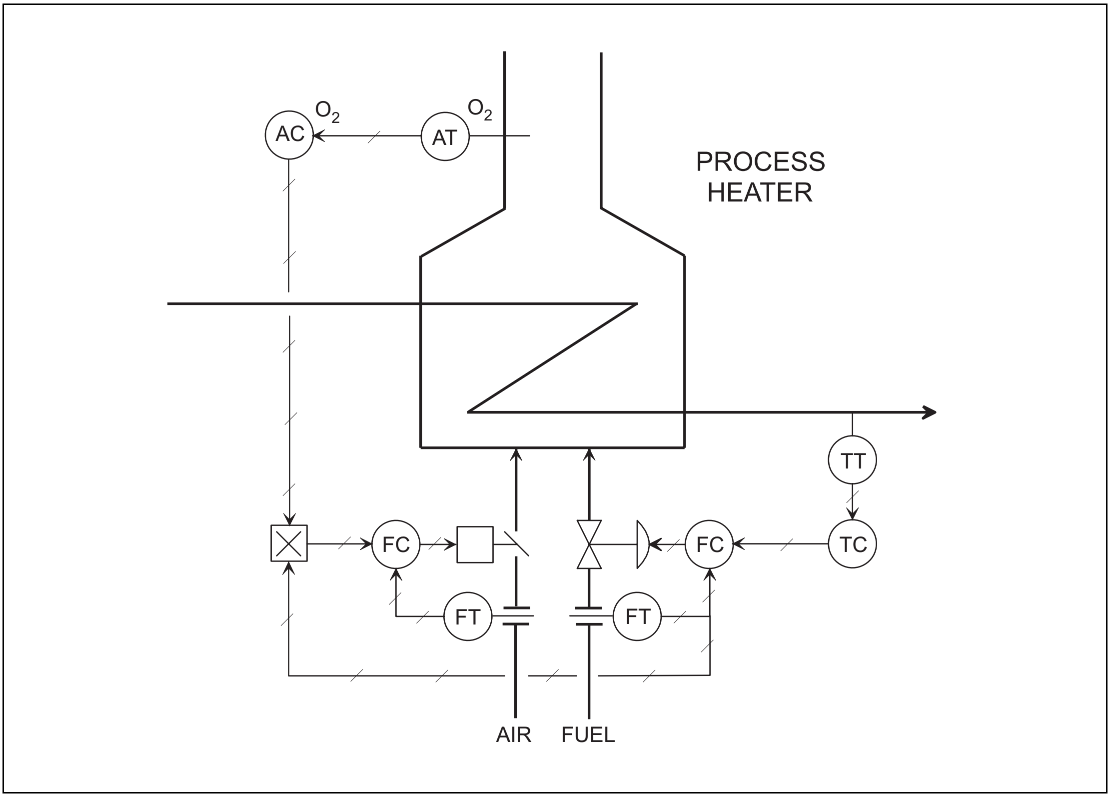


Figure 10-3. P&I Diagram of Automatic Set Ratio Control

ciency will drop because energy is carried away by the inert stack gasses (primarily nitrogen). Maximum combustion efficiency usually occurs at approximately 10 percent excess air, or 2 percent excess oxygen in the stack. Thus, the O_2 composition in the stack can be monitored and controlled by automatically adjusting the air-to-fuel ratio.

This discussion is not intended to be a complete treatment of combustion control, but as an illustration of an application of ratio control. A complete combustion control system might also monitor and control other stack constituents, such as %CO, total combustibles, or opacity. Also, the $\%O_2$ controller set point might not be constant, especially for steam-generating systems. Rather, the set point would be increased at lower boiler loads to assure that fuel and air were adequately mixed within the combustion zone. A complete discussion of combustion control is beyond the scope of this text.

In the control system depicted by Figure 10-4, the air damper will be of the fail-open type; hence, the air-flow controller will be direct-acting. Alternatively, with some digital processor–based systems, the air-flow controller will be reverse-acting, and the analog-output function block will be set with the “reverse output” feature. Since a decrease in the required air-to-fuel ratio will cause a decrease in the measured O_2, the O_2 controller must be reverse-acting. The ultimate effect of a drop in measured O_2 is that the air damper will open.

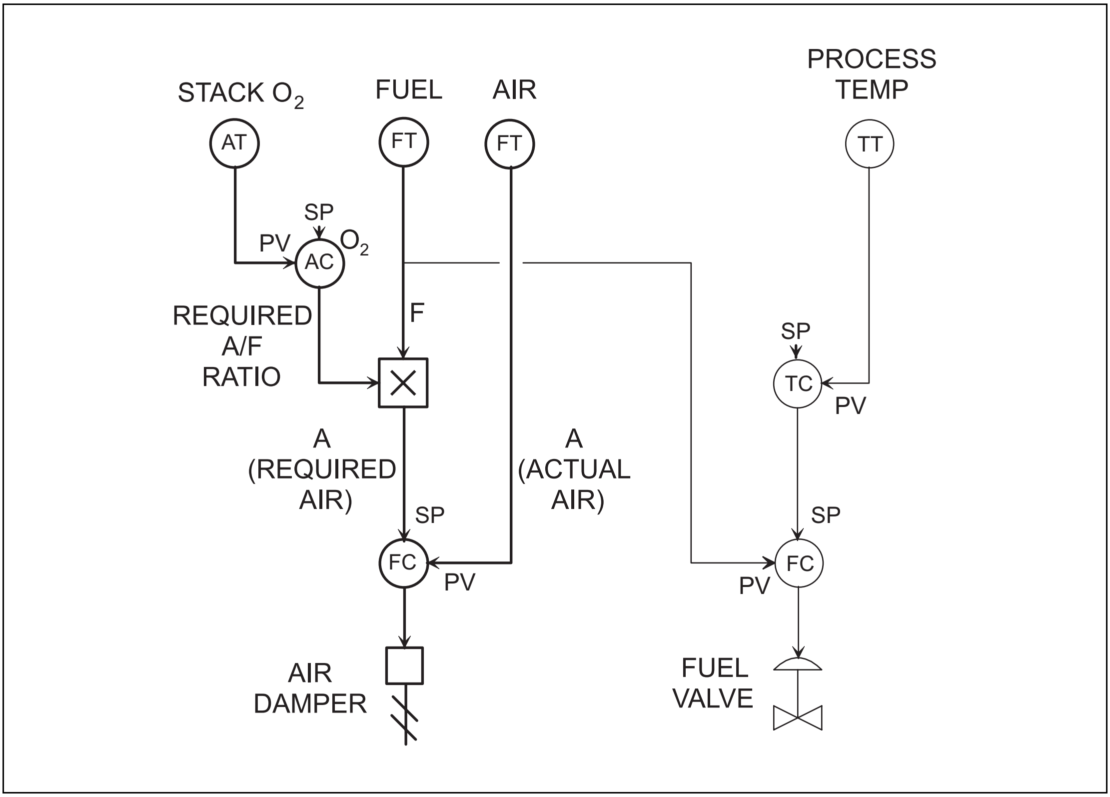


Figure 10-4. Example of Automatically Set Ratio Control

Consider the control scheme of Figure 10-5. This is obviously a simpler control system than Figure 10-4. Furthermore, it possesses all the attributes of a feedback control system, since a drop in measured O_2 will cause the air damper to open. Reflect for a moment on the relative merits of the schemes depicted by Figures 10-4 and 10-5.

Despite its simplicity, the primary deficiency of the scheme of Figure 10-5 is that there must be an error in the feedback loop in order to have a change in the controller output (i.e., in order to change the damper position). If the outlet temperature controller demands more fuel, the additional fuel will act as a load increase on the O_2 control system. This will cause measured O_2 to drop. This is the feedback penalty that must be paid with this control system to change the air flow.

In the control scheme of Figure 10-4, when there is an increase in fuel demand, the air and fuel will rise together if the output of the O_2 controller remains constant. At a constant air-to-fuel ratio, the O_2 in the stack can be expected to remain relatively stable; thus, there will be no need for a change in the controller output. In essence, we have obtained the control action (change in air flow) without having to pay the penalty of an error in the feedback loop.

We shall see later that ratio control is often merely one form of feedforward control.

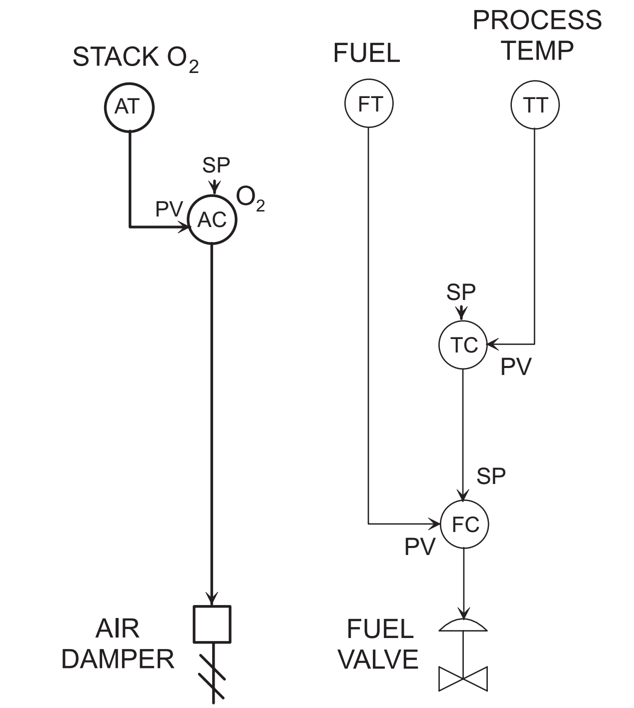


Figure 10-5. Feedback Control of Stack O_2 Requires Paying Feedback Penalty

❖ SCALING THE RATIO CONTROL COMPONENTS

For direct operator-set ratio control systems, the entered value is usually indexed to the ratio actually required, rather than being the required ratio itself. For example, suppose a design ratio is to be maintained between wild flow stream “A” and controlled flow stream “B.” If the flow transmitters are calibrated so the ratio of their ranges is the same as the design ratio, then the incoming signal from the wild flow transmitter requires only a multiplication by 1.0 to give the required set point to the controlled flow controller. The ratio adjustment dial may be calibrated from 0 to 2, or the keypad entry may have software limits of 0 and 2, with “1.0” being the nominal entry. Of course, it may be desirable to limit the entry to a narrower range, for instance, 0.8 to 1.2. Then the operator could not inadvertently enter a value that is far from acceptable.

If the output of a primary controller sets the required ratio automatically, then a simple scaling or computation must be performed so that the 0 – 100 percent output of the controller represents the acceptable range of ratio entry value. When the controller’s output is 50 percent, the required ratio will be the design ratio if the flow transmitters are calibrated so the ratio of their ranges is the same as the design ratio and if the output of the primary controller is multiplied by two. Controller output excursions above or below 50 percent adjust the required ratio to greater or lesser values than the design ratio. The ratio represented by the extremes, 0 percent and 100 percent, must be determined by scaling.

Appendix A presents a general methodology for determining scaling parameters. It also presents an example of an air-to-fuel ratio control system in which the nominal ratio is 11:1. The objective is to limit the controller's authority to the range of 9:1 to 13:1, or ±2 ratio units on either side of the nominal value. For transmitter ranges of 0 to 15,000 scfm (air) and 0 to 1000 scfm (fuel gas), the following required scaling equation is derived:

$$a = 0.267\ (r + 2.25)\, f$$

$$= (0.267\ r + 0.6)\, f$$

where a, f, and r represent *normalized* values (range: 0 – 1) of air, fuel, and the required ratio. Hence, the ratio control system could be implemented as shown in Figure 10-6.

Note that various manufacturers will provide different means to achieve the required computation. Also note that it is highly desirable that an initialization signal be passed back through the computation elements to the primary controller whenever the secondary controller is not fully automatic cascade. This procedure should take the set point or PV of the secondary, invert the computation equations, and set the primary controller output so as to ensure bumpless transfer back to the automatic or cascade mode of the secondary.

❖ RATIO CONTROL USING FOUNDATION™ FIELDBUS FUNCTION BLOCKS

One of the basic function blocks defined by the Fieldbus Foundation document FF-891, Part 2 (Ref. 10-3) is a ratio control block. As a control class block, it supports the back-calculation/ initialization feature. It can be used for ratio control set directly by the operator, as shown in Figure 10-7. Here, the actual ratio would be back-calculated and the ratio block initialized any time the PID block is in the manual mode. If the operator-set ratio is not to be initialized, then the back-calculation mechanism is not required, and a calculate-class block is used in place of the ratio block. Several manufacturers provide this type of block, rather than a ratio block.

Since the ratio block (as defined by Fieldbus Foundation) supports the back-calculation mechanism, it can be used with automatic-set ratio control systems, as shown in Figure 10-8. By setting the scale factors for the primary controller output, the automatically set ratio may be limited in its excursion above and below the nominal value. This accomplishes the purpose of the calculation shown in Figure 10-6. Note that although the ratio block is defined in the FF standard, it may not be supported by all manufacturers.

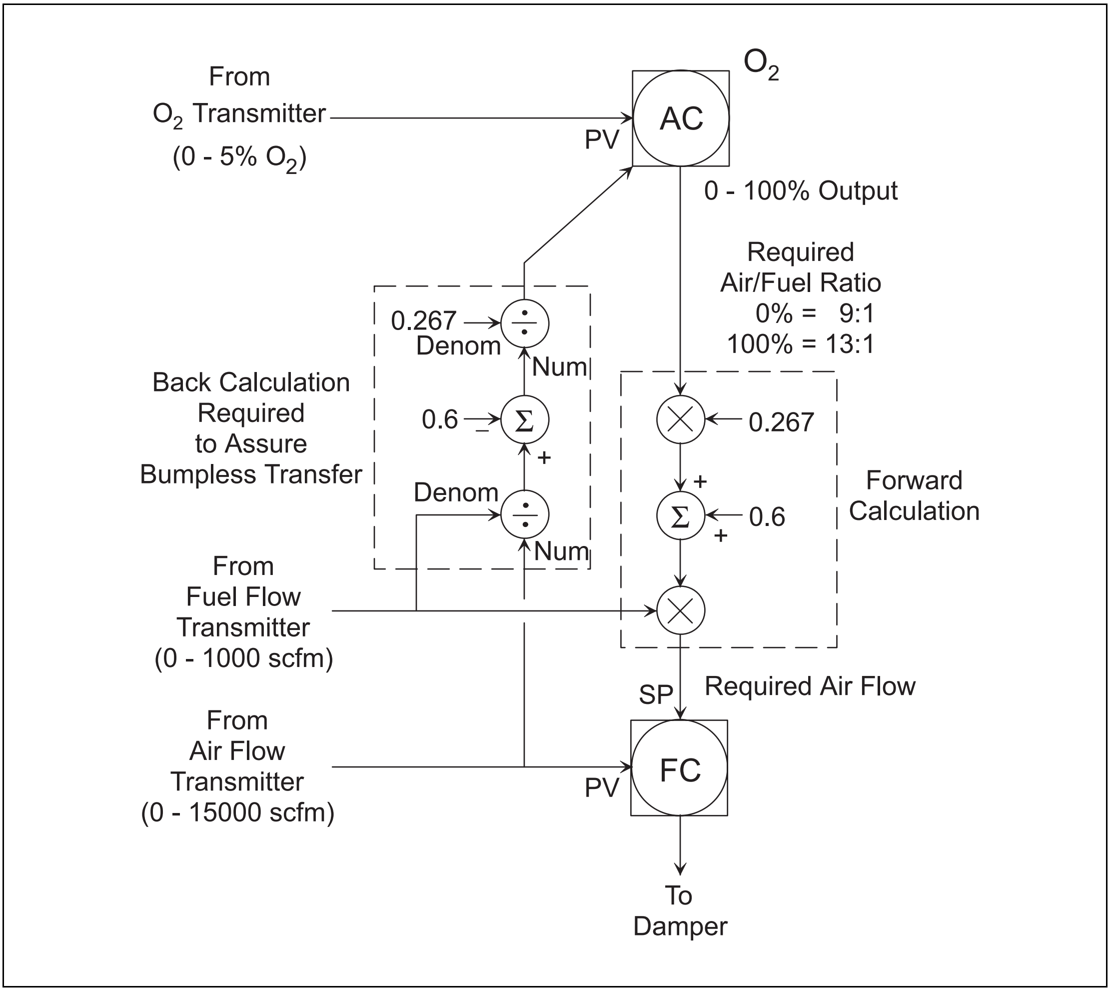


Figure 10-6. Example of Forward and Back Calculation Required for Ratio Control

❖ REFERENCES

10-1. ISA-5.1-1984 (R 1992), *Instrumentation Symbols and Identification*. ISA – The Instrumentation, Systems, and Automation Society, 1992.

10-2. SAMA Standard PMC 22.1-1981, *Functional Diagramming of Instrument and Control Systems*. Scientific Apparatus Makers Association, 1981 (MCAA web site: http://www.measure.org).

10-3. Fieldbus Foundation, *Foundation Specification: Function Block Application Process, Document FF-891, Part 2.*

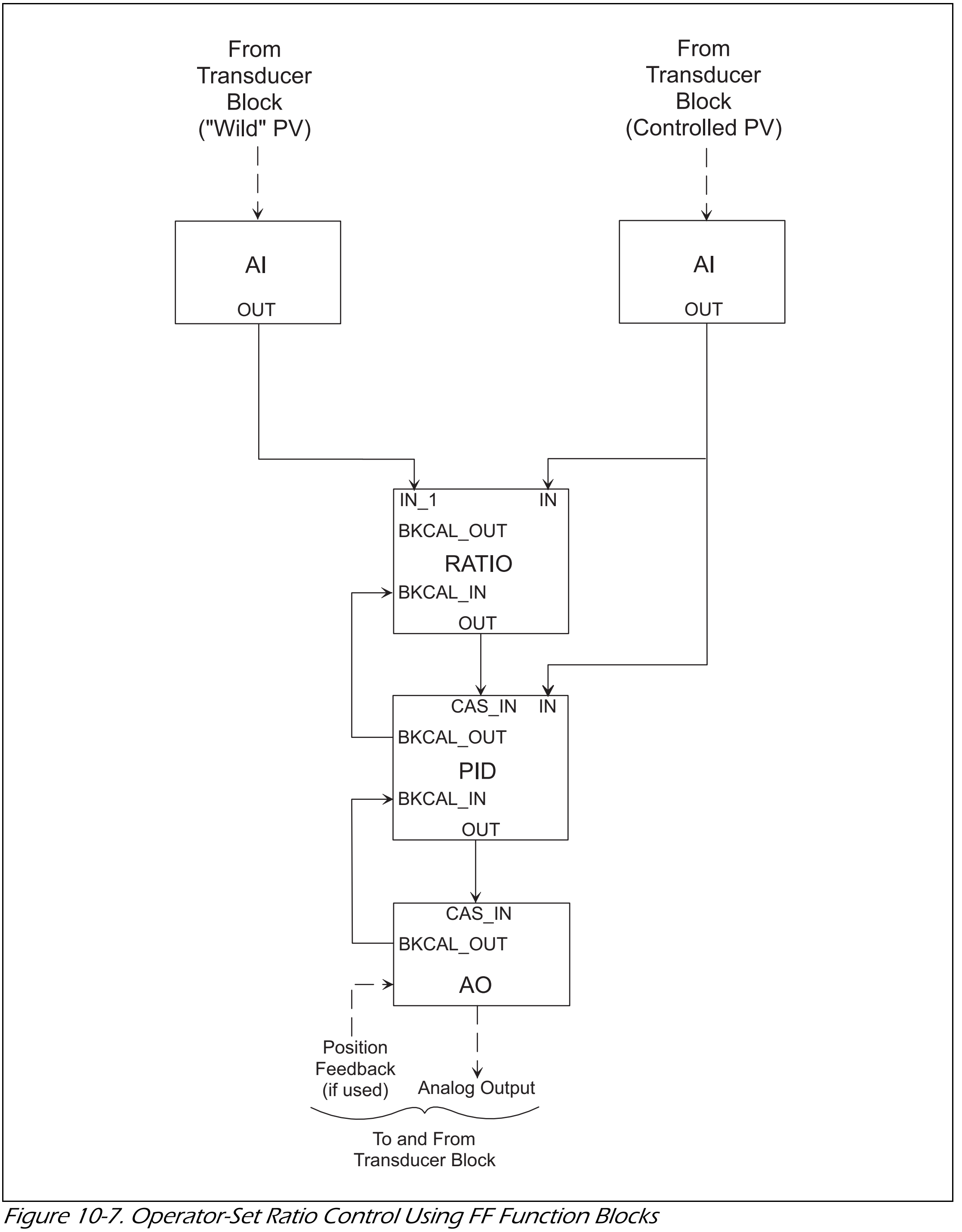


Figure 10-7. Operator-Set Ratio Control Using FF Function Blocks
(Note: If the ratio does not need to be back-calculated, a calculate-class function block can be used in place of the ratio block.)

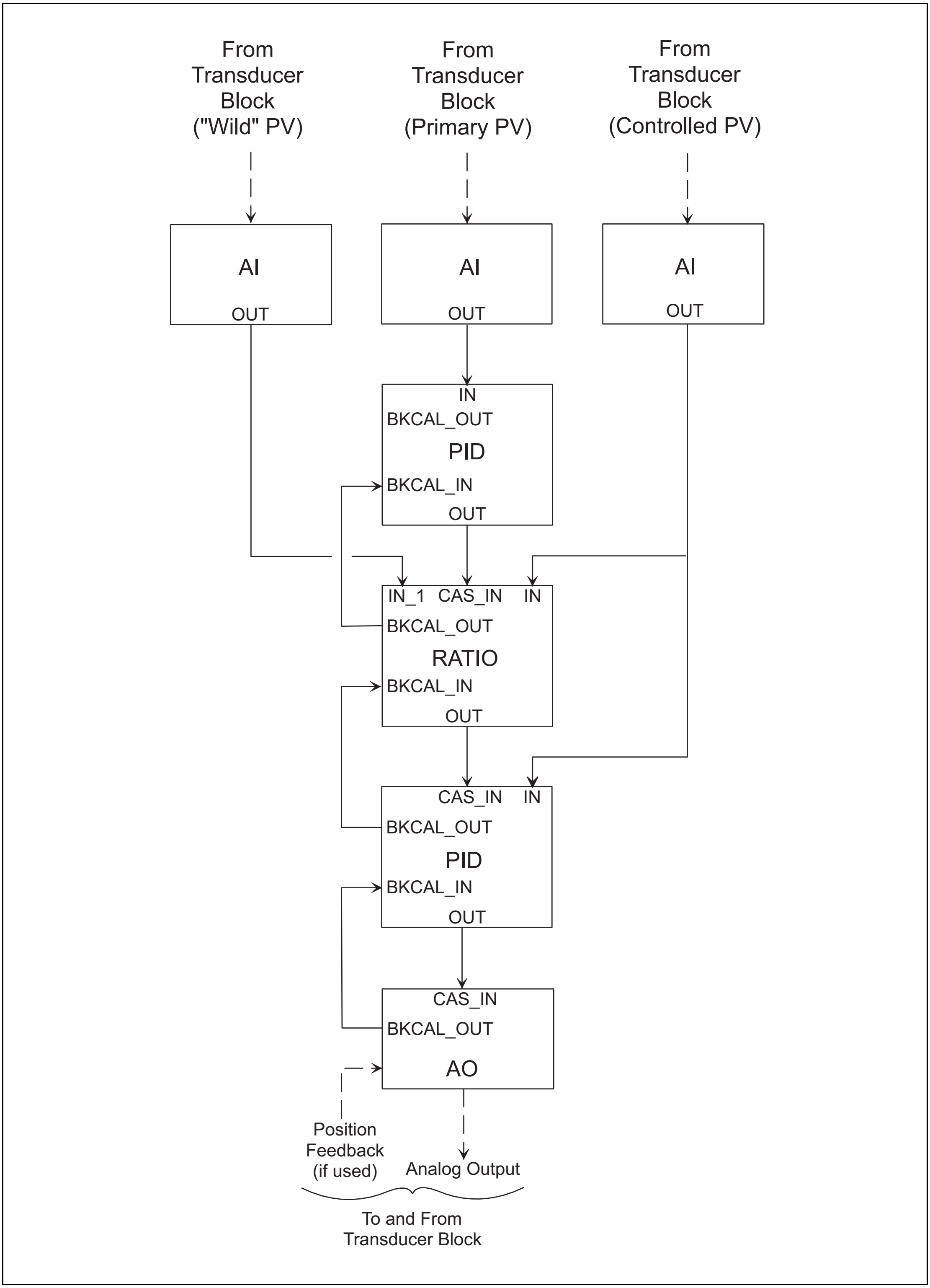


Figure 10-8. Automatic-Set Ratio Control Using FF Function Blocks

11

FEEDFORWARD CONTROL

In my home, the thermostat—the controller for the home heating system—is always a focus of discussion. Either it doesn't match the wallpaper, which detracts from the decor of the room, or someone wants to place a lamp next to it, which causes the thermostat to give a false temperature reading, which in turn causes the room to be too cold.

In the system shown in Figure 11-1a, the thermostat is an on/off device that controls a fuel supply solenoid valve. Suppose the following plan of action is being considered to get rid of the existing thermostat, thereby eliminating a point of discussion in the household:

- The primary cause for variations in the heat delivered to the room is determined to be changes in the outdoor air temperature.
- A suitable location is found for an outdoor air temperature sensor.
- The present on/off valve is replaced with a throttling-type valve that can be positioned anywhere between 0 and 100 percent. (The technological details of this are beyond the scope of this book and are not essential to an understanding of the point being discussed.)
- After considering the heat-transfer characteristics of the home's insulation, a program for required valve position versus outdoor air temperature is determined.

The system is implemented as shown in Figure 11-1b. The thermostat inside the house is eliminated. To control the fuel valve, the control system depends entirely on a measurement of the outdoor air temperature.

If this control scheme were implemented and if all the assumptions were perfect, then we could exactly compensate for the disturbance (changes in outdoor air temperature) and maintain the home temperature at exactly the desired value. The indoor thermostat would not be required.

Certain problems can be foreseen, however. First, the outdoor air temperature may not be the only disturbance. There may be sources of heat within the house, such as cooking and laundry. Or a window or a door may be open. Furthermore, the program determined for the required valve position versus outdoor temperature may not be exactly correct, or it may change with

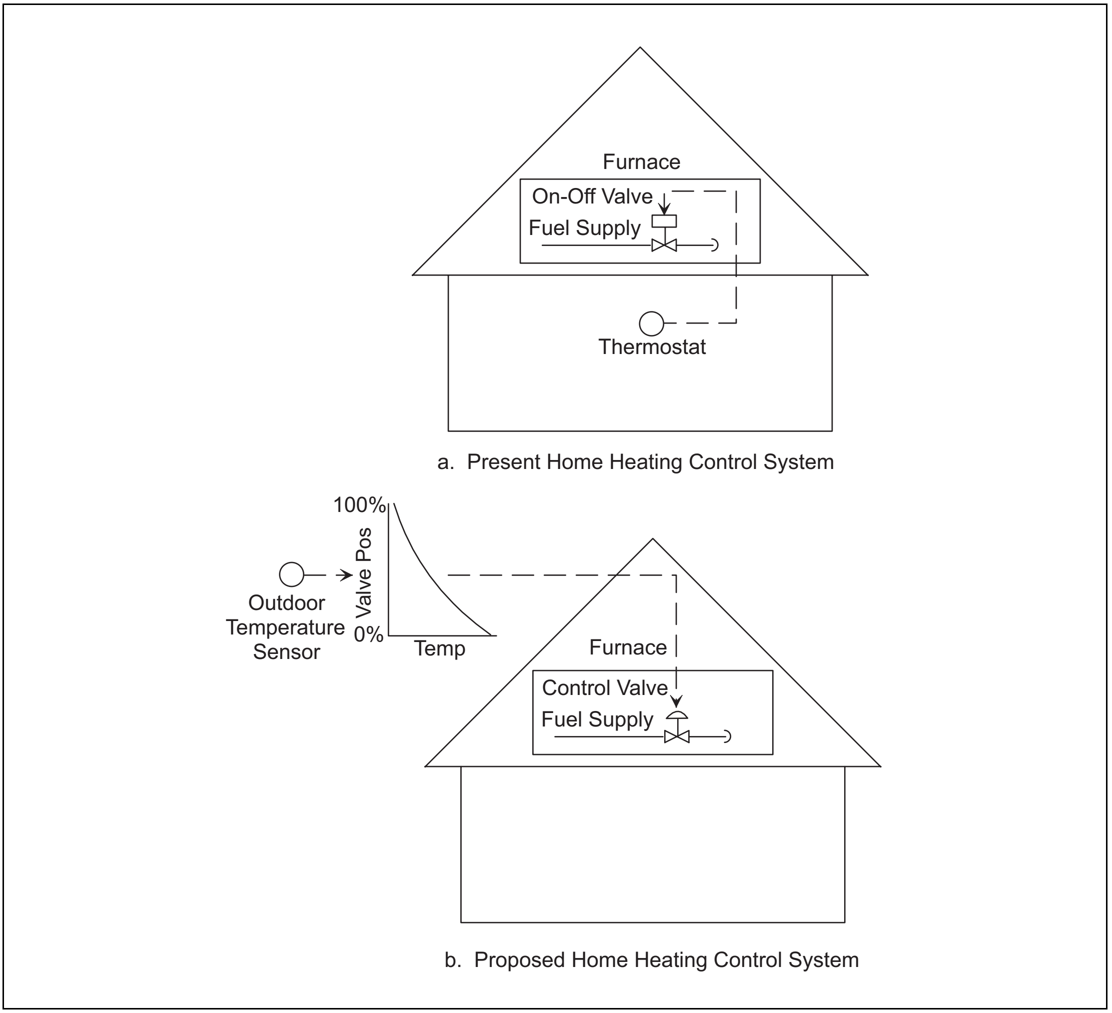


Figure 11-1. Present and Proposed Home Heating Control System

time if the insulation deteriorates. And finally, if there were a sudden drop in outdoor air temperature, the control scheme would immediately increase the furnace's heat delivery, even though significant time would probably pass before the effects of the temperature change were felt inside the house. Thus, in the short term, the house would be overheated.

This trivial illustration is an example of *feedforward* control. With feedforward control, the objective is to drive the controlling device based on a measurement of the disturbance that affects the process rather than on the process variable itself. The controlling device may be a final control element such as a valve. In industrial applications, however, it is preferable, however, to adjust the set point of a lower-level flow control loop.

Several obvious implications can be drawn from this statement of objectives:

- The significant disturbances to the process must be known and measurable.

- The proper corrective action to compensate for the disturbance must be known, both in magnitude and on the correct time schedule.

If our knowledge were perfect about these areas, however, we could formulate a *perfect* feedforward controller. Then, whenever the controller sensed a disturbance, it would take precisely the right control action, at precisely the right time, to compensate for the disturbance. With perfect compensation, the process variable would not deviate at all. We would thereby achieve the goal of driving the final control element without having to pay a feedback penalty.

Idealistic? Perhaps. For real applications, our knowledge of the process, the disturbances, and the correct compensating control action will always be less than perfect, hence our feedforward controller will be less than perfect. For this reason, feedback and feedforward control are usually combined.

If we are eventually going to use feedback control anyway, what is the purpose of using feedforward? Could we not simply implement the control system using only feedback control?

The following argument will answer this question. Admittedly, the argument is qualitative rather than quantitative, but the concept underlying this argument is crucial for understanding and using feedforward control.

Suppose the feedforward controller is only 75 percent perfect. Even so, it will still make 75 percent of the required control action, leaving only the remaining 25 percent as the responsibility of the feedback controller. Since the feedback controller has to do only 25 percent of the work, the feedback penalty will be only 25 percent of what it would have been using feedback control alone.

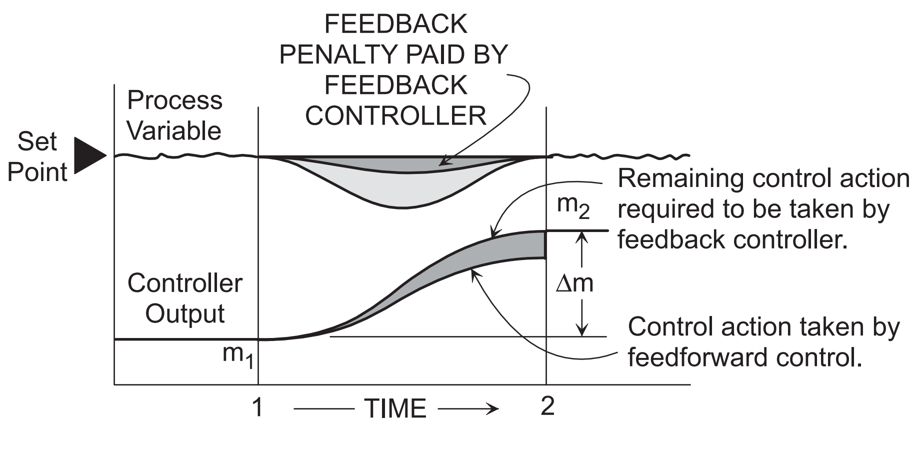


Figure 11-2. Reduction in Feedback Penalty by Feedforward Control

One remaining issue deserves attention before we begin a detailed consideration of the design of feedforward control systems. We can purchase a standard feedback controller from any number of manufacturers, and we can choose a standard feedback controller from a manufacturer's library of control algorithms. But we cannot purchase, or choose, a standard feedforward controller, even though a manufacturer may furnish a rich variety of tools to assist in implementing feedforward control. In other words, each feedforward control system must be designed for a particular application, then implemented using tools (e.g., software function blocks and configuration options) provided by the chosen manufacturer. For this reason, it is best to use examples when providing instruction in the design of feedforward control systems.

❖ DESIGNING FEEDFORWARD CONTROL SYSTEMS

We will use the liquid-phase process heater shown in Figure 11-3 as a vehicle for illustrating the design procedure for feedforward control. Suppose this heater is currently controlled by an outlet temperature controller that is cascaded to a fuel flow controller. With any disturbance to the control loop, an error in the feedback loop must occur in order for the temperature controller to change the set point of the flow controller.

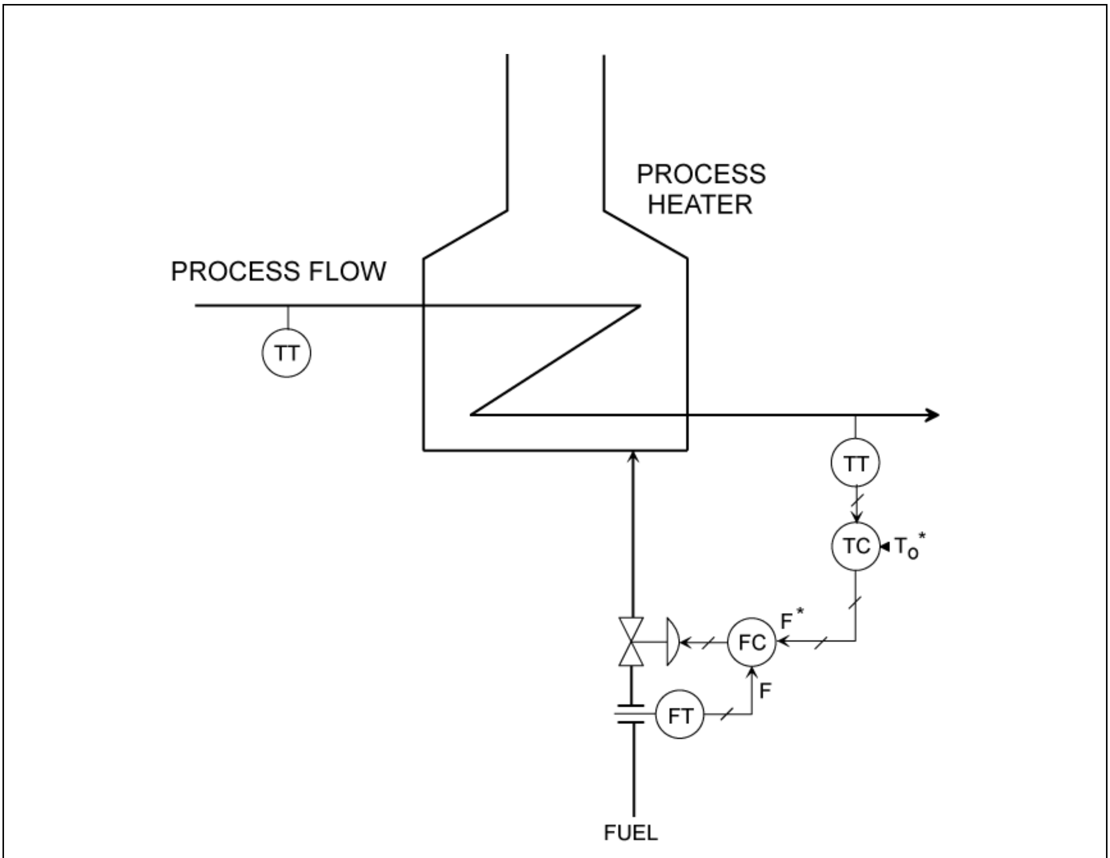


Figure 11-3. Example for Application of Feedforward Control

Suppose we have analyzed the application and have determined that variations in process inlet temperature are the principal disturbances. Hence, the fuel flow controller should be driven by a feedforward controller that senses inlet temperature rather than by an outlet temperature controller. A process model can be formulated, based on a heat balance:

$$F_p \times c_p \times (T_o - T_i) = F \times H_v \times E_{ff} \tag{11-1}$$

where: F_p = Process flow rate
c_p = Process fluid–specific heat
T_o = Outlet temperature
T_i = Inlet temperature
F = Fuel flow rate
H_v = Fuel heating value
E_{ff} = Heater thermal efficiency

Now change the symbol for outlet temperature, T_o, to the symbol T_o*, which represents the *required* outlet temperature. Then solve for the required fuel rate, F*:

$$F^* = \frac{F_p \times c_p \times (T_o^* - T_i)}{H_v \times E_{ff}} \tag{11-2}$$

On the right-hand side of Equation 11-2, T_i is the only variable, by virtue of the fact that the application study showed that inlet temperature was the only disturbance. The rest of the terms are constant. Hence, we can measure the inlet temperature and implement Equation 11-2, using either analog hardware devices or software function blocks, as shown in Figure 11-4.

If the assumption is truly correct that the inlet temperature is the only disturbance and if we need not be concerned with dynamics, this feedforward controller will maintain the correct outlet temperature, T_o*, without T_o having actually been measured and without the use of feedback.

Suppose, however, that we subsequently learn that there are other disturbances. For instance, there may be variations in process flow rate or there may be variations in fuel heating value. (Some processes produce an off-gas that can be burned as fuel. Typically the heating value of this fuel is quite variable.) Using the same feedforward equation, 11-2, these additional disturbances can be measured and incorporated into the control system as shown in Figure 11-5.[1]

Thus, multiple disturbances can be incorporated into a feedforward control scheme, provided they are (1) measurable and (2) corrective action is known.

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1. The measurement of fuel heating value is a subject unto itself. A composition analyzer could be used for this purpose. For a gaseous fuel, a gas-gravity analyzer provides a reasonable indication of fuel heating value, provided there is very little H_2 in the mixture. A Wobbe index analyzer, properly applied, can also be used. It probably would not be configured exactly as shown in Figure 11-5, however, since the product of the Wobbe index and the differential pressure across the fuel measurement orifice plate is itself the heat release rate.

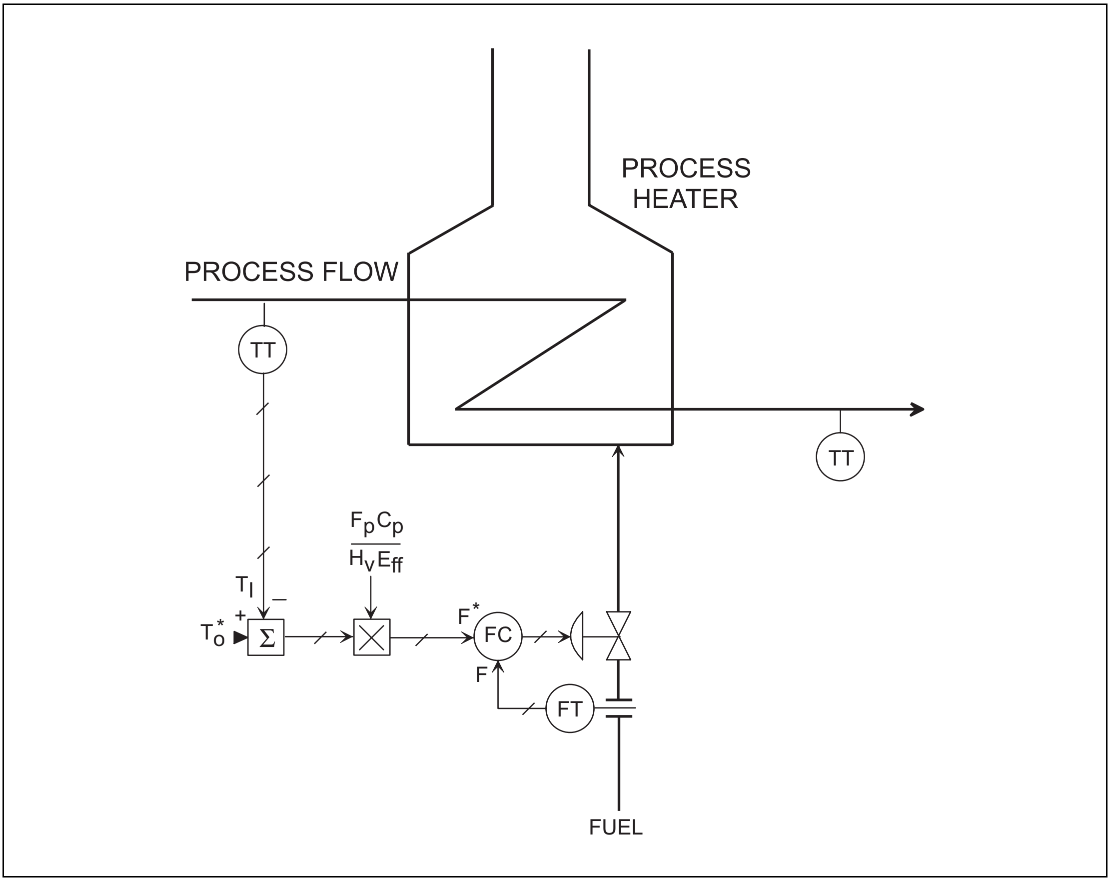

Figure 11-4. Feedforward Control from a Single Disturbance

Are there yet additional disturbances that don't meet these two criteria? Perhaps. The heater efficiency can change. The feedstock can change, causing specific heat to change. There may even be something as mundane as a calibration drift in one of the measuring devices. All of these are unmeasurable disturbances that, in effect, represent an error in the feedforward process model. Although we may have compensated for the most significant disturbances by using feedforward control, we must add feedback to our control scheme to correct for the unmeasurable disturbances.

There are several different methods for combining feedback and feedforward control; these are discussed in the next session, along with some of the advantages and disadvantages of each.

♦ Additive Feedback

The simplest method of combining feedback and feedforward signals is to add the two together, as shown in Figure 11-6. If the feedback controller uses a velocity-mode digital algorithm (see chapter 5 for a discussion of position-mode and velocity-mode digital algorithms), then on each calculation cycle it will calculate a change in output, Δm, which is added to the output of the feedforward controller. Since Δm can be positive, negative, or zero, the feedback

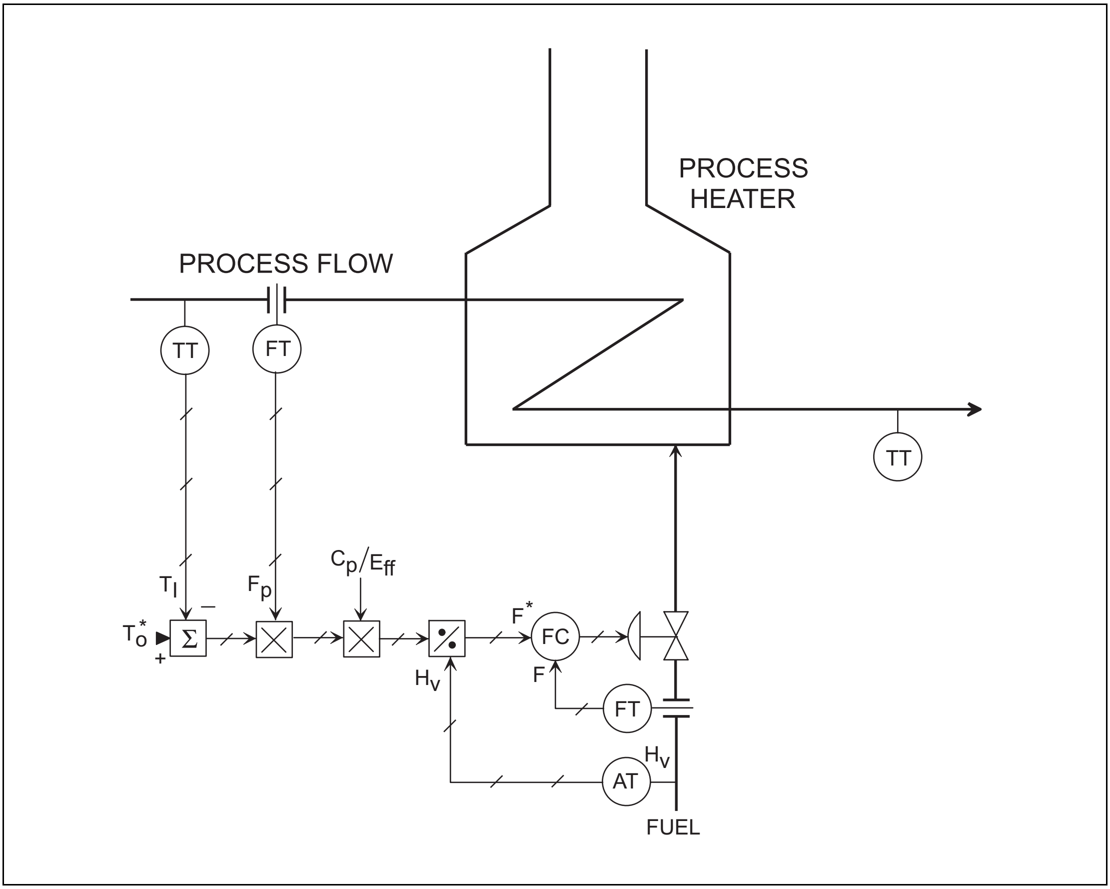

Figure 11-5. Feedforward Control from Multiple Disturbances.

controller can either increase, decrease, or make no change in the output of the feedforward controller. If the feedback controller is a position-mode algorithm, it should be scaled so it represents zero correction to the feedforward signal when it is in the center of its range.

> *(For example, if the control system is being implemented with pneumatic devices, a three-input add-subtract relay could be used to combine the feedback and feedforward signals. The third input would be a constant 9 psi signal, which would be subtracted from the other two. In this way, when the feedback controller output is 9 psi, it is canceled by the constant 9 psi subtraction. If the outlet temperature is too low, the temperature controller output would shift to a value higher than 9 psi, thus adding a positive correction to the value calculated by the feedforward controller. A high outlet temperature would have the opposite effect.)*

Additive feedback, also called "feedback trim," is widely used because of its simplicity. This method may cause problems, however, since the process gain of the feedback loop may vary inversely with process flow rate. If this is so, then the feedback controller may require retuning

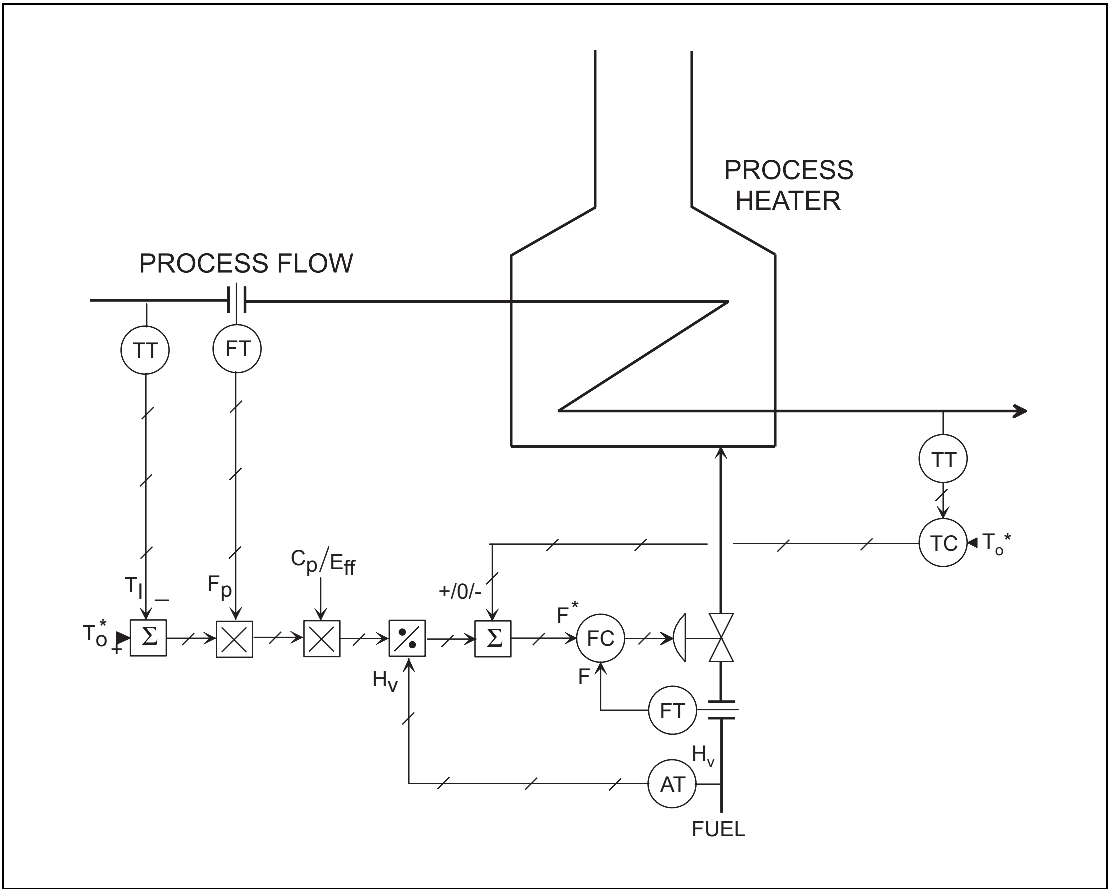

Figure 11-6. Feedforward Control with Additive Feedback Trim

for different process flows. The following discussion illustrates how the process gain of the feedback loop can vary.

The process gain of any feedback loop is given by the following:

$$K_p = \frac{\Delta \text{Measurement}}{\Delta \text{Controller Output}}$$

In the application being illustrated here, the process gain of the feedback loop is given by the following:

$$K_p = \frac{\Delta \text{Temperature}}{\Delta \text{Fuel}}$$

K_p can be calculated by first rearranging Equation 11-1:

$$T_o = T_i + \frac{H_v \times E_{ff}}{c_p} \times \frac{F}{F_p} \qquad (11\text{-}3)$$

$$\text{Now: } \frac{\Delta\,\text{Temperature}}{\Delta\,\text{Fuel}} = \frac{\partial T_o}{\partial F}$$

$$\text{so that } K_p = \frac{H_v \times E_{ff}}{c_p} \times \frac{1}{F_p} \qquad (11\text{-}4)$$

Equation 11-4 shows that with additive feedback trim the process gain (often) varies inversely with the process flow rate. Intuitively, if at 100 percent process flow rate an incremental change in fuel flow has a certain effect on the outlet temperature, then at 50 percent process flow rate, the same incremental change in fuel flow will have roughly twice the effect on outlet temperature.

Whether or not this is a problem to reckon with depends on the circumstances of the particular application. If the process always operates with a very nearly constant process flow rate, then it will not be a problem. If there are wide variations in process flow rate, then the feedback controller will need to be retuned for different load conditions. Even here, if the feedback controller is equipped with a gain-scheduling control algorithm, the problem can be solved simply making the controller gain proportional to the inverse of the process flow rate. In the absence of this special provision, however, the variation in process gain may be a problem. The next method of combining feedback and feedforward control, multiplicative feedback, avoids the problem altogether.

♦ Multiplicative Feedback

In Figure 11-7, the feedback and feedforward signals are multiplied together. The feedback signal can be considered as a multiplying factor, K, which can be less than, equal to, or greater than 1.0. If the feedforward controller is exact, then no correction is necessary. The feedback controller output will then come to equilibrium at a value representing 1.0. If the outlet temperature is low, the temperature controller output will shift to a value greater than 1.0; if it is high, to a value less than 1.0. As a practical matter, the temperature controller output should be scaled so that 0 to 100 percent of signal range represents a limited range of correction, say K = 0.75 to 1.25. Then the feedback controller can correct the computed feedforward results by a factor of +/- 25 percent.

To illustrate the rationale for multiplying the feedback and feedforward signals, suppose that process flow rate is the only disturbance. After all, variations in process flow were the offending element in the additive feedback trim scheme we discussed previously. Then Figure 11-8 will apply. Here the feedback controller output could be scaled to be a corrective value, K, as

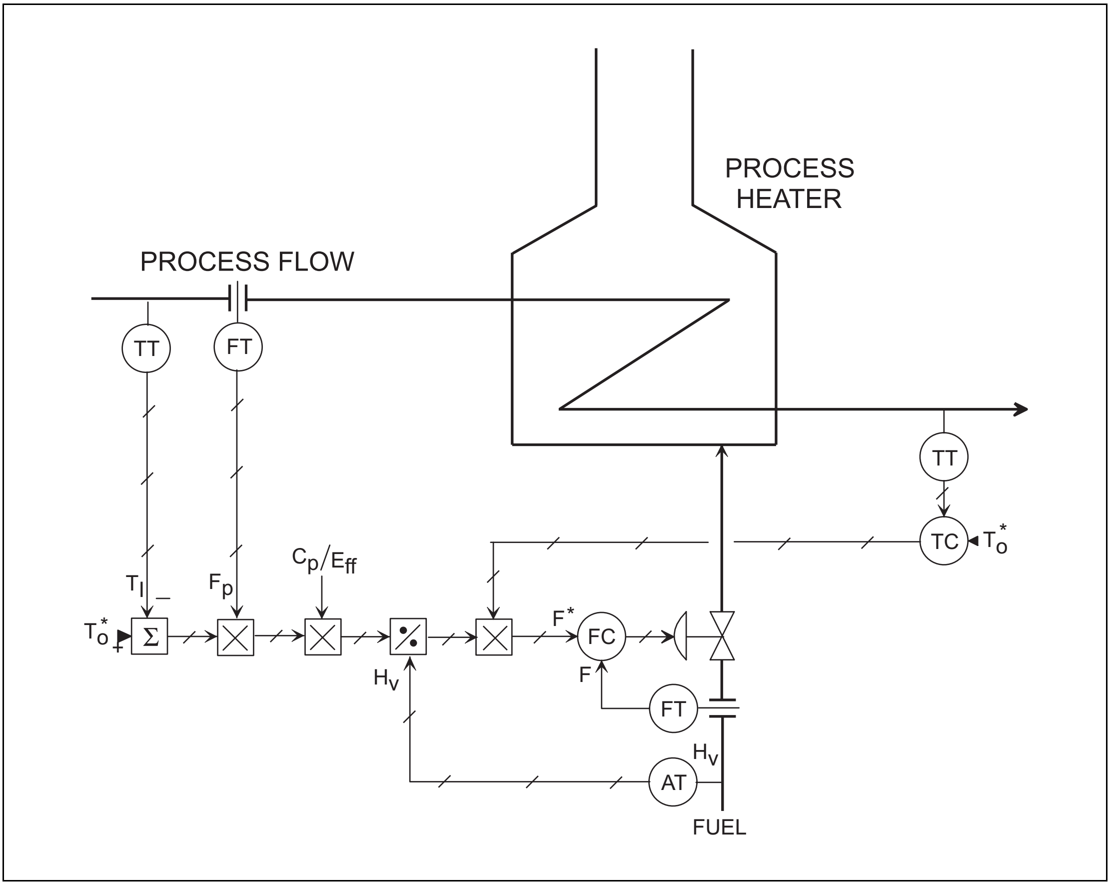

Figure 11-7. Feedforward Control with Multiplicative Feedback

described above. Instead, however, let us suppose that the temperature controller output represents the required ratio of fuel to process flow. That is:

$$\text{FeedbackController Output} = \text{Required}\ \frac{\text{Fuel}}{\text{Process Flow}}\ \text{Ratio}$$

$$= \left(\frac{F}{F_p}\right)^*$$

Then, the required fuel set point is computed as the required fuel-to-process flow ratio times the measured process flow rate, or:

$$F^* = \left(\frac{F}{F_p}\right)^* \times F_p$$

How does this affect the process gain as seen by the temperature controller? As before,

$$K_p = \frac{\Delta \text{Measurement}}{\Delta \text{Controller Output}}$$

Now, however, K_p is given by the following:

$$K_p = \frac{\Delta \text{Temperature}}{\Delta\left(\dfrac{\text{Fuel}}{\text{Process Flow}}\right)}$$

$$= \frac{\partial T_o}{\partial\left(\dfrac{F}{F_p}\right)}$$

Thus, from Equation 11-3,

$$K_p = \frac{H_v \times E_{ff}}{c_p}. \tag{11-5}$$

Equation 11-4 shows that with multiplicative feedback, the process gain of the feedback loop (often) does not vary with changes in process throughput.

The preceding presentation was somewhat rigorous mathematically. We could have made the same argument without assuming that process flow was the only disturbance, but at the penalty of greater mathematical rigor. We would have arrived at the same conclusion—that process gain of the feedback loop does not vary with changes in process flow rate.

Let us try a more “common sense” argument to justify using multiplicative feedback-feedforward rather than additive feedback-feedforward. If process flow increases by a certain amount, say 10 percent, then the fuel should also increase by approximately 10 percent. In other words, process flow and fuel should be scaled up and down together, at approximately a constant ratio. If the temperature controller sets the ratio between fuel and process flow, then only time that control action is required is when the ratio requires adjusting.

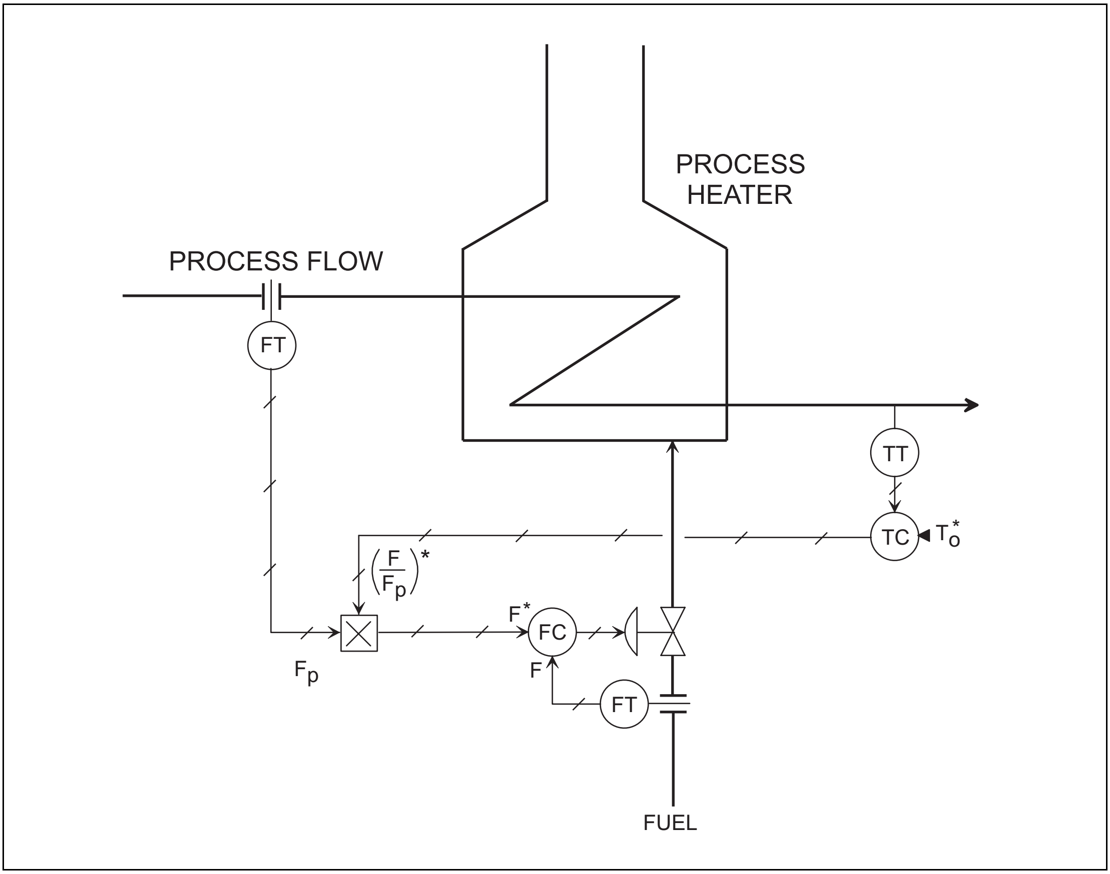

Figure 11-8. Simplified Illustration of Feedforward Control with Multiplicative Feedback

Note that the control scheme shown in Figure 11-8 contains exactly the same components as were depicted in Figure 10-3, which presented air-to-fuel ratio control with the O_2 controller providing feedback adjustment of the ratio. The corresponding elements are:

	Process Heater	Stack O_2 Control
Feedback Controller	Outlet Temperature	Stack O_2 Controller
Disturbance	Process Flow	Fuel Flow
Manipulated Variable	Fuel Flow	Air Flow

This, whether it is called ratio control or feedforward control, is one of the most important control configurations in industrial applications. It finds use in many process industries.

So far, we have presented two methods for combining feedback and feedforward: additive and multiplicative. How does one know which one to use? The answer is that it is never wrong to use additive feedback, but sometimes it may be better to use multiplicative feedback. The applications in which this is the case are those in which both the primary disturbance and the manipulated variable are flows, and where it makes sense that the two should be scaled up or down together. In these cases, use ratio control (i.e., multiplicative feedback). If the primary disturbance is not a flow rate, or if maintaining a constant ratio between the disturbance and manipulated variable does not make sense, then use additive feedback.

These two methods for combining feedback and feedforward constitute a large majority of the applications in actual practice. There is at least one other method, however, that can be used in special circumstances.

♦ Feedback Adjustment of the Feedforward Controller's Reference Value

In the implementation of the feedforward scheme shown in Figures 11-6 and 11-7, a reference value for the feedforward controller is required. This value is sometimes called the set point of the feedforward controller. When we add feedback, the feedback controller also has a set point. How do these two values compare?

With either the additive or multiplicative feedback control schemes shown in Figures 11-6 and 11-7, these reference values should be the same. However, another method for combining feedback and feedforward control is to let the feedback controller adjust the set point of the feedforward controller, as shown in Figure 11-9.

The rationale for this strategy is simple to explain. Suppose you wish to maintain a certain heater outlet temperature, say 500°F. Suppose the feedforward control scheme is configured as shown in Figure 11-5, with 500°F as the reference value for the feedforward controller. But then suppose that, due to unmeasured disturbances, the feedforward controller produces an output temperature of 490°F, 10° less than desired. This scenario is easy to solve. If the feedforward controller produces 10° less than desired, then give it a reference value that is 10° higher than what is really desired. To automate this procedure, let the feedback controller adjust the set point (reference value) of the feedforward controller. The feedback controller output signal should be scaled so that at mid-range its value represents the normal set point of the feedback controller. Then a signal range of 0 to 100 percent will represent a +/- correction, by a reasonable amount, of the feedforward controller reference value.

The advantage of this scheme is that the process gain of the feedback loop is a constant value of 1.0 under any kind of load condition. (When we presented multiplicative feedback, we stated that the process gain of the feedback loop was constant over variations in process throughput. We did not point out that the process gain would vary with changes in required temperature pickup, however. That is, if the inlet temperature varies significantly, the feedback controller gain would have to be adjusted. This is not normally as serious a problem as variations in the process throughput, however.)

If the scheme of combining feedback and feedforward by having the feedback adjust the reference value of the feedforward controller is so good, why isn’t it used more often? Because the opportunity to use does not present itself that often, that’s why. If inlet temperature were not

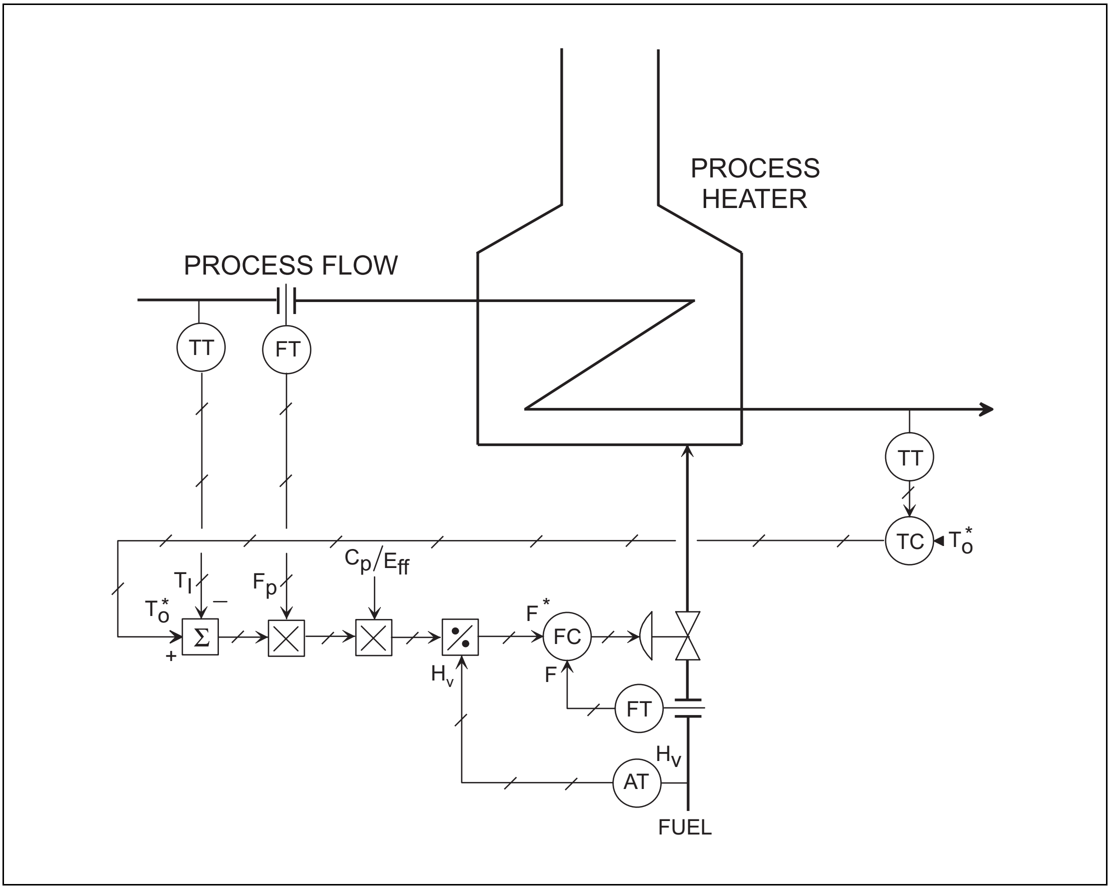

Figure 11-9. Combining Feedforward and Feedback Control by Adjusting the Reference Value of the Feedforward Controller

one of the disturbances considered in Figure 11-9, no opportunity would have arisen to combine feedback and feedforward in this manner. Compare Figure 11-8, for example.

❖ DYNAMIC COMPENSATION

When we presented the homespun example of controlling a domestic furnace based on a measurement of the outdoor air temperature, we identified three possible barriers to implementing perfect feedforward control. These were:

- There may be other disturbances;
- Our process model may be incorrect;
- The dynamics of the process had not been considered.

Combining feedback and feedforward control addressed the first two of these problems. We will now give attention to incorporating process dynamic characteristics into the control scheme. For this purpose, we will take a more abstract view of the process and depict it by the block diagram shown in Figure 11-10. This generic process is subject to two external influences, a load (or disturbance) and a control effort. These represent inputs to the process. One output is produced—the process variable.

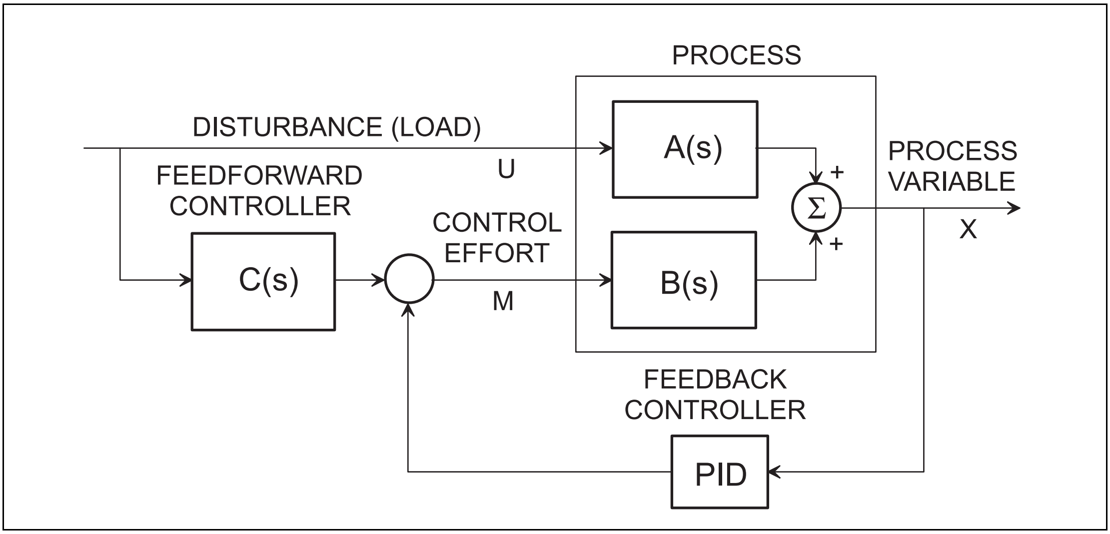

Figure 11-10. An Abstract View of Feedforward Control

If we relate this generic process block diagram to the process heater we have been using, we see that the generic process variable, x, is analogous to the heater outlet temperature; the load, u, is analogous to process flow; and the control effort, m, is analogous to the fuel-flow controller set point.

Why fuel set point, rather than fuel flow itself? In most feedforward and combined feedforward-feedback control schemes, the objective of the higher-level control is to determine the correct set point for a lower-level flow controller, say, for fuel, steam, and the like. As long as the lower-level flow controller fulfills its mission properly, it can be considered to simply be part of the process.

Each of the inputs can be presumed to have a dynamic influence on the process variable. That is, a change in either load or control effort will not (usually) have an immediate and full effect on the process variable. Its full effect, however, will be felt in the longer term. We can represent the effect of the load and control effort as the passing through of two dynamic paths called the "A" path (load to process variable) and the "B" path (control effort to process variable), respectively. The process dynamics can be represented by the transfer functions $A(s)$ and $B(s)$. (With a little more generality, we could consider multiple loads. There would be multiple "A" paths, represented by transfer functions $A_1(s)$, $A_2(s)$, ... , etc. To simplify the presentation, however, we will focus on a single load or disturbance.)

Usually (but not always), $A(s) \neq B(s)$, so to achieve good feedforward control, the control scheme must have some dynamic compensation. This is shown as $C(s)$ in Figure 11-10. We have yet to determine the form of $C(s)$, however.

When the feedforward path containing $C(s)$ is implemented, there are two paths from the disturbance to the process variable. One path is directly through the process, that is, through $A(s)$. The other path is through the feedforward controller, $C(s)$, then through the process transfer function, $B(s)$. The composite transfer function from disturbance to process variable is as follows:

$$H(s) = [A(s) + B(s)C(s)] \tag{11-6}$$

Our objective is to determine $C(s)$ in order to make the composite transfer function from disturbance to process variable equal zero. Why? If the composite transfer function from disturbance to PV equals zero, then disturbances will have no effect on the measurement. Another way to look at this is to say that we want to make the two paths identical but the mirror image of each other so the load signal, passing simultaneously through the two paths, will be canceled out exactly at the process output. We can determine the required $C(s)$ by setting the composite transfer function equal to zero:

$$A(s) + B(s)\, C(s) = 0$$

Therefore, the required dynamic compensation is as follows:

$$C(s) = -\frac{A(s)}{B(s)} \tag{11-7}$$

If we have designed the steady-state feedforward control system using the material we presented earlier in this chapter, then the dynamic compensation should have a unity steady-state gain. If not, then incorporate into $C(s)$ both steady-state and dynamic feedforward control terms.

♦ Determining A(s) and B(s)

Suppose we have a process (heater, distillation tower, etc.) that can be represented by the block diagram representation of Figure 11-10. To determine the process dynamic characteristics, we suspend any feedback and feedforward control action, thus maintaining a constant value for the control effort, m. Both m and the process variable should be near their normal operating points. Then we introduce a step-load change. With a constant value for the control input, we will observe a response due to transfer function $A(s)$ only. If we approximate this as first-order plus dead time (see Figures 6-5 through 6-8), then we can obtain numerical values for the parameters of $A(s)$:

$$A(s) = \frac{K_{pA}\, e^{-\theta_a s}}{\tau_A s + 1} \tag{11-8}$$

Similarly, at a time when there are no load changes, we can make a step change in the control effort and obtain numerical values for the parameters of *B(s)*:

$$B(s) = \frac{K_{pB}\, e^{-\theta_B s}}{\tau_B s + 1} \tag{11-9}$$

Then, by applying Equation 11-6 and simplifying, we get:

$$C(s) = -\frac{K_{pA}}{K_{pB}} \frac{\tau_B s + 1}{\tau_A s + 1} e^{-(\theta_A - \theta_B)s} \tag{11-10}$$

Note that if *A(s)* and *B(s)* have both been approximated as FOPDT, then *C(s)* is made up of three terms:

Feedforward gain $-\frac{K_{pA}}{K_{pB}}$

Lead-lag $\frac{\tau_B s + 1}{\tau_A s + 1}$

Dead time $e^{-(\theta_A - \theta_B)s}$

To provide a more intuitive understanding we will discuss each of these terms.

The steady-state gain term shows a negative sign. This was a result of our formulation of the composite transfer function, Equation 11-5. If one is faced with a practical example, the sign of this term will be obvious. Consider the process heater. If the process flow rate increases at constant fuel rate, then the outlet temperature will decrease, hence, K_{pA} will be a negative number. If the fuel rate increases at a constant load, the outlet temperature will increase, hence K_{pB} will be positive. The negative ratio of the two,

$$-\frac{K_{pA}}{K_{pB}},$$

will be a positive number. When used in a feedforward control scheme, an increase in process flow will cause a corresponding increase in fuel.

The second term is represented by a transfer function called a lead-lag. Although we derived it here as a mathematical expression, the lead-lag represents either a real piece of hardware or a software function block in a digital control system that can be implemented in our control scheme. A lead-lag unit, whether hardware or software, will have two adjustments or tuning parameters: a lead time, T_{LD}, and a lag time, T_{LG}. (The two adjustable parameters provided by some manufacturers are the lag time, T_{LG}, and the lead-lag ratio, $\frac{T_{LD}}{T_{LG}}$.) We will say more later about the input-output response of a lead-lag unit.

The third term is a mathematical expression for dead time or a pure delay. Ignoring any other dynamic effect, suppose there is a pure delay of θ_A time units in path A; and suppose there is a pure delay of θ_B time units in path B. Suppose further that θ_B is less than θ_A. Then, in order for the two signal paths to be identical, the signal through the feedforward controller would have to delayed by the difference between θ_A and θ_B. The last term in Equation 11-9 is merely a formal mathematical confirmation of what our intuitive concept tells us.

Practically all digital-based control systems will include a dead-time function block in their function block library. Thus, Equation 11-9 can easily be implemented in a digital system. In the analog world, however, no practical time delay unit exists that would provide the dead time most industrial applications require. Hence, if we were attempting to implement Equation 11-9 with analog equipment, we would have to approximate the dead time by using multiple lag or lead-lag units.

Note that θ_B must not be longer than θ_A. If it were, the exponent of the third term of Equation 11-9 would become positive, which would represent a "predictor" rather than "delay." This would be impossible to implement. In practical terms, this would imply that, by the time the feedforward controller senses a disturbance, it is already too late to correct for it. Feedforward control can still be applied in this situation to achieve long-term correction, as long as it is recognized that in the short term there will be a deviation that feedforward control is unable to correct.

In summary, we see that if we have approximated each of the two process dynamic paths by a function no more complex than a FOPDT, then the total feedforward controller, including dynamic compensating terms, will be no more complex than a lead-lag plus dead time. Often, we can omit some of these terms. For example, if the dead times through the two process paths are equal, then a dead-time term is not required in the feedforward controller. If all the dynamic terms in the two paths are equal, then the feedforward controller will require steady-state compensation only. An example of this is the air-fuel ratio control system for maintaining constant stack O_2. Both the fuel and the air enter the combustion zone at approximately the same point; neither has an instantaneous effect on the O_2 sensor. But since the effect of changes in air and in fuel travel the same path to the O_2 sensor, we can expect that the dynamics through the A and B paths are equal. Hence, dynamic compensation is not required in the air-fuel ratio (feedforward) control scheme.

Let us summarize the steps for implementing feedforward control with dynamic compensation:

(1) Test each of the process paths. If the process dynamics can be approximated by a FOPDT model, then determine numerical values for process gain, dead time, and

time constant for each path. (The procedure for approximating feedforward controller parameters is more forgiving of uncertainty than that for determining feedback controller tuning parameters.)

(2) Take the ratio of the two transfer functions. The resulting form should be no more complex than a lead-lag plus dead time, perhaps with a nonunity gain. The table in Figure 11-11 summarizes how to determine feedforward controller tuning parameter values from the process data.

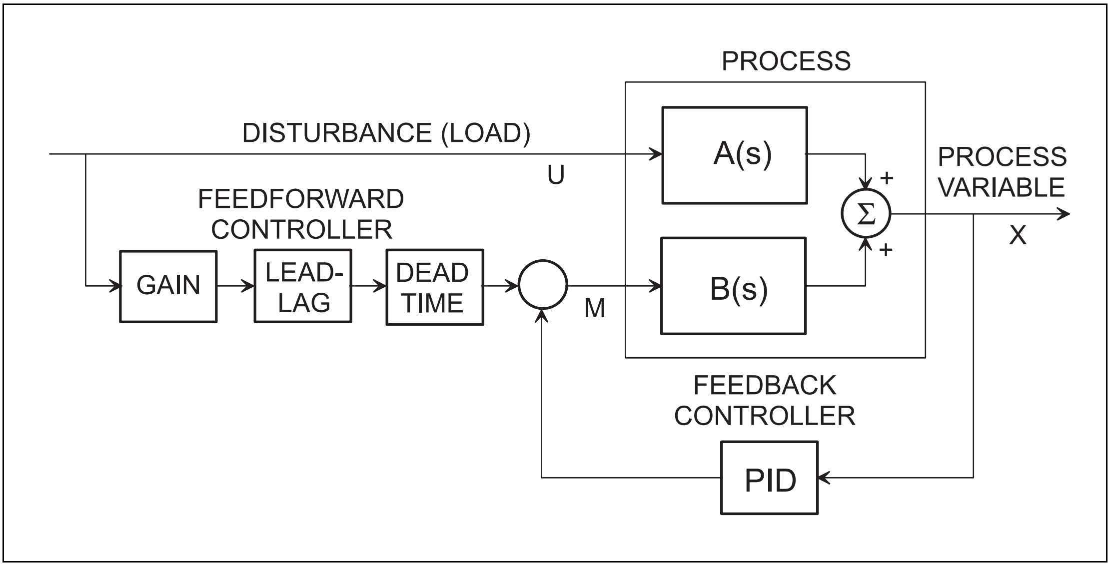

Figure 11-11. Dynamic Compensation Terms for Feedforward Control

Table 11-1. Adjusting Dynamic Compensation Terms

Function Block	**Form**	**Number of Parameters**	**Where do values come from?**
Gain	K_f	1	$K_f = -\frac{K_{pA}}{K_{pB}}$
Lead-Lag	$\frac{T_{LD}s + 1}{T_{LG}s + 1}$	2	$T_{LD} = \theta_B$ $T_{LG} = \theta_A$
Dead Time	$e^{-T_{DTM}s}$	1	$T_{DTM} = \theta_A - \theta_B$

♦ Fine-tuning the Feedforward Controller

Once a feedforward control loop is set up, it may need to be fine-tuned to improve its response. By itself, a feedforward control system will rarely provide perfect compensation for the measured disturbance. Hence, feedforward control will usually be combined with feedback. Since feedback control will correct for any lack of compensation made by the feedforward loop, one may ask what is the incentive for fine-tuning the feedforward loop after it is initially set up?

Recall that feedback control acts only after the fact. It must see an error in order to make a correction. The closer to perfect compensation the feedforward controller provides, the less correction is required by the feedback controller. Hence, the feedback penalty will be reduced.

The method for fine-tuning that we will present here depends upon the ability to test the feedforward controller without using feedback by forcing load changes on the feedforward control system and process. In real applications this may not be possible, or it may be possible only to a limited extent. One may have to work with natural disturbances that occur. Even if it is possible to test the feedforward control system and process, a point of diminishing returns is reached where further testing to improve the feedforward controller is not warranted by the small decrease in error that might be gained. Each application will have to be judged on its own merit. Yet, by understanding the principles of feedforward fine-tuning described later in this chapter, one can decide whether further testing for performance improvement is justified and how to make the best use of the resulting test data.

For tutorial purposes, we will use the liquid-phase process heater as the basis of our discussion, since it is much easier to consider the effects of a real system than an abstract one. We are using additive feedforward control in this tutorial example because it gives us the opportunity to describe the effect of all feedforward tuning parameters, including the feedforward controller gain. For actual applications, however, we recommend multiplicative feedforward control for the heater. If it were used, there would be no gain function block, and only the lead-lag and dead-time function blocks would require adjusting.

For the heater, assume that the main disturbance is change in feed flow rate. Also, assume that we have made the process tests and determined approximate process models $A(s)$ and $B(s)$ and that additive feedforward control has been implemented, as shown in Figure 11-11. If we have approximated both $A(s)$ and $B(s)$ as FOPDT, then the feedforward controller consists of a steady-state term plus dynamic compensation that consists of lead-lag plus dead time, as shown in Figure 11-12.

Now suppose that we make a plant test to determine the response using feedforward control alone, that is, with the feedback controller in manual. We are trying to make the feedforward compensation as good as possible, so at this point we do not want to see the combined effects of feedforward and feedback.

If our compensation were perfect, then a step change in the disturbance would cause no change in the outlet temperature. In practice, such a result will rarely be the case, for the following reasons:

- The process data taken from the initial tests may have been incorrect;
- The process characteristics may have changed;
- The process may be operating at a different point.

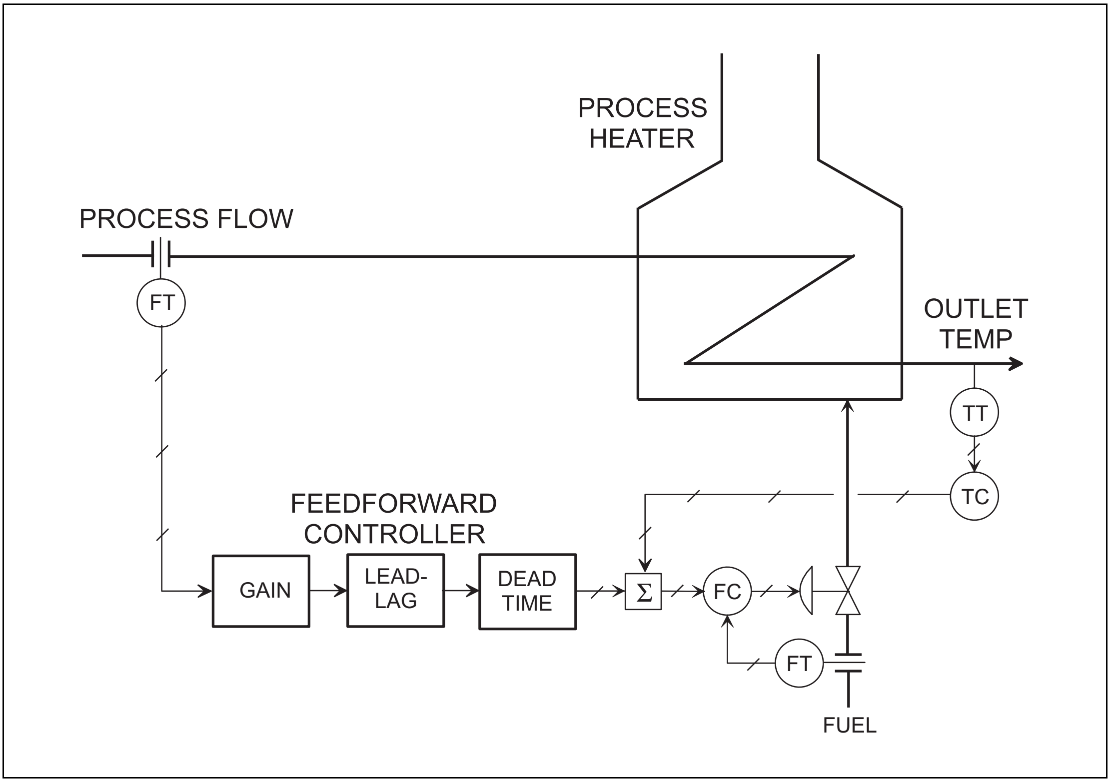

Figure 11-12. Feedforward Control, with Dynamic Compensation, of a Process Heater

With additive feedforward control, four types of adjustments that can be made. (If multiplicative feedforward is used, there are only three. Feedforward gain is not used.) These adjustments are as follows:

- Feedforward controller gain;
- Feedforward controller dead time;
- Feedforward controller lead-lag ratio, at constant lag time;
- Feedforward controller lag time, at constant lead-lag ratio.

The first of these compensates for long-term, steady-state effects. The latter three compensate for dynamic errors. The distinction among these four relates to how soon the error occurs relative to the disturbance or load change.

If the results of our test of the feedforward response (without feedback) shows a long-term change in measurement (outlet temperature), as shown in Figure 11-13, then the required correction is obvious—the feedforward gain needs to be changed. (In Figure 11-13, an increase in feed flow caused an increase in fuel. Since the outlet temperature shows a long-term drop, it is obvious that the fuel was not increased enough, hence the feedforward gain needs to be increased.)

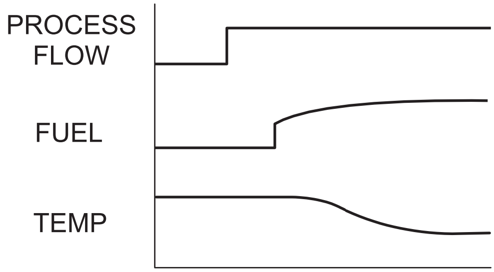

Figure 11-13. PV Response When Feedforward Gain Is in Error

Suppose, however, that the long-term feedforward compensation is correct, but that in the short term there is a deviation, as shown in Figure 11-14. This scenario illustrates that the feedforward gain is correct but that the dynamic compensation terms must be improved. But which term(s)? In which direction and adjusted by how much?

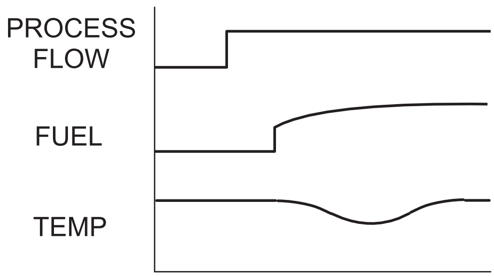

Figure 11-14. PV Response When Dynamic Compensation Is in Error

To better understand the effect of adjustments to dynamic compensation terms, we must understand the behavior of the lead-lag function. Most lead-lag functions (which are implemented either in hardware or as software function blocks) have two parameters that can be adjusted: lead time and lag time. The responses to a step input change for various combinations of lead and lag time are shown in Figure 11-15. The general response for any combination of lead and lag can be stated as an initial jump that is equal to the ratio of lead time to lag time (called the “lead-lag ratio”), followed by a decay to equilibrium at a rate that is determined by the lag time. Thus, as shown in Figure 11-15, the two pertinent parameters for lead-lag tuning are the lead-lag ratio and the lag time, not the lead and lag times individually.

Now return to the problem of fine-tuning the feedforward controller for the response shown in Figure 11-14. To analyze this problem, we will use the principle of superposition. That is, if the same test were repeated but with different dynamic compensation terms, we would obtain a different fuel response. The difference in fuel responses between the two tests would be the

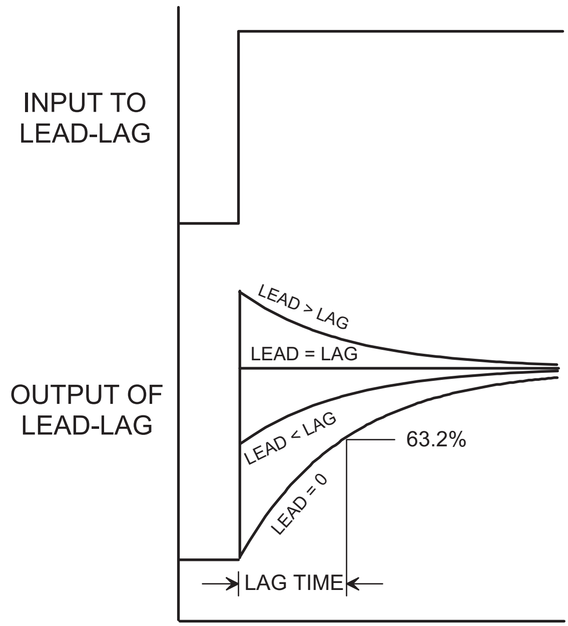

Figure 11-15. Response of Lead-lag to Step Input Change

differential adjustment in fuel. The difference in outlet temperature responses between the two tests would be the differential response in measurement. Our objective is to adjust the dynamic compensating terms so the differential measurement response, when added to the measurement response obtained in Figure 11-14, will result in no deviation in final temperature.

For instance, suppose the dynamic compensator consists only of dead time. Consider the two tests shown in Figure 11-16a. In test #2, the dead time is less than that of test #1. The difference in fuel response (i.e., test #2 - test #1) is approximately a pulse. The incremental response of the outlet temperature to this incremental fuel change is shown on the lower graph.

The incremental effect of a dead-time change is easy to see. The incremental effect of changes in the lead-lag parameters are more difficult to visualize. Figure 11-16b shows the incremental effect of increasing the lead-lag (at constant lag time). Since the increment of fuel is applied over a longer time period than for 11-16a, the incremental effect on temperature is prolonged. Figure 11-16c shows the effect of decreasing the lag time (at constant lead-lag ratio). Initially, there is no incremental change in fuel. The bulk of the incremental fuel change occurs much later; hence, the incremental effect on temperature is both delayed and prolonged. In each of these diagrams the load change is assumed to be an increase in feed-flow rate, and the adjustment is made in a direction that will give a positive incremental fuel change to compensate for the short-term fuel deficiency shown in Figure 11-14.

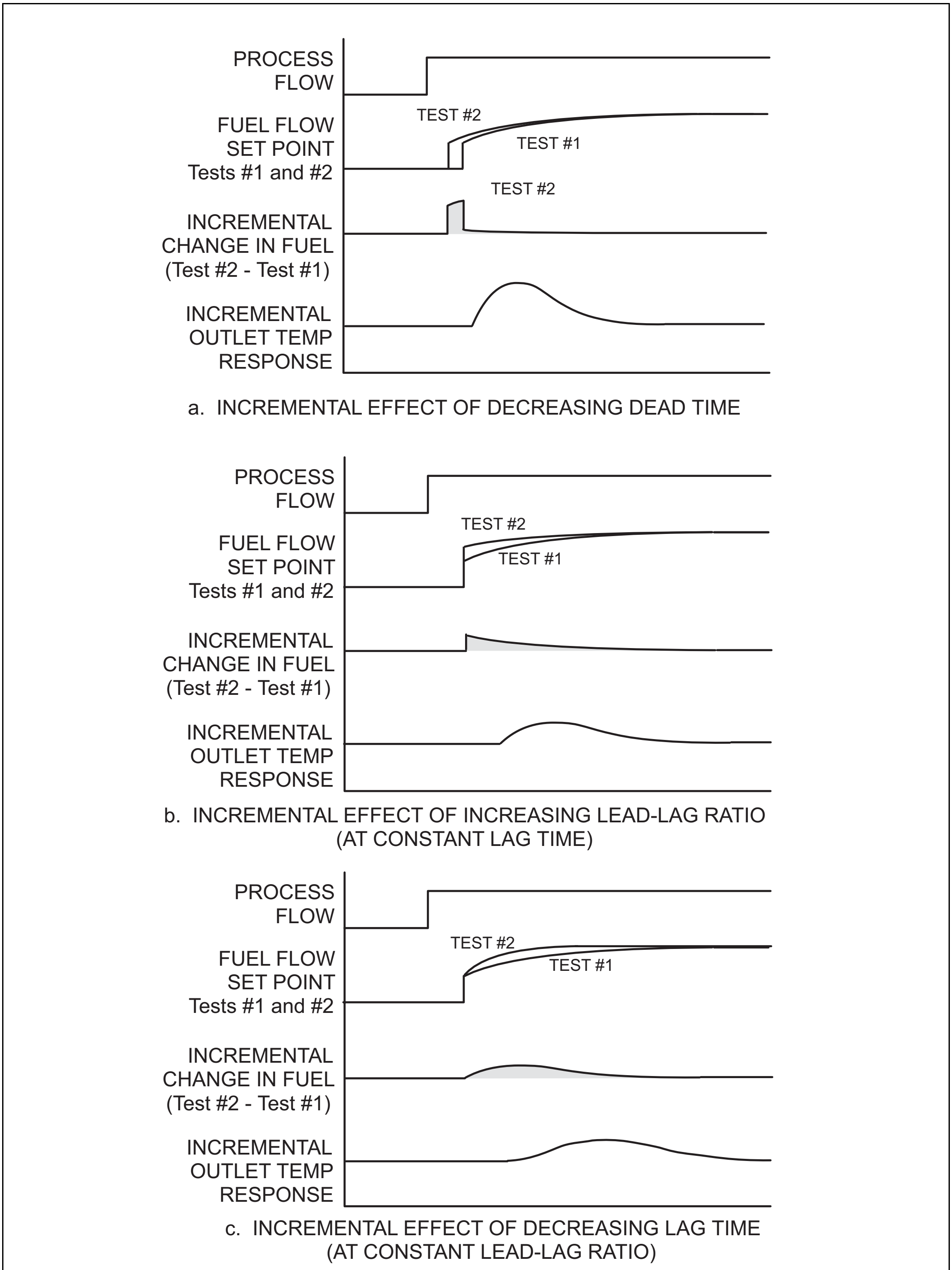

Figure 11-16. Incremental Effect for Various Dynamic Compensation Adjustments

The incremental change in outlet temperature for all three forms of adjustment are plotted on a single graph in Figure 11-17. From this figure we can make the following generalizations:

- A change in dead time only will result in an incremental process response that occurs relatively soon after the load change;
- A change in only the lead-lag ratio (at constant lag time) will result in an incremental process response that is somewhat further out in time from the time of the load change;
- A change in the lag time (at constant lead-lag ratio) will result in an incremental process response that is the furthest away in time.

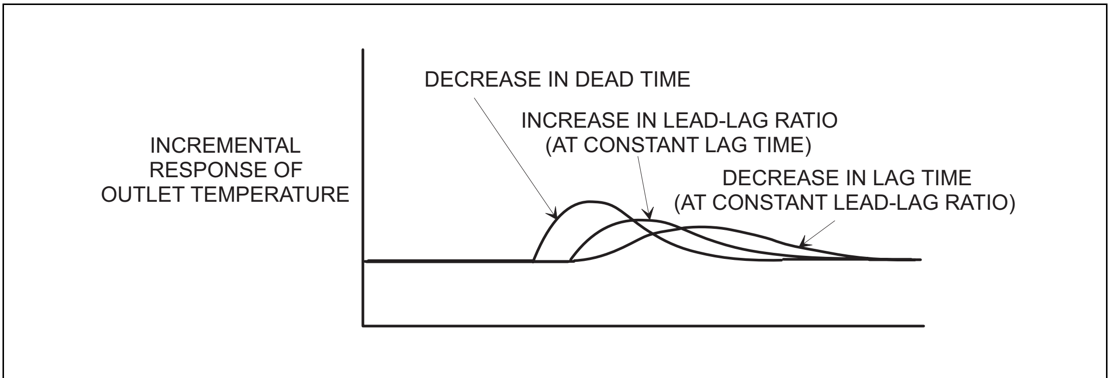

Figure 11-17. Summary of Incremental Effect for Various Dynamic Compensation Adjustments

With these generalities in mind, one can observe the initial response of the feedforward control system to load change, determine the direction and relative time scale of the required incremental corrective process response, and adjust the dynamic compensating terms accordingly.

❖ FURTHER CONSIDERATIONS OF THE FEEDBACK CONTROLLER

The design procedure illustrated in the previous chapter applied feedforward control first, then added feedback. The procedure can be reversed, however. If one is determining a control strategy for a new process unit, it is probably better to first determine the critical process variables and their method of feedback control. Then consider whether or not feedforward control (or cascade or ratio) can be applied to minimize the feedback penalty paid by the feedback controller.

The feedback and feedforward can be combined by any of the methods we have discussed in this chapter. If the feedforward controller has dynamic compensating terms, be sure these are in the feedforward path only, not in the path that results after feedback and feedforward are combined.

The feedback controller should certainly use proportional and reset; it probably will not need derivative. After all, one purpose of derivative was to obtain a better response to a load upset. With feedforward control, we've taken care of the load upset in a different way, hence derivative is not needed.

The primary purpose of feedback is to correct for steady-state errors in the feedforward controller. Thus, the responsibility of the feedback controller is not as great as it would be if it were controlling the process by itself. Therefore, it is likely that you can relax the tuning of the feedback controller (lower gain, longer integral time) in comparison to the tuning that would be used with feedback alone.

❖ FEEDFORWARD: IN PERSPECTIVE

To successfully apply feedforward control, one must first determine which applications are good candidates for feedforward. The candidate application will probably be an important process variable that has a significant effect on product quality, throughput, or efficiency. The major disturbance, or disturbances, must be measurable. There must be a sufficient knowledge about the process dynamic behavior to develop the feedforward control relationship. And finally, one must consider the nature of the disturbances themselves. If the disturbance is of high frequency (relative to the process dynamics), the process itself will filter out the effect of the disturbance. On the other hand, if the disturbance is of low frequency (or slow acting), then feedback control alone can cope with it. (Recall that in chapter 8 we said that if the disturbance was slow, the feedback penalty would be stretched out, so the maximum deviation at any one time would be reduced.)

An example of an application of feedforward control is in distillation column control. One method for controlling product quality is to control the flow rate of a product stream. Consider a simple binary column. If the required separation of the components is high, then the approximate required product flow rates can be computed by measuring the feed-flow rate and composition. Figure 11-18a shows a column separating a light and a heavy component. The approximate required flow rate of the lighter product (distillate) is computed from feed conditions. A product composition controller (which could be an inferred composition, such as from a temperature measurement) then adjusts the required ratio of distillate-to-light component in feed. This is an example of combining feedback and feedforward by ratio control.

Now consider the fact that in many applications column feed rate varies considerably. It may be controlled by a level controller on an upstream process unit. Feedforward control from the feed rate is certainly desirable. The feed composition, however, may not vary as rapidly. Hence, we can dispense with feedforward control from feed composition, thus simplifying our control system and perhaps avoiding the cost of a composition analyzer. This is shown in Figure 11-18b. The product composition controller sets the required ratio of distillate to feed. Should the feed composition change, the required ratio will also have to be changed by paying

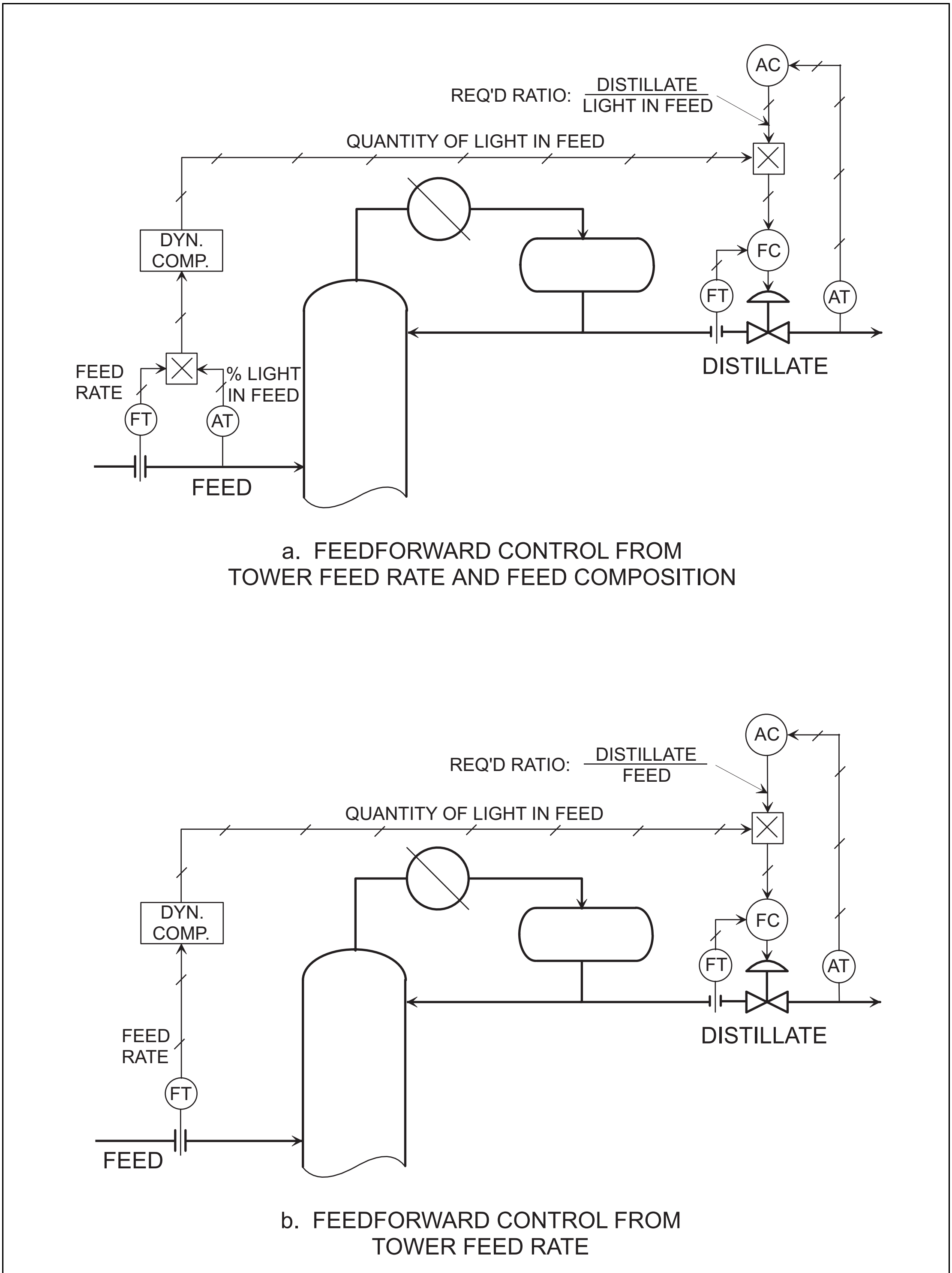

Figure 11-18. Example of Feedforward Control Applied to a Distillation Tower

the feedback penalty. If the feed composition changes slowly enough, however, the feedback controller can change its output gradually over time without incurring an excessive deviation at any one moment.

This last example illustrated the thought process that is required to select suitable feedforward control applications. It was not intended as a dogmatic recommendation for distillation tower control. The principle concept behind feedforward control—obtaining control action without paying the feedback penalty—takes many forms. Some processes that are subject to significant grade changes or changes in feedstock have special programs for driving (often by ramping) the manipulated variables to new operating points during the transition period. If feedback control alone were utilized to drive these manipulated variables to the new operating points, a significant feedback penalty would have to be paid. Very simply, this translates into off-spec production. The author has witnessed a "crude switch" program at a petroleum refinery that reduced the transition time from around eight hours to approximately two hours.

Control loops in which the set point is determined by a schedule or profile generator can benefit from a feedforward approach. Consider the process shown on Figure 11-19a. This could represent any of a number of applications such as a batch chemical process, a food process, annealing furnace, and so on. The set point of the temperature controller is generated by a profile generator. We know that, using this feedback control scheme alone, there is no hope that the temperature will follow the profile exactly since a feedback penalty must be paid in order to move the final control device.

If we had a dual, synchronized profile generator as shown in Figure 11-19b, we could significantly improve the control. One profile generator would generate the required set point trajectory, as before. The other profile generator would generate the approximate valve (final control element) trajectory. This would be a feedforward signal that would be combined with feedback trim. If the major portion of the valve signal is directed by the feedforward signal, then only the feedback controller has to trim this signal, hence we should expect a considerable reduction in the feedback penalty.

How would we arrive at the trajectory for the final control element signal? The answer: Run the process (batch, grade change, etc.) using feedback control alone and record the controller output. That will be a good starting point for the feedforward trajectory.

In a similar vein, the author has on several occasions watched, and deplored, operators who switched a controller to manual, then manually manipulated the controller output. This was often done after a process upset. "Leave it alone," I implored. "Let the controller do its thing. It will eventually bring the measurement back to the set point." True, it would. But what the operators realized, probably only subconsciously, is that in order for the controller to find a new output value a significant feedback penalty will have to be paid. For every bit that they can "help" the valve along, the feedback penalty will be reduced by that much. The success of these operators' action often depended upon the quality of their subconscious feedforward control model.

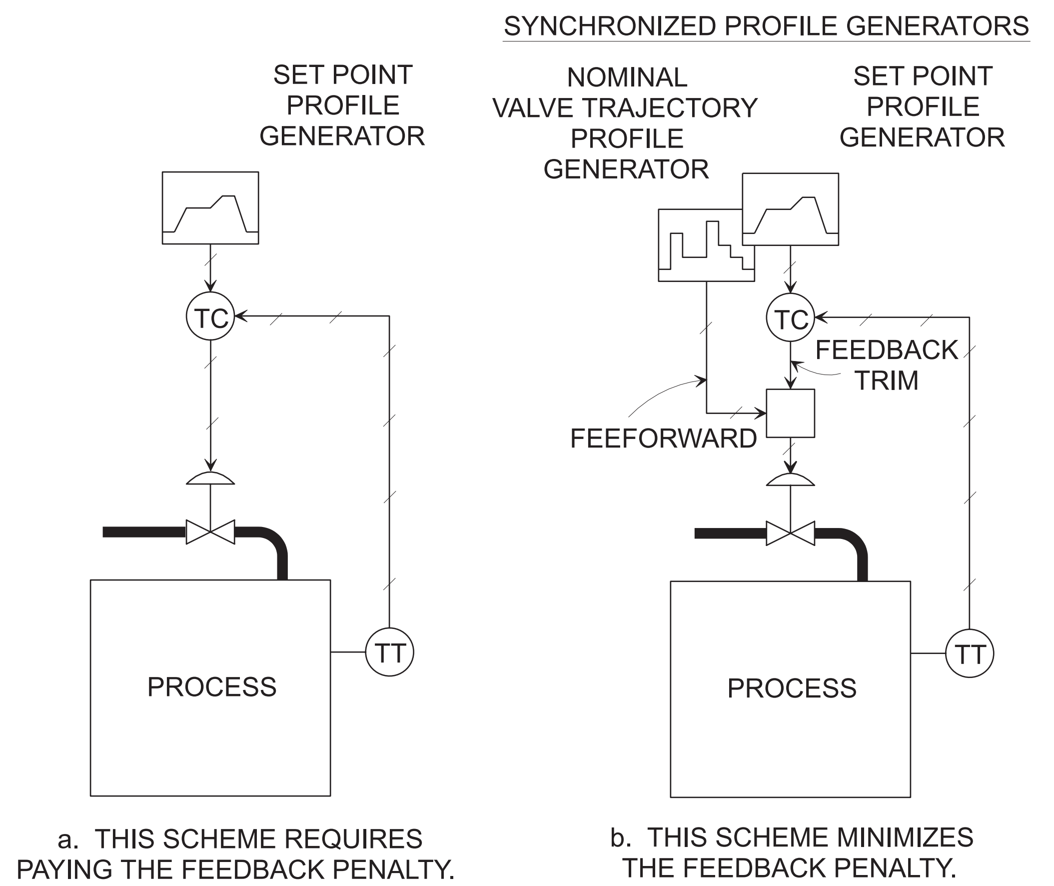

Figure 11-19. A Potential Application for the Feedforward Control Concept

In some cases, it may be desirable to be able to turn feedforward on and off, in response to operating conditions (e.g., feedforward off for startup operations) or simply at the operator's discretion. (In the latter case, there should be some positive logging of "who," "when," and "why" to prevent feedforward from being indiscriminately disabled.) In other cases, such as most boiler-drum level-control systems, feedforward is always on. (Some boiler-drum control systems offer the option of switching between feedback and feedforward with feedback trim as a part of the startup sequence.) There may also be interlocks that turn feedforward off under certain conditions such as analyzer failure.

❖ FEEDFORWARD CONTROL USING FOUNDATION™ FIELDBUS

Application function blocks defined by the Fieldbus Foundation standard (Ref. 11-1) envision *additive* feedforward. The standard PID function block provides for a feedforward input signal that is scaled and added to the PID algorithm output. A feedforward gain value is applied to achieve the desired feedforward contribution. If dynamic compensation is required, external function blocks such as lead-lag and dead time may be configured. Dynamic compensation blocks are in the calculate-class of function blocks, hence they do not support the back-calculation mechanism.

Figure 11-20 depicts a feedback-additive feedforward control scheme with a lead-lag dynamic compensation block. In this figure, the PID output sets a valve position; it could just as well be cascaded to the set point of a lower-level (secondary) PID.

All FOUNDATION™ Fieldbus (FF) devices that support a PID algorithm provide the additive feedforward input. Not all FF devices, however, will provide dynamic compensation blocks.

The FF standard also defines a Ratio block, thus providing for *multiplicative* feedforward control capability. The Ratio block is a control-class block; hence, it supports the back-calculation mechanism. A typical application for this block would be one of the feedforward control schemes for distillation towers, shown in Figures 11-18a and 11-18b. Dynamic compensation blocks can be configured in the feedforward signal to the Ratio block, as shown in Figure 11-21.

Note that not all manufacturers support the Ratio block. Some may not even support the dynamic compensation blocks.

❖ REFERENCES

11-1. Fieldbus Foundation, *Foundation Specification: Function Block Application Process*, *Document FF-891, Part 2.*

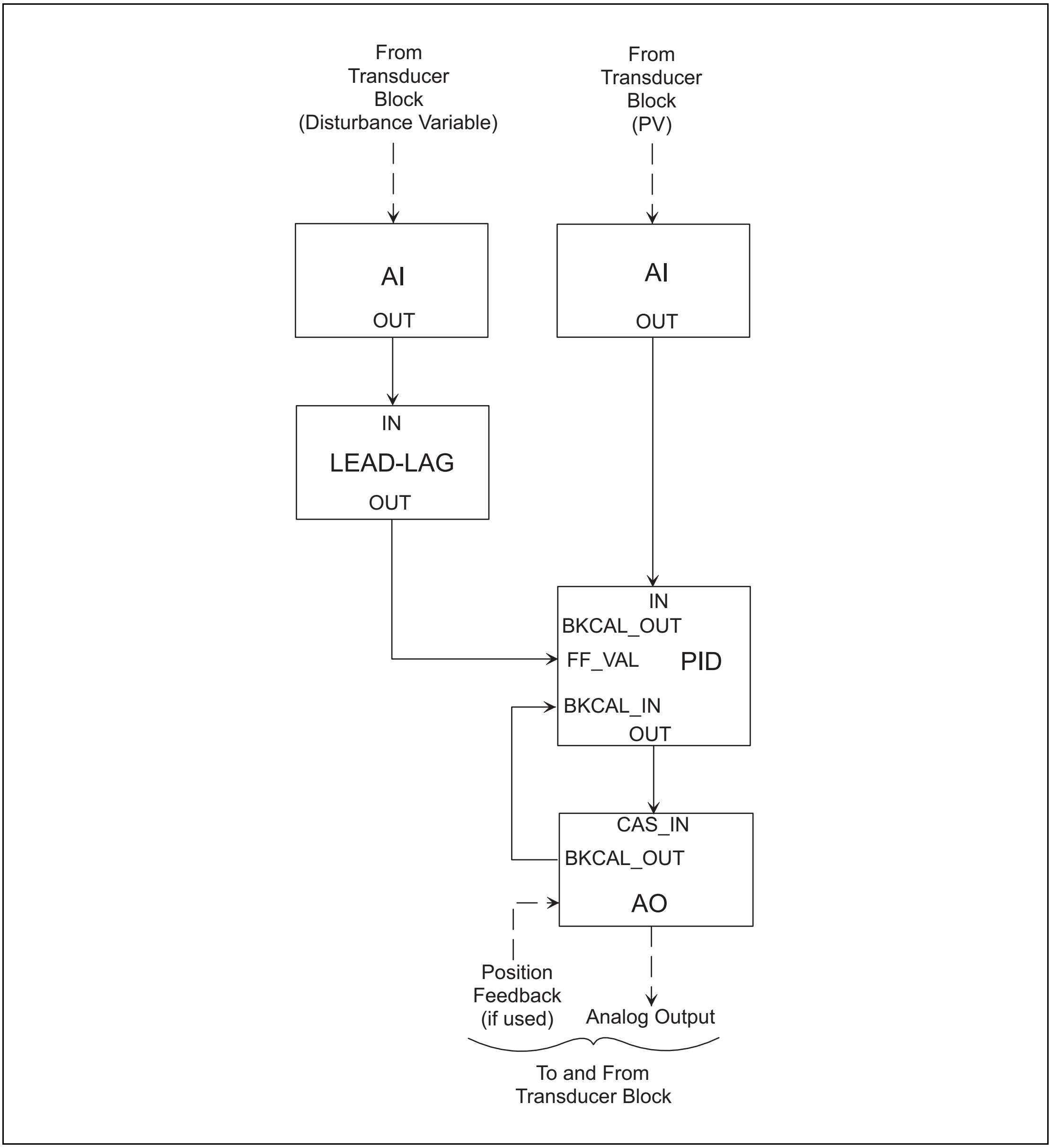

Figure 11-20. Additive Feedforward Using FOUNDATION™ Fieldbus Function Blocks

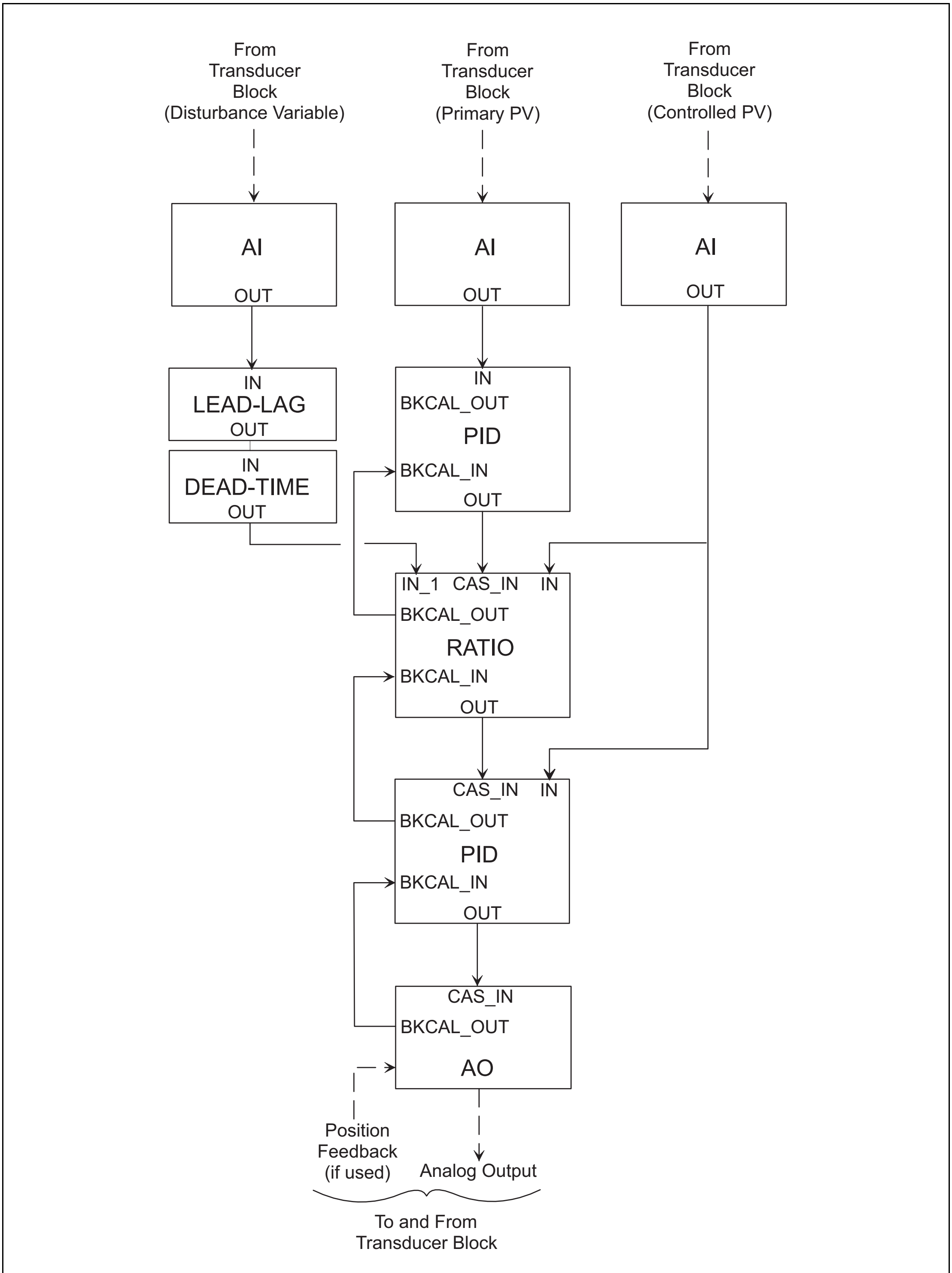

Figure 11-21. Multiplicative Feedforward Using FF Function Blocks

12 OVERRIDE (SELECTOR) CONTROL

❖ OVERRIDE CONTROL

Override control, also called selector control, exists when one process variable is the controlling variable in normal operation. During abnormal operation, however, another process variable assumes control to prevent some safety, process, or equipment limit from being exceeded.

A key element of an override control strategy is a selector switch, implemented either as a hardware device or a software function block. Depending on how it's configured, this selector switch passes the higher or lower of several input signals to its output. There are several ways of using selector switches in a control strategy. One is to select the higher or lower of several measurement signals to be passed on as the process variable to a feedback controller. For example, the highest of several process temperatures may be selected automatically to become the controlling temperature. As process conditions change, the location of the highest temperature may change also. The selector switch assures that, regardless of process conditions, the controlling point is the highest of the measured temperatures.

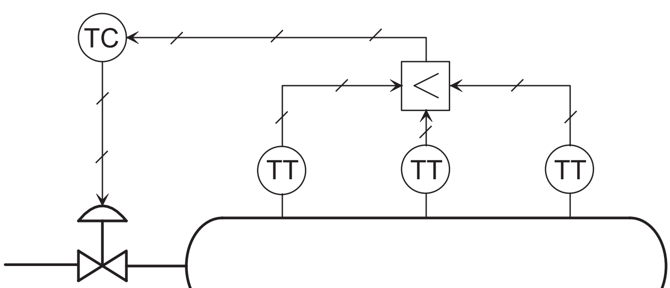

Figure 12-1. Controlling from the Highest Point of a Temperature Profile

Placing a selector switch in the measurement side of a controller, though perhaps important from the vantage point of a particular process application, poses very little technical challenge for the control engineer. If each of the process sensors responds in a similar way to changes in the controller output, then the transition from one sensor to another will be virtually imperceptible.

There are also selectors which select the middle of three inputs. These are used primarily in high-criticality applications, where the failure of a signal, either by going “high” or “low”, could not be tolerated. By taking the middle-of-three inputs, then as long as two of the signals remain viable, the application can continue.

From a control engineers, viewpoint, however, a more interesting application than these occurs when the selector device selects between the higher or lower of several controller outputs.

For example, suppose that the outlet temperature from a process heater is normally controlled by manipulating a fuel-valve position. In normal operation, the outlet temperature should be maintained at its set point. Suppose, however, that the temperature of the heater tubes has an operational limit. If the heater has been properly designed, the tube temperature will remain below the limit during normal operation. On the other hand, abnormal conditions, such as coking in the heater tubes or overloading of the heater, may cause the tube temperature to rise.

If a representative tube temperature is measured, a limiting controller may be used to prevent encroachment on the operational limit. Here, the two controller outputs would be connected to a low signal selector; the controller demanding the lower fuel valve position will override the other. In normal operation, this controller will be the heater outlet temperature controller.

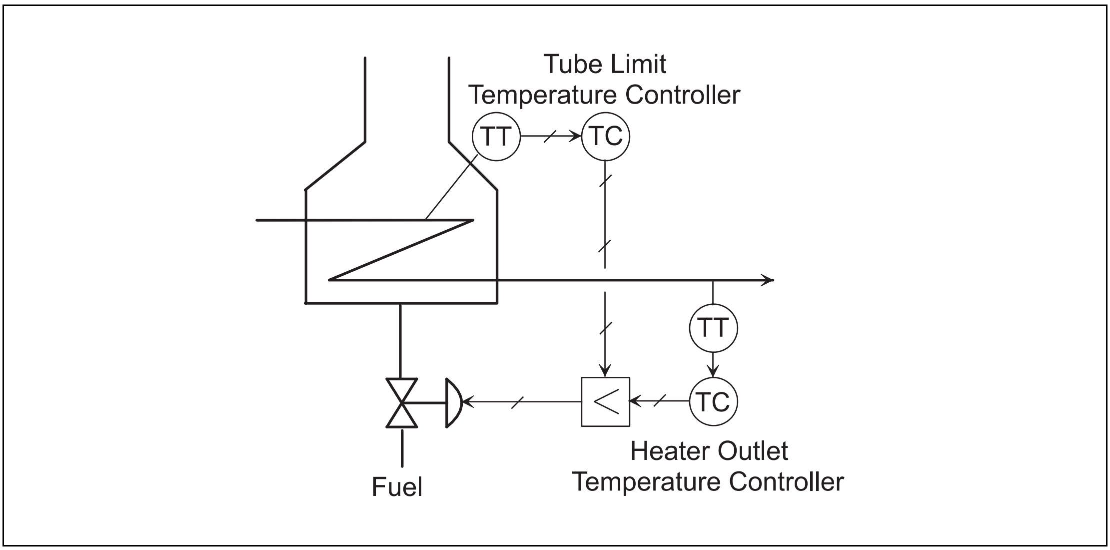

Figure 12-2. An Example of Override (Selector) Control

Note that we present this example merely as an easy-to-visualize application of override control. We will not attempt to explore in depth here the question of how to obtain a representative measurement of tube temperature. One possibility would be to measure several tube temperatures, as in Figure 12-1, and select the higher of the temperatures as the process variable for the limiting controller. We also assume that in an actual installation of this type, one or more alarm points would be set just below the set point of the limiting controller to warn the operator of an impending override. In addition, there would likely be an independent temperature sensor that has a safety shutdown point set just higher than the set point of the limiting controller, in case the override control failed to prevent the tube temperature from rising further. These additional items, however necessary, lie outside the scope of this book.

Let us first suppose that we utilize ordinary PI controllers for this application, as shown in Figure 12-3.[1] Since the fuel valve is undoubtedly a fail-closed type, then both controllers will be set for reverse action. In normal operation, the heater outlet temperature will be near its set point and the tube temperature will be below its set point (limit value). We can assume that the output of the heater outlet temperature controller is between its extreme limits, that is, it is someplace mid scale. If the heater has been in this condition for some time, with the tube temperature at less than its set point, then the integral action in the tube temperature controller will have increased its output to the maximum value. In other words, the heater tube temperature controller will be in a windup condition.

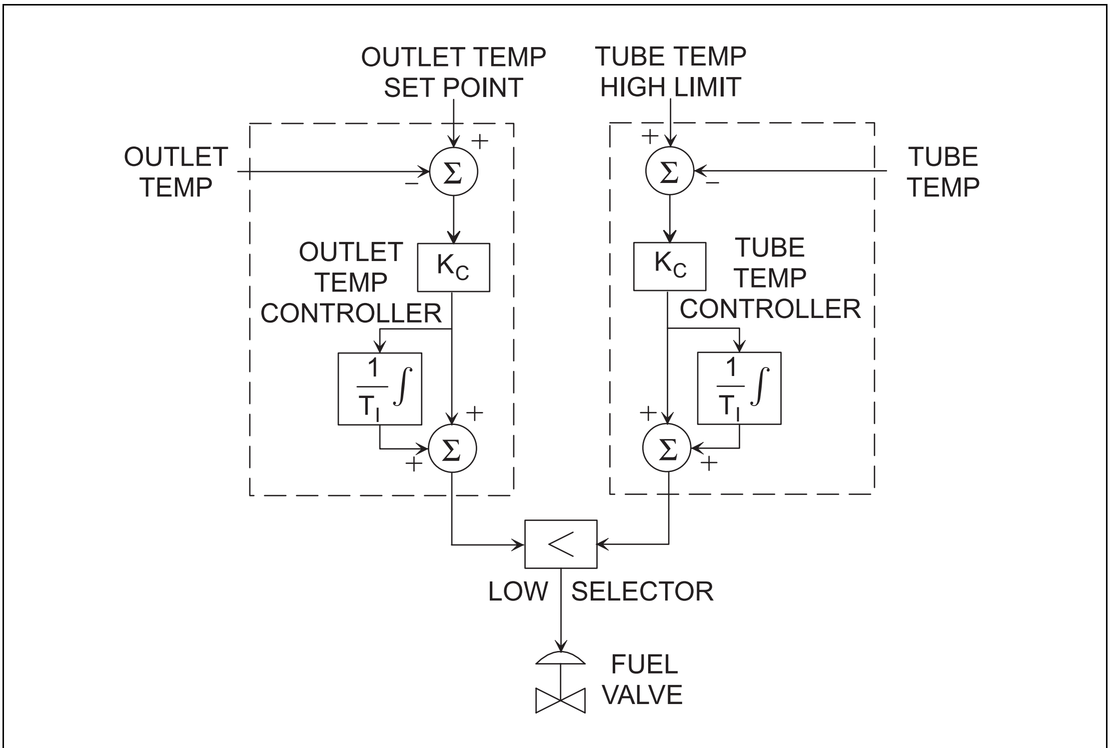

Figure 12-3. Override Control Using Ordinary PI Controllers (Not Recommended)

1. We refer to PI rather than PID because the derivative mode is inconsequential to the technology discussed here.

Since the output of the heater outlet temperature controller is less than that of the tube temperature controller, the outlet temperature controller's output will be selected and will be passed to the fuel valve. As long as the tube temperature stays below its limit, the heater will remain under the control of the outlet temperature controller, just as if the other controller were not present.

Now, suppose that, because of some abnormal process condition, the heater tube temperature rises toward the limit value. For this application, several process conditions could cause the tube temperature to rise. For instance:

- An increase in the heater feed rate would call for additional fuel to be released, consequently a hotter combustion zone and ultimately a higher tube temperature.

- Coking inside the heater tubes, which some steam reformer furnaces experience, will reduce the heat transfer. This will require a higher temperature in the combustion zone (i.e., additional fuel) to transfer sufficient heat to the process fluid. This also results in a higher tube temperature.

- A decrease in heater efficiency will also result in increased fuel firing, and consequently a higher tube temperature.

For the purpose of the discussion here, what caused the tube temperature to rise is unimportant. We only need to know that it is a situation that could occur. If the tube temperature rises slowly toward the limit without exceeding it, the output of the limiting controller will remain at its maximum (wound-up) value and will have no effect on the heater firing rate.

Once the tube temperature crosses the limit, however, the error reverses in sign and the controller output starts decreasing. It must decrease all the way below the output of the heater outlet temperature controller before it will have any effect on the position of the fuel valve. The tube temperature could remain over limit for a significant period before the limit controller overrides and begins reducing the fuel flow. The length of time depends on the rate at which the limit controller output decreases (which in turn depends on the controller tuning as well as the amount by which the limit value is exceeded) and the value of the heater outlet temperature controller output.

To summarize our understanding to this point, our application objective—installing a high-limit controller that can override the normal outlet temperature controller if the temperature of the tube becomes high—is a valid goal. Yet, our application is somewhat flawed because of windup experienced by the nonselected controller.

We will now digress and return to a PID controller modification that we first presented in chapter 5. Recall that Figure 5-12 and the discussion related to it presented the concept of external reset feedback (ERF). A PI controller constructed with ERF has an extra input port (the reset feedback port), which, if it is connected directly to the controller output, causes the controller to behave as an ordinary PI controller. The signal into the reset feedback port can originate from some place other than the controller output, however. We will make use of this feature in the application under discussion.

Suppose that we replace the two ordinary PI controllers with controllers that have ERF. The controller outputs are connected to a low-signal selector as before, but the reset feedback for both controllers originates from the output of the low-signal selector, as shown in Figure 12-4.

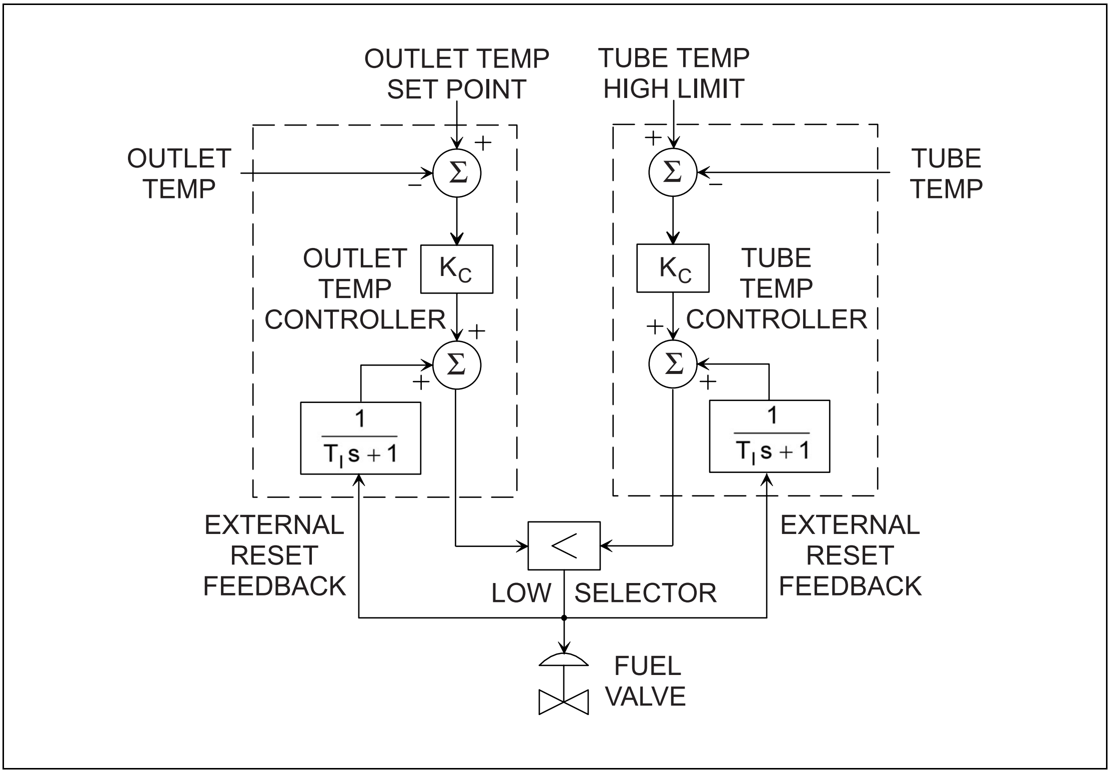

Figure 12-4. Application of External Reset Feedback for the Heater Tube Temperature Example.

We will demonstrate that with this arrangement the nonselected controller will not wind up, even though its set point and measurement may differ. To provide an intuitive insight into this idea, we will use the numerical values shown in Tables 12-1 through 12-5. Transfer each set of these numbers to a diagram like that shown in Figure 12-4 in order to follow a “moving” scenario.

Suppose that the outlet temperature set point is 40 percent of its scale, and the tube temperature limit set point is 70 percent of its scale. Suppose that in normal operation the valve is positioned at 50 percent open, the actual heater outlet temperature is maintained at set point, and the actual tube temperature runs below its set point, say, at 65 percent of its scale. Suppose further that both controllers are tuned with a gain of 1 (PB of 100%).

Table 12-1. Normal Operation

Signal	Heater Outlet Controller	Selector Output	Tube Limit Controller
Set Point	40%		70%
PV	40%		65%
Error	0%		5%
$K_C \times$ Error	0%		5%
Reset Feedback	50%		50%
Output of First-order Lag	50%		50%
Controller Output	50%		55%
Valve Position		50%	

Table 12-1 summarizes the various signal values during normal operation. Note that the system could remain in the normal operation state indefinitely without the nonselected controller (tube limit) winding up. The reason for this is that with the ERF configuration there is no free integrator with a nonzero input that would cause windup. Instead, there is a first-order lag whose input and output signals are equal in the steady state.

Now, suppose that the tube temperature rises slightly, say, to 69 percent of scale. Table 12-2 shows the new scenario. Since the tube temperature has not reached the set point, there is no effect on furnace firing.

Table 12-2. Tube Temperature Rise But Not To Limit

Signal	Heater Outlet Controller	Selector Output	Tube Limit Controller
Set Point	40%		70%
PV	40%		69%
Error	0%		1%
$K_C \times$ Error	0%		1%
Reset Feedback	50%		50%
Output of First-order Lag	50%		50%
Controller Output	50%		51%
Valve Position		50%	

Now suppose that the temperature rises an additional 1 percent, just to the limit set point. Table 12-3 depicts this situation. Now, we are in a “don’t care” situation in which both controller outputs are the same, so the selector switch continues to pass the same values to the fuel valve as well as to the reset feedback ports.

Table 12-3. Tube Temperature Rise To Limit

Signal	Heater Outlet Controller	Selector Output	Tube Limit Controller
Set Point	40%		70%
PV	40%		70%
Error	0%		0%
$K_C \times$ Error	0%		0%
Reset Feedback	50%		50%
Output of First-order Lag	50%		50%
Controller Output	50%		50%
Valve Position		50%	

Again, suppose the temperature rises 1 percent above the limit. The three previous tables have shown a static situation, one that could remain indefinitely, but Table 12-4 is transitory. It depicts the situation right after the tube temperature rise. This is a transitory table because the first-order lags are unbalanced. For the tube limit controller, the output of the first-order lag will fall to match its input. This will force down the controller output, the signal to the valve, and the reset feedback to both controllers. This situation will be repeated as long as the tube temperature is above the limit. In other words, the tube limit controller is now behaving like a normal PI controller.

Table 12-4. Tube Temperature Rise Above Limit

Signal	Heater Outlet Controller	Selector Output	Tube Limit Controller
Set Point	40%		70%
PV	40%		71%
Error	0%		-1%
$K_C \times$ Error	0%		-1%
Reset Feedback	49%		49%
Output of First-order Lag	50%		50%
Controller Output	50%		49%
Valve Position		49%	

Consider the heater outlet temperature controller right after the condition depicted by Table 12-4. Two situations are occurring simultaneously:

- The output of the first-order lag will go down to match its input; this will tend to cause this controller output to go down.

- If the heater outlet temperature has been maintained at set point with a 50 percent valve position, it is logical to assume that with a lower fuel valve position the heater outlet temperature will go down. Thus, the error will go to a positive number; this will tend to cause this controller's output to go up.

Let us assume that with the abnormal condition, the tube limit controller stabilizes ($PV - SP$) when the signal to the valve has decreased to 45 percent. Furthermore, suppose that with that new valve position, the heater outlet temperature drops to 37 percent. (Recall that a valve position of 50 percent was required to maintain heater outlet temperature at its set point of 40 percent.) Table 12-5 depicts the new equilibrium condition.

Table 12-5. Tube Limit Temperature Controller In Control

Signal	Heater Outlet Controller	Selector Output	Tube Limit Controller
Set Point	40%		70%
PV	37%		70%
Error	3%		0%
$K_C \times$ Error	3%		0%
Reset Feedback	45%		45%
Output of First-order Lag	45%		45%
Controller Output	48%		45%
Signal to Flow Cont		45%	

Table 12-5 is again a static table. The control system could remain in this condition indefinitely, as long as the abnormal condition exists. The point is that the situation has reversed from that shown in Table 12-1, in which the outlet temperature was controlling but the tube limit controller was not winding up. The tube limit condition is now controlling. The heater outlet temperature is below set point, but the outlet temperature controller is not winding up. When the abnormal situation is cleared up, the steps above will be reversed: the control will revert to the heater outlet temperature controller.

Following this example carefully reveals the difference in behavior between this control scheme and the control scheme that used ordinary PI controllers (Figure 12-3). In particular, observe Table 12-3. The tube temperature has just reached the limit, and the tube limit controller is on the verge of overriding the other controller and assuming control of the valve. With ordinary PI controllers, the tube limit temperature controller would merely be on the verge of starting to unwind from its maximum output value. It would then have to unwind all the way below the other controller's output before it had any effect on furnace firing.

The foregoing discussion is but one example of override (selector) control. The opportunities for applying this technique are manifold. The characteristics of all such applications are as follows: in normal operation one controller is in control, but in abnormal operation another controller (or perhaps one of several other controllers) overrides. Often the goal is to protect process equipment, even at the sacrifice of normal process control.

Another type of application for override control is to automatically modify the control strategy between process startup and normal operation. In the pulp and paper industry, steam to a batch digester is initially controlled by a flow controller until the pressure rises to the operating point, then a pressure controller overrides and assumes control of the valve (see Figure 12-5).

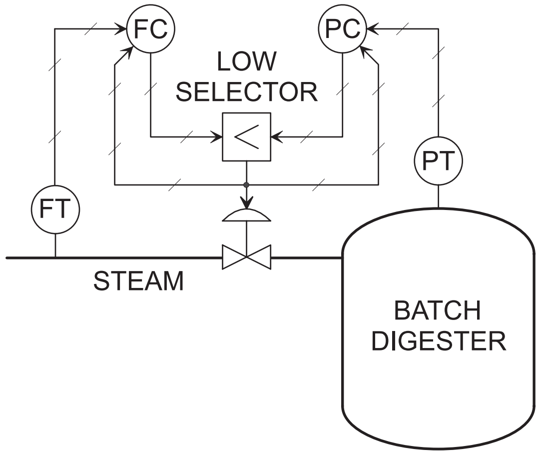

Figure 12-5. An Example of the Use of Override Control for Process Startup

For a generic process, suppose that the feed rate to a process unit is maintained constant by a flow controller. Further, suppose that the process unit uses some utility, for example, steam, as a part of the process. If the steam supply is limited, then the process controller may move the steam valve wide open. Rather than produce off-spec product, the normal procedure would be to reduce feed rate. This procedure can be automated by installing a "valve position controller" (see chapter 16) for the steam valve. Its set point would be near the maximum allowable valve position, say 95 percent. If the process controller drives the valve to that limit, the valve position controller will override the feed-flow controller to reduce feed rate (see Figure 12-6).

For a distillation column, an adverse condition known as "tower flooding" can occur under certain abnormal conditions. This is normally caused by excessive vapor flow within the column. It can often be detected by measuring the differential pressure across a section of the column. If a composition controller (or inferred composition controller, such as temperature) normally controls the steam to the reboiler, then a differential-pressure controller should override and reduce the steam flow. In this case, the selected signal is the set point for a secondary steam-flow controller, not the valve signal itself. The external reset feedback signal could be the flow controller set point. A preferred signal, however, would be the actual flow measurement. This would provide the correct ERF signal regardless of the manual/automatic status of the flow controller.

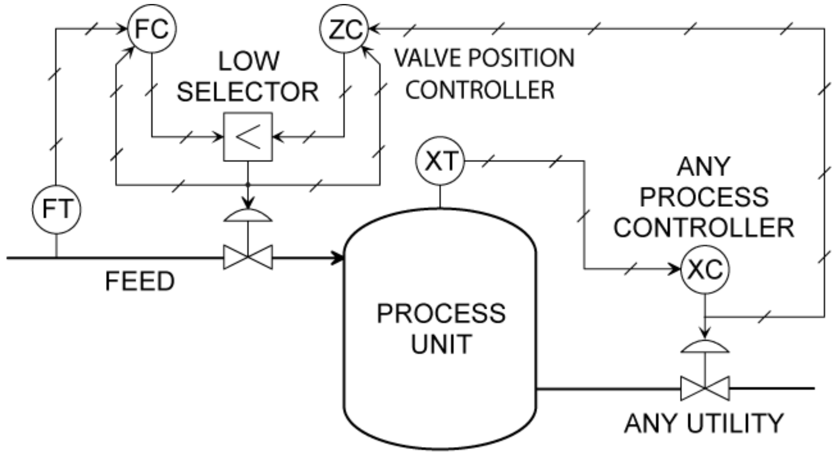

Figure 12-6. An Example of the Use of Override Control for Operating Near Process Limits

In a further level of sophistication, suppose that the temperature controller does not directly set the steam flow, but sets a required steam-to-column feed ratio. For a PI controller with ERF to function as a normal PI controller, its ERF signal must represent the same quantity as its output. (In Figure 12-4, each controller output represents a required valve position; the ERF also represents valve position.) Thus, for the distillation tower it would be incorrect to use a steam-flow signal, whether flow controller set point or actual steam flow, as the ERF to the temperature controller, since its output represents a required steam-to-feed ratio. The correct configuration would be to use the measured steam and feed rates to calculate the actual steam-to-feed ratio, and use this as the ERF for the temperature controller, as shown in Figure 12-7 (Ref. 12-1).

At a compressor station in the pipeline industry, both the suction and discharge pressures are monitored. Whichever pressure demands the lower compressor speed becomes the overriding controller, as shown in Figure 12-8.

For a multi-engine-compressor station, there may be limiting controllers that apply to the station as a whole as well as limiting controllers that apply only to individual engines. Engine speeds are the ultimate manipulated variables; these respond to the lowest of several controller outputs. Figure 12-9 shows controls for two engine-compressor sets. A typical station may have four or more compressors.

Overall station controls are suction pressure, discharge pressure, and station flow. The lowest of these controller outputs sets an upper limit for the speed of all the engines. Individual engine controls, such as torque and engine temperature, may further reduce the speed for any particular engine. In addition, the operator may set a maximum speed setting that overrides all of the other controls.

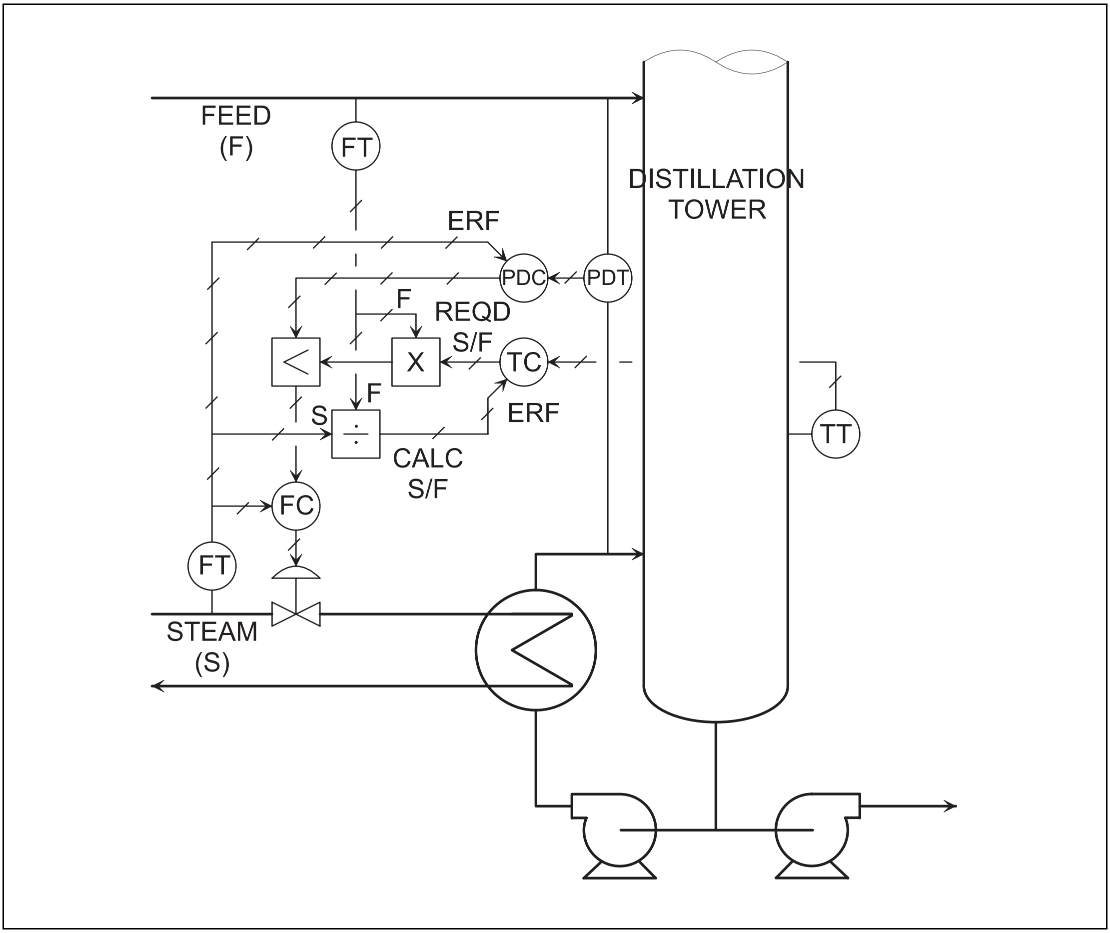

Figure 12-7. An Example of Combining Override, Ratio, and Feedforward Control

To prevent windup of the individual engine controls, conventional external reset feedback is applied to the torque and engine temperature controls. The operator speed setting, being simply a constant entry, cannot wind up; hence it does not require external reset feedback.

Determining an appropriate signal for external reset feedback for the station controllers presents an interesting situation. At any one time, the overriding station controller may be in control of one or more engine speeds. Suppose that the suction pressure controller output in Figure 12-9 is lower than any of the other station controls. Further, suppose that it is also lower than any of the engine controllers, including the operator speed setting, for engine #1. Then the suction pressure controller will be setting the speed for engine #1. Suppose at the same time that the output from one of the controllers from engine #2, say the torque controller, is lower than the suction pressure controller output. The torque controller will be in control of the speed of engine #2. The external reset feedback for the station controllers should be the *higher* of the individual engine speeds. In this case, it would be the signal to the engine #1 speed controller. With this configuration, none of the controllers will wind up regardless of whether an engine controller or a station controller is in control of any particular engine speed.

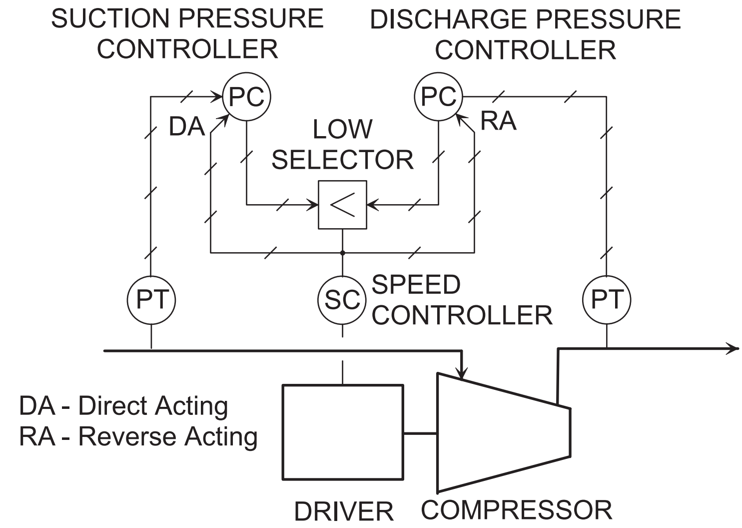

Figure 12-8. An Example of Override Control in the Pipeline Industries

The configuration described here has been recommended for a particular four-engine-compressor station. As far as the author knows, however, no existing implementation embodies this exact configuration.

(As a separate topic, consideration should be given to adjusting the gain of the station controllers, depending upon the number of engine speeds which can be manipulated by a station controller at any one time.)

These examples have presented a broad spectrum of applications for override control, ranging from very simple applications (Figures 12-2, 12-5, and 12-8) to complex configurations (Figures 12-7 and 12-9). From these examples, the reader should be able to draw inspiration in applying override control to other situations.

❖ OTHER METHODS OF IMPLEMENTATION

The discussion in the last section focused on the traditional configuration of override control and began with its development during the age of pneumatic controllers. Traditional override control persists in digital form to the present day with some commercial systems. Other systems, however, use some different form of implementation. The following paragraphs present these other methods of implementation, along with a critique of each.

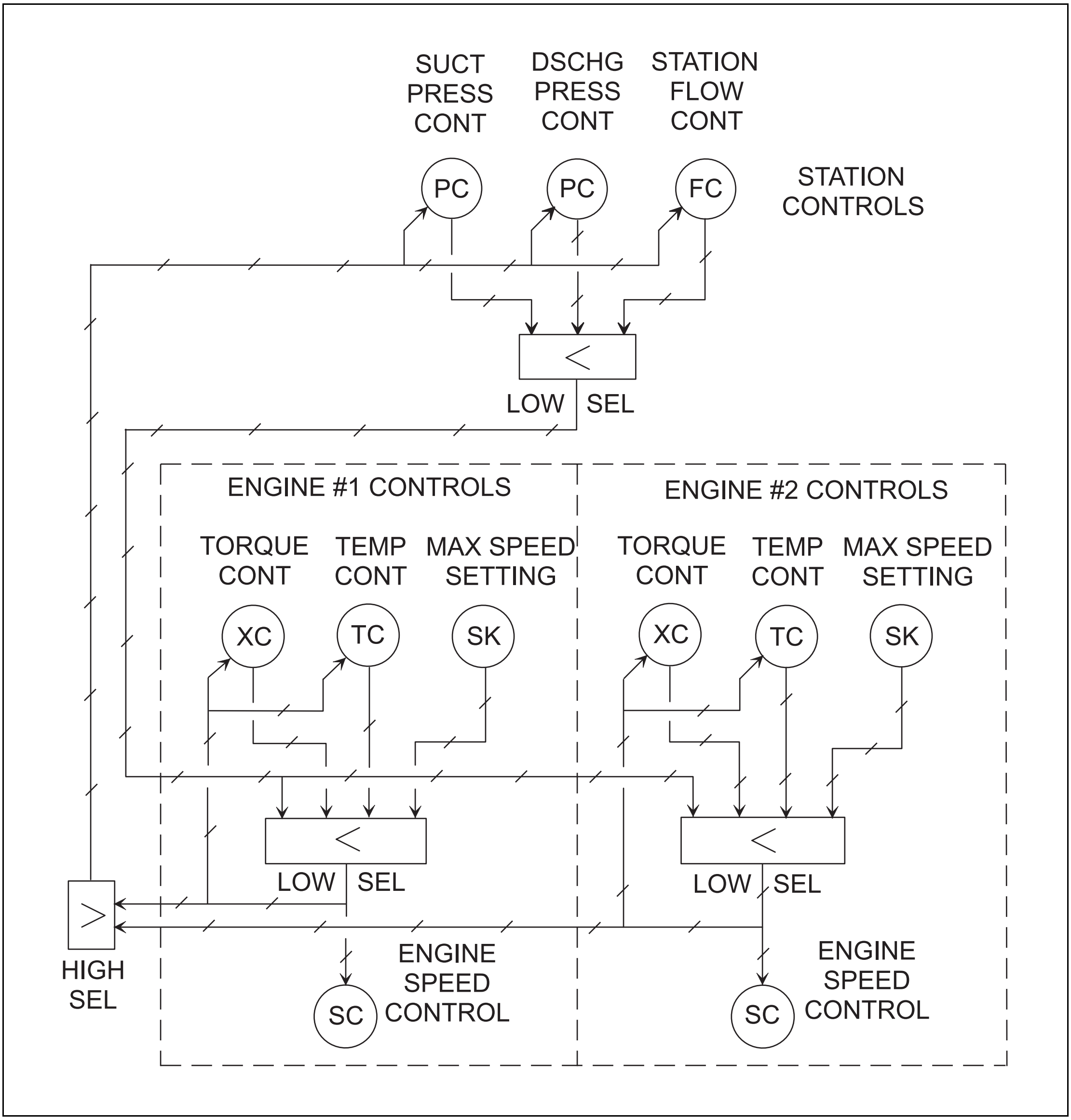

Figure 12-9. A Complex Application of Override Control for a Multiple-Compressor Pipeline Station

♦ "Pass-through" Method

In the traditional configuration of override control, the nonselected controller tracks the output of the other controller through a first-order lag whose time constant is the integral time of the controller. Thus, the tracking speed is governed by the required integral time. Lipták (Ref. 12-2, Section 1.17) states that this can be a problem when two or more controllers have significantly different integral times. After extensive simulation study, the author did not find this to be true, and in fact found the traditional configuration performed better than any proposed solutions when the controller integral times differed significantly.

One of the proposed solutions has been to avoid the tracking speed problem by having external logic detect when a controller is a nonselected controller. Then, the output of the selected controller is immediately "passed through" (or bypasses) the reset lag, as shown in Figure 12-10. The output of the nonselected controller, then, will always be the output of the selected controller, plus gain multiplied by error of the nonselected loop.

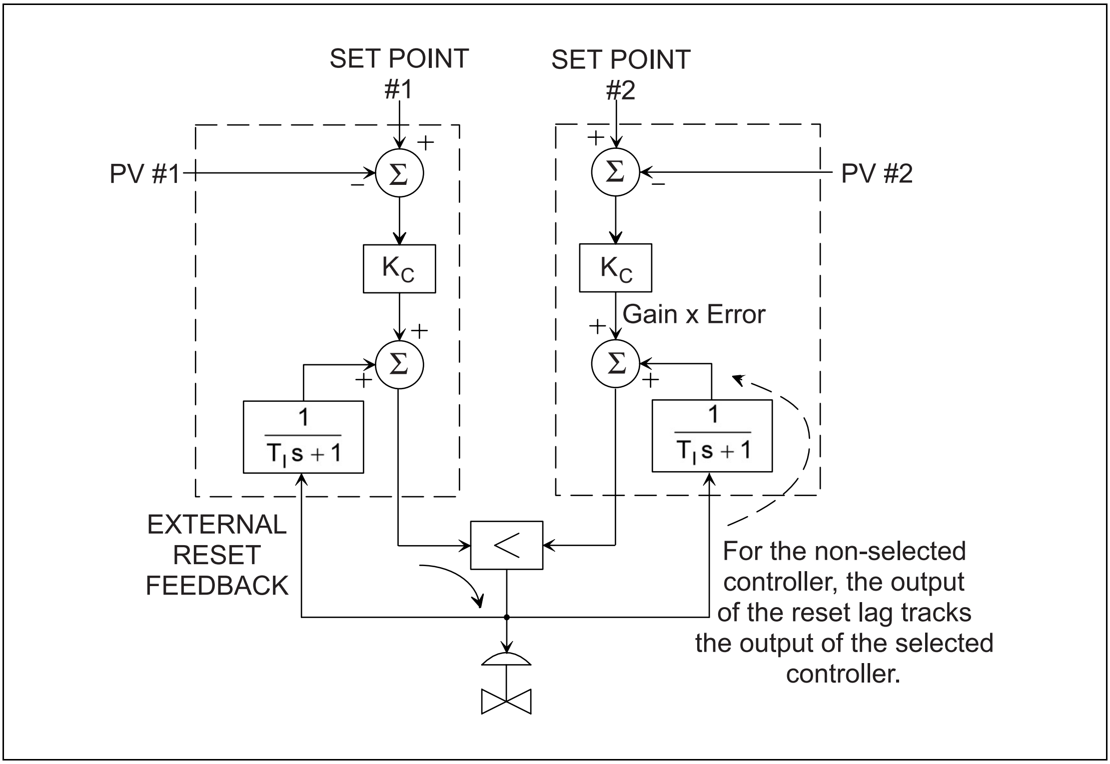

Figure 12-10. Block Diagram of the "Pass-through" Implementation Method for Override Control

Note that with the traditional method of implementation, there is never a jump in either controller output due to switchover. With the "pass-through" method, however, at the moment when one controller switches over to another, the output of the nonselected controller may jump, depending on the magnitude of its error. If the jump is away from the other controller output, then this jump is of no consequence. However, if the jump is in the other direction, it can again override the other controller and move the valve to an unwarranted position, as the Figure 12-11 and the detailed analysis that follows illustrate.

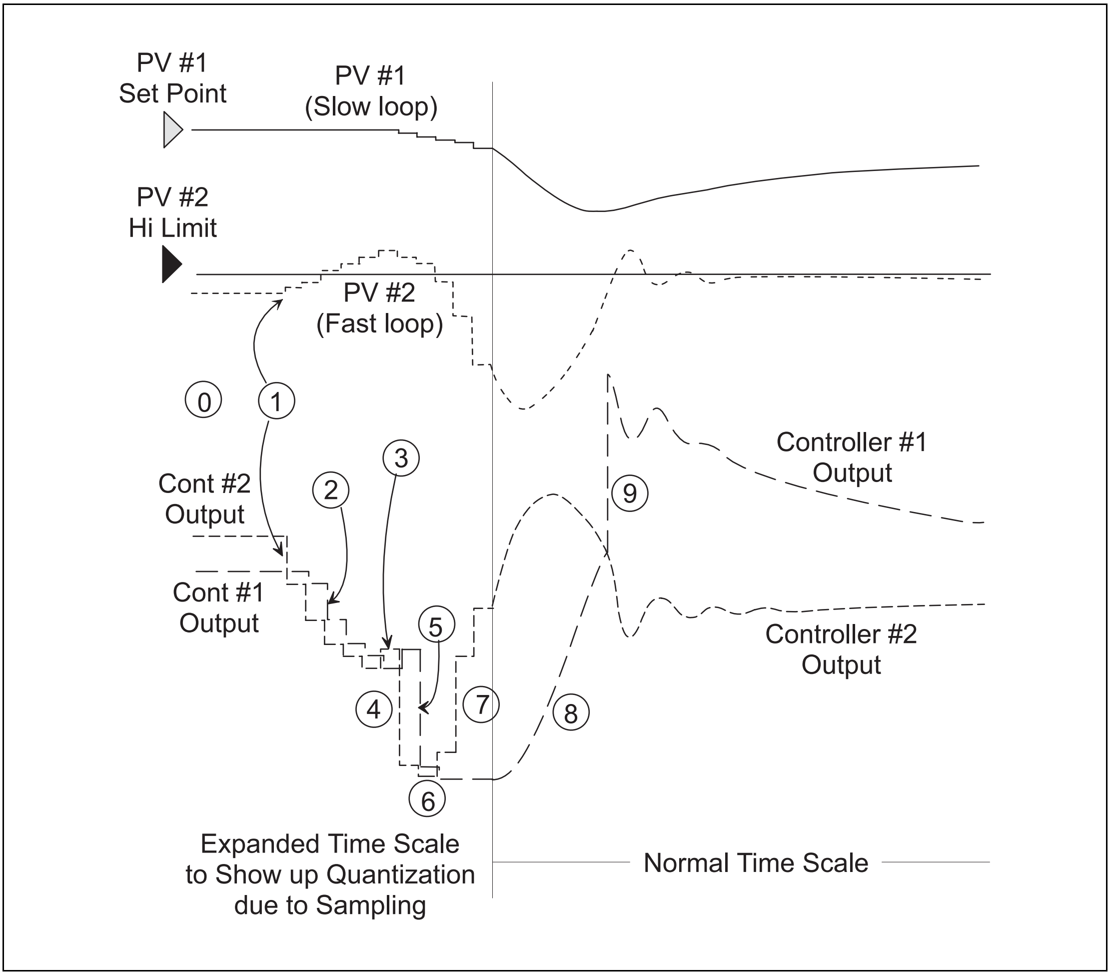

Figure 12-11. Possible Behavior of "Pass-through" Implementation Method

Detailed Analysis of Figure 12-11

(0) Initial Conditions: PV #1 (slow loop) is at set point. PV #2 (fast loop) is below its limit value. Controller #2 output is above Controller #1 output, thus Controller #1 is the selected controller. Both controllers are reverse-acting.

(1) A sudden disturbance causes PV #2 to rise and the output of its controller to fall below the output of Controller #1. Thus, Controller #2 is now the selected controller.

(2) Controller #2 output continues to fall to correct the deviation of PV #2. The output of Controller #1 tracks Controller #2's output, but with one sample time delay. (Actually, Controller #1's output is Controller #2's output plus Gain × Error of loop #1. However, since loop #1 is a slow loop and was previously on set point, then for the example its error is presumed to still be zero.)

(3) At some point, the corrective action to PV #2 will cause its controller output to reverse, then become greater than Controller #1's output. Controller #1 is now the selected controller (for one scan cycle).

(4) Since Controller #2 is now the nonselected controller, its output will be forced to Controller #1 output plus Gain × Error of loop #2. In the example, PV #2 is still above its limit, so the error will be negative. Thus, the output of Controller #2 will make a significant decrease, again overriding Controller #1 and becoming the selected controller.

(5) Controller #1 output tracks (delayed by one sample period) the output of Controller #2, plus Gain × Error of loop #1. In the example, the error of loop #1 is still small, so Gain × Error is essentially zero.

(6) The effect of a significant decrease in Controller #2 output causes PV #2 to change fairly rapidly. This in turn causes an increase in the output of Controller #2, so it becomes greater than Controller #1. Controller #1 is now the selected controller.

(7) Controller #2's output tracks Controller #1's output, plus Gain × Error in loop #2. If by now PV #2 is below its limit, then the error is positive, so there will be a jump upward in Controller output #2, away from Controller output #1.

(8) The valve is now significantly depressed from the condition required by PV #1 or by PV #2. It causes a major upset to PV #1 that is corrected slowly; hence, the output for Controller #1 (the selected controller) increases gradually.

(9) When the output for Controller #1 crosses over the value for the output for Controller #2, the output for Controller #2 is selected. Controller #1's output now makes a jump upward, since PV #1 is below its set point, hence there is a positive error.

This nine-part analysis was observed in a simulation, and shows what *can* happen under adverse circumstances. Under milder circumstances, such as a slow load disturbance on the faster loop, these conditions would probably not be observed. Even under milder conditions, however, noisy systems can adversely affect the logic that determines the selected and nonselected controllers, thus causing unpredictable results.

As we noted earlier, with the traditional method of implementation, there is never a jump in either controller output. Thus, in summary, we can say that the "pass-through" method may either perform comparably or not perform as well as the traditional method. It will never perform better.

♦ Forced Manual for the Nonselected Controller

Some systems force the nonselected controller to the equivalent of the manual mode, and depend on the bumpless transfer feature to prevent windup. If one could guarantee that a switchover would always be done with essentially zero error in the loops, then the performance of this method would be comparable to the traditional method. If there is an error in the nonselected loop, then this method may demonstrate behavior problems similar to the "pass-through" method.

♦ Velocity (Incremental) Mode Control Algorithms

The traditional implementation of override control presented earlier utilizes position-mode PID algorithms. Hence, the final signal to the valve is used for external reset feedback. If the PID control is implemented with a velocity-mode rather than position-mode control algorithm (see chapter 5), then other considerations must be made.

In one of the early computer systems, a velocity-mode PID algorithm was used. In this system, both controllers (assuming that only two controllers participated in the override scheme) calculated a "Δm", or the amount each controller wanted to move the valve. The implementation philosophy was that the controller that should win was the one that wanted to move the valve the greater extent toward closed. The selector switch chose the minimum (algebraic) Δm and applied it to the present position of the valve. For example, if one controller wanted to close the valve by 2% (Δm = -2) and the other controller was presently on set point and did not want the valve to move (Δm = 0), then the Δm of -2 was selected, and the valve decreased in position by 2% (see Figure 12-12).

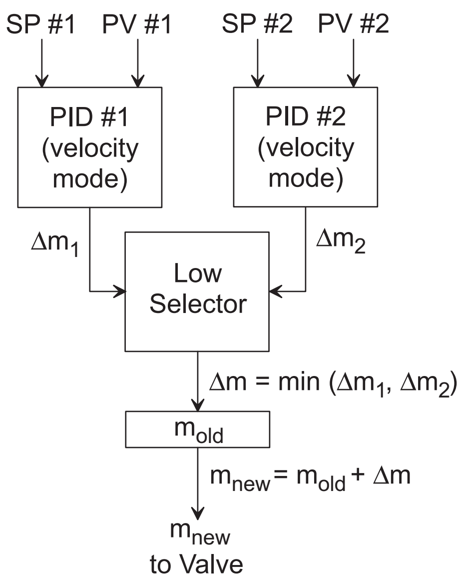

Figure 12-12. Implementation of Override Control with Velocity-mode Algorithms (Not Recommended)

One particular problem observed with this system occurred with a bottoms-stream composition controller on a distillation tower and a tower differential-pressure (inferential of tower flooding) controller, both manipulating a reboiler steam valve. The composition controller, a slow loop, was normally in control while the ΔP controller, a fast loop, normally ran below its limit. Suppose a sudden disturbance, such as a slug of light component in the feed, caused the ΔP to increase. The ΔP controller overrode and decreased the position of the steam valve. Then, if the disturbance rapidly disappeared, the ΔP controller's output became positive (Δm's > 0), but if the composition controller had not yet been upset, its Δm's were zero. Thus, there was no way to recover the former valve position until there was an upset in the column composition. Then, the composition controller recovered the valve position.

With the traditional method of implementation, if a sudden disturbance causes the faster loop to override, then the disturbance disappears and the valve will very quickly recover to the original position, without significant upset to the slower loop.

♦ Pseudo-Velocity Method

One commercial system uses a velocity-mode PID algorithm, then adds the incremental change (Δm) of each controller to a register that represents the previously desired output of that controller. The selector is, in essence, operating between desired positions, not incremental changes. Logic within the selector then adjusts the output of the nonselected controller so it is equal to the output of the selected controller, plus the Gain × Error of the nonselected loop (see Figure 12-13). In essence, then, this system should have the behavior of the "pass-through" method we described earlier.

♦ Selection Based on Error

Other implementers have suggested that the selection criterion be based on the loop errors rather than the loop outputs. The lowest (algebraic) or highest of the error signals, depending on the application, would then be passed to a common PID algorithm. The advantage of this is that there is never a bump in the controller output, as a result of the switchover from one error signal to another. Its disadvantage, however, is that a single set of tuning parameters applies to all of the affected process variables. If there is a slow loop and a fast loop, then a single set of parameters will have to suffice for both.

The author is familiar with a custom implementation that circumvented this problem. In addition to error selection, separate logic determined which error signal was selected and obtained a unique set of tuning parameters from a table for that particular process variable. In essence, this was scheduled tuning. This system reportedly worked quite satisfactorily since the process variables were all relatively noise-free. Furthermore, the disturbances were relatively slow, so all the error signals were near to zero at the time of switchover. Had the "bumpless tuning" provisions described in chapter 4 been utilized, the system would have been immune to problems of switchover even with significant error in the loops.

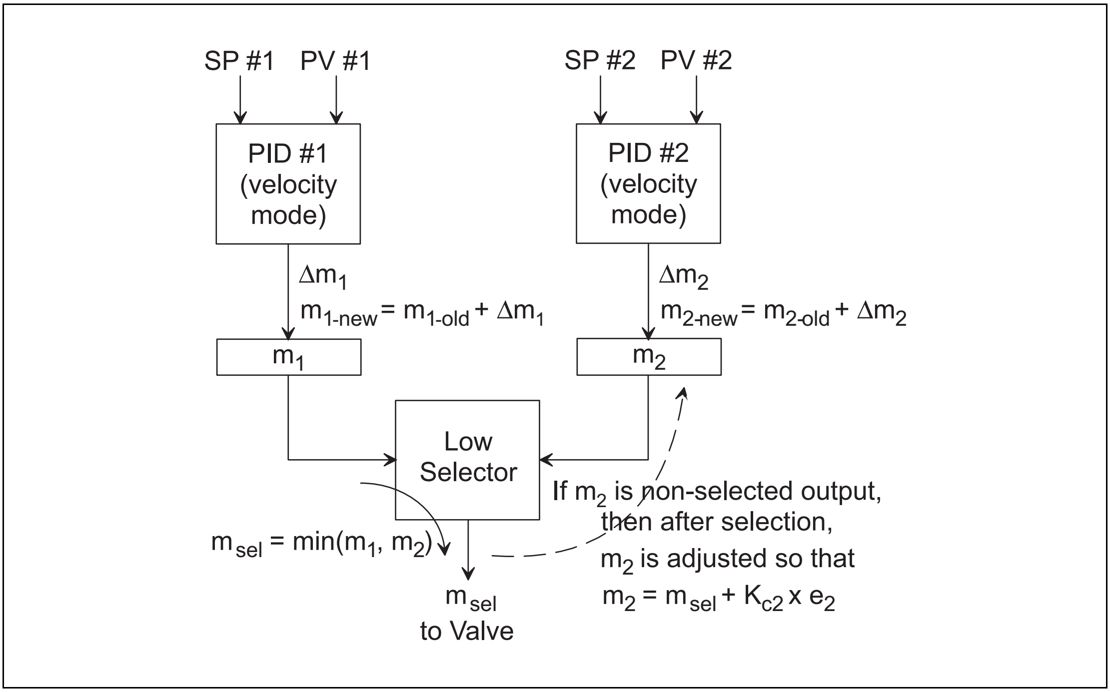

Figure 12-13. Another Implementation of Override Control with Velocity-mode Algorithms

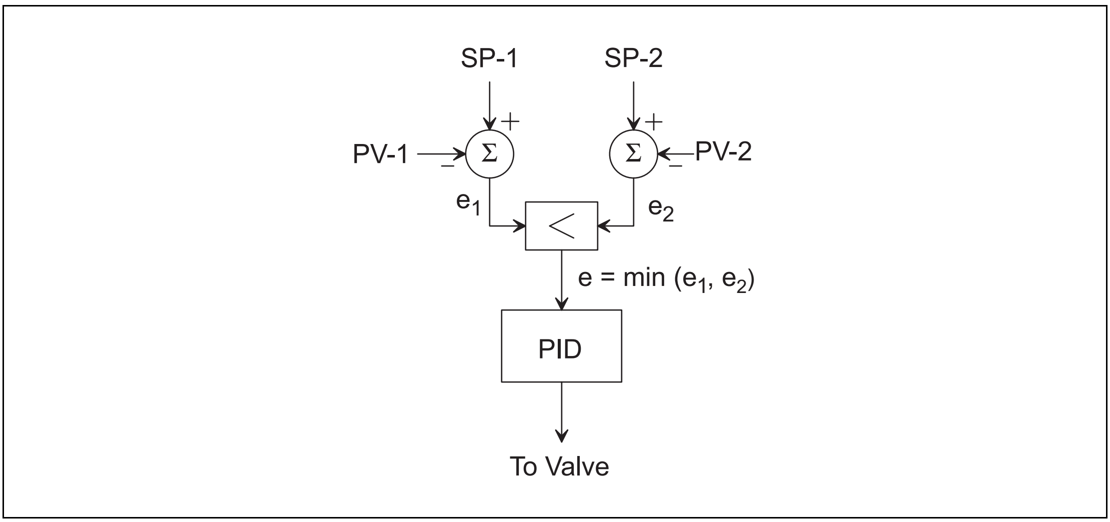

Figure 12-14. Override Control Based on Selection of Errors Rather Than Controller Outputs

❖ OVERRIDE CONTROL USING FOUNDATION™ FIELDBUS

The Fieldbus Foundation standard (Ref. 12-3) defines a control selector (CS) function block. This block is in the control class; hence it supports the back-calculation procedure and is intended for use in override control strategies. The CS block provides for up to three inputs from other control-class blocks, such as PID. (Nonconfigured inputs are ignored.) The block also provides three BKCAL_OUT signals, one for each of the three inputs. The block can be configured as either a high select or a low select. In the automatic mode, the lower (or higher) of the selected input signals is passed to the output. The status returned to the *nonselected* PID via the back-calculation link is "not selected," and the value is the CS output value, which is the same as the selected controller output. The nonselected controller's output is then made equal to this value. In other words, it is equated to the selected controller's output.

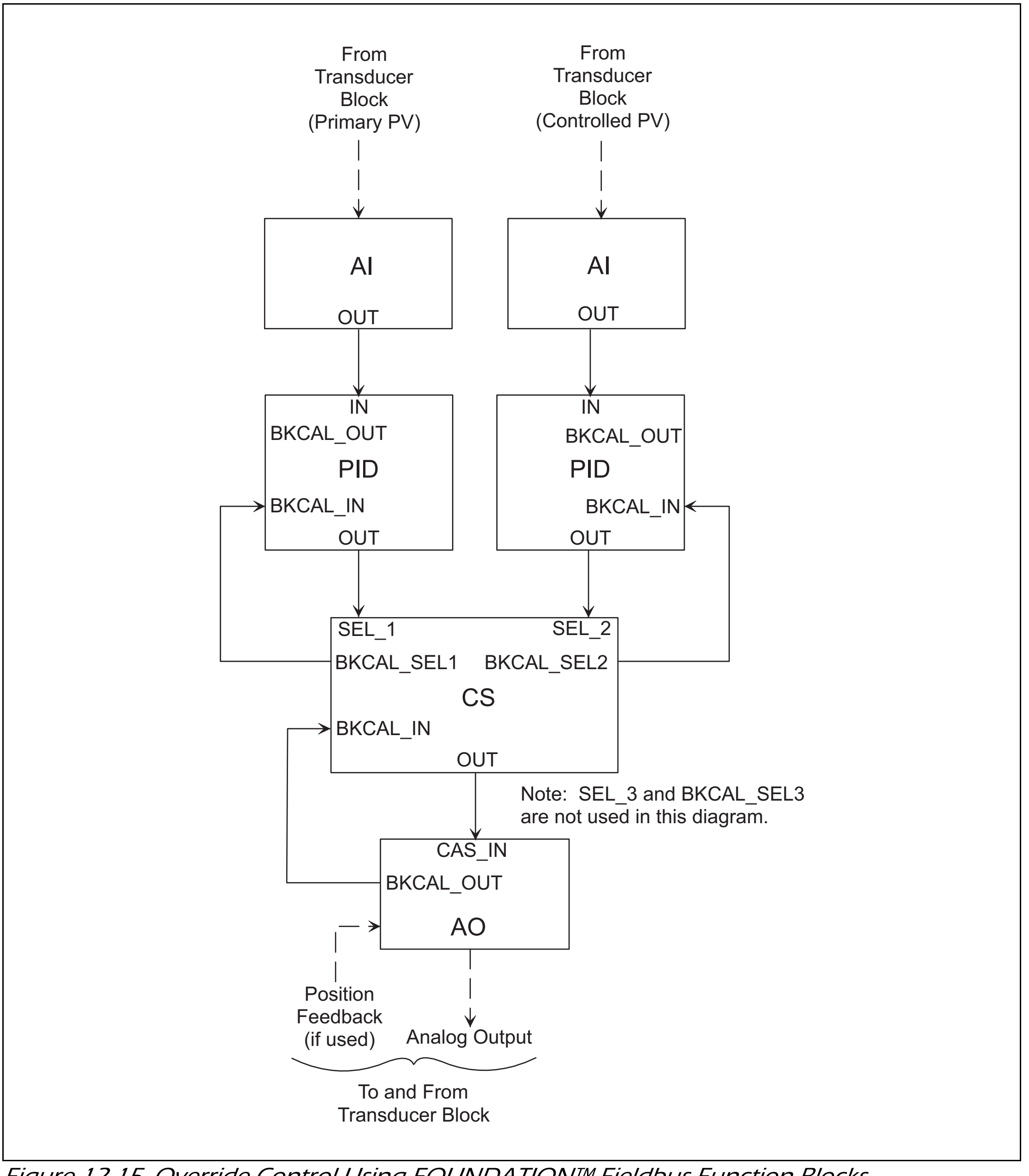

Figure 12-15. Override Control Using FOUNDATION™ Fieldbus Function Blocks

To analyze the behavior of this control scheme, suppose that the CS block is configured for low selection and that the selected controller's PV is at SP. The deviation of the nonselected controller must be such that it is calling for a higher valve position; otherwise, it would have been selected. If process conditions do not change appreciably before the next calculation cycle for each of these blocks, the selected controller will calculate the same valve position as before, but the nonselected controller will again request a higher valve position. The previously selected controller's output will again be selected, and the scenario will repeat.

Suppose, however, that process conditions change, so that each of the controllers requests a new valve position. The CS block will then select the lower of the two new valve positions. Since the controller outputs had previously been forced to be equal, the lower valve position is determined by whichever controller most wants to increment the valve toward closed (or least wants to increment the valve toward open).

In essence, the behavior of this system appears to be similar to the velocity (incremental) mode system we described previously. It may therefore be subject to the same potential problems if the dynamics between the controller output and the two process variables differ greatly, that is, if one PV responds quite rapidly and the other quite slowly.

Some manufacturers may not support the CS block.

❖ REFERENCES

12-1. F. G. Shinskey. *Distillation Control for Productivity and Energy Conservation*, 2d ed., McGraw Hill, Inc., 1984, p. 217.

12-2. Béla G. Lipták, ed. *Instrument Engineers' Handbook: Process Control,* 3d ed., Chilton Book Co., 1995, p. 117.

12-3. Fieldbus Foundation, *Foundation Specification: Function Block Application Process, Document FF-891, Part 2.*

13

CONTROL FOR INTERACTING PROCESS LOOPS

Control loops are said to interact when movement of the final control element of one loop affects not only its own process variable but the process variable of one or more additional control loops as well. If the corrective action by the controllers in the other loops then affects the process variable of the initial loop, an endless cycle of interaction will occur between the loops.

Numerous instances of control loop interaction can be found in industrial process control. Suppose that both the overhead and bottom composition of a distillation column are to be controlled by manipulating the reflux and reboiler steam flow rates. An increase in reflux rate will lower the impurity of the overhead composition and at the same time will increase the amount of heavy component recovered at the bottom of the tower. An increase in reboiler steam rate will lower the impurity of the bottom composition. At the same time, by forcing more of the lighter component to the top of the tower, it will increase the recovery of the lighter component in the overhead stream. The composition control loops in this application, whether using analytical instruments or inferred (e.g., temperature) measurements, may experience severe interaction.

In the heating, ventilating and air conditioning (HVAC) industry, simultaneous control of temperature and humidity is often desired. It is especially critical in clean rooms and other high-precision environmental control applications. If a heating coil is used to control temperature and a cooling coil for dehumidification, severe interaction may occur between the control loops.

For the sake of simplicity, we will limit our discussion to the interaction of two loops. Certainly, there are instances where mutual interaction occurs between more than two loops. Situations where interaction occurs between two loops are probably much more numerous, and the techniques presented here for handling two loops can be extended to the less frequent cases of multiple loop interaction.

There are several methods for coping with interacting process control loops. Perhaps the most universally used method is simply to "detune" one or more of the loops. If one loop has a higher priority than the other, the gain of the lower priority loop can be decreased and the integral time lengthened. This will tend to minimize the interaction between the loops. The penalty for using this approach is that the lower-priority process variable may experience considerable variation.

The first method that warrants our consideration is the judicious pairing of the process variable controllers with the final control elements. After discussing this so-called variable pairing we shall consider a method called *decoupling* in the section that follows. Decoupling involves placing additional intelligence within the loops to compensate for the coupling of the loops through the process.

❖ VARIABLE PAIRING

The interaction of control loops is somewhat similar to the problem of calibrating an instrument. Many instrument technicians have experienced the following situation when calibrating a transmitter or recorder. First, a minimum signal is applied to drive the instrument down scale, then an adjustment labeled "zero" is varied until the instrument reads zero. A full-scale signal is then applied to drive the instrument up scale and an adjustment labeled "span" is varied until the instrument reads properly at the upper end of the scale. Then the minimum signal is reapplied to check zero, and often the zero setting will have to be readjusted. Another check using the maximum signal shows that the span should be readjusted.

The phenomenon just described occurs because the zero and span adjustment knobs are interacting. Sometimes a technician may wonder if the adjustment knobs have been mislabeled. Perhaps the zero knob should be used at the upper end of the scale, and the span knob should be used at the low end. This is an exercise in variable pairing.

Consider the process control example shown in Figure 13-1 (taken from Shinskey [Ref. 13-1]). The objective is to blend two pure streams containing ingredients A and B, respectively, and to maintain a specified total flow rate and specified composition in the mixed stream. If there are two controllers, a flow controller and a composition controller (the method of measuring the composition is irrelevant to this discussion), then the question arises: which is the preferred way to connect the controllers to the two valves—flow controller to valve A and composition controller to valve B, or vice versa?

Of course, one could ask, "Why not use a ratio control technique, and let the composition controller set the ratio?" That is a valid question, but adding ratio control would mean inserting additional intelligence into the loops to achieve decoupling. This procedure is beyond our present scope: the proper pairing of the variables.

Obviously, moving valve A alone affects both the total flow and the composition, as does moving valve B. Hence, the control loops will interact. A bit of reflection, however, leads us to decide that if the required composition of ingredient A is low (say, 5% ingredient A, 95% ingredient B) then the preferred control strategy is to connect the flow controller to valve B, from which most of the flow comes, and to connect the composition controller to valve A so as to dilute the stream with a small portion of ingredient A. If the required composition of ingredient A is high (say, 95% A, 5% B), then the controller-valve connections should be reversed.

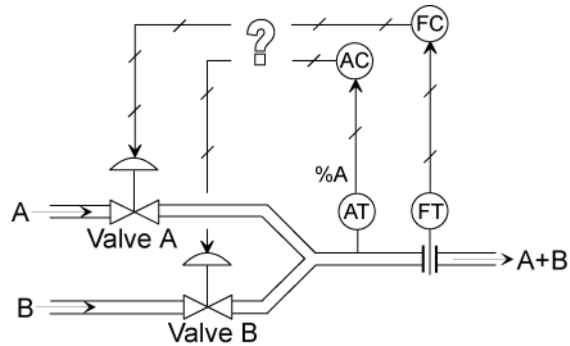

Figure 13-1. A Variable-pairing Problem

In this simple case, the proper pairing of the controllers and valves is obvious from common-sense inspection. What about situations in which common-sense inspection does not yield the proper pairing? In these cases, a more formal procedure is needed to indicate the preferred pairing. Such a procedure, called *relative gain analysis*, was introduced by Bristol (Ref. 13-2).

Relative gain analysis is a means of using the process steady-state gains to determine the preferred pairing of measurements and valves into control loops. Recall from previous chapters that the steady-state process gain is defined as follows:

$$\frac{\Delta \text{ Measurement}}{\Delta \text{ Controller Output}} .$$

Between a particular valve and measurement, however, the apparent steady-state gain when the other loops are in manual may be different than when they are in automatic. Consider Figure 13-2, which shows a two-loop interacting process.

Suppose both control loops are in manual with the process variable at set point. Then suppose a change is made to the output of controller #1, that is, to valve #1. This will have a *direct effect* on measurement #1. A change in the position of valve #1 will also have an effect on measurement #2 because of the interaction through the process. But since controller #2 is in manual, the position of valve #2 is unchanged; hence, there is no further effect on measurement #1. Using the definition of process gain, we can determine the process gain between valve #1 and measurement #1:

$$\left.\frac{\Delta \text{ Measurement \#1}}{\Delta \text{ Controller Output \#1}}\right|_{\text{Loop \#2 in Manual}} .$$

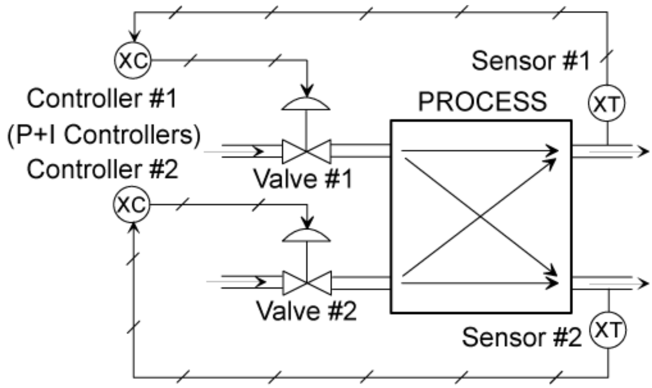

Figure 13-2. Interacting Control Loops

Now, repeat this test, but this time suppose that controller #2 is in automatic with the process variable at set point. Make a change in the output of controller #1; this has the expected direct effect on measurement #1. Now, however, we must consider the effect of this change on measurement #2. Controller #2 senses this deviation from set point and manipulates its output (i.e., valve #2) to return its measurement to set point. (It is assumed that both controllers have integral action.) This change in valve #2 will have a further effect on measurement #1 because of the process interaction. Since this further effect occurs only as a result of the interaction with the other process loop and only when the other controller is in automatic, we call this an *indirect effect*.

The ultimate effect of our change in valve #1 is the sum of the direct and indirect effects. After the process has stabilized, we can determine the steady-state gain between valve #1 and measurement #1 under this condition:

$$\left.\frac{\Delta\ \text{Measurement \#1}}{\Delta\ \text{Controller Output \#1}}\right|_{\text{Loop \#2 in Automatic}}.$$

The *relative gain* is the ratio of these two steady-state gains.

$$\text{Relative gain} = \frac{\begin{array}{c}\text{Apparent process gain with}\\ \text{all other loops in manual}\end{array}}{\begin{array}{c}\text{Apparent process gain with}\\ \text{all other loops in automatic}\end{array}} \tag{13-1}$$

For an n-input, n-output process, there will be n^2 relative gains. These can be arranged in an $n \times n$ relative gain array (RGA), as shown in Equation 13-2, where λ_{ij} represents the relative gain between the i^{th} output (measurement), x_i, and the j^{th} input (valve), m_j:

$$\begin{array}{c|cccc} & m_1 & m_2 & \cdots & m_n \\ \hline x_1 & \lambda_{11} & \lambda_{12} & \cdots & \lambda_{1n} \\ x_2 & \lambda_{21} & \lambda_{22} & & \lambda_{2n} \\ \vdots & \vdots & & \ddots & \vdots \\ x_n & \lambda_{n1} & \lambda_{n2} & \cdots & \lambda_{nn} \end{array} . \qquad (13\text{-}2)$$

If we have determined all the relative gains except one in every row and every column, then the remaining relative gains can readily be calculated using the fact that

$$\text{For any row,} \qquad \sum_{j=1}^{n} \lambda_{ij} = 1 \qquad (13\text{-}3)$$

$$\text{For any column,} \qquad \sum_{i=1}^{n} \lambda_{ij} = 1 \qquad (13\text{-}4)$$

Figure 13-3 depicts a 2×2 process in which numerical values have been indicated for the steady-state gains for each of the process paths. (In an actual application, these steady-state gain values could be determined in one of several ways, including running process tests, performing tests on a simulation model, or evaluating partial derivatives of analytical process equations.) We will use these values to determine all the elements of a relative gain array.

First, assume that a PI controller is connected between measurement #2 and controller #2. This does not necessarily indicate the optimum pairing of variables; it is merely an artifice to help us in the thought process toward determining the relative gain.

Assume that a change is made to valve #1. Consider this as a change of 1 unit, that is, $\Delta m_1 = 1$. The *direct effect* on measurement #1 will be a change of 1 unit .[1]

$$\begin{aligned} \Delta x_{1(direct)} &= k_{11} \times \Delta m_1 \\ &= 1 \times 1 \\ &= 1 \end{aligned}$$

If the PI controller between measurement #2 and input #2 is in manual, then

$$\text{Process gain with other loop in manual} = \frac{\Delta x_{1(direct)}}{\Delta m_1}$$

$$= k_{11} . \qquad (13\text{-}5)$$

1. We will see later that the size of the step change cancels out, so its size is immaterial.

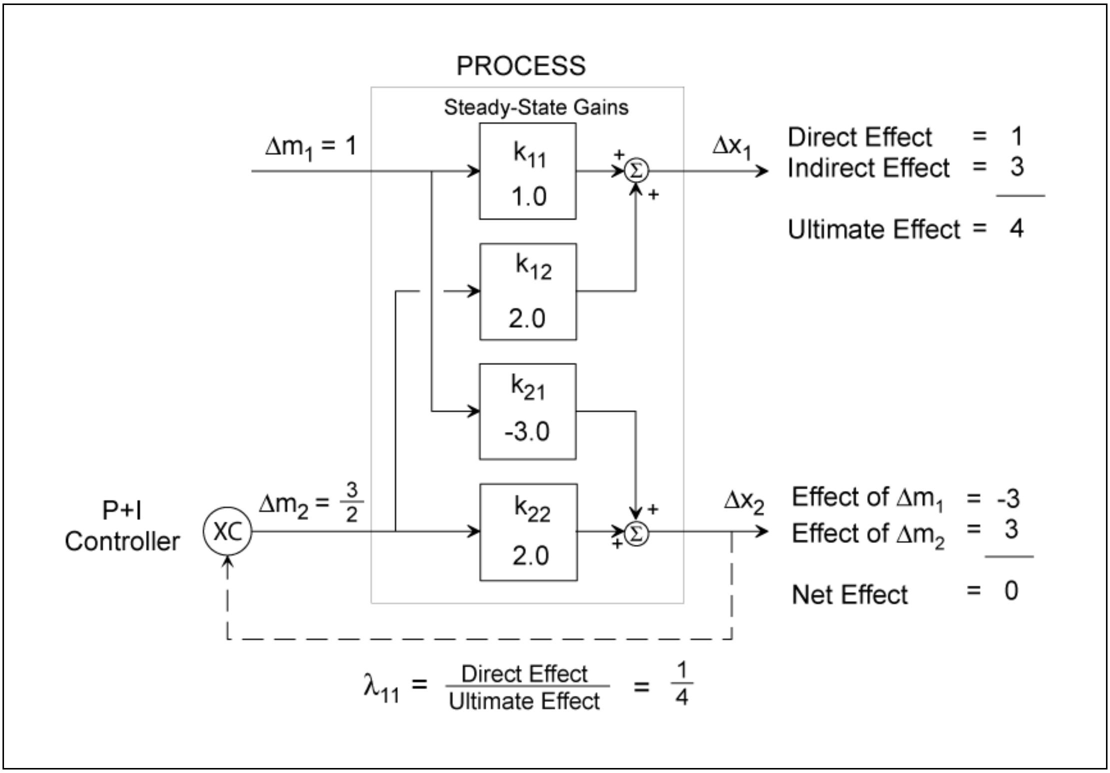

Figure 13-3. Example of Determination of One Relative Gain Element

The change in m_1 also changes measurement #2 by -3 units. That is,

$$\begin{aligned} \Delta x_2 &= k_{21} \times \Delta m_1 \\ &= -3 \times 1 \\ &= -3 \end{aligned}$$

If the PI controller between measurement #2 and input #2 is in automatic, then to compensate for the change to measurement #2, controller #2 must change its output so the total effect on x_2 is zero. The required change for controller #2 output is

$$\begin{aligned} \Delta m_2 &= -\frac{k_{21}}{k_{22}} \Delta m_1 \\ &= -\frac{-3}{2} \times 1 \\ &= \frac{3}{2} \end{aligned}$$

Considering both the change in m_1 and the change in m_2, the total change to x_2 is

$$\begin{aligned}\Delta x_2 &= k_{21}\times\Delta m_1 + k_{22}\times\Delta m_2 \\ &= (-3)\times 1 + 2\times\frac{3}{2} \\ &= 0\end{aligned}$$

Thus, measurement #2 is returned to its initial value. However, the change in m_2 has an additional effect on measurement #1. This *indirect effect* is given by

$$\begin{aligned}\Delta x_{1(indirect)} &= k_{12}\times\Delta m_2 \\ &= 2\times\frac{3}{2} \\ &= 3\end{aligned}$$

Thus, the *ultimate effect* on x_1 of our change in m_1 is the sum of the direct and indirect effects, or 4 units. In general,

$$\begin{aligned}\Delta x_{1(ultimate)} &= \Delta x_{1(direct)} + \Delta x_{1(indirect)} \\ &= k_{11}\times\Delta m_1 + k_{12}\times\Delta m_2 \\ &= k_{11}\times\Delta m_1 + k_{12}\times\left(-\frac{k_{21}}{k_{22}}\right)\Delta m_1 \\ &= \frac{k_{11}k_{22} - k_{12}k_{21}}{k_{22}}\Delta m_1\end{aligned}$$

Hence: Process gain with other loop in auto $= \dfrac{\Delta x_{1(ultimate)}}{\Delta m_1}$

$$= \frac{k_{11}k_{22} - k_{12}k_{21}}{k_{22}}. \qquad (13\text{-}6)$$

Using Equations 13-5 and 13-6 in Equation 13-1 gives the relative gain between x_1 and m_1:

$$\lambda_{11} = \frac{k_{11}k_{22}}{k_{11}k_{22} - k_{12}k_{21}} \qquad (13\text{-}7)$$

(Since Δm_1 is common to both the direct effect and the ultimate effect, we can also say that

$$\lambda_{11} = \frac{\text{Direct Effect}}{\text{Ultimate Effect}} .$$

Using the "k" values given in Figure 13-3, we see that

$$\lambda_{11} = \frac{1}{4} .$$

Then, by using the fact that the sum of the relative gains in any column or any row of the relative gain array must always be 1 (see Equations 13-3 and 13-4), we can now write the complete relative gain array:

	m_1	m_2
x_1	1/4	3/4
x_2	3/4	1/4

Note that relative gains are dimensionless; hence, they are not affected by process variable scaling.

Now that we have calculated the relative gains, let us consider their significance for guiding us in the pairing of controllers and valves, which was our initial objective.

If all the relative gains are between 0 and 1, then the variable pairing that produces the larger relative gain is the preferred pairing. In the example, pairing controller #2 with valve #1 and controller #1 with valve #2 produces relative gains of 3/4 in each loop. Suppose that the loops were connected in this way and that both controllers were in automatic. If we wished to increase measurement #1 by 4 units, we would raise the set point of controller #1 by 4 units. Roughly speaking, controller #1, connected to valve #2, would do 3/4 of the control effort and, due to the interaction between the loops, the other controller would do the remaining 1/4 of the control effort. If the loops were connected the other way (controller #1 - valve #1; controller #2 - valve #2), controller #1 would do 1/4 of the effort in raising its own measurement by 4 units, and it would depend on the other controller to do the remaining 3/4 of the effort. This discussion has illustrated the fact that in this example the first connection, with the relative gains of 3/4, is the preferred pairing of variables. It minimizes the interaction between the loops.

The situation in which all the relative gains are between 0 and 1 is a fortuitous one because the control loops aid each other. The interaction can be much more severe if some of the relative gains are greater than one. In that case, Equations 13-3 and 13-4 require that some of the other relative gains must be less than zero.

Equation 13-6 gives the relative gain for a special but very useful case for n=2, that is, for a 2×2 process. For the general case, suppose we have an $n \times n$ matrix of steady-state gains:

$$\mathbf{K} = \begin{bmatrix} k_{11} & k_{12} & \cdots & k_{1n} \\ k_{21} & \ddots & & \vdots \\ \vdots & & \ddots & \vdots \\ k_{2n} & \cdots & \cdots & k_{nn} \end{bmatrix}$$

Let **Q** be an $n \times n$ matrix that is the inverse of **K**, if **Q** exists. (If **Q** does not exist, then the number of independent process variables is not equal to the number of manipulated variables.) **Q** can be given by its elements as follows:

$$\mathbf{Q} = \mathbf{K}^{-1} = \begin{bmatrix} q_{11} & q_{12} & \cdots & q_{1n} \\ q_{21} & \ddots & & \vdots \\ \vdots & & \ddots & \vdots \\ q_{2n} & \cdots & \cdots & q_{nn} \end{bmatrix}$$

The relative gain between x_i and m_j is given by

$$\lambda_{ij} = k_{ij} q_{ji}$$

Note the reversal of subscripts for the "q" term.

The λ_{ij} relative gains can be arranged in an array as shown in Equation 13-2. In the literature, this is sometimes erroneously referred to as a *relative gain matrix*, although it has none of the mathematical properties of a matrix. The correct term is *relative gain array.*

Suppose that the numerical value for steady-state gain k_{21} in Figure 13-3 had been +3, rather than –3. In that case, the following relative gain array would have been produced:

	m_1	m_2
x_1	-0.5	1.5
x_2	1.5	-0.5

Consider the situation in which the variables are paired so that relative gains of 1.5 are produced. Since the relative gain is the ratio of the direct effect to the ultimate effect, the direct effect must be greater than the ultimate effect. Furthermore, the indirect effect must be in the opposite direction of the direct effect. In other words, the loops oppose each other, but the indirect effect does not completely overcome the direct effect. As we shall see in the following discussion, in this case the pairing shown is not only the preferred pairing, it is the mandatory pairing.

If the loops were connected in such a way that the relative gains were -0.5, then the direct and ultimate effects must be of different signs. This implies that the indirect effect overcomes all of the direct effect, then more. Loops connected in this manner would appear to have positive feedback. When the set point is changed, a controller would initially change its output so as to move its measurement toward the new set point. Then, after the indirect effect—the control action of the other loop—the process variable would be further away from the set point than it was initially. Because of this, connecting the loops across a negative relative gain is prohibited; this implies that the other connection (in a 2×2 system) is mandatory.

Relative gain analysis is based upon the steady-state phenomena of the process. If the relative gains are between 0 and 1, the RGA indicates a preferred pairing. However, since either pairing is feasible—that is, will not lead to positive feedback—taking into account the dynamic characteristics of the process may lead one to choose the nonpreferred pairing. If some of the relative gains are greater than one and others are negative, then the RGA indicates the mandatory pairing, regardless of the process dynamics.

Relative gain analysis can be used in any process application that involves interaction between control loops. In the chemical and petrochemical industry, it has found considerable use in the analysis and design of distillation column control systems. Here there is often a choice between basic control strategies. For a particular control strategy candidate, a relative gain analysis will indicate the preferred pairing of the variables. More than that, by predicting the degree of interaction between control loops, the RGA will enable making a choice between alternative control strategies. Ref. 13-3 contains an in-depth discussion of this aspect.

For any application, the relative gain analysis can indicate significant interaction, for example, all the λ's lying between 0.4 and 0.6, or some of the λ's considerably greater than 1 while others are considerably less than 0. In these cases, consideration should be given to providing additional control components to decouple the control loops.

❖ DECOUPLING

Decoupling refers to the insertion of additional intelligence between the primary controllers and the final control elements so as to make the loops appear to be independent. The final control element may be valves; preferably, however, they will be lower-level flow-control loops.

In general, there are two forms of decoupling. We refer to the conventional form as “forward” decoupling. A less familiar form will be called “inverted” decoupling. Later in this section, we will show that inverted coupling offers certain advantages and also poses potential problems.

♦ Forward Decoupling

Our approach to forward decoupling resembles the approach used for feedforward control. Figure 13-4 shows a process in block diagram form. This process has two inputs and two outputs. The inputs could be either signals to valves or the set points of lower-level flow controllers. Dynamic interaction through the process is indicated by the transfer functions $P_{ij}(s)$, which are assumed to be open-loop stable. Controllers #1 and #2 are used to control the two measurements. If we used conventional feedback control without the decoupling elements $D_{12}(s)$ and $D_{21}(s)$, we could possibly experience significant interaction between the control loops. However, by inserting the decoupling elements between the controller outputs and the process inputs, we hope to make it appear as if the loops are decoupled. In other words, we want to choose $D_{12}(s)$ so that controller output #2 has no effect on measurement #1, and choose $D_{21}(s)$ so that controller output #1 has no effect on measurement #2.

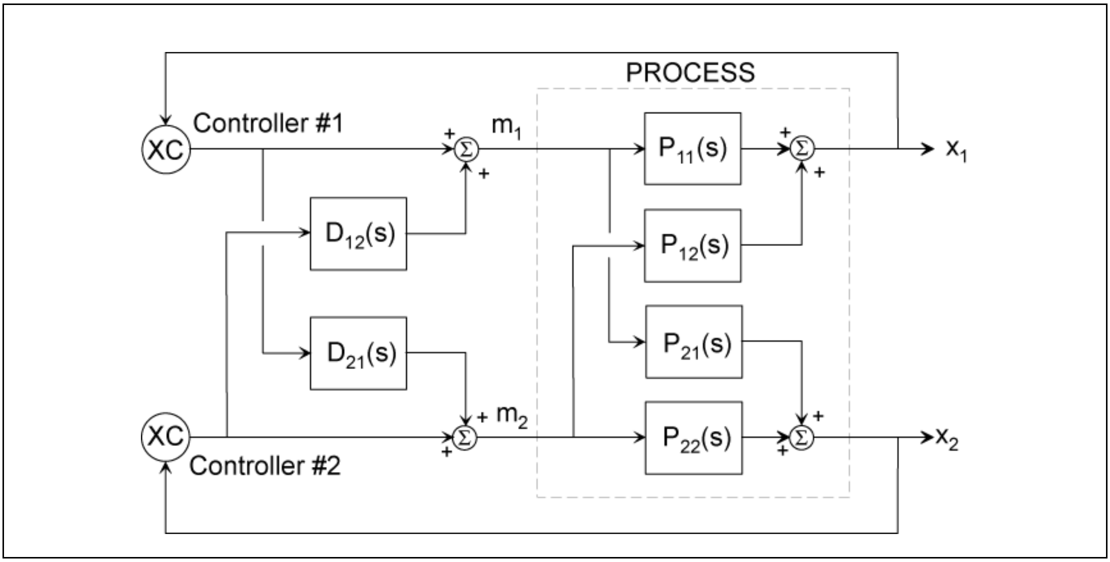

Figure 13-4. Forward Decoupling

Following the same line of reasoning as we used for feedforward control, we see that there are two paths from controller output #1 to measurement #2. One path is to input #1, then through the process path indicated as $P_{21}(s)$. The other path is through the (as yet undefined) decoupling element $D_{21}(s)$ to input #2**,** then through the process path $P_{22}(s)$. The composite transfer function from controller #1 to measurement #2 is $H_{21}(s)$, where $H_{21}(s)$ is given by

$$H_{21}(s) = P_{21}(s) + P_{22}(s)D_{21}(s).$$

We want to choose $D_{21}(s)$ so that this composite transfer function vanishes. That is, we want to make the two signal paths the mirror image of each other and to cancel each other at the measurement. Note that P_{21} is analogous to the "A" path in our feedforward discussion in chapter 11, and P_{22} is analogous to the "B" path. To make the composite transfer function vanish, we choose D_{21} according to the following equation:

$$D_{21}(s) = -\frac{P_{21}(s)}{P_{22}(s)} \tag{13-8}$$

Similarly, we choose $D_{12}(s)$ so that

$$D_{12}(s) = -\frac{P_{12}(s)}{P_{11}(s)} \tag{13-9}$$

Note that with this configuration, each process input is determined by a combination of both controller outputs. If we have approximated each of the process response paths by a simple first-order-plus-dead-time model, then each decoupling element will contain nothing more than a gain, lead-lag plus dead time, just as we utilized in feedforward control. If either the dead time or the time constant through the two paths are equal, then the decoupling element may not require one or both of the dynamic compensation terms.

In formulating the block diagram shown in Figure 13-4, we placed the decoupling elements between controller #1 and input #2 and between controller #2 and input #1, leaving the other paths (e.g., from controller #1 to input #1) as a straight-through connection. As an alternative, we could have placed the decoupling elements between controller #1 and input #1 and controller #2 and input #2, leaving the connections controller #l–input #2 and controller #2–input #1 as straight-through connections. Then the decoupling elements would be the inverse of those given by Equations 13-7 and 13-8. The choice will usually be dictated by process dynamics. If the dead time through P_{11} were longer than the dead time through P_{12} then D_{12} would not be realizable.[1] Similarly, if the dead time through P_{22} were longer than the dead time through P_{21} then D_{21} would not be realizable. If both of these conditions were true, then by using the opposite form of decoupling, the inverses of D_{12} and D_{21} would both be realizable. In Figure 13-5, in which the process blocks have been rearranged, the decoupling elements are designated D_{11} and D_{22}, where

$$D_{22}(s) = D_{12}^{-1}(s) = -\frac{P_{11}(s)}{P_{12}(s)} \tag{13-10}$$

1. A function of the form $\dfrac{\sum_{i=0}^{m} a_i s^i}{\sum_{j=0}^{n} b_j s^j} e^{-\theta s}$ is realizable if (1) $m \le n$, and (2) $\theta \ge 0$. This says that the order of the numerator must not be greater than the order of the denominator, and the dead time term must not be required to predict future values. Since the decoupling elements discussed here are the ratio of two FOPDT functions, the first criterion is inherently satisfied (i.e., $m = n = 1$). Hence the question of realizability focuses upon whether or not θ is positive, negative, or zero.

$$D_{11}(s) = D_{21}^{-1}(s) = -\frac{P_{22}(s)}{P_{21}(s)} \tag{13-11}$$

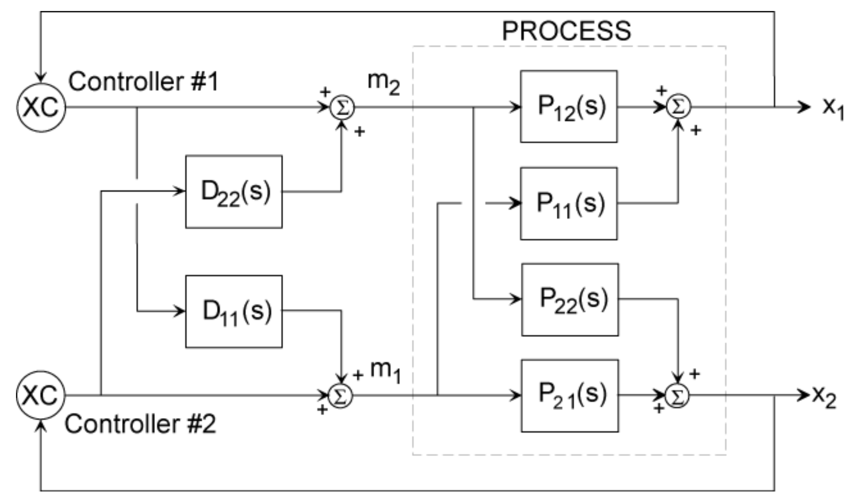

Figure 13-5. Alternative Forward Decoupling with Realizable Elements

Note that controller #1 uses input #2 to control its variable, and controller #2 uses valve #1.

Not all processes can be decoupled in this way. D_{12} may be realizable and D_{21} unrealizable, or vice versa. The inverses would also contain one realizable and one nonrealizable element. In this case, it would be possible to artificially lengthen the dead time through two of the process paths by inserting additional dead time in series with one of the final control elements. For example, suppose that θ_{12}, the dead time through P_{12}, is longer than θ_{11}, the dead time through P_{11} (making D_{12} realizable). Also suppose, however, that θ_{22}, the dead time through P_{22}, is longer than θ_{21}, the dead time through P_{21} (making D_{21} unrealizable). Finally, suppose that

$$\theta_{12} - \theta_{11} > \theta_{22} - \theta_{21}.$$

A dead time element, θ_1, can be inserted in series with input #1, as shown in Figure 13-6, where

$$\theta_1 = \theta_{22} - \theta_{21} .$$

This, in effect, increases the dead time through both P_{11} and P_{21}. When D_{12} and D_{21} are recalculated, both will now be realizable. Ref. 13-4 provides a general procedure for determining additional dead time that may be inserted to decouple a (2×2) process that otherwise could not be decoupled because of nonrealizable elements. Although this reference pertains to inverted decoupling, the procedure is equally applicable for forward decoupling.

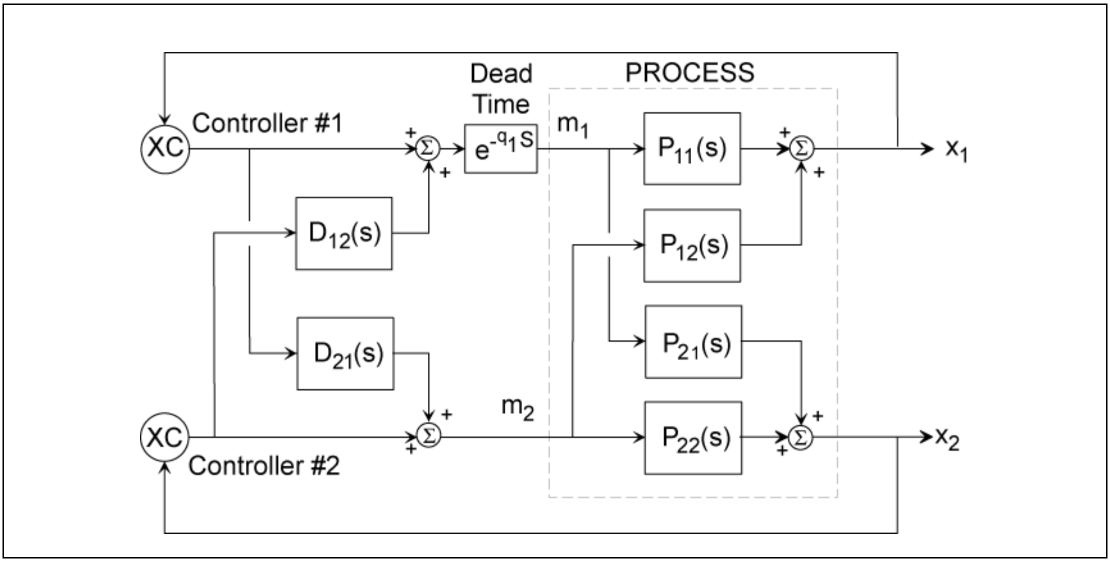

Figure 13-6. Insertion of Additional Dead Time to Force Realizability

Note that we deliberately increased the dead time in one of the control paths so as to achieve realizable decoupling elements. An alternative to this is partial decoupling, which will be described later in this chapter. In partial decoupling, only the naturally realizable element is used.

One drawback to the forward decoupling scheme shown in Figure 13-4 is that a control loop does not have the same response when the other loop is in manual as it does when the other loop is in automatic. For example, if controller #2 is in manual, there is no interaction and the process controlled by controller #1 is merely $P_{11}(s)$. However, when the other loop is in automatic, then controller #1 perceives a combination of all the process elements as the "process." If the $P_{ij}(s)$ functions are exact representations of the process, then the "apparent" process is as follows:

$$\frac{x_i(s)}{m_i(s)} = \frac{P_{11}(s)P_{22}(s) - P_{12}(s)P_{22}(s)}{P_{3-i,3-i}(s)} \quad \text{for } i = 1 \text{ or } 2.$$

Another drawback to the forward decoupling scheme is that if either final control element fails to maintain its integrity (that is, if a valve reaches a limit, or if a lower-level controller is placed in manual), then both control loops are affected. These drawbacks can be overcome by using inverted decoupling, discussed next.

♦ Inverted Decoupling

Figure 13-7 depicts inverted decoupling. With this configuration, each process input is a combination of a controller output and the other process input. The decoupling elements (see Equations 13-7 and 13-8) are of the same form as we used for forward decoupling, except that the subscripts for the decoupling elements are reversed. If each process path has been modeled as FOPDT, then the decoupling elements consist of at most a gain term, lead-lag, and dead time.

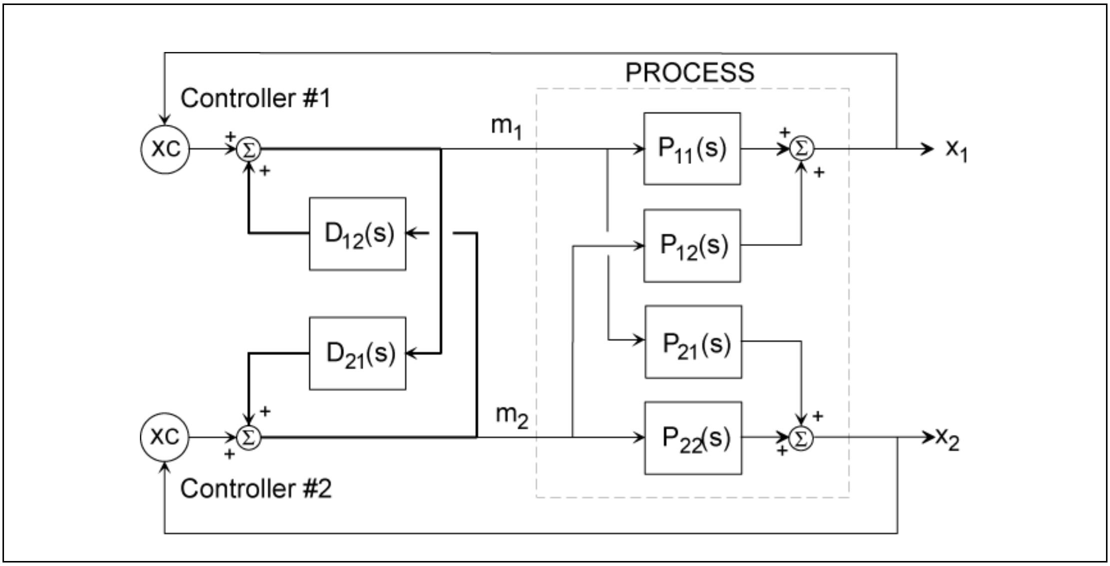

Figure 13-7. Inverted Decoupling

This configuration is especially effective if controllers #1 and #2 are primary controllers that are cascaded to the set points of lower-level flow controllers, as shown in Figure 13-8. Then the decoupling signal can be taken from the process variable of the flow controllers. The decoupling will then be in effect regardless of whether the flow loops are in manual or automatic, or even if a valve reaches a limit. For example, if flow controller #2 is in manual, the signal from the flow transmitter is passed through decoupling element D_{12} and combined with the output of controller #1 to determine the set point of flow controller #1. Later in this chapter, an actual implementation of this technique is shown in an application example.

The advantages of inverted decoupling include:

(1) The apparent process seen by each controller, when decoupling is implemented, is the same as if there were no decoupling and the alternate controller were in manual. For example, in Figure 13-7, the apparent process seen by controller #1 is always P_{11}.

(2) If the decoupler inputs are derived from lower-level control loop transmitters, as shown in Figure 13-8, then each decoupled loop is immune to abnormalities (e.g.,

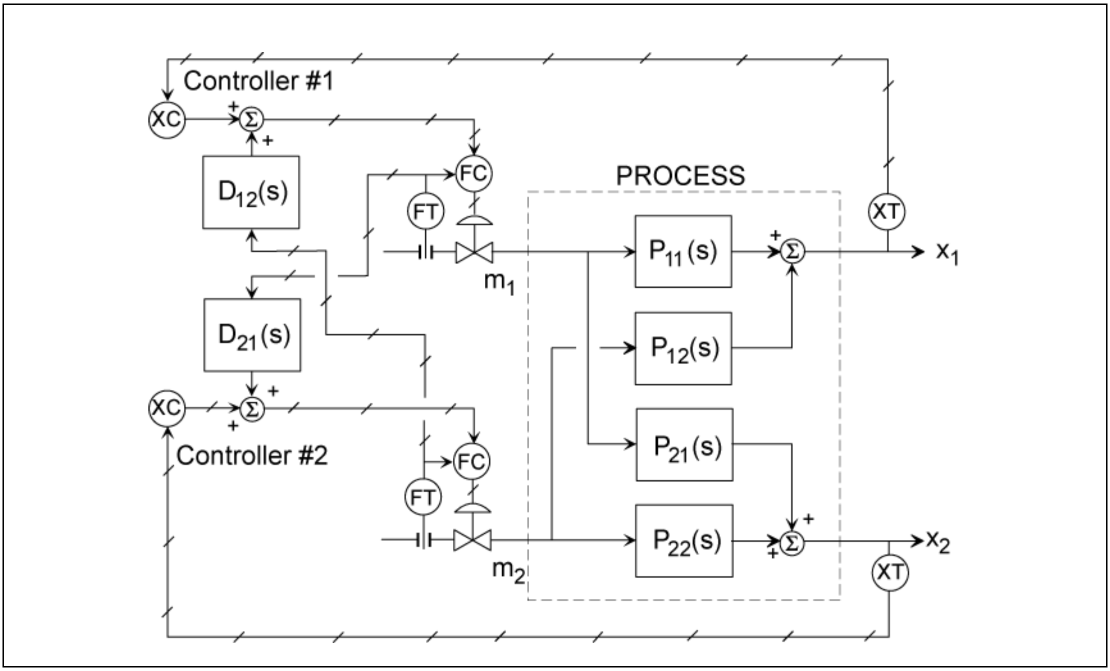

Figure 13-8. Inverted Decoupling Cascaded to Lower-level Controllers

valve at a limit or secondary controller in manual) in the secondary of the opposite control loop.

(3) With digital control system (DCS) or FOUNDATION™ Fieldbus PID function blocks inverted decoupling can often be implemented using the feedforward input and the internal summing junction. This will automatically provide such features as initialization and bumpless transfer between manual and automatic.

We now turn to three issues that are key to inverted decoupling: realizability, stability, and robustness.

Realizability

Realizable decoupling elements are elements in which the required dead time does not require that future values of the input be predicted (see the footnote on page 278 for further discussion of realizability). The question of realizability of the decoupling elements is the same for inverted decoupling as for forward decoupling. The same approaches as we described previously can therefore be used. These are:

(1) If both decoupling elements are nonrealizable, then the inverses of both decoupling elements are realizable. Compare Figures 13-4 and 13-5 as well as the text related to these figures.

(2) It may be possible to add dead time in series with one of the final control elements to force realizability of the decoupling elements. Since the input to the decoupler must be taken before this inserted dead-time element, however, the advantage

offered by taking the signal from a lower-level transmitter output is lost (see Figure 13-9). An alternative to this may be partial decoupling, which we describe later in this chapter.

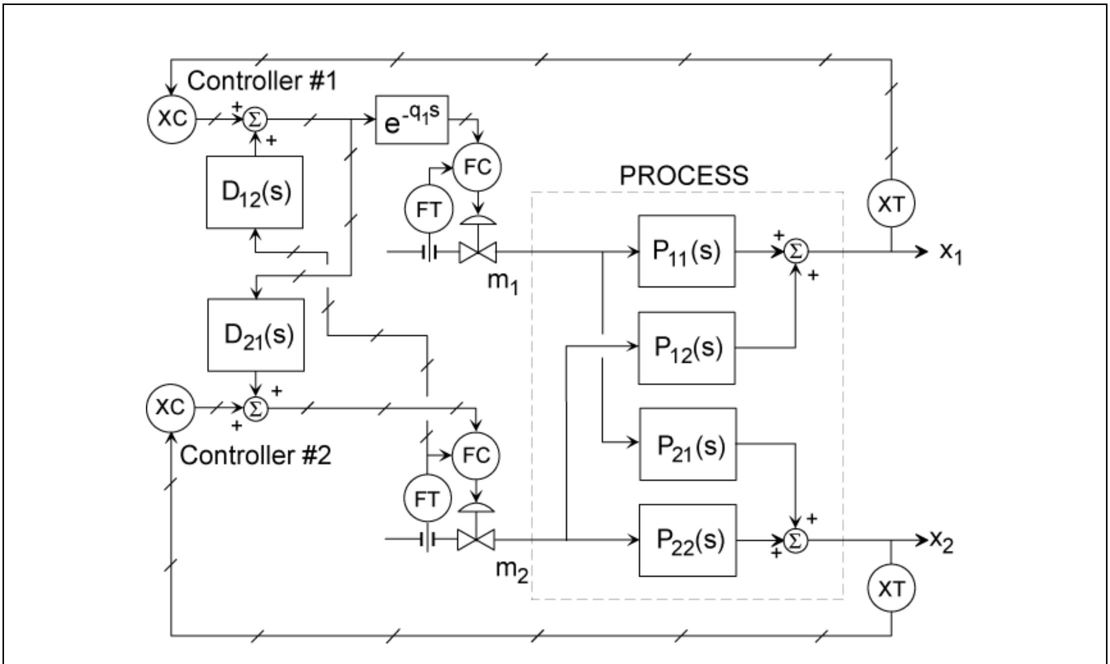

Figure 13-9. Insertion of Dead Time to Assure Realizability in Inverted Decoupling

Stability

One possible drawback to the inverted decoupling technique is that it creates a closed loop through the decoupling elements, as shown by the heavy black line of Figure 13-7. Under some circumstances, this can be an unstable loop. However, since all the elements of this loop are determined entirely by the control systems designer, its stability can be verified before it is actually installed.

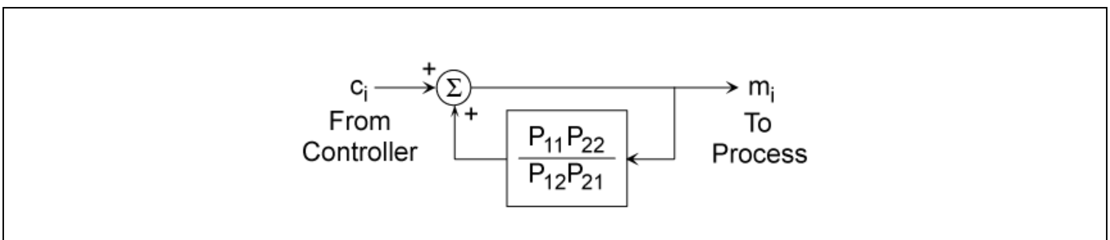

Figure 13-10. Equivalent Feedback Loop Created by Inverted Decoupling

For this analysis, assume that the realizability question has been addressed and the transfer functions and variables renumbered so that Figure 13-7 represents a decoupling control scheme with realizable elements. The feedback loop through the decoupling elements is not necessarily stable, however. Figure 13-10 shows a diagram of the equivalent feedback loop formed by the inverted decoupler. If each P_{ij} has been modeled as FOPDT, that is, if

$$P_{ij}(s) = \frac{K_{ij}\, e^{-\theta_{ij} s}}{\tau_{ij} s + 1},$$

then the loop equation is the following:

$$\frac{m_i(s)}{c_i(s)} = \frac{1}{1 - \dfrac{K(\tau_{11}s+1)(\tau_{22}s+1)}{(\tau_{12}s+1)(\tau_{21}s+1)} e^{-\theta s}} \tag{13-12}$$

$$\text{where } K = \frac{K_{12}K_{21}}{K_{11}K_{22}}, \tag{13-13}$$

$$\text{and } \theta = \theta_{12} - \theta_{11} + \theta_{21} - \theta_{22}. \tag{13-14}$$

Note that by assuming the realizability of the decoupling elements, $\theta \geq 0$.

The question of stability, then, lies in the solutions of the characteristic equation:

$$(\tau_{12}s+1)(\tau_{21}s+1) - K(\tau_{11}s+1)(\tau_{22}s+1)e^{-\theta s} = 0 \tag{13-15}$$

Note that we are not investigating stability in order to find the limits for setting a tuning parameter. Instead, we seek solutions to Equation 13-15 to provide the upper and lower limits that the value *K*, given by the process parameters in Equation 13-13, must meet in order for the inverted decoupler to be stable. Since *K* can be either positive or negative, we must have both upper and lower limits.

Ref. 13-4 gives a rigorous procedure for finding limits for *K* that involves solving a transcendental equation. If θ is zero or small relative to any of the process time constants, then approximate limits for *K* are as follows:

$$-\frac{\tau_{12}\tau_{21}}{\tau_{11}\tau_{22}} < K < min\left(1, \frac{\tau_{12}\tau_{21}}{\tau_{11}\tau_{22}}\right) \tag{13-16}$$

Robustness

McAvoy has reported that "ideal decoupling is sensitive to modeling errors" (Ref. 13-5). Since inverted decoupling is a form of ideal decoupling, this statement would appear to apply. In applications in which the loops are inherently closely coupled, say those in which some relative gain numbers are much greater than one and others are much less than zero, K will be a positive number and can easily approach an upper stability limit stated by Equation 13-16. In these cases, there is very little tolerance for modeling error; hence, McAvoy's conclusion is probably justified. In other applications, such as those in which all relative gain numbers are between zero and one, K will be a negative number. If the dynamics (time constants) of the process are not adverse, then K will be well within the limits stated by Equation 13-16, thus providing tolerance for modeling error.

♦ Partial Decoupling

Partial decoupling means "doing only half the job." For instance, suppose one of the measurements represents an attribute (say, composition) of a high-value product, whereas the other measurement represents a similar attribute of a lower-valued by-product. Here it might be highly desirable to prevent the lower-valued product control loop from affecting the other while at the same time tolerating disturbances from the higher-valued loop to the by-product loop. Thus, a decoupler can be provided in only one direction, as shown in Figure 13-11. The question of realizability can be checked immediately; the question of stability of the decoupler is not applicable. Hence, although only one decoupling element is used (consequently, the configuration effort is halved), the overall problems, including tuning the decoupler, are reduced by more than half. This is an approach that should be considered before embarking on a full decoupling project.

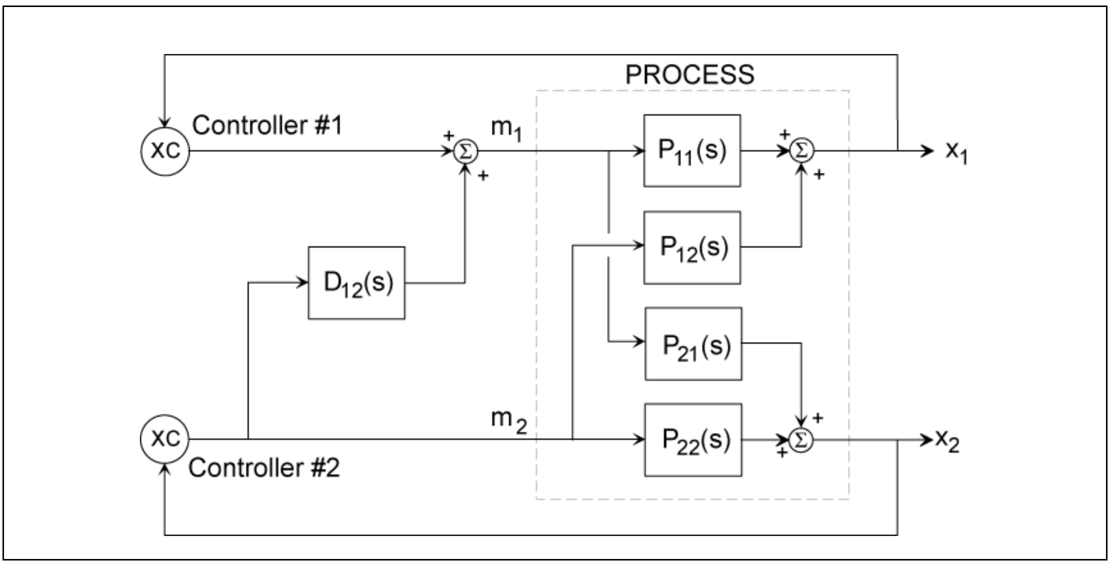

Figure 13-11. Partial Decoupling

❖ DECOUPLING APPLICATION EXAMPLES

In the last section we presented the formal concepts behind decoupling. In actual practice, these concepts may require modification on an ad hoc basis to fit actual circumstances. The following two subsections relate two actual installations where a decoupling approach was employed. These examples are presented not in the expectation that the reader will have an identical, or even similar, application to which these techniques can be applied, but to show how a decoupling approach has been used to solve real-world problems.

♦ Petroleum Refinery Heater

In the petroleum refining industry, two large fired heaters in the process stream are the crude heater, which heats the crude-oil charge to the crude (or atmospheric) distillation tower, and the vacuum heater, which heats the residue from the crude tower for charge to the vacuum distillation tower. These are both large capital equipment items as well as significant energy consumers.

At one refinery, these heaters needed to be replaced. Management decided to replace both heaters with a single dual-purpose heater. This novel arrangement undoubtedly saved on the initial capital outlay, but it ignored potential control problems.

The internal piping divided the heater into four quadrants. Three of the quadrants served as the "crude heater"; the fourth served as the "vacuum heater." Large upflow burners were installed in the base of each quadrant. The fuel flow to the crude heater quadrant burners was regulated by a crude-charge temperature controller, as the previous separate crude heaters had been. Likewise, the fuel flow to the vacuum heater burners was regulated by a vacuum-charge temperature controller (see Figure 13-12).

After construction and startup, a serious control problem was discovered. An inspection of the geometry of the heater revealed that all of the absorbed heat from burners #1 and #2 and approximately three-fourths from burner #3 goes to crude charge. The remainder goes to the vacuum charge. Similarly, approximately three-fourths of the heat from burner #4 goes to the vacuum charge; the remainder goes to crude charge. This caused the temperature loops to interact. If the crude-charge temperature controller calls for more fuel, the side effect is to raise the vacuum-charge temperature. That temperature controller then calls for less fuel, which creates the side effect of lowering the crude-charge temperature.

An on-the-spot study attempted to quantify the degree of interaction between the temperature control loops. Both temperature controllers were placed in manual, and step changes were introduced separately to the fuel flow controller set points; the response of each of the charge temperatures was recorded. Not surprisingly, the dynamics through each of the paths were approximately equal; hence, decoupling involved steady-state terms only. An inverted decoupler configuration was employed. A certain fraction of the measured fuel flow to each burner or set of burners was subtracted from the output of the opposite temperature controller, as shown in Figure 13-12. Thus, the set point of each fuel flow controller was reduced to compensate for heat contributed by the other heater's burner.

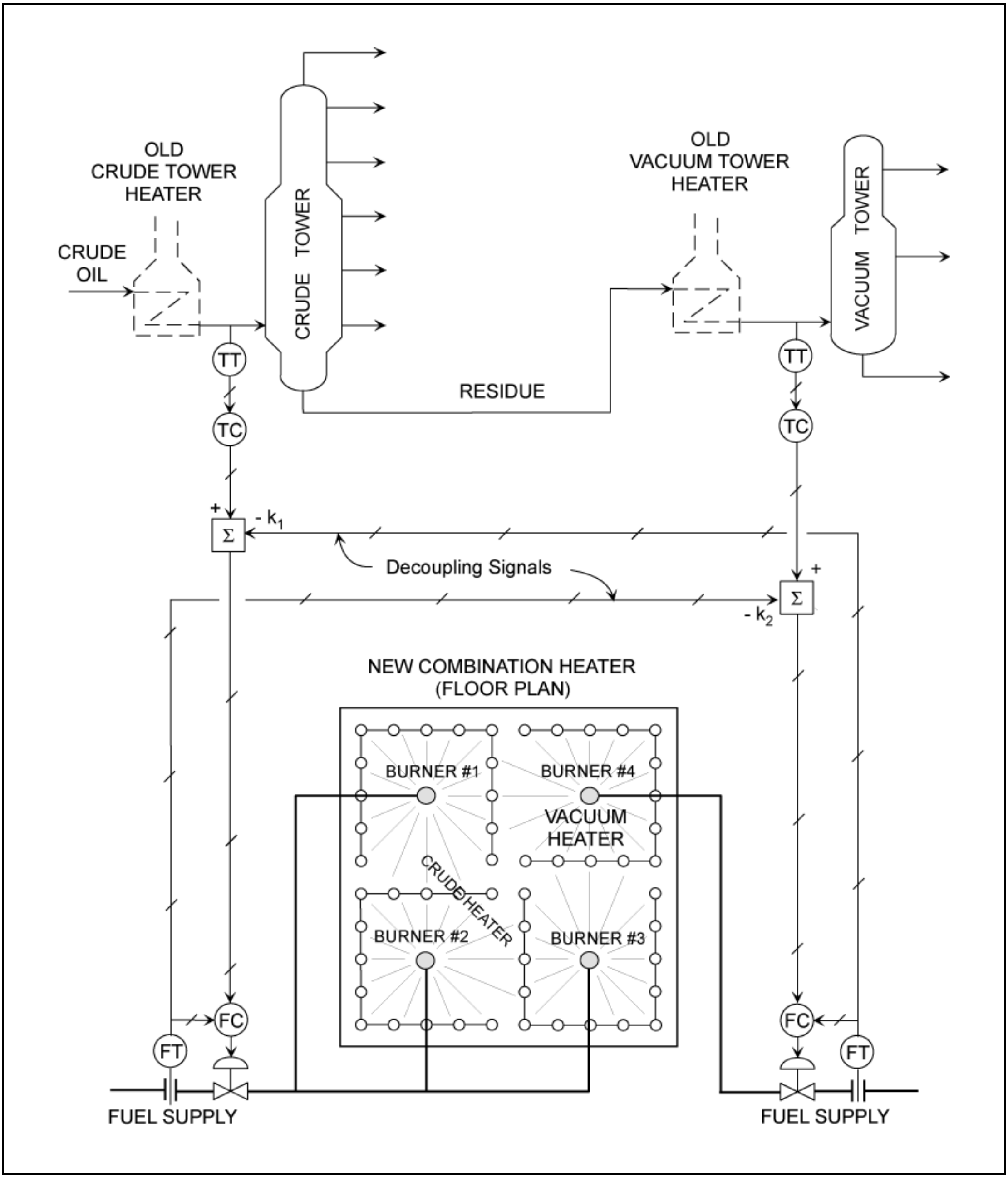

Figure 13-12. An Interaction Problem: Decoupling Solution at an Oil Refinery

Though the decoupling approach was an improvement over the original control scheme, which ignored the interaction, unpredictable convective air currents within the heater reduced the hoped-for improvement. The lesson to be learned from this example is that a poor process design cannot be overcome by a sophisticated control strategy, no matter how clever it is.

♦ Spray Water Temperature and Flow Control

One process plant needed to closely control the flow and temperature of a spray water. The flow rate and temperature directly affected the final product quality. The spray water was blended from hot- and cold-water supply headers, as shown in Figure 13-13. Total flow and temperature were measured, but not the flow rates of the separate hot and cold streams.

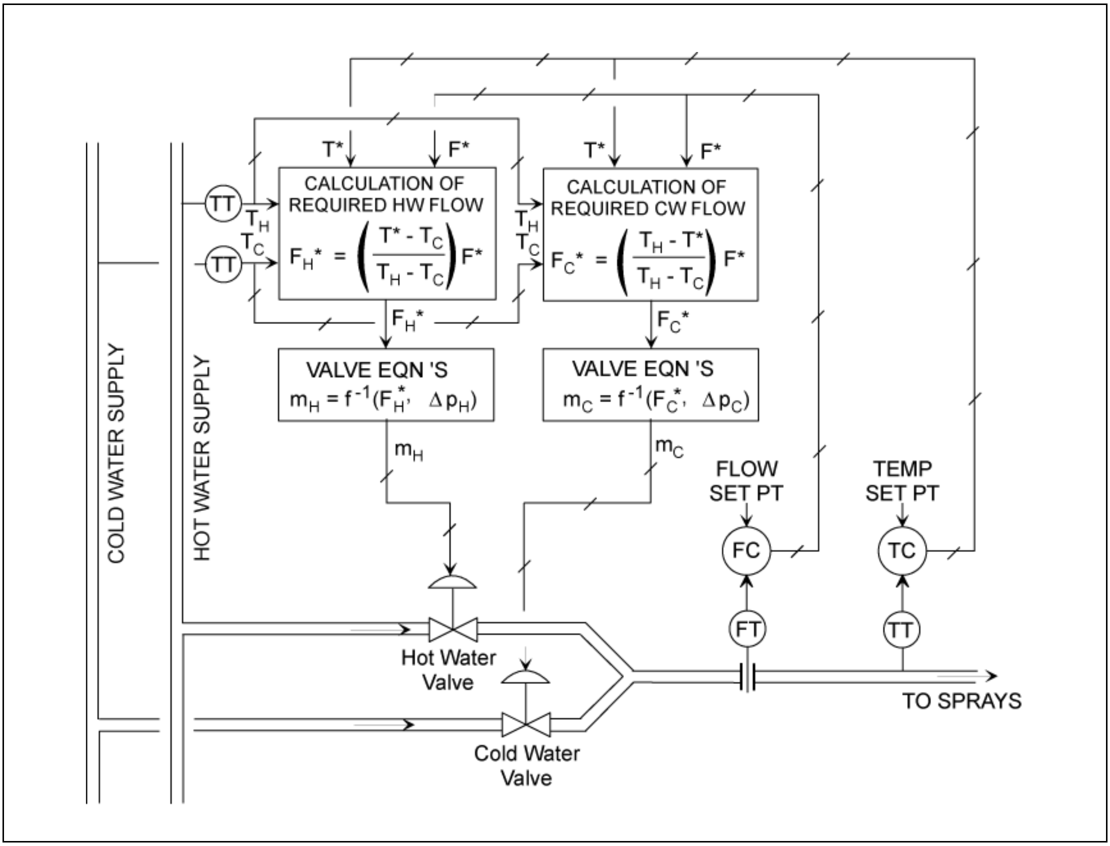

Figure 13-13. Decoupling Control of Spray Water Flow and Temperature

Different product recipes specified various requirements for the spray water temperature and flow rate. Furthermore, the temperatures of the hot- and cold-water supply headers also varied. As a result, it was impossible to make a consistently wise choice of the proper pairing of the variables, as we did in the discussion of Figure 13-1. Hence, a decoupling approach based on heat and mass balance equations was employed. With known values of the required temperature and flow rates for the spray water and known temperature of the headers, the required flows of cold and hot water could be calculated:

$$F_H^* = \left(\frac{T^* - T_C}{T_H - T_C}\right) F^* \qquad (13\text{-}17)$$

$$F_C{}^* = \left(\frac{T_H - T^*}{T_H - T_C}\right) F^* \tag{13-18}$$

where T^* = required spray water temperature
F^* = required spray water flow rate
$F_H{}^*$ = required hot-water flow rate
$F_C{}^*$ = required cold-water flow rate
T_H = hot-water supply header temperature
T_C = cold-water supply header temperature

Note that incorporating measured values of T_H and T_C introduces an element of feedforward into the calculations.

If there were flow controllers on the hot- and cold-water streams, the required flow rates could have been used directly as set points for these controllers. However, the plant only had control valves with equal-percentage valve characteristics. Given the required flow rates, the required valve positions were calculated by inverting the following equations:

Traditional liquid flow equation: $F = C_v \sqrt{\Delta p}$ (13-19)

Equal percentage valve equation: $C_v = R^{\frac{m}{100} - 1} C_{vmax}$ (13-20)

where: m = valve position, % open
R = rangeability factor (typically 50)
F = flow rate, gpm of water
Δp = pressure drop across valve, psig
C_v = valve coefficient, variable with stem position
C_{vmax} = tabulated C_v of valve, wide open

Combining Equations 13-19 and 13-20 and solving for m yields the following:

$$m = \left[ln(F) - \frac{1}{2} ln(\Delta p) + ln\left(\frac{Cvmax}{R}\right)\right] \frac{100}{ln(R)} \tag{13-21}$$

These calculations were performed for each of the control valves. The pressures up and downstream of the valves were not measured, therefore a constant value of Δp was assumed in the calculations. This undoubtedly introduced one source of error, although apparently not a significant one.

So far in the discussion, the control scheme has been based only on calculations, without any involvement of the feedback flow and temperature controllers. The nominal required values for spray water flow and temperature were introduced into the calculations as T^* and F^*. To incorporate feedback, the required values for temperature and flow were introduced as set points of feedback controllers. The outputs of these controllers then adjusted the nominal values of T^* and F^* into the calculations as necessary to maintain the correct temperature and flow. (This technique of combining feedback and feedforward was described in chapter 11. See the discussion related to Figure 11-9.)

The effective action of the control scheme is as follows. If the temperature is low, the temperature controller output, passing through both calculation paths, opens the hot water valve and closes the cold water valve in the amounts necessary to maintain a constant flow rate. If the flow is low, the flow controller output, passing through the same two calculation paths, opens both valves in the right amount. This maintains a constant ratio between the two flow rates, and hence maintains a constant mixed water temperature. The performance of this control scheme represented a considerable improvement over the original control scheme, which was based on a conventional, nondecoupled approach.

❖ REFERENCES

13-1. F. G. Shinskey. *Process Control Systems,* 4th edition, McGraw-Hill, 1996.

13-2. E. H. Bristol. "On a New Measure of Interaction for Multivariable Process Control," *IEEE Trans. Automatic Control,* January 1966.

13-3. F. G. Shinskey. *Distillation Control for Productivity and Energy Production,* 2nd edition, McGraw-Hill, 1984.

13-4. H. L. Wade. "Inverted Decoupling: A Neglected Technique," *ISA Transactions*, 36 (1) 1997.

13-5. T. J. McAvoy. *Interaction Analysis, Principles and Applications,* ISA – The Instrumentation, Systems, and Automation Society, 1983.

14

DEAD-TIME COMPENSATION AND MODEL-BASED CONTROL

In the approach for feedback control outlined earlier in this book, very little was required in the way of process knowledge. The only knowledge that was absolutely necessary was the directional response of the process to a change in controller output so the direct or reverse action of the controller could be properly set. Other than that, controller tuning, however determined, was used in feedback control to match the controller to the process. Indeed, the remarkable feature of the PID controller is its ability to bring the measurement to a desired value in the presence of set point or load changes with only minimal knowledge of the process. This feature is the contribution of the integral mode.

However, if knowledge of the process, and specifically some type of process model, is taken into account more precise controllers can be designed. Simple model-based controllers have been in use for some time. Currently, the amount of activity in developing and applying more sophisticated forms of model-based control is increasing. The motivation behind it is the desire to provide better control for more difficult applications, such as processes with a long dead time relative to the dominant time constant, processes with unusual response characteristics such as inverse response, processes in which control loops interact significantly, and processes that are subject to multiple physical or operational constraints.

This chapter presents an overview of model-based control. We start with the Smith predictor controller, which has been in use for many years, and proceed through more mathematically and computationally intensive techniques, including Dahlin's algorithm and internal model control. The following chapter continues this discussion and provides an overview of multivariable model predictive control. To be consistent with the theme stated at the beginning of this book, our objective is to provide readers with an intuitive understanding of these techniques and their capabilities rather than the mathematically rigorous details. Nevertheless, some mathematical derivations will be required. The reader may consult other references for a more extensive mathematical treatment.

We have already encountered a form of model-based control in feedforward and the closely related decoupling control. With those techniques, however, a standard PID controller was used, and feedforward or decoupling was an "add-on." In the techniques presented here, the process model becomes a part of the control algorithm itself.

❖ SMITH PREDICTOR CONTROL

Smith predictor control (Ref. 14-1) was developed for processes that have a long dead-time-to-time-constant relationship. One example of an application where this technique has enjoyed considerable success is in controlling the wet end of a paper machine. Thick stock containing approximately ½ percent fibers, water, and additives such as filler, sizing, dye, and resins, is fed to a "headbox." From the headbox it flows onto an endless, moving screen and becomes a mat. As the mat moves away from the headbox, the water drains away, eventually producing a mat that is dry and stable enough to be drawn to additional equipment for further drying. At that point, the fiber content per unit area is measured—this is called the "basis weight." Using this sensor data, a controller manipulates a mechanism at the headbox that controls the basis weight to a set value.

The actual papermaking process and controls are far more complex than this, but this description provides the essence of the application problem. It can be seen that the length of time it takes for the mat to travel from the headbox to the basis-weight sensor represents a pure dead time that greatly exceeds the time constant represented by the headbox. In chapter 2, we saw that as the ratio of dead time to time constant increased, the loop became more difficult to control.

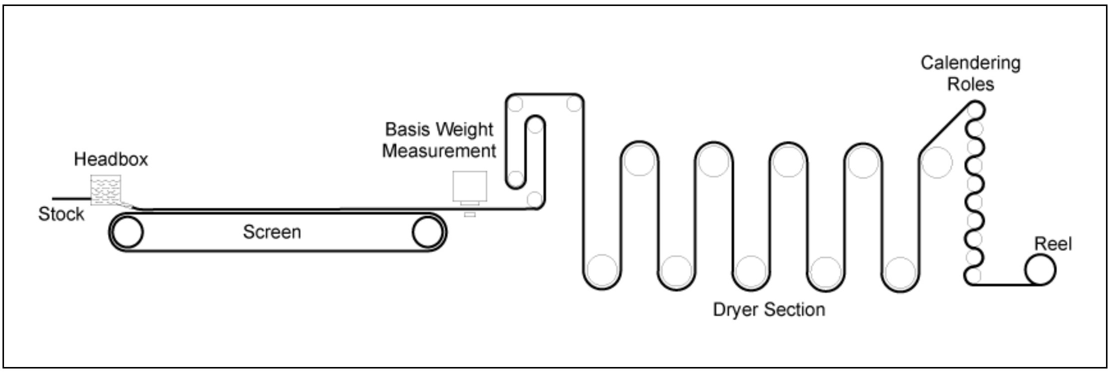

Figure 14-1. Paper Machine: An Application for Smith Predictor Control

To apply Smith predictor control, the process is modeled in two parts, the non-dead-time portion and the dead time. Hence, the process model can be represented in Laplace transform notation as:

$$\hat{P}(s)e^{-\hat{\theta}s} \tag{14-1}$$

where $\hat{P}(s)$ can be (and in practice usually is) as simple as an open-loop stable first-order lag, and $e^{-\hat{\theta}s}$ represents the process dead time. (The ^ indicates that these are process model values that are estimates of the true process functions.)

A block diagram (see Figure 14-2) of the control scheme shows that the controller output goes both to the process and to the first part of the process model, $\hat{P}(s)$. The output of the first portion of the process model then becomes a prediction of what the actual process measurement will be $\hat{\theta}$ minutes into the future. This predicted output can be used as a feedback signal.

If the model contained a perfect representation of the non-dead-time portion of the process and if there were no other disturbances, then this feedback is all that would be required. In the real world, neither of these two idealistic conditions will be true. Hence, the Smith predictor algorithm delays the predicted output by the time $\hat{\theta}$, which itself is a prediction of the actual dead time in the process. This delayed signal is a prediction of what the measurement value should be at the present time. This is compared with the actual measurement and the difference is added to the feedback signal. In the block diagram of the Smith predictor, Figure 14-2, all components within the broken lines comprise the control algorithm.

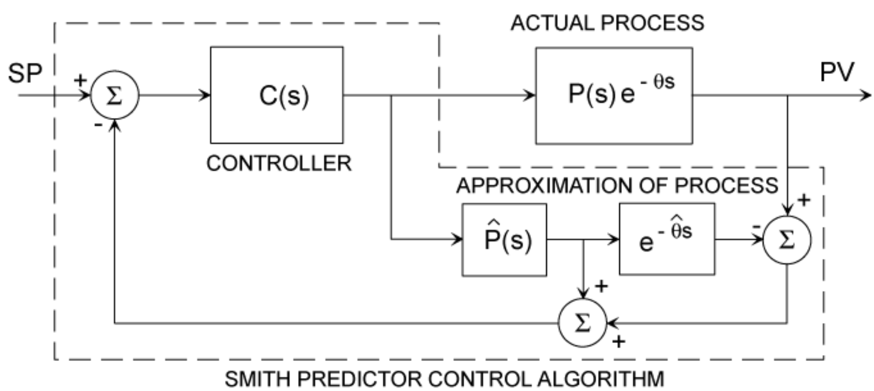

Figure 14-2. Smith Predictor Control Algorithm

We have not yet spoken of the block within the broken line that is labeled "controller." This is the feedback controller embedded in the control algorithm. At this point in our discussion, it could be a PI, PID, or any other chosen form. As such, it would require the usual controller tuning. In the following section, we will discuss the controller block in greater detail.

Meanwhile, let's consider the process knowledge required. If the process is modeled as first-order plus dead time, then the model in the Smith predictor is comprised of a first-order lag, dead time, and summation blocks. The parameters that must be entered are those of the approximating process model, specifically:

$\hat{K}_p$ - Estimated process gain

$\hat{\tau}$ - Estimated process time constant

$\hat{\theta}$ - Estimated process dead time

How accurately must these three parameter values be estimated? Of the three, the control scheme is the least forgiving of error in estimating the process dead time. If the process gain and time constant are exact but the dead time is in error, a lack of synchronization will characterize the relationship between a predicted response and the actual response to a controller output. The result will be a “double bump” in the feedback signal, separated by the time error in the dead-time estimate. For this reason, the Smith predictor has found the most ready applications in strip and sheet processes such as rolling mills and rubber processes, where the dead time is known with a fair degree of certainty. The paper machine certainly fits this category since the dead time is a direct function of the traveling screen speed, which is either known or can be directly measured. Other potential applications of Smith predictor control are in composition control for distillation columns (particularly if the time for moving a product sample from a process line to an analyzer represents a significant contribution to the dead time) and desuperheater control on a steam generator.

♦ Algorithm Synthesis

In our discussion of the Smith predictor, we said that we were free to choose the form of the controller function block, such as PI, PID, and so on. Having made this choice, our next task would be to determine controller tuning parameters. Alternatively, we can take a more formal approach, one that will determine both the form of the controller and its tuning parameters. In essence, our approach will “synthesize,” or design, the controller for us.

Note that with the Smith predictor, if the estimated process model were perfect, the apparent control loop would contain the non-dead-time portion of the process only. The loop would appear as if the dead time had been removed and placed downstream of the feedback control loop, as shown in Figure 14-3. (The true process variable would still be affected by the dead time, however.)

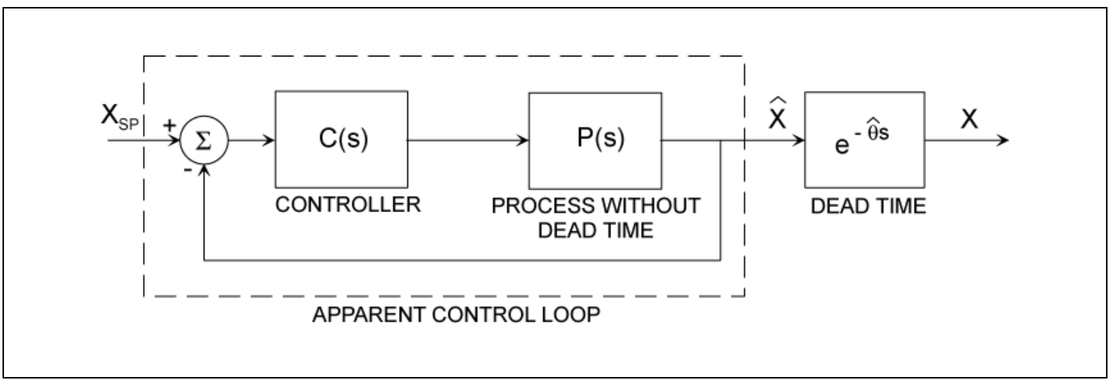

Figure 14-3. Apparent Control Loop Using Smith Predictor Control Algorithm with Exact Process Model

Suppose we consider the apparent feedback loop shown within the dotted line of Figure 14-3. Suppose also that we require this loop to have a certain response to a set point change. This is equivalent to requiring that the true process variable have that response when the dead time expires. (If the actual process contains dead time, there is no control strategy that will obtain a response to set point change in less time than that dead time.)

Call the output of the apparent feedback loop $\hat{x}$, and the actual process variable x. We can specify the required response of $\hat{x}$ to a set point step change as transfer function $F(s)$. That is, we require that

$$\frac{\hat{x}(s)}{x_{SP}(s)} = F(s) \tag{14-2}$$

$$\text{or} \quad \frac{x(s)}{x_{SP}(s)} = F(s)e^{-\theta(s)}. \tag{14-3}$$

Although we have the freedom to choose $F(s)$ to represent any type of function we wish, for simplicity we will choose $F(s)$ to be a first-order lag. That is, we want the apparent feedback control loop output to respond to a set point change as a first-order lag. The first-order lag must have unity gain since we want to start and end on the old and new set points, respectively. Hence, the only parameter we need to completely specify the desired response is the time constant, which we call λ. (We deliberately use the same symbol as we used for lambda tuning in chapter 6 since the concepts are quite similar.)

Now the question is, what type of controller, both in form and tuning parameters, will produce the specified response? We proceed by writing the transfer function of the feedback loop shown in the block diagram of Figure 14-3. By block diagram algebra, it can be shown that this transfer function is:

$$\frac{\hat{x}(s)}{x_{SP}(s)} = \frac{C(s)\hat{P}(s)}{1 + C(s)\hat{P}(s)} \tag{14-4}$$

(We could have used either $P(s)$ or $\hat{P}(s)$ in Equation 14-4 since by the assumption of a perfect process model, $P(s) = \hat{P}(s)$.)

Equate the right-hand sides of Equations 14-2 and 14-4, and solve for the unknown controller transfer function $C(s)$:

$$F(s) = \frac{C(s)\hat{P}(s)}{1 + C(s)\hat{P}(s)}$$

Hence:

$$C(s) = \frac{F(s)}{[1 - F(s)]\hat{P}(s)} \tag{14-5}$$

Since we know the form and the parameters of *P(s)* and *F(s)*, we can evaluate Equation 14-5 and determine both the structure and parameters required for *C(s)*.

Before we give an example, let's recapitulate our steps so far:

- We first formulated the Smith predictor control algorithm.
- We noted that if the process model were exact, then the apparent control loop would appear as Figure 14-3, with a feedback loop that contains no dead time.
- If we want the feedback loop to have a particular form of response to a set point change, then write that response as a transfer function *F(s)*. Since *F(s)* is within our freedom to choose, both in form and parameters, *F(s)* is known.
- We equated the expression for the closed-loop transfer function to *F(s)*. This expression involved the desired closed-loop response, *F(s)*; the known process transfer function (with no dead time), *P(s)*; and the unknown controller transfer function *C(s)*. We then solved the resulting equation and thus derived both the form and parameter values for the feedback controller, *C(s)*.

$$\text{Suppose } P(s) = \frac{K_p}{\tau s + 1}$$

$$\text{and } F(s) = \frac{1}{\lambda s + 1}$$

where the parameters of *P(s)* and *F(s)* are known quantities. The parameters of *P(s)* are determined from the actual process response; the time constant, λ, of *F(s)*, the desired closed-loop response, can have any value of our choosing. Then, applying Equation 14-5 yields,

$$C(s) = \frac{\tau s + 1}{K_p \lambda s}$$

This can be rewritten in the following form:

$$C(s) = \frac{\tau}{K_p \lambda}\left(1 + \frac{1}{\tau s}\right)$$

which is recognized as the form of the transfer function for a PI controller, with the tuning parameters related to the process model parameters and the time constant of the desired response, λ.

$$K_C = \frac{\tau}{K_p \lambda} \tag{14-6}$$

$$T_I = \tau \tag{14-7}$$

In this example, we replaced the requirement to determine two tuning parameters for the PI controller with the requirement to enter a single tuning parameter, λ, that specifies the desired closed-loop response. This is a characteristic feature of many of the model-based control schemes—single parameter tuning.

Figure 14-4 presents a comparison of a Smith Predictor controller, tuned according to Equations 14-6 and 14-7, with a PI and PID controller tuned according to Ziegler-Nichols rules. In each diagram, the process model is FOPDT, and a perfect model is assumed for the control algorithm. Figures 14-4a, 14-4b, and 14-4c differ in that the value of λ used is equal to the process time constant (14-4a), to one-half the process time constant (14-4b), and to twice the process time constant (14-4c). Note that in 14-4a the controller output makes one move, then because of the assumption of a perfect process model, it makes no further moves. In 14-4b, the controller output initially overdrives, then cuts back, as a result of the requirement that the closed-loop response be twice as fast as the open-loop response. In 14-4c, the opposite requirement is in effect; hence, the controller output makes a partial step, then gradually moves to its final value.

Figures 14-4d and 14-4e depict the performance of a PI and PID controller, respectively. A PI or PID controller is used for comparative purposes, even though the process dead time is much longer than the process time constant ($\theta = 1.6\ \tau$). (Normally, PID control is not recommended if the dead time exceeds the time constant.) Notice that in Figure 14-4d, the slow integral action occasioned by the long dead time causes the process variable to approach the set point very slowly. In 14-4e, the presence of derivative permits the controller gain to be higher and the integral action to be faster, so the achievement of set point is improved.

Except for one problem, the Smith predictor example could be implemented with analog control equipment, using the block diagram of Figure 14-2. The problem is that it is not possible to find an analog device that will perform the time delay required by the approximating process model. It can, however, be implemented in a digital-based system that has fast sampling function blocks that mimic analog devices. The function block types required are a first-order lag, dead time, summation blocks, and PI controller whose tuning parameters can be calculated from the process parameters and desired response time constant. If the sampling time is significant (say, longer than one-fifth the process time constant), then the approach given in the following section can be used.

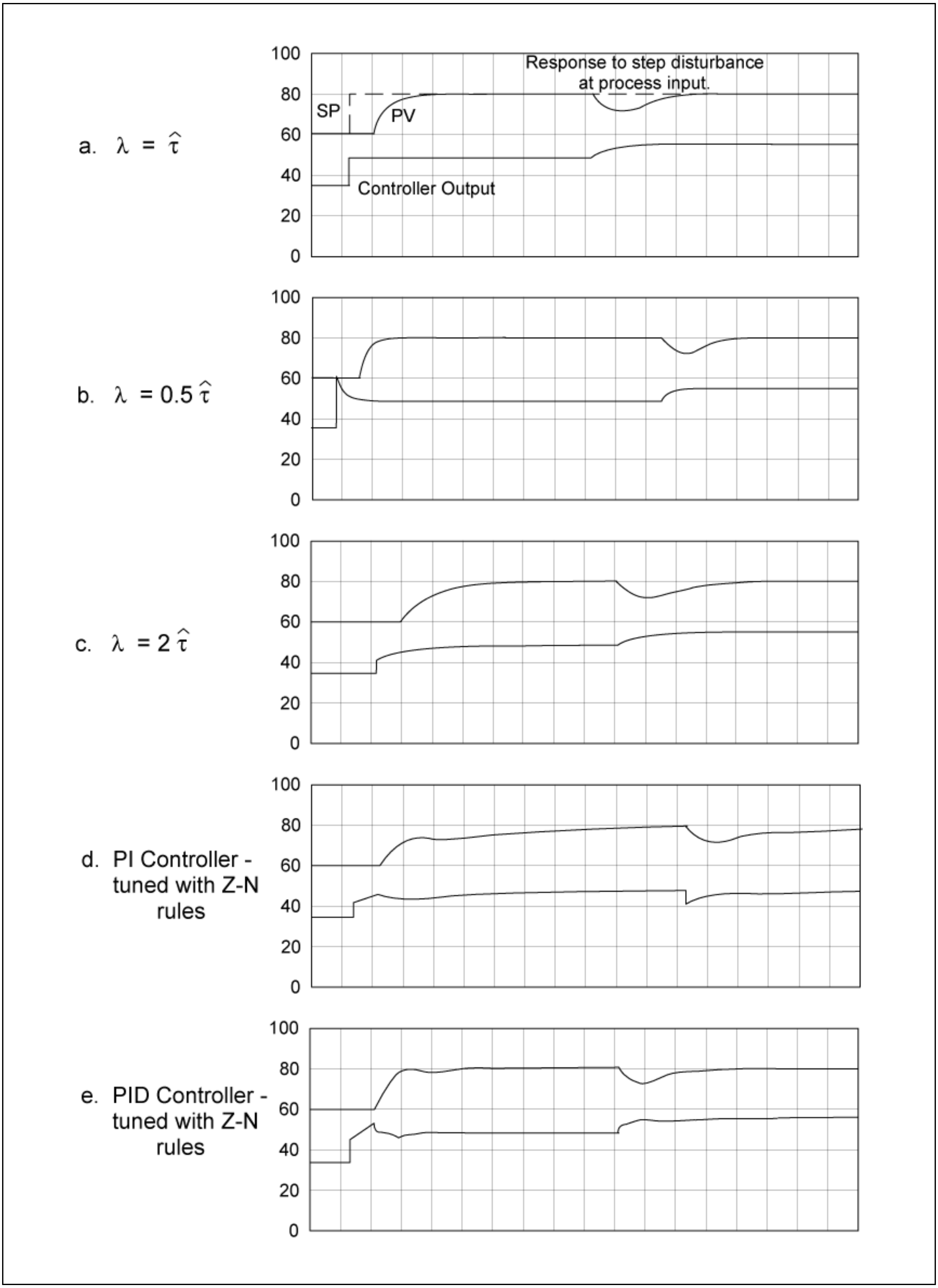

Figure 14-4. Smith Predictor, PI and PID Performance Compared

❖ DAHLIN'S ALGORITHM

Numerous circumstances occur where we cannot assume that the sampling time is insignificant and that function blocks that mimic analog devices will be sufficient. For instance, the process variable itself may be sampled by a sensor that only provides periodic data. Or we might want to reduce computer loading by reducing the frequency at which the control algorithm is processed. Hence, we must take into account the sampling period itself when determining the control algorithm.

What we seek is a discrete algorithm of the following form:

$$m_i = c_0 e_i + c_1 e_{i-1} + \cdots + d_1 m_{i-1} + d_2 m_{i-2} + \cdots$$

This algorithm computes the controller output at time i as a function of the error values at time i, i-1, ... , and so on, plus previous values of the controller output itself at times i-1, i-2, … , and so on. The terms c_0, c_1, … , d_1, d_2, ... are weighting coefficients. Our algorithm synthesis procedure must determine how far back in time to go with the weighted sums, as well as values for the coefficients. In other words, we must determine both the form and the parameters of the control algorithm.

The algorithm synthesis procedure by the use of a mathematical technique known as "Z-transforms" (so called because the letter "Z" is used to represent the operation of shifting a discrete variable in time by one sample period). It is beyond the scope of this work to present the complete theoretical background for Z-transforms; a very brief and nonrigorous overview is presented within the boxed text on the next two pages. Let us say at the outset that we need this much theoretical background only for the design of the control algorithm. Once the algorithm is designed and implemented, the user of the algorithm needs only to enter certain parameters. Note that much of the development parallels the controller synthesis technique that we presented earlier for the Smith predictor algorithm.

If the process can be modeled as FOPDT, that is, if

$$P(s) = \frac{K_p e^{-\theta s}}{\tau s + 1}$$

then, by taking into account the sample-and-hold operation of the controller output to the process, and by using Equations 14-A6 and 14-A7 from the boxed text on pages 300 and 301, the following discrete representation is obtained:

$$P(z) = Z\{H\,P(s)\} \qquad (14\text{-}8)$$

$$= \frac{K_p (1 - a) z^{-(N+1)}}{1 - a^{z-1}}$$

AN OVERVIEW OF Z-TRANSFORM THEORY

(1) *Z*-transforms bear the same relation to difference, or discrete, equations as Laplace transforms do to differential equations. For example, the differential equation

$$\tau \frac{d}{dt} x(t) = -x(t) + Ky(t) \tag{14-A1}$$

can be transformed into a Laplace equation by replacing the functions $x(t)$ and $y(t)$ with their transforms, $X(s)$ and $Y(s)$, and by replacing the symbol for derivative operation, d/dt, with the symbolic operator "s".

$$\tau s X(s) = -X(s) + KY(s)$$

After rearranging, this becomes $\dfrac{X(s)}{Y(s)} = \dfrac{K}{\tau s + 1}$ (14-A2)

We can transfer between Equations 14-A1 and 14-A2 by inspection, without becoming involved in the theoretical background.

(2) Similarly, the difference (or discrete) equation

$$x_{i+1} = c x_i + d y_i \tag{14-A3}$$

can be transformed into a *Z*-transform equation by replacing the sequences $\{x_i, x_{i+1}, \cdots\}$ and $\{y_i, y_{i+1}, \cdots\}$ with their transforms $X(z)$ and $Y(z)$, and by using the symbolic operator "z" as a forward shift operator. That is:

$$z(x_i) = x_{i+1}$$

Hence, $$z X(z) = d X(z) + c Y(z)$$

or after rearranging: $\dfrac{X(z)}{Y(z)} = \dfrac{c z^{-1}}{1 - d z^{-1}}$ (14-A4)

We can transfer between Equations 14-A3 and 14-A4 by inspection, without becoming involved in the theoretical background.

(3) A continuous function, $x(t)$, or its continuous transform, $X(s)$, can describe a continuous physical signal at all points. On the other hand, a sequence, $\{x_i, x_{i+1}, \cdots\}$, or its discrete transform, $X(z)$, describes the physical signal only at discrete instances in time. The time between sampling instants is called the *sample period*. (We assume that the sample period is constant.) The sequence provides no information about the behavior of the continuous physical signal between sample instants.

(4) A transfer function, such as a first-order lag, has an equivalent *Z*-transfer function. The reader is asked to accept the following without proof.

(a) If

$$F(s) = \frac{K}{\tau s + 1}$$

then

$$F(z) = Z\{F(s)\} = \frac{K(1 - b)}{1 - bz^{-1}} \quad (14\text{-A5})$$

where

$$b = e^{-\frac{\Delta T}{\tau}}$$

and ΔT = sample period

(b) If $G(s) = e^{-\theta s}$

and if ΔT = sample period

and if $\theta = N \Delta T$

(i.e., there are exactly *N* sample periods in the dead time),

then

$$G(z) = Z\{G(s)\} = z^{-N} \quad (14\text{-A6})$$

(5) On the other hand, suppose that a continuous process, *P(s)*, is driven by a computer control algorithm, which computes its output periodically and holds the output value constant between times. Furthermore, suppose the process variable is sampled at the same rate as the processing of the control algorithm. These are common conditions in computer control applications. We cannot say, however, that the control algorithm sees the discrete equivalent of *P(s)*. Rather, we have to take the sample-and-hold operation into account. If the symbol *H* represents the sample-and-hold operation, then the control algorithm sees the discrete representation of the process as $P(z) = Z\{H\,P(s)\}$.

If

$$P(s) = \frac{K_p e^{-\theta s}}{\tau s + 1}$$

then

$$P(z) = Z\{HP(s)\} = \frac{K_p(1 - a)z^{-(N+1)}}{1 - a z^{-1}} \quad (14\text{-A7})$$

where

$$\Delta T = \text{sample period}$$

$$a = e^{-\frac{\Delta T}{\tau}}$$

$$N = \frac{\theta}{\Delta T}$$

$$\text{where:} \quad a = e^{-\frac{\Delta T}{\tau}} \tag{14-9}$$

$$\Delta T = \text{sample period}$$

$$N = \frac{\theta}{\Delta T}$$

(The equation for N assumes that there are an exact number of multiples of the sampling periods in the dead time. An extension of Z-transform theory, called "modified Z-transforms," treats the case where the dead time does not have an integral number of sample periods (see, for instance, Ref. 14-3).

Although the process variable will be a continuous function, we will have knowledge of it only at discrete sampling instants. Hence, our desired response can be stated only in terms of a discrete model. Suppose we want the response of the process variable to match the trajectory of a first-order lag at the sampling instants. However, since the controller output will be applied on a sample-and-hold basis and since the process contains dead time, we must also take these into account when formulating the desired response. Thus, the desired response is the discrete counterpart of a sample-and-hold, first-order lag plus dead time:

$$F(z) = Z\left\{\frac{1}{\lambda s + 1}e^{-\theta s}\right\} \tag{14-10}$$

$$= \frac{(1 - b)z^{-(N+1)}}{1 - bz^{-1}}$$

$$\text{where} \quad b = e^{-\frac{\Delta T}{\lambda}} \tag{14-11}$$

$$N = \frac{\theta}{\Delta T}$$

and λ is the time constant of our desired response trajectory. In fact, λ is our only tuning parameter.

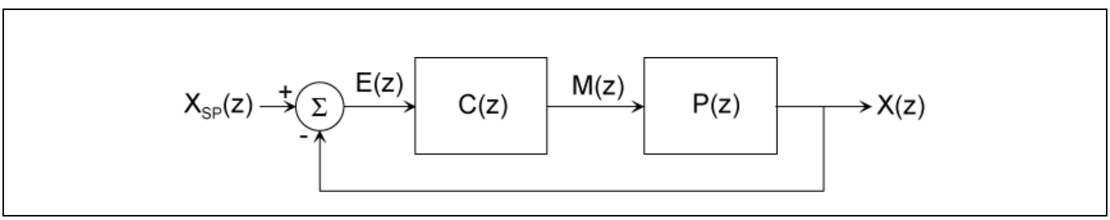

Figure 14-5. Feedback Control Loop for Deriving Dahlin's Algorithm

The feedback control loop is shown in Figure 14-5. The closed-loop function is as follows:

$$\frac{X(z)}{X_{SP}(z)} = \frac{C(z)P(z)}{1 + C(z)P(z)} \tag{14-12}$$

The desired, response, however, is given by

$$\frac{X(z)}{X_{SP}(z)} = F(z) \tag{14-13}$$

Equate the right-hand sides of Equations 14-12 and 14-13 and solve for the controller algorithm, *C(z)*:

$$\frac{C(z)P(z)}{1 + C(z)P(z)} = F(z) \tag{14-14}$$

$$\text{Therefore: } C(z) = \frac{F(z)}{[1 - F(z)]P(z)} \tag{14-15}$$

(Compare Equations 14-14 and 14-15 with Equations 14-4 and 14-5.)

After substituting Equations 14-8 and 14-10 for *P(z)* and *F(z)*, Equation 14-15 can be simplified to:

$$C(z) = \frac{c_0 + c_1 z^{-1}}{1 - d_1 z^{-1} + d_{N+1} z^{-(N+1)}} \tag{14-16}$$

$$\text{where } c_0 = \frac{1 - b}{K_p(1 - a)} \tag{14-17}$$

$$c_1 = -c_0 a \tag{14-18}$$

$$d_1 = b \tag{14-19}$$

$$d_{N+1} = 1 - b \tag{14-20}$$

and *a* and *b* are given by Equations 14-9 and 14-11.

Since the control algorithm's input is a sequence of variables representing the error and the output is a sequence of variables going to the process, we can write

$$\frac{M(z)}{E(z)} = C(z) \qquad (14\text{-}21)$$

From Equations 14-16 and 14-21 we can write the final form of Dahlin's control algorithm (Ref. 14-2), this time not in *Z*-transform notation but as a discrete algorithm (see item 2 in the boxed text on page 300):

$$m_i = c_0 e_i + c_1 e_{i-1} + d_1 m_{i-1} + d_{N+1} m_{i-(N+1)} \qquad (14\text{-}22)$$

This states that the present value of the controller output is to be computed as a weighted sum of the present and immediately past values of the error, the immediately past value of the controller output, and a stored value of the controller output from $N+1$ sample times ago. The coefficients c_0, c_1, d_1, and d_{N+1}, given by Equations 14-17 through 14-20, are merely the weighting coefficients. The *Z*-transform technique was used as a means for determining actual values for these coefficients.

Let us review what we have done:

- We first approximated the process as a first-order lag plus dead time and determined values for the parameters K_p, τ, and θ. We also determined a sample period, ΔT, so it was an integer divisor of the dead time, θ. (We could just as well have determined the sample period first, then estimated the dead time to be a integer multiple of the sample period.)

- Using the values K_p, τ, θ, and ΔT, we calculated the parameters of the discrete process model, *P(z),* using Equation 14-8.

- We made a choice for the time constant, λ, of the desired response (first-order lag plus dead time).

- From the values for λ and ΔT, we calculated parameters of the discrete closed-loop response model, F*(z),* using Equation 14-10.

- From the parameters of *P(z)* and *F(z)*, we calculated the coefficients of the control algorithm, using Equations 14-17 through 14-20. The algorithm was then implemented as Equation 14-22.

We are now ready for a trial of the control algorithm. The parameter λ represents the single tuning parameter by which we can affect the behavior of the algorithm.

If we have already been implemented steps 2, 3, and 5, that is, programmed them into a computer or microprocessor device, then our only responsibility is to determine and enter the process model parameters, the sampling period, and the tuning parameter λ. The tedious mathematical and computational aspects are done for us. This leads to an observation that applies equally to this and other model-based control techniques: designing and implementing the algorithm requires a relatively high skill level in mathematical theory and programming, higher than that required to, say, design a conventional control algorithm. The user of the algorithm, however, is insulated from those details as well as from the necessity of conventional tuning of the controller. On the other hand, if the algorithm does not perform as expected, the user has no intuitive feel for corrective action, other than to make adjustments to the single tuning parameter, λ. It would certainly not be expected that one could successfully modify individual coefficients of the control algorithm.

Figure 14-6 shows a simulated process that was first tested in the open loop to determine the process model parameters, then implemented and tested for a closed-loop response to a set point change and to a disturbance. The process model is not exact. The process was approximated by the same model used for Figure 14-4. The value of λ is chosen to be equal to (Figure 14-6a), equal to twice the value of (Figure 14-6b), and equal to half the value of (Figure 14-6c), the process time constant, τ.

❖ INTERNAL MODEL CONTROL

Internal model control (Ref. 14-4) is quite similar in concept to the control ideas discussed in previous sections. The structure of the control scheme it produces will differ, however. Like the others, internal model control (IMC) also requires an approximating process model. The process model can be of many forms. For simplicity's sake, we will again assume that the process can be modeled by an open-loop stable first-order lag plus dead time.

We start by redrawing the Smith predictor control algorithm shown in Figure 14-2. In place of a generic controller *C(s)*, we show the controller derived by algorithm synthesis in Equation 14-5**.** The redrawing is shown in Figure 14-7.

We then proceed through a sequence of block diagram manipulations, from Figure 14-7 to 14-8a, Figure 14-8b to 14-8c, then to 14-8d. Each block diagram is functionally identical with the previous diagram. In Figure 14-8c, the feedback controller is shown within a broken line. This in itself is a feedback loop that can be collapsed into a single block by the usual procedure:

$$\frac{\dfrac{F(s)}{\left[1-F(s)\right]\hat{P}(s)}}{1+\dfrac{F(s)}{\left[1-F(s)\right]\hat{P}(s)}} = \frac{F(s)}{\hat{P}(s)}$$

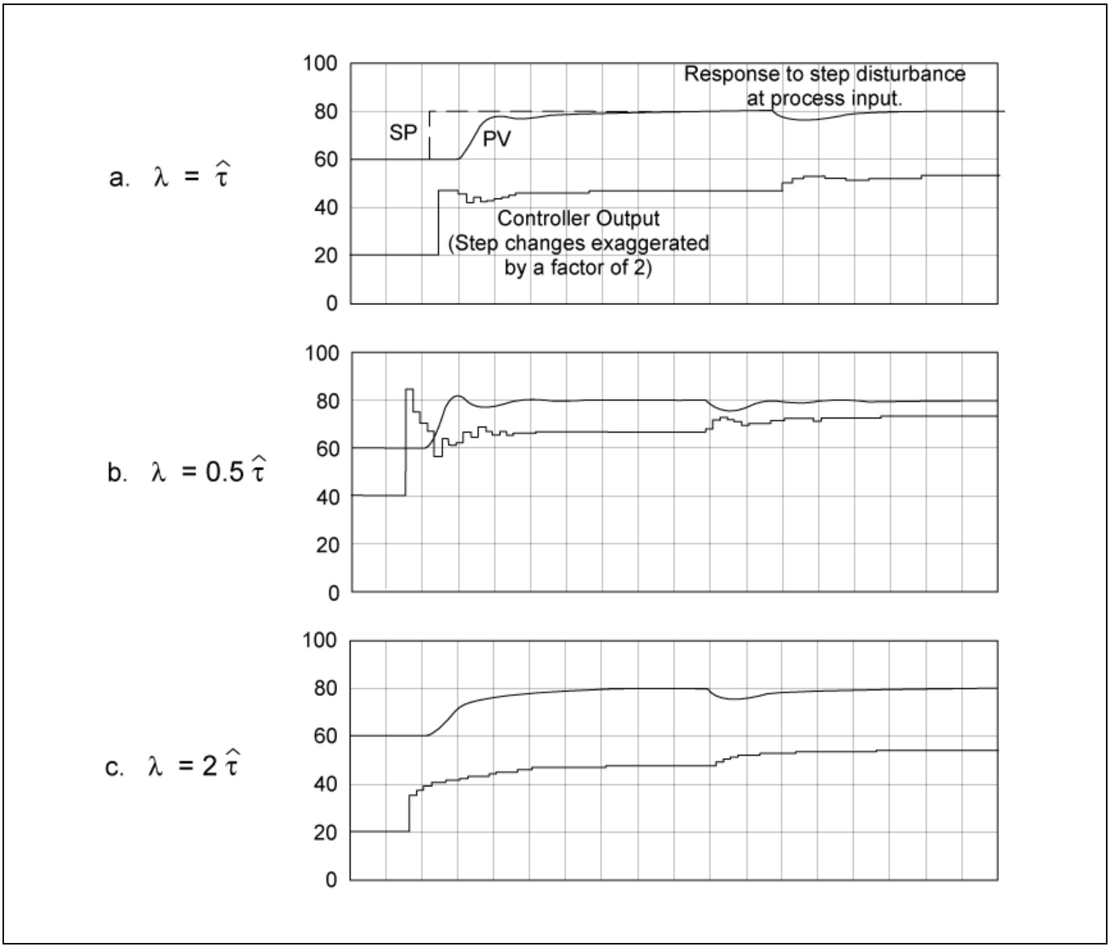

Figure 14-6. Performance of Dahlin's Algorithm
(Note: The process model is only an approximation of the true process.)

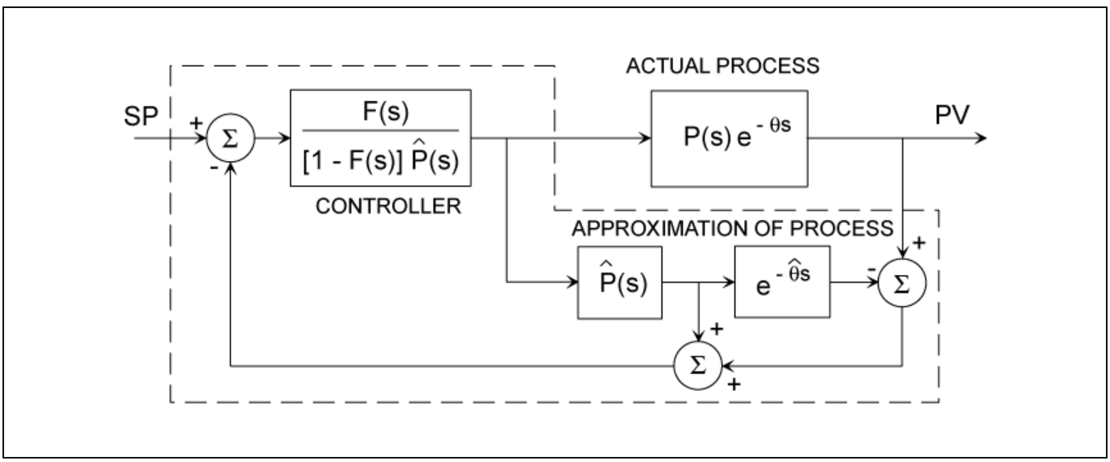

Figure 14-7. Smith Predictor Control Algorithm with Synthesized Controller

If the non-dead-time portion of the process is modeled as a first-order lag and the desired response is also a first-order lag[1], then

$$\hat{P}(s) = \frac{\hat{K}p}{\hat{\tau}s + 1}$$

$$F(s) = \frac{1}{\lambda s + 1}$$

and the controller becomes:

$$\frac{F(s)}{\hat{P}(s)} = \frac{\hat{\tau}s + 1}{\hat{K}_p(\lambda s + 1)} \tag{14-23}$$

This controller is shown in Figure 14-8d.

We make the interesting observation that, if *P(z)* and *F(z)* are both first-order, the feedback controller merely has a lead-lag, with no integrator. Actually, we should not perhaps call it a feedback controller at all, since if process and model match perfectly (and there are no disturbances), the output of the summation block on the right-hand side of Figure 14-8d is zero. This indicates that there is no feedback. The control loop is an open-loop system, in which the controller determines the output signal (i.e., valve signal) that produces the response we want. Notice that the term in the numerator of the controller (the "zero") cancels the lag term ("pole") in the denominator of the process. This leaves it to the denominator of the lag with time constant of λ to produce the desired response. The equivalent open-loop block diagram is shown in Figure 14-9.

In real applications, we are likely to encounter both a model mismatch and disturbances. The feedback signal biases the set point by just the right amount to cause the controller to calculate the correct output value so the process variable equals the set point. As an illustration, consider Figure 14-10, which shows only the steady-state components. Suppose the actual process gain is K, but an erroneous value, $\hat{K}$, has been used, where $\hat{K} \neq K$. Suppose also that there is some additive disturbance, d, to the process output. We start by calculating x and $\hat{x}$, then go forward around the loop (using the symbol shown in Figure 14-10):

1. If the non-dead-time portion of the process, *P(z)*, is modeled with higher than first order, then the desired response, *F(z)*, must also be of higher order so that the controller will be realizable.

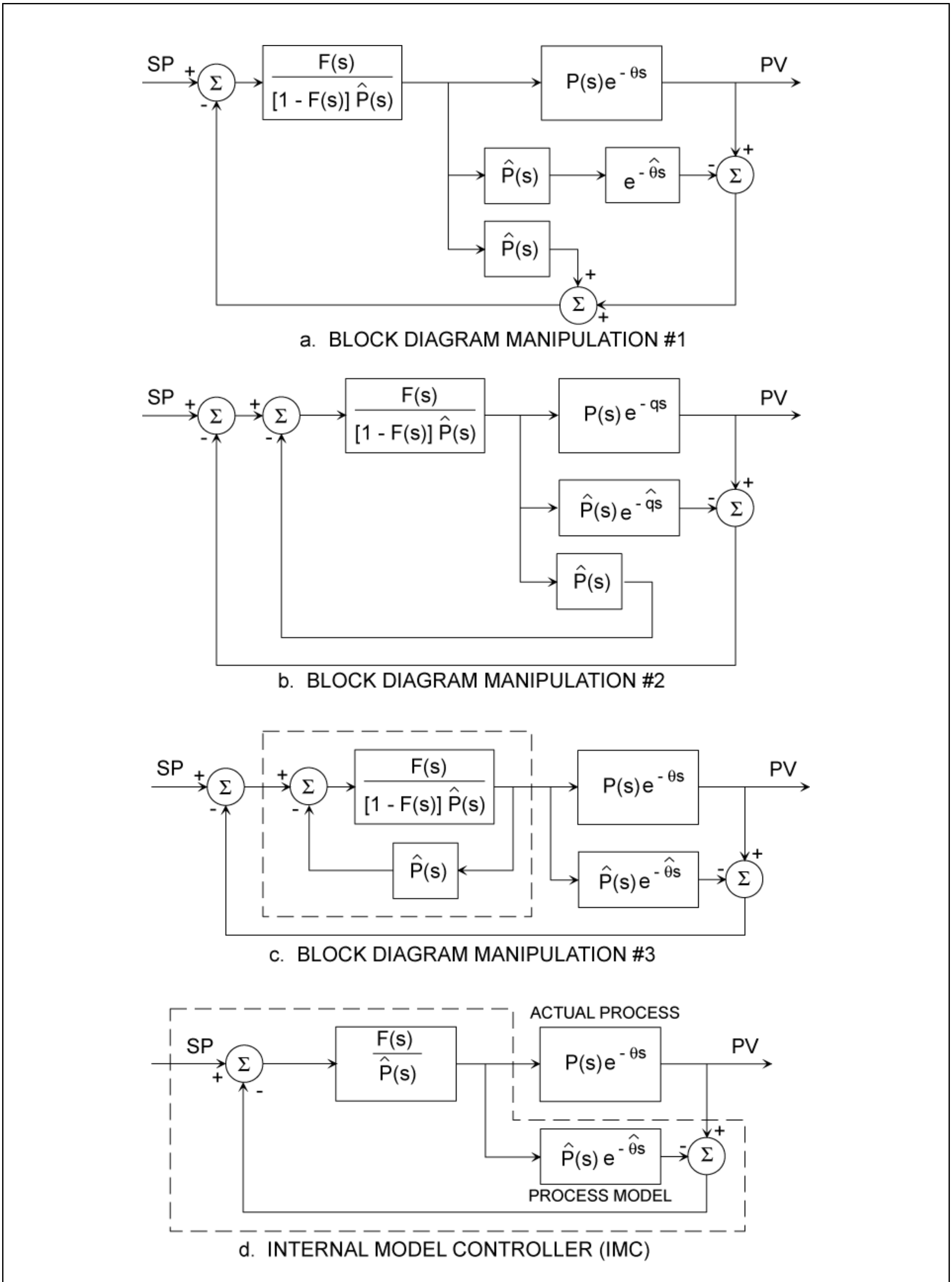

Figure 14-8. Evolution of Smith Predictor Controller to IMC by Block Diagram Manipulation

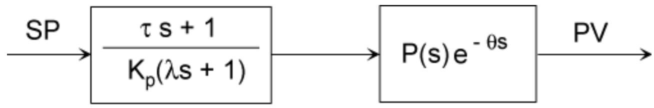

Figure 14-9. Equivalent Open-loop Diagram Assuming Perfect FOPDT Model

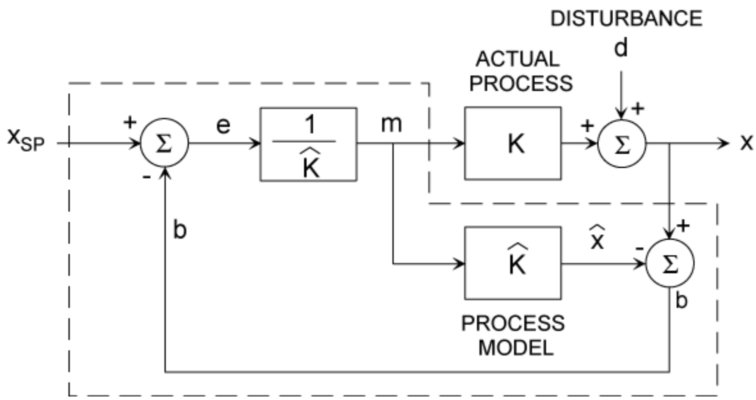

Figure 14-10. Diagram for Steady-state Analysis of Internal Model Control

$$
\begin{aligned}
\hat{x} &= \hat{K}m \\
b &= x - \hat{x} \\
&= x - \hat{K}m \\
e &= x_{SP} - b \\
&= x_{SP} - x + \hat{K}m \\
m &= \frac{1}{\hat{K}}e \\
e &= x_{SP} - x + \frac{1}{\hat{K}}\hat{K}e
\end{aligned}
$$

Therefore: $0 = x_{SP} - x$

This last equation demonstrates the fact that x and x_{SP} are equal, despite the mismatch in the process model and an unknown additive disturbance.

Note that the Smith predictor with algorithm synthesis, the Dahlin algorithm, and the internal model controller were all formulated on the same premise. That is, the process model was known and the desired response of the closed loop, except for the dead time, could be specified. Hence, these three formulations should all produce similar results.

In chapter 15, we present a different approach to model-based control that uses the time domain response rather than transfer functions.

❖ REFERENCES

14-1. O. J. M. Smith. "Closer Control of Loops with Dead Time," *Chem. Engr. Prog.*, vol. 53, 5 (1957): pp. 217-19.

14-2. E. B. Dahlin. "Designing and Tuning Digital Control Systems," *Instr. Cont. Syst.,* 42, (1968): pp. 77-83.

14-3. P. B. Deshpande and R.H. Ash. *Computer Process Control with Advanced Control Applications,* ISA – The Instrumentation, Systems, and Automation Society, 1988.

14-4. C. E. Garcia and M. Morari. "Internal Model Control 1 – A Unifying Review and Some New Results," *Ind. Engr. Chen. Process Des. Dev.,* 25 (1986): pp. 252-65.

15

MULTIVARIABLE MODEL PREDICTIVE CONTROL

❖ REAL-WORLD PROBLEMS

The real world presents many problems to the control system engineer. For example:

- Processes may not be well behaved; that is, they may have an inverse response or a long dead time. Hence, they may not be amenable to approximation as first-order plus dead time.
- The processes may consist of multiple-input, multiple-output (MIMO) interacting control loops.
- Rigorous dynamic models suitable for control purposes may not be tractable.
- The process may be subject to random disturbances.
- Several forms of constraints may be present:
  - on process variables,
  - on manipulated variables,
  - on auxiliary variables,
  - on rate-of-change of variables,
  - on combinations of variables.
- The constraints may vary with time or with operating conditions.
- No feasible solution may be available that will satisfy all the constraints simultaneously.
- Processes are not necessarily "square"; they may be "fat" or "thin," where
  - Square means same number of outputs as inputs (zero degrees of freedom),
  - Fat means more inputs than outputs (excess degrees of freedom; this usually leads to an optimization opportunity),

  - Thin means more outputs than inputs (insufficient degrees of freedom; all of the outputs cannot be controlled simultaneously),
- A valve going into saturation, a controller being placed in manual, or an analyzer failure may transpose a square or fat system into a thin system.

Many of these problems were discussed in earlier chapters. For example, feedforward control was applied to handle measurable disturbances. Override control was applied to handle constraints, though we did not note that situations may arise for which there is no feasible solution that satisfies all the constraints simultaneously. Decoupling control was used for interacting processes. However, if all these problems were encountered simultaneously, a more systematic approach would be desirable. Model predictive control (MPC) is one such approach.

MPC uses a sampled-data form of the process model to predict future values of a process variable based on past values of controller output. MPC compares the predicted values with future values of the set point to calculate both predicted future error values and predicted encroachment on constraints. From these predictions, MPC calculates a sequence of controller output values that will minimize some function of the predicted error as well as avoid encroaching upon the constraint. MPC is usually (but not always) applied to MIMO processes, subject to numerous disturbances and dynamically varying constraints. The technology thus encompasses feedback, feedforward, decoupling, and constraint control in one comprehensive package.

MPC involves a number of terms, which are represented in this chapter with the following symbols:

$\mathbf{v}$	column vector of n elements (size of $\mathbf{v}$ is nx1)
$\mathbf{v}^{\mathbf{T}}$	row vector (superscript T indicates transpose of the vector)
v_i	i^{th} element of vector v. $i = 1,2, \ldots, n$
$\mathbf{P}$	matrix of n rows and m columns (size of $\mathbf{P}$ is nxm)
$\mathbf{P}^{\mathbf{T}}$	transpose of matrix $\mathbf{P}$ (size of $\mathbf{P}^{\mathbf{T}}$ is mxn)
Pij	element of matrix $\mathbf{P}$, in the i^{th} row, j^{th} column
Δm	change in controller output (control move)
x	actual value of a controlled variable
$\hat{x}$	predicted value of a controlled variable
CV	controlled variables
MV	manipulated variables
DV	disturbance variables
AV	auxiliary variables
K	control horizon (number of future samples to calculate control moves)
N	prediction horizon (number of future samples to predict values of CV)
λ	time constant for reference trajectory

The following symbols are used for MIMO processes:

R	number of CVs
S	number of MVs
T	number of DVs

❖ HISTORY

MPC grew out of the need of industrial process control for a more comprehensive approach to the control problems mentioned at the beginning of this chapter. It was introduced through two seminal papers, both from industry rather than from academia (Refs. 15-4 and 15-11). The concepts presented in these papers, though very pragmatic, lacked a theoretical underpinning in such areas as stability and robustness. However, academic researchers in control technology soon related this MPC technique to internal model control (see chapter 14), and added the necessary theoretical basis (Refs. 15-5, 15-6, 15-7, 15-9).

Several companies commercialized the MPC technique by offering costly proprietary software packages, each with some variation on the basic technology. These early commercial packages required significant computing power for implementation. Because of the cost of the hardware and software, as well as the engineering services, on-site preparation, and commissioning costs, the initial market for MPC was generally limited to large, complex processing operations where the economic benefit was sufficient to justify the cost. Thus, the early MPC market was largely in petroleum refining operations. In a 1996 survey (Ref. 15-10), twenty-two hundred MPC installations were reported worldwide, of which approximately two-thirds were in the petroleum refining industry.

A typical early application for MPC in a petroleum refinery was for controlling the catalytic cracking unit. This is a large, highly interactive process with numerous constraints. Because of its impact on the refinery's overall economics, it is an ideal candidate for optimization. The MPC program accepted target values (for conversion rates, compositions, temperatures, etc.) from a steady-state optimization program, then generated set points for first-level controllers, such as flow control loops. By coping with the interactive nature of the process and its constraints, the MPC program made possible the steady-state optimization.

Several trends have been notable since the introduction and early application of MPC:

- The technology itself has been refined ("downsized") to be more efficient, thereby reducing the need for computing power.

- The cost of computational hardware has decreased dramatically.

- Expertise in applying MPC technology has become widespread among control system engineers.

- More vendors have appeared, each offering its own technological innovation. (One vendor even offers MPC packaged as a single-input, single-output (SISO) function block on their proprietary DCS.)

As a result of these trends, as well as market saturation in the petroleum refining industry, the MPC technique has enjoyed widespread use in other applications, including batch processes. If a survey were conducted today, it would undoubtedly reveal a large increase in the number of applications. Its use now extends across the entire industrial processing sector rather than being dominated by petroleum refining.

❖ UNCONSTRAINED MPC FOR SISO PROCESSES

Clearly, a control technique that is powerful enough to address all the problems we have listed deserves far more discussion than the scope of this work permits. Nevertheless, we will attempt to present the essence of the technique here both in mathematical terms and in graphical form, first for simple single-input, single-output (SISO) processes. Then, we will indicate how it can be extended to MIMO applications. Later in the chapter, we will discuss additional issues regarding MPC. For greater depth of coverage, see Refs. 15-3 and 15-12.

For consistency with the terminology used in the MPC literature, we will use the terms *process variable*, *control variable*, and *CV* interchangeably. Likewise, *process input*, *controller output*, *manipulated variable*, and *MV* are interchangeable, as are *disturbance variable* and *DV.* auxiliary variables, which are usually associated with constraints, will be termed *AV*s.

♦ Process Model

MPC begins by maintaining a sampled data step-response model of the process. Compare this model with the step-response model we have used several times in this work. We made a step change to the process input, observed the response, then approximated that response with three parameters, representing process gain, dead time, and time constant. With MPC, all of the sampled data collected through the step testing is retained in a series of memory locations, called a "vector." For example, in Figure 15-1, the sequence of values (p_1, p_2, … ,p_N) that result from a step-input change of one unit would be retained as the step-response model. This data vector is called **p** in Equation 15-1.

$$\mathbf{p} = \begin{bmatrix} p_1 \\ p_2 \\ \vdots \\ p_N \end{bmatrix} \tag{15-1}$$

Note that this technique is as valid for irregular responses such as those shown in Figure 15-2 as is it for the well-behaved process shown in Figure 15-1. Also note that for a self-regulating process, the response reaches an equilibrium when the sampled values stop changing. Hence, only a finite number of samples need to be retained. With commercial packages, *N* can be as small as 30 or as large as 120.

Note also that if the step-input change is something other than one unit, the data should be normalized so it represents the response that would result from a one-unit input change.

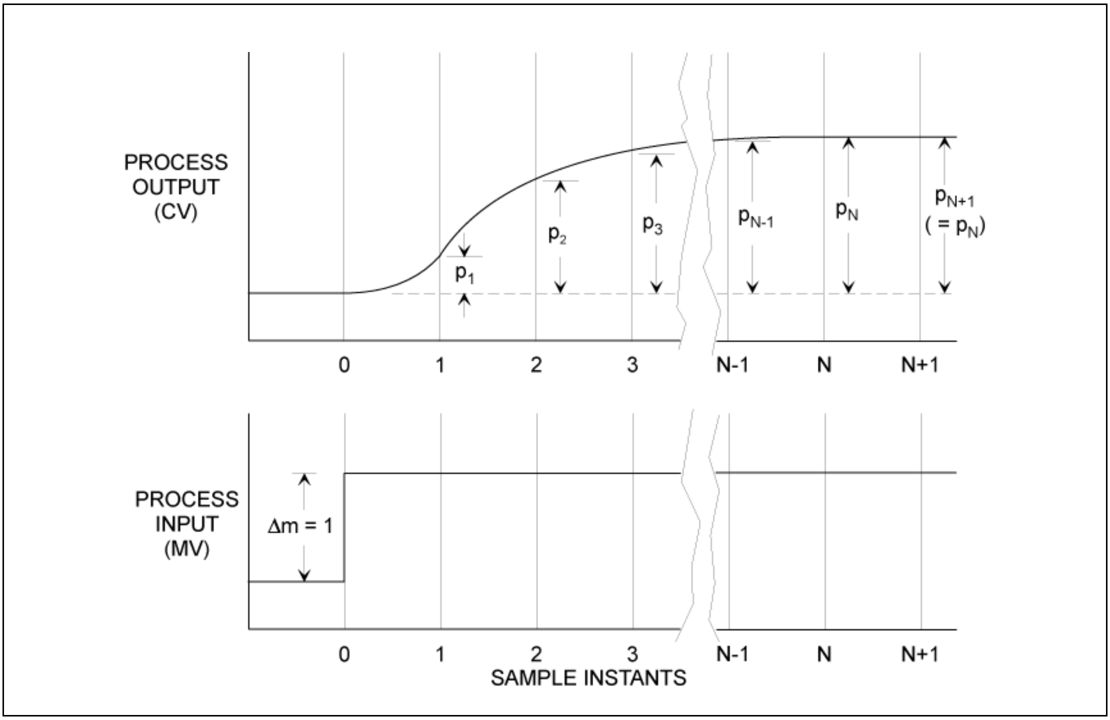

Figure 15-1. Step-response Model

Finally, note that what we have called "process input" may in fact be a set point change to a lower-level controller, such as a flow controller. In this case, we would be considering the flow loop merely as a part of the process.

While the procedure just described is valid in concept, in actual practice a more elaborate procedure may be employed for obtaining the step-response model. For instance, the data may be prefiltered to eliminate noise, there may be a series of alternate direction steps of varying lengths, and so on. One such test procedure is called a pseudo-random binary test sequence (Ref. 15-1). Determination of the process model is called the "identification" phase and may be a proprietary procedure for a particular commercial package.

❖ PREDICTION

For the next step in the exposition, let us assume that during some sample and control period, we know the current value of the *CV*; call this x_0. Furthermore, assume that we also have a sequence of values, $\left(\hat{x}_1, \hat{x}_2, \cdots, \hat{x}_N\right)$, or collectively, vector $\hat{\mathbf{x}}$. This represents our current prediction of what the *CV* will be for the next *N* sample periods, based on prior values of the process input as well as on the assumption that there will be no further changes in process input or disturbances to the process. The maximum index, *N*, is the number of sample values in our step-response model. This is called the "prediction horizon."

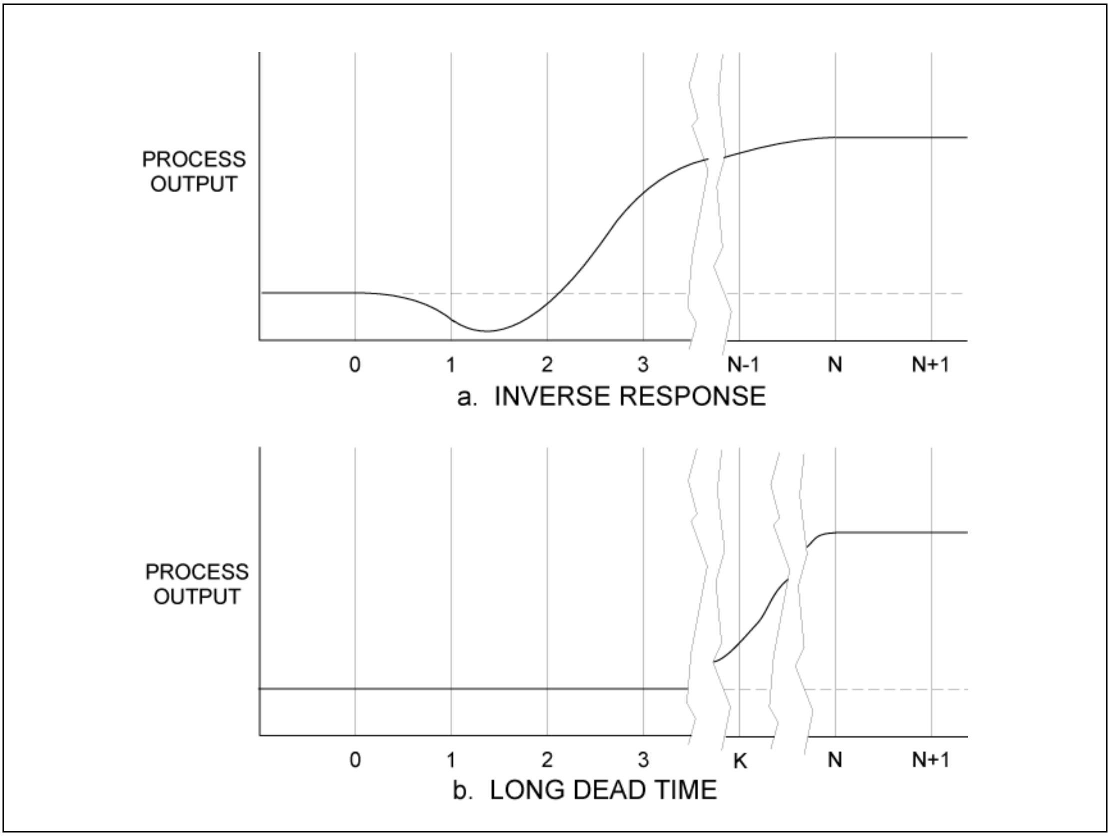

Figure 15-2. Step-response Model for Irregular Processes

Suppose also that we know the next *K* controller output moves (changes in the *MV*) that we intend to make. (Go along with us on this; don't worry about how we happen to know. Later, we will see how these moves are determined.) Call this sequence of moves $\left(\Delta m_0, \Delta m_1, \cdots, \Delta m_{K-1}\right)$, or collectively, vector $\mathbf{\Delta m}$. *K* is called the "control horizon"; *K* is much less than *N*, perhaps one-third of *N*.

Each control move will change our prediction of the future profile of the control variable. This is depicted by Figure 15-4 and shown by Equation 15-2. By the principle of superposition, the change to the predicted profile will be the magnitude of the control move multiplied by the step-response vector. (Recall that the step-response vector is the step response to a process input change of one unit.) For the control horizon, the predicted values of *CV* are given by the following:

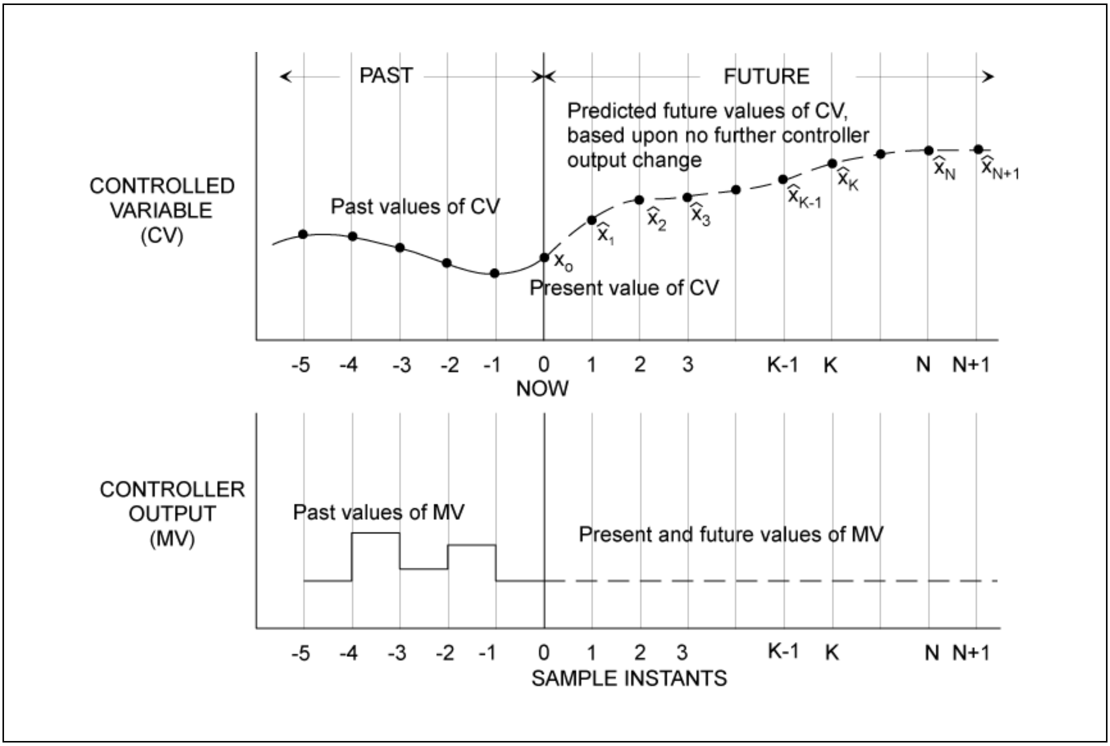

Figure 15-3. Predicted Profile, Based Only on Prior Control Moves

$$\begin{aligned}
\hat{x}_1 &= x_0 + p_1\Delta m_0 \\
\hat{x}_2 &= x_0 + p_2\Delta m_0 + p_1\Delta m_1 \\
\hat{x}_3 &= x_0 + p_3\Delta m_0 + p_2\Delta m_1 + p_1\Delta m_2 \\
\vdots &= \qquad \vdots \\
\hat{x}_K &= x_0 + p_K\Delta m_0 + p_{K-1}\Delta m_1 + \cdots + p_1\Delta m_{K-1} \\
\vdots &= \qquad \vdots \qquad \vdots \\
\hat{x}_N &= x_0 + p_N\Delta m_0 + p_{N-1}\Delta m_1 + \cdots + \cdots + p_{N-K+1}\Delta m_{K-1}
\end{aligned} \tag{15-2}$$

Equation 15-2 can be written in expanded vector-matrix notation, as shown in Equation 15-3, or written in compact vector-matrix notation, as shown by Equation 15-4.[1] $\mathbf{x}_0$ is an *N-vector* comprised of elements x_0, the current value of the control variable.

1. Readers unfamiliar with vector-matrix notations can consult any number of textbooks on this subject.

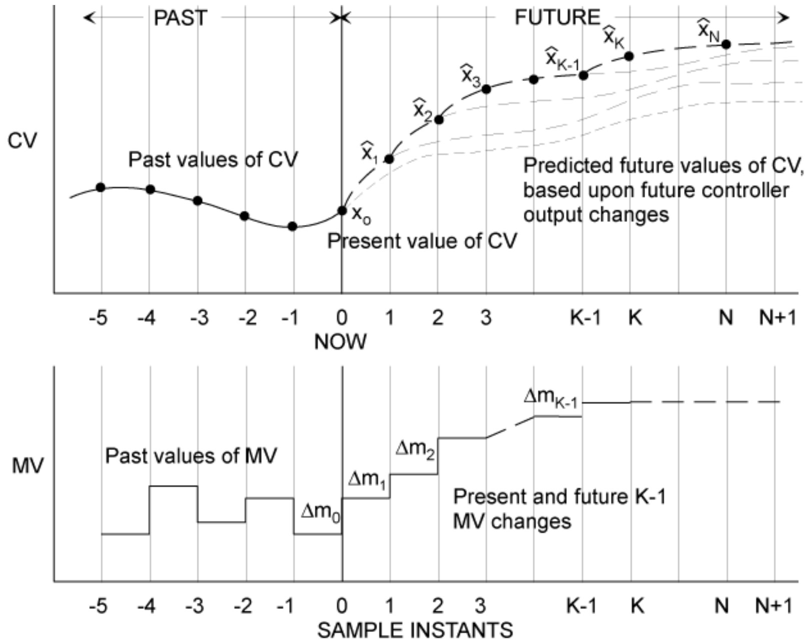

Figure 15-4. Modification of Predicted Profile by Current and Future Controller Moves

$$\begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \\ \hat{x}_3 \\ \vdots \\ \hat{x}_K \\ \vdots \\ \hat{x}_N \end{bmatrix} = \begin{bmatrix} x_0 \\ x_0 \\ x_0 \\ \vdots \\ x_0 \\ \vdots \\ x_0 \end{bmatrix} + \begin{bmatrix} p_1 & 0 & 0 & \cdots & 0 \\ p_2 & p_1 & 0 & \cdots & 0 \\ p_3 & p_2 & p_1 & & \vdots \\ \vdots & \vdots & & \ddots & \vdots \\ p_K & p_{K-1} & & p_2 & p_1 \\ \vdots & \vdots & & & \vdots \\ p_N & p_{N-1} & \cdots & \cdots & p_{N-K+1} \end{bmatrix} \begin{bmatrix} \Delta m_0 \\ \Delta m_1 \\ \Delta m_2 \\ \vdots \\ \Delta m_{K-1} \end{bmatrix} \qquad (15\text{-}3)$$

$$\hat{\mathbf{x}} = \mathbf{x_0} + \mathbf{P}\Delta\mathbf{m} \qquad (15\text{-}4)$$

♦ Calculating Control Moves

We will now address the question of how to determine the current and future control moves. We assumed previously that we knew the control moves; hence, we could correct the predicted profile of the control variable. If this were true and we knew the set point during the prediction horizon, as shown in Figure 15-5, then we could also predict the error values in the future. Call the sequence of error values $(\hat{e}_1, \hat{e}_2, \cdots, \hat{e}_N)$, or collectively, the vector $\hat{\mathbf{e}}$.

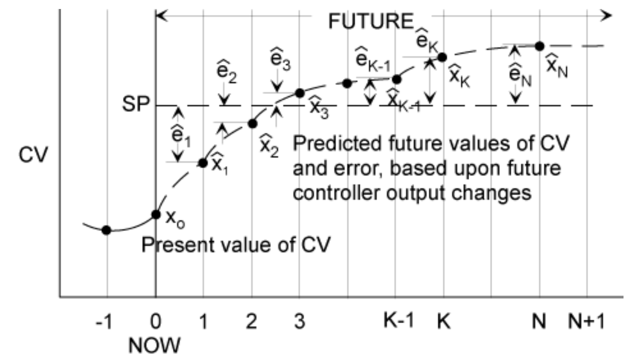

Figure 15-5. Predicted Error Profile

$$\hat{e}_i = x_{SP,i} - \hat{x}_i$$

Hence,

$$\begin{aligned}\hat{\mathbf{e}} &= \mathbf{x_{SP}} - \hat{\mathbf{x}} \\ &= \mathbf{x_{SP}} - \mathbf{x_0} - \mathbf{P\Delta m} \\ &= \mathbf{e_0} - \mathbf{P\Delta m}\end{aligned} \qquad (15\text{-}5)$$

where $\mathbf{e_0} = \mathbf{x_{SP}} - \mathbf{x_0}$

To calculate the control moves, we will minimize a cost functional, *J*, which is the sum of the squares of the predicted errors:

$$\begin{aligned}J &= \sum_{1}^{N} \hat{e}_i^{\,2} \\ &= \hat{\mathbf{e}}^T \hat{\mathbf{e}}\end{aligned} \qquad (15\text{-}6)$$

After incorporating Equation 15-5, this becomes:

$$\begin{aligned}J &= \left[\mathbf{e_0}^T - \mathbf{\Delta m}^T \mathbf{P}^T\right]\left[\mathbf{e_0} - \mathbf{P\Delta m}\right] \\ &= \mathbf{e_0}^T \mathbf{e_0} - 2\mathbf{e_0}^T \mathbf{P\Delta m} + \mathbf{\Delta m}^T \mathbf{P}^T \mathbf{P\Delta m}\end{aligned} \qquad (15\text{-}7)$$

The usual minimization procedure is to set the derivative to zero, hence:

$$\frac{\partial J}{\partial \Delta \mathbf{m}} = \begin{bmatrix} \dfrac{\partial J}{\partial \Delta m_0} \\ \dfrac{\partial J}{\partial \Delta m_1} \\ \vdots \\ \dfrac{\partial J}{\partial \Delta m_{K-1}} \end{bmatrix} \tag{15-8}$$

$$= \mathbf{P}^{\mathrm{T}}\mathbf{P}\Delta\mathbf{m} - \mathbf{P}^{\mathrm{T}}\mathbf{e_0}$$

Set the right-hand side of this equation to zero and solve for $\Delta\mathbf{m}$:

$$\Delta\mathbf{m} = \left[\mathbf{P}^{\mathrm{T}}\mathbf{P}\right]^{-1}\mathbf{P}^{\mathrm{T}}\mathbf{e_0} \tag{15-9}$$

The matrix **P** is a tall, slender matrix (size NxK) and cannot be inverted, but $[\mathbf{P^TP}]$ is a square matrix (KxK); hence, in general, it can be inverted. The matrix **P** is determined initially by the process model, Equation 15-1.

We do not need to invert $[\mathbf{P^TP}]$ at each control sample period, but only at the time **P** is determined. In fact, the entire matrix manipulation, $[\mathbf{P^TP}]^{-1}\mathbf{P^T}$, can be performed at that time. Furthermore, let matrix **W** be a (KxN) matrix, defined by the following:

$$\mathbf{W} = \left[\mathbf{P}^{\mathrm{T}}\mathbf{P}\right]^{-1}\mathbf{P}^{\mathrm{T}} \tag{15-10}$$

Matrix **W** is comprised of a series of K row vectors, each of N elements:

$$\mathbf{W} = \begin{bmatrix} - & \mathbf{w}_1{}^{\mathrm{T}} & \rightarrow \\ - & \mathbf{w}_2{}^{\mathrm{T}} & \rightarrow \\ \vdots & \vdots & \vdots \\ - & \mathbf{w}_K{}^{\mathrm{T}} & \rightarrow \end{bmatrix}$$

The current control move to be made, Δm_0, is calculated using only the top row, $[\mathbf{w_1}]^{\mathrm{T}}$, of **W.**

$$\Delta m_0 = [\mathbf{w_1}]^{\mathrm{T}}\mathbf{e_0} \tag{15-11}$$

The other control moves, $\Delta m_1, \Delta m_2, \cdots, \Delta m_{K-1}$, are not required, since after making the control move and correcting the predicted profile, we are going to step forward one sample period and repeat the procedure. This further reduces the computation burden at each calculation step.

The astute reader will observe that several ingredients are missing from our discussion of unconstrained MPC for SISO processes thus far:

- Feedback has not been utilized;
- No provisions for controller tuning have been presented;
- Furthermore, we have not utilized our knowledge of measurable disturbances.

These omissions will now be corrected.

♦ Incorporating Feedback

After making the control move calculated by Equation 15-11, the predicted profile must be corrected. This corrected profile includes a prediction of the value of the control variable at the next sample instant, $\hat{x}_1$. Once we have advanced to the next sample instant, that prediction becomes the prediction of the value of the control variable at the present time, $\hat{x}_0$. The actual value of the variable, x_0, is determined and the difference between the actual and predicted values is then calculated:

$$\Delta x_0 = x_0 - \hat{x}_0 \tag{15-12}$$

The entire profile, including the predicted current value, is then shifted by this difference, as shown in Figure 15-6:

$$\text{For } i = 0, 1, \cdots, N \qquad \hat{x}_i \leftarrow \hat{x}_i + \Delta x_0 \tag{15-13}$$

where "←" means "replaced by." Academic research (Refs. 15-5, 15-6, and 15-7) shows that this step is equivalent to adding an integrator into the control loop, thus assuring that the process variable will eventually come to set point.

♦ Tuning

There are two common techniques for tuning MPC. Some commercial systems use one or the other; some use both. These techniques are

- Move suppression
- Reference trajectory

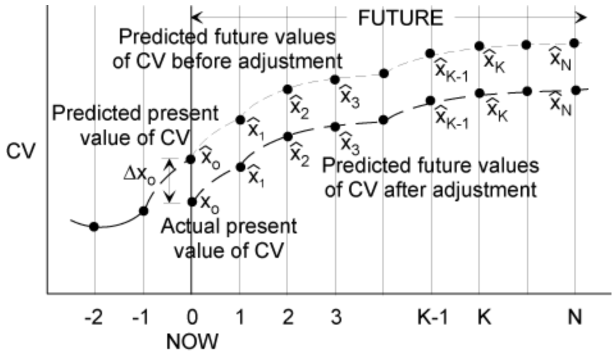

Figure 15-6. Incorporation of Feedback

Other design parameters that have an effect on performance, and hence could be considered as tuning parameters, include the sample time, the prediction horizon (N sample periods), and the control horizon (K sample periods).

Move Suppression

With this technique, the cost functional J, given by Equation 15-6, is augmented by the weighted sum of squares of the control moves.

$$J = \sum_{1}^{N} \hat{e}_i^{\,2} + \sum_{1}^{K} q_i \, \Delta m_{i-1}^{\;2} \tag{15-14}$$

$$= \hat{\mathbf{e}}^T \hat{\mathbf{e}} + \Delta\mathbf{m}^T \mathbf{Q} \, \Delta\mathbf{m}$$

$$= \left[\mathbf{e_0}^T - \Delta\mathbf{m}^T \mathbf{P}^T\right]\left[\mathbf{e_0} - \mathbf{P}\Delta\mathbf{m}\right] + \Delta\mathbf{m}^T \mathbf{Q} \, \Delta\mathbf{m}$$

$$= \mathbf{e_0}^T \mathbf{e_0} - 2\mathbf{e_0}^T \mathbf{P} \Delta\mathbf{m} + \Delta\mathbf{m}^T \left[\mathbf{P}^T \mathbf{P} + \mathbf{Q}\right] \Delta\mathbf{m}$$

Equations 15-15, 15-16, and 15-17 are analogous to Equations 15-8, 15-9, and 15-10:

$$\frac{\partial J}{\partial \Delta \mathbf{m}} = \begin{bmatrix} \dfrac{\partial J}{\partial \Delta m_0} \\ \dfrac{\partial J}{\partial \Delta m_1} \\ \vdots \\ \dfrac{\partial J}{\partial \Delta m_{K-1}} \end{bmatrix} = \left[\mathbf{P}^T \mathbf{P} + \mathbf{Q}\right] \Delta \mathbf{m} - \mathbf{P}^T \mathbf{e_0} \tag{15-15}$$

$$\Delta \mathbf{m} = \left[\mathbf{P}^{\mathrm{T}} \mathbf{P} + \mathbf{Q}\right]^{-1} \mathbf{P}^{\mathrm{T}} \mathbf{e_0} \tag{15-16}$$

$$\mathbf{W} = \left[\mathbf{P}^{\mathrm{T}} \mathbf{P} + \mathbf{Q}\right]^{-1} \mathbf{P}^{\mathrm{T}} \tag{15-17}$$

Note that Equation 15-11 is still applicable for the calculation of Δm_0.

In practice, the q_i weights in Equation 15-14 are usually selected to be the same value, q, thus leading to a single tuning parameter for move suppression. A larger value of q will lead to a more conservative controller.

Reference Trajectory

A reference trajectory for exponential return to set point from the present value is established by specifying a time constant, λ. Pseudo set point values are the values of this reference trajectory at the future sampling instances, using Equation 15-18. The error vectors used in Equation 15-6 or 15-14 are the differences between these pseudo set point values and the predicted values for the *PV*. This technique provides an additional parameter for tuning. A small value for λ will cause the controller to be aggressive; a larger value will result in a more conservative controller.

$$x_{SP,i} = \left(x_{SP} - x_0\right)\left(1 - \exp\left(-i\Delta T / \lambda\right)\right) \tag{15-18}$$

♦ Incorporating Feedforward Control

If there is a measurable disturbance, then a unit step-response model is determined for the effect of this disturbance on the process variable, similar to that shown in Figure 15-1. This disturbance model is characterized by a sequence of values, (d_1, d_2, ..., d_N), or collectively by the vector, **d**. At the beginning of each calculation cycle, the disturbance variable is measured, and the change in the disturbance since the last sample instant, u, is determined. Then the predicted profile is corrected to account for this change. Equations 15-19, 15-20, and 15-21 are modifications of Equations 15-3, 15-4, and 15-5 to incorporate this feature. The vector $\mathbf{e_0}$ can then be used in Equation 15-9 or 15-16 to compute the control moves.

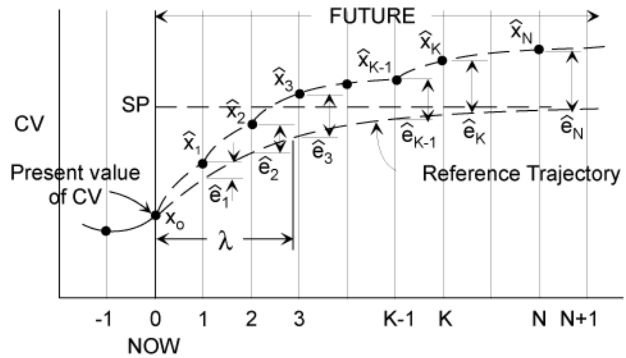

Figure 15-7. Reference Trajectory for Return to Set Point

$$\begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \\ \hat{x}_3 \\ \vdots \\ \hat{x}_K \\ \vdots \\ \hat{x}_N \end{bmatrix} = \begin{bmatrix} x_0 \\ x_0 \\ x_0 \\ \vdots \\ x_0 \\ \vdots \\ x_0 \end{bmatrix} + \begin{bmatrix} d_1 \\ d_2 \\ d_3 \\ \\ d_K \\ \\ d_N \end{bmatrix} \Delta u + \begin{bmatrix} p_1 & 0 & 0 & \cdots & 0 \\ p_2 & p_1 & 0 & \cdots & 0 \\ p_3 & p_2 & p_1 & & \vdots \\ \vdots & \vdots & & \ddots & \vdots \\ p_K & p_{K-1} & & p_2 & p_1 \\ \vdots & \vdots & & & \vdots \\ p_N & p_{N-1} & \cdots & \cdots & p_{N-K+1} \end{bmatrix} \begin{bmatrix} \Delta m_0 \\ \Delta m_1 \\ \Delta m_2 \\ \vdots \\ \Delta m_{K-1} \end{bmatrix} \quad (15\text{-}19)$$

$$\hat{\mathbf{x}} = \mathbf{x_0} + \mathbf{d}\Delta \mathrm{u} + \mathbf{P}\Delta\mathbf{m} \quad (15\text{-}20)$$

$$\mathbf{e_0} = \mathbf{x_{SP}} - \mathbf{x_0} - \mathbf{d}\Delta \mathrm{u} \quad (15\text{-}21)$$

Summary Diagram

A diagram showing in detail the computations made in one calculation cycle and the effect on the memory locations holding the data vector $\hat{\mathbf{x}}$ is shown in Figure 15-8. This diagram starts with the status of $\hat{\mathbf{x}}$ at the end of one calculation cycle, then proceeds to the beginning of the next calculation cycle and on through the completion of that cycle.

❖ UNCONSTRAINED MPC FOR MIMO PROCESSES

Conceptually, it is a simple matter to scale up from the SISO process used in the previous section to a process that has multiple inputs, multiple outputs, and multiple disturbances, as shown in Figure 15-9.

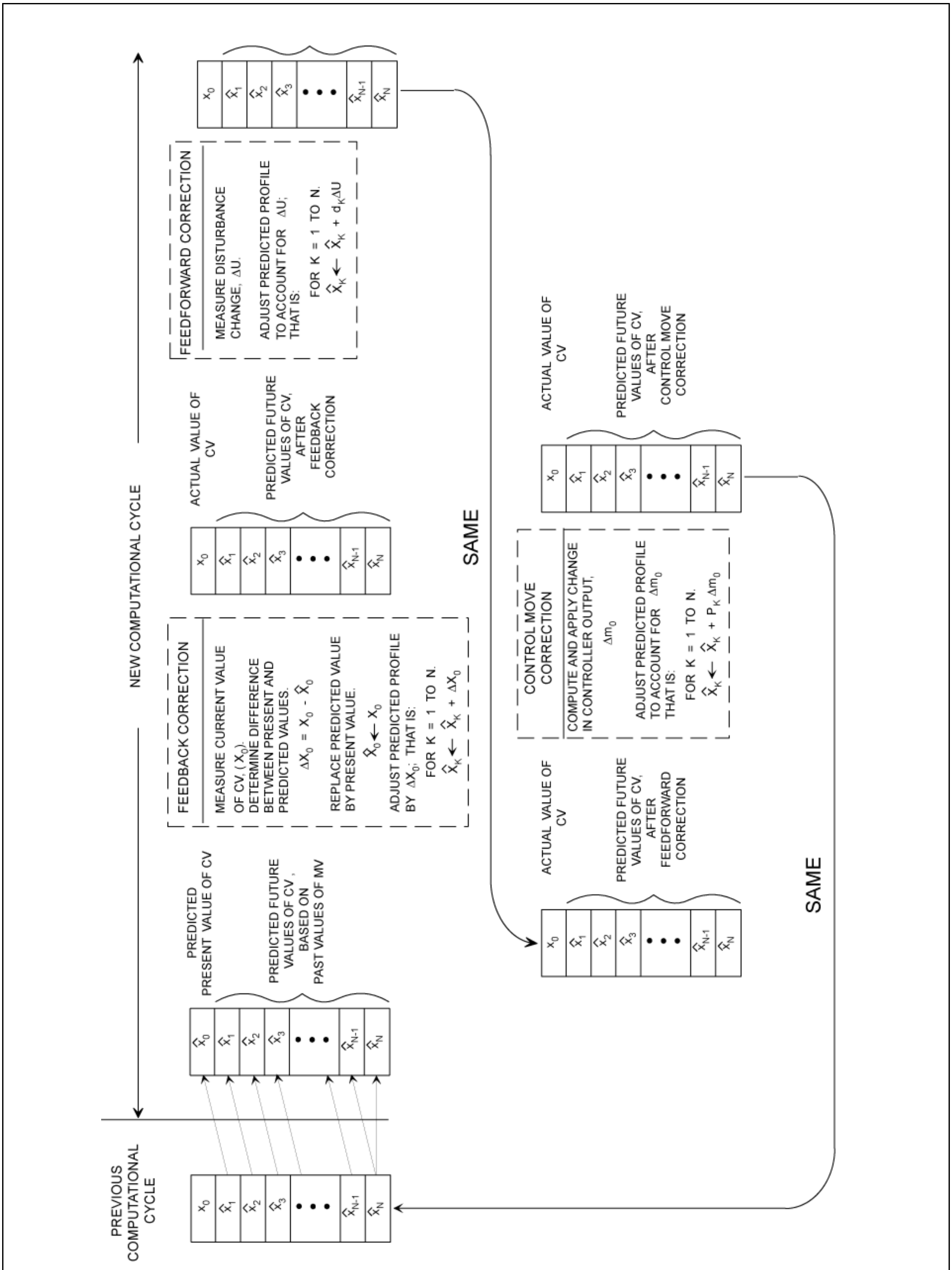

Figure 15-8. Stored Values and Computation during a Typical MPC Sample and Calculation Cycle

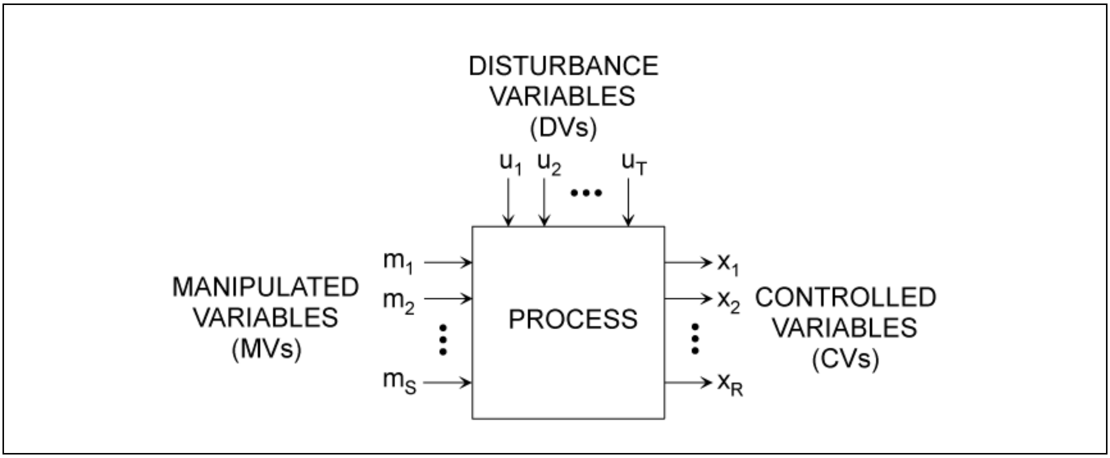

Figure 15-9. Multiple-Input, Multiple-Output Process

In this section, we shall assume that

R = number of *CVs*
S = number of *MVs*
T = number of *DVs*

In practice, R may equal S; that is, the process control system may be a "square" system. We will continue to let K represent the number of sample instances in our control horizon and N the number of sample instances in our prediction horizon.

From each *MV* to each *CV* there will be a step-response model similar to that shown in Figure 15-1. (Some may be null; that is, not every *MV* will affect all of the *CVs*.) These models are designated $\mathbf{p}_{ij}$, where subscript "i" represents "to *CV*" and subscript "j" represents "from *MV*." Hence,

For $i = 1,\ldots,R$; $j = 1,\ldots, S$

$$\mathbf{p}_{ij} = \begin{bmatrix} p_{ij,1} \\ p_{ij,2} \\ \vdots \\ p_{ij,N} \end{bmatrix} \qquad (15\text{-}22)$$

and

$$\mathbf{P}_{ij} = \begin{bmatrix} p_{ij,1} & 0 & 0 & \cdots & 0 \\ p_{ij,2} & p_{ij,1} & 0 & \cdots & 0 \\ p_{ij,3} & p_{ij,2} & p_{ij,1} & & \vdots \\ \vdots & \vdots & & \ddots & \vdots \\ p_{ij,K} & p_{ij,K-1} & & & p_{ij,1} \\ \vdots & \vdots & & & \vdots \\ p_{ij,N} & p_{ij,N-1} & \cdots & \cdots & p_{ij,N-K+1} \end{bmatrix} \tag{15-23}$$

From each *DV* to each *CV* there will be a similar step-response model (some may be null), designated $\mathbf{d}_{ik}$, where subscript "i" represents "to *CV*" and subscript "k" represents "from *DV*."

For $i = 1,\ldots, R$; $k = 1,\ldots,T$

$$\mathbf{d}_{ik} = \begin{bmatrix} d_{ik,1} \\ d_{ik,2} \\ \vdots \\ d_{ik,N} \end{bmatrix} \tag{15-24}$$

The vectors representing the current values and predicted profiles of the *CVs* are as follows:

For $i = 1,\ldots, R$

$$\mathbf{x}_{i,0} = \begin{bmatrix} x_{i,0} \\ x_{i,0} \\ \vdots \\ x_{i,0} \end{bmatrix} \qquad \hat{\mathbf{x}}_i = \begin{bmatrix} \hat{x}_{i,1} \\ \hat{x}_{i,2} \\ \vdots \\ \hat{x}_{i,N} \end{bmatrix} \tag{15-25}$$

The vectors representing future control moves are as follows:

For $j = 1$ to S

$$\mathbf{\Delta m}_j = \begin{bmatrix} \Delta m_{j,0} \\ \Delta m_{j,1} \\ \\ \Delta m_{j,K-1} \end{bmatrix} \tag{15-26}$$

Using these definitions, the predicted profile for each of the *CVs* is given by the following equation:

For $i = 1$ to R
$$\hat{\mathbf{x}}_i = \mathbf{x}_{i,0} + \sum_{k=1}^{T} \mathbf{d}_{ik} \Delta u_k + \sum_{j=1}^{S} \mathbf{P}_{ij} \Delta \mathbf{m}_j \qquad (15\text{-}27)$$

Now define the following "super vectors" (vector of vectors) and "super matrix" (matrix of matrices) (the size of each vector or matrix is indicated):

$$\hat{\mathbf{x}} = \begin{bmatrix} \hat{\mathbf{x}}_1 \\ \hat{\mathbf{x}}_2 \\ \vdots \\ \hat{\mathbf{x}}_R \end{bmatrix} \quad \mathbf{x}_0 = \begin{bmatrix} \mathbf{x}_{1,0} \\ \mathbf{x}_{2,0} \\ \vdots \\ \mathbf{x}_{R,0} \end{bmatrix} \quad \Delta\mathbf{u} = \begin{bmatrix} \Delta \mathrm{u}_1 \\ \Delta \mathrm{u}_2 \\ \vdots \\ \Delta \mathrm{u}_T \end{bmatrix} \quad \Delta\mathbf{m} = \begin{bmatrix} \Delta\mathbf{m}_1 \\ \Delta\mathbf{m}_2 \\ \\ \Delta\mathbf{m}_S \end{bmatrix}$$

(RN x 1) (RN x 1) (T x 1) (SK x 1)

$$\mathbf{P} = \begin{bmatrix} \mathbf{P}_{11} & \mathbf{P}_{12} & \dots & \mathbf{P}_{1S} \\ \mathbf{P}_{21} & \mathbf{P}_{22} & \dots & \vdots \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{P}_{R1} & \mathbf{P}_{R2} & \dots & \mathbf{P}_{RS} \end{bmatrix} \quad \mathbf{D} = \begin{bmatrix} \mathbf{d}_{11} & \mathbf{d}_{12} & \cdots & \mathbf{d}_{1T} \\ \mathbf{d}_{21} & \mathbf{d}_{22} & \dots & \vdots \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{d}_{R1} & \mathbf{d}_{R2} & \cdots & \mathbf{d}_{RT} \end{bmatrix}$$

(RN x SK) (RN x T)

Analogous to Equations 15-20 and 15-21 we have:

$$\hat{\mathbf{x}} = \mathbf{x}_0 + \mathbf{D}\Delta\mathbf{u} + \mathbf{P}\Delta\mathrm{m} \qquad (15\text{-}28)$$

$$\mathbf{e_0} = \mathbf{x_{SP}} - \mathbf{x_0} - \mathbf{D}\Delta\mathbf{u} \qquad (15\text{-}29)$$

After minimizing the sum of squares of the errors, the vector of current and future control moves is given by:

$$\Delta\mathbf{m} = \mathbf{W}\mathbf{e_0} \qquad (15\text{-}30)$$

where
$$\mathbf{W} = \left[\mathbf{P}^{\mathrm{T}}\mathbf{P} + \mathbf{Q}\right]^{-1} \mathbf{P}^{\mathrm{T}} \qquad (15\text{-}31)$$

Note that not every element of $\Delta\mathbf{m}$ needs to be computed; only the current move for each of the *MV*s is required. Therefore, we can compute:

For $j = 1$ to S

$$\Delta m_{j,0} = \mathbf{w}_j^T \mathbf{e_0} \tag{15-32}$$

where $\mathbf{w}_j^T$ is the j^{th} row vector of the matrix **W**.

❖ CONSTRAINED MPC

In real-world applications, there will be constraints on process variables, manipulated variables, and auxiliary variables. There may also be constraints on the rate of change of these variables. Furthermore, some of the constraints may be hard constraints; others may be soft. Hard constraints are established by physical limits and include valves that cannot go beyond saturation limits or controllers whose set point cannot be moved outside the measured range. Soft constraints are based on process design, equipment limits, and safety considerations. An example of a soft constraint is the tube temperature limit used in chapter 12 in the discussion of override control. As long as there is a feasible solution (i.e., an operating point) that satisfies all constraints, then hard and soft constraints can be treated equally.

However, suppose no feasible solution satisfies all constraints simultaneously. For example, returning to the process heater example used in chapter 12, suppose there is an upper limit on tube temperature and a lower limit on fuel gas-flow rate. Suppose also that there is an operating condition that tends to cause the tube temperature to rise above the limit when the fuel gas flow rate is already at a minimum. These two limiting conditions cannot both be satisfied. If we maintain minimum fuel flow, the tube temperature will exceed the limit. If we reduce fuel to achieve the tube temperature limit, we will be operating below the minimum fuel limit.

Both of these limits are soft constraints. One strategy is to permit each of these constraints to be violated by a small amount. But rather than leave it up to the process operator to make an ad hoc decision regarding how much each constraint can be violated, it may be preferable to have a “graceful violation” of each limit in a planned fashion (perhaps up to some other hard limit).

First, assume that there is a feasible solution that will satisfy all the constraints. The objective will be to minimize the functional J, subject to constraints:

$$\min_{\Delta m} J = \hat{\mathbf{e}}^T\hat{\mathbf{e}} + \mathbf{\Delta m}^T\mathbf{Q}\mathbf{\Delta m} \tag{15-33}$$

Subject to:

$$\mathbf{X_L} \le \hat{\mathbf{x}} \le \mathbf{X_U}$$
$$\mathbf{M_L} \le \mathbf{m} \le \mathbf{M_U}$$
$$\mathbf{Y_L} \le \hat{\mathbf{y}} \le \mathbf{Y_U}$$

where $\hat{\mathbf{y}}$ refers to the predicted value of auxiliary variables and the subscripts “L” and “U” refer to lower and upper limits for each class of variable. If there are rate-of-change limits, these should also be included in the constraint set.

A problem of the type that Equation 15-33 describes is called a "quadratic programming" (QP) problem (Ref. 15-8). There are standard iterative techniques for solving problems of this type; most MPC vendors include a QP solver within their software package.

If there are no feasible solutions, then the proper course of action is process-dependent. One approach is to assign an additional cost to the encroachment of each soft constraint, and then to minimize this cost. MPC vendors typically provide tools so the user can design the proper strategy in the event the normal control solution is not feasible.

It is generally recognized that constraints are dynamic. That is, they shift with time and operating conditions. Furthermore, the most economic operating condition is normally at a constraint, or at the intersection of two or more constraints. Hence, one economic benefit of MPC is its ability to push the operating point to the constraints.

Simple override control (see chapter 12) had a similar objective for a fixed, small number of constraints. However, override control can take action only after a constraint is encountered. MPC can predict in advance that a constraint is going to be encountered and begin to take "evasive" action.

❖ VARIATIONS IN MPC VENDOR OFFERINGS

The previous sections presented the general technology of multivariable MPC. The decreasing cost of computing resources, technological developments in both industry and academia, and the dissemination of knowledge about MPC have led to a significant increase in the number of vendors offering commercial MPC packages. This has naturally led to a differentiation in the available technology, though all adhere to the basic concepts (Refs. 15-3 and 15-12). In addition to the differentiation in technology, there is also a differentiation in the size and scope of targeted applications as well as in the ancillary services offered.

One of the major differences in technology is in the form of the process model and the requirements for obtaining it. Figure 15-1 described a step-response model. Quite similar to this is the pulse-response model, which is depicted conceptually in Figure 15-10. In fact, both the step- and the pulse-response models can be derived from each other, as verified by Equations 15-34 and 15-35.

$$
\begin{aligned}
p_1 &= h_1 \\
p_2 &= h_2 + h_1 \\
&\vdots \\
p_N &= h_N + h_{N-1} + \cdots h_2 + h_1
\end{aligned}
\qquad (15\text{-}34)
$$

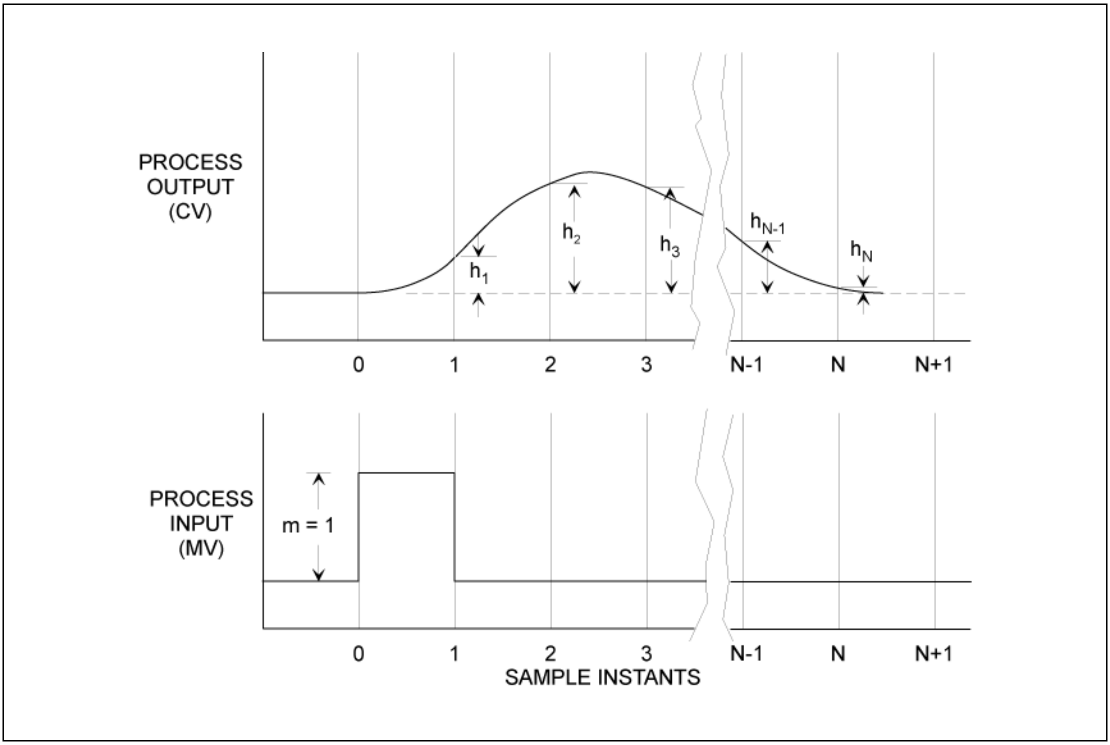

Figure 15-10. Pulse-response Model

$$\begin{aligned} h_1 &= p_1 - p_0 \\ h_2 &= p_2 - p_1 \\ &\vdots \\ h_N &= p_N - p_{N-1} \end{aligned} \qquad (15\text{-}35)$$

where **p** and **h** represent the step- and pulse-response data vectors, and $p_0 = 0$. Just as p_i is assumed to equal p_N, for $i > N$, the h_i values are assumed to equal 0 for $i > N$.

The step- and pulse-response models are called convolution models. They are used in similar ways in MPC technology. The values of the *MV*, rather than the control moves, are calculated using the pulse-response model, however.

The problem with determining convolution models directly is that the identification procedure must determine many parameters. To reduce the number of parameters required for identification, several vendors have incorporated different forms of models. Some vendors use transfer function models of the following form:

$$P(z) = \frac{a_1 z^{-1} + a_2 z^{-2} + \cdots a_n z^{-n}}{b_0 + b_1 z^{-1} + b_2 z^{-2} + \cdots b_d z^{-d}}$$

or equivalently:

$$x_t - b_0 x_{t-1} - b_1 x_t - 2 - \cdots - b_d x_{t-d} = a_1 m_{t-1} + a_2 m_{t-2} + \cdots + a_n m_{t-n}$$

This requires the identification of far fewer parameters, which consequently reduces the process testing requirements. It does, however, require that an *a priori* assumption be made about the form of the model.

One vendor approximates the process model as the weighted sum of Laguerre polynomials. Only the weighting coefficients are required by identification. With any of the reduced identification methods, the step- or pulse-response model is then calculated off line, and the remainder of the MPC technology proceeds as described.

Many vendors include proprietary identification procedures as a part of their MPC package. Often these can use either historical data or data from process tests. The HMI provided with the vendor's package will also provide graphical and statistical aids for evaluating the data and the resulting process model (see Ref. 15-2 for a discussion of statistical evaluation).

Other vendors take an entirely different approach to process modeling. One vendor uses neural networks as the basis for process modeling. One vendor offers "model-free" predictive control. Fuzzy logic has also been proposed as a basis for predictive modeling.

Another distinction between vendors lies in how the control law is calculated. In our discussion in this chapter, we incorporated a selectable time-constant reference trajectory. However, some vendors use another approach in which they establish one or more future *coincidence points.* The control moves are calculated so as to force the predicted profile to pass through these coincidence points.

Although not a part of MPC technology per se, most vendors also include a steady-state economic optimization procedure for calculating the set points for the MPC program.

❖ MPC IN PERSPECTIVE

As we mentioned previously, MPC can encompass feedback, feedforward, decoupling, and constraint control. For many processes, these functions could also have been accomplished using a control strategy that was based on the extensive use of function blocks, as presented in earlier chapters. The control systems engineer must consider many facets of each approach. Table 15-1 presents some subjective comparisons of the two approaches. See also Ref. 15.1 for an excellent discussion on assessing the benefits of MPC and preparing for it.

Table 15-1. Subject Comparison of Advanced Regulatory Control (via Function Blocks) and Model Predictive Control

Advanced Regulatory Control	Model Predictive Control
Can handle feedback, feedforward, decoupling, and constraints for relatively simple dynamics.	Can handle feedback, feedforward, MIMO, and competing constraints for almost any dynamics, including ill-behaved and nonlinear processes.
Can recognize constraints when they are hit and take control action to avoid further encroachment.	Can predict consequences of past control action and adjust current and future actions to avoid hitting constraints.
Does not handle competing constraints where there is no feasible solution.	Can permit prioritized encroachment on soft constraints when there is no feasible solution that satisfies all constraints.
Uses standard library of control algorithms.	Typically uses proprietary, software-intensive shell program.
Requires engineering-intensive handcrafting of control strategy, using function blocks.	Once the shell program is established and the process models are obtained, fairly easy to implement the control strategy.
Process models can be obtained by process testing or estimation. Can be fine-tuned after startup.	Process models obtained by process testing or from historical data. Most vendor packages provide separate programs for identifying models, as well as graphical aids for assessing process model quality. Can be fine-tuned after startup.
Tuning the feedback control follows established and well-known procedures. Advanced regulatory control can be fine-tuned by considering physical phenomena.	Tuning capabilities are more limited, and the technique for tuning is typically more obscure.
May or may not provide operators with a “feel” for the way the process and control strategy work.	Does not provide the operator with a “feel” for the process and control strategy work.
Users with moderate skill level can partially disable control strategy and apply manual intervention if something goes wrong.	May require high skill level to support operations if something goes wrong. Usually has safety features that revert, either manually or automatically, to regulatory (i.e., PID) control if the MPC fails or is deemed inadequate.
Requires moderate skill level for ongoing technical maintenance, such as adjusting for new operating conditions.	Requires high skill level for ongoing technical maintenance, such as adjusting for new operating conditions.
Makes no attempt to determine optimum operating point. Optimum set points must be provided by some other entity.	Often includes a steady-state optimizer as a part of a vendor’s package.

❖ REFERENCES

(Refs. 3 and 12 both contain extensive bibliographies of MPC literature. Each lists over one hundred references.)

15-1. T. L. Blevins, G. K. McMillan, W. K. Wojsznis, and M. W. Brown. *Advanced Control Unleashed.* ISA – The Instrumentation, Systems, and Automation Society, 2003.

15-2. G. E. P. Box, G. M. Jenkins, and G. C. Reinsel. *Time Series Analysis, Forecasting and Control*, 3d ed. Prentice-Hall, 1994.

15-3. E. F. Camancho and C. Bordons. *Model Predictive Control.* Springer-Verlag, 1999.

15-4. C. R. Cutler and B. L. Ramacher. *Dynamic Matrix Control – A Computer Control Algorithm,* Paper No. 51b, AIChe 86th National Meeting, April 1979.

15-5. C. E. Garcia and M. Morari. "Internal Model Control 1 - A Unifying Review and Some New Results," *Ind. Eng. Chem. Process Des. Dev.,* 21 (1982): 308-23.

15-6. C. E. Garcia and M. Morari. "Internal Model Control 2 - Design Procedures for Multivariable Systems," *Ind. Eng. Chem. Process Des. Dev.,* 24 (1985): 472-84.

15-7. C. E. Garcia and M. Morari. "Internal Model Control 3 - Multivariable Control Law Computation and Tuning Guidelines," *Ind. Eng. Chem. Process Des. Dev.,* 24 (1985): 484-94.

15-8. A. Morshedi, C. Cutler, and T. Skrovanek. "Optional Solution of Dynamic Matrix Control with Quadratic Programming Techniques (QDMC)," *ISA Natl. Meet.,* paper no. 85-0732. October 1985.

15-9. D. M. Prett and C. Garcia. *Fundamental Process Control.* Butterworths, 1988.

15-10. S. J. Qin and T. A. Badgwell. "An Overview of Industrial Model Predictive Control Technology," *Fifth International Conference on Chemical Process Control,* AIChE and CACHE, 1997: 232-56.

15-11. J. Richelet, A. Rault, J. L. Testud, and J. Papon. "Model Predictive Heuristic Control: Application to Industrial Processes," *Automatica,* 14 (1978): 413-28.

15-12. J. A. Rossier. *Model-Based Predictive Control, A Practical Approach.* CRC Press, 2003.

16 OTHER CONTROL TECHNIQUES

This chapter discusses a number of useful control techniques that are not covered in previous chapters. The topics included are:

- Split-range control
- Cross-limiting control
- Floating control
- Techniques for increasing effective valve rangeability
- Time proportioning control

❖ SPLIT-RANGE CONTROL

A common application for split-range control is when a temperature control loop must at times apply heat to a process and at other times must apply cooling. This is normally accomplished by using two valves, one called the "heating" valve, the other the "cooling" valve. In the usual installation, each valve operates through one-half of the controller's output range, applying the maximum heating or cooling at the extremes of the controller's output range. At the midpoint of the range, neither heating nor cooling is applied.

Figure 16-1 shows a traditional installation using pneumatic instrumentation. The 3–15 psi controller output goes to both valves in parallel. From 3–9 psi (0 – 50%), the cooling valve strokes from open to closed. From 9–15 psi (50 -100%), the heating valve strokes from closed to open. If valve-spring ranges alone are used to achieve this schedule, then the heating valve is selected to be fail-closed and the cooling valve to be fail-open.

Alternatively, valve positioners can be used, rather than relying on the valve-spring ranges to achieve the schedule. Not only will this provide more accurate and precise range splitting, but it will also make the overall design of the installation more flexible. For instance, if the application requires that both valves be fail closed and also operate in split-range fashion during normal operation, then one of the positioners can be reverse-acting, as shown in Figure 16-2.

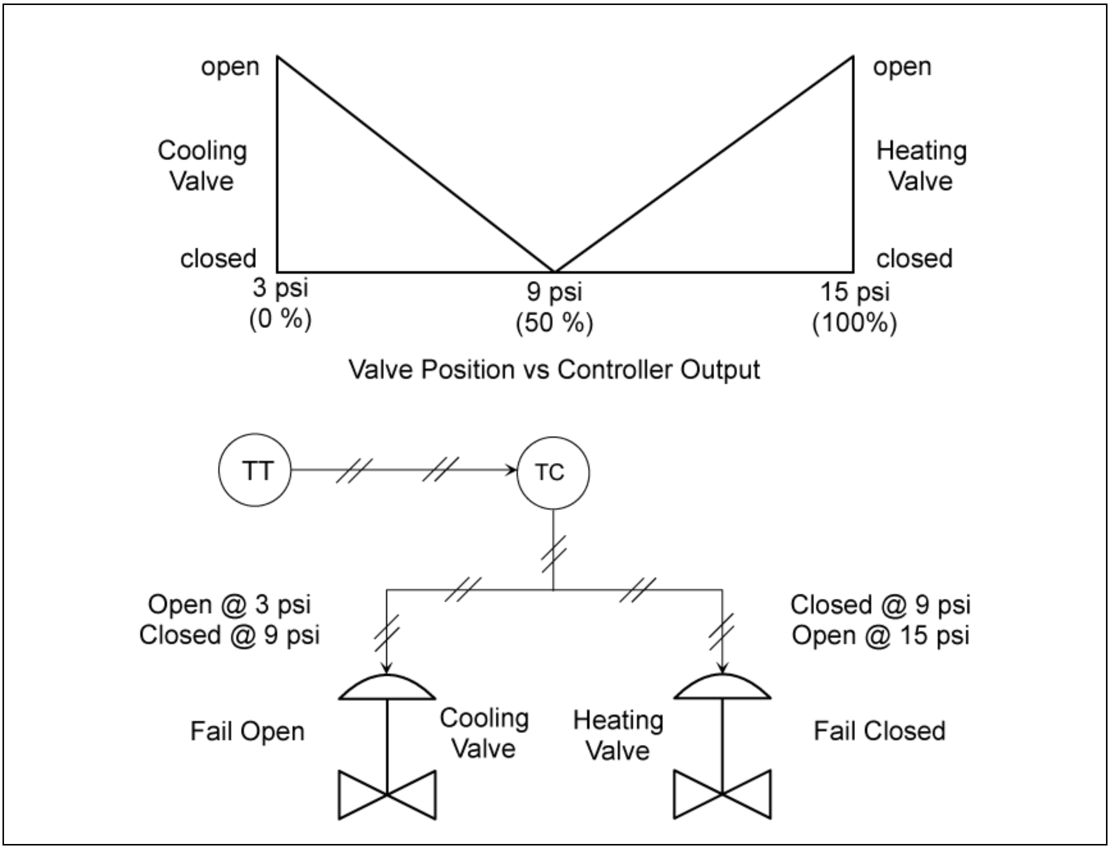

Figure 16-1. Typical Split-range Control Application

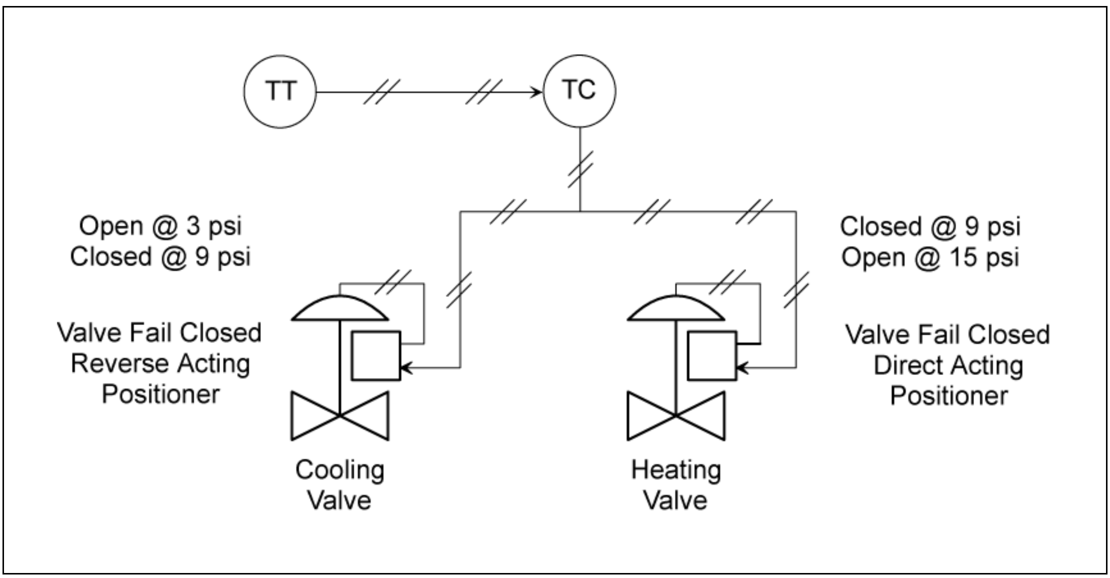

Figure 16-2. Split-range Control with Positioners and Fail-closed Valves

Both Figures 16-1 and 16-2 are based on the traditional method of designing split-range applications and on the use of traditional hardware. Let us rethink both of these.

First, consider the way the controller output signal is split. Many control engineers have never considered anything other than a 0–50 and 50–100 split. But suppose the "strength" of the heating valve's control effect is much greater than that of the cooling valve. For example, suppose the maximum flow rate through each valve is the same, but the temperature differential between the heating fluid and the process is much greater than the temperature differential between the cooling fluid and the process. In this case, the *process gain* when the controller output is in the heating valve range is much greater than when it is in the cooling valve range. If "one-size-fits-all" tuning is used, then the temperature controller will have to be tuned for the worst case, that is, for the temperature range when the process gain is the highest. In the cooling range, the control loop may be too sluggish.

A preferred method for scheduling the valve operation would be to shift the split point toward the cooling side. For example, let the cooling valve operate from 0% (valve open) to 33% (valve closed) and the heating valve from 33% (valve closed) to 100% (valve open). This will reduce heating range process gain by 25% and increase the cooling range process gain by 50%. Some other split point may be chosen experimentally. The objective is for the process gains to be approximately the same for both the heating and cooling regime, and for transition between the heating and cooling regimes to be transparent to the control loop behavior.

In general, the split can be made in the field, using one analog output and adjustments to the valves or positioners. However, it may be very difficult, if not impossible, to shift the split point to some arbitrary value as described above. With a digital processor (DCS or PLC), however, the split can be performed in software, for instance by the use of characterizer function blocks as shown in Figure 16-3. There are significant advantages to using two analog outputs and performing the split in the software. It is much easier to adjust in software than on field equipment. Also, by not adjusting the field equipment maintenance personnel can more easily replace valves and positioners without having to readjust for the split.

There are undoubtedly other configurations which will achieve the same concept shown in Figure 16-3, depending upon the type of function blocks available in a particular processor and also upon whether or not valve positioners are used. The split could even be performed in fieldbus function blocks in the valve actuators, if a fieldbus communication architecture is used.

One further point: Some control engineers prefer that there be a slight overlap in the valve operation, as shown in Figure 16-4. The advantages of an overlap are that it insures that there will not be a dead band between operation of the valves and that it provides an artificial live load on the control loop when the valves are operating near the closed position. The disadvantage is the cost of the additional utilities as the price being paid for improved controllability in the transition region.

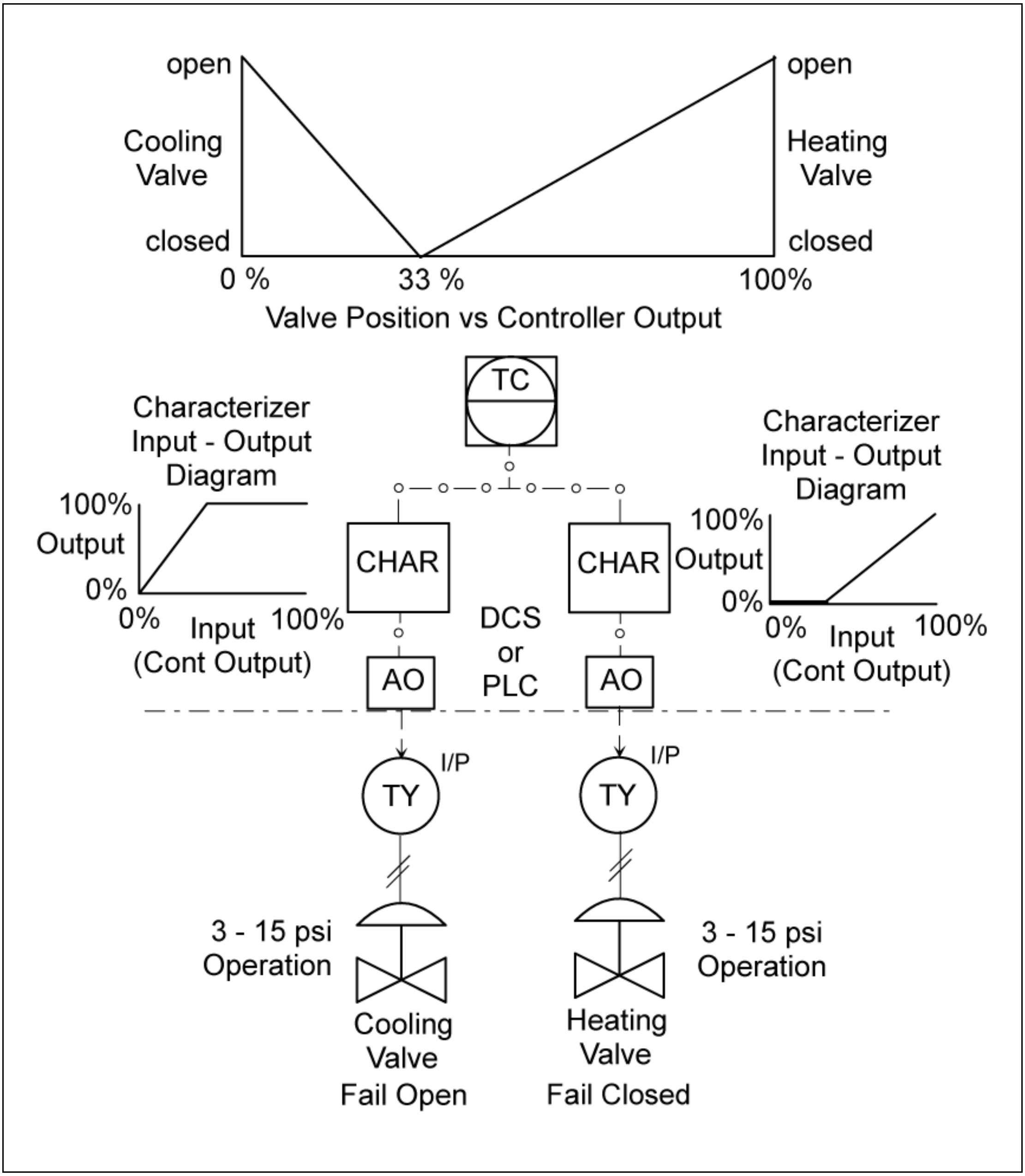

Figure 16-3. Flexible Split Range Control Implemented in a Digital Processor

❖ CROSS-LIMITING CONTROL

Cross-limiting control (often called "lead-lag" control) is a technique applied to large combustion units such as steam generators and process heaters. During steady operation, a nominal fuel-to-air ratio is maintained. This ratio can be manually set; more than likely, however, it will be set from a stack analyzer control system (see chapters 9 and 11). During transient con-

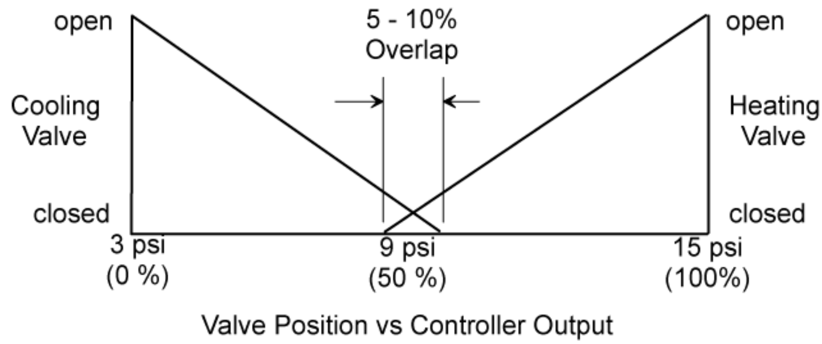

Figure 16-4. Split Range Control with Overlap

ditions, the objective is to avoid a fuel-rich situation, which could be a safety hazard. Thus on an increase in firing demand, the air flow is increased first (air "leads"). When there has been a proven increase in air, the fuel rate is increased (fuel "lags") to satisfy the air-to-fuel ratio control requirements. On a decrease in firing demand, the fuel is decreased first; when there has been a proven decrease in fuel, the air flow rate is decreased.

The configuration shown in Figure 16-5 will accomplish this objective. The scaling of the components in the output of the O_2 controller should be such that when its output is 50%, it is requesting a 1-to-1 nominal air-to-fuel ratio. Its allowable range of request should be limited. For instance, at 0% output, the required air-to-fuel ratio could be 80% of nominal; at 100% output, the ratio could be 120% of nominal.

❖ FLOATING CONTROL

Floating control refers to a technique whereby the set point of a process controller is varied automatically in response to changes in demand on the control loop. The usual objective is energy savings. This technique is closely related to sliding pressure control for steam boilers. We will first illustrate the concept for pressure control of a distillation column, and then cite several additional candidate applications. Our objective is not to provide complete details for implementation, but to provide a sufficient number of examples so as to provoke thought as to how this concept can be utilized, possibly in completely different applications.

♦ Floating Pressure Control for Distillation Columns

Traditionally, distillation columns are operated at a constant pressure, perhaps by varying the rate of condensation of overhead vapors. If there is a water-cooled condenser, then there may be a considerable variation in cooling water temperature, say, between summer and winter. If the condenser and control valve are sized for worst-case conditions, then when the cooling water temperature is high, the valve will be essentially wide open. At lower cooling water temperatures, the pressure controller will maintain the valve in a partially closed position. In other

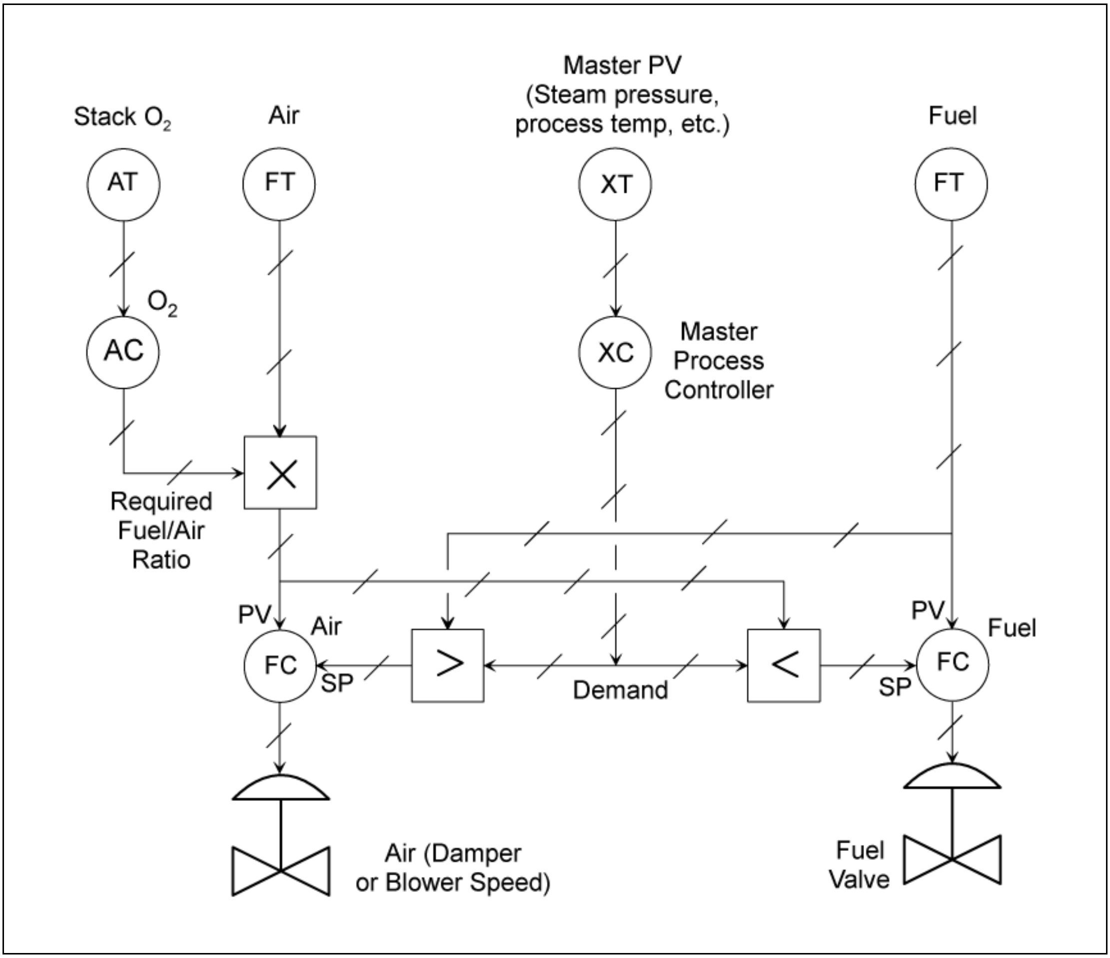

Figure 16-5. Cross-limiting Control

words, during times when the cooling water temperature is less than maximum, only a portion of the condensing capacity is being utilized.

For many hydrocarbon feedstocks, it is economically advantageous to operate at the lowest possible pressure. At lower pressures, there is an increase in relative volatility of the components so that the material can be separated to the same product specifications with less energy input (see Ref. 16-1).

One possible strategy would be to arbitrarily decide upon a nearly wide open valve position (say, 95%), then issue instructions to the operator to "gradually lower the pressure controller set point when the controller output is running less than 95%, and gradually increase the set point tends to run more than 95%." The reasons for choosing a value less than 100% are that we want to have some valve movement available for short-term pressure control, and we also want to be able to detect when there is a need to increase the set point. If this strategy were diligently followed, we would be operating at all times at the lowest pressure consistent with available condensing capacity; hence, we would be operating at minimum utility usage.

Rather than relying upon operator diligence, however, we can automate this strategy by employing another controller called a "valve position controller". Its process variable is the output of the pressure controller; its output adjusts the set point of the pressure controller. The set point of the valve position controller is the desired long-term valve position, say 95%. The valve position controller should have integral-only control action to prevent passing proportional steps back to the pressure controller. It should also be tuned for very slow response so that, in the short term, the pressure controller reacts as if it had a constant set point. These requirements, an integral action controller and tuning for slow response, are common to all floating control applications.

Figure 16-6 depicts a control loop configuration for floating pressure control.

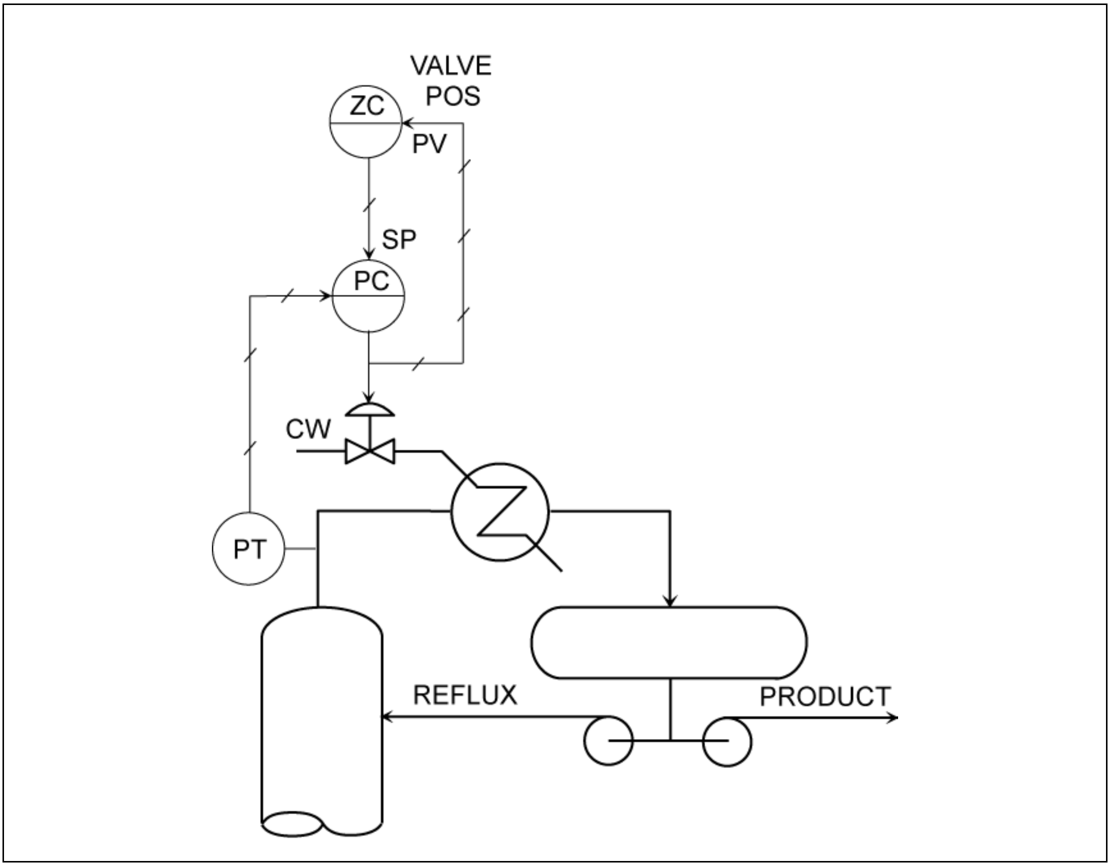

Figure 16-6. Floating Pressure Control for a Distillation Column

Two other points should be made regarding floating pressure control for distillation. The technique will only have economic benefit if the product quality for all draw streams is maintained constant. If the composition of one stream, say the bottom product stream, is not controlled, then at reduced column pressure, the purity of that stream will merely increase with no savings in utilities. The other point is that if temperature measurements are used as indicators of composition, these measurements must be pressure compensated.

♦ Floating Pressure Control for Steam Systems

If a utility steam boiler is serving a number of process users, for example, heat exchangers, reboilers and turbine-driven equipment, the steam pressure need be no higher than that required by the most demanding user. Take-off signals from all the process controller outputs can be collected and passed through a high signal selector, as shown in Figure 16-7. The output of the high selector represents the most demanding user; this becomes the process variable for a valve position controller whose output adjusts the set point of the master pressure controller. Energy conservation derives from the elimination of energy loss due to excessive throttling (see Ref. 16-1).

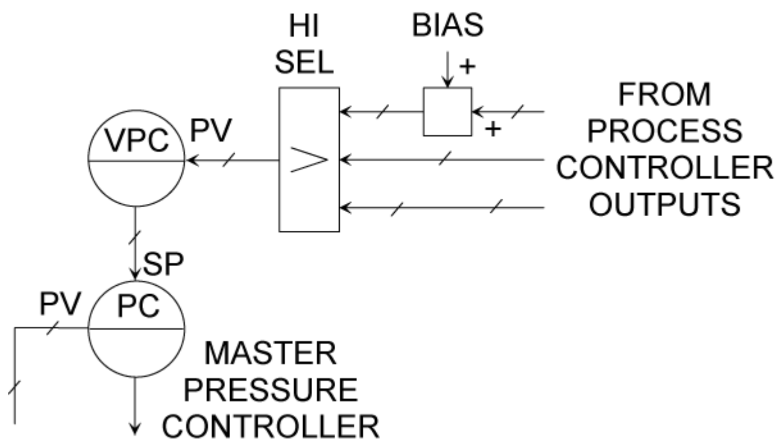

Figure 16-7. Floating Pressure Control for a Simple Steam System

The choice of set point for the valve position controller (VPC) must be made judiciously. Since the pressure controller set point will be adjusted rather slowly, there must be enough "room" above the set point of the VPC to accommodate rapid increases in load on a particular process unit. For example, if the set point for the VPC is set at 70%, then the most demanding user would have an additional 30% valve travel to accommodate short term load increases. If a few of the users are more susceptible to large load increases than the others, then the demand signal from those users can be artificially biased, thus allowing a higher set point for the valve position controller.

This concept can be extended to systems with multiple headers, where low pressure headers are provided by exhaust steam from a higher pressure header, plus supplemental steam from a pressure reducing station. The output of a reducing station pressure controller, along with the demand signal from other users is auctioneered through a high signal selector to determine the demand of the most demanding user; this becomes the PV for the VPC setting the pressure set point for the higher level header (see Figure 16-8).

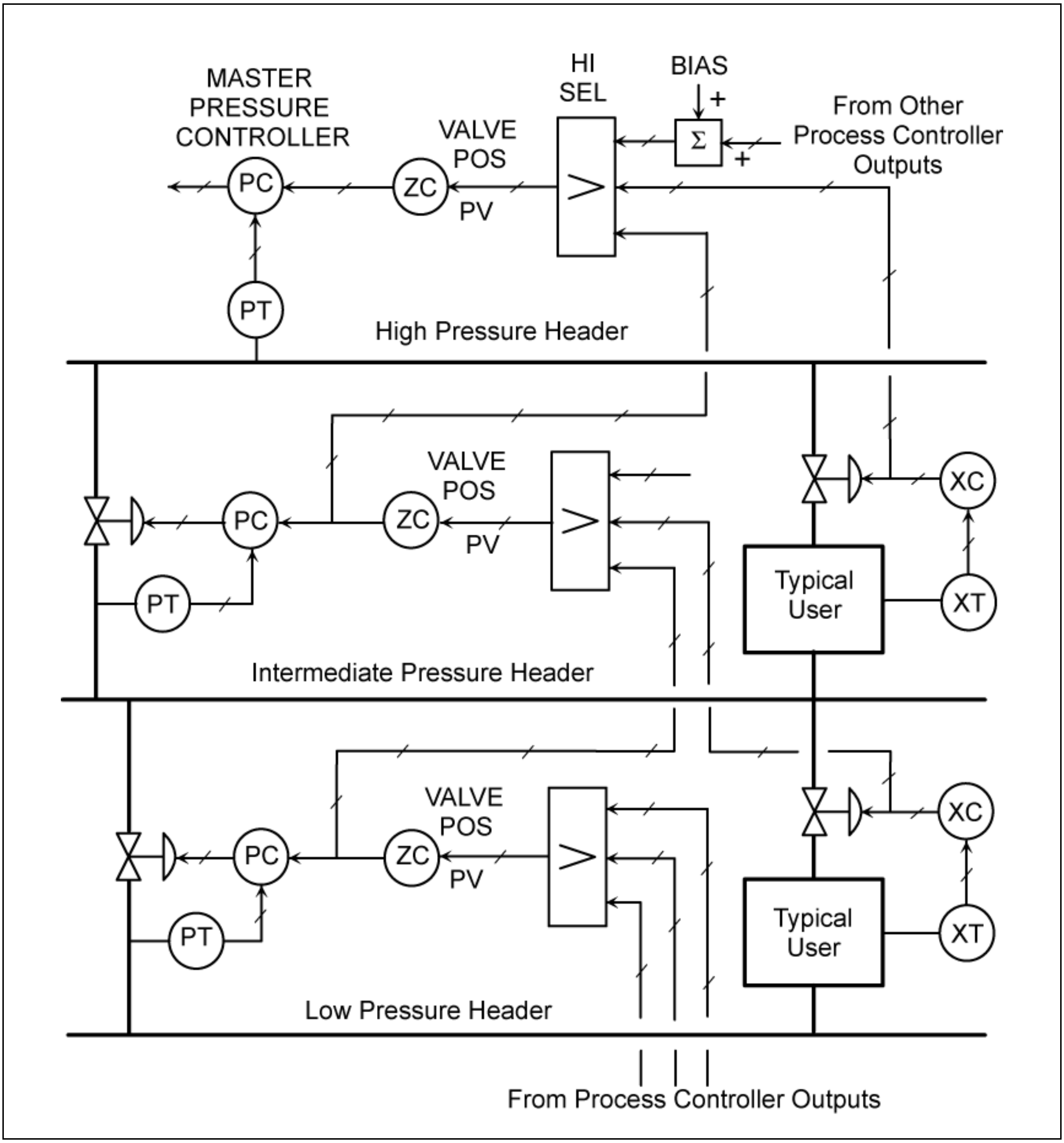

Figure 16-8. Floating Pressure Control Extended to Multiple Header Steam Systems

❖ HOT OR CHILLED WATER SUPPLY SYSTEMS

If a central utility supplies heated (or chilled) water, it will be more economical if the water is supplied at a temperature no higher (or lower) than that required by the most demanding user.

A floating temperature control concept, very similar to the floating pressure control concept described for steam systems, can be employed. The demand signals (process controller output signals) from all users of the utility is collected and auctioneered through a high selector. The output, representing the demand of the most demanding user, becomes the PV for the valve position controller, which in turn adjusts the temperature controller set point. The same considerations apply here as stated previously:

- The VPC should be an integral only controller, and tuned for slow response.
- The set point for the VPC should be chosen judiciously so as to allow for rapid increases in demand.

❖ COOLING TOWER SYSTEMS

Lipták (Ref. 16-2, p. 1151) describes a control scheme for minimizing the sum of fan and pump operating costs. The essence of this scheme is shown, in Figure 16-9, depicts the use of both floating control and override control (see chapter 12). For this discussion, it is assumed that both the fan and pump speeds can be modulated. Fan speed is modulated by controlling the difference between ambient wet bulb temperature and cooling water supply temperature. Pump speed is normally controlled by the difference between cooling water return and supply temperatures. As subsequently described, however, the differential temperature controller can be overridden by a differential pressure controller sending supply and return cooling water pressures.

For a given ambient wet bulb temperature, an increase in cooling water supply temperature would result in a decrease in fan speed, consequently a decrease in fan operating cost. However, this increase in cooling water supply temperature would cause each of the utility users to open its temperature control valve. Consequently, a selector, selecting the most demanding user, would cause the valve position controller to raise the differential pressure controller set point, which in turn would increase the pump speed to maintain the most demanding user valve position at some desired maximum (say 90% open). This increase in pump speed results in an increase in pumping cost.

An optimization program calculates the minimum of the sum of fan and pump operating cost, and sets the optimum set points for both the approach temperature controller (difference between ambient wet bulb and cooling water supply temperature) and the range temperature controller (difference between return and supply temperature). In normal operation the range temperature controller sets the pump speed. If, however, the most demanding user's valve position becomes excessively open, the differential pressure controller will increase pump speed, thus insuring that the demand of the users is given priority over the optimization program. Additional details, as well as alternatives, may be found in Ref. 16-2.

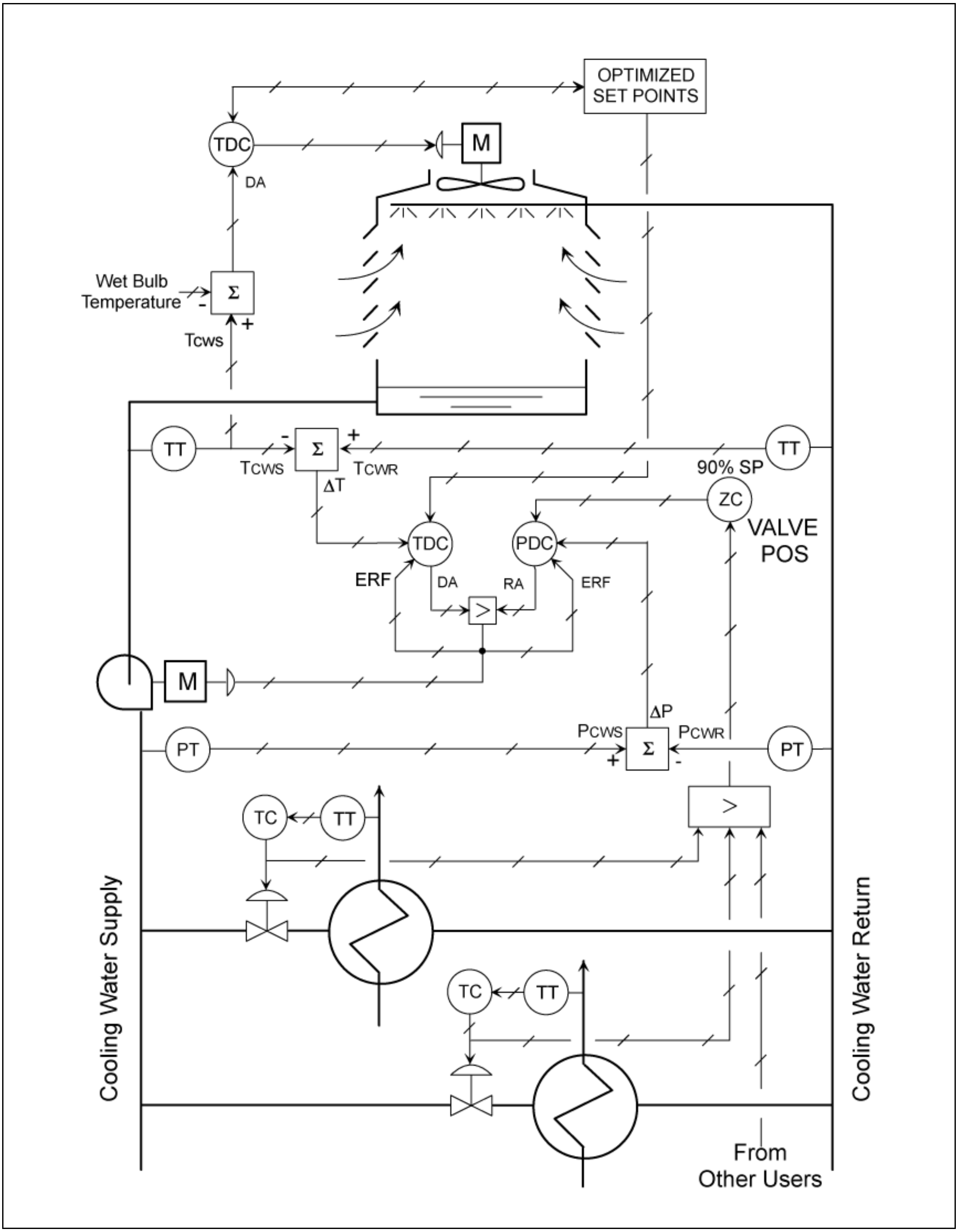

Figure 16-9. A Cooling Tower Optimization Control Scheme (Adapted from Ref. 16-2, Figure 8.10k.)

❖ INCREASING VALVE RANGEABILITY

The term "rangeability" refers to the ratio of the maximum-to-minimum C_v of a process control valve. The most common rangeability stated by manufacturers is 50. Some valves have a stated rangeability of 33, and there are special-purpose valves with stated rangeability much greater than 50. The installed characteristics of the valve, determined by other restrictions to flow in series with the valve, will cause the actual rangeability to be less than the manufacturer's stated rangeability (see chapter 3).

Occasionally the turndown requirements of the process require an effective valve rangeability greater than can be provided by a standard control valve. Two possible solutions to this problem include:

- Install a small valve and a large valve to operate in parallel
- Install a small valve and a large valve to operate in sequence

♦ *Small and Large Valves Operating in Parallel*

Figure 16-10 depicts a valve configuration in which a small valve and a large valve are used in parallel. The small valve is used for moment-to-moment control of the process. The large valve, operated by a valve position controller, attempts to keep the small valve within its operating range by increasing (decreasing) its position whenever the small valve approaches the upper (lower) end of its range. The valve position controller is an integral-only controller with a set point of 50%. In addition to the integral-only control feature, it is recommended that the control algorithm should have a gap (see chapter 4), centered on either side of 50%, in which the effective error is zero. Reasonable limits for the gap are 20% and 80%. If the small valve position is within these limits, the large valve remains stationary; if the position of the small valve encroaches on either of these limits, then the effective error will become non-zero; the integral action of the valve position controller will cause the large valve to move gradually. In doing so, the action of the process controller will be to move the small valve back to within the limits, thus stopping the large valve at the new position until the small valve position again encroaches on a limit.

The purpose of the gap is to permit the small valve to have a reasonably wide range of travel without causing a hunting movement of the large valve. If the digital processor used for controls strategy implementation does not have a standard gap action algorithm, the action can be duplicated by configuration of other function blocks. An alternative method of implementation would be to use pressure switches, set at the limits and with low differential. The pressure switches could energize solenoid valves which, operating through supply and bleed restrictions, would cause the large valve to slowly open or close. This would be an especially attractive scheme if the large valve were operated by a hydraulic cylinder, rather than an air-operated actuator.

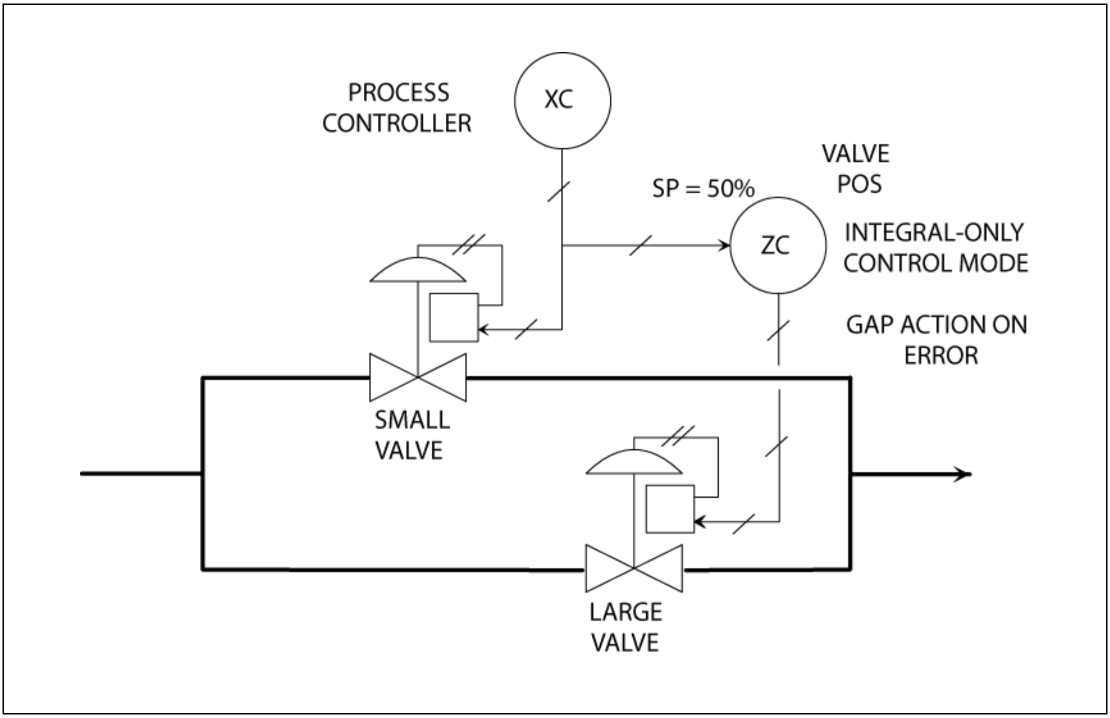

Figure 16-10. Parallel Valve Operation for Increased Rangeability

♦ Small and Large Valves Operating in Sequence

If the requirements call for equal percentage (see chapter 3) valve characteristics with wider rangeability than can be provided by a standard control valve, then two equal percentage valves, one large and the other small, can be operated in sequence. With a standard rangeability of 50, two valves together can theoretically provide a maximum rangeability of 2500. In practical applications, it is rarely necessary, nor desirable, to go to that extreme.

Suppose an application requires an equal percentage valve with a rangeability that is considerably in excess of that available in a standard control valve. A large valve is chosen with a maximum C_v exceeding the maximum of the application requirement. A small valve is then chosen so that its minimum C_v (maximum C_v divided by its rangeability) is less than the minimum C_v of the application requirement. There should be a significant overlap between the minimum C_v of the larger valve and the maximum C_v of the smaller valve.

For example, suppose an application requires a maximum C_v of 650 and a rangeability of 500. For the chosen family of valves, suppose that the nearest size which includes the required C_v of 650 has a maximum C_v of 808 and a rangeability of 50. The minimum C_v required by the application is 1.3. The small valve should have a minimum C_v less than 1.3, or a maximum C_v less than 1.3 x 50, or 65. The nearest listed size with a C_v less than 65 has a tabulated C_v of 56.7. Thus we have the ranges of C_v's shown in Figure 16-11.

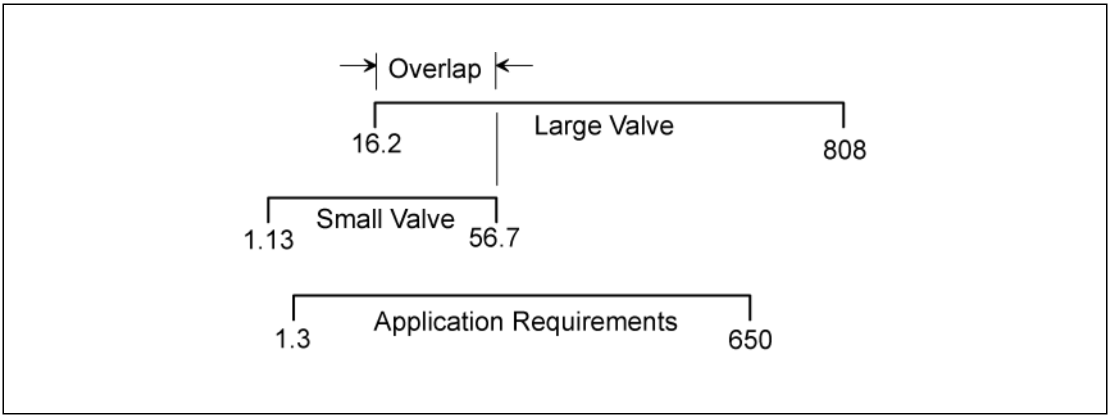

Figure 16-11. Ranges of C_V's for Application and for Valves

To determine the operating ranges for the valves, the C_V's versus valve positions can be plotted on a semi-log graph, as shown in Figure16-12. The objective is for the valves to appear to have one continuous, equal percentage operating range. To determine the extremes for the operating range for each valve, use the following equation:

$$C_V = C_{V\max} R^{m-1}$$

where: C_V is the actual value at any position of the valve;
$C_{V\,max}$ is the maximum value (tabulated value in vendor's catalog)
R is the rangeability.

For this application, the required rangeability is given by:

$$R = \frac{808}{1.13} = 715$$

Hence, calculating m from the equation above gives the following:

- The small valve should operate over a range of 0 to 59.6%, resulting in its C_V varying from 1.12 to 56.7

- The large valve should operate over a range of 40.5% to 100%, resulting in its C_V varying from 16.2 to 808.

At any value of controller output, only one of the valves should be open. As the controller output rises from 0%, the small valve should open and the large valve should remain closed until the controller output reaches to just below 59.6%. At that point switching logic should open the large valve to approximately 32.2% and close the small valve. If the switching logic were not present, the large valve would have started to open when the controller output was 40.5%. When the controller output is falling from 100%, the small valve should be closed and the large valve stroked closed until the controller output reaches just above 40.5%. At that point, the switching logic should close the large valve and open the small valve to approximately 68%. Alternatively, the lower switching point could be raised, thus avoiding operating the large valve in the lower portion of its range, a region where its positioning resolution is likely to be poor.

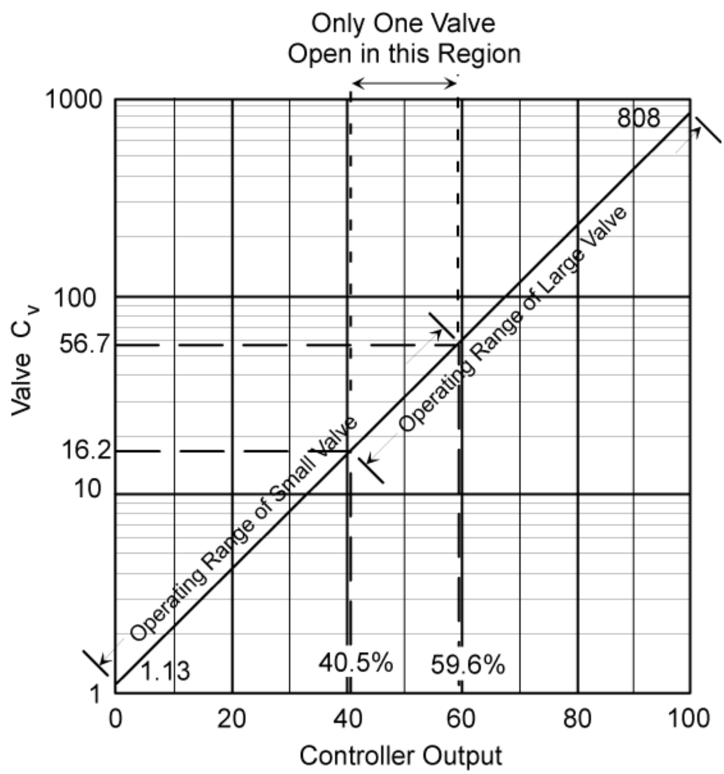

Figure 16-12. Example of Operating Ranges for Sequenced Valves

If a valve characteristic other than equal percentage is required, Ref. 16-3, p. 64, recommends configuring a characterizer function block in the output signal of the controller.

❖ TIME PROPORTIONING CONTROL

In some applications, primarily temperature control, final control is provided by an electric heater which is controlled in an on-off fashion through contactors or solid-state relays. A typical application is that of plastic extrusion, where the extruder barrel is heated by one or more zones of electric heating elements.

Rather than a simple on-off temperature controller, the controller has normal proportional-integral-derivative functions. Its output, a 0 to 100% analog (or digital modulating) signal, is then converted to a "time proportioning" signal, as shown in Figure16-13. The time proportioning signal is an on-off signal with a fixed cycle time, such as 10 seconds. The "on" time is a fraction of this time cycle, determined by the controller output's analog value, as shown in Figure 16-14. For instance, if the controller output is 36%, then the electric heaters would be on for 36% of the cycle time and off for 64%. The mass of the extruder barrel provides a filtering effect so that the heat delivery to the product is a relatively even 36% of maximum.

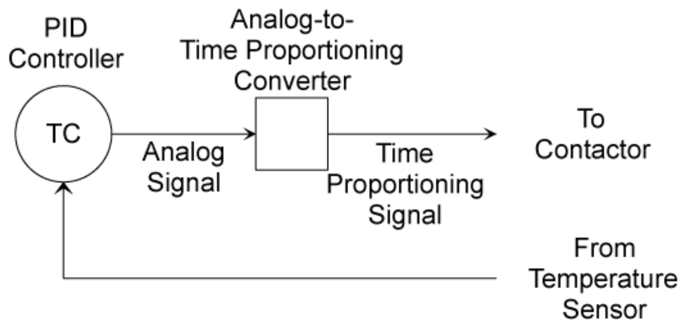

Figure 16-13. Functional Equivalent of a Time Proportioning PID Controller

In commercial time proportioning controllers, the PID controller and the analog-to-time proportioning converter may be integrated into one element, so that the analog signal is not externally accessible.

Simulation results indicate that the controller can be tuned using any of the tuning techniques discussed in chapter 6. Prior to controller tuning, however, the time cycle must be set. If the time cycle is too long, the process filtering device (for instance, the extruder barrel in plastic extrusion applications) will be unable to adequately filter the signal and a "ripple" will be produced in the measured temperature. If the process exhibits an underdamped response to a set point or disturbance, then there should be, at a minimum, 12 to 15 time cycles per process cycle.

On the other hand, there will likely be a minimum time resolution of the digital processor containing the time proportioning controller. If the time cycle is too short, the control loop will exhibit a ripple due to time resolution error. As a general guide, the time cycle should be at least 100 times as long as the minimum time resolution of the digital processor. For instance, if the cycle time of the digital processor is 100 ms, then the time cycle for time proportioning control should be 10 seconds or longer.

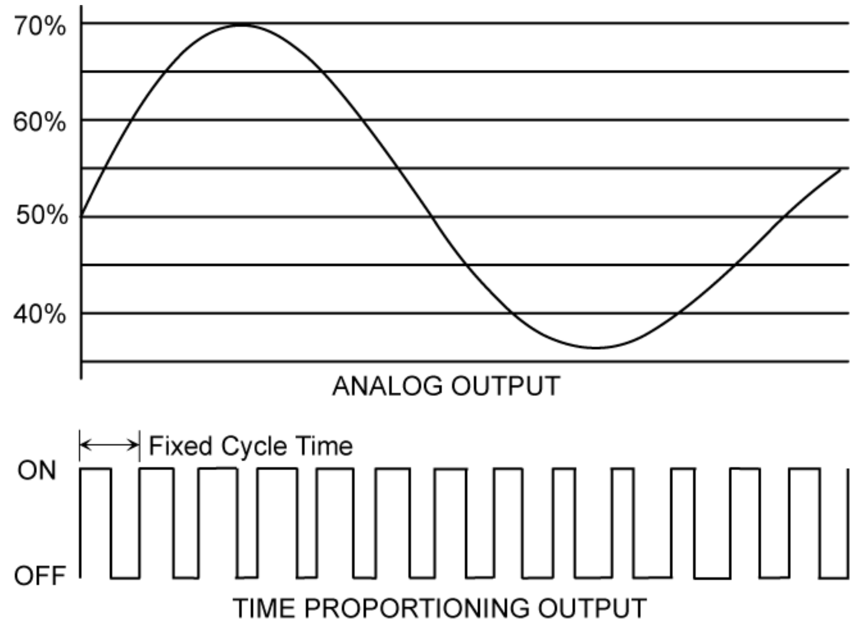

Figure 16-14. Typical Analog Output and Its Corresponding Time Proportioning Signal

❖ REFERENCES

16-1. F. G. Shinskey. *Energy Conservation Through Control.* Academic Press, 1978.

16-2. Bela G. Lipták, ed. *Instrument Engineers' Handbook. Process Control.* 3d ed. Chilton Book Co., 1995.

16-3. F. G. Shinskey. *Process Control Systems: Application, Design and Tuning.* 4th ed. McGraw Hill, 1996.

APPENDIX

A

SIGNAL SCALING

In many control schemes some type of computation using process variable or controller output signals is performed. For instance, a flow rate may be compensated for temperature, pressure or density effects, a required air-to-fuel ratio may be multiplied by a fuel rate to determine an air flow controller set point, or two temperatures may be subtracted in order to control from a differential temperature. The implementation of these computations is straightforward if the signals have been digitized, converted to engineering units and made available in floating-point format to a high-level software program. If, however, the computation is to be done with analog hardware, say by combining two or more 3–15 psig or 4–20 mA signals using discrete devices called adding relays, multipliers, etc., then the signals must be properly scaled so that the resultant signal has the correct range in engineering units. The same problem exists if the computation is performed in a digital device in which all signals are represented by a integer value within a common range. In both of these cases, we can consider that the signal has a range of 0 – 100%, or a normalized range of 0 to 1.

Quite often, the proper signal scaling values are obvious. At other times, a more formal methodology for signal scaling will be very beneficial. This appendix presents such a method which will work in all cases, although for simple scaling problems, it is perhaps an overkill.

Every analog signal has both a signal value and an interpretation in engineering units. In the case of transmitters, the meaning of the signal is fixed by the transmitter range. In other cases, we may have the freedom to choose the engineering range of the signal.

Example: Stack oxygen is sensed and transmitted by an 0_2 transmitter to an 0_2 controller. Both the transmitter output and the controller output are 0 – 100% signals. If the 0_2 transmitter is calibrated for 0 - 5% stack oxygen, then the interpretation of the signal in engineering units is obvious; 0% signal means 0% 0_2, 50% signal means 2 1/2% 0_2, etc.

The engineering units for the output of the 0_2 controller are *not* % 0_2; the engineering units for a controller's output will rarely be the same as the engineering units for its process variable. The output indicator on the front of the instrument or faceplate display may read "0 - 100%", but we have to ask "percent of what?" For a controller output, we must normally look downstream and see how the signal is used before we can determine the engineering units meaning of the signal.

If the 0_2 controller output is multiplied by a fuel flow rate to determine a set point for an air flow controller (see Figure 10-3 and 10-4) then the 0_2 controller output must represent the required air-to-fuel ratio. But what actual ratio? We have the freedom to choose the numerical range of that ratio, then enforce that by the scaling of the multiplier unit.

For a pure methane fuel, the stoichiometric ratio of air-to-fuel is about 10-to-1; that is, there are 10 standard volumetric units of air per 1 standard volumetric unit of fuel. Allowing for 10% excess air, a nominal ratio would be 11-to-1. We could choose mid-scale to represent 11, but how much variation to allow the controller output? We might choose to let a +/- 50% deviation from mid-scale represent a variation of +/- 2 ratio units, so that 0% and 100% controller output represents required air-to-fuel ratios of 9-to-1 and 13-to-1, respectively. Note that this last choice is an arbitrary one, based upon an engineering judgment as to how must variability to allow the 0_2 controller output.

Having established the engineering range for the controller output, we must now determine compatible scaling parameters for the multiplier. We will first present the general methodology for signal scaling, then return to this example and apply that method.

This scaling method is applicable to any signal media, pneumatic, analog or digital with a fixed integer range. For all of these, we will simply define a *normalized* signal to be within the range or 0-to-1.

Pneumatic:	3–15 psig
Electronic:	4–20 mA
Digital (example):	0–4000 counts
Common terminology	0–100%
Normalized value:	0–1

It will be beneficial to adopt standard symbols to represent the signal, both in the normalized value and in engineering units. We will use *uppercase letters* to represent the engineering value, and *lowercase letters* to represent the normalized value. For example, if we are speaking of a pressure transmitter output, we will use the letter "*P*" to represent the pressure in psig, and "*p*" to represent the normalized signal. For convenience, we let the symbol "$\underline{P}$" (*P* underbar) represent the low end of the transmitter span, and "$\overline{P}$" (*P* overbar) represent the upper end of the transmitter span. Obviously, for any measured value of pressure within the range of the transmitter,

$$\underline{P} \le P \le \overline{P}$$

$$0 \le p \le 1$$

Formal conversion equations between engineering units and normalized values are:

$$P = \underline{P} + p\left(\overline{P} - \underline{P}\right) \tag{A-1}$$

$$p = \frac{P - \underline{P}}{\overline{P} - \underline{P}} \tag{A-2}$$

If the transmitters are "zero-based", then $\underline{P} = 0$. In this case the equations simplify to:

$$P = p\overline{P} \tag{A-3}$$

$$p = \frac{P}{\overline{P}} \tag{A-4}$$

We are now ready to begin the step-by-step procedure for signal scaling for a computational device (or software function block) which combines several input signals.

(1) Write the equation, in engineering units, that the device or function block is to perform. If there are three input signals, *X*, *Y*, and *Z*, then this equation will be of the form:

$$M = f(X, Y, Z) \tag{A-5}$$

where *M* represents the output and *f (...)* represents some functional relationship determined by the engineering application. You may wish to substitute other symbols that are more indicative of the true process variables, such as *T*, *P*, *F*, or *L* for the symbols *X*, *Y*, *Z* and *M*.

(2) Write the equation, given in the manufacturer's literature, for the device or function block you intend to use. This will relate the normalized output signal, *m*, to normalized input signals, *x*, *y*, *z*, plus adjustable scaling parameters k_0, k_1, … This equation will be of the form:

$$m = f(x, y, z, k_0, k_1, ...) \tag{A-6}$$

If you substituted other symbols process variable-indicative symbols, such as *T*, *P*, *F*, or L, etc. for the symbols *X*, *Y*, *Z* and *M*, then you should make a similar substitution, i.e., *t*, *p*, *f*, or *l*, for the symbols *x*, *y*, *z*, and *m*.

(3) For each symbol appearing in Equation A-5, substitute an equation of the type A-1 or A-3. After this substitution, Equation A-5 should contain only symbols representing normalized values plus symbols representing signal span limits in engineering units.

(4) Rearrange the modified Equation A-5, resulting from Step 3, into the same form as the manufacturer's Equation, A-6. If the modified Equation A-5 cannot be rearranged to the same form as A-6, then the selected manufacturer's device or function block is inadequate to perform the required computation. A combination of function blocks may be required.

(5) Each of the scaling parameters in the manufacturer's equation should now be related to an expression containing only signal span limits. These expressions are the proper values for the scaling parameters.

(6) Check your results with one or more numerical cases.

Continuing our example, the 0_2 controller's output, 0–100%, represents a required air-to-fuel (volumetric) ratio of 9-to-1 to 13-to-1. Suppose the fuel flow transmitter range is 0–1000 scfm and the air flow transmitter range is 0–15,000 scfm. The multiplier multiplies the fuel rate by the 0_2 controller output and produces a signal which is the set point of the air flow controller. This signal must have the same range as the air flow transmitter, or 0–15,000 scfm.

We are speaking of three signals—air, fuel, and ratio—so we can use the symbols *A*, *F*, and *R*. The signal range data is as follows:

$$\underline{A} = 0 \qquad \overline{A} = 15{,}000$$
$$\underline{F} = 0 \qquad \overline{F} = 1{,}000$$
$$\underline{R} = 9.0 \qquad \overline{R} = 13.0$$

Applying the methodology, we have:

(1) $A = RF$ (A-7)

(The required air rate, scfm, equals the required ratio, scfm per scfm, times the fuel rate, scfm.)

(2) We choose a multiplier for which the manufacturer's literature provides the following functional equation:

$$m = k_0 (x_1 + k_1)(x_2 + k_2) + k_3 \qquad \text{(A-8)}$$

Converting to the normalized symbols relating to air, fuel and ratio, this is:

$$\begin{aligned} a &= k_0 (r + k_1)(f + k_2) + k_3 \qquad \text{(A-9)} \\ &= k_0\, r\, f + k_0\, k_1\, f + k_0\, k_2\, r + k_0\, k_1\, k_2 + k_3 \end{aligned}$$

(3) Relate symbols representing engineering units to the normalized signals and the signal range limits:

$$A = a\,\underline{A}$$

$$F = f\,\underline{F}$$

$$R = \underline{R} + r(\overline{R} - \underline{R})$$

Substitute these into Equation A-7:

$$a\,A = \left[\,\underline{R} + r\left(\overline{R} - \underline{R}\right)\right] f\,\overline{F} \qquad \text{(A-10)}$$

(4) Rearrange: $$a = \frac{\left(\overline{R} - \underline{R}\right)\overline{F}}{\overline{A}}\,r\,f + \frac{\underline{R}\,\overline{F}}{\overline{A}}\,f \qquad \text{(A-11)}$$

(5) Comparing Equations A-9 and A-11, we see that:

$$\begin{aligned} k_0 &= \frac{\left(\overline{R} - \underline{R}\right)\overline{F}}{\overline{A}} \\ &= \frac{(13.0 - 9.0)\,1{,}000}{15{,}000} \\ &= 0.267 \end{aligned}$$

$$\begin{aligned} k_0\,k_1 &= \frac{\underline{R}\,\overline{F}}{\overline{A}} \\ k_1 &= \frac{\underline{R}}{\overline{R} - \underline{R}} \\ &= \frac{9.0}{13.0 - 9.0} \\ &= 2.25 \end{aligned}$$

$$\begin{aligned} k_0\,k_2 &= 0 \\ k_2 &= 0 \end{aligned}$$

$$\begin{aligned} k_0\,k_1\,k_2 + k_3 &= 0 \\ k_3 &= 0 \end{aligned}$$

(6) Suppose that the fuel flow is 600 scfm and the 0_2 controller output is 50%, representing a demand for 11:1 air-to-fuel ratio. The multiplier calculates the normalized air flow:

$$a = 0.267\ (0.5 + 2.25)\ 0.6$$

$$= 0.44$$

Hence, the required air rate is 0.44 x 15000, or 6600 scfm, providing the required 11:1 ratio.

APPENDIX

B

DERIVATION OF EQUATIONS FOR INSTALLED VALVE CHARACTERISTICS

This appendix derives the equations used for presenting the graphs for installed valve characteristics for linear and equal percentage valves, Figures 3-6 and 3-7. We will use the following symbols:

C_V	Valve coefficient
C_L	An index of the resistance to flow offered by the line and fittings, plus the internal friction of the pump.
C_E	An equivalent index, analogous to C_V, representing the combined effects of C_V and C_L
F	Flow rate, gallons per minute
m	Valve position, percent open
R	Valve rangeability (R = 50 used for Figures 3-6 and 3-7).
β	Ratio of minimum to maximum pressure drop across valve
ΔP_V	Valve pressure drop, psi
ΔP_L	Dynamic pressure drop loss due to pipe friction and internal pump friction
ΔP_T	Available system pressure drop, psi
ρ	Fluid specific gravity.

An idealized model for a valve installation is shown in Figure B-1. Here the following assumptions are made:

(1) The supply pressure (blocked discharge pressure of pump) and discharge pressure are constant. The difference, ΔP_T, represents a constant available system pressure drop.

(2) The pressure drop due to internal friction loss in the pump and the pressure drop due to friction loss through the pipe and fittings can be combined into a single pressure drop, ΔP_L. This pressure drop varies with flow.

(3) The pressure drop through the valve, ΔP_V, also varies with flow—becoming a minimum when the valve is wide open and reaching a maximum, equal to ΔP_T, when the valve is fully closed.

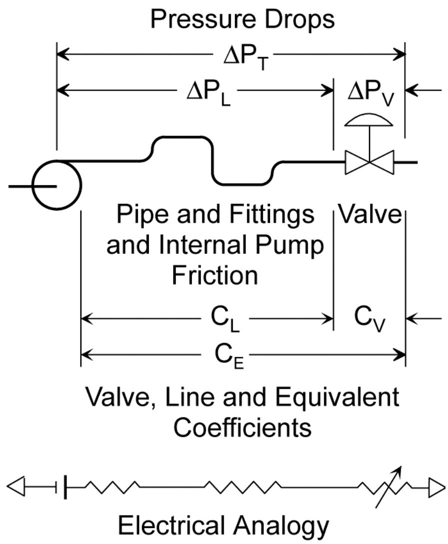

Figure B-1. Idealized Valve Installation

For incompressible fluids, the basic flow equation through a valve is:

$$F = C_V \sqrt{\frac{\Delta P_V}{\rho}} . \qquad \text{(B-1)}$$

Without loss of generality, we can assume $\rho = 1$, so that it need not appear in further equations. (If ρ were used in the derivation, in the end they would all cancel out.)

The valve coefficient, C_V, varies with valve position, from a minimum to a maximum value, $C_{V\,max}$, when the valve is wide open. $C_{V\,max}$ is the value for C_V normally tabulated in valve manufacturers' literature. The relation between C_V and valve position is called the manufactured characteristics of the valve, and varies by valve type:

For a linear valve:
$$C_V = \frac{m}{100} C_{V\,max} . \qquad \text{(B-2)}$$

For an equal percentage valve:
$$C_V = R^{\frac{m}{100} - 1} C_{V\,max} . \qquad \text{(B-3)}$$

When the valve is wide open, the flow is given by

$$F_{max} = C_{V\,max} \sqrt{\Delta P_{V\,min}} , \qquad \text{(B-4)}$$

Hence, the minimum valve pressure drop is given by

$$\Delta P_{V\,min} = \left(\frac{F_{max}}{C_{V\,max}}\right)^2. \tag{B-5}$$

For the maximum pressure drop, $\Delta P_{v\max} = \Delta P_T$. (B-6)

When the flow is maximum, the line and pump head loss is also maximum, hence

$$\Delta P_{L\,max} = \Delta P_T - \Delta P_{V\,min} \tag{B-7}$$

Let C_L, an index similar to C_V but related to the restriction to flow offered by the line, fittings and internal pump friction, be defined by the following equation:

$$F_{max} = C_L\sqrt{\Delta P_{L\,max}} \tag{B-8}$$

Note that C_L is a constant.

Also, let C_E, an index similar to C_V but which combines the effect of C_L and C_V, be defined by the following equation:

$$F_{max} = C_E\sqrt{\Delta P_T} \tag{B-9}$$

Note that C_E is a variable since C_V varies with valve position.

At any non-zero valve position $\Delta P_V = \left(\frac{F}{C_V}\right)^2,$

$$\Delta P_L = \left(\frac{F}{C_L}\right)^2$$

$$\Delta P_T = \left(\frac{F}{C_E}\right)^2$$

Since $\Delta P_T = \Delta P_L + \Delta P_V,$

then $$\left(\frac{F^2}{C_E{}^2}\right) = \left(\frac{F^2}{C_L{}^2}\right) + \left(\frac{F^2}{C_V{}^2}\right).$$

Cancel the common F^2 and solve for C_E:

$$C_E = \frac{C_L C_V}{\sqrt{C_L{}^2 + C_V{}^2}} \tag{B-10}$$

Now, introduce β, the ratio between the minimum and maximum valve pressure drops:

$$\beta = \frac{\Delta P_{V\,min}}{\Delta P_{Vmax}} \tag{B-11}$$

Since $P_{Vmax} = P_T$, then:

$$\begin{aligned}
\beta &= \frac{\Delta P_{V\,min}}{\Delta P_T}, \\
&= \frac{\Delta P_{V\,min}}{\Delta P_{L\,max} + P_{V\,min}}, \\
&= \frac{\left(\frac{F_{max}}{C_{V\,max}}\right)^2}{\left(\frac{F_{max}}{C_L}\right)^2 + \left(\frac{F_{max}}{C_{V\,max}}\right)^2}, \\
&= \frac{1}{\left(\frac{C_{V\,max}}{C_L}\right)^2 + 1}.
\end{aligned}$$

Hence $$C_L = C_{V\,max}\sqrt{\frac{\beta}{1-\beta}}. \tag{B-12}$$

Combining Equations B-10 and B-12 results in:

$$C_E = \frac{\sqrt{\beta}\left(\dfrac{C_V}{C_{V\,max}}\right)C_{V\,max}}{\sqrt{(1-\beta)\left(\dfrac{C_V}{C_{V\,max}}\right)^2 + \beta}} \tag{B-13}$$

When the valve is wide open, $C_{E\,max} = \sqrt{\beta}\, C_{V\,max}$.

Therefore

$$\frac{C_E}{C_{E\,max}} = \frac{\left(\dfrac{C_V}{C_{V\,max}}\right)}{\sqrt{(1-\beta)\left(\dfrac{C_V}{C_{V\,max}}\right)^2 + \beta}}. \tag{B-14}$$

then, by combining Equations B-9 and B-14:

$$\frac{F}{F_{max}} = \frac{C_E}{C_{E\,max}}, \tag{B-15}$$

$$= \frac{\left(\dfrac{C_V}{C_{V\,max}}\right)}{\sqrt{(1-\beta)\left(\dfrac{C_V}{C_{V\,max}}\right)^2 + \beta}}.$$

By substituting $\left(\dfrac{C_V}{C_{V\,max}}\right)$ from Equation B-2 or B-3 for the appropriate type of valve, then the fractional flow versus valve position for selected values of β.