CHAPTER 4
DEVELOPMENT OF CONTROL FUNCTION

4.1 Introduction

In this chapter, control function of the clutch actuator, according to the dynamic model created before, is built up.

In general, steps to develop a control function are firstly to understand the process and translate dynamic performance requirements into time, frequency, or pole-zero specification, and then make a linear approximate model. Base on the linear approximate model, sketch a frequency response (Bode plot) and a root locus to form an initial estimate control function of the complexity of the design model. Finally, if the trial-and-error compensators do not give entirely satisfactory performance, consider a design based on optimal control. The symmetric root locus shows possible root locations from which to select locations for the control poles that meet the response specification [10].

In this study, instead of a linear approximate model, a complete model which is more close to the real system is exercised, with an expectancy of better outcome than general process.

To deal with such problem, the dynamic model is seen as a black box. The first estimate of the control parameters can base on turning methods such as Ziegler-Nichols (Z-N) method and Internal Model Control (IMC) method, which use some strategies to forecast the system dynamic characters, and then using optimization methods to optimize the control parameters to be able to control the process more close to the setpoint. Theoretically, such manner can provide a better control performance than general process, because of the complete controlled model, a model more close to reality, is applied.

According to these processes, control function with firstly estimated control parameters is built up in this chapter. And in the next chapter, optimization will be implemented to modify the control parameters using Matlab®. Such combination of dynamic modeling, dynamic simulation, part optimization, control function design, control function optimization, and system simulation within a unit of interface are the speciality of this study.

4.2 Control Strategies

According to the dynamic model created in CHAPTER 2, the clutch actuator seems can be controlled immediately with an open loop, since it looks as though that the dynamic model has included all the outside affections on the clutch actuator, the clutch loadings. However, in reality, vibration, heat affection, rotation speed, etc. may affect the clutch resistance force that applies on clutch actuator as an outside loading. The dynamic model just gives a general case of operations. Thus, a feedback control is required absolutely for a reliable control.

To define the control problem, ascribing the affections from engine, transmission shaft, vibration, heat affection, rotation speed, etc., which work upon the clutch, as outside loadings of the clutch/clutch-actuator system, the clutch/clutch-actuator system can be seen as an independent dynamic system to be controlled.

To formulate the problem, first, the control signal of the electrical motor is chosen to be input of the dynamic system, or output of the control function. The DC motor used on the clutch actuator is generally controlled by Pulse-Width Modulation (PWM). Such frequency control with a fixed input voltage can be seen as a transformation from voltage control. To be convenient, the DC motor control signal is chosen to be voltage control. Second, for output of the dynamic system, or feedback signal of the control function, the best choice is the torque transmit ability of the clutch T_{cf}, which is a final output of the clutch/clutch-actuator system.

However, such parameter is not easy to be measured in practice, so does the normal force N_{clu} which has a direct relation with T_{cf}. Thus, a most reliable choice for the dynamic system output, or feedback of the control function, is travel position of the clutch, or travel position of the clutch actuator D_{out}, which is also practical in general AMT clutch control.

However, control of such dynamic system with a single loop is not enough. In general, there should be one more loop for the control of the driving motor. According to Eq.(2.2-3), motor rotation angle θ_m, which has an immediate relation with system output D_{out}, is controlled directly by armature current i_a. Since the control input is voltage of the motor v_a, the transformation from v_a to i_a can be seen as another control loop with a plant gain as Eq.(2.2-5), and suffers outside loadings from motor speed $\dot{\theta}_m$, as stated in Eq.(2.2-5), and other environment effects.

The block diagram of the control loop is shown in Figure 4.2-1.

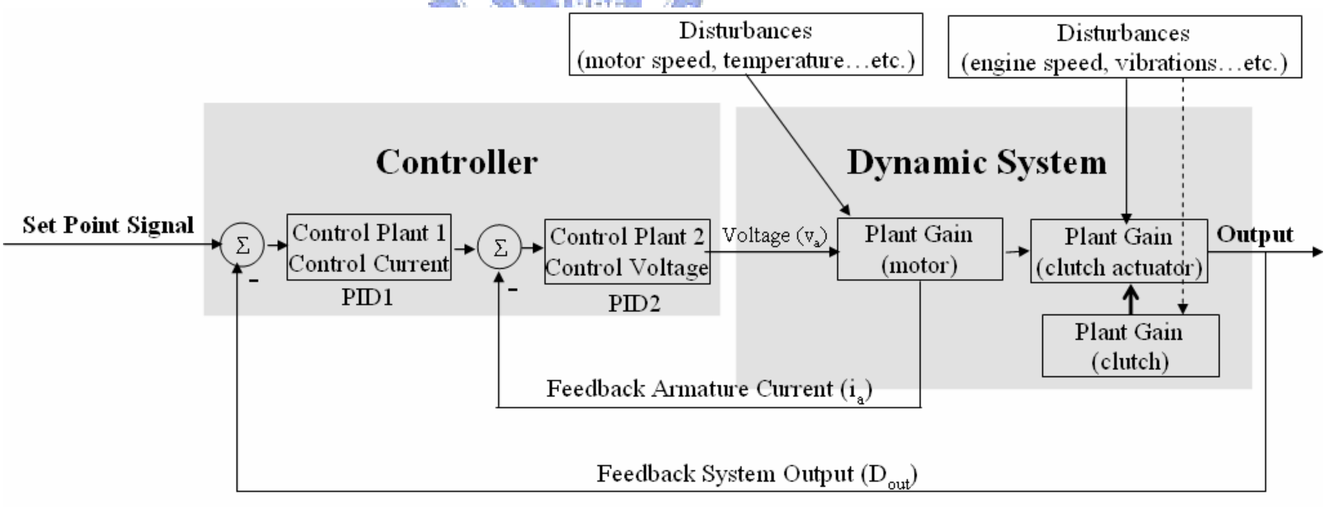

Figure 4.2-1 Control Loops

For the strategy of the control plants, Proportional-Integral-Derivative (PID) control method is used. There are many methods to control a feedback loop, such as PI control, PD control, PID control, and many other sophisticated control methods. In general, PI controller and PID controller are most practical in commercial, and thus with lower costs. PI control is adequate for all processes where the dynamics are essentially of the first order. If the step response of the dynamic system looks like that of a first-order system or, more precisely, if the Nyquist curve lies in the first and the forth quadrants only, then PI control is sufficient. PID control is sufficient for process where dominant dynamics are the second order. And it is beneficial when tight control of a higher-order system is required. It can also speeds up the transient response of the system. A sophisticated control, a higher order control, can provide high quality to the controlled system, however, such strategy always be of a significant economic value. Since clutch control of AMT system requires a fine transient response, a low cost for commercial, and dynamic system of the clutch/clutch-actuator is a high order system more than two, as Eq.(2.3-3) and Eq.(2.3-4), PID control is chosen to be the control method. Thus, in the control function, the two control plants, as shown in Figure 4.2-1, are both dominated by PID control strategy.

The PID controller was first described by Callender et al. (1936). This technology was based on extensive experimental works and simple linearized approximations to the system dynamic. It led to standard experiments suitable to application in the field and eventually to satisfactory “turning” of the coefficients of the PID controller. It also results in the development of a comprehensive set of technologies for the design of servomechanisms, which is a control method that focuses on the control of output position as in this study [10]. The standard form, or ISA form, of PID algorithm is shown in Eq.(4.2-1), where $y_{sp}(t)$ is the set point, $y(t)$ is the real output, which is also feedback signal of the control function, $e(t)$ is the control error, $u(t)$ is output of the control function or input of the dynamic system, and K, T_i, and T_d are the three main parameters of PID control.

$$u(t) = K(e(t) + \frac{1}{T_i}\int_0^t e(\tau)d\tau + T_d \frac{de(t)}{dt}) \qquad (4.2\text{-}1)$$
$$e(t) = y_{sp}(t) - y(t)$$

The three main parameters in PID control, K, T_i, and T_d, are ratios that adjust the control error, time integral of the error, and the time rate of change of the error. The increase of the control error, which means increase of K, is effective in reducing the error of the system. However, an over-large K may typically leads to instability. The increase of integral of the error, which means decrease of T_i, is effective in reducing or eliminating constant steady-state errors, but this benefit typically comes with the cost of worse transient response. The increase of differential of the error, which means increase of T_d, can improve the stability of the system, or physically means increase of the damping effect. Such property of differential is able to "predict" the system output. However, as the physical meaning, a larger T_d will leads to a slower system response, thus a longer rise time. The block diagram in Matlab® of such standard form is shown in Figure 4.2-2.

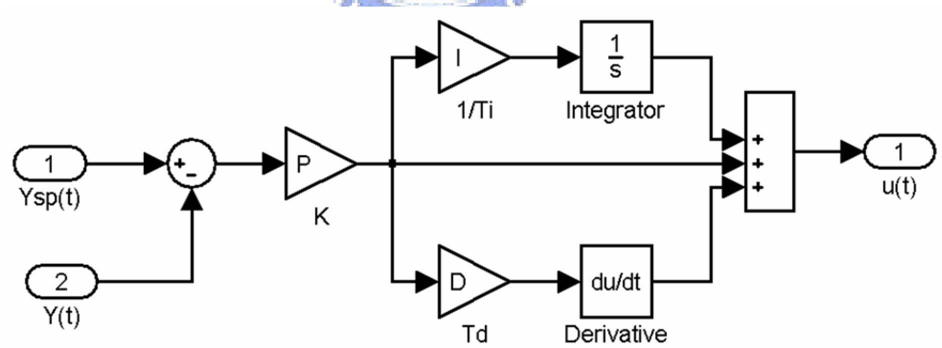

Figure 4.2-2 Standard Form of PID Control

Since the differential term in PID control has the ability of "predict", a better performance by a modified PID algorithm is obtained if the "predict" term interacts other terms. Block diagram of such algorithm is shown in Figure 4.2-3, and the modified transfer function is shown in Eq.(4.2-2). Such algorithm is also more commercial in practice.

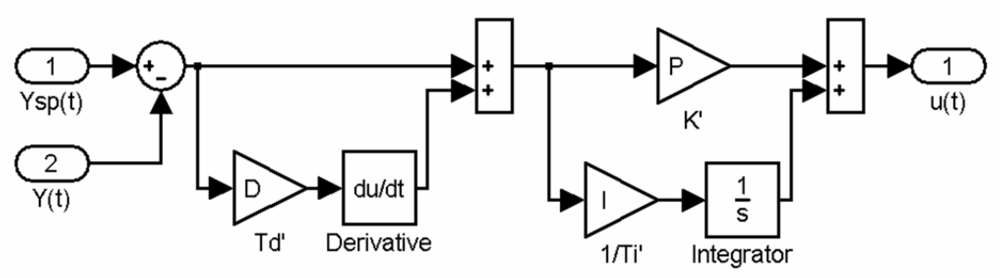

Figure 4.2-3 Interacting Form of PID Control

$$G'(s) = K'(1 + \frac{1}{aT_i'})(1 + sT_d') \qquad (4.2\text{-}2)$$

The derivative term in PID control may result in difficulties that should be prevented. If the control error $e(t)$ is sinusoidal, for example: $a \sin \omega t$, and high frequency, the differential term may leads to an arbitrarily large amplitude:

$$u_D(t) = KT_d \frac{de(t)}{dt} = aKT_d \omega \cos \omega t$$

Such high frequency gain of the derivative term is therefore limited to avoid noise amplification. This is done by implementing the derivative term as Eq.(4.2-3). And the transfer function form is shown in Eq.(4.2-4).

$$u_D(t) = \frac{T_d}{N}\frac{dD}{dt} + KT_d \frac{de(t)}{dt} \qquad (4.2\text{-}3)$$

$$D(s) = \frac{sKT_d}{1 + s\, T_d/N} e(t) \qquad (4.2\text{-}4)$$

Where N is a constant with typical value from 8 to 20.

Defining of setpoint weightings are also improvable of reducing steady-state errors and large transients. Setpoint weightings are used to provide different control errors $e(t)$ to each proportional, integral, and derivative term. For example, in general form of PID control, the control function should be modified to Eq.(4.2-5).

$$u(t) = K(e_p(t) + \frac{1}{T_i}\int_0^t e(\tau)d\tau + T_d \frac{de_d(t)}{dt}) \qquad (4.2\text{-}5)$$

where:

$$e_p(t) = by_{sp}(t) - y(t)$$
$$e_d(t) = cy_{sp}(t) - y(t)$$

b and c are constants of setpoint weighting. But note that these two constants are typically 0 or 1 in commercial controllers.

From the general form, and modify with interacting modification, derivative limitation, and setpoint weighting, an improved PID control algorithm, which is used in the clutch actuator control function, is expressed in Eq.(4.2-6).

$$U(s) = K((b + \frac{1}{sT_i'})\frac{1+scT_d'}{1+\frac{sT_d'}{N}}Y_{sp}(t) - (1+\frac{1}{sT_i'})\frac{1+sT_d'}{1+\frac{sT_d'}{N}}Y(t)) \qquad (4.2\text{-}6)$$

Besides, both control plants, as shown in Figure 4.2-1, have saturate outputs. In the control function of the clutch actuator, the first control plant, PID1 in Figure 4.2-1, is limited at 25/-25, which is the maximum current restricted for the motor, and the second control plant, PID2 in Figure 4.2-1, is limited at 12/-12, which is the maximum voltage provideable to the motor.

Integrator windup is also established in the control function, which gives upper and lower limits for integrators. The limits are the same with the saturate outputs of each control plants.

The turning of these control parameters should depend on the system dynamic characters. Methods used in this study are introduced in the next section. According to these methods, initial estimates of the control parameters are obtained and are used as initial values of the optimization implemented in CHAPTER 5.

4.3 Turning of PID Parameters

In this section, parameters of the control functions, the two control plants with PID algorithm as shown in Eq.(4.2-6), are determined. Such parameters are used to be the initial values of the optimization implemented in CHAPTER 5.

To design the parameters of the PID control function, it is necessary to understand what the primary goal the control is. The two common types are setpoint tracking and disturbances adjusting. In the control of the clutch actuator, in general operation, most disturbances are not abruptly, which always affect gradually with gradual variations, for example: temperature increase. Setpoint of the clutch control always works with sudden changes, which requires clutch to be suddenly disengaged and engaged. And the most difficulty for the control of clutch actuator is to cope with the resistant force from clutch, which is in a high order form. Obviously, the clutch control should focus on setpoint tracking more than disturbances adjusting. Such tendency is taken as a base to choice turning methods for the PID control functions.

In the control function, Zinger-Nichols (Z-N) method is used to design parameters within the first control plant, and Internal Model Control (IMC) method is used to design parameters within the second control plant. There are many turning methods for PID control, for example: Z-N, IMC, ISE, ITAE, Cohen-Coon…etc. Both Z-N method and IMC method are good for setpoint tracking [30].

IMC method is very good for setpoint tracking, and disturbances can be completely eliminated theoretically. Steady state error can be almost zero if the controlled plant is realized thoroughly. But the premise is the system can not be too complicate, otherwise the controller will be in very high order terms. The control loop of PID2 from motor voltage to motor current is a first order system, as shown in Eq.(2.2-5). IMC method suits for such system very well.

Z-N method is widely used and is easy to find appropriate parameters from a complex model. But it is sometimes insufficient for perfect solution. However, for initial values of optimization, it is enough and efficient. Therefore it is chosen to be the turning method of first control plant that controls the clutch position.

Besides the turning methods design the primary parameters K, T_i, and T_d, other parameters, derivative limitation N and setpoint weighting b and c, should also be defined. The typical values of N are 8 to 20, an average value of 14 is chosen in initial. And both b and c are 0 or 1 in commercial controller, both b and c are set to 1 in initial.

The two turning methods used to design parameters of the two control plants are introduced and implemented in the following subsections.

4.3.1 Internal Model Control

The internal model principle is a general method for design of control systems that can be applied to PID control. A block diagram of such a system is shown in Figure 4.3-1.

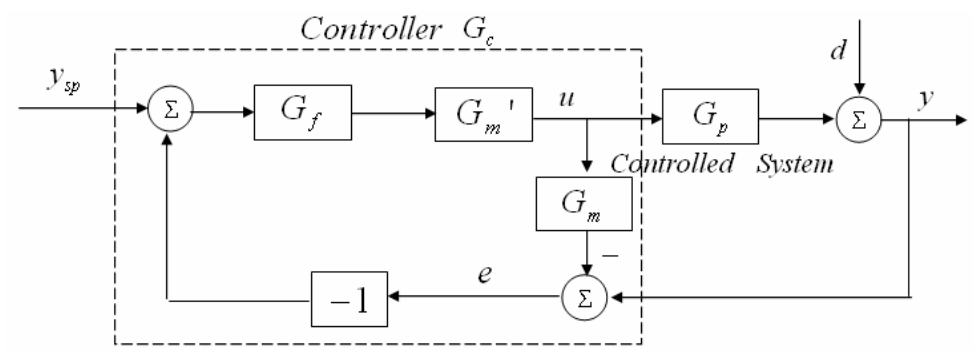

Figure 4.3-1 Block Diagram of Internal Model Control

In the diagram, it is assumed that all disturbances acting on the controlled system are reduced to an equivalent disturbance d at the controlled system output. In the figure, G_m denotes a model of the controlled system, G_m' is an approximate inverse of G_m, and G_f is a low-pass filter.

If the model match the controlled system, i.e., $G_m = G_p$, the signal e is equal to the disturbance d for all control signals u. If $G_f = 1$ and G_m' is an exact inverse of the process, then the disturbance d will be canceled perfectly. The filter G_f is introduced to obtain a system that is less sensitive to modeling error. A common choice is $G_f = 1/(1+sT_f)$, where T_f is a design variable.

The controller obtained by the internal model principle can be represented as an ordinary series controller with the transfer function shown in Eq.(4.3-1)

$$G_c = \frac{G_f G_m'}{1 - G_f G_m' G_m} \qquad (4.3\text{-}1)$$

From this expression, it follows that controller of this type cancels poles and zeros of the controlled system.

By making special assumptions, it is possible to obtain PI or PID controller from the principle. The approximated equations are [29]:

PI:

$$G_c(s) = \frac{1+sT}{K_p s(L+Tf)}$$

PID:

$$G_c(s) = \frac{(1+sT)(1+sL/2)}{K_p s(L+T_f)}$$

where the special assumptions are [29]:

$$G_p(s) = \frac{K_p}{1+sT} e^{-sL}$$
$$G_m'(s) = \frac{1+sT}{K_p}$$
$$G_f(s) = \frac{1}{1+sT_f}$$

Such Internal Model Control (IMC) method is used in the second control plant as shown in Figure 4.2-1, because the controlled system G_p is in a simple and reliable term as stated in Eq.(2.2-5).

From Eq.(2.2-5), the transfer function of G_p can be expressed as Eq.(4.3-2), where the term $K_e\dot{\theta}_m$ in Eq.(2.2-5) is seen as an disturbance.

$$G_p = \frac{1}{L_a s + R_a} \qquad (4.3\text{-}2)$$

Since the system equation of the electrical model, Eq.(2.2-5), is reliable based on many experiences, and the parameters according to experiences are verified as in Table 2.4-11 and Figure 2.4-15, it is reasonable to choice T_f to be 0 for the confidence from experiments and theories. However, even experiences can have errors and the dynamic characters may change with working condition. Still a small constant 1e-3 is assigned to T_f.

From Eq.(4.3-2), transfer functions of G_m and G_m' are shown in Eq.(4.3-3).

$$G_m = \frac{1}{L_a s + R_a}$$
$$G_m' = L_a s + R_a \qquad (4.3\text{-}3)$$

Where $L_a = 0.001355$ and $R_a = 0.2638$ according to experiences as discussed in 2.4.3 .

Block diagram of the control loop is shown in Figure 4.3-2.

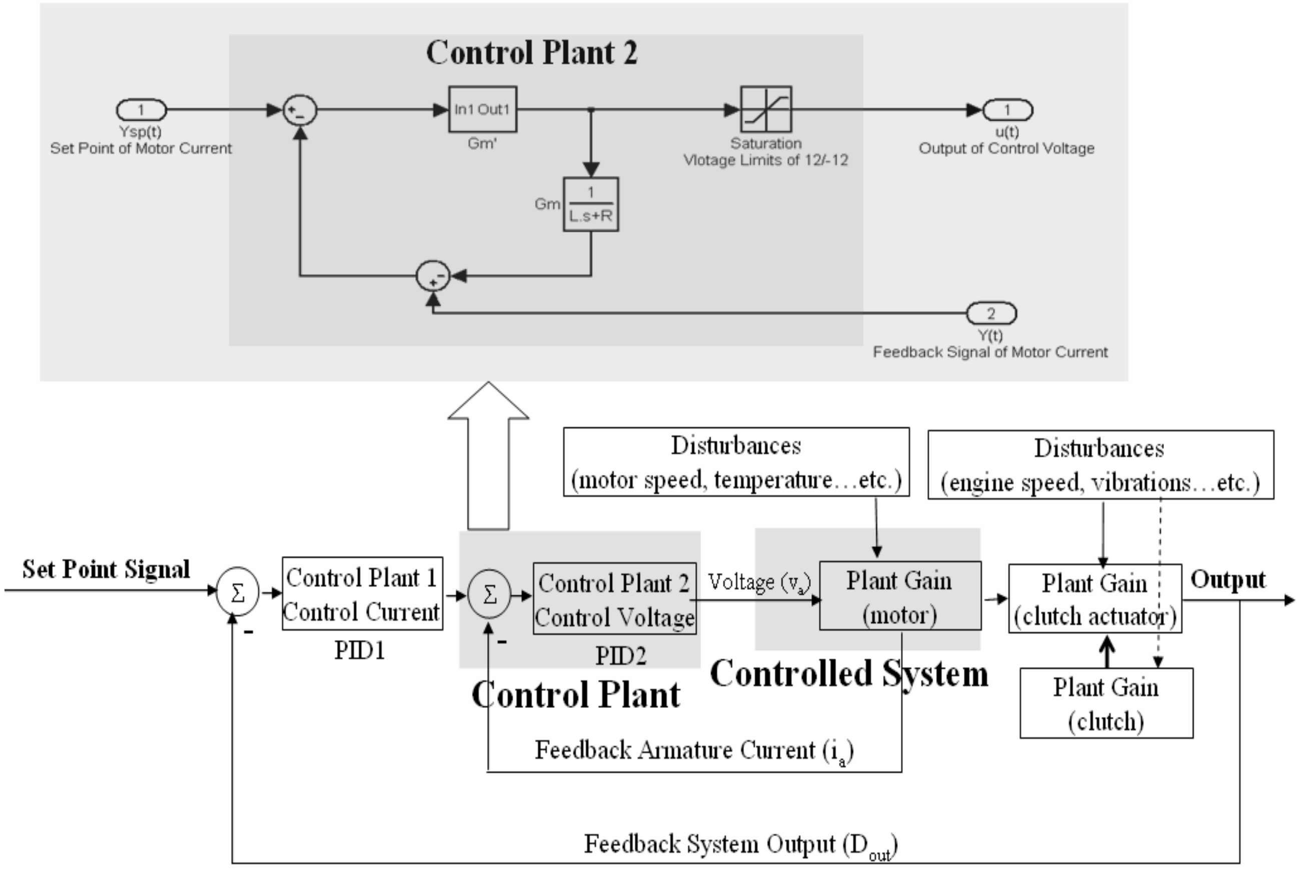

Figure 4.3-2 Block Diagram of Second Control Plant

Figure 4.3-3 shows simulation result of the control plant, where a sinusoid setpoint and an impulse setpoint are implemented. It is obvious that errors are small in both transient state and steady state. It also shows a good disturbance adjusting ability. In the scale views, where clutch actuator traveled to the upper limit and suffered a collision, the controller can eliminate the interference quickly and steadily.

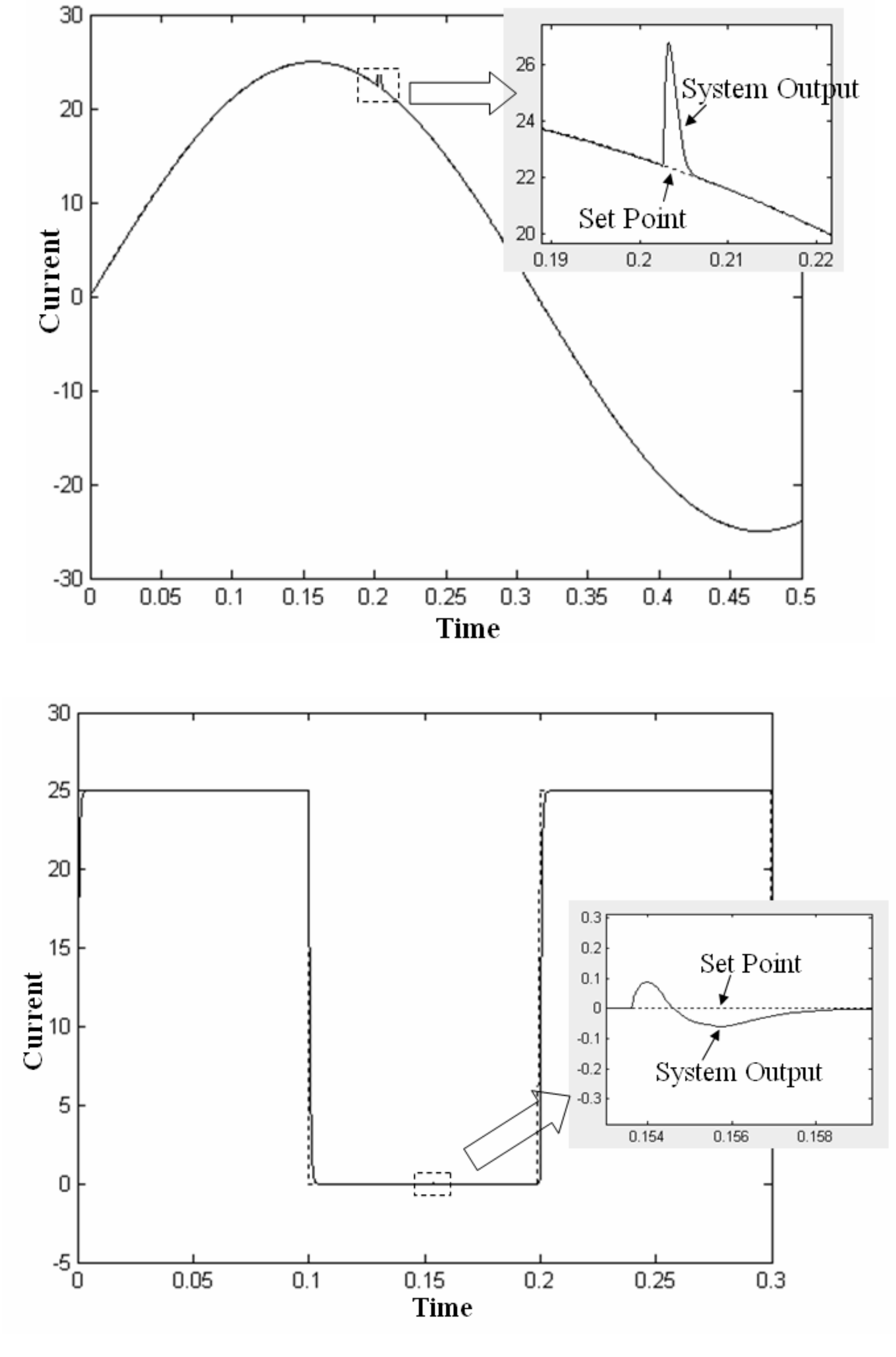

Figure 4.3-3 Simulation Results of Second Control Plant

Since the performance of the control plant designed by internal model method is good, such strategy and parameters are adopted directly into the final design, and is not required to submit into optimization implementation in CHAPTER 5.

4.3.2 Ziegler-Nichols

Two classical methods for determining the parameters of PID controller were presented by J.G. Ziegler and N.B. Nichols in 1942. These methods are still widely used today. They often form the basis for turning procedures used by controller manufacturers and process industry. The methods are based on determination of some features of the controlled dynamics.

The first method is step response method. Z-N recognized that the step response of a large number of process control system exhibits a process reaction curve like that shown in Figure 4.3-4.

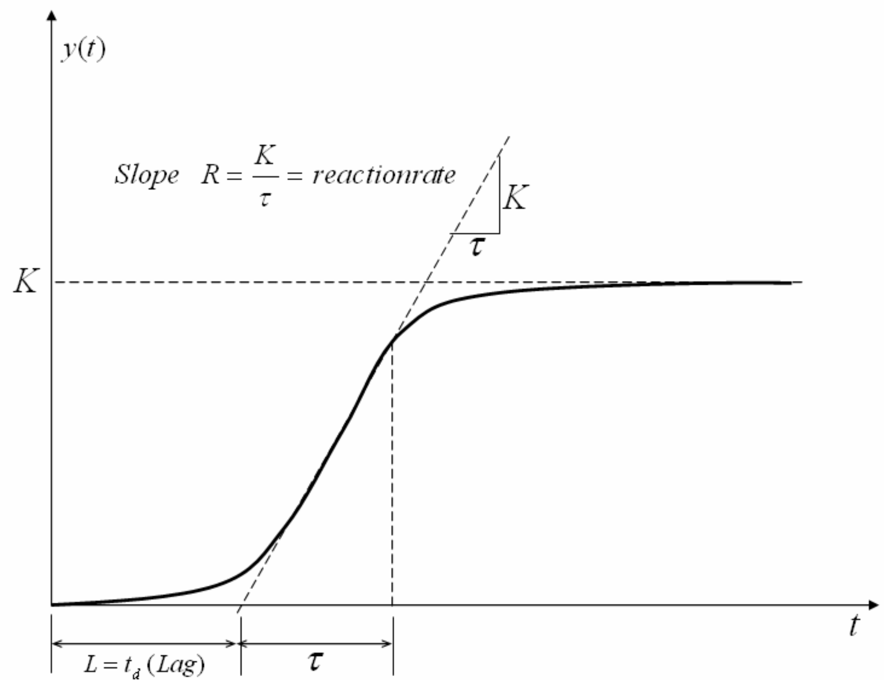

Figure 4.3-4 System Reaction Curve

This curve can be generated from either experimental data or dynamic simulation of the plant. The S-shape of the curve is characteristic of many high-order system, and such plant transfer function may be approximated by Eq.(4.3-4).

$$\frac{Y(s)}{U(s)} = \frac{Ke^{-t_d s}}{\tau s+1} \tag{4.3-4}$$

which is simply a first-order system plus a time delay of t_d seconds. The constants in Eq.(4.3-4) can be determined from the unit step response of the controlled system. If a tangent is drawn at the inflection point of the reaction curve, then the slope of the line is $R = K/\tau$ and the intersection of the tangent line with the time axis identifies the time delay $L = t_d$.

Such method is based on a decay ratio of approximately 0.25. This means that a dominant transient decays to a quarter of its value after one period of oscillation. A quarter decay corresponds to ξ=0.21 and is a good compromise between quick response and adequate stability margins. The regulator parameters suggested by Z-N for the common controller terms are shown in Table 4.3-1 [10].

PID Controller	
K	$1.2/RL$
T_i	$2L$
T_d	$0.5L$

Table 4.3-1 Step Response Method [10]

The second method is frequency response method. This method is based on a simple characterization of the controlled dynamics. The design is based on knowledge of the point on the Nyquist curve of the process transfer function where the Nyquist curve intersects the negative real axis. This point is characterized by the parameters K_u and T_u, which are called the ultimate gain and the ultimate period. These parameters can be determined in the following way. Connect a controller to the controlled system, set the parameter so that action is proportional, i.e., $T_i = \infty$ and $T_d = 0$. Increase the gain slowly until the process starts to continuously oscillate. The gain when this occurs is K_u and the period of the oscillation is T_u. The formulas for the parameter of the controller in terms of the ultimate gain and the ultimate period of the controller are shown in Table 4.3-2 [29].

PID Controller	
K	$0.6K_u$
T_i	$0.5T_u$
T_d	$0.125T_u$

Table 4.3-2 Frequency Response Method [29]

Z-N method is used to design parameters of Eq.(4.2-6), the first control plant of the control function where the setpoint input $Y_{sp}(t)$ is signal of the desired travel of the clutch, the feedback signal $Y(t)$ is clutch position measured by position sensor, and the output $U(s)$ is the setpoint of DC motor current. Block diagram of the control loop according to this control plant is shown in Figure 4.3-5.

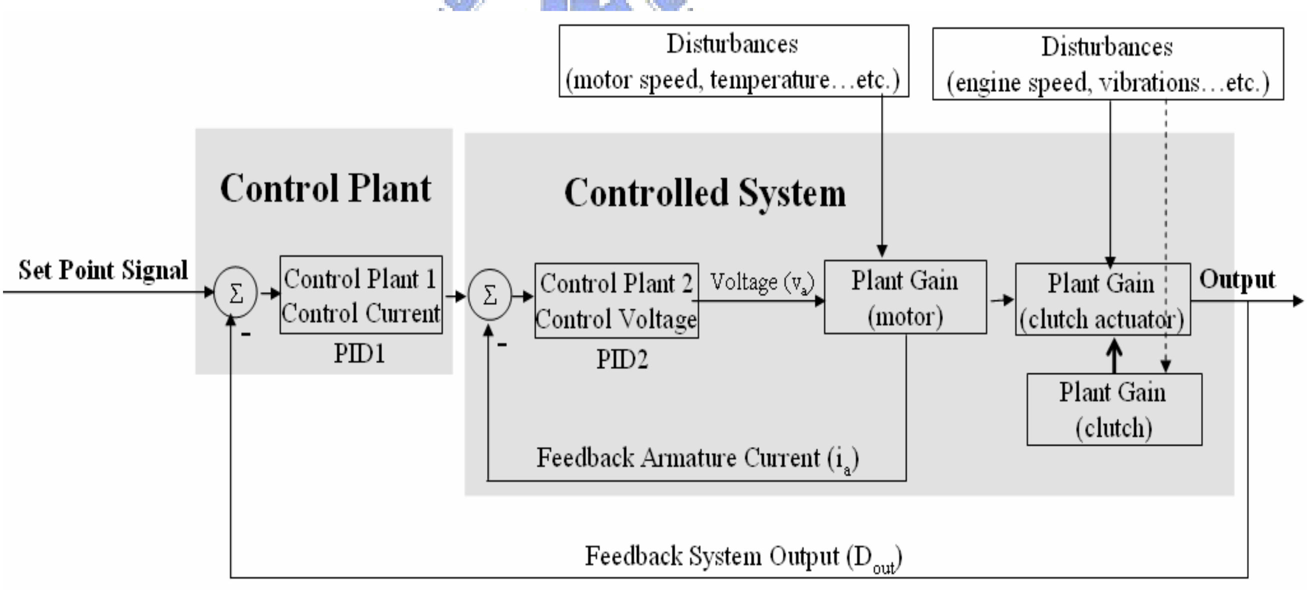

Figure 4.3-5 Block Diagram of First Control Plant

For step response method, a unit step input to the controlled system is not workable.

Physically, the unit input of one ampere is not enough to lend an impetus to the actuator because of the friction restriction. The same problem is emerged in the simulated model. To be convenient, frequency method is selected to be used.

By the criteria of frequency response method, proportional gain K is adjusted with other parameters set to $T_i = \infty$ and $T_d = 0$. The adjustment results are shown in Figure 4.3-6, which shows that the controlled system continuously oscillates with a period of 0.0078 second with $K = 7.9$.

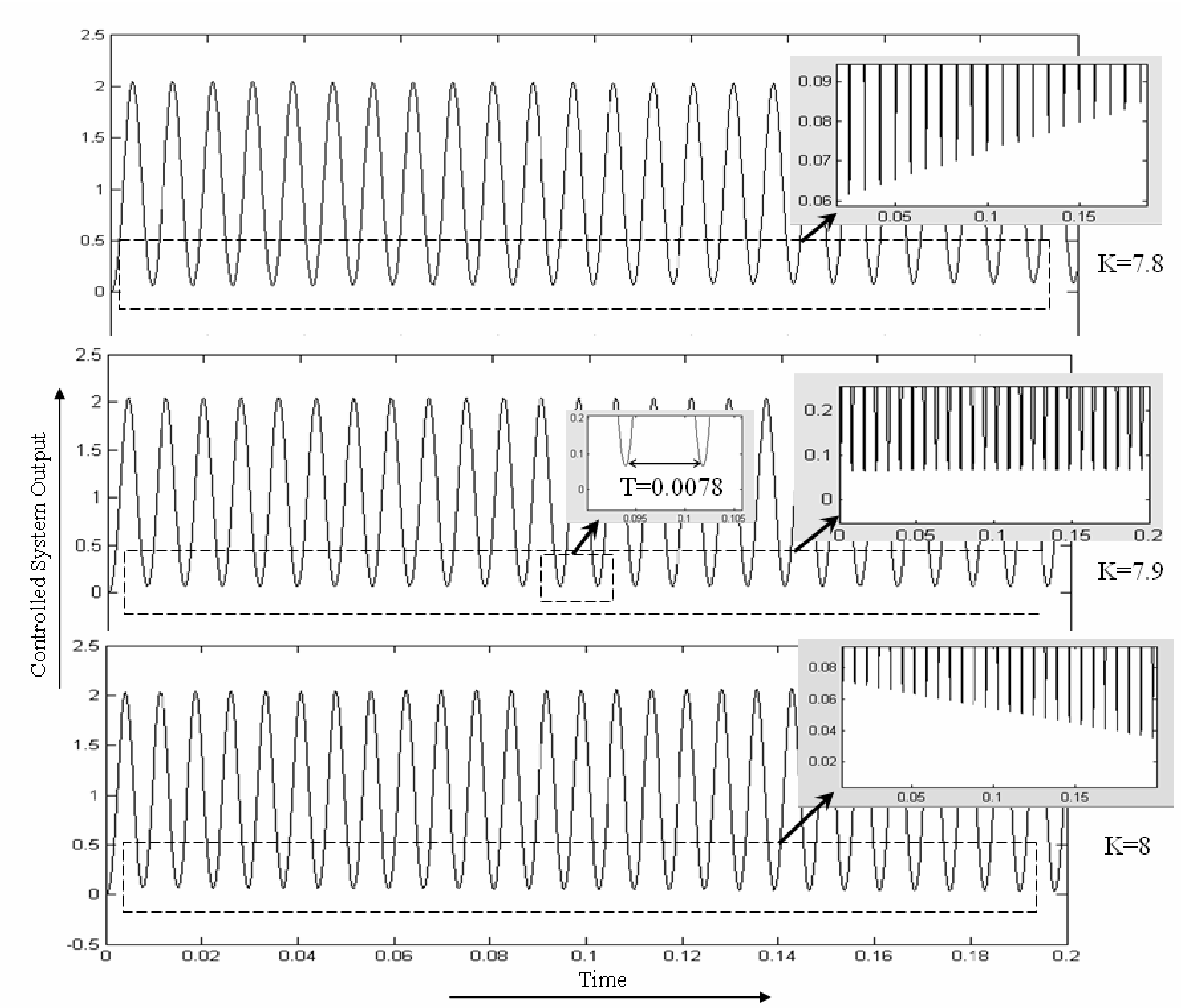

Figure 4.3-6 Frequency Response

From Figure 4.3-6, $K_u = 7.9$ and $T_u = 0.0078$ is obtained. From Table 4.3-2, parameters within the first control plant based on Eq.(4.2-6) are shown in Table 4.3-3.

First Control Plant	
K	*4.74*
T_i	*0.0039*
T_d	*0.000975*
N	*14*
b	*1*
c	*1*

Table 4.3-3 Parameters of First Control Plant

The control function with parameters designed before is simulated as shown in Figure 4.3-7. It is obvious that overshoot is conspicuous. Such instability may be caused from a too large proportional gain K or a too small derivative gain T_d. Such defects will be amended in the next chapter using optimization method.

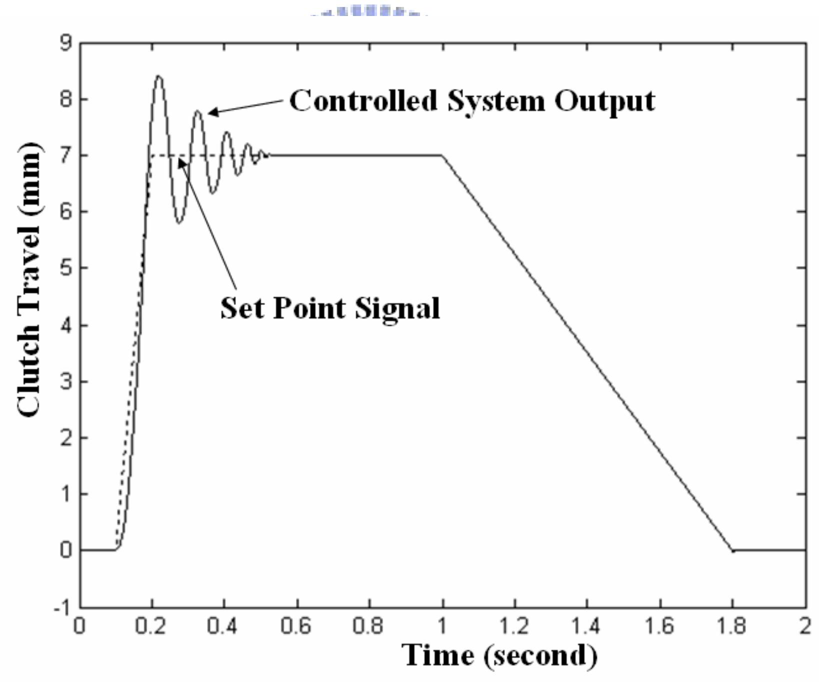

Figure 4.3-7 Simulation Result with Control Function