Chapter 2
Design of First Order Controllers for Unstable Infinite Dimensional Plants

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Abstract A design method for first order controllers is presented for a class of unstable infinite dimensional plants, including systems with time delays, fractional order systems, and systems represented by PDEs. The design restricts the controllers to be in the form of PI, PD and lead or lag controllers. The approach is based on the small gain theorem and requires minimization of an H_∞ norm of a transfer function over a low number of parameters. The gain margin optimization problem is solved for PD controllers. For PI controllers, optimization of the integral action gain is also discussed.

2.1 Introduction

This work deals with the design of different types of first order controllers for infinite dimensional plants whose transfer functions contain single unstable pole. In this context PI, PD, lead and lag controllers are investigated. The basic idea is to put the characteristic equation of the feedback system into a form where the small gain theorem can be applied. For this purpose, algebraic manipulations similar to those used in [6, 17] play a crucial role. Once the controller structure is fixed, the range of allowable controller gain is estimated by computing the H_∞ norm of an infinite dimensional transfer function which contains a free parameter. Optimization of this free parameter is helpful for reducing the conservative results obtained in [17].

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It should be noted that when there are only small number of free parameters in the controller, classical stability checks (e.g. Nyquist criterion) can be used to determine the set of all stabilizing controller parameters. However, this brute force method may not be numerically very attractive, especially when the plant considered is unstable and infinite dimensional.

In particular, for time delay systems there are several numerically feasible methods for finding low order controllers, see e.g. [4, 5, 14, 19]. For applications to communication networks see [11] and [20]. The method of [17] has been extended to cover fractional order systems with time delays in [16], see also [3].

This chapter is organized as follows. Section 2.2 contains several examples of engineering applications where plant model falls within the framework of the present study. A sufficient condition for the stability of the feedback system (based on the small gain theorem) is derived in Sect. 2.3. Then in Sect. 2.4 different types of controllers are designed based on this condition. Conclusions and future works are given in Sect. 2.5.

2.2 Problem Definition and Examples of Plants Considered

As mentioned above, the plants considered here have transfer functions in the form

$$P(s) = \frac{1}{s-p}\, G(s) \tag{2.1}$$

where $p \geq 0$ is the unstable pole and $G \in H_\infty$ is the stable part of the plant. Note that $G(s)$ can be irrational (plant is infinite dimensional). The factorization in the form (2.1) also implies that the plant is strictly proper.

The controllers to be designed have the following common structure

$$C(s) = K_p + \frac{K_d\, s}{\tau s + 1} + \frac{K_i}{s} \quad , \quad K_p, K_d, K_i \in \mathbb{R}, \quad \tau \geq 0. \tag{2.2}$$

Note that PD, PI, lead and lag controllers are special case of (2.2):

$$C_{pd}(s) = K_p\,(1 + \widetilde{K}_d\, s), \qquad \widetilde{K}_d = \frac{K_d}{K_p}, \tag{2.3}$$

$$C_{pi}(s) = K_p\left(1 + \frac{\widetilde{K}_i}{s}\right), \qquad \widetilde{K}_i = \frac{K_i}{K_p}, \tag{2.4}$$

$$C_\ell(s) = K_p\left(\frac{1 + \alpha\,\tau\, s}{1 + \tau\, s}\right), \qquad \alpha\tau = \tau + \widetilde{K}_d. \tag{2.5}$$

Clearly C_ℓ is a lead controller if $\widetilde{K}_d > 0$ and it is a lag controller if $0 > \widetilde{K}_d > -\tau$.

Definition 1 The feedback system formed by the controller C and the plant P is stable if $S := (1 + PC)^{-1}$, CS and PS are stable, i.e., they are transfer functions in H_∞. If this is the case, then the controller C is said to stabilize the plant P. The set of all controllers stabilizing a given plant P is denoted by $\mathscr{C}(P)$.

The goal of this chapter is to determine controllers $C(s)$, in the form (2.3)–(2.5), stabilizing a given unstable infinite dimensional plant $P(s)$ whose transfer function is given by (2.1). There are several applications where plant transfer functions have this structure; specific examples are given below.

Example 1 Integrating systems with transport delay:

$$P(s) = \frac{K\ e^{-hs}}{s}\ , \quad K > 0, \quad h > 0\ , \tag{2.6}$$

i.e., in this case, $p = 0$ and $G(s) = Ke^{-hs}$. There are many application examples and control methods for this plant, [12, 21]. Application examples include oil/gas pipelines, communication networks, manufacturing plants, storage systems, etc., see e.g., [13, 18].

Example 2 Abstract model of an aircraft:

$$P(s) = \frac{e^{-hs}}{s - p}\ , \quad h > 0, \quad p > 0, \quad G(s) = e^{-hs}. \tag{2.7}$$

This model is used for the purpose of controlling the high frequency longitudinal dynamics (short period) of an aircraft. Dynamics due to elasticity, sensor, actuator, sampling, contribute to the time delay. The product $h \cdot p$ represents how difficult it is to control this open loop unstable system. Depending on the operating regime, it is observed that $0.06 < h \cdot p < 0.37$ for an X-29 aircraft [2].

Example 3 Flexible beam with non-collocated actuator and sensor: Typically, mathematical models of flexible beams are given by partial differential equations, [1], and their transfer functions are irrational. For the free-free beam model (with normalized material parameters) shown in Fig. 2.1, the following infinite product expansion of $G(s)$ converges in H_∞ (see [9, 10]):

$$P(s) = \frac{1}{s}\ G(s)\ , \qquad G(s) = \frac{2e^{-hs}}{(\tau_v s + 1)} \prod_{n=1}^{\infty} \left(\frac{1 + \varepsilon s - s^2/\omega_n^2}{1 + \varepsilon s + s^2/\widetilde{\omega}_n^2} \right)\ , \tag{2.8}$$

where $\tau_v > 0$ is the sensor parameter, $h > 0$ is the input delay, $\varepsilon > 0$ is the damping parameter of the beam and $\omega_n,\ \widetilde{\omega}_n > 0$ with $\omega_n \to 2\left(\frac{\pi}{4} + n\pi\right)^2$ and $\widetilde{\omega}_n \to \left(\frac{\pi}{2} + n\pi\right)^2$ as $n \to \infty$.

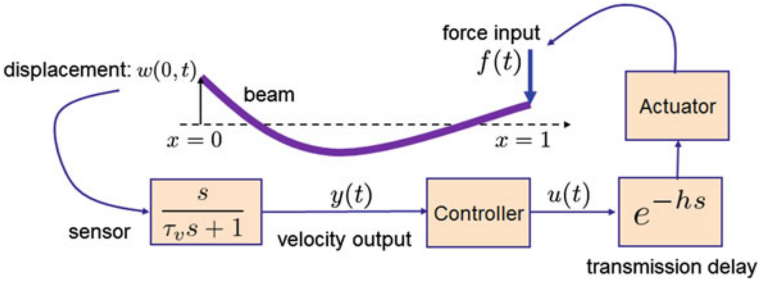

Fig. 2.1 Flexible beam control loop under delayed point force input and velocity feedback

Example 4 Interconnected systems with time delays:

$$P(s) = \frac{e^{-hs}}{s+2}\left(\frac{s+1+2(s-1)e^{-2s}}{s+1-2e^{-0.4s}}\right) = \frac{1}{s-p}G(s), \qquad (2.9)$$

where $h > 0$ and $p \approx 0.5838$ is the unique root of $(s+1-2e^{-0.4s}) = 0$ in $\overline{\mathbb{C}}_+$. So,

$$G(s) = e^{-hs}\left(\frac{(s+1)+2(s-1)e^{-2s}}{s+2}\right)\left(\frac{s-p}{s+1-2e^{-0.4s}}\right).$$

Example 5 A non-laminated magnetic suspension system: The following fractional order plant model is taken from [8]:

$$P(s) = \left((s^\alpha)^5 + (s^\alpha)^4 - c\right)^{-1}, \quad \alpha = 0.5, \quad c > 0. \qquad (2.10)$$

It has been shown that P can be factored as in the standard form (2.1), see [7]:

$$P(s) = \frac{1}{s-p}G(s) \quad \text{with} \quad p = r^2, \quad G(s) = \frac{(s^\alpha + r)(s^\alpha - r)}{(s^\alpha)^5 + (s^\alpha)^4 - c}$$

where $r > 0$, is the unique root of $(z^5 + z^4 - c) = 0$ on $\mathbb{R}_+$.

2.3 A Sufficient Condition for Feedback System Stability

In this section the controller is taken to be in the form C_{pd} or C_ℓ. Such a controller is stabilizing a plant in the form (2.1) if and only if there exists a constant $a > 0$ such that U_a is unimodular (i.e. $U_a, U_a^{-1} \in H_\infty$):

$$U_a(s) := \frac{s-p}{s+a} + \frac{K_p}{s+a} G(s)C_0(s),$$

where $C_0(s) := \frac{1+\alpha\tau s}{1+\tau s}$ when $C = C_{pd}$ or $C_0(s) := (1 + \widetilde{K}_d\ s)$ when $C = C_\ell$. Define

$$K_p := (p+a)G(0)^{-1} \quad \text{and} \quad G_0(s) := G(s)G(0)^{-1} \tag{2.11}$$

then

$$U_a(s) = 1\ +\ (p+a)\ \frac{s}{s+a}\ \Psi_0(s) \quad \text{where} \quad \Psi_0(s) = \frac{1}{s}\ (\ G_0(s)C_0(s) - 1\ )\,.$$

Thus, using the fact that $\left\| \frac{s}{s+a} \right\|_\infty \leq 1$, U_a is unimodular if

$$(p+a) < \|\Psi_0\|_\infty^{-1}. \tag{2.12}$$

The condition (2.12) was derived earlier in [6, 17]. Clearly, a less conservative condition for U_a to be unimodular is

$$(p+a) < \|\Psi_a\|_\infty^{-1}, \tag{2.13}$$

where

$$\Psi_a(s) := \frac{1}{s+a}\ (\ G_0(s)C_0(s) - 1\ )\,.$$

Note that

$$\|\Psi_a\|_\infty \leq \|\Psi_0\|_\infty \quad \forall\, a > 0.$$

Therefore, the controller defined as above is a stabilizing controller for the plant if

$$pG(0)^{-1} < K_p < (p+a_o)G(0)^{-1},$$

where $a_o > 0$ is the largest $a > 0$ satisfying (2.13).

In order to illustrate the computations involved in the above discussion, let us consider the plant defined by (2.6) with $K = 1$ and $h > 0$. Let $C_0(s) = 1$ (i.e., consider proportional control only). The exact value of the upper bound of the controller gain can be easily computed as

$$K_p = a < \frac{\pi}{2h} \approx 1.57h^{-1}.$$

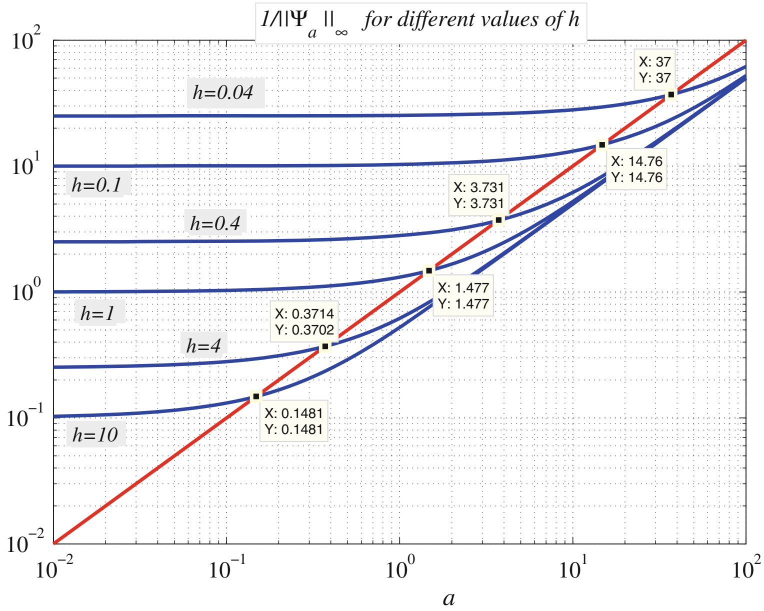

Fig. 2.2 The graph of $1/\|\Psi_a\|_\infty$ for different values of h: the largest a satisfying (2.13) is $1.48/h$

If one uses the condition (2.12), the conservative upper bound of the controller gain is

$$K_p = a < \frac{1}{\|\Psi_0\|_\infty} = h^{-1}.$$

On the other hand, if (2.13) is used, then

$$K_p = a < \tau_o h^{-1} \qquad \text{with} \qquad \tau_o \approx 1.48,$$

where τ_o is computed as shown in Fig. 2.2.

Clearly, when $C_0(s) \neq 1$, for example, $C_0(s) = \frac{1+\alpha\tau s}{1+\tau s}$ or $C_0(s) = (1+\widetilde{K}_d\, s)$, the free parameters (α, τ) or $\widetilde{K}_d$ can be used to further maximize a_o, the largest $a > 0$ satisfying (2.13).

Remark 1 There are some plants which do not admit a feasible stabilizing controller in the form C_{pd} or C_ℓ. For example, if the plant does not satisfy the parity interlacing property (PIP), then there does not exist a *stable* stabilizing controller. In order to illustrate this point, consider the plant

$$P(s) = \frac{1}{s-p}\left(\frac{1-s/z}{1+\tau s}\right) \quad p > 0\,,\, z > 0,\, \tau > 0,$$

for which there exist a stable stabilizing controller if and only if $p < z$. If the proportional controller is defined as $C(s) = K_p = (p + a)$ then, using the notation set above,

$$\Psi_a(s) = \frac{-s\ (\tau + z^{-1})}{(s + a)(1 + \tau s)} \quad \Rightarrow 1/\|\Psi_a\|_\infty = z\left(\frac{1 + \tau a}{1 + \tau z}\right).$$

So, the condition (2.13) becomes

$$p + a < z\left(\frac{1 + \tau a}{1 + \tau z}\right) = \frac{z}{1 + \tau z} + \frac{\tau z}{1 + \tau z}a,$$

which is stronger than the PIP, i.e., $p < z$.

2.4 PD and PI Controller Designs

2.4.1 PD Controller Design

Recall that for the plant (2.1), a PD controller is in the form $C_{pd}(s) = K_p C_0(s)$ where $K_p = (p + a)G(0)^{-1}$ and $C_0(s) = (1 + \widetilde{K}_d\ s)$. Based on the results of Sect. 2.3 the largest $a > 0$ satisfying (2.13) should be computed depending on $\widetilde{K}_d \in \mathbb{R}$. For this purpose define $G_0(s) = G(s)G(0)^{-1}$, $Q := \widetilde{K}_d \in \mathbb{R}$ and

$$\gamma(Q, a) := \left\| \frac{G_0(s) - 1}{s + a} + \ Q\ \frac{s}{s + a} G_0(s) \right\|_\infty . \tag{2.14}$$

In order to maximize the gain margin (GM) of the system one should try to minimize $\gamma(Q, 0)$ (the conservative approach) or try to find the largest a satisfying (2.13). See [17] for a detailed discussion on the computation of the optimal Q minimizing $\gamma(Q, 0)$ for the conservative approach. The main idea can be extended to the case $a > 0$ easily; see the algorithm given below.

Initialize: Determine a range of $Q \in [Q_{\min}\ , Q_{\max}] \subset \mathbb{R}$

Step 1. For each fixed Q in this interval
if it exists find the largest $a_{\max}(Q)$ such that

$$(p + a) < 1/\gamma(Q, a) \quad \forall\ a < a_{\max}(Q).$$

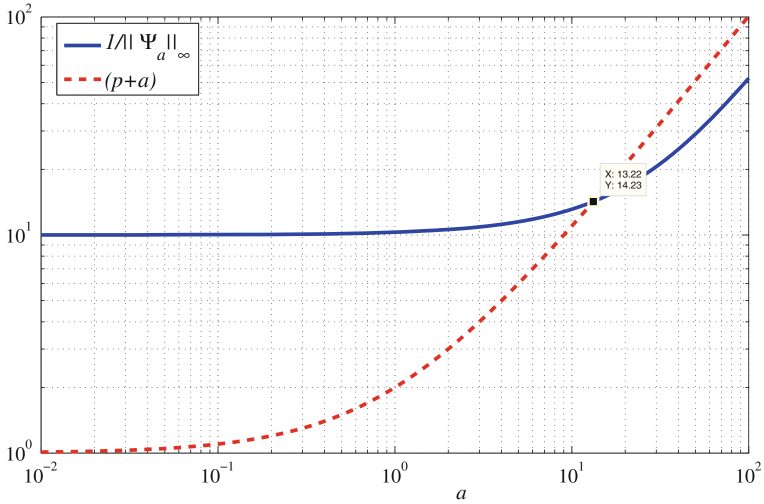

Fig. 2.3 $(p + a)$ and $1/\|\Psi_a\|_\infty$ versus a for the plant (2.7) with $h = 0.1$ and $p = 1$

Step 2. Plot Q versus $a_{\max}(Q)$ find the maximum of $a_{\max}(Q)$ and define

$$Q_{opt} := \arg\ \max\{a_{\max}(Q)\}.$$

End: An allowable range of the controller gain K_p is

$$pG(0)^{-1} < K_p < (p + a_o)G(0)^{-1} \quad \text{with} \quad a_o := a_{\max}(Q_{opt}).$$

For $p > 0$, gain margin optimizing (see [15]) PD controller parameters are

$$\widetilde{K}_{d,opt} = Q_{opt}\ , \quad K_{p,\text{GMopt}} = \sqrt{p\ (p + a_o)}G(0)^{-1}.$$

Alternatively, one can choose the *least fragile* proportional gain

$$K_{p,\text{LF}} = \left(p + \frac{a_o}{2}\right) G(0)^{-1}.$$

Step 1 of the algorithm involves drawing a graph like the one shown in Fig. 2.2. To illustrate the numerical computations, consider the plant (2.7) with $h = 0.1$ and $p = 1$. If proportional controller is used, then $Q = 0$ and $a_{\max}(0) = 13.2$ as seen in Fig. 2.3; that means the allowable range of the gain is $1 < K_p < 14.2$.

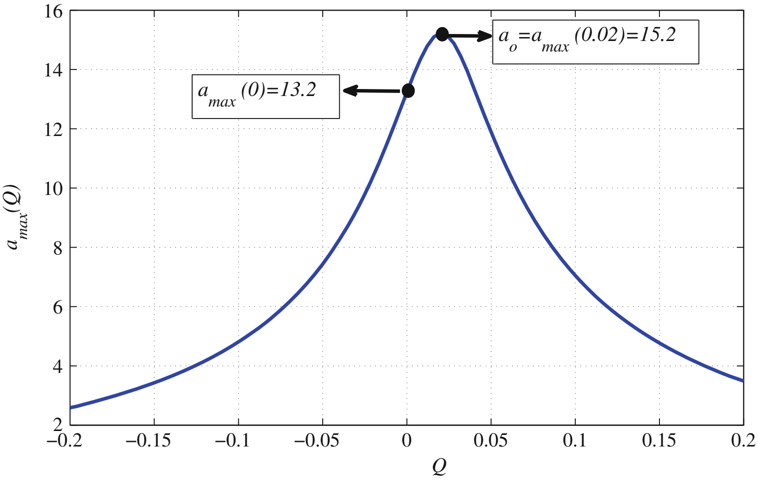

Fig. 2.4 $a_{\max}$ versus Q for the plant (2.7) with $h = 0.1$ and $p = 1$

On the other hand, it is possible to enlarge this interval by adding a derivative action. Figure 2.4 shows how $a_{\max}$ change as a function of Q. Clearly, the optimal choice is $\widetilde{K}_{d,opt} = Q_{opt} = 0.02$ and that leads to $a_o = \max a_{\max}(Q) = 15.2$ which means that the allowable gain is in the interval $1 < K_p < 16.2$ and

$$C_{pd,\text{GMopt}}(s) = 4.025\ (1 + 0.02\ s) \quad \text{and} \quad C_{pd,\text{LF}}(s) = 8.6\ (1 + 0.02\ s).$$

Remark 2 On Lead-Lag Controller Design.
Recall that for the lead or lag controller design $C_0(s)$ is in the form

$$C_0(s) = 1 + \frac{Q_1\ s}{1 + Q_2 s} \ \text{ with } \ Q_1 := \widetilde{K}_d > -\tau, \quad Q_2 := \tau > 0.$$

Then, similar to the PD controller design, the parameter a which determines the controller gain should be such that $(p + a) < 1/\gamma_a$, where

$$\gamma_a(Q_1, Q_2) = \left\| \frac{G_0(s) - 1}{s + a} + \left(\frac{Q_1}{1 + Q_2\ s} \right) \frac{s}{s + a} G_0(s) \right\|_\infty .$$

So, to find $a_{\max}(Q_1, Q_2)$, in Step 1 of the corresponding gain margin optimization algorithm, the computations are done for two parameters in nested loops. Then in Step 2, a surface plot of $a_{\max}(Q_1, Q_2)$ is obtained and its maximum is determined.

2.4.2 PI Controller Design

Consider the design of a PI controller in the form

$$C_{pi}(s) = C_1(s) + \frac{K_i}{s}, \tag{2.15}$$

where $C_1(s) = K_p$ is such that $C_1 \in \mathscr{C}(P)$. In other words, a controller C_1 is already designed to stabilize P and now the integral action is to be added to the controller. The following discussion is valid for $C_1 = C_{pd}$ as well, in that case the addition of integral term will give a PID controller $C_2 = C_{pid}$.

Since $C_1 \in \mathscr{C}(P)$ the following statement holds:

$$H_1(s) := \frac{P(s)}{1 + C_1(s)P(s)} \quad \text{is in } H_\infty.$$

The characteristic equation of the feedback system formed by C_2 and P is

$$1 + C_1(s)P(s) + \frac{K_i}{s}P(s) = (1 + C_1(s)P(s))\left(1 + \frac{K_i}{s}H_1(s)\right) = 0\,.$$

Using the fact that $C_1 \in \mathscr{C}(P)$ it can be concluded that

$$C_2 \in \mathscr{C}(P) \quad \Longleftrightarrow \quad V_1^{-1} \in \mathscr{H}_\infty \;\; \text{with} \;\; V_1(s) = \left(1 + \frac{K_i}{s}H_1(s)\right).$$

Now define

$$b := K_i H_1(0)\,,$$

then V_1 can be re-written as

$$V_1(s) = \left(1 + \frac{b}{s}\right)\left(1 + \left(1 + \frac{b}{s}\right)^{-1} b \left(\frac{H_1(s)H_1(0)^{-1} - 1}{s}\right)\right). \tag{2.16}$$

Let us now assume that $b > 0$ (this is without loss of generality, since the sign of K_i can be adjusted according to the sign of $H_1(0)$). Then, note that

$$\left(1 + \frac{b}{s}\right)^{-1} = \frac{s}{s+b} \in H_\infty \quad \text{with} \quad \|\frac{s}{s+b}\|_\infty = 1.$$

The following result can be derived from the small gain theorem: $V_1^{-1} \in H_\infty$, i.e., $C_2 \in \mathscr{C}(P)$, if b satisfies

$$0 < b < 1/\|\Phi_0\|_\infty \quad \text{where} \quad \Phi_0(s) = \left(\frac{H_1(s)H_1(0)^{-1} - 1}{s}\right). \tag{2.17}$$

In fact, a careful examination of (2.16) shows that, rather than (2.17), the following less conservative sufficient condition on b can be used for C_2 to be in $\mathscr{C}(P)$:

$$0 < b < 1/\|\Phi_b\|_\infty \quad \text{where} \quad \Phi_b(s) = \left(\frac{H_1(s)H_1(0)^{-1} - 1}{s+b}\right). \tag{2.18}$$

Clearly, there is an analogy between Ψ_a and Φ_b, and the conditions (2.13) and (2.18). Note that Φ_b depends on K_p which is assumed to be in $\mathscr{C}(P)$. So, the optimal PI controller $C_{pi,opt}(s) = K_{p,opt} + (K_{i,opt}/s)$ can be designed as follows.

For each fixed $K_p \in \mathscr{C}(P)$, find the largest allowable $b > 0$ satisfying (2.18) and let it be denoted as $b_{\max}(K_p)$. Accordingly, define

$$K_{p,opt} := \arg\ \max\ \{\, b_{\max}(K_p) \ : \ K_p \in \mathscr{C}(P)\,\}.$$

Then, the least fragile integral action gain is

$$K_{i,opt} = \frac{b_{\max}(K_{p,opt})}{2} H_1(0)^{-1}.$$

In order to illustrate the computations involved in the design method described above, let us consider once more the plant (2.7) with $h = 0.1$ and $p = 1$. Recall from Fig. 2.3 that $C_1(s) = K_p$ is a stabilizing controller if $K_p \in (1\ ,\ 14.2)$. For each fixed K_p in this interval, define

$$H_1(s) = \frac{e^{-0.1s}}{s - 1 + K_p e^{-0.1s}}\ , \quad \text{clearly} \quad H_1(0) = \frac{1}{K_p - 1}.$$

Simple computations give Φ_b as:

$$\Phi_b(s) = \frac{H_1(s)H_1(0)^{-1} - 1}{s+b} = \left(\frac{1}{s+b}\right)\left(\frac{1 - s - e^{-0.1s}}{s - 1 + K_p e^{-0.1s}}\right).$$

Following the above procedure, for each $K_p \in (1, 14.2)$ and $b > 0$, the H_∞-norm $\|\Phi_b\|_\infty$ is computed. Then, from the graph of $1/\|\Phi_b\|_\infty$ versus b, the largest b, denoted by $b_{\max}(K_p)$, satisfying (2.18) is determined. Figure 2.5 shows $b_{\max}(K_p)$ versus K_p. Clearly, the largest $b_{\max}(K_p) = 5.7$ is achieved at $K_p = K_{p,opt} = 4.8$. For the least fragile integral gain, let $K_{i,opt} = H_1(0)^{-1} b_{\max}(K_{p,opt})/2 = 10.8$. The resulting controller is

$$C_{pi,\mathrm{LF}}(s) = 4.8 + \frac{10.8}{s}.$$

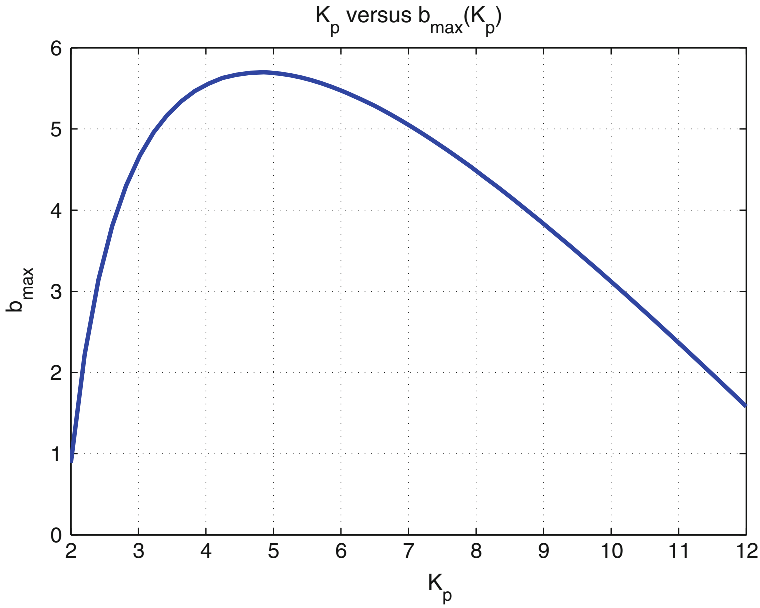

Fig. 2.5 $b_{\max}(K_p)$ versus K_p for the plant (2.7) with $h = 0.1$ and $p = 1$

2.5 Conclusions and Future Extensions

In this chapter of the book, a method is proposed for the design of stabilizing first order controllers (PD, PI and lead or lag controllers) for a class of infinite dimensional plants. The main assumption is that the plant has a single unstable pole (at the origin, or on the positive real axis). Examples from several applications are given to justify the plant model considered. These examples include systems with time delays, fractional order systems, and systems represented by PDEs.

The approach is based on the small gain theorem and requires minimization of the H_∞ norm of an infinite dimensional stable transfer function over a low number of parameters.

Another way to obtain $C_\ell(s) = K_p C_0(s)$ with a large gain margin would be to find a first order approximation of an infinite dimensional stable controller determined from the following H_∞ control problem. For a fixed $a > 0$, first, solve the one block problem

$$\gamma_o(a) = \inf_{Q \in \mathscr{H}_\infty} \| \frac{1}{s+a} \ (1 - G_0(s)Q(s)) \|_\infty .$$

If

$$(p + a) < 1/\gamma_o(a), \tag{2.19}$$

then define

$$Q_a(s) := \arg \gamma_o(a).$$

Now, all controllers in the form $K_0 G(0)^{-1} Q_a(s)$ stabilize the plant, which is given by $P(s) = \frac{1}{s-p} G(s)$, provided that the gain is in the interval

$$p < K_0 < (p + a).$$

Thus, to maximize the allowable controller gain the maximum a defined below should be determined:

$$a_{\max} = \arg \max\{ a \ : \ a \in \mathbb{R}_+ \text{ and (2.19) holds } \}.$$

The least fragile *stable* controller, in this framework, is

$$C_{s,LF}(s) = (p + \widehat{a}) G(0)^{-1} Q_{\widehat{a}}(s) \quad \text{where} \quad \widehat{a} := \frac{a_{\max}}{2}.$$

Approximation of $Q_{\widehat{a}}(s)$ by a first order controller, then, gives a lead or lag controller in the form $C_\ell(s)$. The above approach (and other alternative methods of approximating the plant first and then designing a low order controller) must be further compared with the proposed design of Sect. 2.4.1 on practical application examples. Currently, this is left open for a future study.

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