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Umberto Viaro

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Chapter 7

Stability margin design

A problem considered with interest in the recent control literature (see, e.g., [1] $\div$ [4]) is that of determining a controller that maximizes a given stability margin or achieves a tradeoff between different stability margins so as to confer robustness to the control system. The solution to this problem is expectedly more difficult when the plant transfer function $G_p(s)$ has both poles and zeros in the right half–plane (RHP). In this case, in fact, the achievable margins may not exceed certain upper bounds depending on the location of the RHP poles and zeros of $G_p(s)$. These bounds have been expressed in [2, p. 202] in terms of the $\mathcal{H}_\infty$ norm of the closed–loop transfer function (complementary sensitivity function)

$$W(s) = \frac{G_c(s)G_p(s)}{1 + G_c(s)G_p(s)}, \tag{7.1}$$

where $G_c(s)$ is the controller transfer function.

For instance, assuming that the RHP poles p_{p_i} and zeros z_{p_i} of $G_p(s)$ are simple, the upper bound m_φ^* on the absolute value of the achievable phase margin m_φ is

$$m_\varphi^* = 2 \arcsin \left[\frac{1}{\gamma_{opt}} \right], \tag{7.2}$$

where

$$\gamma_{opt} = \inf \|W(s)\|_\infty \tag{7.3}$$

subject to

$$W(z_{p_i}) = 0, \;\; W(p_{p_i}) = 1. \tag{7.4}$$

It has been observed (see, e.g., [5] ÷ [8]) that a synthesis approach based on a single optimization criterion may give rise to controllers that are fragile with respect to other criteria. This difficulty arises, for example, when a near–optimal phase margin is sought, since this criterion usually leads to very small gain margins m_g. Therefore, it is more reasonable to seek a balance between m_φ and m_g, especially when the attainable margins are tiny. In the case of unstable plants, such a balance is obtained when the distance from the critical point $-1 + \jmath 0$ of the intersections of the Nyquist diagram of the loop function

$$L(s) = G_c(s)G_p(s) \tag{7.5}$$

with the real axis is (almost) equal to the distance of its intersections with the unit circle centred at the origin.

However, both the phase and the gain margins can be large and yet the Nyquist diagram of the loop transfer function can pass close to the critical point. A better measure of the stability robustness is provided by the distance δ_c from the critical point to the nearest point on the Nyquist plot of the loop transfer function. This distance is given by

$$\delta_c = \frac{1}{\sup_\omega |S(\jmath\omega)|} \tag{7.6}$$

which is the reciprocal of the $\mathcal{H}_\infty$ norm of the sensitivity function

$$S(s) = \frac{1}{1 + L(s)} = 1 - W(s). \tag{7.7}$$

The minimization of the $\mathcal{H}_\infty$ norm of $S(s)$ leads to an allpass $S(s)$ [1, Lemma 1], that is, with $|S(\jmath\omega)|$ constant. In turn, this fact implies that the Nyquist diagram of $L(\jmath\omega)$ is a circle with centre at $-1 + \jmath 0$, so that the distance from the critical point to all of the points on the Nyquist plot of the loop function $L(\jmath\omega)$ is the same.

Section 7.1 provides a characterization of the internally stabilizing controllers that ensure a circular Nyquist plot of the loop function around the critical point. On the basis of this characterization, a simple analytic procedure to find the desired controller is derived in Section 7.2 for an exactly–proper plant. In Section 7.3, the procedure is extended to strictly–proper plants. Section 7.4 shows how a controller that maximizes either m_φ or m_g can be obtained from the controller that produces a circular Nyquist plot. The procedures are illustrated by numerical examples. The following exposition follows closely those in [9] and [10].

7.1 Controller characterization

Consider first a plant characterized by an exactly–proper transfer function $G_p(s)$ with all of its zeros and poles in the RHP, and assume that these poles and zeros are simple (more general cases will be considered later).

According to the Nyquist stability criterion, the Nyquist diagram of the loop function $L(s)$ must encircle the critical point counterclockwise a number of times equal to the number n_L of the RHP poles of $L(s)$. The desired circular shape is attained only if $L(s)$ is exactly proper (excluding the case in which $L(\jmath\omega)$ travels along the circle with unit radius and centre in $-1 + \jmath 0$ and arrives at the origin as $\omega \to \infty$). Therefore, the controller transfer function $G_c(s)$ turns out be exactly proper such as $G_p(s)$.

If the diagram of $L(\jmath\omega)$ is a circle centred at $-1+\jmath 0$, then the diagram of the return difference

$$D(s) = 1 + L(s), \tag{7.8}$$

is a circle centred at the origin, which implies that $D(s)$ is an allpass function of the form:

$$D(s) = K_D \frac{\prod_{i=1}^{n_L}(s + q_i)}{\prod_{i=1}^{n_L}(s - q_i)} \tag{7.9}$$

whose poles q_i are the negatives of its zeros $-q_i$. According to [2, pp. 36–37], the following lemma can be stated.

Lemma 7.1.1 *Under the previous assumptions, the control system is internally stable if and only if: (i) $\mathbb{R}\mathrm{e}[q_i] > 0$, (ii) all of the n_p RHP poles p_{p_i} of $G_p(s)$ are also poles of $D(s)$, and (iii) at the n_p RHP zeros z_{p_i} of $G_p(s)$ the interpolation conditions $D(z_{p_i}) = 1$ are satisfied.*

An immediate consequence of condition (i) in Lemma 7.1.1 is that all of the $n_c = n_L - n_p$ poles of $G_c(s)$, if any (a proportional controller might be enough), are in the RHP, as stated by the next lemma.

Lemma 7.1.2 *If the control system is internally stable, the zeros of $G_c(s)$ are in the RHP.*

Proof Consider the locus described by the roots of

$$1 + D(s) = 0 \tag{7.10}$$

as K_D varies over the extended real axis $\overline{\mathbb{R}}$. This root locus is formed by n_L branches, each of which passes through one pole q_i of $D(s)$ for $K_D = 0$ and through its zero $-q_i$ for $K_D = \infty$, and is symmetric with respect to the imaginary axis which is crossed for $|K_D| = 1$ (in particular, the point at infinity is crossed for $K_D = -1$). Therefore, for $|K_D| \neq 1$ all of the n_L roots of (7.10) are in the same half–plane. Since the zeros of $G_c(s)$ are crossed by locus branches for the same value of K_D as the RHP zeros of $G_p(s)$, they too must belong to the RHP. □

If the process and controller transfer functions are denoted, respectively, by

$$G_p(s) = K_p \frac{\prod_{i=1}^{n_p}(s - z_{p_i})}{\prod_{i=1}^{n_p}(s - p_{p_i})}, \tag{7.11}$$

$$G_c(s) = \frac{K_c s^{n_c} + \sum_{i=0}^{n_c-1} b_i s^i}{s^{n_c} + \sum_{i=0}^{n_c-1} a_i s^i}, \tag{7.12}$$

then $D(s)$ can be expressed as

$$D(s) = \frac{\left[s^{n_c} + \sum_{i=0}^{n_c-1} a_i s^i\right] \prod_{i=1}^{n_p}(s - p_{p_i}) + \left[K_c s^{n_c} + \sum_{i=0}^{n_c-1} b_i s^i\right] K_p \prod_{i=1}^{n_p}(s - z_{p_i})}{\left[s^{n_c} + \sum_{i=0}^{n_c-1} a_i s^i\right] \prod_{i=1}^{n_p}(s - p_{p_i})} \tag{7.13}$$

whose high–frequency gain (ratio of the coefficients of the highest power of s at the numerator and denominator) is given by

$$K_D = 1 + K_c K_p. \tag{7.14}$$

7.2 Synthesis procedure

In order for (7.13) to exhibit the form (7.9), the following polynomial identity must be satisfied:

$$\left[s^{n_c} + \sum_{i=0}^{n_c-1} a_i s^i\right] \prod_{i=1}^{n_p}(s - p_{p_i}) + \left[K_c s^{n_c} + \sum_{i=0}^{n_c-1} b_i s^i\right] K_p \prod_{i=1}^{n_p}(s - z_{p_i})$$

$$= (-1)^{n_c+n_p}(1 + K_c K_p)\left[(-s)^{n_c} + \sum_{i=0}^{n_c-1} a_i(-s)^i\right] \prod_{i=1}^{n_p}(-s - p_{p_i}). \quad (7.15)$$

Equating the coefficients of s^i, $i = 0, \ldots, n_c + n_p - 1$, on both sides of (7.15) (the coefficients of $s^{n_c+n_p}$ are necessarily equal) leads to a set of $n_e = n_c+n_p$ equations in the $n_x = 2n_c+1$ unknown parameters of (7.12). These equations are nonlinear because the unknown high–frequency gain K_c of (7.12) is multiplied by the a_is.

An efficient procedure to arrive at a controller of minimal order n_c compatible with a $D(s)$ of the desired form consists of the following steps:
(i) set $n_c = n_p - 1$ (in general, (7.15) does not admit solutions for $n_c < n_p - 1$);
(ii) by considering K_c as a parameter, obtain from (7.15) a set of $2n_c+1$ equations linear in the $2n_c$ unknown coefficients a_i and b_i,
(iii) find the values of K_c that annihilate the determinant of the $(2n_c + 1) \times (2n_c+1)$ system matrix (only for these values a solution may exist);
(iv) for every real value of K_c, if any, check whether the set of equations admits a solution and, in this case, whether this solution leads to an unstable and nonminimum–phase controller (which ensures that the Nyquist diagram of $L(s)$ encircles counterclockwise the critical point $n_c + n_p$ times, as required by the Nyquist criterion);
(v) if more admissible solutions exist, choose the one corresponding to the largest radius $\rho = |K_D| = |1 + K_c K_p|$ of the circular Nyquist diagrams for both $D(s)$ and $L(s)$; if no admissible solution exists, resort to a more complicated controller and repeat the procedure (see later).

Essentially, point (iii) requires the solution of an algebraic equation and point (iv) the solution of sets of $2n_c$ equations in the $2n_c$ unknown coefficients a_i and b_i. Of course, an admissible solution might entail cancellations in the resulting controller transfer function, which implies that the actual order of the controller is less than n_c. This rare situation occurs for particular pole–zero distributions of the plant transfer function [9]. Concerning point (v), to increase the controller complexity, it is reasonable to resort to a new controller transfer function of immediately higher order. To this purpose, a fixed far–off RHP pole is added to the n_c poles to be determined. In this way, the number of denominator parameters in not increased whereas the numerator contains an additional unknown coefficient. Therefore, the number of equations still matches the number of unknowns.

Example If the plant transfer function is [3]

$$G_p(s) = \frac{(s-1)(s-3)}{(s-2)(s-4)}, \tag{7.16}$$

eqn. (7.15) with $n_c = n_p - 1 = 1$ particularizes to

$$\begin{aligned}&(s+a_0)(s-2)(s-4) + (K_c s + b_0)(s-1)(s-3)\\ &= -(1+K_c)(-s+a_0)(-s-2)(-s-4).\end{aligned} \tag{7.17}$$

Equating the coefficients of the equal powers of s on both sides of (7.17) leads to

$$\begin{aligned}(2+K_c)a_0 + b_0 &= 10K_c + 12,\\ 6K_c a_0 - 4b_0 &= 5K_c,\\ (16+8K_c)a_0 + 3b_0 &= 0.\end{aligned} \tag{7.18}$$

The system matrix is then:

$$A = \begin{bmatrix} 2+K_c & 1 & -(10K_c+12) \\ 6K_c & -4 & -5K_c \\ 16+8K_c & 3 & 0 \end{bmatrix} \tag{7.19}$$

whose determinant is equal to 0 for $K_c = -1.44428$ and $K_c = -1.01286$. The latter value leads to the acceptable solution: $a_0 = -1.13746$, $b_0 = 2.99421$. Therefore, the transfer function of the internally stabilizing controller ensuring the desired circular form of the loop Nyquist diagram, depicted in Fig. 7.1, is:

$$G_c(s) = -1.01286\,\frac{s-2.95619}{s-1.13746}. \tag{7.20}$$

The tiny radius of this diagram is $\rho = |K_D| = |1 + K_c K_p| = 0.01286$. Correspondingly, the $\mathcal{H}_\infty$ norm of the sensitivity function is $\|S(s)\|_\infty = 1/0.01286 = 77.760497$, which coincides with its lower bound. The absolute value of the upper and lower phase margins (related, respectively, to the lower and upper intersections of the loop Nyquist diagram with the unit circle centred at the origin) is equal to $0.7368°$. The upper and lower gain margins (related, respectively, to the right and left intersections with the real axis) are $m_{g,u} = 0.112425$ dB and $m_{g,l} = -0.110988$ dB.

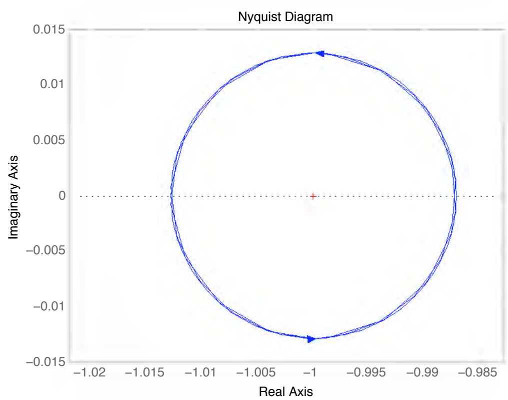

Figure 7.1: Tiny circular Nyquist diagram of the loop function $L(s) = G_c(s)G_p(s)$ with $G_p(s)$ and $G_c(s)$ equal to (7.16) and (7.20), respectively .

7.3 Extension to strictly–proper plants

It is shown next that the procedure of Section 7.2 can be extended to the case of plants that are characterized by a strictly–proper transfer function and exhibit left half–plane (LHP) poles and zeros, too.

Assume first that all of the poles and zeros of $G_p(s)$ are in the RHP and denote by η its pole–zero excess. Then, the procedure of Section 7.2 can be applied to

$$\hat{G}_p(s) = \frac{1}{(-z)^\eta} G_p(s)(s - z)^\eta, \tag{7.21}$$

where the additional RHP zero z is much greater that the largest magnitude of all the poles and zeros of $G_p(s)$. In this way, the zero at infinity of multiplicity η of $G_p(s)$ is converted into a finite but large RHP zero of equal multiplicity of $\hat{G}_p(s)$, while the low–frequency gain is retained. The resulting controller transfer function $G_c(s)$ will make the Nyquist diagram of $\hat{L}(s) = G_c(s)\hat{G}_p(s)$ circular around the critical point. Correspondingly, a long initial part of the Nyquist diagrams of $L(s) = G_c(s)G_p(s)$ and $\hat{L}(s)$ will almost overlap, and their stability margins will be practically the same.

The LHP poles and zeros of $G_p(s)$, if any, can safely be cancelled without endangering internal stability. However, if the number of LHP poles exceeds that of the LHP zeros, a remote pole of suitable multiplicity must be included in the controller to ensure that its transfer function remains proper.

Example Assume that the plant transfer function is

$$G_p(s) = \frac{(s-2)(s+4)}{(s-1)(s-3)(s+5)}. \tag{7.22}$$

Form the "augmented" plant transfer function:

$$\hat{G}_p(s) = -\frac{1}{z}\, G_p(s)\, \frac{(s-z)(s+5)}{s+4} \tag{7.23}$$

with z large compared to the magnitude of the other poles and zeros. By applying the procedure of Section 7.2 to (7.23) with $z = 1000$, the following internally stabilizing controller transfer function is found:

$$G_c(s) = \frac{933.77s - 1400}{s - 874.8336} \tag{7.24}$$

leading to a perfectly circular Nyquist diagram of $\hat{L}(s) = G_c(s)\hat{G}_p(s)$ with upper and lower phase margin equal to $\hat{m}_\varphi = \pm 3.792°$, and upper and lower gain margins equal, respectively, to $\hat{m}_{g,u} = 0.5952$ dB and $\hat{m}_{g,l} = -0.5568$ dB.
The controller transfer function for the actual plant (7.22) can be chosen as:

$$\hat{G}_c(s) = \frac{s+5}{s+4}\, G_c(s). \tag{7.25}$$

The Nyquist diagram of $L(\jmath\omega) = \hat{G}_c(\jmath\omega)G_p(\jmath\omega)$ tends to the origin after encircling the critical point the required number of times. However, its

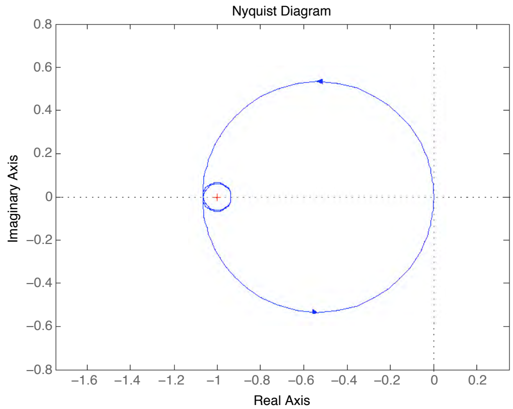

Figure 7.2: Nyquist diagram for the loop function $L(s) = \hat{G}_c(s)G_p(s)$ with $G_p(s)$ and $\hat{G}_c(s)$ equal to (7.22) and (7.25), respectively.

stability margins differ only slightly from those for $\hat{L}(s)$. Precisely they are: $m_\varphi = \pm 3.534°$, $m_{g,u} = 0.5952$ dB and $m_{g,l} = -0.5471$ dB. The Nyquist diagram for $L(s)$ is shown in Fig. 7.2.

7.4 Margin maximization

If the Nyquist diagram of $L(\jmath\omega)$ is a circle centred at $-1+\jmath 0$ with radius ρ, the absolute value of the upper and lower gain margins are not equal because the one corresponding to the intersection with the real axis at the right of the critical point is larger than the one corresponding to the intersection at the left of the same point. On the other hand, the Nyquist diagram retains the circular shape if $L(s)$ is multiplied by a factor k, which modifies its gain. Simple geometric considerations show that the absolute values of both gain margins become equal, so that the smaller of the two is maximized, for

$$k = \frac{1}{\sqrt{1-\rho^2}}. \tag{7.26}$$

Moreover, for this value of k the derivative dm_φ/dk equals zero and the absolute value $|m_\varphi|$ of the upper and lower phase margins reaches its maximum. For instance, by multiplying (7.20) by $k = 1/\sqrt{1-0.01286^2} =$

1.0000827, the absolute value of both phase margins for the example of Section 7.2 becomes 0.737° and that of both gain margins 0.112 dB. Clearly, in this delicate case, the improvements are tiny.

If it is no longer required that the loop Nyquist diagram be circular, either the phase or the gain margins can be broadened. A simple procedure to maximize one of them is outlined next.

Let $G_c(s)$ be the internally stabilizing controller transfer function that leads to a circular loop Nyquist plot according to the procedure of Section 7.2, and denote by m_φ the absolute value of the related phase margins. Replacing such $G_c(s)$ by

$$G_{c,a}(s) = -G_c(s)[G_c(-s)G_p(-s)]^{-1} = -G_c(s)L^{-1}(-s) \tag{7.27}$$

the loop transfer function becomes

$$L_a(s) = -L(s)L^{-1}(-s) \tag{7.28}$$

which has the same number of RHP poles as $L(s)$ but whose order is $2n_L$.
The magnitude and phase of $L_a(\jmath\omega)$ are

$$|L_a(\jmath\omega)| = 1, \forall\omega, \text{ and } \arg[L_a(\jmath\omega)] = -\pi + 2\arg[L(\jmath\omega]. \tag{7.29}$$

Therefore, the Nyquist diagram of (7.28) is superimposed on the arc of the unit circumference centred at the origin whose phase is included between $-\pi - 2m_\varphi$ and $-\pi + 2m_\varphi$ (like the arc inside the narrow ellipse represented in Fig. 7.3). By indicating with ω_i the $n_L - 1$ positive angular frequencies such that $\arg[L_a(\jmath\omega)] = -\pi$, the $2n_L$ poles of the closed–loop function

$$W_a(s) = \frac{L_a(s)}{1 + L_a(s)} \tag{7.30}$$

are: $0, \pm\omega_i$ $(i = 1, \ldots, n_L - 1)$ as well as ∞. These points correspond to the $2n_L$ intersections with the imaginary axis of the $2n_L$ branches of the root locus for the equation

$$1 + K_a L_a(s) = 0 \tag{7.31}$$

whose departure and arrival points are:
(i) the n_L RHP poles of $L(s)$ and the n_L LHP poles of $L^{-1}(-s)$ (that are the negatives of the n_L RHP zeros of $L(s)$);

(ii) the n_L RHP zeros of $L(s)$ and the n_L LHP zeros of $L^{-1}(-s)$ (that are the negatives of the n_L RHP poles of $L(s)$).
For $K_a = -1$ in (7.31) n_L branches arrive at the imaginary axis from the left and n_L branches arrive there from the right. Indeed, controller (7.27) leads to such a critical situation in which the gain margins are equal to zero but the absolute value of the phase margins is maximal. From this controller, however, an internally stabilizing controller leading to a near–maximal value of the phase margins can be obtained by slightly perturbing the parameters of (7.27) as follows.

If the LHP base points of $L_a(s)$, that is, the poles and zeros of $L^{-1}(-s)$, are shifted slightly towards the left, for $K_a = -1$ the points on the related new root locus are all in the LHP, except for one point at infinity. Therefore, by setting either $K_a = -1 + \varepsilon$ or $K_a = -1 - \varepsilon$ (whatever applies), with ε suitably small, it is possible to locate all of the poles of the closed–loop transfer function in the LHP, as shown by the following example.

Example Consider again the plant transfer function (7.22). For $G_c(s)$ as in (7.20), the controller transfer function (7.27) particularizes to

$$G_{c,a}(s) = -G_c(s)L^{-1}(-s) = \frac{(s-2.95619)(-s-1.13746)(-s-2)(-s-4)}{(s-1.13746)(-s-2.95619)(-s-1)(-s-3)}. \tag{7.32}$$

Correspondingly, the Nyquist diagram of the loop transfer function $L_a(s)$ is an arc passing through $-1 + \jmath 0$ (see arc in Fig 7.3) and all of the poles of the closed–loop transfer function are purely imaginary. Shifting the LHP poles and zeros of (7.32) to the left by 0.01 and making its high–frequency gain equal to $-1 - \varepsilon = -1.000005$ lead to the controller transfer function

$$G_{c,\varphi}(s) = -1.000005\,\frac{(s-2.95619)(s+1.14746)(s+2.01)(s+4.01)}{(s-1.13746)(s+2.96619)(s+1.01)(s+3.01)} \tag{7.33}$$

which results in $m_\varphi = 1.447°$, only a little smaller than its upper bound 1.4739°. The resulting loop Nyquist diagram is similar to the narrow ellipse surrounding the critical point shown in Fig. 7.3. Observe, by way of comparison, that the controller derived in [3] for the same plant with the objective of making m_φ greater than $1°$ leads to the following lower and upper phase margins: $1.0428°$ and $-1.0654°$.

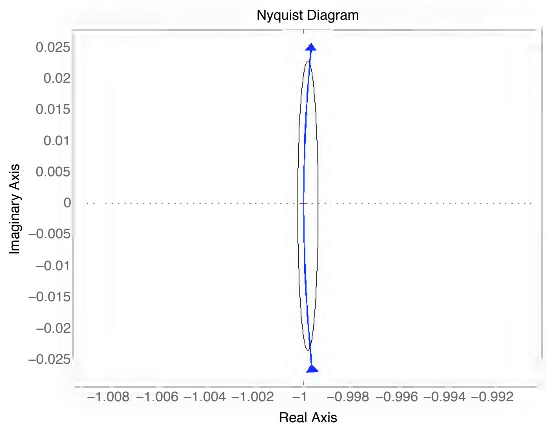

Figure 7.3: Nyquist diagram for the loop function $L_a(s)$ defined in (7.28) with $G_p(s)$ and $G_c(s)$ given by (7.16) and (7.20), respectively (arc inside the ellipse), and Nyquist diagram for the slightly modified loop function corresponding to (7.33) (narrow ellipse).

A procedure similar to the one used for maximizing m_φ can be conceived for maximizing the absolute value of the gain margins. In this case, it is enough to shift slightly to the left the LHP poles and zeros of

$$L_b(s) = -L(s)L(-s) \tag{7.34}$$

whose Nyquist diagram is superimposed to a horizontal *segment* passing through the critical point.

7.5 Concluding remarks

Stability margins are important measures of stability robustness. However, the gain and phase margins can be large and yet the Nyquist diagram of the loop function $L(\jmath\omega)$ can pass close to the critical point. A more meaningful measure of stability robustness, especially when the plant is unstable and nonminimum–phase, is provided by the minimal distance of this diagram from the critical point, which is the inverse of the $\mathcal{H}_\infty$ norm of the sensitivity function $S(s)$. Indeed, the minimization of this norm leads to $|S(\jmath\omega)| = \text{const}$ and, thus to $|1 - L(\jmath\omega)| = \text{const}$. In agreement with this result, a stabilization procedure that ensures a circle–shaped loop Nyquist diagram around $-1 + \jmath 0$ has been suggested.

The procedure, which leads to a minimal–order internally stabilizing controller, entails the solution of a simple algebraic equation and a set of linear equations. Both the case of an exactly–proper plant and that of a strictly–proper plant have been considered. It has been shown how from the controller obtained according to the aforementioned procedure another controller can be constructed that almost doubles either the gain margin or the phase margin, thus approaching their respective upper bounds. However, the order of the resulting controller is higher than that of the controller that ensures a balance between the two margins and the robustness of the design decreases.

The procedure has been illustrated by means of a rather critical example [3] for which the phase margin may not exceed 1.4739°.

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Often, short papers tend to be sharper than longer works because they focus on a single theme without lingering on unessential aspects, thus showing clearly the signicance of a contribution or an idea. The author of this book had the privilege of collaborating for over a quarter of a century with Antonio Lepschy (1931-2005), a recognized leader of the Italian control community.

Lepschy had a liking for the brief paper format, so that many results obtained by his research team were published in this way. The present compilation tells a few of these short stories, duly updated, trying to preserve their original flavour.

Umberto Viaro (http://umbertoviaro.blogspot.com/) has been professor of System and Control Theory at the University of Udine, Italy, since 1994. His 25-year-long collaboration with Antonio Lepschy resulted in more than 100 joint papers and two books. An essential role in this research activity was played by Wiesław Krajewski of the Systems Research Institute, Polish Academy of Sciences. The current research interests of Umberto Viaro concern optimal model reduction, robust control, switching and LPV control. He is the author or coauthor of 4 books and about 180 research papers.

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