

For reference, we give here the formulas for addition on an elliptic curve in the so-called Weierstrass's normal form. If $P_1 = (x_1, y_1)$ and $P_2 = (x_2, y_2)$ are points of the curve

$$Y^2 + a_1XY + a_3Y - (X^3 + a_2X^2 + a_4X + a_6),$$

then

$$-P_1 = (x_1, -y_1 - a_1x_1 - a_3), \quad P_1 + P_2 = P_3^* = (x_3, -y_3),$$

where

$$x_3 = \lambda(\lambda + a_1) - a_2 - x_1 - x_2, \quad y_3 = (\lambda + a_1)x_3 + \mu + a_3,$$

and

$$\begin{aligned} \lambda &= (y_2 - y_1)/(x_2 - x_1), \quad \mu = (y_1x_2 - y_2x_1)/(x_2 - x_1) \quad \text{if } x_1 \neq x_2; \\ \lambda &= (3x_1^2 + 2a_2x_1 + a_4 - a_1y_1)/N, \quad \mu = (-x_1^3 + a_4x_1 + 2a_6 - a_3y_1)/N, \end{aligned}$$

with $N = 2y_1 + a_1x_1 + a_3$ if $x_1 = x_2$, $P_2 \neq -P_1$.

An algorithm for obtaining a minimal model of an elliptic curve is described in Laska [40]. Other algorithms connected with elliptic curves are given in Cremona [21].

The original conjecture of Birch and Swinnerton-Dyer was generalized by Tate [54] and Bloch [11]. For a first introduction to the theory of modular forms see Serre [49], and for a second see Lang [39].

Hasse actually showed that, if \mathcal{E} is an elliptic curve over any finite field \mathbb{F}_q containing q elements, then the number N_q of \mathbb{F}_q -points on \mathcal{E} (including the point at infinity) satisfies the inequality

$$|N_q - (q + 1)| \leq 2q^{1/2}.$$

For an elementary proof, see Chahal [17]. Hasse's result is the special case, when the genus $g = 1$, of the Riemann hypothesis for function fields, which was mentioned in Chapter IX, §5.

It follows from the result of Siegel (1929), mentioned in §9 of Chapter IV, and even from the earlier work of Thue (1909), that an elliptic curve with integral coefficients has at most finitely many *integral* points. However, their method is not constructive. Baker [6], using the results on linear forms in the logarithms of algebraic numbers which he developed for the theory of transcendental numbers, obtained an explicit upper bound for the magnitude of any integral point in terms of an upper bound for the absolute values of all coefficients. Sharper bounds have since been obtained, e.g. by Bugeaud [15]. (For modern proofs of Baker's theorem on the linear independence of logarithms of algebraic numbers, see Waldschmidt [59]. The history of Baker's method is described in Baker [7].)

For information about the proof of Mordell's conjecture we refer to Bloch [12], Szpiro [53], and Cornell and Silverman [19]. The last includes an English translation of Faltings' original article. As mentioned in §9 of Chapter IV, Vojta (1991) has given a proof of the Mordell conjecture which is completely different from that of Faltings. There is an exposition of this proof, with simplifications due to Bombieri (1990), in Hindry and Silverman [32].

For congruent numbers, see Volume II, Chapter XVI of Dickson [23], Tunnell [57], and Noda and Wada [43]. The survey articles of Goldfeld [30] and Oesterlé [44] deal with Gauss's class number problem.

References for earlier work on Fermat's last theorem were given in Chapter III. Ribet [47] and Cornell *et al.* [20] provide some preparation for the original papers of Wiles [60] and Taylor and Wiles [56]. For the *TW*-conjecture, see also Darmon [22]. For Euler's conjecture, see Elkies [25].

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Notations

 $\in, \notin, =, \neq, \emptyset, \subseteq, \subset, \cup, \cap, 2$
 $B \setminus A, A^c, 3$
 $A \times B, A^n, aRb, 3$
 $R_a, f : A \rightarrow B, f(a), f(A), 4$
 $i_A, g \circ f, 4$
 $f^{-1}, \mathbb{N}, 1, S(n), 5$
 $s_m(n), 6$
 $a + b, p_m(n), a \cdot b, 7$
 $<, \leq, >, \geq, 8$
 $I_n, 9$
 $\#(E), 10$
 $\sim, \mathbb{Z}, +, 11$
 $0, -a, b - a, \cdot, 1, 12$
 $P, -P, 13$
 $P + P, P \cdot P, a^2, a < b, 14$
 $a/b, \mathbb{Z}^\times, \sim, \mathbb{Q}, 15$
 $+, \cdot, a^{-1}, 16$
 $P, -P, 16, 17$
 $\mathcal{P}, A < B, 18$
 $A + B, 19$
 $AB, 20$
 $\mathbb{R}, \sim, 22$
 $\sqrt[n]{a}, a^{1/2}, b^n, \sqrt[n]{a}, a^{1/n}, \mathbb{R}, 23$
 $\lim_{n \rightarrow \infty}, a_n \rightarrow l, n \rightarrow \infty, 24$
 $\inf, \sup, \underline{\lim}_{n \rightarrow \infty}, \overline{\lim}_{n \rightarrow \infty}, 24$
 $[a, b], 26$
 $|a|, d(a, b), 27$
 $\beta_\delta(x), \overline{A}, \text{int } A, \mathbb{R}^n, |a|, 28$
 $|a|_1, |a|_2, \mathbb{F}_2^n, 28$
 $\mathcal{C}(I), |f|, |f|_1, |f|_2, \mathcal{C}(\mathbb{R}), \mathbb{F}_2^\infty, 29$
 $\lim_{n \rightarrow \infty} a_n = a, a_n \rightarrow a, 30$
 $\overline{E}, 31$
 $L(I), L^2(I), 32$
 $\varphi'(x_0), 33$
 $|A|, 34$
 $B_r, 35$
 $e, 38, 187$
 $e^t, 39, 45$
 $\ln x, \log x, 39$
 $\mathbb{C}, i, 40$
 $\bar{z}, \Re z, \Im z, 41$
 $\cos z, \sin z, 45$
 $\pi, 46, 48, 186, 217, 364, 509$
 $\mathbb{H}, \overline{A}, n(A), t(A), 49$
 $i, j, k, 51$
 $V(u), 52$
 $\langle x, y \rangle, SU_2(\mathbb{C}), SO_3(\mathbb{R}), S^3, P^3(\mathbb{R}), 53$
 $SO_4(\mathbb{R}), \mathbb{O}, \varepsilon, 53$
 $\overline{\alpha}, n(\alpha), 54$
 $e, a^{-1}, 55$
 $ab, HK, 56$
 $\mathcal{S}_n, \text{sgn}(\alpha), \mathcal{A}_n, 57$
 $Ha, G/H, 58$
 $a^n, \langle a \rangle, \langle S \rangle, 59$
 $N_a, G \times G', 60$
 $M_n(\mathbb{Z}), \mathcal{P}(X), A + B, AB, 61$
 $na, a^{-1}, \mathbb{R}^\times, 62$
 $R/S, 63$
 $R \oplus R', av, v + w, D^n, 64$
 $\mathcal{C}(I), O, \mathcal{C}^1(I), U_1 + U_2, U_1 \oplus U_2, 65$
 $\langle S \rangle, 66$
 $\dim V, [E : F], e_1, \dots, e_n, Tv, TS, 67$
 $S + T, GL(V), M_n(F), 68$
 $V \otimes V', T \otimes T', 68$
 $M_n(D), 69$
 $\langle u, v \rangle, \|v\|, 71$

\mathcal{B} , 74 $\ell^2, L^2(I)$, 75 $*\mathbb{R}$, 76 $b|a, b \nmid a, \times, (a, b)$, 83 $[a, b]$, 84 $a \wedge b, a \vee b$, 85 $(a_1, a_2, \dots, a_n), [a_1, a_2, \dots, a_n]$, 86 $K[t]$, 87, 96 $K(t)$, 88, 262 ${}^m C_n$, 92, 111 $\partial(f), |f|, R[t], R[[t]]$, 96 $K[t, t^{-1}]$, 98 $c(f)$, 100 $R[t_1, \dots, t_m]$, 101 $\Phi_p(x)$, 102 f' , 103 $\delta(a)$, 104 $\mathbb{Q}(\sqrt{d}), \mathcal{O}_d$, 105 $a \equiv b \pmod{m}, \not\equiv$, 106 $\mathbb{Z}_{(m)}$, 107 $\mathbb{Z}_{(m)}^\times$, 108 $\mathbb{F}_p, \varphi(m)$, 109 $\Phi_n(x)$, 111, 112 $\tilde{f}(x)$, 111 \mathbb{F}_p^\times , 114 $\mathcal{G}, N(\gamma)$, 119 $\mathcal{H}, \tilde{\gamma}$, 120 $N(\gamma), (\alpha, \beta)_r$, 121 $\lfloor x \rfloor, g(k), w(k)$, 122 $G(k)$, 123 $K[[t_1, \dots, t_m]]$, 124 \mathbb{F}_q , 125 $(a/n), \operatorname{sgn}(\pi_a)$, 130 (a/p) , 133 $G(m, n)$, 137 $\mathbb{Q}(\sqrt{d}), \alpha'$, 140 $N(\alpha), \omega, \mathcal{O}_d, \mathcal{G}, \mathcal{E}$, 141 $(a_1, \dots, a_m), AB$, 145 A' , 146 $h(d), \mathcal{O}(K)$, 151 $f * g$, 152 $\delta(n), \mathcal{A}, |f|$, 153 $i(n), j(n)$, 154 $\tau(n), \sigma(n)$, 155 $\mu(n), \hat{f}(n)$, 156 M_p , 158 γ , 159, 380 F_n , 160 $GL_n(\mathbb{Z})$, 162 $A \oplus B$, 163, 301 $M_1 \cap M_2, M_1 + M_2$, 166 Δ_k , 171 $|a|$, 173 (f/g) , 174 $\lfloor \xi \rfloor, \xi_n, \tau$, 179 $[a_0, a_1, a_2, \dots]$, 179, 182, 212 p_n, q_n , 180 p_n/q_n , 181, 212 $[a_0, a_1, \dots, a_N]$, 182 $M(\xi)$, 190 D , 191 ζ' , 192 $[a_0, \dots, a_{m-1}, \overline{a_m, \dots, a_{m+h-1}}]$, 192 \mathcal{C} , 198 \mathcal{H} , 201 $\Gamma, SL_2(\mathbb{Z}), T(z), S(z)$, 202 $R(z), \partial(F)$, 203 \tilde{F} , 204 $\tau(f)$, 206 $h^\dagger(D)$, 206, 207 \mathcal{H}/Γ , 209, 218 $\Gamma(n), \xi^{(k)}, \mu_k$, 210 $|f|, \lfloor f \rfloor, \{f\}$, 211 $PSL_2(\mathbb{R})$, 218 $\det A$, 224-229 $M_n, \operatorname{diag}[\alpha_{11}, \alpha_{22}, \dots, \alpha_{nn}]$, 225 $SL_n(F), A^t$, 226, 229 $\|v\|$, 229, 234 $A \otimes B$, 231 J_m , 232, 248 e_m , 233, 248 $2\text{--}(v, k, \lambda)$, 247 $t\text{--}(v, k, \lambda)$, 250 $C_p, \mathcal{A}_n, PSL_n(q)$, 251 $M_{12}, M_{11}, M_{24}, M_{23}, M_{22}$, 251 $|x|$, 254 $[n, k, d], C(H), G_{24}, R(1, m)$, 255 $\mathbb{C}_\zeta((z)), \mathbb{C}_\zeta[[z]], |a|$, 261 $|a|_\infty, v_p(a), |a|_p, K(t)$, 262

$|f|_\infty, v_p(f), |f|_p, 262$
 $K((t)), 263$
 $\overline{F}, 270, 271$
 $\mathbb{Q}_p, \|a\|, 271$
 $R, M, U, 274$
 $\mathbb{Z}_p, 275$
 $\pi, k, 276$
 $\overline{f}(x), 280$
 $F^\times, F^{\times 2}, (u, v), 292$
 $f \sim g, \det V, U^\perp, V_1 \perp V_2, 293$
 $\text{ind } V, 296$
 $\text{ind}^+ V, \text{ind}^- V, 297$
 $\tau_w, 300$
 $A \oplus B, 301$
 $V \approx V', \mathscr{W}(F), (a, b)_F, 303$
 $f_a, G_a, \mathbb{Q}_\infty, (a, b)_\infty, (a, b)_p, 304$
 $f_{a,b}, G_{a,b}, 307$
 $s_F(f), s_p(f), 310$
 $\mathbb{Q}_v, |\cdot|_v, (a, b)_v, 313$
 $(a, b/F), 324$
 $\chi(x), \lambda(S), 327$
 $\kappa_n, 328, 348, 380$
 $\|q\|, 329$
 $< Y >, 331$
 $d(A), \Pi, 333$
 $\text{int } S, A^*, 334$
 $B_\varepsilon, 337$
 $\mu_i(K, A), \Delta(K), 339$
 $K^*, 340$
 $|x|, 341$
 $d(y, z), \|x\|, (y, z), 342$
 $H_x, G_x, \overline{G}_x, 342$
 $V(x_0), \beta_r(x_0), 342$
 $V(A), 344$
 $B_R, m(A), 345$
 $\gamma(A), \gamma_n, 347$
 $\delta(K), \delta_n, 348$
 $e_1, \dots, e_n, A_n, D_n, E_8, E_7, E_6, 350$
 $A_{24}, 351$
 $\mathcal{L}_n, A_k \rightarrow A, 352$
 $\text{h}(K, K'), 352, 353$
 $\pi(x), \log x, 364$
 $\log^n x, Li(x), p_n, 365$
 $\theta(x), \psi(x), 367$
 $\lfloor y \rfloor, 368$
 $\Lambda(n), \zeta(s), 370$
 $\sigma, t, 371$
 $\hat{k}(u), k(t), 374, 375$
 $\vartheta(x), \Gamma(z), z!, 379$
 $B(x, y), \kappa_n, \gamma, Z(s), 380$
 $\alpha^*, 382$
 $\gamma_n, \tilde{\gamma}_n, 384$
 $|A|, \zeta_K(s), 384$
 $\pi_K(x), Z_K(s), 385$
 $N(P), N(A), \zeta_L(s), 386$
 $\tau(n), 388$
 $M(x), 390$
 $\pi_2(x), L_2(x), C_2, 392$
 $\pi(x; m, a), \theta(x; m, a), 400$
 $\psi(x; m, a), e, \chi, 400$
 $\chi_1, \hat{G}, g, 401$
 $G_m, 403$
 $L(s, \chi), 404$
 $\Lambda(s, \chi), 409$
 $\rho, 410, 434$
 $\rho_R, 410$
 $\rho \otimes \sigma, \rho_U, 411$
 $\text{tr} A, 414$
 $\chi(s), 414, 434$
 $g, \delta_{il}, \alpha_{ij}^{(\mu)}, n_\mu, 415$
 $\chi_\mu, \chi_R, 416$
 $\mathcal{C}_k, h_k, \chi_{ik}, 418$
 $C_k, C'_k, 418$
 $\sigma, \tilde{\sigma}, \hat{A}(s), 419$
 $\tilde{\psi}(s), 420$
 $\lambda, \psi_i^v, 421$
 $\mathcal{S}_n, \mathcal{A}_n, C_n, \omega, 423$
 $K_\rho, 426$
 $\mathcal{C}_0(G), M(f), 432$
 $\tilde{f}, (f, g), \mu(E), L^p(G), f * g, 433$
 $\mathcal{H}, \rho, \chi(s), \hat{G}, 434$
 $\hat{f}, \hat{\mu}, 435$
 $\mathcal{C}(G), 436$
 $SU(2), S^3, 437$
 $SO(3), 438$
 $\mathbb{T}^n, 439$
 $GL(n), O(n), U(n), Sp(n), 440$
 $[u, v], \mathfrak{gl}(n, \mathbb{R}), \mathfrak{gl}(n, \mathbb{C}), 440$
 $A_n, B_n, C_n, D_n, L(G), 440$
 $\tilde{G}, 441$
 $G_2, F_4, E_6, E_7, E_8, 442$
 $SU(n), SO(n), Spin(n), 442$

$[\xi], \{\xi\}, I, \varphi_{\alpha, \beta}(N), 448$
 $\chi_{\alpha, \beta}, 449$
 $e(t), 450$
 $\varphi_{a, b}(N), I^d, \{x\}, m \cdot x, 452$
 $D_N^* = D_N^*(\xi_1, \dots, \xi_N), \varphi_\alpha(N), 459$
 $\delta_N, 462$
 $D_N^*(x_1, \dots, x_N), 464$
 $\mathcal{B}, A \triangle B, \sigma(\mathcal{A}), \mu(B), 465$
 $(X, \mathcal{B}, \mu), \text{a.e.}, \int_X f d\mu, 466$
 $L(X, \mathcal{B}, \mu), T^{-1}B, 466$
 $\lambda, T_a, 472$
 $RA, 473$
 $p_1, \dots, p_r, [a_{-m}, \dots, a_m], 476$
 $\sigma, B_{p_1, \dots, p_r}, 477$
 $C_{i_1 \dots i_k}^{a_1 \dots a_k}, 478$
 $\tau, B_{p_1, \dots, p_r}^+, T, 479$
 $T_1 M, 482$
 $(X, d), \bar{A}, 485$
 $g(x), 496$
 $I_0, I_n, J_n(\gamma), 497$
 $g_\lambda(x), 498, 510$
 $U, V, 499, 531$
 $a_n, b_n, M(a, b), 503$
 $\mathcal{K}(a, b), 504, 505$
 $\mathcal{E}(a, b), e_n, 505$
 $\mathcal{P}(a, b, p), 505$
 $p_n, q_n, 506$
 $c, \mathcal{K}(a, c), c_n, 507$
 $\mathcal{E}(a, c), K(\lambda), E(\lambda), 508$
 $f_\lambda(x), 510$
 $S(t) = S(t, \lambda), 510, 526$
 $E(u), 516, 530$
 $\Pi(u, a), 517, 530$
 $q, z, \theta(v) = \theta(v; \tau), 519$
 $\theta_{\alpha, \beta}(v) = \theta_{\alpha, \beta}(v; \tau), 520$
 $\theta_{00}(v), \theta_{01}(v), \theta_{10}(v), \theta_{11}(v), 520$
 $\vartheta_1(\pi v, q), \vartheta_2(\pi v, q), 521$
 $\vartheta_3(\pi v, q), \vartheta_4(\pi v, q), Q_0, 521$
 $\text{sn } u, \text{cn } u, \text{dn } u, u = \pi \theta_{00}^2(0)v, 525$
 $\lambda(\tau), 525, 531$
 $\mathbf{K}(\tau), \mathbf{K}'(\tau), 527$
 $\Theta(u), \mathbf{E}(\mathbf{K}), 530$
 $\mathcal{H}, U, V, T, S, 531$
 $\mathcal{T}, \mathcal{T}', 533$
 $\mathcal{D}, \mathcal{D}^*, \bar{\mathcal{D}}, 534$
 $F(\alpha, \beta; \gamma; z), 537$
 $\theta_A(\tau), 538$
 $r_4(m), 541$
 $\sigma(m), \sigma'(m), 542$
 $r_2(m), 543$
 $d_1(m), d_3(m), r_s(m), p(n), 544$
 $(a)_0, (a)_n, (a)_\infty, 546$
 $\eta(\tau), 549$
 $\mathcal{C}, \bar{\mathcal{C}}, 550$
 $\mathcal{W} = \mathcal{W}(a_1, \dots, a_6), 552, 558$
 $\mathcal{C}_{a, b}, 553, 558$
 $\theta, d, 553$
 $P_1 + P_2, -P, 556, 583$
 $E = E(\mathbb{Q}), h(P), 559$
 $\hat{h}(P), 560, 561$
 $(P, Q), 563$
 $C_{A, B}, D, E, N, 565$
 $E, E^t, E^f, 569$
 $\Delta, b_2, b_4, b_6, b_8, 570$
 $\mathcal{W}_p, N_p, c_p, L(s) = L(s, \mathcal{W}), 571$
 $c_n, N = N(\mathcal{W}), f_p, A(s), 572$
 $r, E(\mathcal{W}, \mathbb{Q}), 572$
 $\Gamma_0(N), 573$
 $C_n, 577$
 $A_+(n), A_-(n), 578$
 $\mathcal{E}_{A, B}, 579$
 $\mathcal{W}_{A, B}, 580$
 $[\frac{n}{m}]_q, 581$

The *Landau order symbols* are defined in the following way: if $I = [t_0, \infty)$ is a half-line and if $f, g: I \rightarrow \mathbb{R}$ are real-valued functions with $g(t) > 0$ for all $t \in I$, we write

$f = O(g)$ if there exists a constant $C > 0$ such that $|f(t)|/g(t) \leq C$ for all $t \in I$;

$f = o(g)$ if $f(t)/g(t) \rightarrow 0$ as $t \rightarrow \infty$;

$f \sim g$ if $f(t)/g(t) \rightarrow 1$ as $t \rightarrow \infty$.

The *end of a proof* is denoted by \square .

Axioms

(A1)–(A3), 7	(N1)–(N3), 5
(A4),(A5), 12	(O1)–(O4), 8
(AM1), 7	(O4)', 14
(AM2), 13	(P1)–(P3), 13
(B1), (B2), 465	(P4), 19
(C1)–(C4), 106	(Pr1)–(Pr3), 465
(D1)–(D3), 27	(V1)–(V3), 261
(M1)–(M4), 7, 83	(V3)', 261
(M5), 16	

Index

- abelian
 - group, 55, 79, 172, 196, 400, 434, 441, 555, 556
 - Lie algebra, 441
- absolute value, 27, 41, 211, 261
- addition, 60
 - of integers, 11
 - of natural numbers, 6–7
 - of points of elliptic curve, 555–556, 582
 - of rational numbers, 16
- addition theorem for
 - elliptic functions, 511, 527, 555
 - exponential function, 38, 45, 46, 77
 - theta functions, 523
 - trigonometric functions, 46, 528
- adeles, 443, 444
- affine
 - conic, 549
 - cubic, 549
 - line, 549
 - plane curve, 549–552
- AGM algorithm, 502–509, 512, 535, 536
- algebra, texts, 78
- algebraic, 214
 - addition theorem, 511, 537
 - function field, 386–388, 395
 - integer, 123, 141, 151
 - number, 174, 214
 - number field, 151, 174, 384–385
- algebraic number theory, texts, 174
- algebraic topology, 76, 78, 388
- algebraically closed field, 77
- almost
 - all, 466
 - everywhere, 466
 - periodic function, 75, 79, 489
- alternating group, 57, 228, 251, 423–425
- analysis
 - complex, 48, 77
 - quaternionic, 77
 - real, 26, 76
- analytic continuation, 48, 381, 390, 405, 515, 520, 527
- angle, 47, 209
- anisotropic subspace, 295, 297, 299, 302
- approximation theorem for valuations, 267
- arc length
 - in hyperbolic geometry, 208
 - of ellipse, 493, 494
 - of lemniscate, 495
- archimedean absolute value, 262, 264, 284
- Archimedean property, 22, 26, 76
- argument of complex number, 47
- arithmetic
 - of elliptic curves, texts, 582
 - of quaternions, 120–121, 126, 541
 - progression, 312–313, 399, 443, 485, 490, 492

- arithmetic-geometric mean, 502–509, 512, 535–536
- arithmetical function, 152–154, 175
- arithmetical functions, 153, 175
- Artin's
 - primitive root conjecture, 124–125, 385, 388
 - reciprocity law, 174
- associative, 4, 60
 - algebra, 68–69, 78, 79
 - law, 7, 55, 83
- asymptote, 550
- automorphism group of
 - code, 255
 - Hadamard matrix, 252
 - t -design, 250
- automorphism of
 - \mathbb{H} , \mathbb{O} , 77
 - group, 59
 - Hadamard matrix, 252
 - quadratic field, 140
 - ring, 64
 - torus, 473
- badly approximable number, 190, 194, 217, 463, 489
- Baire's category theorem, 487
- Baker's theorem, 583
- baker's transformation, 482
- balanced incomplete block design, 247
- Banach algebra, 434
- basis of
 - lattice, 332, 333
 - module, 164, 169, 173
 - vector space, 66
- Bateman–Horn conjecture, 393–395
- Beal's conjecture, 581
- Bernoulli number, 151
- Bernoulli shift, one-sided, 479, 483
- Bernoulli shift, two-sided, 477, 483
- Bernstein's theorem, 394
- Bertrand's postulate, 366
- Bessel's inequality, 74
- best approximation
 - in inner product space, 73
 - of real number, 186, 217
- beta integral, 380
- Bézout
 - domain, 95, 101, 124, 168
 - identity, 91, 92, 109, 118, 121, 167
- Bieberbach's theorems, 346, 358
- bijection, 4
- bijective map, 4, 10
- binary
 - digit ('bit'), 255
 - linear code, 255, 256
 - operation, 55, 60
 - quadratic form, 205–207, 218
 - relation, 3
- binomial
 - coefficient, 92–94, 111
 - theorem, 110, 116, 537
- birational
 - equivalence, 216, 558, 575, 582
 - transformation, 216, 558
- Birkhoff's
 - ergodic theorem, 464–471, 479, 489
 - recurrence theorem, 485
- Blaschke's selection principle, 353, 356, 357, 360
- Blichfeldt's theorem, 334
- block, 247, 250
- Bohl's theorem, 448, 451, 458
- Bolzano–Weierstrass theorem, 25, 42
- Boolean algebra, 75–76
- Boolean ring, 61
- Borel subset, 472, 473, 479, 481
- bounded
 - sequence, 24, 26
 - set, 30
 - variation, 461–462, 489
- bracket product, 440
- Brahmagupta's identity, 195–196, 304
- Brauer group, 324
- Brauer's theorem, 425
- Brouwer's fixed point theorem, 76
- Bruck–Ryser–Chowla theorem, 249, 321, 325
- Brun's theorem, 392, 395
- BSD* conjecture, 570, 572–573, 578, 583
- Burnside's theorem, 427–428
- calendars, 119, 187

- cancellation law, 7, 76, 83, 147
- canonical height, 561–565
- Cantor's construction of reals, 18, 26, 269
- Carathéodory extension of measure, 477
- Cardano's formula, 39, 76
- cardinality, 10
- Carmichael number, 116, 125
- Cartesian product, 3
- cascade, 465
- Casimir operator, 443
- Catalan number, 93
- Cauchy sequence, 25, 30
- Cauchy's theorem, 48, 534
- Cauchy–Schwarz inequality, 28, 72
- Cayley number, 53
- Cayley–Hamilton theorem, 50
- central limit theorem, 76, 391, 395, 489
- centralizer, 60
- centre, 49, 53, 68–70
- chain condition, 90, 95, 96, 98, 101, 148
- chain rule, 33
- character of
 - abelian group, 400, 401, 434
 - representation, 414, 437
- character theory, texts, 443
- characteristic
 - function, 327
 - of ring, 62–63, 103, 111
 - polynomial, 283, 419
- Chebyshev's functions, 367–370, 372, 400
- Chevalley–Warning theorem, 116, 125
- Chinese remainder theorem, 119, 125, 267, 316
- chord and tangent process, 555
- circle method, 123, 392, 549
- class
 - field theory, 125, 325
 - function, 414, 417, 437
 - number, 151, 207, 218, 578, 584
- classical mechanics, 429, 465, 481, 490
- classification of
 - finite simple groups, 124, 251, 258
 - simple Lie algebras, 442, 444
- Clebsch–Gordan formula, 438
- Clifford algebra, 78, 324
- closed
 - ball, 35, 344, 348
 - sets, 28
- closure, 28, 204, 337, 485
- codes, 255, 388
- codeword, 256
- coding theory, texts, 258, 395
- coefficients, 66, 97
- combinatorial line, 488
- common divisor, 83
- common multiple, 84
- commutative
 - group, 55, 59, 79, 109
 - law, 7, 55, 83
 - ring, 60–61, 75, 78
- compact
 - abelian group, 434, 483
 - group, 436, 437, 442
 - metric space, 485
 - set, 28, 43, 286–287, 432
- complement of set, 3, 61
- complete
 - elliptic integral, 496, 503–509, 517, 537
 - metric space, 30–32, 75, 270
 - ordered field, 23, 26
 - quotient, 181–183, 212, 479
- completed zeta function, 381
- completion of
 - metric space, 31
 - valued field, 271
- complex
 - analysis, 48, 77
 - conjugate, 41
 - integration, 77
 - multiplication, 537, 574, 578
 - number, 39–48, 69
- composite mapping, 4
- composite number, 88, 124
- composition of solutions, 196, 198, 199
- conductor of elliptic curve, 572
- conformal equivalence, 218
- congruence
 - of integers, 106–119

- of symmetric matrices, 293
- subgroup, 210
- congruent, 106
- congruent numbers, 575–578, 584
- conic, 549, 550
- conjugacy class, 60, 125, 417
- conjugate
 - character, 421
 - complex number, 41
 - element of quadratic field, 140
 - group elements, 60, 417
 - ideal, 146
 - octonion, 54
 - quadratic irrational, 192, 209–210
 - quaternion, 49
 - representation, 421
- connected set, 28, 44
- constant, 386
 - coefficient, 97
 - sequence, 24, 269
- contains, 2
- continued fraction
 - algorithm, 179, 181, 211–212, 217
 - expansion of Laurent series, 212, 219
 - expansion of real number, 179–182
 - map, 479
- continued fractions in higher dimensions, 217
- continued fractions, texts, 217
- continuous function (map), 26, 28, 29, 43, 65
- continuously differentiable, 34, 36, 65, 67
- contraction principle, 32–33, 35, 36, 76, 285
- contragredient representation, 411
- convergence
 - in measure, 29
 - of compact sets, 353
 - of lattices, 352–357
- convergent of Laurent series, 212
- convergent of real number, 181, 182, 185–188, 480
- convergent sequence, 24–26, 30, 32, 33, 269
- convex set, 327, 341
- convolution product, 152, 433
- Conway's groups, 351
- coordinate, 3, 41, 47
- coprime, 85
- coset, 58, 63–64, 107
 - representative, 58, 78
 - right, left, 58
- countably infinite, 10
- covering, 337, 359
- critical
 - determinant, 339, 348, 357
 - lattice, 339, 357
- cross-polytope, 339
- cross-ratio, 500
- crystal, 346, 358
- crystallographic group, 346–347, 358
- crystallography, 358
- cube, 339
- cubic
 - curve, 549–558
 - polynomial, 39, 76
- cusp, 554, 571, 572
 - form, 573
- cut, 18–21
- cyclic group, 59, 62, 110, 114–116, 198, 251, 263, 423
- cyclotomic
 - field, 151
 - polynomial, 102, 111, 112, 135, 426
- cylinder set
 - general, 478
 - special, 476
- decimal expansion, 18, 107, 476, 509
- decomposable lattices, 348, 349
- Dedekind
 - construction of reals, 18–22, 76
 - eta function, 549
 - zeta function, 384–385, 395, 409, 573
- Dedekind–Peano axioms, 5, 76
- deduced representation, 420
- degree of
 - affine curve, 549
 - algebraic number, 214
 - extension field, 67

- polynomial, 96–97
- representation, 410, 414
- De Morgan's laws, 3
- dense, 17–18, 275, 447
 - sequence, 448, 452
 - subset of metric space, 31, 270
- densest
 - lattice, 348, 350–351, 359, 362
 - packing, 359–360
- density of lattice packing, 348
- derivative, 33–36, 67, 103
- designs, 247–251, 254–258
- determinant, 53, 180, 181, 224–229, 256
 - of lattice, 333, 347, 356
 - of quadratic space, 293
- diagonal matrix, 225
- diagonalization of quadratic form, 237–239, 294, 297
- difference of sets, 3
- differentiable map, 33–34, 43, 75
- differential form, 256
- dimension of vector space, 67
- Diophantine
 - approximation, 185, 212, 217, 219, 329, 358
 - equation, 161, 178, 195–201, 215–217, 219
- direct product of groups, 53, 60, 117, 172, 401, 429, 443
- direct sum of
 - rings, 64, 117
 - vector spaces, 65
- Dirichlet
 - L -function, 404–409, 443
 - character, 403–404, 574
 - domain, 342
 - product, 152–156, 175, 389
 - series, 175, 389, 572
- Dirichlet's
 - class number formula, 218
 - convergence criterion, 138
- Dirichlet's theorem on
 - Diophantine approximation, 329
 - primes in arithmetic progression, 218, 313, 316, 400, 443
- units in number fields, 176
- discrepancy, 459–464, 488
- discrete
 - absolute value, 275, 276
 - group, 203, 330
 - set, 342
 - subgroup, 330, 490
- discriminant of
 - binary quadratic form, 205–207
 - elliptic curve, 553, 570, 579, 580
 - lattice, 333
 - quadratic irrational, 191
- disjoint sets, 2
- distance, 18, 27–31
- distributive
 - lattice, 85
 - law, 7, 60, 85
- divisibility tests, 107
- divisible, 83, 87, 146
- division
 - algebra, 54, 78
 - algorithm, 14, 90, 98
 - ring, 62–70, 125
- divisor, 83, 386
 - of zero, 14, 62
- doubly-periodic function, 514–517, 527, 538
- dual
 - 2-design, 249
 - convex body, 340
 - group, 401, 434, 443
 - lattice, 334, 340, 538
- duality theorem, 435
- dynamical system, 30, 388, 395, 447
- Dynkin diagrams, 359
- e , 38, 187
- echelon form, 164
- eigenvalue, 239, 257, 430
- eigenvector, 239
- Eisenstein
 - integers, 141–143
 - irreducibility criterion, 102, 124
- element, 2
- elementary group, 425
- ellipse, 237, 493–494
- ellipsoid, surface area of, 495–496

elliptic

curves, 555–558, 569–574, 582–583

functions, 509–516, 525–530

elliptic integral, 496–503, 536–537

of first kind, 498, 505, 514–516, 537

of second kind, 498, 505, 516–517, 530, 537

of third kind, 498, 506, 517, 530, 537

empty set, 2, 61

endomorphism of torus, 473, 483

energy surface, 465, 481

entropy, 482–483

equal, 2

equivalence class, 4

equivalence of

absolute values, 266

complex numbers, 184, 191, 201, 210

fundamental sequences, 26, 31

Hadamard matrices, 252

ideals, 151

matrices, 171

quadratic forms, 293, 297–298, 311, 317, 325

representation, 410, 416

equivalence relation, 4, 11, 26, 58, 106, 184

Erdős–Turan inequality, 462, 489

ergodic, 465

hypothesis, 464–465, 470

measure-preserving transformation, 470, 472–473, 478–479

theorems, 466, 489

ergodic theory, texts, 489

error-correcting code, 28, 256, 258

Euclidean

algorithm, 91–92, 98, 104, 109, 121, 181, 560

distance, 28, 72, 346

domain, 104–106, 119–120, 124, 142, 171

metric, 342

norm, 229, 234, 342

prime number theorem, 134, 363

Euler's

angles, 439

conjecture, 580–581, 584

constant, 159, 380

criterion for quadratic residues, 112, 135, 136, 159

formulas for $\cos z$ and $\sin z$, 45, 77

parametrization of rotations, 51–53

pentagonal number theorem, 545, 581

phi-function, 109–110, 114, 154–157, 389, 399, 403

prime number theorem, 363–364

product formula, 371, 381, 405

theorem on homogeneous functions, 551

even

lattice, 349, 351

permutation, 57, 130, 224, 227

eventually periodic, 18, 277

continued fraction, 192

exceptional simple Lie algebra, 442, 444

existence theorem for ordinary differential equations, 36–38, 76, 510

exponential

function, 38–39, 45–47

series, 38, 45

sums, 388, 489

extended Riemann hypothesis, 385, 395, 409

extension of absolute value, 273, 283–284, 290

extension of field, 45, 67, 283–284, 288, 386

exterior algebra, 256

extreme point, 236

face, 343

facet, 343–345

vector, 344–346

factor, 83, 146

group, 58–59

factorial domain, 90, 101, 124, 151, 153, 175

factorization, 124, 127

Faltings' theorem, 215, 575, 583

Fano plane, 248, 250

Fermat equation, 142, 151, 575, 576, 579

- Fermat number, 160–161, 175
- Fermat prime, 160–161
- Fermat's
 - last theorem, 151–152, 175, 579–580, 584
 - little theorem, 110–112, 116
- Fibonacci numbers, 95
- field, 23, 40, 62–63, 109, 125
 - of fractions, 88, 100, 261
- field theory, texts, 79
- finite
 - dimensional, 66, 67
 - field, 109, 125, 232, 289, 298, 386–387, 570–571
 - field extension, 45, 67, 283–284, 288, 386
 - group, 57
 - set, 10
- finitely generated, 63, 66, 95, 168, 172, 176, 559
- Fischer's inequality, 239, 243
- fixed point, 32, 35, 37
 - theorems, 32, 76
- flex, 551–553, 557
- flow, 465, 481, 483
- formal
 - derivative, 103, 112, 278
 - Laurent series, 211, 219, 263, 271, 289, 358
 - power series, 96, 124, 275, 287
- Fourier
 - integrals, 435–436, 443
 - inversion formula, 435–436
 - series, 79, 138, 378, 436, 443, 450
 - transform, 374, 378, 435–436
- fraction, 15
- fractional part, 448, 479
- free
 - action, 218
 - product, 203
 - subgroup, 172, 569
 - submodule, 172
- Fresnel integral, 139
- Frobenius
 - complement, 429
 - conjectures, 124, 211
 - group, 429, 443
 - kernel, 429
 - reciprocity theorem, 420
 - theorem on division rings, 69
- Fuchsian group, 218
- function, 4; 386
- function fields, 386–388, 395
 - and coding theory, 388, 395
- functional equation of
 - L -functions, 409, 443, 572
 - zeta functions, 380–381, 385, 387, 443
- fundamental
 - domain, 203, 333, 346, 499
 - sequence, 25, 26, 30–33, 269, 270, 274
 - solution, 197–201
- fundamental theorem of
 - algebra, 42–45, 69, 77, 98, 413
 - arithmetic, 88–90, 124, 145
- Furstenberg's theorems, 484–487, 490
- Furstenberg–Katznelson theorem, 485
- Furstenberg–Weiss theorem, 485
- Galois theory, 79, 160
- gamma function, 328, 379–380, 394, 572
- Gauss
 - class number problem, 218, 578, 584
 - invariant measure, 479
 - map, 479–480, 483, 489
 - sum, 135–140, 174
- Gauss–Kuz'min theorem, 480, 489
- Gaussian integer, 119–120, 141, 145, 542
- Gaussian unitary ensemble, 384
- GCD domain, 87–90, 98, 100–101
- gear ratios, 187
- Gelfand–Raikov theorem, 434
- general linear group, 68, 251, 440
- generalized
 - character, 428
 - trigonometric polynomial, 75, 458
 - upper half-plane, 218–219
- generated by, 59, 60, 63, 66, 91, 114, 161, 331, 465
- generating function, 544, 581

- generator
 - matrix, 255
 - of cyclic group, 114
- genus of
 - algebraic curve, 216, 395, 558
 - field of algebraic functions, 387, 388
- geodesic, 208–210, 219, 388, 395, 482
 - flow, 388, 395, 481, 483
- geometric
 - representation of complex numbers, 41, 47
 - series, 34
- geometry of numbers, texts, 357
- Golay code, 255, 256
- golden ratio, 179
- good lattice point, 464
- Good–Churchhouse conjectures, 391–392, 395
- graph, 30
- Grassmann algebra, 256
- greatest
 - common divisors, 83–87, 89–92, 148
 - common left divisor, 167
 - common right divisor, 121
 - lower bound, 19, 22, 24
- group, 55–60, 109
 - generated by reflections, 359, 442, 444
 - law on cubic curve, 555–556, 565, 582, 583
- group theory, texts, 78
- Haar
 - integral, 432–433, 443
 - measure, 358, 435, 474
- Hadamard
 - design, 250–251
 - determinant problem, 223, 243–247, 250, 257
 - inequality, 223, 229–230
 - matrix, 223, 230–233, 250–257, 306, 321, 350
- Hales–Jewett theorem, 488, 490
- Hall’s theorem on solvable groups, 428, 443
- Hamiltonian system, 481, 483
- Hamming distance, 28, 255
- Hardy–Ramanujan expansion, 549, 582
- Hasse
 - invariant, 310, 311, 324
 - principle, strong and weak, 317, 325
- Hasse–Minkowski theorem, 312–316, 324–325
- Hasse–Weil (*HW*) conjecture, 572
- Hausdorff
 - distance, 352–353
 - maximality theorem, 485
 - metric, 353, 356, 357
- heat conduction equation, 519
- height of a point, 559–561
- Hensel’s lemma, 277–282, 290
- Hermite
 - constant, 347–348, 351, 359
 - normal form, 175, 332
- highest coefficient, 97
- Hilbert field, 306–311, 324
- Hilbert space, 75, 79, 433
- Hilbert symbol, 303–307, 313, 324
- Hilbert’s problems
 - 5th, 440, 443
 - 9th, 174
 - 10th, 217
 - 17th, 323–324, 325
 - 18th, 358, 359–360
- H*-matrix, 230–231, 244, 245
- holomorphic function, 48, 124, 372, 377, 404, 405, 512
- homogeneous linear equations, 68, 166
- homomorphism of
 - groups, 52–53, 58–59, 566
 - Lie algebras, 441
 - Lie groups, 441
 - rings, 63–64, 99, 111
 - vector spaces, 67
- Horner’s rule, 99
- Hurwitz integer, 120–121, 126, 541
- hyperbolic
 - area, 209
 - geometry, 208–209, 219
 - length, 208
 - plane, 298, 299, 302
- hypercomplex number, 77

- hypergeometric function, 537
- hyperreal number, 76
- ideal, 63, 90–91, 145, 148
 - class group, 151
 - in quadratic field, 146–151, 207
 - of Lie algebra, 441
- identity
 - element, 12, 20, 55–56, 60, 65, 83
 - map, 4
- Ikehara's theorem, 367, 373, 385, 389, 390, 408
- image, 4
- imaginary
 - part, 40
 - quadratic field, 140
- incidence matrix, 247–249
- included, 2
- indecomposable lattice, 348–349
- indefinite quadratic form, 205
- indeterminate, 96
- index of
 - quadratic space, 296, 297
 - subgroup, 58, 60, 419, 423–424, 568
- indicator function, 327, 358, 449, 470
- individual ergodic theorem, 466
- induced representation, 419–423, 425
- induction, 9
- infimum, 19, 22
- infinite order, 59
- inflection point, 551
- inhomogeneous Lorentz group, 442
- injection, 4
- injective map, 4, 9, 68
- inner product space, 71–75, 79, 433
- integer, 10–15, 17
 - of quadratic field, 106, 141
- integrable, 466
 - in sense of Lebesgue, 32, 75, 327, 435
 - in sense of Riemann, 327, 358, 449, 450
- integral
 - divisor, 386
 - domain, 62, 87, 96
 - equations, 74, 79, 223
 - lattice, 349, 538
 - representation for Γ -function, 380, 573
- interior, 28, 333, 337, 343
- intersection
 - of modules, 166
 - of sets, 2–3, 61
 - of subspaces, 65
- interval, 26, 29, 65
- invariant
 - factor, 171
 - mean, 437
 - region, 481
 - subgroup, 58
 - subset, 485
 - subspace, 411
- inverse, 12, 16, 55, 62, 153
 - class, 417
 - element, 55–56, 62, 556, 583
 - function theorem, 34–36, 76
 - map, 4–5
- inversion
 - of elliptic integral, 514, 516
 - of order, 57, 129
- invertible
 - element of ring, 62
 - matrix of integers, 162
 - measure-preserving transformation, 466, 479
- involutory automorphism, 41, 140
- irrational number, 22, 179, 448, 451
- irrationality of $\sqrt{2}$, 18, 99
- irreducible
 - character, 414, 416–418, 437
 - curve, 551, 552–553
 - element, 89–90, 145
 - ideal, 148
 - polynomial, 98, 102, 111–112
 - representation, 411–418, 434, 437
- irredundant representation, 344
- isometric
 - metric spaces, 31
 - quadratic spaces, 299
- isometry, 31, 75, 208, 299–303, 346, 430
- isomorphism, 5, 14, 17, 22, 23
 - of groups, 59

- of measure-preserving transformations, 482–483
 - of rings, 64
 - of vector spaces, 68
- isotropic
 - subspace, 295
 - vector, 295
- Jacobi symbol, 130–134, 139, 174
- Jacobi's
 - imaginary transformation, 377, 520
 - triple product formula, 518, 536, 545, 548, 582
- Jacobian elliptic functions, 525–530
- join, 2
- Jordan–Hölder theorem, 124
- Kepler conjecture, 359–360
- kernel of
 - group homomorphism, 53, 59
 - linear map, 68
 - representation, 426
 - ring homomorphism, 63
- Kervaire–Milnor theorem, 78
- Kingman's ergodic theorem, 489
- K -point
 - affine, 549
 - projective, 550, 555
- kissing number, 351, 359, 362
- Kronecker
 - approximation theorem, 452, 488
 - delta, 415
 - field extension theorem, 45
 - product, 68, 231, 233, 250, 255, 411
- Lagrange's theorem
 - on four squares, 120–122, 218, 541
 - on order of subgroup, 58, 110, 114
- Landau order symbols, 194, 365, 590
- Landau's theorem, 405–406
- Landen's transformation, 529, 536, 566
- Langlands program, 174, 178, 582
- Laplace transform, 373, 394, 405
- lattice, 85, 123; 141, 332–334, 348
 - in locally compact group, 490
 - packing, 348, 359
 - packing of balls, 348, 350, 359
 - point, 328, 332
 - translates, 337–338
- Laurent
 - polynomial, 98, 216
 - series, 48, 211, 263, 287, 289, 358, 373
- law of
 - iterated logarithm, 391, 395, 489
 - Pythagoras, 17, 73–74, 108
 - quadratic reciprocity, 129, 133–136, 151, 174, 314
 - trichotomy, 8, 13, 21
- least
 - common multiple, 84–87, 89
 - common right multiple, 167
 - element, 8
 - non-negative residue, 107, 115
 - upper bound, 19, 22, 24
- least upper bound property (**P4**), 19, 22, 26
- Lebesgue measurable, 32, 75
- Lebesgue measure, 32, 327, 465, 472, 473, 480
- Leech lattice, 350–351, 359, 362
- Lefschetz fixed point theorem, 76
- left
 - Bézout identity, 121
 - coprime matrices, 167–168
 - coset, 58
 - divisor, 167
- Legendre
 - interchange property, 517
 - normal form, 498
 - polynomials, 74
 - relation, 508
 - symbol, 129, 133, 135, 149, 232, 305
 - theorem on ternary quadratic forms, 312
- lemniscate, 495, 509, 536, 537
- less than, 8
- L -function, 404, 443, 571–574, 578
- Lie
 - algebra, 440–442, 443–444
 - group, 439–442, 443–444
 - subalgebra, 440

- subgroup, 441
- limit, 24, 30, 269
- linear
 - code, 255, 388
 - combination, 65
 - differential system, 172–173
 - Diophantine equation, 91, 161, 165–166
 - fractional transformation, 179, 208, 500, 531, 535
 - map, 67
 - systems theory, 176
 - transformation, 67
- linear algebra, texts, 79
- linearly dependent, 66
- linearly independent, 66
- Linnik's theorem, 409
- Liouville's
 - integration theory, 537
 - theorem in complex analysis, 77, 517
 - theorem in mechanics, 481
- Lipschitz condition, 460
- Littlewood's theorem, 383, 395
- LLL-algorithm, 358
- L^2 -norm, 72
- local-global principle, 317–318, 325
- locally compact, 28
 - group, 358, 432–436, 444, 490
 - topological space, 432
 - valued field, 284–290
- locally Euclidean topological space, 440
- logarithm, 39, 364
- lower
 - bound, 14, 19, 22
 - limit, 24
 - triangular matrix, 229
- Lucas–Lehmer test, 158, 175
- Mahler's compactness theorem, 357, 360, 362
- map, 4
- mapping, 4
- Markov
 - spectrum, 210, 219
 - triple, 210–211, 222
- marriage theorem, 78
- Maschke's theorem, 411, 431
- Mathieu groups, 251, 254, 255
- 'matrix', 162, 168
- matrix theory, texts, 79, 257
- maximal ideal, 64, 148, 269, 274
- maximal totally isotropic subspaces, 296
- Mazur's theorem, 570
- mean motion, 458, 489
- measurable function, 29, 466
- measure theory, texts, 489
- measure zero, 29, 32, 482
- measure-preserving transformation, 466–473, 477–484
- meet, 2
- Mellin transform, 573
- Méray–Cantor construction of reals, 18, 26
- Merkur'ev's theorem, 324
- meromorphic function, 48, 263, 512, 538
- Mersenne prime, 158–159, 175
- Mertens' theorem, 364
- method of successive approximations, 32, 36, 38
- metric space, 27–32, 72, 255, 268
- Meyer's theorem, 313, 316
- minimal
 - basis, 173
 - model, 571, 580, 583
 - polynomial, 283
 - vector, 345, 346
- minimum of a lattice, 345–347, 356, 357
- Minkowski's theorem on
 - discriminant, 330, 358
 - lattice points, 328–330, 338–339, 358
 - linear forms, 328
 - successive minima, 339–341, 358
- minor, 171
- mixing transformation, 480, 483
- Möbius
 - function, 156, 390–392, 395
 - inversion formula, 156–157, 175
- modular
 - elliptic curve, 573–574, 578, 584, 586
 - form, 258, 544, 573–574, 578, 583

- function, 531–536, 538
- group, 202–205, 531
- transformation, 201
- module, 161, 166–169, 171–172, 175
- modulo m , 106
- monic polynomial, 97, 151, 174, 262
- monotonic sequence, 24–26
- Monster sporadic group, 258
- Montgomery's conjecture, 383–384, 395
- Mordell conjecture, 215, 216, 575, 583
- Mordell's theorem, 176, 559, 565–569
- multiple, 83
- multiplication, 60
 - by a scalar, 64
 - of integers, 12
 - of natural numbers, 7
 - of rational numbers, 16
- multiplicative
 - function, 154–155, 175
 - group, 62, 114–115, 125, 292
 - inverse, 16
- Nagell–Lutz theorem, 569, 578
- natural
 - logarithm, 39, 364
 - number, 5–10, 14
- nearest neighbour conjecture, 384
- negative
 - definite quadratic space, 296
 - index, 297
 - integer, 13
- neighbourhood, 33
- Nevanlinna theory, 215, 219
- Newton's method, 277, 290
- node, 554, 571, 572
- non-archimedean absolute value, 261, 264, 273–276
- non-associative, 53–54, 78
- non-Euclidean
 - geometry, 208–209, 219
 - line, 208
 - triangle, 209, 533
- non-negative linear functional, 432
- nondecreasing sequence, 24, 25
- nondegenerate lattice, 332
- nonincreasing sequence, 24, 25
- non-singular
 - cubic curve, 555
 - linear transformation, 68
 - matrix, 227
 - point, 549, 550
 - projective curve, 387
 - projective variety, 387
 - quadratic subspace, 293
- norm of
 - n -tuple, 28
 - complex number, 119
 - continuous function, 28, 29
 - element of quadratic field, 106, 140–141
 - ideal, 384
 - integral divisor, 386
 - linear map, 34
 - octonion, 54
 - prime divisor, 386
 - quaternion, 50–51, 121
 - vector, 71, 271, 341
- norm-Euclidean domain, 106
- normal
 - form for cubic curve, 552–553, 556–558
 - frequencies, 430
 - modes of oscillation, 430
 - number, 474, 476, 489
 - subgroup, 58–59, 79, 421, 427
 - vector, 474–476, 489
- normed vector space, 271, 287, 341
- n -th root of
 - complex number, 44, 47, 77
 - positive real number, 23
- n -tuple, 3, 28, 64
- nullity of linear map, 68
- nullspace of linear map, 68
- number theory, texts, 123
- numbers, 1, 74
- numerical integration, 460–464, 489
- octave, 53
- octonion, 53–55, 77, 82, 442
- odd permutation, 57, 129, 130, 224, 227
- one (1), 5, 60
- one-to-one, 4

- correspondence, 4
- open
 - ball, 28, 33, 342
 - set, 28, 43
- operations research, 78
- Oppenheim's conjecture, 324, 325
- order in natural numbers, 8
- order of
 - element, 59, 113, 114
 - group, 57, 109, 113, 114
 - Hadamard matrix, 230
 - pole, 48
 - projective plane, 248
- ordered field, 23, 26, 41, 76, 79, 280, 296–297, 308
- ordinary differential equation, 36–38, 76, 510
- orientation, 225, 347
- Ornstein's theorem, 483
- orthogonal
 - basis, 74, 294, 335
 - complement, 293
 - group, 440
 - matrix, 52, 238
 - set, 73
 - sum, 293, 349
 - vectors, 73, 293
- orthogonality relations, 402, 415–416, 418, 437
- orthonormal set, 73–75
- Oseledets ergodic theorem, 489
- Ostrowski's theorems, 266, 284, 290, 313
- packing, 337, 359–360
- p -adic
 - absolute value, 262
 - integer, 275, 277, 287
 - number, 18, 271, 275, 277, 287, 288, 310, 313, 436
- pair correlation conjecture, 383–384, 395
- Paley's construction, 231–233, 255, 350
- parallelogram law, 73, 79, 561, 564
- parallelotope, 229, 333
- parametrization, 51–52, 217, 219, 554–555, 558, 574
- Parseval's equality, 74–75, 335–336
- partial
 - fractions, 497
 - order, 85
 - quotient, 181, 190, 212, 480
- partition of
 - positive integer, 544–549
 - set, 4, 58
- partition theory, texts, 581
- Pascal triangle, 94
- path-connected, 44, 437, 441
- Peano axioms, 5, 76
- Pell equation, 144, 196–201, 217
 - for polynomials, 213, 219
- pendulum, period of, 494
- Pépin test, 160–161
- percolation processes, 489
- perfect number, 157–159, 175
- period of continued fraction, 192–194, 197–199, 217
- periodicity of
 - continued fraction, 192–194, 209, 217
 - elliptic functions, 513–516, 527, 538
 - exponential function, 46–47
- permutation, 57, 129–131, 227
- perpendicular, 73
- Perron–Frobenius theorem, 480
- Pfister's multiplicative forms, 323
- π (π), 46, 48, 186, 217, 222, 364, 509
- Picard's theorem, 538
- pigeonhole principle, 10, 57
- Plancherel theorem, 435
- Poincaré
 - model, 208, 219
 - recurrence theorem, 483–484
- point, 247, 250, 549, 550
 - at infinity, 550, 552, 553
- pointwise ergodic theorem, 466
- Poisson summation, 138, 378, 394, 435, 538
- polar
 - coordinates, 47, 495
 - lattice, 334
- pole of order n , 48
- poles of elliptic functions, 526–527
- polynomial, 96–103

- part, 211
- ring, 87, 104
- polytope, 343, 358
- Pontryagin–van Kampen theorem, 435
- positive
 - index, 297
 - integer, 13–14
 - measure, 433
 - rational number, 16–17
 - real number, 18, 22
 - semi-definite matrix, 235, 239
- positive definite
 - matrix, 235, 239
 - quadratic form, 205
 - quadratic space, 296
 - rational function, 323–324
- power series, 45, 46, 48
- primality testing, 124, 127
- prime
 - divisor, 386
 - element, 89, 145
 - ideal, 148–151, 384–385
 - ideal theorem, 385, 393, 395
 - number, 88–89
- prime number theorem, 365–367, 369–377, 390, 394–395
 - for arithmetic progressions, 394, 400, 403–408
- primitive
 - Dirichlet character, 409
 - polynomial, 100
 - quadratic form, 206
 - root, 115–116, 124–125, 385
 - root of unity, 111, 112, 114
- principal axes transformation, 238, 257
- principal character, 401
- principal ideal, 91, 145
 - domain, 92, 95–96, 98, 104, 105–106, 168–172
- principle of the argument, 533
- probability
 - measure, 466
 - space, 466
 - theory, 29, 75, 76, 391, 395, 582
- problem of moments, 220
- problem, $3x + 1$, 490
- product
 - formula for theta functions, 519, 521
 - formula for valuations, 267
 - measure, 476
 - of ideals, 145
 - of integers, 12
 - of linear maps, 67
 - of natural numbers, 7
 - of rational numbers, 16
 - of representations, 411
 - of sets, 3
- projective
 - completion, 550–551, 552–553
 - conic, 550
 - cubic, 550
 - equivalence, 551–553
 - line, 550
 - plane, 248, 250, 321, 325
 - plane curve, 550
 - space, 53
- proper
 - divisor, 89
 - subset, 2
- properly equivalent
 - complex numbers, 184, 201
 - quadratic forms, 205–206
- properly isomorphic, 347
- public-key cryptography, 124
- Puiseux expansion, 44
- pure
 - imaginary complex number, 41
 - quaternion, 51, 52
- Pythagoras' theorem (or law), 17, 73–74, 108
- Pythagorean triple, 108, 217
- q -binomial coefficient, 95, 581
- q -difference equation, 546
- q -hypergeometric series, 581
- q -integral, 581
- quadratic
 - field, 105–106, 124, 140–151, 174, 207, 218
 - form, 205–207, 291–293, 563
 - irrational, 191–194, 206, 209–210, 214, 537
 - nature, 129, 133

- non-residue, 112–113, 121, 129, 132–133, 385
- polynomial, 42, 283
- residue, 112–113, 121, 129, 132–133, 280
- space, 292–303, 324
- quadratic spaces, texts, 324
- quantum group, 582
- quartic polynomial, 76–77
- quasicrystal, 359
- quasiperiodic tiling, 359
- quaternion, 48–53, 69, 77–82, 120–122, 541
- quaternionic analysis, 77
- Quillen–Suslin theorem, 176
- quotient, 15, 48, 90
 - group, 58
 - ring, 63, 107, 384
 - space, 209, 210, 218
- Rådström’s cancellation law, 353, 360
- Ramanujan’s tau-function, 388, 395
- random matrices, 384, 395, 489
- range of linear map, 68
- rank of
 - elliptic curve, 569–570, 572
 - linear map, 68
- rational
 - function, 88, 212, 262, 386
 - number, 15–17, 181–182, 277
 - transformation, 558
- real
 - analysis, 26, 76
 - number, 22–26
 - part, 40
 - quadratic field, 140
- reciprocal lattice, 334
- reciprocity for Gauss sums, 137
- recurrence for number of partitions, 545
- recursion theorem, 5–6
- reduced
 - automorphism group, 252
 - lattice basis, 358
 - quadratic form, 206, 207
 - quadratic irrational, 192–194
- reducible
 - curve, 551
 - polynomial, 282
 - representation, 411
- Reed–Muller code, 255–256
- refinement theorems, 86, 123–124
- reflection, 53, 300
- reflexive relation, 4
- regular prime, 151
- regular representation, 410, 416–417, 434, 437
- relatively
 - dense set, 343
 - prime, 85, 167
- relevant vector, 344
- remainder, 90
 - theorem, 99
- replacement laws, 106
- representation of
 - compact group, 436–439
 - finite group, 410–413, 437, 443
 - group, 410, 442–444
 - locally compact group, 434–436
- representative of
 - coset, 58, 78
 - residue field, 276
- representatives, distinct, 78
- represented by quadratic form, 294, 295
- residue, 48, 371, 390
 - class, 107, 400
 - field, 274–276, 386
- resolution of singularities, 558
- restriction of map, 4
- Ribet’s theorem, 580
- Riemann
 - integrable, 327, 358, 449, 450, 452
 - normal form, 498–502, 510, 555–556
 - surface, 218
 - zeta function, 366, 370–373, 380–384, 390–392, 394
- Riemann hypothesis, 381–383, 391–392, 395, 398
 - for algebraic varieties, 388, 395
 - for elliptic curves, 571, 574, 583
 - for function fields, 387–388, 395, 583
- Riemann–Lebesgue lemma, 376

- Riemann–Roch theorem, 341, 358, 387
- Riemannian manifold, 388, 481–482, 483
- Riesz representation theorem, 433
- Riesz–Fischer theorem, 75
- right
 - coset, 58–60
 - multiple, 167
 - vector spaces, 68
- ring, 60–64, 68, 96, 107
- ring theory, texts, 78
- Rogers–Ramanujan identities, 546–549, 582
- root, 99–100, 219, 277
 - lattice, 348–351, 359
- Roth’s theorem on algebraic numbers, 123, 214–216, 219
- ruler and compass constructions, 160, 175
- scalar, 64
- schemes, 582
- Schmidt’s
 - discrepancy theorem, 463–464, 489
 - subspace theorem, 214–215, 219
- Schmidt’s orthogonalization process, 74, 230
- Schreier’s refinement theorem, 123–124
- Schur’s lemma, 413, 415, 417, 418
- Schwarz’s inequality, 28, 72, 234, 373, 453
- self-dual lattice, 334, 335, 340
- semi-simple
 - Lie algebra, 441–442, 444
 - Lie group, 441–442, 444
- semi-stable elliptic curve, 572, 574, 580
- semidirect product, 428
- semigroup, 76
- Serre’s
 - ε -conjecture, 580
 - conjecture, 175–176
- set, 2–4, 61
- set of representatives, 276, 289
- shift map, 477, 479, 487
- Siegel’s
 - formula, 335
 - lemma, 340, 358
- modular group, 218–219
- theorem on Diophantine equations, 216, 217, 219, 583
- sigma algebra, 433, 465
- sign of a permutation, 57, 130–131, 227
- signed permutation matrix, 242, 246, 252
- simple
 - associative algebra, 69
 - basis, 349–350
 - group, 58, 251, 258, 427, 428
 - Lie algebra, 441–442, 444
 - Lie group, 77, 251
 - pole, 48
 - ring, 63
- simply-connected, 53, 437
 - covering space, 53, 441–442
 - Lie group, 441–442
- simultaneous diagonalization, 239, 429
- singular matrix, 227, 228
- small
 - divisor problems, 219–220
 - oscillations, 257, 429–430
- Smith normal form, 169–173, 176
- sojourn time, 470
- solvable
 - by radicals, 39, 79
 - group, 428
 - Lie algebra, 443–444
- spanned by, 66
- special, 53
 - linear group, 202, 229
 - orthogonal group, 53, 438–439, 442
 - unitary group, 53, 437–438, 442
- spectrometry, 236, 257
- spherical trigonometry, 538
- sporadic simple group, 251, 258, 351
- square, 14
 - class, 292, 293, 565–566
 - design, 249, 258, 321
- square 2-design, 249, 258, 321
- square root of
 - complex number, 39, 42
 - positive real numbers, 22–23, 24–25
- square-free

- element, 90
 - integer, 90, 140
 - polynomial, 103, 112
- square-norm, 342, 345
- star discrepancy, 459, 464
- Steiner system, 250
- step-function, 449
- Stieltjes integral, 220, 368, 394
- Stirling's formula, 328, 380
- Stone's representation theorem, 62, 76
- Stone–Weierstrass theorem, 488
- strictly proper part, 211
- strong
 - Hasse principle, 317–318
 - triangle inequality, 30, 32, 262
- structure theorem
 - for abelian groups, 172, 569
 - for modules, 172
- subadditive ergodic theorem, 489
- subgroup, 56–57
- subset, 2, 61
- subspace, 65–68
- successive
 - approximations, 32, 36, 38
 - minima, 339–341, 358
- successor, 5
- sum of
 - linear maps, 68
 - modules, 166–167
 - natural numbers, 6–7
 - points of elliptic curve, 556, 583
 - representations, 411, 412
 - subspaces, 65
- sum of squares, 51, 55, 78, 126, 544, 581
 - four, 51, 120–122, 218, 541–542
 - three, 108, 120, 318–319
 - two, 108, 119–120, 199–201, 218, 247, 542–544
 - for polynomials, 322–323, 325
 - for rational function, 323–324, 325
- supplements to law of quadratic reciprocity, 133, 314
- supremum, 19, 22
- surface
 - area of ellipsoid, 495–496
 - of negative curvature, 388–389, 395, 482, 483
- surjection, 4
- surjective map, 4, 10, 68
- Sylvester's law of inertia, 297
- symmetric
 - difference, 61, 465
 - group, 57, 227, 423–425
 - matrix, 232, 238–239, 292–294, 301
 - relation, 4
 - Riemannian space, 219
 - set, 328, 341
- symmetric 2-design, 249
- symmetry group, 346, 430
- symmetry operation, 430
- symplectic matrix, 219, 440, 442
- systems of distinct representatives, 78
- Szemerédi's theorem, 484–485, 490
- tangent
 - space, 440, 481–482
 - to affine curve, 550
 - to projective curve, 550
- taxicab number, 117
- Taylor series, 48, 373, 406
- t -design, 250, 254–255
- tensor product, 68, 303
- theta functions, 379, 519–525, 530, 535, 541–544
 - of lattice, 538
- tiling, 204, 337, 343, 344, 346, 358, 359, 499
- topological
 - entropy, 388
 - field, 268
 - group, 432, 440, 442, 443
- topology, 28, 268
- torsion
 - group of elliptic curve, 569–570, 577
 - subgroup, 59, 172
 - submodule, 172
- torsion-free, 346, 358
- torus, n -dimensional, 439, 441, 472–473
- total
 - order, 8, 18, 24
 - variation, 461–462, 489
- totally isotropic subspace, 295–296, 299

- totient function, 110
- trace of
 - matrix, 414
 - quaternion, 49–50
- transcendental element, 386
- transcendental number, 174, 578, 583
- transformation formulas
 - for elliptic functions, 512, 528–529, 538
 - for theta functions, 377–379, 520, 522
- transitive
 - law, 8
 - relation, 4
- translation, 346
 - of torus, 472, 483
- transpose of a matrix, 227, 229
- triangle inequality, 27, 262
- triangular matrix, 229
- trichotomy law, 8
- trigonometric
 - functions, 45–47, 77
 - polynomial, 450, 488
- trivial
 - absolute value, 262, 273
 - character, 401
 - representation, 410
 - ring, 62
- TW*-conjecture, 574, 578, 580, 584, 586
- twin prime, 392–393, 395
- twisted *L*-functions, 574
- 2-design, 247–250, 258
- type (A) Hilbert field, 308–310
- type (B) Hilbert field, 308–311
- ultrametric inequality, 262, 273, 277
- uniform distribution, texts, 488
- uniformization theorem, 218
- uniformly distributed mod 1, 448–458
- union of sets, 2–3, 61
- unique factorization domain, 90
- unit, 62, 87, 108, 144–145, 153
- unit circle, 47–48
- unit tangent bundle, 481–482
- unitary
 - group, 440, 442
 - matrix, 53, 437
 - representation, 412, 434, 437
 - symplectic group, 440, 442
- universal quadratic form, 295, 297
- upper
 - bound, 19, 22
 - density, 484
 - half-plane, 201, 208–209, 519, 531, 534
 - limit, 24
 - triangular matrix, 229
- valuation
 - ideal, 274–276
 - ring, 88, 274–276
- valuation theory, texts, 290
- value, 4
 - group, 262–263, 274, 276
- valued field, 261–264
- van der Corput's
 - difference theorem, 454, 458
 - sequence, 463, 464
- van der Waerden's theorem, 485–488, 490
- vector, 64
 - space, 64–70
- vertex of polytope, 343
- volume, 327
- von Mangoldt function, 370, 372
- Voronoi cell, 342–346, 349, 359
 - of lattice, 344–346, 353–357, 359
- Voronoi diagram, 358
- Waring's problem, 122–123, 126
- weak Hasse principle, 317
- Wedderburn's theorem on
 - finite division rings, 125
 - simple algebras, 69
- Weierstrass approximation theorem, 74, 450, 488
- weighing, 233–236, 257
 - matrix, 234–236, 257
- weight of vector, 254–256
- Weil conjectures, 387–388, 395
- Weyl's criterion, 451
- Wiener's Tauberian theorem, 367, 394
- Wiles theorem, 575, 580, 584
- Williamson type, 233

Wilson’s theorem, 112, 122

Witt

 cancellation theorem, 302, 303, 317

 chain equivalence theorem, 310

 equivalence, 303

 extension theorem, 302

 ring, 303, 323, 324

zero, 12, 60

zeros of elliptic functions, 526–527

zeta function, 366, 370–373, 380–384,
 394, 398

 generalizations, 389, 395

 of function field, 386–387

 of number field, 384–385, 395, 409

CONTROL THEORY (TSRT09, TSRT06)

Exercises & solutions

9 september 2025

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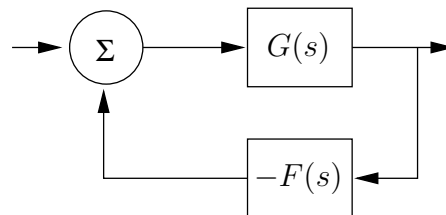
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1 Introduction

1.1

Consider the linear feedback control system given by the figure below.



Show that if the small gain theorem (Swe: lågförstärkningssatsen) is fulfilled the Nyquist criterion is also fulfilled.

Solution

1.2

Consider a static nonlinear system described by an ideal relay given by the function

$$y(t) = f(u(t)) = \begin{cases} 1, & u > 0 \\ 0, & u = 0 \\ -1, & u < 0 \end{cases}.$$

What is the gain of the relay?

Solution

1.3

Consider the system initially modelled as

$$Y(s) = G(s)U(s) \quad G(s) = \frac{1}{(s+1)^2}$$

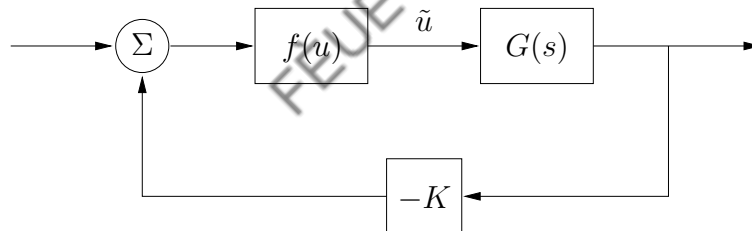
The control signal however goes through a valve with saturation

$$\tilde{u}(t) = \begin{cases} 1, & \text{if } u(t) > 2 \\ \frac{1}{2}u(t), & \text{if } |u(t)| \leq 2 \\ -1, & \text{if } u(t) < -2 \end{cases}$$

The real system is thus

$$Y(s) = G(s)\tilde{U}(s).$$

The system is controlled using proportional feedback, i.e. $u(t) = -Ky(t)$.



For what values of K is the closed-loop system guaranteed to be stable according to the small gain theorem?

Solution

1.4

Compute the norms $\|\cdot\|_\infty$ and $\|\cdot\|_2$ of the continuous-time signals

(a)

$$y(t) = \begin{cases} a \sin(t), & t > 0 \\ 0, & t \leq 0 \end{cases}$$

(b)

$$y(t) = \begin{cases} \frac{1}{t}, & t > 1 \\ 0, & t \leq 1 \end{cases}$$

(c)

$$y(t) = \begin{cases} e^{-t}(1 - e^{-t}), & t > 0 \\ 0, & t \leq 0 \end{cases}.$$

Solution**1.5**

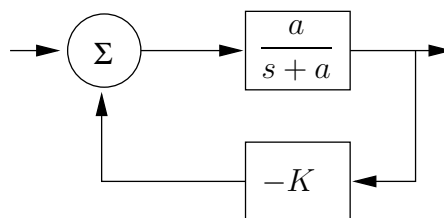
Consider the linear system

$$G(s) = \frac{\omega_0^2}{s^2 + 2\zeta\omega_0 s + \omega_0^2}.$$

Compute the system gain $\|G\|$ for all values of $\omega_0 > 0$ and $\zeta > 0$.

Solution**1.6**

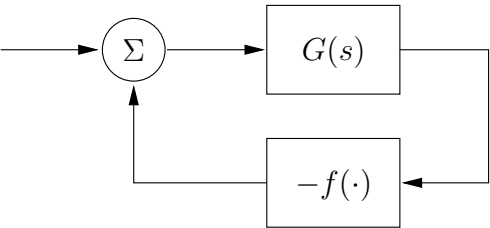
Analyze stability of the following system by using the small gain theorem and by computing the poles of the closed-loop system. Explain differences in the results.



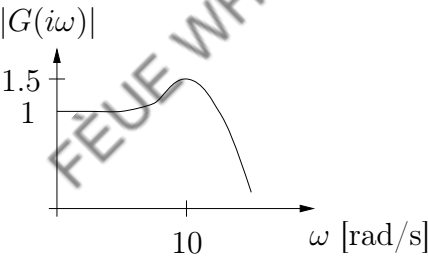
Solution

1.7

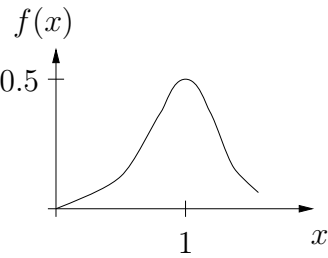
Consider the feedback control system



where $G(s)$ is a linear system with the magnitude plot



and $f(\cdot)$ is an amplifier with the following input-output relationship



Is the closed-loop system stable?

Solution

1.8

Consider a DC motor given on state-space form

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= -ax_2 + au \\ y &= x_1\end{aligned}$$

The inverse time constant a can vary as

$$a = 1 + \rho, \quad |\rho| < \delta.$$

The system is controlled using a proportional controller $u(t) = -Ky(t)$. Suppose that a is constant. Give a sufficient condition on K such that the closed-loop system is stable for all a .

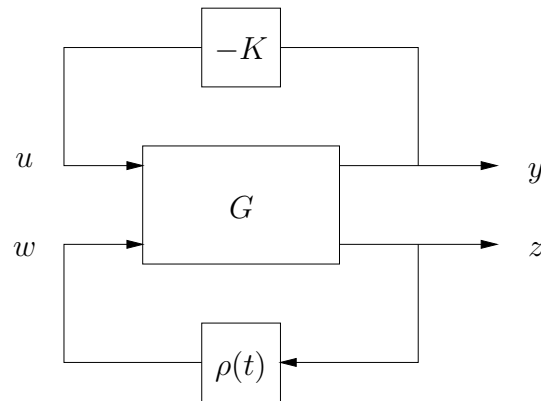
Solution

1.9

Once again consider the DC motor in exercise 1.8, but now assume that the parameter a can vary arbitrarily fast with time

$$a = a(t) = 1 + \rho(t), \quad |\rho(t)| < \delta, \quad \forall t$$

- (a) Introduce a new, artificial input signal w and a new artificial output signal z such that the system can be described by the feedback connection below



- (b) Consider the time-varying, static system from $z(t)$ to $w(t)$:

$$w(t) = \rho(t) \cdot z(t), \quad |\rho(t)| < \delta, \quad \forall t.$$

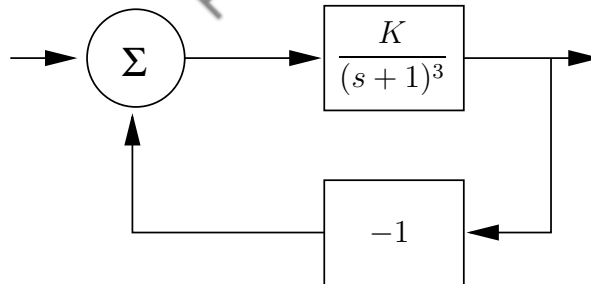
Show that the gain of this system (according to Definition 1.1 in the textbook) is at most δ .

- (c) Give a sufficient condition on K , for instance an inequality that implicitly characterizes K , for the closed-loop system to be stable no matter how $a(t)$ varies with time.

Solution

1.10

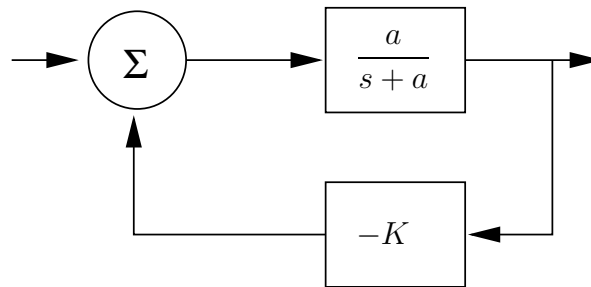
Analyze stability of the closed loop system below with respect to the feedback gain $K > 0$ using nyquist theory (use `nyquist` in MATLAB to understand what the curve looks like conceptually, but derive the stability condition analytically). Verify the results by computing closed-loop poles in MATLAB.



Solution

1.11

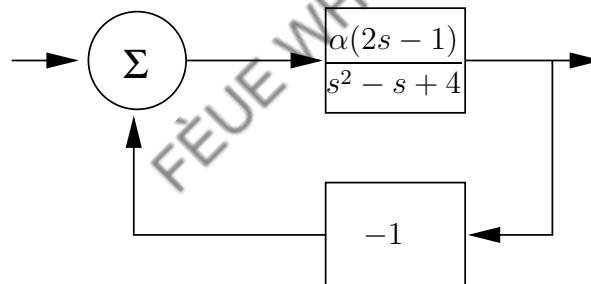
Analyze stability of the closed loop system below using nyquist theory. Verify the results by computing the closed-loop poles analytically.



Solution

1.12

Consider the closed loop system below.



- Derive the closed-loop transfer function and analyze stability with respect to the parameter $\alpha > 0$.
- Analyze closed-loop stability using nyquist theory.
- Confirm the structural findings using a root loci (`rlocus` in MATLAB).

Solution

2 Representation of Linear Systems

2.1

A simplified model of an alternating-current generator can be described as follows. The input signals to the system are the magnetizing current I_m , which is fed into the armature winding, and the driving torque M which is applied to the rotor axis. The rotation speed of the generator is ω , and the change in rotation speed is given by

$$J\dot{\omega} = M - M_e$$

where

$$M_e = K_e \cdot \omega \cdot I_f$$

is the electrical torque due to the emf. I_f is the current in the stator winding, given by the relationship

$$e = R \cdot I_f$$

where the voltage e is generated in the stator winding according to

$$e = C_e \cdot I_m \cdot \omega$$

and R is the load resistance applied to the stator winding. Consider e and ω as output signals, M , I_m and R as input signals. Set $K_e = C_e = J = 1$ and find a state-space representation for this system.

Linearize around the stationary point

$$\omega_0 = R_0 = I_{m0} = M_0 = 1$$

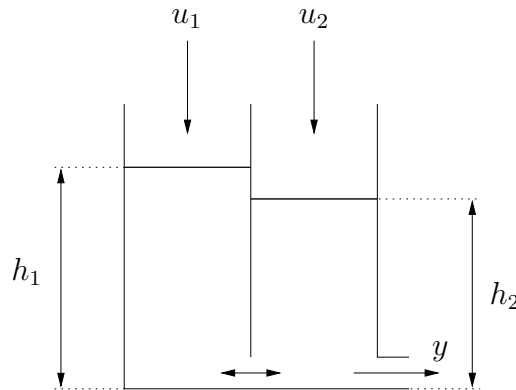
and derive the transfer function matrix from

$$u = \begin{bmatrix} \Delta M \\ \Delta I_m \\ \Delta R \end{bmatrix} \text{ to } y = \begin{bmatrix} \Delta \omega \\ \Delta e \end{bmatrix}$$

Solution

2.2

Consider the system consisting of two coupled tanks described in the figure below.



The flow of water into the left and right halves of the tank are denoted u_1 and u_2 respectively. These flows are the input signals. The water levels in the two halves are denoted h_1 and h_2 respectively. The flow y out from the tank is assumed to be proportional to the water level in the right half of the tank

$$y(t) = \alpha h_2(t)$$

The flow between the two halves is proportional to the difference between the levels

$$f(t) = \beta(h_1(t) - h_2(t))$$

where a flow from left to right is considered positive. Let h_i, u_i and y be deviations from nominal values. Thus, they can have negative values. Assume that the area of the halves are $A_1 = A_2 = 1$.

- Derive the transfer function from u_1, u_2 to y .
- Compute the maximum and minimum singular values of $G(0)$ and give an intuitive explanation to the corresponding input signals.

Solution

2.3

Find a state-space realization of the system

$$G(s) = \begin{bmatrix} \frac{1}{(s+1)(s+2)} & \frac{s+3}{(s+1)(s^2+s+1)} \end{bmatrix}$$

Solution

2.4

Find a state-space realization of the system

$$y(t) = \frac{p}{p^2 + 4p + 4}u_1(t) + \frac{p-1}{p^2 + 5p + 6}u_2(t)$$

Solution

2.5

A system is described by the differential equation

$$\ddot{y} + a_1\dot{y} + a_2y = b_{11}\dot{u}_1 + b_{12}u_1 + b_{21}\dot{u}_2 + b_{22}u_2.$$

Find a state-space realization.

Solution

2.6

Consider the system

$$\begin{cases} \dot{y}_1 + y_2 & = \dot{u} + 2u \\ \dot{y}_2 + y_2 + y_1 & = u \end{cases}$$

Find a state-space realization.

Solution

2.7

- (a) Consider $Y_1(s) = \frac{1}{s+1}U_1(s)$ and $Y_2(s) = \frac{1}{s+2}U_2(s)$. Introduce suitable definitions of $Y(s)$, $U(s)$ and $G(s)$ to write this as the multi-variable system $Y(s) = G(s)U(s)$.
- (b) Consider $Y_1(s) = \frac{1}{s+1}U_1(s) + \frac{1}{s+2}U_2(s)$. Introduce suitable definitions of $Y(s)$, $U(s)$ and $G(s)$ to write this as the multi-variable system $Y(s) = G(s)U(s)$.
- (b) Consider $Y_1(s) = \frac{1}{s+1}U_1(s)$ and $Y_2(s) = \frac{1}{s+2}U_1(s)$. Introduce suitable definitions of $Y(s)$, $U(s)$ and $G(s)$ to write this as the multi-variable system $Y(s) = G(s)U(s)$.

Solution

2.8

Extract a state-space model in the cases below (assuming the equations indeed are equations in a state-space model), i.e., define the states, inputs and outputs, and derive the 4 matrices defining the system. Conclude by writing the data in the format $\left[\begin{array}{c|c} A & B \\ \hline C & D \end{array} \right]$ to visually verify that all matrix sizes match. All variables in the equations below are scalars.

- (a) $\dot{x}_1 = -x_1 + 2x_2 + 3u_2 + 4u_3$, $\dot{x}_2 = 5u_1$, $y_1 = 6x_1 - 7x_2$, $y_2 = 8x_2 + 9u_1 + 10u_2 + 11u_3$
- (b) $\dot{v} = v + 2u + w$, $y = 3v + u - w$
- (c) $\dot{y} = u$, $\dot{x} = v$, $\dot{q} = h$, $z = y + 2x + 3q$.

Solution**2.9**

In a state-space representation $\dot{x} = Ax + Bu, y = Cx + Du$, the matrix D has width 4, the matrix A has height 3 and the output y has length 2. What are the dimensions of A, B, C, D, x, y and u .

Solution**2.10**

- (a) Consider the following model

$$\begin{aligned}\dot{\eta} &= Mq + R(p - \eta) + S\phi \\ \dot{q} &= N(\eta - q) - Sp\end{aligned}$$

Identify the states in this model, and assume all other signals are external inputs. All signals can be multi-variable, meaning M, R, S and N can be matrices. We are interested in the signals $\phi + q - \eta$ and q as outputs. Derive a suitable definition of states, inputs and outputs, and derive a standard state-space representation with 4 suitably defined matrices (we commonly call the state, input and output x, u and y , but that is of course not necessary, and in some cases even possibly confusing, but do that here)

- (b) In the model you end up with, there will be blocks of zeros and identity matrices. Explicitly describe the size of these, assuming p has length b while η, q and ϕ all have length a .

Solution

2.11

A controller implemented as a linear feedback from observed (estimated) states and feedforward from a reference is a multivariable system, taking reference r and measurements y as inputs, and outputting the control signal u (what you call input and output is in the eye of the beholder, in a controlled system the control signal is the input, but for the controller the control signal is the output)

The control signal is computed as $u = -L\hat{x} + l_0r$, and the state-estimate is evolving in the observer according to $\dot{\hat{x}} = A\hat{x} + Bu + K(y - C\hat{x})$. Identify how you can write this as a standard state-space representation modelling the relationship from inputs r and y to the output u .

Solution

2.12

Consider a model with two inputs and one output written in the following form

$$Y(s) = \frac{1}{s+1}U_1(s) + \frac{1}{s+1}U_2(s)$$

- Derive a state-space realization which uses two states. Explicitly write down the dimension of all 4 matrices in the state-space representation.
- Derive a state-space realization which only has one state! Explicitly write down the dimension of all 4 matrices in the state-space representation.
- If the model $\frac{1}{s+1}$ is a model from flow of water to the obtained water level in a tank, what would the two different models represent physically?
- Create the model in (a) in MATLAB using the command `ss`. Then convert that object to a transfer function representation by applying the command `tf` and verify that it creates something reasonable. Now take that transfer function and convert it to a state-space model by applying

`ss`. What is the result? What has happened? Use the commands `obsv` and `ctrb` which computes observability and controllability matrices (see their help on how to use them) to think of what might have occurred and what that tells us about the system. To obtain the matrices of a state-space representation in MATLAB of a system `G` you can simply write `G.a` etc.

- (e) Write the model in the form $Y(s) = G(s)U(s)$ and use Proposition 3.5 to analytically compute the poles of the transfer function. How does the result relate to the previous findings.

Solution

3 Properties of Linear Systems

3.1

Consider the transfer function matrix

$$G(s) = \begin{pmatrix} \frac{1}{s+2} & -\frac{1}{s+2} & \frac{1}{s+2} \\ \frac{1}{s+2} & \frac{s+1}{s+2} & \frac{1}{s+2} \end{pmatrix}$$

Derive the pole and the zero polynomials of the system? What is the dimension of a minimal state-space realization?

Solution

3.2

Find the poles and zeros of

$$G(s) = \frac{1}{(s+1)(s+3)} \begin{pmatrix} 1 & 0 \\ -1 & 2(s+1)^2 \end{pmatrix}$$

Solution

3.3

Find the poles of

$$G(s) = \frac{1}{(s+1)^2} \begin{pmatrix} 1-s & \frac{1}{3}-s \\ 2-s & 1-s \end{pmatrix}.$$

What is the dimension of a minimal realization?

Solution

3.4

(a) Consider the system

$$G(s) = \begin{pmatrix} \frac{s+5}{s^2+3s+2} & \frac{1}{s+2} \\ \frac{1}{s+4} & \frac{1}{s+2} \end{pmatrix}$$

What is the dimension of a minimal state-space realization?

(b) Consider the system

$$G(s) = \begin{pmatrix} \frac{s+5}{s^2+3s+2} & \frac{1}{s+2} \\ \frac{1}{s+4} & \frac{1}{s+4} \end{pmatrix}$$

What is the dimension of a minimal state-space realization?

Solution

3.5

A system has the following input-output relation

$$\begin{cases} \dot{y}_1 + y_1 - \dot{y}_2 &= u_1 - u_2 \\ \dot{y}_2 + \dot{y}_1 + y_2 &= u_1 + u_2 \end{cases}.$$

Find a matrix fraction description, $y(t) = A(p)^{-1}B(p)u(t)$, and compute the poles and zeros of the system.

Solution

3.6

Consider the MIMO system:

$$\begin{aligned}\dot{x}(t) &= \begin{pmatrix} -2 & 1 & 0 \\ 0 & -1 & 0 \\ 0 & 0 & -3 \end{pmatrix} x(t) + \begin{pmatrix} 0 & 1 \\ 1 & 0 \\ 1 & 0 \end{pmatrix} u(t) \\ y(t) &= \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix} x(t)\end{aligned}$$

Find a minimal realization of the system, i.e. a realization that is controllable and observable.

Solution

3.7

Consider the multivariable system

$$Y(s) = G(s)U(s)$$

where

$$G(s) = \begin{pmatrix} \frac{1}{s+1} & \frac{3}{s+2} \\ \frac{2}{s+3} & \frac{1}{s+4} \end{pmatrix}$$

- Determine the maximum and minimum singular value of the frequency response at the frequency $\omega = 2$ rad/s.
- Determine also the input vectors, in terms of their Fourier transforms, corresponding to the largest and smallest gain of the system at $\omega = 2$.
- Generate, in Matlab, an input vector that corresponds to the largest gain of the system and simulate the system using this input.

Hint: Use sinusoidal input signals and use the following properties of a sinusoidal signal considered over a finite time interval.

- The Fourier transform is proportional to the amplitude of the sinusoidal signal, i.e. for $u_1(t) = A \sin \omega t$ the Fourier transform $U_1(i\omega)$ is proportional to A .
 - Time delay of a signal corresponds to a change of the argument of the Fourier transform, i.e. if $u_1(t) = A \sin \omega t$ has the transform $U_1(i\omega)$ the signal $A \sin(\omega t + \phi)$ has the transform $U_1(i\omega)e^{i\phi}$.
- (d) Verify that the obtained output signals correspond to largest gain of the system.

Solution

3.8

Find the poles and zeros of

$$G(s) = \begin{pmatrix} \frac{1}{s+1} & 0 & \frac{s-1}{(s+1)(s+2)} \\ \frac{-1}{s-1} & \frac{1}{s+2} & \frac{1}{s+2} \end{pmatrix}.$$

Solution

5 Disturbance Models

5.1

A continuous-time signal $u(t)$ has the power spectrum $\Phi_u(\omega)$. For the power spectra below, find linear filters such that the processes can be represented as a white signal fed through those filters.

$$(a) \quad \Phi_u(\omega) = \frac{a^2}{\omega^2 + a^2}$$

$$(b) \quad \Phi_u(\omega) = \frac{a^2 b^2}{(\omega^2 + a^2)(\omega^2 + b^2)}$$

$$(c) \quad \Phi_u(\omega) = \frac{\omega^2 + c^2}{(\omega^2 + a^2)}$$

- (d) Sketch the amplitude gain of the linear system in (c) for the cases $c > a > 0$ and $a > c > 0$ (in a logarithmic scale, i.e., corresponding to using `bodemag` in MATLAB). Verify with MATLAB.

Solution

5.2

A position sensor is mounted on a machine that vibrates with a frequency around 5 Hz, and this causes that a disturbance $n(t)$ affects the position measurement. In order to include the properties of the measurement disturbance in the control design one formulates a model that describes the properties of the disturbance as a filtered white signal V . The following models are suggested

$$(i) \quad N(s) = \frac{1}{s + 0.001} V(s)$$

$$(ii) \quad N(s) = \frac{900s}{s^2 + 6s + 900} V(s)$$

$$(iii) \quad N(s) = \frac{25}{s^2 + s + 25} V(s)$$

Which disturbance model is the best choice? Confirm your reasoning using MATLAB to study the amplitude gain.

Solution

5.3

Consider a missile with mass m propelled by the thrust u . The missile's position is z . A simplified model for the air drag is

$$f = k_1 \cdot \dot{z} + v$$

where v represent wind gusts.

- (a) Derive a state-space representation and an input-output representation for how the controlled output z depends on u and v .
- (b) The system disturbance v has the spectral density

$$\Phi_v(\omega) = k_0 \cdot \frac{1}{\omega^2 + a^2}$$

Modify the state-space representation in (a) to make it possible to express the wind gusts disturbance v using a white signal. What can be said about the modelled frequency content of the wind gusts?

Solution

5.4

Assume, in exercise 5.3, that the position z is measured with an error

$$y(t) = z(t) + n(t)$$

Derive a state-space model for the missile if

(a)

$$\Phi_n(\omega) = 0.1$$

(b)

$$\Phi_n(\omega) = 0.1 \frac{\omega^2}{\omega^2 + b^2}.$$

(c)

$$\Phi_n(\omega) = 0.1 \frac{1}{\omega^2 + b^2}.$$

Solution**5.5**

A system has the state-space representation

$$\begin{aligned}\dot{x} &= Ax + Bu + Nw \\ y &= Cx + n\end{aligned}$$

We assume that the system disturbance w changes stepwise and that the measurement noise is periodical with a frequency of about 2 Hz.

Modify the state-space representation to make it possible to model the disturbances.

Solution**5.6**

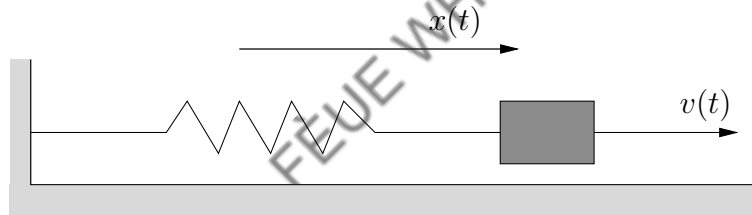
In airplanes it is common to measure acceleration as well as speed. The acceleration is measured using accelerometers and the speed is calculated from measurements of air data, such as dynamical pressure et cetera. Thus, the measurements are independent, but of course they are related to each other.

- (a) Derive a state-space model for the speed and acceleration. Let the measured speed and acceleration be output signals and assume that the derivative of the acceleration is white noise. Furthermore, assume that the measurement errors in speed and acceleration are white noises, independent of each other.
- (b) Discuss how we can get better estimates of the speed and acceleration using Kalman filtering.

Solution

5.7

The depicted dynamical system is described by the differential equation



$$\ddot{x}(t) + \dot{x}(t) = v(t)$$

The external force $v(t)$ is white noise with

$$\begin{aligned} E v(t) &= 0 \\ E v(t)v(s) &= \delta(t - s) \end{aligned}$$

We want to estimate the position $x(t)$ and speed $\dot{x}(t)$ at every time instant. We have sensors for both speed and position but for economical reasons we only want to use one sensor. We can choose between

Alternative I: The measured signal is

$$y_1(t) = x(t) + e_1(t)$$

Alternative II: The measured signal is

$$y_2(t) = \dot{x}(t) + e_2(t)$$

The measurement errors are $e_1(t)$ and $e_2(t)$. For simplicity we assume that they are both white noises with

$$Ee_1(t) = Ee_2(t) = E[e_1(t)e_2(s)] = 0$$

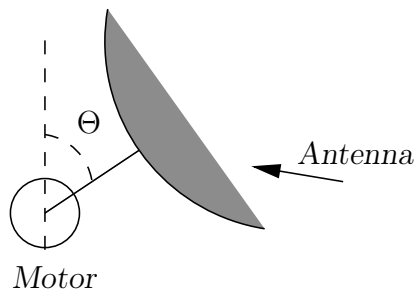
$$Ee_1(t)e_1(s) = Ee_2(t)e_2(s) = \delta(t - s)$$

For each alternative derive the linear filter that, in steady state, yields the best estimate of $x(t)$ and $\dot{x}(t)$, in the sense of smallest variance of the estimation error, from measurements up to and including time t . State, with an explanation, which alternative you think is the best.

Solution

5.8

Consider the depicted radar antenna.



From noisy measurements of the position of the antenna Θ_m we want to estimate the true position Θ . To be able to do this we need a model of the system. To this end, describe the dynamics of the antenna with

$$J\ddot{\Theta}(t) + B\dot{\Theta}(t) = \tau(t) + \tau_d(t),$$

where J is the moment of inertia for the moving parts of the antenna, B is the coefficient of viscous friction, $\tau(t)$ is the torque produced by the motor, and $\tau_d(t)$ is the torque caused by the wind. Assume that $\tau_d(t)$ can be modeled as white noise. Furthermore, assume that the torque $\tau(t)$ is proportional to the motor voltage, $\mu(t)$, i.e.

$$\tau(t) = k\mu(t)$$

Finally, let us for simplicity, assume that the measurement error can be modeled as additive white noise $e_m(t)$. Hence, the output signal is

$$\Theta_m(t) = \Theta(t) + e_m(t).$$

Discuss how $\Theta(t)$ can be estimated from $\Theta_m(t)$ using a Kalman filter.

Technical data:

$$B/J = 4.6 \text{ s}^{-1}$$

$$k/J = 0.787 \text{ rad/Vs}^2$$

$$J = 10 \text{ kg m}^2$$

$$E \tau_d(t)\tau_d(s) = v_d\delta(t-s) = 10 \text{ N}^2\text{m}^2 \cdot \delta(t-s)$$

$$E e_m(t)e_m(s) = v_m\delta(t-s) = 10^{-7} \text{ rad}^2 \cdot \delta(t-s)$$

Solution

5.9

Consider an electric motor with transfer operator

$$G(p) = \frac{1}{p(p+1)}$$

from input voltage to actual angular displacement. The motor operates in two disturbance modes:

(i)

$$y(t) = G(p)(u(t) + w(t))$$

(ii)

$$y(t) = G(p)u(t) + w(t)$$

In both cases we have $w(t) = \frac{1}{p}v(t)$ where $v(t)$ is a unit disturbance, for example an impulse.

- (a) Realize both cases on state-space form. For case (ii) it is assumed that the states caused by the disturbance are separate from the ones describing the motor dynamics.
- (b) For both cases, give examples of physical phenomena that can be modeled with the disturbance $w(t)$.
- (c) Study the two state-space realizations. Are all states controllable? Can states corresponding to $w(t)$ be made unobservable? Can the influence of $w(t)$ on $y(t)$ be eliminated?

Solution

5.10

Consider the movement of a swing due to the wind. The swing is described by the transfer operator

$$y(t) = \frac{1}{p^2 + p + 1}u(t)$$

where the output signal $y(t)$ is the angular displacement and the input signal $u(t)$ is the torque about the point of suspension. The influence of the wind can be modeled as

$$u(t) = Kv(t)$$

where $v(t)$ is a Gaussian distributed disturbance with the spectrum

$$\Phi_v(\omega) = \frac{2\alpha}{\alpha^2 + \omega^2}, \quad \alpha > 0.$$

K quantifies the strength of the wind and α quantifies the gustiness of the wind.

- (a) Does α increase or decrease when the gustiness increases, i.e. when the wind changes direction more frequently?
- (b) Derive and interpret conditions on α and K such that the swing has an angular displacement of more than 1.15 at least a quarter of the time. This is equivalent to the output having a variance greater than 1.

Hint:

$$\begin{aligned} \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{|b_2(i\omega)^2 + b_1i\omega + b_0|^2}{|(i\omega)^3 + a_2(i\omega)^2 + a_1i\omega + a_0|^2} d\omega \\ = \frac{b_2^2 a_0 a_1 + (b_1^2 - 2b_0 b_2) a_0 + b_0^2 a_2}{2a_0(-a_0 + a_1 a_2)} \end{aligned}$$

Solution

6 The Closed-Loop System

6.1

For a given system G and a given controller F we have defined four transfer functions as

$$\begin{aligned} G_{wu} &= (I + FG)^{-1}, & G_{wu} &= -(I + FG)^{-1}F \\ G_{wy} &= (I + GF)^{-1}G, & G_{wy} &= (I + GF)^{-1} \end{aligned}$$

All four transfer functions have to be stable for the closed-loop system to be internally stable.

Show that

$$\begin{pmatrix} G_{wu} & G_{wu} \\ G_{wy} & G_{wy} \end{pmatrix} = \begin{pmatrix} I & F \\ -G & I \end{pmatrix}^{-1}$$

Solution

6.2

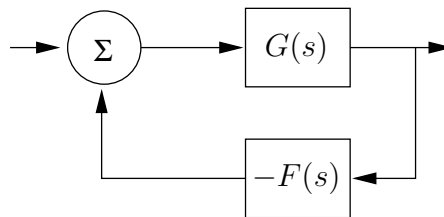
The system

$$G(s) = \frac{s-1}{s+1}$$

and the controller

$$F(s) = \frac{s+2}{s-1}$$

are used in the feedback connection depicted below.



Compute G_c , T and S . Are they stable? Is the closed-loop system internally stable?

Solution

7 Limitations in Control Design

7.1

Given the system

$$G(s) = \frac{s - 3}{s + 1}.$$

we want the complementary sensitivity function to be

$$T(s) = \frac{5}{s + 5}.$$

- (a) Compute a controller $F_r = F_y = F$ which results in this T . Will this controller really work?
- (b) Suggest an alternative T , still having the bandwidth 5 rad/s, but resulting in an internally stable system with $F_r = F_y = F$.
- (c) A rule of thumb for control of non-minimum phase systems states that the bandwidth of the closed-loop system cannot realistically be greater than half the value of the non-minimum phase zero. In this case 1.5 rad/s. Have we circumvented this rule of thumb in the above design or does the closed-loop system have any disadvantages?

Solution

7.2

A continuous-time system has a zero at $s = 3$ and a time-delay of 1.0 second. What is the upper limit of the realistic bandwidth/crossover frequency if the magnitude curve of the open-loop system decreases monotonically?

Solution

7.3

Give an example of a system for which there exists no controller having all three properties: a stable closed-loop system, small magnitude of the sensitivity function at low frequencies and small amplification of measurement errors at high frequencies.

Solution

7.4

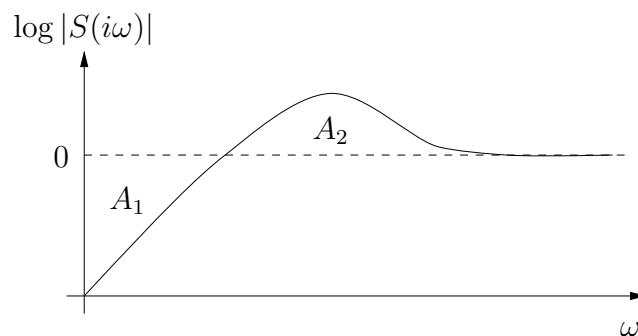
A multivariable system is supposed to attenuate system disturbances (w) at least a factor 10 for frequencies under 0.1 rad/s. Furthermore, measurement disturbances (n) should be attenuated at least a factor 10 for frequencies above 2 rad/s. Constant system disturbances should be attenuated at least a factor 100 in steady state.

- Formulate conditions on the singular values of S and T which will guarantee that the requirements are fulfilled.
- Translate the specifications into requirements on the loop gain GF_y .
- Formulate the requirements using $\|\cdot\|_\infty$ and frequency weights W_S och W_T .
- Which crossover frequency and phase margin would we expect, having the weights i (b), had the system been a SISO system? What lower bound on $\|T\|_\infty$ does this result in?
- Is this lower bound on $\|T\|_\infty$ consistent with the requirements in (c)?

Solution

7.5

A control system has the sensitivity function S , depicted below



What can be stated about the open-loop system if the surface A_2 is larger than the surface A_1 ?

Solution

7.6

For a certain feedback system we demand that:

- (i) output disturbances, with frequencies under 2 rad/s, should be attenuated at least a factor 1000.
- (ii) the system should remain stable despite a model uncertainty

$$|\Delta G| \leq 100|G|$$

for frequencies above 20 rad/s. G is the frequency response of the nominal system and ΔG is the absolute error in the frequency response.

Can this be accomplished using a linear, time-invariant controller?

Solution

7.7

We have the following specifications on a SISO system

$$\begin{aligned} |S(i\omega)| &\leq 10^{-3}, & \omega &\leq 1 \\ |T(i\omega)| &\leq 10^{-3}, & \omega &\geq 100 \end{aligned}$$

- (a) State two non-constant frequency weights W_S and W_T which would guarantee that the specifications are met.
- (b) Trying to find a controller fulfilling the design criteria, for example using the methods presented in Chapter 10 in the textbook, we fail. Should this have been anticipated from the very beginning?

Solution

8 Controller Structure and Control Design

8.1

Let

$$G(s) = \begin{pmatrix} \frac{1}{s+2} & \frac{10}{s+1} \\ \frac{1}{s+5} & \frac{5}{s+3} \end{pmatrix}.$$

- Compute $\text{RGA}(G(0))$.
- Which input-output pairing should be avoided?

Solution

8.2

Given the multivariable system

$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \frac{1}{0.1s + 1} \begin{pmatrix} \frac{0.6}{s+1} & -0.4 \\ 0.3 & 0.6 \end{pmatrix} \begin{pmatrix} u_1 \\ u_2 \end{pmatrix}.$$

Assume that we want the controller to be diagonal and that we use the *relative gain array* (RGA) to decide what input should control what output. Furthermore, assume that we want a crossover frequency of $\omega_c = 10$ rad/s. Decide how the signals should be paired.

Solution

8.3

Study the multivariable system

$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \begin{pmatrix} \frac{1}{10s+1} & \frac{-2}{2s+1} \\ \frac{1}{10s+1} & \frac{s-1}{2s+1} \end{pmatrix} \begin{pmatrix} u_1 \\ u_2 \end{pmatrix}.$$

- (a) Decide, using RGA analysis, which input signal should control which output signal.
- (b) Assume that we want to use decentralized control, i.e. we want a controller on the form

$$F = W_1 F^{\text{diag}}(s) W_2, \text{ where } F^{\text{diag}}(s) = \begin{pmatrix} F_{11}(s) & 0 \\ 0 & F_{22}(s) \end{pmatrix}.$$

Furthermore, assume that we do not want the steady-state error in one channel to affect the steady-state error in the other channel. Give the structure of a controller $F(s)$, expressed in $F^{\text{diag}}(s)$, that will accomplish this, using static decoupling (statisk frikoppling). Verify numerically in MATLAB that the loop-gain is decoupled (use any proportional controllers you want).

Solution

8.4

Design a controller, using the IMC method, for a stable first order process

$$G(s) = \frac{K}{\tau s + 1}, \quad \tau > 0.$$

What type of controller do we get? Compute the sensitivity function and the complementary sensitivity function and sketch the Bode plot of the sensitivity function. What does Bode's integral theorem state for this case?

Solution

8.5

Design a controller, using the IMC method, for the system

$$G(s) = \frac{6 - 3s}{s^2 + 5s + 6}.$$

What type of controller do we get?

Solution

8.6

Consider the DC motor

$$y = \frac{1}{p(p+1)}u$$

Compute an IMC based controller for this system. Write the controller on the form $u = -F_y(p)y$, and sketch the Bode plot for $F_y(p)$. Approximately what type of controller do we get when we want a high bandwidth for the closed-loop system?

Solution

8.7

Given the multivariable system

$$G(s) = \frac{1}{s/20 + 1} \begin{pmatrix} \frac{9}{s+1} & 2 \\ 6 & 4 \end{pmatrix}.$$

- What are the poles and zeros of $G(s)$?
- Compute an IMC based controller for the system.

Solution

8.8

Consider the system

$$G(s) = \begin{pmatrix} \frac{2}{s+1} & \frac{3}{s+2} \\ \frac{1}{s+1} & \frac{1}{s+1} \end{pmatrix}$$

(Example 1.1 in the textbook)

Show how an IMC based controller can be computed for this system. Give an explicit expression for the corresponding sensitivity function.

Solution

8.9

Consider the multivariable system

$$Y(s) = G(s)U(s)$$

where

$$G(s) = \begin{pmatrix} \frac{2}{s+1} & \frac{3}{s+2} \\ \frac{\alpha}{s+1} & \frac{1}{s+1} \end{pmatrix}$$

and $\alpha > 0$.

- Determine the zero of the multivariable system. How does the zero depend on the value of α ?
- Assume that one would like to achieve complete decoupling of the system $G(s)$ such that

$$G(s)F(s) = \begin{pmatrix} \frac{1}{(s+1)^2} & 0 \\ 0 & \frac{1}{(s+1)^2} \end{pmatrix}$$

Are there any cases when this is not a good idea? Motivate!

- Assume that one instead chooses to use a static decoupling such that $G(s)F(s)$ is decoupled for $\omega = 0$. Are there any values of α for which this is not a good idea? Motivate!

Solution

8.10

Consider the multivariable system

$$Y(s) = G(s)U(s)$$

where

$$G(s) = \begin{pmatrix} \frac{1}{s+2} & \frac{2}{s+4} \\ \frac{1}{s+1} & \frac{1}{s+2} \end{pmatrix}$$

- (a) Determine the RGA at $\omega = 0$.
- (b) Assume that the system is going to be controlled by the diagonal regulator

$$U(s) = F(s)(R(s) - Y(s))$$

where

$$F(s) = \begin{pmatrix} K & 0 \\ 0 & K \end{pmatrix}$$

Use the result from a) to judge how successful this will be. Compute the poles (using MATLAB) of the closed loop system for the case $K = 5$.

- (c) How can the controller (or system model) be modified such that a diagonal (one input controls one output) $F(s)$ can be used? Verify that the closed loop system is stable for $K = 5$ for the modified setup, and perform a step-response.

Solution

9 Minimization of Quadratic Criteria: LQG

9.1

Consider the system

$$G(s) = \frac{1}{s-1}$$

represented on state-space form with noise as

$$\begin{aligned}\dot{x}(t) &= x(t) + u(t) + v_1(t) \\ z(t) &= x(t) \\ y(t) &= x(t) + v_2(t)\end{aligned}$$

The noises $v_i(t)$ are white with intensities R_i . We use the criterion

$$V = \int Q_1 x^2(t) + Q_2 u^2(t) dt,$$

and want to find the LQG controller.

- Show that the controller is a function of $\alpha = Q_1/Q_2$ and $\beta = R_1/R_2$ only.
- Compute the poles of the closed-loop system as a function of α and β .

Solution

9.2

Consider the system

$$\begin{aligned}z &= \frac{1}{p+1}u + \frac{1}{p+1}v \\ y &= z + e\end{aligned}$$

where v and e are scalar white noise disturbances with $Ev^2 = r_1$ and $Ee^2 = r_2$. We minimize the criterion

$$V = \int q_1 z^2(t) + u^2(t) dt$$

- (a) Compute the loop gain of the feedback connection.
- (b) How do r_1 and q_1 influence the loop gain?
- (c) Sketch the magnitude of the frequency response. What happens when $r_1 \rightarrow \infty$ and when $q_1 \rightarrow \infty$ respectively?

Solution

9.3

Consider the double integrator

$$\ddot{z}(t) = u(t).$$

We want to find a controller such that the criterion

$$\int_0^\infty (z^2(t) + \eta \cdot \dot{z}^2(t)) dt$$

is minimized for some $\eta > 0$. We assume that $z(t)$ and $\dot{z}(t)$ are both known (and need not to be estimated).

Where are the poles of the optimal closed-loop system located? How is the control signal affected when η is decreased?

Solution

9.4

Consider the antenna in Exercise 5.8. We want to control it and a suitable measure on the performance of closed-loop system is given by the criterion

$$J = E \{ \Theta^2(t) + \rho \mu^2(t) \}$$

where ρ is a constant we can choose. Derive an optimal control signal and discuss how it is to be combined with the Kalman filter.

Solution

9.5

Consider control of the DC motor

$$G(s) = \frac{1}{s(s+1)}$$

We want to use the motor together with a system that has a resonance peak at approximately 0.5 rad/s. Other than that, we do not know much about the system. Describe how we can compute an LQG controller with good robustness qualities, i.e. small complementary sensitivity gain, at this frequency.

Solution

9.6

A system has static gain G_0 . It is influenced by system disturbances, with all energy concentrated at zero frequency, i.e.

$$\Phi_\nu(\omega) = \delta(\omega)$$

The reference signal is zero, as is the measurement noise. We choose a controller that minimizes

$$E \{y^2(t) + \alpha u^2(t)\}$$

What is the value of the sensitivity function at zero frequency?

Solution

9.7

Consider the system

$$\begin{aligned} z &= \frac{1}{p+1}u + \frac{1}{p+1}\nu \\ y &= z + e \end{aligned}$$

where ν is noise of very low frequency,

$$\nu = \frac{1}{p + \varepsilon} v,$$

v and e are noises with $\Phi_v(\omega) \equiv \Phi_e(\omega) \equiv 1$.

- (a) Find a controller that minimizes

$$E \{ z^2 + u^2 \}$$

when $\varepsilon \rightarrow 0$.

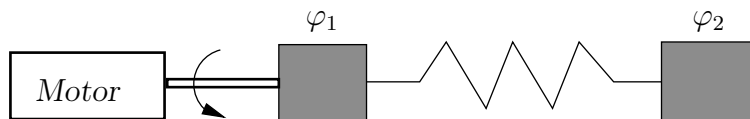
What is the static gain of the sensitivity function?

- (b) Use output-LTR (LTR(y)) to compute L . What is the static gain of the sensitivity function?

Solution

9.8

Consider a motor driving two rotating masses connected by a flexible shaft:



The angular displacements of the masses are φ_1 and φ_2 respectively and ω_1 and ω_2 are the angular velocities. The moments of inertia are 10 for both masses. The spring rate of the shaft is k and the damping factor is 0.1. The input is the voltage applied to the motor. With the states $x_1 = \varphi_1 - \varphi_2$, $x_2 = \omega_1$ and $x_3 = \omega_2$ we get the state-space representation

$$\dot{x} = \begin{pmatrix} 0 & 1 & -1 \\ -\frac{1}{2}\omega_0^2 & -0.01 & 0.01 \\ \frac{1}{2}\omega_0^2 & 0.01 & -0.01 \end{pmatrix} x + \begin{pmatrix} 0 \\ \omega_0 \\ 0 \end{pmatrix} u$$

$$z = (0 \quad 0 \quad 1) x$$

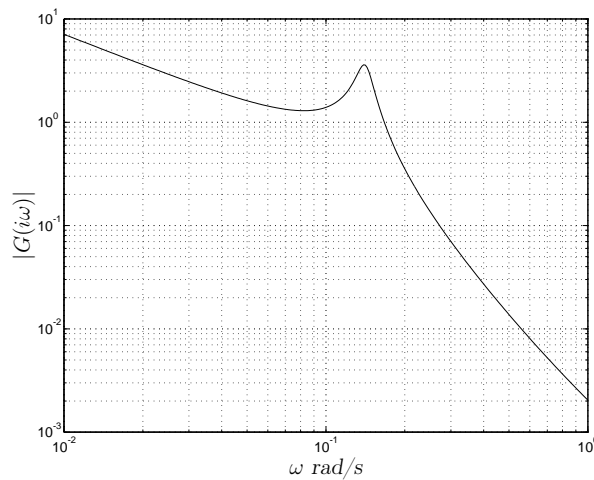
where

$$\omega_0^2 = \frac{k}{50}$$

The Bode plot, when $k = 1$, is shown in the figure below. There is a resonance peak at the frequency ω_0 . The spring rate is not exactly known, but has a value close to 1. We want to design a controller that yields a stable closed-loop system despite variations in k .

How can the above model be extended with a model for the noise to assure robustness for an uncertain value of k when we use LQG controller design? Give an actual example of such an extended system.

Solution



9.9

Consider the system

$$\dot{x}(t) = \begin{pmatrix} 1 & 0 \\ 0 & -2 \end{pmatrix} x(t) + \begin{pmatrix} 3 \\ 2 \end{pmatrix} u(t)$$

Show that

$$u(t) = - \begin{pmatrix} 2 & -3 \end{pmatrix} x(t)$$

cannot be an optimal state feedback for any quadratic criterion on the form

$$\min \int (x^T(t)Q_1x(t) + Q_2u^2(t)) dt$$

where Q_1 is a positive definite matrix.

Solution

9.10

Consider the system

$$\begin{aligned} \dot{x} &= \begin{pmatrix} 1 & -1 \\ 2 & 4 \end{pmatrix} x + \begin{pmatrix} -4 \\ 8 \end{pmatrix} u \\ y &= (1 \quad 1)x \end{aligned}$$

We want to minimize the criterion

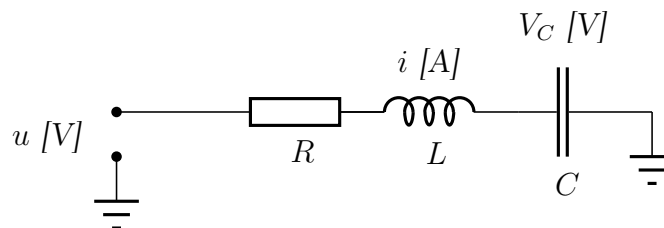
$$V(T) = \int_0^T x^T(t)x(t) + u^2(t) dt$$

Is it possible to find a state feedback $u = -Lx$ such that $V(T) < \infty$ when $T \rightarrow \infty$?

Solution

9.11

The figure below shows a simple electrical circuit.



Introduce the state variables $x_1 = V_C$ and $x_2 = i$. With the component values

$$R = 5 \, \Omega, \quad L = 0.1 \, H, \quad C = 1000 \, \mu F$$

we get the state-space representation

$$\begin{aligned}\dot{x}(t) &= \begin{pmatrix} 0 & 1000 \\ -10 & -50 \end{pmatrix} x(t) + \begin{pmatrix} 0 \\ 10 \end{pmatrix} u(t) \\ y(t) &= (1 \ 0) x(t)\end{aligned}$$

Compute a state feedback that minimizes

$$J = \int_0^\infty (x_2^2(t) + 0.01u^2(t)) \, dt$$

This criterion aims at limiting the power loss without getting too large signals.

Solution

9.12

A system has the state-space representation

$$\begin{aligned}\dot{x}(t) &= \begin{pmatrix} -2 & -1 & 1 \\ 1 & 0 & 0.5 \\ 0 & 0 & A \end{pmatrix} x(t) + \begin{pmatrix} 1 \\ 0.5 \\ 0 \end{pmatrix} u(t) + \begin{pmatrix} 0 \\ 0 \\ 1 \end{pmatrix} v(t) \\ z(t) &= (1 \ 0 \ 0) x(t) \\ y(t) &= z(t) + e(t)\end{aligned}$$

where $e(t)$ and $v(t)$ are unit disturbances.

The controller, a feedback from reconstructed states, minimizes

$$E [z^2(t) + u^2(t)]$$

How does the value of A affect the sensitivity function?

Solution

9.13

A simplified model for how the elevator angle affects the movements of an airplane is given by

$$\dot{x} = \begin{bmatrix} -0.01 & 0.03 & -10 \\ 0 & -1 & 300 \\ 0 & 0 & -0.5 \end{bmatrix} x + \begin{bmatrix} 4 \\ -20 \\ -10 \end{bmatrix} u$$

where

$$x = \begin{bmatrix} \text{roll angle} \\ \text{yaw angle} \\ \text{pitch-angle velocity} \end{bmatrix}$$

In particular we are interested in the control of the pitch-angle velocity and choose the controlled variable to be

$$z = \begin{bmatrix} 0 & 0 & 1 \end{bmatrix} x$$

All state variables are measured

$$y = x + e$$

We want to design a feedback from reconstructed states using LQG methodology. It is especially important that the sensitivity function has a small gain for frequencies around 1 rad/s. Show how to modify the model of the airplane to achieve such a sensitivity function.

Solution

9.14

Consider the system

$$\dot{x}(t) = \alpha x(t) + u(t) \quad x(0) = x_0 \quad (1)$$

The system is controlled by the feedback

$$u(t) = -Lx(t) \quad (2)$$

where L is chosen such that

$$J = \int_0^\infty x^2(t) + \rho u^2(t) dt \quad (3)$$

is minimized.

- (a) Determine L as function of ρ and α .
- (b) If it is desired to keep $u(t)$ small, this can be achieved by choosing ρ large. What is the resulting L when $\rho \rightarrow \infty$? Consider, for example, the cases $\alpha = 1$ and $\alpha = -1$, respectively. Why is it not optimal to choose $L = 0$, i.e. $u(t) = 0$, in both cases?

Solution

9.15

An electrical motor has the transfer functions

$$Y(s) = \frac{1}{s(s+1)}U(s)$$

and it is controlled using state feedback

$$u(t) = -Lx(t) \quad (r(t) = 0)$$

where angle $x_1(t) = y(t)$ and angular velocity $x_2(t) = \dot{y}(t)$. The gain vector L is determined by minimizing the criterion

$$J = \int_0^\infty x^T(t)Q_1x(t) + Q_2u^2(t)dt$$

Figure 1 shows the simulation results when the system starts in the initial condition $x(0) = (1 \ 1)^T$ for some different choices of Q_1 and Q_2 (can you identify x_1 vs x_2). Combine the figures with the cost/weight-matrices. Yes, two of the responses look suspiciously similar...

(i)

$$Q_1 = \begin{pmatrix} 1 & 0 \\ 0 & 0 \end{pmatrix} \quad Q_2 = 0.1$$

(ii)

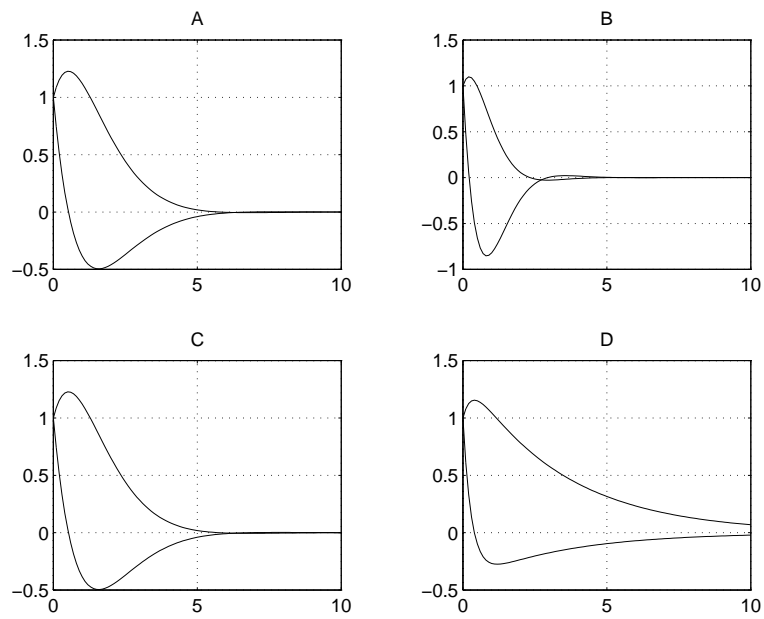
$$Q_1 = \begin{pmatrix} 1 & 0 \\ 0 & 10 \end{pmatrix} \quad Q_2 = 1$$

(iii)

$$Q_1 = \begin{pmatrix} 0.1 & 0 \\ 0 & 0 \end{pmatrix} \qquad Q_2 = 0.1$$

(iv)

$$Q_1 = \begin{pmatrix} 1 & 0 \\ 0 & 0 \end{pmatrix} \qquad Q_2 = 1$$

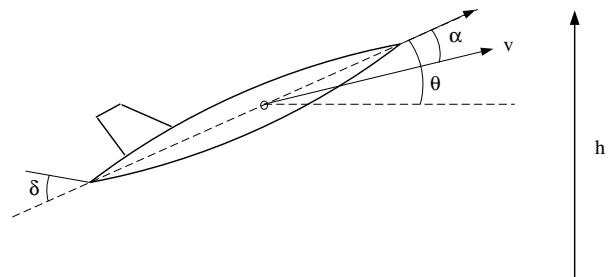


Figur 1.

Solution

9.16

Consider the simplified description of an aircraft in the figure below.



Using the state space variables

$$\begin{aligned}
x_1(t) &= \alpha(t) && \text{angle of attack (rad)} \\
x_2(t) &= \dot{\theta}(t) && \text{pitch rate (rad/s)} \\
x_3(t) &= \theta(t) && \text{pitch angle (rad)} \\
x_4(t) &= h(t) && \text{height (deviation from an operating point)}
\end{aligned}$$

the input signal

$$u(t) = \delta(t) \quad \text{control surface angle (rad)}$$

and the primary state to be controlled is the height $h(t)$

The dynamics is described by

$$\dot{x} = Ax + Bu$$

where

$$A = \begin{pmatrix} -0.17 & 1 & 0 & 0 \\ -0.56 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ -2.22 & 0 & 2.22 & 0 \end{pmatrix}, \quad B = \begin{pmatrix} 0.011 \\ 0.56 \\ 0 \\ 0 \end{pmatrix}$$

- (a) Is the system asymptotically stable?
- (b) Assume that the system has the initial state

$$x_0 = (0 \ 0 \ 0.1 \ 1)^T$$

and that the system is controlled by the state feedback)

$$u = -Lx + r$$

(with reference r assumed to be 0 in our experiments here) where the gain vector L is chosen such that the criterion

$$\int_0^\infty x^T(t)Q_1x(t) + u^T(t)Q_2u(t) dt$$

is minimized. Note that Q_1 is a penalty on the states, i.e., it is following MATLABs notation, or alternatively, we have defined the controlled variables to simply be $z = x$. Assume that the matrices are chosen as

$$Q_1 = \begin{pmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 \end{pmatrix}, \quad Q_2 = 1$$

Determine the poles of the closed loop system. Simulate the closed loop system (either using Simulink or the command `initial`) and study the achieved response of states and the control signal.

- (c) Assume that Q_2 is varied. How does that affect the location of the closed loop poles and the properties of x and u ?
- (d) Assume now that the following conditions shall be fulfilled:
 - $|x_1| < 0.2$ all the time.
 - $|x_4| < 0.1$ after 25 seconds.
 - $|u| < 0.5$ after one second.

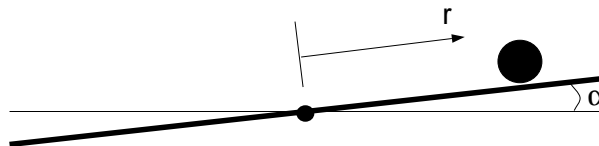
Determine Q_1 and Q_2 such that these conditions are satisfied. What is the resulting location of the closed loop poles?

- (e) When only focusing on and penalizing a single state (or a few, or a linear combination of some states) it is easier to introduce a controlled output $z = Mx$ and then decide on a suitable penalty on z instead, i.e., use the cost $\int_0^\infty z^T(t)Q_1z(t) + u^T(t)Q_2u(t) dt$. Do that for the setup in (a) and confirm that the resulting penalty on the state will match.

Solution

9.17

The figure below illustrates a system consisting of a ball on a plane. The variable r denotes the position of the ball relative to the center of the plane, and α represents the angle of the plane. The input signal is the torque that rotates the plane.



Figur 2: Ball on plane.

The system is represented by the state variables

- $x_1(t)$ - position, $r(t)$
- $x_2(t)$ - velocity, $\dot{r}(t)$
- $x_3(t)$ - plane angle, $\alpha(t)$
- $x_4(t)$ - plane angular velocity, $\dot{\alpha}(t)$

and torque is the input signal $u(t)$. The state space model is

$$\dot{x}(t) = Ax(t) + Bu(t)$$

where

$$A = \begin{pmatrix} 0 & 1 & 0 & 0 \\ 0 & 0 & -7 & 0 \\ 0 & 0 & 0 & 1 \\ 0 & 0 & 0 & 0 \end{pmatrix} \quad B = \begin{pmatrix} 0 \\ 0 \\ 0 \\ 1 \end{pmatrix} \quad C = (1 \quad 0 \quad 0 \quad 0)$$

(a) Assume that the system starts in the initial state

$$x(0) = (0.1 \quad 0 \quad -0.1 \quad 0)^T$$

i.e. the ball is positioned to the right of the center, and the plane leans downwards on the right side. Assume that all state variables can be measured. Determine a state feedback such that the following requirements are fulfilled:

- $|x(t)| \rightarrow 0$ when $t \rightarrow \infty$.
- $|x_1(t)| \leq 0.2 \quad \forall t$.
- $|u(t)| \leq 2.5 \quad \forall t$

Determine also the absolute value of the poles of the closed loop system.

(b) Verify that all sensors that measure the states have to work in order to obtain a stable closed loop system.

Hint: The characteristic equation of the closed loop system is given by

$$\lambda^4 + l_4\lambda^3 + l_3\lambda^2 - 7l_2\lambda - 7l_1 = 0$$

Missing a sensor is equivalent to setting the corresponding feedback l_i to zero.

Solution

10 Loop Shaping

10.1

Consider the system

$$y = \frac{1}{p+1}u$$

We want to create a closed-loop system with S , T and G_{wu} , such that

$$\int \left| \frac{S(i\omega)}{i\omega} \right|^2 + |0.5 T(i\omega)|^2 + |5 G_{wu}(i\omega)|^2 d\omega$$

is minimized. Compute the controller.

Solution

10.2

Consider the system

$$y = \frac{1}{p+1}u$$

We want to create a closed-loop system with S , T and G_{wu} , such that

$$\begin{aligned} |S(i\omega)| &< \gamma\omega \\ |T(i\omega)| &< 2\gamma \\ |G_{wu}(i\omega)| &< 0.2\gamma \end{aligned}$$

Write down the equations that determines the controller.

Solution

10.3

Consider the SISO system $G(s)$ with state-space realization

$$\begin{aligned}\dot{x} &= Ax + Bu \\ y &= Cx\end{aligned}$$

We want to use loop shaping with the weights

$$W_S = \frac{1}{s}, \quad W_T = 1, \quad W_u = 1$$

- State the equations that determine the optimal controller in \mathcal{H}_2 and \mathcal{H}_∞ respectively.
- Explicitly write down the observer for the extended state vector and show that the optimal controller can be written as

$$u(t) = \frac{\alpha}{1 + L(pI - A)^{-1}B} \int_0^t y(\tau) d\tau$$

for some L , where $\alpha = 1$ for the \mathcal{H}_2 controller and $\alpha > 1$ for the \mathcal{H}_∞ controller. State the equation determining L .

- Show that the controller will have a pole at the origin unless the system does itself has a pole at the origin.

Solution

10.4

Once again consider the system in Exercise 9.8.

- Suggest frequency weights W_S , W_T and W_u , for \mathcal{H}_2 and \mathcal{H}_∞ design, such that we get robustness against uncertain values of k .
- State the extended system from u and w to z on state-space form.

Solution

10.5

A DC-motor has transfer function

$$G(s) = \frac{1}{s(s+1)}$$

and it going to be controlled using proportional feedback

$$U(s) = K(R(s) - Y(s))$$

The properties of the closed loop system is specified via the requirement

$$|S(i\omega)| < |W_S^{-1}(i\omega)| \quad \forall \omega$$

The figure below shows three alternatives for the weight function $W_S^{-1}(i\omega)$. Which alternative is the best? Motivate the answer.

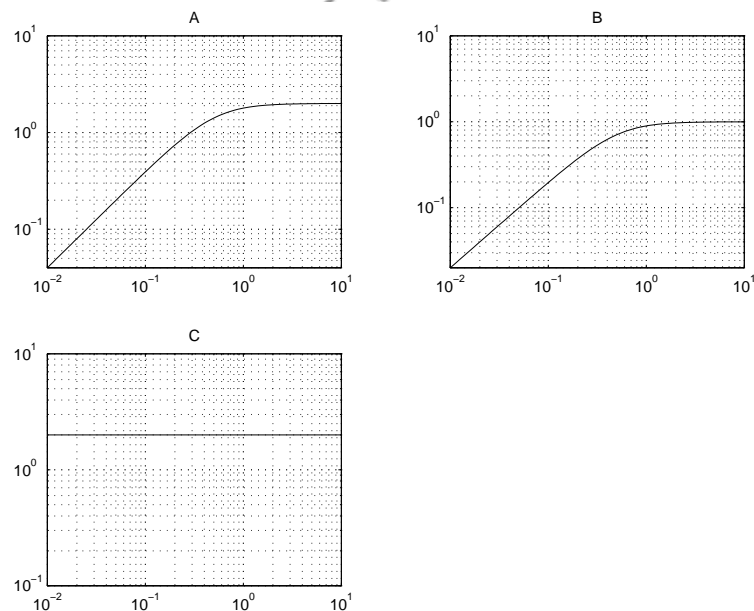


Figure 3: Suggestions for $|W_S^{-1}(i\omega)|$.

Solution

10.6

The system

$$Y(s) = \frac{1}{s+1}U(s)$$

is going to be controlled by the proportional feedback

$$U(s) = K(R(s) - Y(s))$$

- (a) Derive $S(s)$, $T(s)$ and $G_{ru}(s)$ respectively, i.e. the sensitivity function, the complementary sensitivity function and the transfer function from reference to input signal.
- (b) The properties of the control system are specified using the weight function according to

$$|S(i\omega)W_S(i\omega)| < 1 \quad \forall \omega$$

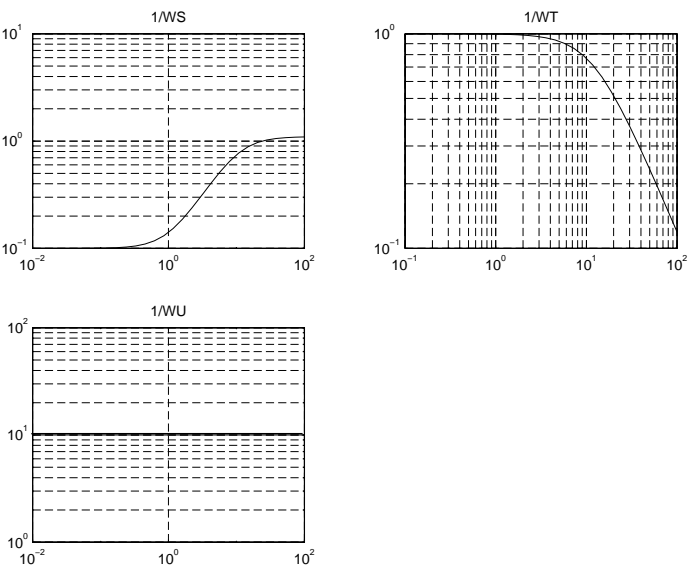
$$|T(i\omega)W_T(i\omega)| < 1 \quad \forall \omega$$

$$|G_{ru}(i\omega)W_u(i\omega)| < 1 \quad \forall \omega$$

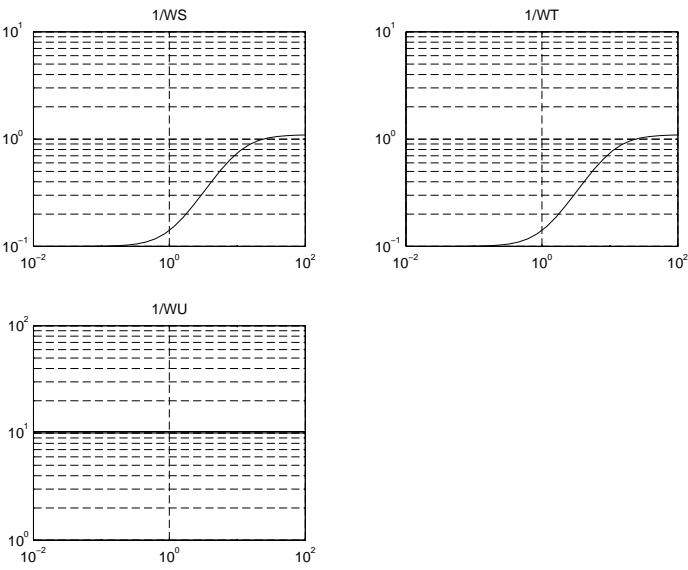
The figures below show three suggestions for weight functions W_S , W_T and W_U . Two of the alternatives are unrealistically or incorrectly specified. Which are the two incorrect alternatives? Motivate the answer.

- (c) Consider the alternative in b) that is realistically specified. Is it possible to choose K such that all requirements are fulfilled?

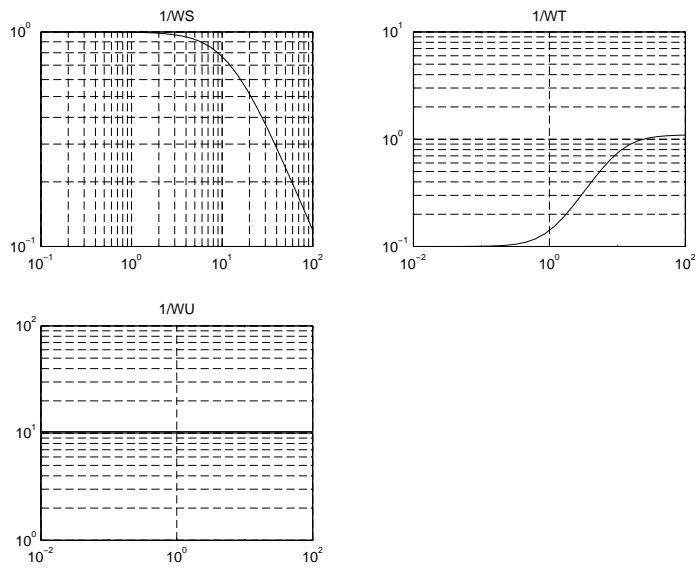
Solution



Figur 4: Alternative I



Figur 5: Alternative II



Figur 6: Alternative III

12 Stability of Nonlinear Systems

12.1

Given the nonlinear differential equation

$$\ddot{y} + 0.2(1 + \dot{y}^2)\dot{y} + y = 0$$

let the state variables be $x_1 = y$ and $x_2 = \dot{y}$. Try to show that the origin is a stable equilibrium by using the Lyapunov function candidate

$$V = \frac{1}{2}(x_1^2 + x_2^2).$$

Solution

12.2

Consider the system

$$\begin{aligned}\dot{x}_1 &= \sin x_1 + x_2^3 \\ \dot{x}_2 &= x_1 - x_2\end{aligned}$$

Is it possible to use the function

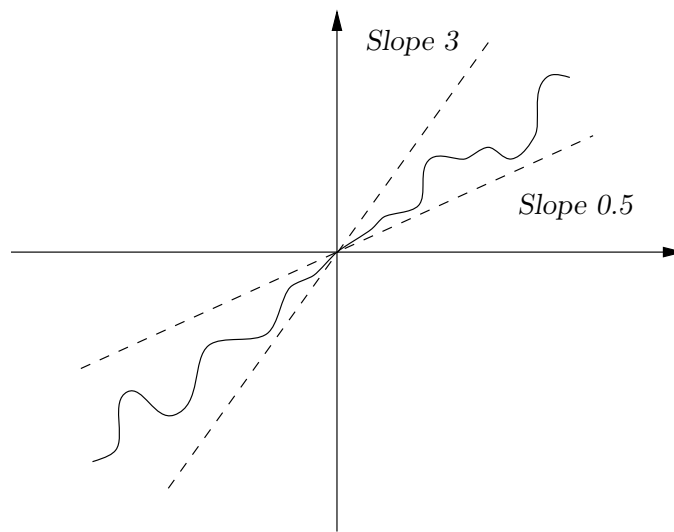
$$V(x_1, x_2) = -\frac{1}{2}x_1^2 + \frac{1}{4}x_2^4$$

to prove Lyapunov stability for the above system? Motivate your answer.

Solution

12.3

A nonlinear function lies in the sector depicted in the figure below

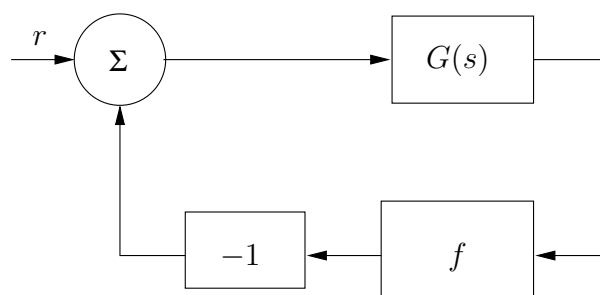


According to the circle criterion, what forbidden circle in the complex plane corresponds to this nonlinearity?

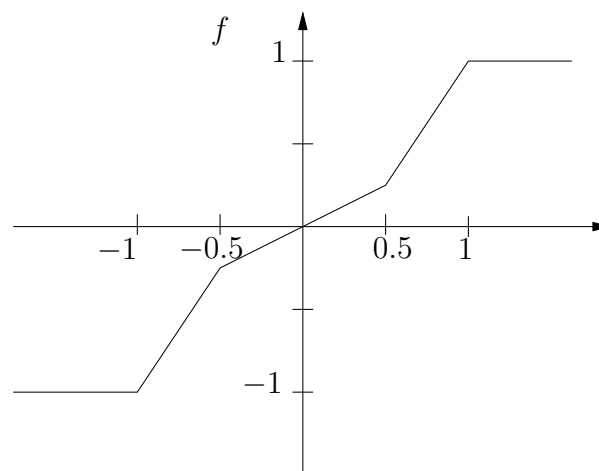
Solution

12.4

A nonlinear system is described by the following block diagram



where $G(s)$ is a linear system and the static nonlinearity f is given in the figure below (the saturations at -1 and 1 extends to $-\infty$ and ∞).

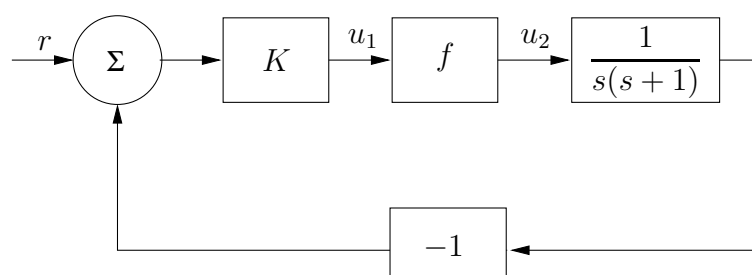


What assumptions on $G(s)$ must be fulfilled in order to prove that the feedback system is stable according to the circle criterion?

Solution

12.5

Consider the system below.

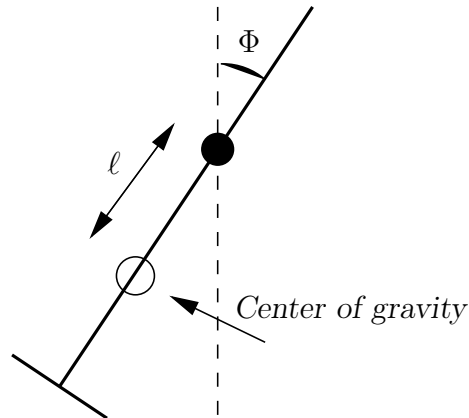


The nonlinearity f is such that u_2 has the same sign as u_1 but is otherwise not known. For what values of $K > 0$ is the feedback system stable according to the circle criterion?

Solution

12.6

Consider the swing depicted below.



The movement of the swing is described by the equation

$$J \frac{d^2 \Phi}{dt^2} + mg\ell \sin \Phi = 0$$

where m is the mass and J is the moment of inertia. The swing can be controlled by alternating between bending and stretching the knees while standing on the swing. The control signal is the location of the center of gravity ℓ . We assume that J is constant.

Show that the control signal

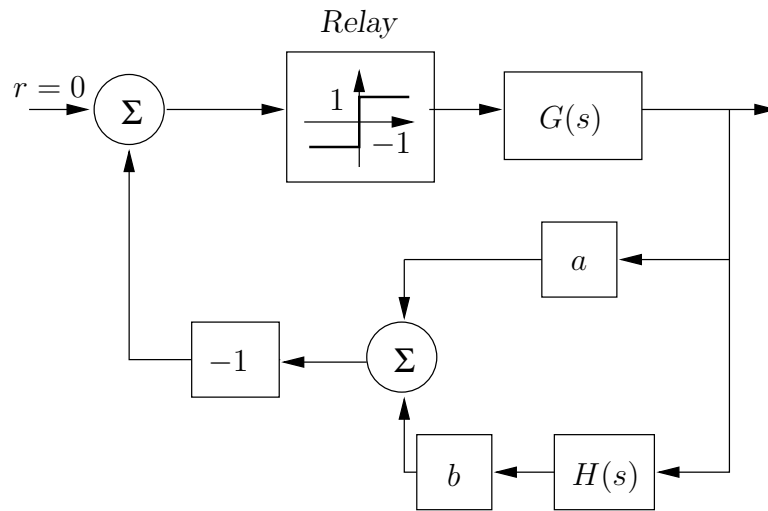
$$\ell = \ell_0 + \varepsilon \Phi \dot{\Phi}, \quad \varepsilon > 0$$

will bring the swing to rest in $\Phi = 0$.

Solution

12.7

The block diagram below is given.



We have that

$$H(s) = s \quad \text{and} \quad G(s) = \frac{1}{(s+1)(s+2)}.$$

How shall the feedback coefficients a and b be chosen to guarantee Lyapunov stability?

Hint: Use a quadratic Lyapunov function candidate.

Solution

12.8

A servo system contains a nonlinearity where the relationship between the input signal u and the output signal y is

$$y = u + \arctan(u)$$

What requirements on the linear part of the servo system must be fulfilled in order to prove stability using the circle criterion? Start by drawing the nonlinearity to get a hint about what you need to know about the arctan function.

Solution**12.9**

A simplified model for the movements of an airplane is given by

$$\dot{x} = \begin{bmatrix} -0.01 & 0.03 & -10 \\ 0 & -1 & 300 \\ 0 & 0 & -0.5 \end{bmatrix} x + \begin{bmatrix} 4 \\ -20 \\ -10 \end{bmatrix} u$$

All states are measured and the control signal is

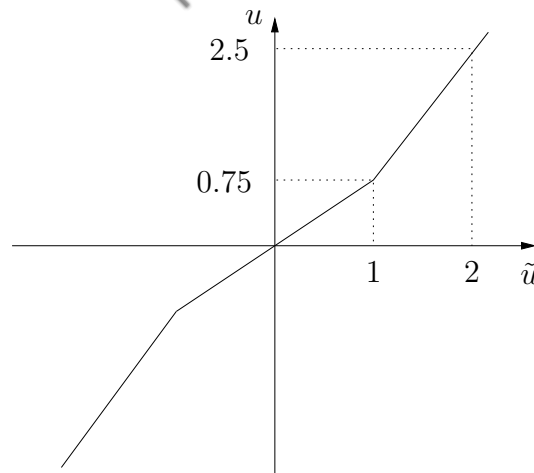
$$\tilde{u} = -Lx + \tilde{r}$$

where L is the feedback that minimizes

$$\int (x^T(t)Q_1x(t) + u^2(t)) dt$$

for $Q_1 = 10 \cdot I$

The requested control signal \tilde{u} is different from the actual u , due to the hydraulic servo dynamics. The relationship between \tilde{u} and u is



Will the closed-loop system be stable? Motivate your answer.

Solution

12.10

The Lorenz system was initially developed to model atmospheric convection and is given by the following three state nonlinear state-space model

$$\frac{dx}{dt} = \sigma(y(t) - x(t))$$

$$\frac{dy}{dt} = x(t)(\rho - z(t)) - y(t)$$

$$\frac{dz}{dt} = x(t)y(t) - \beta z(t)$$

where σ , ρ , and β are positive constants.

- Derive the stationary points.
- Compute the Jacobian of state-space model to derive linearizations in the stationary points.
- Use MATLAB to investigate stability around the stationary points for the case $\sigma = 1$, $\rho = 2$ and $\beta = 3$.

Solution

12.11

Analyze Exercise [1.3](#) using the circle criteria. A numerical study is sufficient.

If you want a challenge, you might want to find all values of interest using analytic computations. To do that, remember that the derivative of a function $\frac{n(\omega)}{d(\omega)}$ is given by $\frac{(\frac{d}{d\omega}n(\omega))d(\omega) - n(\omega)(\frac{d}{d\omega}d(\omega))}{d(\omega)^2}$.

Solution

13 Phase Plane Analysis

13.1

Given the differential equation

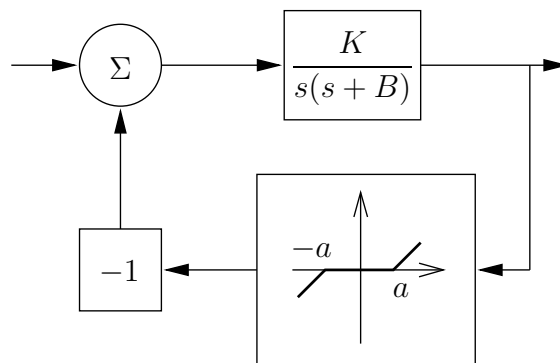
$$\ddot{y} - (0.1 - \frac{10}{3}\dot{y}^2)\dot{y} + y + y^2 = 0$$

- Convert the model to a nonlinear state-space model.
- Find the stationary points.
- Compute linearizations around the stationary points and analyze stability.
- Draw the phase portrait of this system.

Solution

13.2

Draw the phase portrait of the depicted position servo.

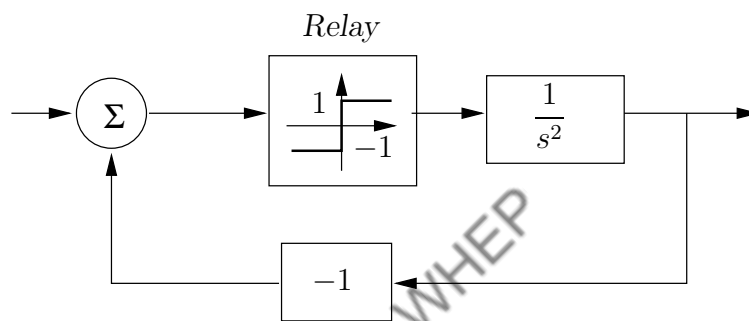


The position is measured using an E-transformer, which can be described as a dead zone. Assume that $K > \frac{B^2}{4}$.

Solution

13.3

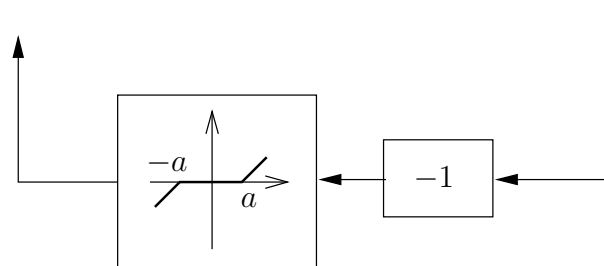
The following system is given



- (a) With zero input signal the output of the relay is $+1$ or -1 , depending on the history of the input signal. The relay does not switch until the input signal has changed polarity.

Draw a phase portrait of the system.

- (b) Due to imperfections the actual feedback loop is

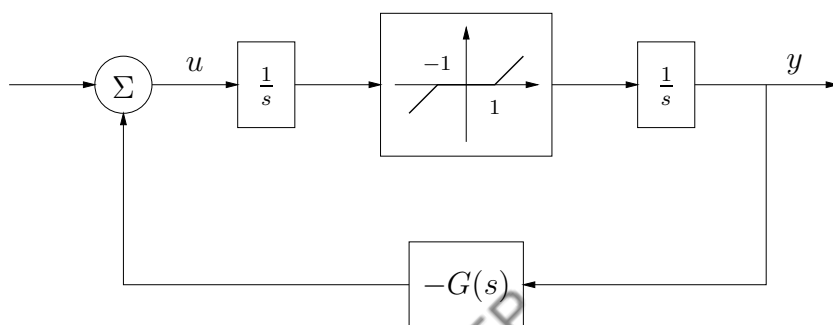


Draw the phase portrait of this system.

Solution

13.4

Linus is on his way home after an exam. On the highway outside of Linköping a gust of wind makes the car drift from the desired path. Your task is to, using phase plane analysis, decide how the movement of the car will progress. Will it return to the desired path? If the car has a constant speed in the direction of travel the system can be described by the following block diagram



The torque applied to the steering wheel is u . The backlash comes from a gear unit in the steering. The output signal y is the deviation from the desired path. $G(s)$ is the transfer function from Linus' visual perception to the torque he applies to the steering wheel.

Distinguish between the cases:

- (a) $G(s) = 1$ (there was a party after the exam)
- (b) $G(s) = 1 + s$ (there was not a party after the exam)

Solution

13.5

A simple ecological system consists of two species of fish. The first kind eats algae and the second kind eats the first kind. Let x_1 denote the number of algae eating fish and x_2 denote the number of predatory fish. Then we have

$$\begin{aligned}\dot{x}_1 &= 2x_1 - \frac{x_1x_2}{1 + \frac{1}{6}x_1} - 0.2x_1^2 \\ \dot{x}_2 &= -3x_2 + \frac{x_1x_2}{1 + \frac{1}{6}x_1}\end{aligned}$$

- (a) From these equations, calculate the stationary points.
- (b) Classify the stationary points and sketch the phase portraits in a surrounding of them. It is sufficient to consider a linearised version of the equations.
- (c) Without any further calculations, merge the phase portraits you have made around the stationary points in a fashion that seems reasonable. Only consider $x_1 > x_2 > 0$.

An interpretation of the given equations is:

If the algae eating fish have an infinite amount of food and lack enemies, their number will grow exponentially as

$$\dot{x}_1 = 2x_1$$

As there is a limited amount of algae the growth saturates according to

$$\dot{x}_1 = 2x_1 - 0.2x_1^2.$$

If there are predatory fish x_2 present the algae eaters will be devoured at the rate

$$\frac{x_1x_2}{1 + \frac{1}{6}x_1}$$

The interpretation of this term is that if x_1 is large every predatory fish can eat until it is full. This corresponds to 6 algae eating fish per time unit. On the other hand, if the number x_1 is relatively small the predatory fish will eat less.

The second equation says that if the supply of food is unlimited ($x_1 = \infty$) the predatory fish will multiply according to

$$\dot{x}_2 = 3x_2.$$

If food is lacking ($x_1 = 0$) the predatory fish will expire as

$$\dot{x}_2 = -3x_2$$

Solution

13.6

A mass is suspended from a spring. Its position $y(t)$ satisfies the differential equation

$$\ddot{y}(t) + y(t) = f(t)$$

where $f(t)$ is an external force acting on the mass. Draw a phase portrait of the system when

$$f(t) = \begin{cases} -1 & \text{if } \dot{y}(t) > 0 \\ +1 & \text{if } \dot{y}(t) < 0 \end{cases}$$

Will the system reach an equilibrium?

Solution

13.7

Consider the system

$$\dot{x} = \begin{pmatrix} -x_1^3 + u \\ x_1 \end{pmatrix}$$

- (a) Sketch a phase portrait when $u = 0$.
- (b) Use the Lyapunov function

$$V(x) = x_1^2 + x_2^2$$

to compute a control signal

$$u = f(x_1, x_2)$$

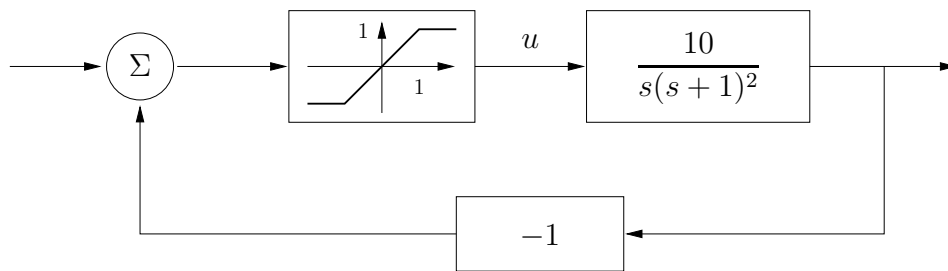
which will make the origin globally asymptotically stable. Sketch a phase portrait, in a neighborhood of the origin, for the closed-loop system.

Solution

14 Oscillations and Describing Functions

14.1

Consider the feedback control system including an input saturation according to the figure below.

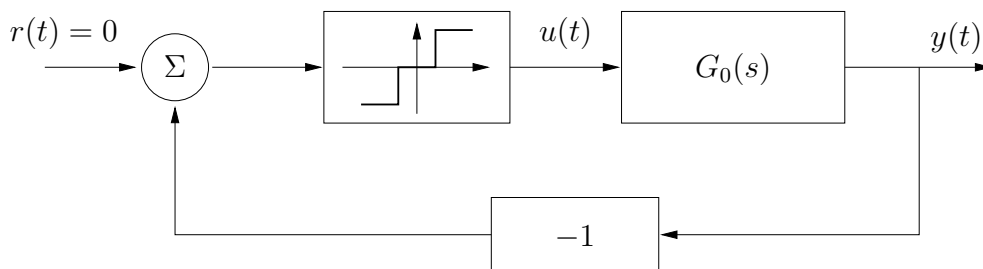


- Investigate stability of the closed loop system using describing functions. If a periodical solution exists, determine its frequency and amplitude.
- Build a simulation model of the control system and investigate the validity of the results from a).

Solution

14.2

A temperature control system, depicted below, contains a relay with dead zone.



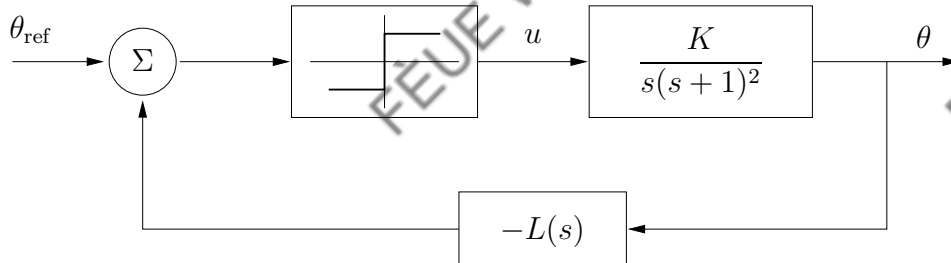
$G_0(s) = \frac{1}{s(1+s)^2}$, $\pm D$ is the width of the dead zone and $\pm H$ is the output level of the relay. The values of the dead zone and output level are such that a stable oscillation just barely can exist. If H is increased or if D is decreased an oscillation will not be possible. The amplitude of the oscillation is 2.5 units. Compute D , H and the frequency of the oscillation. The describing function for a relay with dead zone is

$$\begin{aligned}\operatorname{Re}\{Y_N(C)\} &= \frac{4H}{\pi C} \sqrt{1 - D^2/C^2}, & C \geq D \\ \operatorname{Im}\{Y_N(C)\} &\equiv 0\end{aligned}$$

Solution

14.3

A relay servo is given by

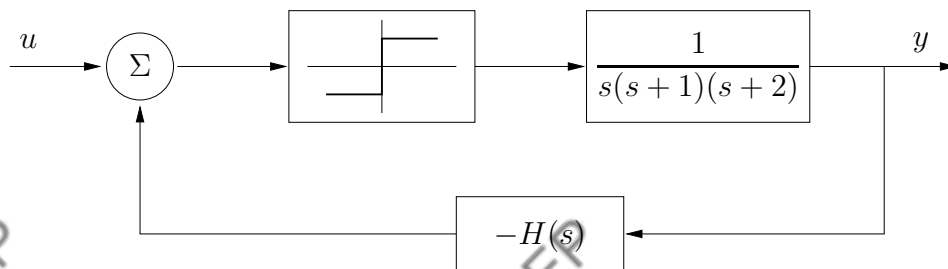


The gain K is strictly positive.

- The feedback used is $L(s) = 1$. Show that there is an oscillation for all values of K .
- To avoid too much wear on the system we do not want the amplitude of the oscillation in θ to be greater than 0.1. For what values of K is this fulfilled?
- We want to use a gain K that is larger than what is possible in (b). State a feedback $L(s)$ with $L(0) = 1$ that makes this feasible. No details are necessary. Just motivate why the feedback should solve the problem.

Solution**14.4**

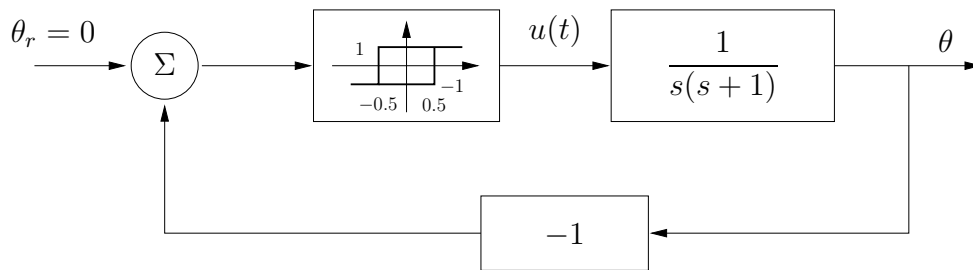
Consider the nonlinear system



- (a) If proportional control is used, i.e. $H(s) = 1$, a stable oscillation occurs. Find the amplitude and frequency of the oscillation.
- (b) To eliminate the oscillation we use proportional and derivative control, i.e. $H(s) = 1 + Ks$. Show how K can be chosen to eliminate the oscillation.

Solution**14.5**

Consider the feedback control system where a motor is controlled using a relay with hysteresis.

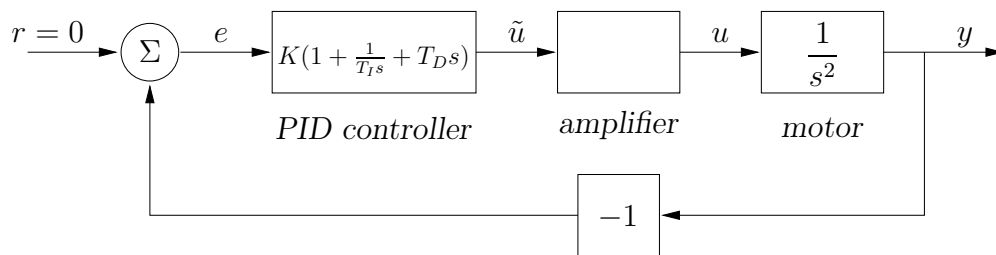


- Investigate the stability of the system using the describing function method. If a periodical solution exists, determine its frequency and amplitude.
- Build a simulation model of the control system and investigate the validity of the results from a).
- Introduce suitable state variables and sketch a phase portrait.

Solution

14.6

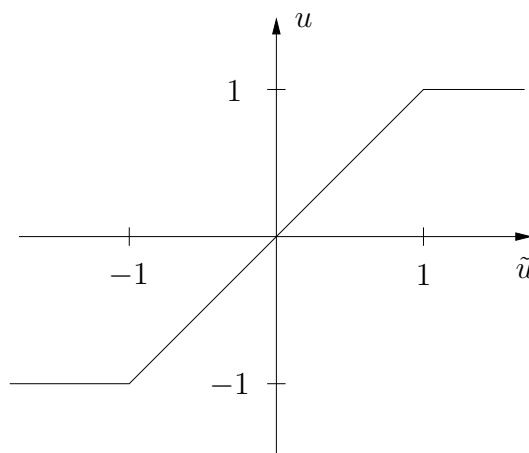
Consider the following servo system



The PID controller has $K = 2$, $T_I = 2$ and $T_D = 0.5$.

- The tuning of the controller was done assuming that the amplifier has the transfer function 1. Show that, if this assumption is true, this results in an asymptotically stable closed-loop system.

(b) The actual amplifier contains a saturation



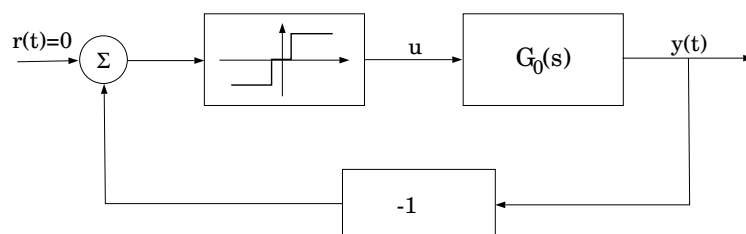
State the amplitude, frequency and stability properties of possible oscillations.

(c) Discuss, based on the results from (b), under what circumstances the servo system will function as intended. Especially investigate the influence of different signal amplitudes.

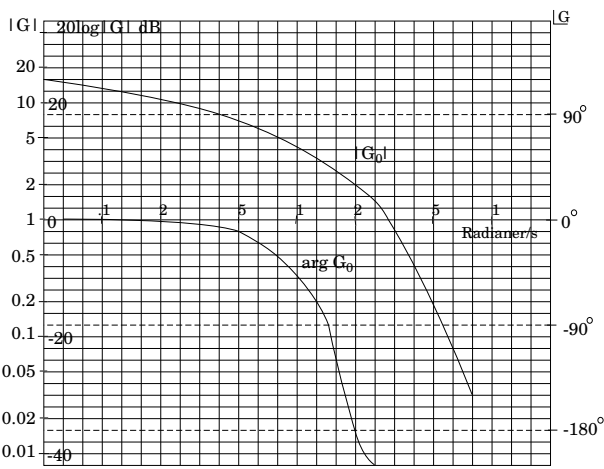
Solution

14.7

Z Consider the feedback control system



When the system is simulated a limit cycle occurs. Determine the amplitude and frequency of the limit cycle. The Bode diagram for the linear part $G_0(s)$ of the control system is given in the figure below.



The describing function of a relay with deadzone is given by

$$Y_f(C) = \frac{4}{\pi C} \sqrt{1 - 1/C^2} \quad C \geq 1$$

Solution

17 To Compensate Exactly for Nonlinearities

17.1

Find a feedback which makes the system

$$\begin{aligned}\dot{x}_1 &= -x_1 + 7x_2 \\ \dot{x}_2 &= -x_2 + \cos x_1 + u\end{aligned}$$

linear.

Solution

17.2

Find an output feedback, $u = f(y)$, which makes the system

$$\begin{aligned}\dot{x}_1 &= x_3 + 8x_2 \\ \dot{x}_2 &= -x_2 + x_3 \\ \dot{x}_3 &= -x_3 + x_1^4 - x_1^2 + u \\ y &= x_1\end{aligned}$$

linear.

Solution

17.3

Find a feedback which makes the system

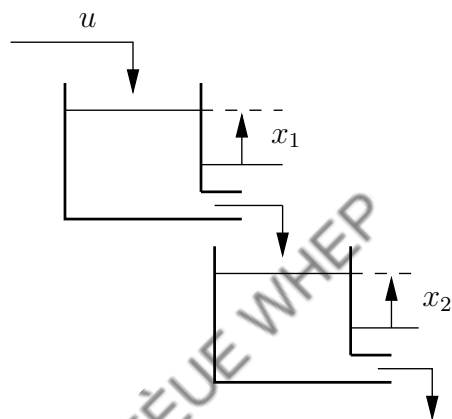
$$\begin{aligned}\dot{x}_1 &= x_1^2 + x_2 \\ \dot{x}_2 &= u \\ y &= x_1\end{aligned}$$

linear.

Solution

17.4

Consider the two tank system:



The dynamics of the system is described by

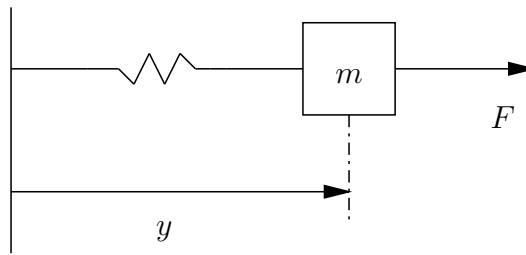
$$\begin{aligned}\dot{x}_1 &= 1 + u - \sqrt{1 + x_1} \\ \dot{x}_2 &= \sqrt{1 + x_1} - \sqrt{1 + x_2}\end{aligned}$$

Which state should be chosen as output to achieve a strong relative degree 2? Do a feedback linearization of the system.

Solution

17.5

A mass m is suspended from a spring:



The force F is generated by the control signal u fed through an actuator such that

$$F = \frac{1}{s+1}u$$

The position of the mass is y . The spring rate and the viscous damping are nonlinear. Thus the force is

$$-k(y) - d(\dot{y}).$$

- (a) Realize this system on state-space form. The input signal is u and the output signal is y .
- (b) Can the system in (a) be made linear using feedback? If so, compute such a feedback.

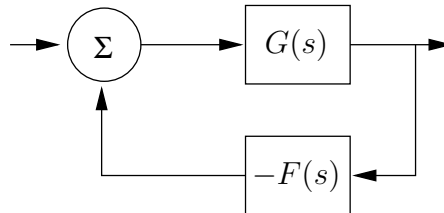
Solution

Solutions

1 Introduction

1.1

The small gain theorem for linear systems can be stated as follows: Assume that both $G(s)$ and $F(s)$ are stable transfer functions, and interconnected according to the figure below.



Then the closed-loop system is stable if

$$|G(i\omega)| \cdot |F(i\omega)| < 1, \quad \forall \omega.$$

The transfer function of the closed-loop system is

$$G_c(s) = \frac{G(s)}{1 + G(s)F(s)}$$

$G_c(s)$ is stable according to the Nyquist criterion if the Nyquist curve for $G(i\omega)F(i\omega)$ does *not* encircle the point -1 . Since we know from the small gain theorem that

$$|G(i\omega)F(i\omega)| \leq |G(i\omega)| \cdot |F(i\omega)| < 1,$$

the Nyquist curve can not encircle the point -1 , and hence the Nyquist criterion is fulfilled.

Note that input-output stability follows from asymptotic stability. Input-output stability is the concept used in the general small gain theorem.

Go back

1.2

We have that $y(t) = f(u(t))$ where $f(\cdot)$ is the function describing the ideal relay. The gain is defined as

$$\|f\| = \sup_{u \neq 0} \frac{\|y\|_2}{\|u\|_2}$$

We have that $|f(u)| \equiv 1$, $\forall u(t) \neq 0$, and this yields

$$\|y\|_2^2 = \int_{-\infty}^{\infty} y^2(t) dt = \lim_{T \rightarrow \infty} \int_{-T}^T [f(u(t))]^2 dt = \lim_{T \rightarrow \infty} 2T = \infty$$

for *all* choices of $u(t) \neq 0$ such that $0 < \|u\|_2 < \infty$. Take for example $u(t) = \frac{1}{t}$. This means that an ideal relay has *infinite* gain.

Go back

1.3

The closed-loop system is stable if the product of gains in the loop is smaller than 1. To obtain a notation similar to the book, we can define the nonlinear part as S_1 and move the gain K into the linear system and define S_2 as the system $KG(s)$. The small gain theorem requires $\|S_1\| \cdot \|S_2\| < 1$.

$$\text{We have that: } \begin{cases} \|S_1\| = \|f(u)\| \\ \|S_2\| = \|KG\| = |K| \|G\| = |K| \sup_{\omega} |G(i\omega)| \end{cases}$$

A Bode plot reveals that the amplitude gain of G is monotonically decreasing, which indeed is what we get analytically

$$|G(i\omega)| = \frac{1}{|(1+i)^2|} = \frac{1}{|1+i|^2} = \frac{1}{\sqrt{\omega^2+1}^2} = \frac{1}{\omega^2+1}$$

and consequently the largest value is 1 and thus $\|G\| = 1$.

To compute the gain of the nonlinear part, we use the definition and upper bound the nonlinearity

$$\begin{aligned}\|f(u)\|^2 &= \frac{\|\tilde{u}\|_2^2}{\|u\|_2^2} = \frac{\int_{-\infty}^{\infty} (f(u(t)))^2 dt}{\|u\|_2^2} \leq \left[|f(u(t))| \leq \frac{1}{2}|u(t)| \right] \\ &\leq \frac{\frac{1}{4}\|u\|_2^2}{\|u\|_2^2} = \frac{1}{4} \quad \Rightarrow \quad \|f(u)\| \leq \frac{1}{2}\end{aligned}$$

This is a natural result. The largest gain is obtained when the nonlinearity acts linearly, and the gain is then 1/2. For larger and smaller inputs, the saturation outputs a signal with amplitude smaller than 1/2 of the input.

$$\|S_1\| \cdot \|S_2\| \leq \frac{1}{2} \cdot |K| < 1$$

i.e. , we must choose $|K| < 2$ to be able to guarantee input-output stability.

Go back

1.4

- (a) $\|y\|_{\infty} = |a|, \quad \|y\|_2 = \infty$
- (b) $\|y\|_{\infty} = 1, \quad \|y\|_2 = 1$
- (c) $\|y\|_{\infty} = \frac{1}{4}, \quad \|y\|_2 = \frac{1}{6}\sqrt{3}$

Go back

1.5

The gain of the system is

$$\|G\| = \sup_{\omega} |G(i\omega)| = \sup_{\omega} \frac{\omega_0^2}{\sqrt{(\omega_0^2 - \omega^2)^2 + 4\zeta^2\omega_0^2\omega^2}}$$

By differentiating $|G(i\omega)|$ we see that the magnitude of $G(i\omega)$ has its maximum at $\omega = 0$ if $\zeta > \frac{1}{\sqrt{2}}$. This results in the gain

$$\|G\| = 1$$

If $0 < \zeta < \frac{1}{\sqrt{2}}$ the maximum of $|G(i\omega)|$ is attained at $\omega = \omega_0 \sqrt{1 - 2\zeta^2}$. This results in the gain

$$\|G\| = \frac{1}{2\zeta \sqrt{1 - \zeta^2}}$$

Go back

1.6

The closed-loop system from input (call it r) to the output (call it y) is given by $Y(s) = \frac{a}{s+a}U(s) = G(s)U(s)$, $U(s) = R(s) - KY(s)$ and thus $Y(s) = \frac{G(s)}{1+KG(s)}R(s)$ which simplifies to $\frac{a}{s+a+Ka}R(s)$. The characteristic equation of the closed-loop system is thus given by

$$s + a + Ka = 0$$

which implies the pole $s = -(K + 1)a$.

One has to distinguish between the cases $a > 0$ and $a < 0$ respectively.

- (i) For $a > 0$ the system $G(s)$ is stable and the small gain theorem is applicable. The system $G(s)$ has gain one ($|G(i\omega)|$ is decreasing with $G(0) = 1$), and the small gain theorem hence gives $1 \cdot |K| < 1$.

For positive a the pole is in the negative left half-plane for $K > -1$. Hence the small-gain theorem is conservative.

- (ii) For $a < 0$ the system $G(s)$ is not stable and the small gain theorem is not applicable. The pole of the closed-loop is in the left half-plane if $K < -1$.

Go back

1.7

The linear part, represented by $G(s)$, has gain 1.5 according to the figure. For the nonlinear part we assume that $f(x)$ is an odd function, such that $f(-x) = -f(x)$. The nonlinearity can hence be bounded by

$$|f(x)| \leq 0.5 |x|$$

and hence the gain is 0.5. Since $1.5 \cdot 0.5 < 1$ the closed loop system is stable according to the small gain theorem.

[Go back](#)

1.8

Using proportional control $u = -Ky = -Kx_1$ we get

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= -aKx_1 - ax_2\end{aligned}$$

The characteristic equation is $s^2 + as + aK = s^2 + (1 + \rho)s + (1 + \rho)K = 0$. The closed-loop system is stable if $(1 + \rho) > 0$ and $(1 + \rho)K > 0$, i.e. it is stable for all $K > 0$ when $\delta < 1$.

[Go back](#)

1.9

(a) With $a = 1 + \rho$ it holds that

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= -a(Kx_1 + x_2) = -Kx_1 - x_2 - \underbrace{\rho(Kx_1 + x_2)}_{w=\rho z}\end{aligned}$$

The open system with input signal w and output signal z is given by

$$\begin{aligned}\dot{x} &= \begin{pmatrix} 0 & 1 \\ -K & -1 \end{pmatrix} x + \begin{pmatrix} 0 \\ -1 \end{pmatrix} w \\ z &= (K \quad 1) x,\end{aligned}$$

that is

$$G_{wz}(s) = -\frac{s+K}{s^2+s+K}.$$

(b) The relationship

$$w(t) = \rho(t)z(t)$$

gives

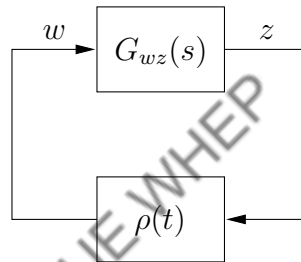
$$\|w(t)\|_2^2 = \int_{-\infty}^{\infty} \rho^2(t)z^2(t)dt < \delta^2 \|z(t)\|_2^2$$

and

$$\frac{\|w(t)\|_2}{\|z(t)\|_2} < \delta$$

which implies that the gain is at most δ .

(c) The closed-loop system is depicted in the figure below.



A sufficient condition for stability, according to the low gain theorem, is

$$\|G_{wz}\| \cdot \|\rho\| < 1 \quad \Leftrightarrow \quad \|G_{wz}\|^2 < 1/\delta^2$$

The magnitude, of the linear systems frequency response, squared is

$$|G_{wz}(i\omega)|^2 = \frac{\omega^2 + K^2}{(K - \omega^2)^2 + \omega^2}$$

What is the maximum of this function?

$$\sup_x \frac{x + K^2}{(K - x)^2 + x} = \frac{\sqrt{K^4 + 2K^3}}{2K^4 + 4K^3 - (2K + 2K^2 - 1)\sqrt{K^4 + 2K^3}}$$

for $x = -K^2 + \sqrt{K^4 + 2K^3}$. Thus, an implicit condition on K to guarantee stability of the closed-loop system is

$$\frac{\sqrt{K^4 + 2K^3}}{2K^4 + 4K^3 - (2K + 2K^2 - 1)\sqrt{K^4 + 2K^3}} < 1/\delta^2$$

Go back

1.10

Since the poles of the loop gain $G_o(s) = K \frac{1}{(s+1)^3} = KG(i\omega)$ has no poles in the right half-plane, we can use the simplified nyquist criteria, which says that the loop gain cannot encircle -1 for the closed-loop to remain stable. Plotting the nyquist curve $G(i\omega)$ shows that it satisfies this for $K = 1$ but a sufficiently large K will scale the curve $KG(i\omega)$ large enough to enclose the critical point.

```
>> s = tf('s');
>> G = 1/(s+1)^3;
>> nyquist(G,2*G,10*G);
```

To find the critical value of K analytically, we much find the frequency w_p for which $G(i\omega)$ it crosses the negative real line, and then by knowing the value of $G(i\omega_p)$ find the value of K such that the crossing is moved sufficiently far to the left leading to $KG(i\omega_p) < -1$ thus enclosing -1 indicating instability. This can be solved by either solving $\text{Im } G(i\omega_p) = 0$ or $\arg G(i\omega_p) = -\pi$.

Using the argument, we have $\arg \frac{1}{(i\omega+1)^3} = \arg 1 - \arg(i\omega+1)^3 = -3 \arg(i\omega+1)$ and we thus have $-3 \arctan\left(\frac{\omega}{1}\right) = -\pi$ leading to $\omega_p = \tan(\pi/3) = \sqrt{3}$.

From this we can compute the value of $G(i\omega_p)$ at the crossing. The long cumbersome way is $G(i\omega_p) = \frac{1}{(\sqrt{3}i+1)^3} = \frac{(-\sqrt{3}i+1)^3}{(|\sqrt{3}i+1|^2)^3} = \frac{(-\sqrt{3}i+1)^3}{(3+1)^3} \dots = -\frac{1}{8}$. The fast way to do it is to realize that $G(i\omega_p)$ is a negative real number, and therefor $G(i\omega_p) = -|G(i\omega_p)| = -\frac{1}{|(\sqrt{3}i+1)^3|} = -\frac{1}{|\sqrt{3}i+1|^3} = -\frac{1}{\sqrt{3^2+1^2}^3} = -\frac{1}{8}$.

Having the crossing of the real line at $-\frac{1}{8}$ it follows that $K > 8$ will lead to encircling, and thus instability of the closed-loop system.

We can verify in MATLAB that the closed-loop system has unstable poles for $K > 8$ while smaller values of the feedback gain leads to stability.

```
>> pole(feedback(G*7))
>> pole(feedback(G*9))
```

The phase ω_p is called the *phase crossover frequency* (frequency at which phase drops below -180°) and the value $1/|G(i\omega_p)|$ the *amplitude margin*. These values are studied in Bode plots when simplified stability analysis is done via phase margin/amplitude margin/cross over frequency/phase cross over frequency, which thus is an application of the simplified nyquist criteria.

```
>> bode(G)
>> margin(G)
```

Go back

1.11

The closed-loop system from input (call it r) to the output (call it y) is given by $Y(s) = \frac{a}{s+a}U(s) = G(s)U(s)$, $U(s) = R(s) - KY(s)$ and thus $Y(s) = \frac{G(s)}{1+KG(s)}R(s)$ which simplifies to $\frac{a}{s+a+Ka}R(s)$. The characteristic equation of the closed-loop system is thus given by

$$s + a + Ka = 0$$

which means the closed-loop pole is $s = -(K+1)a$. Hence, for closed loop stability we require $(K+1)a > 0$.

If a is positive (i.e. the open-loop system is stable) the pole is in the negative left half plane for $K > -1$. If a is negative (i.e. the controlled system is unstable) the pole is in the negative left half plane for $K < -1$.

Using MATLAB we can plot the Nyquist curve for various values of a to realize that $\frac{a}{s+a}$ always generates a circle in the right half-plane, touching the points 0 and 1. The only difference is the directions of the arrows, which will be important later.

```
>> s = tf('s');
>> subplot(1,2,1);nyquist(1/(s+1));
>> subplot(1,2,2);nyquist(-3/(s+3));
```

The nyquist curve $K \frac{a}{i\omega + a}$ which has to be analyzed for closed-loop stability is thus a scaled circle.

When a is positive, the open loop system is stable, and we can use the simplified Nyquist criterion which says that the point -1 should not be encircled. When $K > 0$ the Nyquist curve will simply be scaled but remain in the right half plane, and thus certify stability of the closed-loop. When $K < 0$ it will not only be scaled, but also flipped into the left half-plane. If $K < -1$ it will thus be flipped and large enough to encircle -1 indicating instability. Hence stability can be guaranteed for $K > -1$ for the case $a > 0$.

When a is negative, the open loop system is unstable, and the general Nyquist criterion now says we *must* encircle the point -1 once to guarantee closed loop stability (see below). Following the same reasoning as above, this can never happen for $K > 0$, but requires a $K < -1$.

To conclude, we arrive a precisely the same results as an analysis of the closed loop poles does, as expected.

Finally, a word on the exact Nyquist result which says that with P_o unstable poles in the open loop system, and P_s unstable poles in the closed-loop system, it holds that $P_s - P_o$ is equal to the number of times $G(i\omega)$ circles around the point -1, when the key-hole contour is traversed counter-clockwise. Unfortunately, this is in conflict with MATLABs notation which draws the arrows on the nyquist curve in the opposite direction. Hence, when counting the argument change (thus encirclements) you must count the number of times you encircle -1 with counter-clockwise being positive direction, going in the *opposite direction of the arrow*.

For instance, we just learned that $K = -4$ will stabilize an unstable system $\frac{-3}{s+3}$. Draw the nyquist plot of the loop gain for a stabilizing case such as $K = -4$ with a pole in -3

```
>> nyquist(-4*-3/(s+3));
```

It shows that the curve encircles -1 once in the positive counter-clockwise direction following the arrows, but we must switch the direction when counting and thus note that we make -1 encirclements. This is consistent with the

criteria which becomes $(0 \text{ unstable closed-loop poles}) - (1 \text{ unstable open loop poles}) = (-1 \text{ encirclements})$

Go back

1.12

- a) With loop gain $G_0(s) = \alpha \frac{2s-1}{s^2-s+4}$ the closed-loop is given by $G_c(s) = \frac{G_o(s)}{1+G_o(s)} = \frac{\alpha(2s-1)}{s^2-s+4+K(2s-1)} = \frac{\alpha(2s-1)}{s^2+(2\alpha-1)s+4-\alpha}$. The closed loop is thus stable for $1/2 < \alpha < 4$
- b) We draw the nyquist curve for $\alpha = 1$. The poles for the open-loop system are given by $\frac{1}{2} \pm i\frac{\sqrt{15}}{2}$, i.e., there are $P_o = 2$ open-loop unstable poles

```
>> s = tf('s');
>> G = (2*s-1)/(s^2-s+4)
>> pole(G)
>> nyquist(G)
```

Following the curve (opposite direction to the MATLAB arrows!) and measuring the angle to -1 (counter-clockwise positive!), we end up with a total argument change of -720° , so -2 encirclements. Consequently, the number of unstable poles for the case $\alpha = 1$ is given by $P_s = 2 + (-2) = 0$. The closed-loop system is stable.

The nyquist curve crosses the real axis at -2 and -0.25, and thus $\alpha < 0.25$ and $\alpha > 2$ will change in what way the point -1 is encircled by the curve. For $\alpha < 1/2$ the point -1 will be outside the curve completely, meaning there are 0 encirclements, leading to instability as $P_s = 2+0 = 2$, while a large $\alpha > 4$ will lead to a situation where we count -360° argument change and thus $P_s = 2 + -1 = 1$. We can confirm the unstable pole count numerically, or by hand

```
>> pole(feedback(3*G,1))
>> pole(feedback(0.2*G,1))
```

- c) To draw the root-loci for the closed-loop characteristic equation $s^2 - s + 4 + \alpha(2s - 1) = 0$ we identify it as $P(s) + \alpha Q(s)$ and note that the command in MATLAB wants the object $\frac{Q(s)}{P(s)}$ which is nothing but our system $G(s)$. The root loci tells us there are two unstable starting point (i.e. unstable for small α) and then it moves into the left half-plane indicating a stable situation, but for large α a pole moves out into the right half-plane again. Exactly what we have found previously.

```
>> rlocus(G)
```

[Go back](#)

2 Representation of Linear Systems

2.1

Putting $x = \omega, u_1 = M, u_2 = I_m, u_3 = R, y_1 = \omega$ and $y_2 = e$ give the state equation

$$\dot{x} = u_1 - \frac{x^2 u_2}{u_3}$$

and

$$y_1 = x$$

$$y_2 = u_2 x$$

The input vector $u_{1,0} = u_{2,0} = u_{3,0} = 1$ and the state x_0 is a stationary point, giving the stationary output $y_{1,0} = y_{2,0} = 1$. Introducing

$$\Delta x = x - x_0 \quad \Delta u = u - u_0 \quad \Delta y = y - y_0$$

and linearizing gives

$$\frac{d}{dt} \Delta x = -2 \Delta x + (1 \quad -1 \quad 1) \Delta u$$

$$\Delta y = \begin{pmatrix} 1 \\ 1 \end{pmatrix} \Delta x + \begin{pmatrix} 0 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix} \Delta u$$

Computing the transfer function and renaming the variables give

$$\begin{pmatrix} \Delta \omega \\ \Delta e \end{pmatrix} = \frac{1}{s+2} \begin{pmatrix} 1 & -1 & 1 \\ 1 & s+1 & 1 \end{pmatrix} \begin{pmatrix} \Delta M \\ \Delta I_m \\ \Delta R \end{pmatrix}$$

[Go back](#)

2.2

(a)

$$y = \alpha h_2, \quad f = \beta(h_1 - h_2)$$

$$\dot{h}_1 = \frac{1}{A_1}(u_1 - f), \quad \dot{h}_2 = \frac{1}{A_2}(u_2 + f - y)$$

$$\dot{h} = \begin{pmatrix} -\frac{1}{A_1}\beta & \frac{1}{A_1}\beta \\ \frac{1}{A_2}\beta & -\frac{1}{A_2}(\beta + \alpha) \end{pmatrix} h + \begin{pmatrix} \frac{1}{A_1} & 0 \\ 0 & \frac{1}{A_2} \end{pmatrix} u$$

$$y = (0 \quad \alpha) h$$

$$G(s) = \frac{1}{s^2 + (2\beta + \alpha)s + \alpha\beta} \begin{pmatrix} \alpha\beta & \alpha(s + \beta) \end{pmatrix}$$

(b) The result above gives

$$G(0) = \begin{pmatrix} 1 & 1 \end{pmatrix}$$

Singular values in $\omega = 0$ (i.e., for constant input signals) are given by the square roots of the largest and smallest eigenvalues of the matrix

$$G(0)^T G(0) = \begin{pmatrix} 1 \\ 1 \end{pmatrix} \begin{pmatrix} 1 & 1 \end{pmatrix} = \begin{pmatrix} 1 & 1 \\ 1 & 1 \end{pmatrix}$$

Solving $\det(\lambda I - G(0)^T G(0)) = 0$ yields eigenvalues 0 and 2, which implies

$$\bar{\sigma}(G(0)) = \sqrt{2} \quad \underline{\sigma}(G(0)) = 0$$

The maximum singular value corresponds the input signal vector

$$u_{max} = \begin{pmatrix} 1 \\ 1 \end{pmatrix}$$

which gives the steady state output signal $y = 2$, while the minimum singular value corresponds the input signal vector

$$u_{min} = \begin{pmatrix} 1 \\ -1 \end{pmatrix}$$

which gives the steady state output signal $y = 0$.

Go back

2.3

Find the common denominator of the system

$$\begin{aligned} Y(s) &= \frac{(s^2 + s + 1)}{(s + 1)(s + 2)(s^2 + s + 1)} U_1(s) + \frac{(s + 2)(s + 3)}{(s + 1)(s + 2)(s^2 + s + 1)} U_2(s) \\ &= \frac{(s^2 + s + 1)}{(s^4 + 4s^3 + 6s^2 + 5s + 2)} U_1(s) + \frac{(s^2 + 5s + 6)}{(s^4 + 4s^3 + 6s^2 + 5s + 2)} U_2(s) \end{aligned}$$

It is now straightforward to realize the system on observer canonical form

$$\begin{aligned} \dot{x}(t) &= \begin{pmatrix} -4 & 1 & 0 & 0 \\ -6 & 0 & 1 & 0 \\ -5 & 0 & 0 & 1 \\ -2 & 0 & 0 & 0 \end{pmatrix} x(t) + \begin{pmatrix} 0 & 0 \\ 1 & 1 \\ 1 & 5 \\ 1 & 6 \end{pmatrix} u(t) \\ y(t) &= (1 \ 0 \ 0 \ 0) x(t) \end{aligned}$$

[Go back](#)

2.4

Find the common denominator

$$y(t) = \frac{1}{p^3 + 7p^2 + 16p + 12} \begin{pmatrix} p^2 + 3p & p^2 + p - 2 \end{pmatrix} \begin{pmatrix} u_1(t) \\ u_2(t) \end{pmatrix}$$

Observer canonical form yields

$$\begin{aligned} \dot{x}(t) &= \begin{pmatrix} -7 & 1 & 0 \\ -16 & 0 & 1 \\ -12 & 0 & 0 \end{pmatrix} x(t) + \begin{pmatrix} 1 & 1 \\ 3 & 1 \\ 0 & -2 \end{pmatrix} \begin{pmatrix} u_1(t) \\ u_2(t) \end{pmatrix} \\ y(t) &= (1 \ 0 \ 0) x(t) \end{aligned}$$

[Go back](#)

2.5

Laplace transformation yields

$$Y(s) = \frac{(b_{11}s + b_{12})}{(s^2 + a_1s + a_2)}U_1(s) + \frac{(b_{21}s + b_{22})}{(s^2 + a_1s + a_2)}U_2(s)$$

The system on observer canonical form is

$$\begin{aligned}\dot{x}(t) &= \begin{pmatrix} -a_1 & 1 \\ -a_2 & 0 \end{pmatrix} x(t) + \begin{pmatrix} b_{11} & b_{21} \\ b_{12} & b_{22} \end{pmatrix} u(t) \\ y(t) &= \begin{pmatrix} 1 & 0 \end{pmatrix} x(t)\end{aligned}$$

[Go back](#)

2.6

Laplace transformation yields

$$A(s)Y(s) = B(s)U(s)$$

where

$$A(s) = \begin{pmatrix} s & 1 \\ 1 & (s+1) \end{pmatrix} \quad B(s) = \begin{pmatrix} s+2 \\ 1 \end{pmatrix}$$

Multiplication by $A^{-1}(s)$ results in

$$Y(s) = A^{-1}(s)B(s)U(s)$$

i.e.

$$Y(s) = \begin{pmatrix} \frac{s^2+3s+1}{s^2+s-1} \\ \frac{-2}{s^2+s-1} \end{pmatrix} U(s) = \begin{pmatrix} \frac{2s+2}{s^2+s-1} + 1 \\ \frac{-2}{s^2+s-1} \end{pmatrix} U(s)$$

The system on controller canonical form

$$\begin{aligned}\dot{x}(t) &= \begin{pmatrix} -1 & 1 \\ 1 & 0 \end{pmatrix} x(t) + \begin{pmatrix} 1 \\ 0 \end{pmatrix} u(t) \\ y(t) &= \begin{pmatrix} 2 & 2 \\ 0 & -2 \end{pmatrix} x(t) + \begin{pmatrix} 1 \\ 0 \end{pmatrix} u(t)\end{aligned}$$

[Go back](#)

2.7

- (a) We have $p = 2$ outputs with $y = \begin{pmatrix} y_1 \\ y_2 \end{pmatrix}$ and $m = 2$ inputs $u = \begin{pmatrix} u_1 \\ u_2 \end{pmatrix}$.
Hence $G(s)$ must have dimension 2×2 and to obtain the wanted definitions it must be $G(s) = \begin{pmatrix} \frac{1}{s+1} & 0 \\ 0 & \frac{1}{s+2} \end{pmatrix}$ since $Y_1(s) = 1 \cdot \frac{1}{s+1} U_1(s) + 0 \cdot U_2(s)$ and $Y_2(s) = 0 \cdot U_1(s) + 1 \cdot U_2(s)$.
- (b) We have $p = 1$ output with trivially $y = (y_1)$ and $m = 2$ inputs $u = \begin{pmatrix} u_1 \\ u_2 \end{pmatrix}$. Hence $G(s)$ must have dimension 1×2 be $G(s) = \begin{pmatrix} \frac{1}{s+1} & \frac{1}{s+2} \end{pmatrix}$.
- (c) We have $p = 2$ outputs with $y = \begin{pmatrix} y_1 \\ y_2 \end{pmatrix}$ and $m = 1$ inputs $u = (u_1)$.
Hence $G(s)$ must have dimension 2×1 and be $G(s) = \begin{pmatrix} \frac{1}{s+1} \\ \frac{1}{s+2} \end{pmatrix}$.

Go back

2.8

- (a) The model is already in a commonly used notation in terms of names.

With states $x = \begin{pmatrix} x_1 \\ x_2 \end{pmatrix}$, inputs $u = \begin{pmatrix} u_1 \\ u_2 \\ u_3 \end{pmatrix}$ and outputs $y = \begin{pmatrix} y_1 \\ y_2 \end{pmatrix}$, we arrive at

$$\begin{aligned} \dot{x} &= \begin{pmatrix} -1 & 2 \\ 0 & 0 \end{pmatrix} x + \begin{pmatrix} 0 & 3 & 4 \\ 5 & 0 & 0 \end{pmatrix} u \\ y &= \begin{pmatrix} 6 & 7 \\ 0 & 8 \end{pmatrix} x + \begin{pmatrix} 0 & 0 & 0 \\ 9 & 10 & 11 \end{pmatrix} u \end{aligned}$$

Data representing the system can be gathered as $\left[\begin{array}{cc|ccc} -1 & 2 & 0 & 3 & 4 \\ 0 & 0 & 5 & 0 & 0 \\ \hline 6 & 7 & 0 & 0 & 0 \\ 0 & 8 & 9 & 10 & 11 \end{array} \right]$

- (b) The only differential to be found is \dot{v} , hence this is the only state. In the differential equation for \dot{v} there are two other variables, which thus must be external inputs. There are no other variables to be found, hence all inputs are gathered in, e.g., $f = \begin{pmatrix} u \\ w \end{pmatrix}$ (we cannot call the input in this system u , as that name is taken by a particular element of the input vector). There is only one output which happens to be called y . Our model is

$$\begin{aligned}\dot{v} &= v + \begin{pmatrix} 2 & 1 \end{pmatrix} f \\ y &= 3v + \begin{pmatrix} 1 & -1 \end{pmatrix} f\end{aligned}$$

The data representing the system is thus $\left[\begin{array}{c|cc} 1 & 2 & 1 \\ \hline 3 & 1 & -1 \end{array} \right]$

- (c) The states are $p = \begin{pmatrix} y \\ x \\ q \end{pmatrix}$, the inputs are $w = \begin{pmatrix} u \\ v \\ h \end{pmatrix}$ and the output is z .

$$\begin{aligned}\dot{p} &= \begin{pmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix} p + \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix} w \\ z &= \begin{pmatrix} 1 & 2 & 3 \end{pmatrix} p + \begin{pmatrix} 0 & 0 & 0 \end{pmatrix} w\end{aligned}$$

All data can be collected as $\left[\begin{array}{ccc|ccc} 0 & 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 \\ \hline 1 & 2 & 3 & 0 & 0 & 0 \end{array} \right]$

[Go back](#)

2.9

Let n , p and m denote the dimensions of the vectors x , y and u . Graphically, the following algebra must hold

$$\begin{bmatrix} \dot{x}^{n \times 1} \\ y^{p \times 1} \end{bmatrix} = \left[\begin{array}{c|c} A^{n \times n} & B^{n \times m} \\ \hline C^{p \times n} & D^{p \times m} \end{array} \right] \begin{bmatrix} x^{n \times 1} \\ u^{m \times 1} \end{bmatrix}$$

The width of B and D must be the height of u as they are multiplied hence $m = 4$. The height of C and D must be the height of y which is $p = 2$. The number of states n must match the height of A so $n = 3$. The matrix A will be square and is thus of width 3. Since $n = 3$ the matrix C must be of width 3. The height of B must match the height of x and is thus 3.

$$\begin{bmatrix} \dot{x}^{3 \times 1} \\ y^{2 \times 1} \end{bmatrix} = \left[\begin{array}{c|c} A^{3 \times 3} & B^{3 \times 4} \\ \hline C^{2 \times 3} & D^{2 \times 4} \end{array} \right] \begin{bmatrix} x^{3 \times 1} \\ u^{4 \times 1} \end{bmatrix}$$

Go back

2.10

(a) To begin with, states in a state-space model are the ones we have first-order differentials on, and consequently the states are $x = \begin{pmatrix} \eta \\ q \end{pmatrix}$. The remaining signals, which must be external inputs to the system, are thus $u = \begin{pmatrix} \phi \\ p \end{pmatrix}$. We have two (possibly vector-valued) outputs which when stacked together leads to the output $y = \begin{pmatrix} -\eta + q + \phi \\ q \end{pmatrix}$

To prepare for identification of a compact matrix representation, we start by writing it more explicitly using the states and inputs in the order we have defined them in their vector format in a very explicit format, including the definition of the output

$$\dot{\eta} = -R\eta + Mq + S\phi + Rp \quad (4)$$

$$\dot{q} = N\eta - Nq + 0\phi - Sp \quad (5)$$

$$y = \begin{pmatrix} -\eta + q + \phi + 0p \\ 0\eta + q + 0\phi + 0p \end{pmatrix} \quad (6)$$

In matrix format

$$\begin{pmatrix} \dot{\eta} \\ \dot{q} \end{pmatrix} = \begin{pmatrix} -R & M \\ N & -N \end{pmatrix} \begin{pmatrix} \eta \\ q \end{pmatrix} + \begin{pmatrix} S & R \\ 0 & -S \end{pmatrix} \begin{pmatrix} \phi \\ p \end{pmatrix} \quad (7)$$

$$y = \begin{pmatrix} -I & I \\ 0 & I \end{pmatrix} \begin{pmatrix} \eta \\ q \end{pmatrix} + \begin{pmatrix} I & 0 \\ 0 & 0 \end{pmatrix} u \quad (8)$$

which with our introduced notation can be written using our 4 sought matrices as

$$\dot{x} = \begin{pmatrix} -R & M \\ N & -N \end{pmatrix} x + \begin{pmatrix} S & R \\ 0 & -S \end{pmatrix} u \quad (9)$$

$$y = \begin{pmatrix} -I & I \\ 0 & I \end{pmatrix} x + \begin{pmatrix} I & 0 \\ 0 & 0 \end{pmatrix} u \quad (10)$$

Note that we have to use I and not 1 since we have vectors. Also note that 0 is a matrix of zeros, with dimension consistent with the vector it multiplies with, and the object it generates. MATLABs command `ss` accepts a sloppy scalar 0 in multivariable system but Simulinks block **State-space model** does not, so it is crucial to understand the dimensions on all objects.

- (b) If we use superscript to denote dimensions on identity matrices and zeros, our model is

$$\begin{aligned} \dot{x} &= \begin{pmatrix} -R & M \\ N & -N \end{pmatrix} x + \begin{pmatrix} S & R \\ 0^{a \times a} & -S \end{pmatrix} u \\ y &= \begin{pmatrix} -I^{a \times a} & I^{a \times a} \\ 0^{a \times a} & I^{a \times a} \end{pmatrix} x + \begin{pmatrix} I^{a \times a} & 0^{a \times b} \\ 0^{a \times a} & 0^{a \times b} \end{pmatrix} u \end{aligned}$$

Blocks multiplying with p must have width b , all other blocks have width a . Since all blocks are used in expressions where the resulting object must have height a , the blocks must have height a .

[Go back](#)

2.11

The dynamics is expanded to $\dot{\hat{x}} = A\hat{x} + B(-L\hat{x} + l_0 r) + K(y - C\hat{x})$ which we can gather as $\dot{\hat{x}} = (A - KC - BL)\hat{x} + Bl_0 r + Ky$. We gather the external signals as $\begin{pmatrix} r \\ y \end{pmatrix}$ and write the dynamics and output definition as

$$\begin{aligned} \dot{\hat{x}} &= (A - KC - BL)\hat{x} + (Bl_0 \quad K) \begin{pmatrix} r \\ y \end{pmatrix} \\ u &= -L\hat{x} + \begin{pmatrix} 0 & 0 \end{pmatrix} \begin{pmatrix} r \\ y \end{pmatrix} \end{aligned}$$

In other words, the 4 matrices defining the state-space model from (r, y) to u are

$$\left[\begin{array}{c|cc} A - KC - BL & Bl_0 & K \\ \hline -L & l_0 & 0 \end{array} \right]$$

Go back

2.12

The issue we illustrate in this exercise is that a model might have different interpretations, which leads to different representations.

- (a) Considering that we are asked to use two states, it is natural to guess that we can work with $X_1(s) = \frac{1}{s+1}U_1(s)$ and $X_2(s) = \frac{1}{s+1}U_2(s)$, as these are first-order differential equations (denominator has degree 1).

Write as $(s+1)X_1(s) = U_1(s)$ and $X_2(s) = \frac{1}{s+1}U_2(s)$ which leads us to $\dot{x}_1 + x_1 = u_1$ and $\dot{x}_2 + x_2 = u_2$. With that, we can write the state-space model as

$$\dot{x}_1(t) = -x_1(t) + u_1(t)$$

$$\dot{x}_2(t) = -x_2(t) + u_2(t)$$

$$y(t) = x_1(t) + x_2(t)$$

Define the state $x(t) = \begin{pmatrix} x_1(t) \\ x_2(t) \end{pmatrix}$ and input $u(t) = \begin{pmatrix} u_1(t) \\ u_2(t) \end{pmatrix}$ to write it using our 4 matrices.

$$\dot{x}(t) = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} x(t) + \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} u(t) \quad (11)$$

$$y(t) = \begin{pmatrix} 1 & 1 \end{pmatrix} x(t) + \begin{pmatrix} 0 & 0 \end{pmatrix} u(t) \quad (12)$$

We have $n = 2$ states, $m = 2$ inputs and $p = 1$ outputs. Consequently, our " A "-matrix is size $n \times n = 2 \times 2$, " B " is size $n \times m = 2 \times 2$, " C " is size $p \times n = 1 \times 2$ and " D " is size $p \times m = 1 \times 2$.

- (b) To arrive at only 1 state, we must recognize that the pole in the two transfer functions are the same.

This allows us to write the model as $Y(s) = \frac{1}{s+1}(U_1(s) + U_2(s))$. We let this expression be the state, $X_1(s) = \frac{1}{s+1}(U_1(s) + U_2(s))$ to arrive at $(s+1)X_1(s) = U_1(s) + U_2(s)$ corresponding to $\dot{x}_1 + x_1 = u_1 + u_2$. We thus write down the state space model

$$\begin{aligned}\dot{x}_1(t) &= -x_1(t) + u_1(t) + u_2 \\ y(t) &= x_1(t)\end{aligned}$$

Here we have $n = 1$ as $x(t) = x_1$, input is still $u(t) = \begin{pmatrix} u_1(t) \\ u_2(t) \end{pmatrix}$ with $m = 2$, and there is still only one output $p = 1$. Written in matrix form,

$$\begin{aligned}\dot{x}(t) &= (-1) x(t) + \begin{pmatrix} 1 & 1 \end{pmatrix} u(t) \\ y(t) &= \begin{pmatrix} 1 \end{pmatrix} x(t) + \begin{pmatrix} 0 & 0 \end{pmatrix} u(t)\end{aligned}$$

As $n = 1$ our " A "-matrix is of size $n \times n = 1 \times 1$ (a scalar), " B " is size $n \times m = 1 \times 2$, " C " is a scalar $p \times n = 1 \times 1$ and " D " is size $p \times m = 1 \times 2$.

- (c) The first model represents a physical setup with two identical tanks with different flow to each tank, and the output is the sum of the water levels in the two tanks, i.e., the total amount of water we currently have stored in our two tanks. The second model represents a single tank which has two input flows.
- (d) As expected, creating the model and converting to a transfer model gives us a model where the transfer function from every input to the output is $\frac{1}{s+1}$

```
A = [-1 0;0 -1]
B = [1 0;0 1];
C = [1 1];
D = [0 0];
G = ss(A,B,C,D)
G = tf(G)
```

```
G =
```

```
From input 1 to output:
```

```
1
```

```
-----
```


$$s + 1$$

From input 2 to output:

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

However, when we take this model and go back to a state-space model, it has lost one state!

ans =

A =

$$\begin{matrix} & x1 \\ x1 & -1 \end{matrix}$$

B =

$$\begin{matrix} & u1 & u2 \\ x1 & 0.5 & 0.5 \end{matrix}$$

C =

$$\begin{matrix} & x1 \\ y1 & 2 \end{matrix}$$

D =

$$\begin{matrix} & u1 & u2 \\ y1 & 0 & 0 \end{matrix}$$

Note that the model is not exactly the same as the one we derived with one state. The difference is that MATLAB has moved a factor 2 between B and C (divided B by 2, and multiplied C by 2). This means that the state is not the same state as we used, but that MATLAB has decided to use a state which represents half the water level instead (since y is the water level, and the model says $y = 2x$, it consequently means x must represent twice the water level).

When converting, MATLAB has derived a minimal representation. A minimal representation is a model which exhibits exactly the same input-output behaviour but uses the least number of states possible. A model can be non-minimal if it lacks controllability or observability.

This can be checked by the controllability and observability matrices which must be of full rank. Let us do that on the initial model with two states.

The system is controllable, as we easily can see there are two independent columns (full rank means there are n independent columns). Consequently, if we had both states measured, we would be able to control both states independently, and select a linear state-feedback which led to arbitrary poles in the closed-loop system.

```
>> ctrb(A,B)

ans =

1      0     -1      0
0      1      0     -1
>> rank(ctrb(A,B))
```

```
ans =

2
```

However, it is not observable (full rank means n independent rows)

```
>> obsv(A,C)

ans =

1      1
-1     -1

>> rank(obsv(A,C))

ans =

1
```

From a physical point of view, the system we have modelled is problematic as we cannot deduce the levels in the two tanks by simply observing the sum of the water levels, and consequently we cannot

create an observer to estimate the two individual levels if we want to control them. Nevertheless, the model with two states could be sensible to use in simulations etc if that actually is how we built the system but it then has to be kept in mind that it has these issues.

- (e) The transfer function is $\begin{pmatrix} \frac{1}{s+1} & \frac{1}{s+1} \end{pmatrix}$. To derive the pole polynomial, we have to compute the minors (sv. underdeterminanter) and maximal minors (sv. maximala underdeterminanter). There are two maximal minors as the largest square matrix we can obtain is by removing one column to arrive at a (trivial) square 1×1 matrix, thus leading to $\frac{1}{s+1}$ and $\frac{1}{s+1}$. The minors are simply the same two objects. Consequently the pole polynomial which is the least common denominator to all minors is $s + 1$ and thus there is 1 pole in -1. The analysis has derived the pole of a minimal realization of the model.

[Go back](#)

3 Properties of Linear Systems

3.1

The transfer function matrix has the minors

$$-\frac{1}{(s+2)^2} - \frac{(s+1)}{(s+2)^2} = -\frac{1}{s+2}$$

when the first column is deleted,

$$\frac{1}{(s+2)^2} - \frac{1}{(s+2)^2} = 0$$

when the second column is deleted and

$$\frac{(s+1)}{(s+2)^2} + \frac{1}{(s+2)^2} = \frac{1}{(s+2)}$$

when the third column is deleted. In addition, the elements of the transfer function are themselves minors. The pole polynomial, i.e. the least common denominator to all minors is thus

$$p(s) = (s+2)$$

The system has a pole in $s = -2$. Hence, the system can be realized as a state-space system of order one.

The maximal minors are

$$-\frac{1}{s+2}, \quad 0, \quad \frac{1}{(s+2)}$$

Thus, the zero polynomial is a constant. The system lacks zeros.

[Go back](#)

3.2

The transfer function matrix has the determinant

$$\det G(s) = \frac{2}{(s+3)^2}$$

and the minors

$$\frac{1}{(s+1)(s+3)}, \quad \frac{-1}{(s+1)(s+3)}, \quad \frac{2(s+1)}{(s+3)}$$

The pole polynomial, i.e. the least common denominator of the minors, is

$$p(s) = (s+1)(s+3)^2,$$

Hence the poles are -1 , -3 and -3 . The maximal minor is

$$\frac{2}{(s+3)^2}$$

If we normalize with the pole polynomial we get

$$\frac{2(s+1)}{(s+1)(s+3)^2}.$$

The zero polynomial is thus

$$n(s) = (s+1)$$

There is a zero in -1 .

[Go back](#)

3.3

Minors:

$$\underbrace{\frac{1-s}{(s+1)^2}}_{2 \text{ st}}, \quad \frac{2-s}{(s+1)^2}, \quad \frac{1/3-s}{(s+1)^2}, \quad \frac{1/3}{(s+1)^3}$$

There are 3 poles in $s = -1$ as the least common denominator is $(s+1)^3$.
Thus a minimal realization must be of order three.

[Go back](#)

3.4

- (a) The determinant of the transfer function matrix is

$$\frac{(s+5)}{(s^2+3s+2)(s+2)} - \frac{1}{(s+2)(s+4)} = \frac{6(s+3)}{(s+1)(s+2)(s+2)(s+4)}$$

and the minors are

$$\frac{(s+5)}{s^2+3s+2}, \quad \frac{1}{(s+2)}, \quad \frac{1}{(s+4)}, \quad \frac{1}{(s+2)}$$

Thus, the pole polynomial is

$$p(s) = (s+1)(s+2)(s+2)(s+4)$$

which means that the poles are located at -1 , -2 , -2 and -4 . We need four states to realize the system.

- (b) The determinant of the transfer function matrix is

$$\frac{(s+5)}{(s+4)(s^2+3s+2)} - \frac{1}{(s+2)(s+4)} = \frac{4}{(s+1)(s+2)(s+4)}$$

The pole polynomial is

$$p(s) = (s+1)(s+2)(s+4)$$

which means that the poles are located at -1 , -2 and -4 . We need three states to realize the system.

Go back

3.5

The system can be written on the form

$$A(p)y(t) = B(p)u(t)$$

where

$$A(p) = \begin{pmatrix} (p+1) & -p \\ p & (p+1) \end{pmatrix} \quad B(p) = \begin{pmatrix} 1 & -1 \\ 1 & 1 \end{pmatrix}$$

Multiplication with $A^{-1}(p)$ yields

$$G(p) = A^{-1}(p)B(p) = \frac{1}{2p^2 + 2p + 1} \begin{pmatrix} (2p+1) & -1 \\ 1 & (2p+1) \end{pmatrix}$$

The transfer function matrix, $G(s)$, has the determinant

$$\det G(s) = \frac{(2s+1)^2}{(2s^2 + 2s + 1)^2} + \frac{1}{(2s^2 + 2s + 1)^2} = \frac{2}{(2s^2 + 2s + 1)}$$

and the minors

$$\frac{(2s+1)}{(2s^2 + 2s + 1)} \quad \frac{-1}{(2s^2 + 2s + 1)} \quad \frac{1}{(2s^2 + 2s + 1)}$$

This results in the pole polynomial

$$p(s) = 2s^2 + 2s + 1$$

Hence, the poles are located at $-\frac{1}{2} \pm i\frac{1}{2}$.

The maximal minor is

$$\frac{2}{(2s^2 + 2s + 1)}$$

Thus, there are no zeros of the system.

Go back

3.6

Alt. 1: The output signal y only depends on x_1 and x_2 . The states x_1, x_2 do not depend on x_3 due to the structure of the matrix A . Hence, the state x_3 is unobservable and can be eliminated from the state-space form:

$$\dot{\tilde{x}}(t) = \begin{pmatrix} -2 & 1 \\ 0 & -1 \end{pmatrix} \tilde{x}(t) + \begin{pmatrix} 0 & 1 \\ 1 & 0 \end{pmatrix} u(t), \quad y(t) = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \tilde{x}(t),$$

where $\tilde{x} = (x_1, x_2)^T$. The controllability matrix and the observability matrix both have full rank and hence the realization is minimal.

Alt. 2: The transfer function matrix can be computed as $G(s) = C(sI - A)^{-1}B$

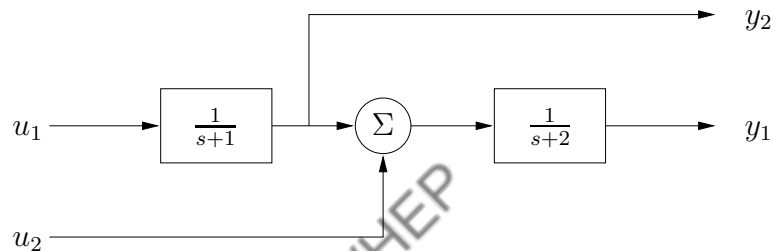
$$G(s) = \begin{pmatrix} \frac{1}{(s+2)(s+1)} & \frac{1}{s+2} \\ \frac{1}{s+1} & 0 \end{pmatrix},$$

i.e.

$$y_1 = \frac{1}{(s+2)(s+1)}u_1 + \frac{1}{s+2}u_2$$

$$y_2 = \frac{1}{s+1}u_1$$

This results in the block diagram:



Introduce a state after each block, for example $x_1 = y_1$ and $x_2 = y_2$. This results in the same minimal realization as in Alt.1.

Go back

3.7

- (a) The singular values at $\omega = 2$ can be determined in two ways, and for both alternatives we start by entering the transfer function matrix in Matlab.

```
>> s=tf('s');
>> G=[1/(s+1) 3/(s+2); 2/(s+3) 1/(s+4)];
```

Alternative (i): The frequency function $G(i\omega)$ at the angular frequency $\omega = 2$ can be computed according to


```
>> G2 = freqresp(G,2)
```

```
G2 =
```

```
    0.2000 - 0.4000i    0.7500 - 0.7500i
    0.4615 - 0.3077i    0.2000 - 0.1000i
```

The eigenvalues and eigenvector of $G(i\omega)^*G(i\omega)$ are now obtained from

```
>> [V,D]=eig(G2'*G2)
```

```
V =
```

```
    0.8288 + 0.2392i    0.4860 + 0.1403i
   -0.5059             0.8626
```

```
D =
```

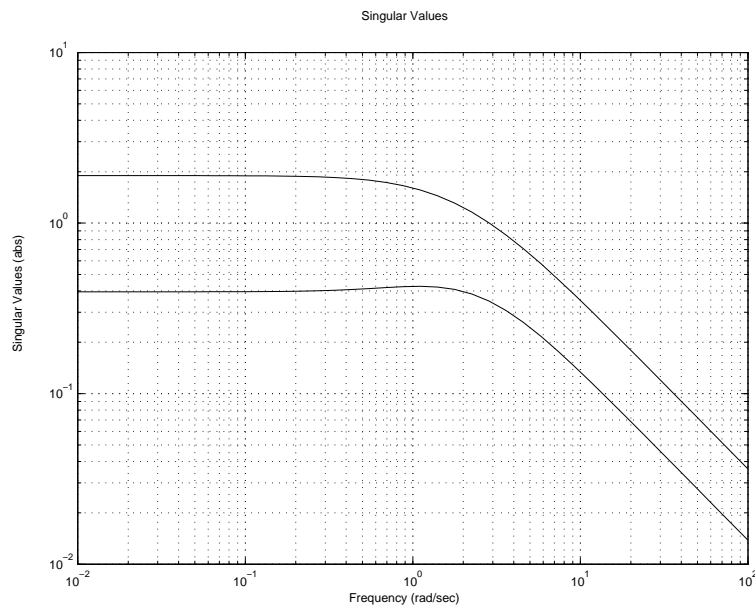
```
    0.1579    0
         0    1.5248
```

The smallest singular value is hence $\underline{\sigma}(G(i2)) = \sqrt{0.1579} \approx 0.40$ and the largest is $\bar{\sigma}(G(i2)) = \sqrt{1.5248} \approx 1.24$.

Alternative (ii): The singular values can be determined graphically using the command

```
>> sigma(G)
```

which gives the result



- (b) The second column of the matrix V defines the Fourier transform of the input vector that corresponds to the largest gain of the system, i.e. the input vector is such that the input components fulfill

$$|U_1(i\omega)| = \sqrt{0.486^2 + 0.1403^2} \approx 0.51$$

$$\arg U_1(i\omega) = \arctan(0.1403/0.486) \approx 0.28 \text{ rad}$$

and

$$|U_2(i\omega)| \approx 0.86 \quad \arg U_2(i\omega) = 0$$

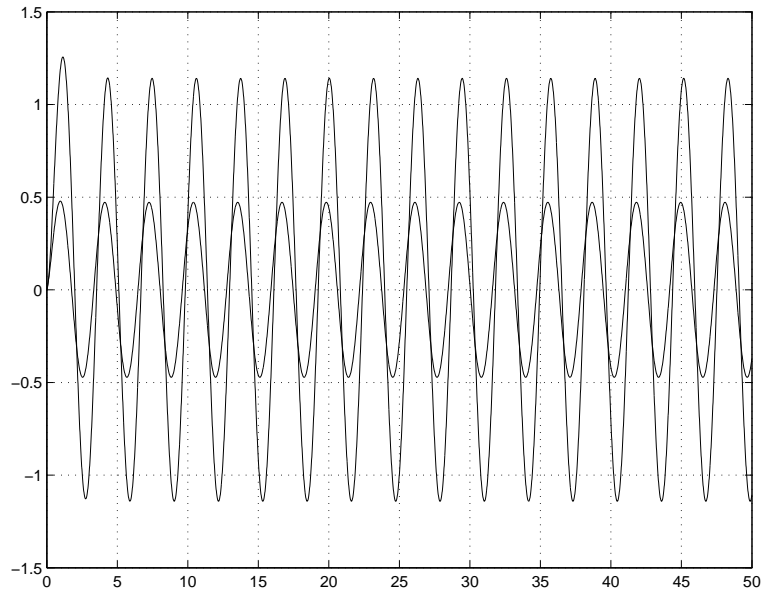
- (c) Using the hints an input vector can be generated using the command sequence

```
>> t=(0:0.01:50).';
>> u12=0.51*sin(2*t+0.28);
>> u22=0.86*sin(2*t);
```

The system can then be simulated using

```
>> y2=lsim(G,[u12 u22],t);
>> plot(t,y2)
```

This gives the result below showing that the output components have amplitudes 1.14 and 0.48.



- (d) Using the fact that the Fourier transform at the studied frequency are proportional to the signal amplitude the ratio of the norms of the output and input vectors becomes

$$\frac{\sqrt{1.14^2 + 0.48^2}}{\sqrt{0.51^2 + 0.86^2}} \approx 1.24$$

which corresponds to the largest singular value.

Go back

3.8

The minors of order 1 are

$$\frac{1}{s+1}, \quad \frac{s-1}{(s+1)(s+2)}, \quad \frac{-1}{s-1}, \quad \frac{1}{s+2}$$

The minors of order 2 are

$$\frac{-(s-1)}{(s+1)(s+2)^2}, \quad \frac{2}{(s+1)(s+2)}, \quad \frac{1}{(s+1)(s+2)}.$$

The least common denominator yields the pole polynomial

$$p(s) = (s + 1)(s + 2)^2(s - 1),$$

and the poles are therefore -1 , -2 , -2 , 1 . The maximal minors, normalized with the pole polynomial, are then given by

$$\frac{-(s - 1)^2}{(s + 1)(s + 2)^2(s - 1)}, \quad \frac{2(s - 1)(s + 2)}{(s + 1)(s + 2)^2(s - 1)}, \quad \frac{(s - 1)(s + 2)}{(s + 1)(s + 2)^2(s - 1)},$$

and the gcd of the numerators is thus $z(s) = s - 1$ and the only zero is 1 .

[Go back](#)

5 Disturbance Models

5.1

$\Phi_u(\omega)$ is a non-negative rational function in ω^2 which means a spectral factorization should be possible. Do the decomposition $\Phi_u(\omega) = G(i\omega)G(-i\omega)\Phi_e(\omega)$ where $G(s)$ has all poles and zeros in the left-half plane and $\Phi_e = 1$.

(a)

$$\Phi_u(\omega) = \frac{a^2}{\omega^2 + a^2} \Phi_e(\omega) = \frac{a}{i\omega + |a|} \cdot \frac{a}{-i\omega + |a|}$$

Thus the linear filter is

$$G(s) = \frac{a}{s + |a|}$$

(b) We note the spectrum appears to be a product of two functions, and both of these have the same structure as in (a). We must have

$$\begin{aligned} \Phi_u(\omega) &= \frac{a^2 b^2}{(\omega^2 + a^2)(\omega^2 + b^2)} \Phi_e(\omega) \\ &= \frac{a^2}{(i\omega + |a|)(-i\omega + |a|)} \frac{b^2}{(i\omega + |b|)(-i\omega + |b|)} \\ \Rightarrow G(s) &= \frac{ab}{(s + |a|)(s + |b|)} \end{aligned}$$

(c) Only structural difference compared to (a) is that we note there must be zero

$$\Phi_u(\omega) = \frac{\omega^2 + c^2}{\omega^2 + a^2} \Phi_e(\omega) = \frac{(i\omega + |c|)(-i\omega + |c|)}{(i\omega + |a|)(-i\omega + |a|)}$$

Thus the linear filter is

$$G(s) = \frac{s + |c|}{s + |a|}$$

(d) For both systems, the amplitude gain starts at c/a for $\omega = 0$ and they both asymptotically tend to 1 as $\omega \rightarrow \infty$. A zero in $-c$ means the derivative of the amplitude gain starts to increase (when viewed in a logarithmic scale) just before c rad/s. A pole in $-a$ means

the derivative of the amplitude gain starts to drop off just before a rad/s. Consequently, when $c > a$ the amplitude gain starts at c/a and then starts dropping off around a rad/s and then levels out after c rad/s and reaches the gain 1. For $c < a$ the amplitude gain starts at $c/a < 1$ and then starts increasing around c rad/s and then levels out after a rad/s and reaches the gain 1. In MATLAB, we can, e.g., study `bodemag((s+1)/(s+10), (s+10)/(s+1))`.

[Go back](#)

5.2

Consider the disturbance model

$$N(s) = H(s)V(s)$$

where V is white. In (i) the transfer function is of low pass character, which means that N will be of low frequency character. The disturbance is located around 5 Hz, i.e. 10π rad/s. The magnitude curve of model (ii) has a peak around this angular frequency, which means that this model is the most appropriate one. In model (iii) the peak is located around 5 rad/s but it also has low-frequency content.

```
s = tf('s');
bodemag(1/(s+0.001), 900*s/(s^2+6*s+900), 25/(s^2+s+25)); grid on
```

[Go back](#)

5.3

(a) We are given

$$f = k_1 \dot{z} + v$$

The force is $m\ddot{z} = u - f$, where m is the mass of the missile and u is the thrust.

On input-output form:

$$\ddot{z} + \frac{k_1}{m}\dot{z} = \frac{1}{m}(u - v)$$

State-space form: Let $x_1 = z$, $x_2 = \dot{z} \Rightarrow \dot{x}_1 = x_2$,

$$\dot{x}_2 = \frac{1}{m}(u - f) = \frac{1}{m}(u - k_1x_2 - v)$$

That is

$$\dot{x} = \begin{pmatrix} 0 & 1 \\ 0 & -\frac{k_1}{m} \end{pmatrix} x + \begin{pmatrix} 0 \\ \frac{1}{m} \end{pmatrix} u + \begin{pmatrix} 0 \\ -\frac{1}{m} \end{pmatrix} v$$

$$z = \begin{pmatrix} 1 & 0 \end{pmatrix} x$$

(b) Description of v :

$$\Phi_v(\omega) = |H(i\omega)|^2 \Phi_e(\omega)$$

Thus $H(s) = \frac{\sqrt{k_0}}{s+|a|}$, i.e. $\dot{v} + |a|v = \sqrt{k_0}e$. Introduce an extra state $x_3 = v$ which results in a new state-space form with $\dot{x}_3 = -|a|x_3 + e$:

$$\dot{x} = \begin{pmatrix} 0 & 1 & 0 \\ 0 & -\frac{k_1}{m} & -\frac{1}{m} \\ 0 & 0 & -|a| \end{pmatrix} x + \begin{pmatrix} 0 \\ \frac{1}{m} \\ 0 \end{pmatrix} u + \begin{pmatrix} 0 \\ 0 \\ \sqrt{k_0} \end{pmatrix} e$$

$$z = \begin{pmatrix} 1 & 0 & 0 \end{pmatrix} x$$

The input-output form is

$$(p^2 + \frac{k_1 p}{m}) z = \frac{1}{m} \left(u - \frac{\sqrt{k_0}}{p + |a|} e \right)$$

The frequency content of the wind gusts are described by the amplitude gain of $H(s)$ which means it is considerable in frequencies up to $|a|$ and then drops off.

Go back

5.4

- (a) With $\{A, B, C, N\}$ according to exercise 5.3 we get

$$\begin{aligned}\dot{x} &= Ax + Bu + Ne \\ y &= Cx + n\end{aligned}$$

where n has spectral density $\Phi_n \equiv 0.1$.

- (b) A noise signal with the desired spectral density can be generated by a system with transfer function $G_n(s) = \frac{s}{s+|b|}$. The input is white noise with spectral density $\Phi_{w_n} = 0.1$. On state-space form we get

$$\begin{aligned}\dot{x}_4 &= -|b|x_4 + |b|w_n \\ n &= -x_4 + w_n\end{aligned}$$

The extended state-space form is

$$\begin{aligned}\dot{x} &= \begin{pmatrix} A & 0 \\ 0 & -|b| \end{pmatrix} x + \begin{pmatrix} B \\ 0 \end{pmatrix} u + \begin{pmatrix} N & 0 \\ 0 & |b| \end{pmatrix} \begin{pmatrix} e \\ w_n \end{pmatrix} \\ y &= \begin{pmatrix} C & -1 \end{pmatrix} x + w_n\end{aligned}$$

- (c) Following the same procedure as in (b) we get a transfer function $G_n(s) = \frac{1}{s+|b|}$. The input is white noise with spectral density $\Phi_{w_n} = 0.1$. On state-space form we get

$$\dot{x}_4 + |b|x_4 = w_n.$$

The extended state-space form is

$$\begin{aligned}\dot{x} &= \begin{pmatrix} A & 0 \\ 0 & -|b| \end{pmatrix} x + \begin{pmatrix} B \\ 0 \end{pmatrix} u + \begin{pmatrix} N & 0 \\ 0 & 1 \end{pmatrix} \begin{pmatrix} e \\ w_n \end{pmatrix} \\ y &= \begin{pmatrix} C & 1 \end{pmatrix} x\end{aligned}$$

Go back

5.5

- (1) A model for w : A stepwise change results from $w = \frac{1}{s}v$ where v is a number of impulses.
Introduce the state x_w , $\dot{x}_w = v$.

- (2) A model for n : Use a second order system with a resonance peak at $\omega_0 = 2\pi \cdot 2 = 4\pi$ rad/s and damping $\xi = 0.01$

$$n = \frac{\omega_0^2}{p^2 + 2\xi\omega_0 p + \omega_0^2} e$$

Introduce the states $x_{n1} = n$ and $x_{n2} = \dot{n}$

$$\dot{x}_n = \begin{pmatrix} \dot{x}_{n1} \\ \dot{x}_{n2} \end{pmatrix} = \underbrace{\begin{pmatrix} 0 & 1 \\ -\omega_0^2 & -2\xi\omega_0 \end{pmatrix}}_{A_n} x_n + \underbrace{\begin{pmatrix} 0 \\ 1 \end{pmatrix}}_{B_n} e$$

We get the extended model

$$\dot{x}_u = \begin{pmatrix} \dot{x} \\ \dot{x}_w \\ \dot{x}_n \end{pmatrix} = \begin{pmatrix} A & N & 0 \\ 0 & 0 & 0 \\ 0 & 0 & A_n \end{pmatrix} x_u + \begin{pmatrix} B \\ 0 \\ 0 \end{pmatrix} u + \begin{pmatrix} 0 & 0 \\ 1 & 0 \\ 0 & B_n \end{pmatrix} \begin{pmatrix} v \\ e \end{pmatrix}$$

[Go back](#)

5.6

- (a) Choose the states $x_1 =$ acceleration and $x_2 =$ speed. This results in the state-space form

$$\dot{x} = \begin{pmatrix} 0 & 0 \\ 1 & 0 \end{pmatrix} x + \begin{pmatrix} 1 \\ 0 \end{pmatrix} e$$

$$y = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} x + \begin{pmatrix} v_1 \\ v_2 \end{pmatrix}$$

- (b)

$$\dot{\hat{x}} = \left[\begin{pmatrix} 0 & 0 \\ 1 & 0 \end{pmatrix} - \begin{pmatrix} k_{11} & k_{12} \\ k_{21} & k_{22} \end{pmatrix} \right] \hat{x} + \begin{pmatrix} k_{11} & k_{12} \\ k_{21} & k_{22} \end{pmatrix} y$$

The matrix K is determined from the algebraic Riccati equation.

[Go back](#)

5.7

We get the state-space form

$$\begin{pmatrix} \dot{x}_1 \\ \dot{x}_2 \end{pmatrix} = \underbrace{\begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}}_{=A} \begin{pmatrix} x_1 \\ x_2 \end{pmatrix} + \underbrace{\begin{pmatrix} 0 \\ 1 \end{pmatrix}}_{=N} v_1$$

where $x_1 = x$, $x_2 = \dot{x}$.

The Kalman filter:

$$\dot{\hat{x}} = \begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix} \hat{x} + \begin{pmatrix} k_1 \\ k_2 \end{pmatrix} (y - C\hat{x})$$

where $C = (1 \ 0)$ (Case I) or $C = (0 \ 1)$ (Case II). The noise intensity is

$$R = \begin{bmatrix} R_1 & R_{12} \\ R_{21} & R_2 \end{bmatrix} = \begin{cases} E \begin{bmatrix} v(t) \\ e_1(t) \end{bmatrix} \begin{bmatrix} v(t) & e_1(t) \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} & \text{in case I, and} \\ E \begin{bmatrix} v(t) \\ e_2(t) \end{bmatrix} \begin{bmatrix} v(t) & e_2(t) \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} & \text{in case II.} \end{cases}$$

The Kalman gain is

$$K = \begin{pmatrix} k_1 \\ k_2 \end{pmatrix} = PC^T R_2^{-1},$$

where $P = E[\tilde{x}(t)\tilde{x}^T(t)]$ is given by the ARE

$$AP + PA^T - PC^T R_2^{-1} CP + NR_1 N^T = 0.$$

Case I:

$$P = \begin{pmatrix} 0.910 & 0.414 \\ 0.414 & 1.287 \end{pmatrix}$$

Case II:

$$P = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}$$

The position x_1 is estimated more accurately in case I and the speed x_2 is estimated more accurately in case II.

Go back

5.8

Introduce the states

$$x(t) = \begin{pmatrix} \Theta(t) \\ \dot{\Theta}(t) \end{pmatrix}$$

and denote $\alpha = B/J$, $H = k/J$ and $\gamma = 1/J$. A state space model of the system is

$$\begin{aligned} \dot{x}(t) &= \underbrace{\begin{pmatrix} 0 & 1 \\ 0 & -\alpha \end{pmatrix}}_A x(t) + \underbrace{\begin{pmatrix} 0 \\ H \end{pmatrix}}_B \mu(t) + \underbrace{\begin{pmatrix} 0 \\ \gamma \end{pmatrix}}_N \tau_d(t) \\ y(t) &= \underbrace{\begin{pmatrix} 1 & 0 \end{pmatrix}}_C x(t) + e_m(t) \end{aligned}$$

The Riccati equation used to compute the Kalman gain K is

$$AP + PA^T + NR_1N^T - (PC^T + NR_{12})R_2^{-1}(PC^T + NR_{12}) = 0.$$

Using $R_{12} = 0$ (process- and measurement noise independent), A , N and C as above, $R_1 = v_d$ and $R_2 = v_m$ gives

$$\begin{pmatrix} 0 & 1 \\ 0 & -\alpha \end{pmatrix} P + P \begin{pmatrix} 0 & 0 \\ 1 & -\alpha \end{pmatrix} + \begin{pmatrix} 0 \\ \gamma \end{pmatrix} v_d (0 \quad \gamma) - P \begin{pmatrix} 1 \\ 0 \end{pmatrix} v_m^{-1} (1 \quad 0) P = 0.$$

With

$$P = \begin{pmatrix} p_{11} & p_{12} \\ p_{12} & p_{22} \end{pmatrix},$$

the components of this matrix equation become

$$\begin{aligned} 2p_{12} - \frac{p_{11}^2}{v_m} &= 0 \\ p_{22} - \alpha p_{12} - \frac{p_{11}p_{12}}{v_m} &= 0 \\ -2\alpha p_{22} + \gamma^2 v_d - \frac{p_{12}^2}{v_m} &= 0 \end{aligned}$$

If we eliminate p_{12} and p_{22} we get

$$\frac{p_{11}^4}{4v_m^3} + \frac{\alpha p_{11}^3}{v_m^2} + \frac{\alpha^2 p_{11}^2}{v_m} - \gamma^2 v_d = 0$$

Now introduce

$$p_{11} = v_m \cdot p'_{11}$$

which yields

$$\begin{aligned} p_{11}'^4 + 4\alpha p_{11}'^3 + 4\alpha^2 p_{11}'^2 - 4\gamma^2 \frac{v_d}{v_m} &= 0 \\ (p_{11}'^2 + 2\alpha p_{11}')^2 - 4\gamma^2 \frac{v_d}{v_m} &= 0 \\ p_{11}'^2 + 2\alpha p_{11}' - 2\gamma \sqrt{\frac{v_d}{v_m}} &= 0 \end{aligned}$$

Define $\beta = \gamma \sqrt{\frac{v_d}{v_m}}$ This results in

$$p_{11}' = -\alpha + \sqrt{\alpha^2 + 2\beta}$$

The solution is

$$P = v_m \begin{pmatrix} -\alpha + \sqrt{\alpha^2 + 2\beta} & \alpha^2 + \beta - \alpha \sqrt{\alpha^2 + 2\beta} \\ \alpha^2 + \beta - \alpha \sqrt{\alpha^2 + 2\beta} & -\alpha^3 - 2\alpha\beta + (\alpha^2 + \beta) \sqrt{\alpha^2 + 2\beta} \end{pmatrix}$$

The steady state Kalman gain $K = PC^T R_2^{-1}$ becomes

$$K = \begin{pmatrix} -\alpha + \sqrt{\alpha^2 + 2\beta} \\ \alpha^2 + \beta - \alpha \sqrt{\alpha^2 + 2\beta} \end{pmatrix}$$

and using the numerical values given we get

$$K = \begin{pmatrix} 40.36 \\ 814.3 \end{pmatrix}$$

The covariance matrix for the estimation error is

$$P = \begin{pmatrix} 40.36 \cdot 10^{-7} & 814.3 \cdot 10^{-7} \\ 814.3 \cdot 10^{-7} & 366.1 \cdot 10^{-5} \end{pmatrix}$$

Hence the filter for estimating Θ is

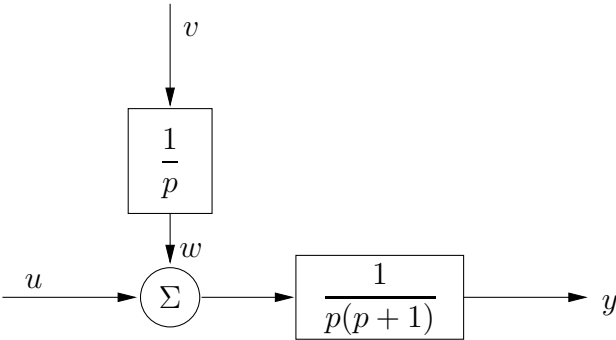
$$\dot{\hat{x}} = \begin{pmatrix} 0 & 1 \\ 0 & -\alpha \end{pmatrix} \hat{x} + \begin{pmatrix} 0 \\ H \end{pmatrix} \mu(t) + K (y - (1 \ 0) \hat{x})$$

with K as above.

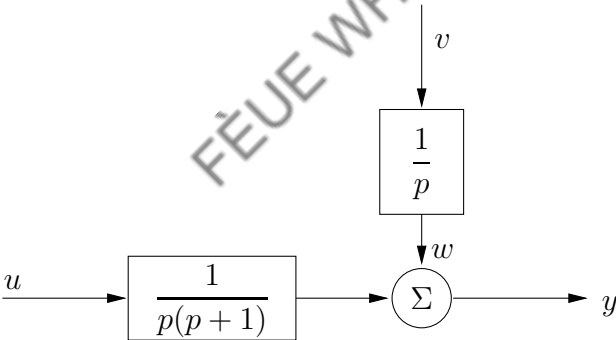
Go back

5.9

(i)



(ii)



$v(t)$ unit disturbance

(a) (i)

$$\begin{aligned} \dot{x} &= \overbrace{\begin{pmatrix} 0 & 1 & 0 \\ 0 & -1 & 1 \\ 0 & 0 & 0 \end{pmatrix}}^A x + \overbrace{\begin{pmatrix} 0 \\ 1 \\ 0 \end{pmatrix}}^B u + \begin{pmatrix} 0 \\ 0 \\ 1 \end{pmatrix} v \\ y &= \underbrace{\begin{pmatrix} 1 & 0 & 0 \end{pmatrix}}_C x. \end{aligned}$$

(ii)

$$\dot{x} = \overbrace{\begin{pmatrix} 0 & 1 & 0 \\ 0 & -1 & 0 \\ 0 & 0 & 0 \end{pmatrix}}^A x + \overbrace{\begin{pmatrix} 0 \\ 1 \\ 0 \end{pmatrix}}^B u + \begin{pmatrix} 0 \\ 0 \\ 1 \end{pmatrix} v$$

$$y = \underbrace{\begin{pmatrix} 1 & 0 & 1 \end{pmatrix}}_C x.$$

- (b) (i) Offset in the motor voltage, step disturbance in the load
(ii) Measurement disturbance – error in the sensor for angular displacement

(c) (i)

$$S = (B \quad AB \quad A^2B) = \begin{pmatrix} 0 & 1 & -1 \\ 1 & -1 & 1 \\ 0 & 0 & 0 \end{pmatrix} \quad \text{not full rank}$$

(ii)

$$S = \begin{pmatrix} 0 & 1 & -1 \\ 1 & -1 & 1 \\ 0 & 0 & 0 \end{pmatrix} \quad \text{not full rank}$$

In (i) we can make x_3 unobservable by choosing $u = -Lx$ with $\ell_3 = 1$. This is not possible in (ii).

Go back

5.10

- (a) The spectrum of the wind has low pass characteristics with bandwidth α . When α increases $v(t)$ behaves more and more like white noise, i.e. the gustiness increases. This can also be seen by studying the covariance function

$$R_v(\tau) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \Phi_v(\omega) e^{i\omega\tau} d\omega = e^{-\alpha|\tau|}, \quad \alpha > 0.$$

The covariance function gets more narrow when α increases, i.e. the correlation with neighboring values of $v(t)$ decreases and the gustiness increases.

- (b) Using spectral factorization, the influence from the wind can be described as white noise $e(t)$ with intensity 1 filtered through a linear system with transfer function

$$H(s) = \frac{\sqrt{2/\alpha}}{1 + s/\alpha}$$

. We get $y = G(s) H(s)e$ where

$$G(s) H(s) = \frac{K\sqrt{2\alpha}}{(\alpha + s)(s^2 + s + 1)} = \frac{K\sqrt{2\alpha}}{s^3 + (1 + \alpha)s^2 + (1 + \alpha)s + \alpha}.$$

The variance of the output signal is

$$\begin{aligned} \text{Var}(y) &= \frac{1}{2\pi} \int_{-\infty}^{\infty} |G(i\omega)H(i\omega)|^2 d\omega \\ &= \frac{1}{2\pi} \int_{-\infty}^{\infty} \left| \frac{K\sqrt{2\alpha}}{(i\omega)^3 + (1 + \alpha)(i\omega)^2 + (1 + \alpha)i\omega + \alpha} \right|^2 d\omega \\ &= \frac{K^2(1 + \alpha)}{1 + \alpha + \alpha^2}. \end{aligned}$$

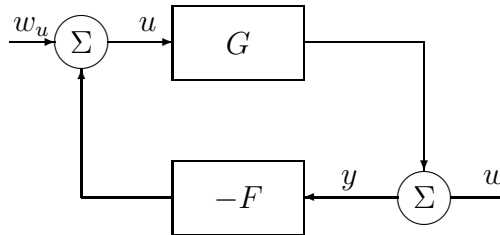
Thus the requirement can be formulated as $\frac{K^2(1+\alpha)}{1+\alpha+\alpha^2} > 1$.

[Go back](#)

6 The Closed-Loop System

6.1

Consider the block diagram



We have the relationships

$$y = (I + GF)^{-1}(w + Gw_u) = G_{wy}w + G_{wu}w_u$$

and

$$u = (I + FG)^{-1}(w_u - Fw) = G_{wu}w_u + G_{uw}w$$

which results in the input-output model

$$\begin{bmatrix} u \\ y \end{bmatrix} = \begin{bmatrix} G_{wu} & G_{uw} \\ G_{wy} & G_{yu} \end{bmatrix} \begin{bmatrix} w_u \\ w \end{bmatrix}.$$

We also have

$$w_u = u + Fy$$

and

$$w = y - Gu$$

which results in the transfer function matrix

$$\begin{bmatrix} w_u \\ w \end{bmatrix} = \begin{bmatrix} I & F \\ -G & I \end{bmatrix} \begin{bmatrix} u \\ y \end{bmatrix}$$

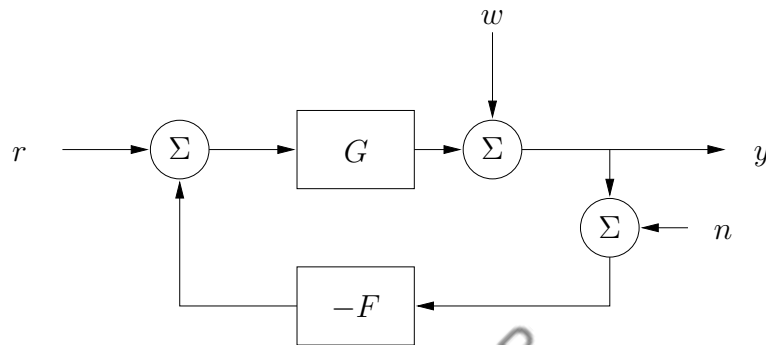
Thus, we have shown that

$$\begin{bmatrix} G_{wu} & G_{uw} \\ G_{wy} & G_{yu} \end{bmatrix}^{-1} = \begin{bmatrix} I & F \\ -G & I \end{bmatrix}$$

Alternative solution: Show that the matrix product is the identity matrix.

[Go back](#)

6.2



With the transfer functions

$$G = \frac{s-1}{s+2}, \quad F = \frac{s+2}{s-1}$$

we get

$$Y = G(R - F(Y + N)) + W \Rightarrow (1 + GF)Y = GR - GFN + W$$

$$\Rightarrow Y = (1 + GF)^{-1}GR - (1 + GF)^{-1}GFN + (1 + GF)^{-1}W$$

The closed-loop system, the sensitivity function and the complementary sensitivity function are

$$G_c = G_{ry} = (1 + GF)^{-1}G = \frac{s-1}{2s+3}$$

$$S = G_{wy} = (1 + GF)^{-1} = \frac{s+1}{2s+3}$$

$$T = 1 - S = \frac{s+2}{2s+3}$$

and are all stable.

Internal stability?

Check the following transfer functions

$$G_{w_u u} = (1 + FG)^{-1} = \frac{s+1}{2s+3}$$

$$G_{w_u} = -(1 + FG)^{-1}F = -\frac{(s+2)(s+1)}{(s-1)(2s+3)}$$

$$G_{w_u y} = (1 + GF)^{-1}G = \frac{s-1}{2s+3}$$

$$G_{w_y} = (1 + GF)^{-1} = \frac{s+1}{2s+3}$$

The systemet is not internally stable as G_{w_u} is unstable.

Go back

7 Limitations in Control Design

7.1

- (a) The complementary sensitivity function is given by

$$T(s) = \frac{F(s)G(s)}{1 + F(s)G(s)}$$

To compute the controller $F(s)$ which results in the desired $T(s)$ we express $F(s)$ as a function of $T(s)$ and $G(s)$ as follows. The above expression yields

$$F(s) = G^{-1}(s) \frac{T(s)}{1 - T(s)}$$

With the given $T(s)$ and $G(s)$ this results in

$$F(s) = \frac{5(s+1)}{s(s-3)}$$

The zero located in $s = 3$ will be cancelled. However, the transfer function from reference signal to control signal

$$U(s) = \frac{5(s+1)}{(s-3)(s+5)} R(s)$$

has a pole located in $s = 3$ and is unstable.

- (b) We can get a bandwidth of 5 rad/s if we keep the right-half plane zero and add a pole in $s = -3$, i.e.

$$T(s) = \frac{5}{s+5} \cdot \frac{3-s}{3+s}$$

In this case, the relationship

$$F(s) = G^{-1}(s) \frac{T(s)}{1 - T(s)}$$

yields

$$F(s) = -\frac{5(s+1)}{s(s+13)}$$

No pole-zero cancellation occurs and all closed-loop system transfer functions are stable.

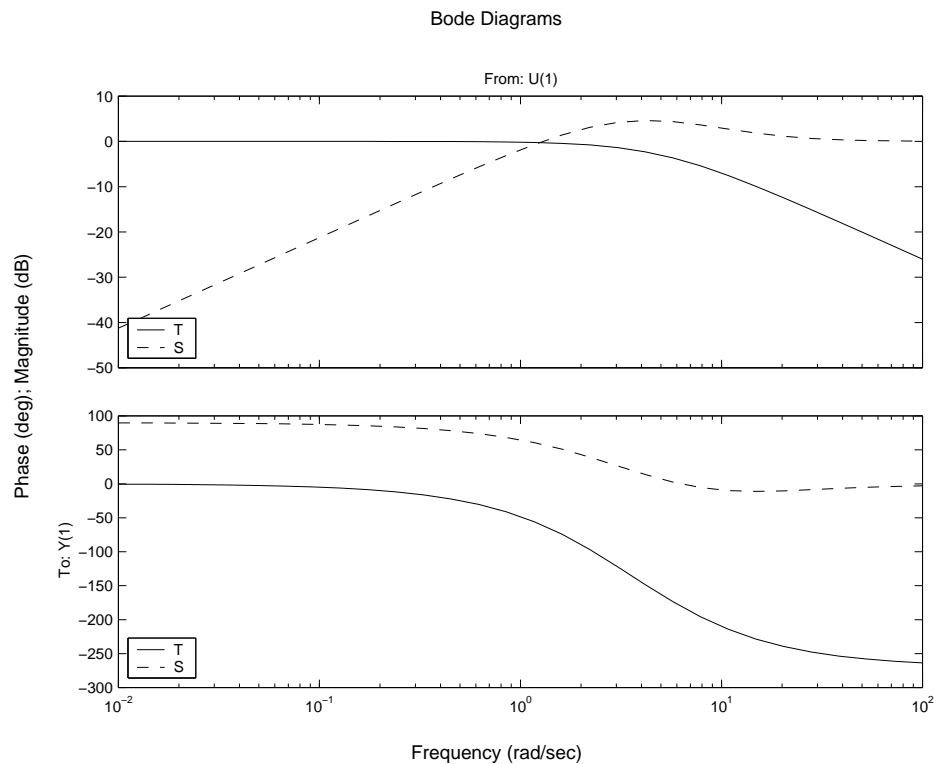
(c) With $F_r = F_y = F$ we get

$$G_c(s) = T(s) = \frac{5}{s+5} \cdot \frac{3-s}{3+s}$$

The sensitivity function is

$$S(s) = 1 - T(s) = \frac{s(s+13)}{(s+3)(s+5)}$$

The Bode diagrams for those transfer functions are shown in the figure below.



The bandwidth for G_c (i.e. T) is 5 rad/s as desired. However, the phase is -90° already at 2 rad/s. This means that the output signal will not follow the reference signal for frequencies above this. In addition, $|S(i\omega)| > 1$ for $\omega \geq 1.3$ rad/s, i.e. system disturbances are *amplified* for those frequencies.

[Go back](#)

7.2

According to the rules of thumb presented in the textbook the bandwidth of the closed-loop system cannot be greater than (1) half the magnitude of the right-half plane zero, in this case $3/2 = 1.5$ rad/s, (2) the inverse of the time-delay, here $1/1 = 1$ rad/s.

As a comparison we can study the crossover frequency. The crossover frequency is often close to the bandwidth. The transfer function for a system with a zero in $s = 3$ and a time-delay of 1 second can be expressed as

$$G(s) = e^{-s}(3 - s)\bar{G}(s)$$

or

$$G(s) = e^{-s} \frac{(3 - s)}{(3 + s)} (3 + s) \bar{G}(s)$$

The argument of the frequency response is

$$\arg G(i\omega) = -\omega - 2 \arctan \frac{\omega}{3} + \arg((3 + i\omega)\bar{G}(i\omega))$$

According to the assumptions the magnitude curve decreases monotonically and according to Bode's relation we get

$$\arg((3 + i\omega)\bar{G}(i\omega)) \leq 0$$

This implies that

$$\arg G(i\omega) \leq -\omega - 2 \arctan \frac{\omega}{3}$$

and the phase margin is

$$\varphi_m = \pi + \arg G(i\omega_c) \leq \pi - \omega_c - 2 \arctan \frac{\omega_c}{3}$$

Let us study the case $\varphi_m = 0$ under the assumption that equality holds in the above inequality. Then

$$0 = \pi - \omega_c - 2 \arctan \frac{\omega_c}{3}$$

i.e.

$$\omega_c \approx 2$$

Hence, the crossover frequency cannot be greater than 2 rad/s.

Go back

7.3

Assume that we have a zero close to the origin and a pole far from the origin in, both in the right-half plane. For example, we could have ($\epsilon \ll 1$)

$$G(s) = \frac{-\epsilon + s}{-\frac{1}{\epsilon} + s}$$

According to Theorem 7.4 in the textbook, the magnitude of the sensitivity function must have a peak in a neighborhood of ϵ . In addition, Theorem 7.6 says that the magnitude of the complementary sensitivity function must have a peak in the neighborhood of $1/\epsilon$.

Go back

7.4

- (a) The requirements on $|S(i\omega)| = \bar{\sigma}(S(i\omega))$ and $|T(i\omega)| = \bar{\sigma}(T(i\omega))$ can be formulated as

$$\begin{aligned} |S(i\omega)| &\leq \frac{1}{10}, & \omega &\leq 0.1, & |T(i\omega)| &\leq \frac{1}{10}, & \omega &\geq 2 \\ |S(0)| &\leq \frac{1}{100} \end{aligned}$$

- (b) The corresponding requirements on the loop gain GF_y is

$$\begin{aligned} |G(0)F_y(0)| &> 100 \\ |G(i\omega)F_y(i\omega)| &> 10, & \omega &\leq 0.1 \\ |G(i\omega)F_y(i\omega)| &< \frac{1}{10}, & \omega &\geq 2 \end{aligned}$$

- (c) The requirements in (a) can be reformulated using weighting functions W_S and W_T such that

$$\begin{aligned} |S(i\omega)| &\leq |W_S^{-1}(i\omega)|, & \forall \omega \\ |T(i\omega)| &\leq |W_T^{-1}(i\omega)|, & \forall \omega \end{aligned}$$

If W_S^{-1} and W_T^{-1} are first order transfer functions

$$W_S^{-1}(s) = a_1 \left(1 + \frac{s}{b_1} \right), \quad W_T^{-1}(s) = \frac{a_2}{s} \left(1 + \frac{s}{b_2} \right)$$

we get, for example,

$$W_S^{-1}(s) = \frac{1}{100}(1 + 30\sqrt{11}s), \quad W_T^{-1}(s) = \frac{\sqrt{2}}{10s} \left(1 + \frac{s}{2}\right)$$

- (d) The minimal slope of the magnitude of the loop gain in the interval $[0.1, 2]$ is approximately given by the line tangent to the forbidden regions in (b).

$$\text{Slope in the Bode plot: } \frac{\log 0.1 - \log 10}{\log 2 - \log 0.1} \approx -1.53$$

This implies

$$\frac{\log 1 - \log 10}{\log \omega_c - \log 0.1} = -1.53 \quad \Rightarrow \quad \omega_c = 0.45 \text{ rad/s}$$

From Bode's relation we get

$$\arg GF_y \lesssim -1.53 \cdot \frac{\pi}{2} = -138^\circ$$

which results in a phase margin of approximately 40° .

A lower bound on $\|T\|_\infty$?

$$G(i\omega_c)F_y(i\omega_c) = 1 \cdot e^{-i \cdot 138^\circ} = -0.743 + 0.669i$$

$$|T(i\omega_c)| = \left| \frac{G(i\omega_c)F_y(i\omega_c)}{1 + G(i\omega_c)F_y(i\omega_c)} \right| \approx 1.4$$

$$\begin{aligned} \|T\|_\infty &= \sup_{\omega} |T(i\omega)| \quad \Rightarrow \quad \|T\|_\infty \geq |T(i\omega)|, \quad \forall \omega \\ &\Rightarrow \quad \|T\|_\infty \geq 1.4 \end{aligned}$$

(e)

$$|T(i\omega_c)| = 1.4$$

$$|W_T^{-1}(i\omega_c)| = \frac{0.14}{0.45} \sqrt{1 + \frac{0.45^2}{2^2}} = 0.32$$

It is impossible to find a feasible solution using this choice of weighting functions. Try weighting functions of higher order.

Go back

7.5

If the surface A_2 is greater than the surface A_1 we have that $\int_0^\infty \log |S(i\omega)| d\omega > 0$. According to Theorem 7.3, the loop gain $G(s)F_y(s)$ has unstable poles.

Go back

7.6

The first requirement implies that

$$|S(i\omega)| < 10^{-3} \quad \omega \leq 2$$

where

$$S(s) = \frac{1}{1 + F(s)G(s)}$$

When $|F(i\omega)G(i\omega)|$ is large we approximately have

$$|S(i\omega)| \approx \frac{1}{|F(i\omega)G(i\omega)|}$$

which results in

$$|F(i\omega)G(i\omega)| > 10^3 \quad \omega \leq 2$$

Furthermore, the system should be stable in spite of the model uncertainty

$$|\Delta G(i\omega)| \leq 100|G(i\omega)| \quad \omega \geq 20$$

where $\Delta G(s)$ is the absolute model error in $G(s)$. Thus, the relative model error fulfills the inequality

$$\left| \frac{\Delta G(i\omega)}{G(i\omega)} \right| \leq 100$$

To preserve stability we must have

$$|T(i\omega)| < 10^{-2} \quad \omega \geq 20$$

where

$$T(s) = \frac{F(s)G(s)}{1 + F(s)G(s)}$$

When $|F(i\omega)G(i\omega)|$ is small we approximately have

$$|T(i\omega)| \approx |F(i\omega)G(i\omega)|$$

and this results in

$$|F(i\omega)G(i\omega)| < 10^{-2} \quad \omega \geq 20$$

To fulfill this requirement, the loop gain must decrease from 10^3 to 10^{-2} between the frequencies $\omega = 2$ to $\omega = 20$, i.e. 100 dB in a decade (slope -5). According to Bode's relation we have $\arg G(i\omega) \approx -5 \cdot 90^\circ$ in this interval. This results in an unstable closed-loop system. We can not fulfill the requirements.

[Go back](#)

7.7

[Go back](#)

8 Controller Structures and Control Design

8.1

(a)

$$\text{RGA}(G(0)) = G(0) \cdot * G^{-T}(0) = \begin{pmatrix} -\frac{5}{7} & \frac{12}{7} \\ \frac{12}{7} & -\frac{5}{7} \end{pmatrix}$$

(b) Avoid the pairs $u_1 \leftrightarrow y_1$ and $u_2 \leftrightarrow y_2$.

[Go back](#)

8.2

$$\text{RGA}(G(s)) = \begin{pmatrix} \frac{3}{s+4} & \frac{s+1}{s+4} \\ \frac{s+1}{s+4} & \frac{3}{s+4} \end{pmatrix}$$

For the frequency zero we get

$$\text{RGA}(G(0)) = \begin{pmatrix} 3/4 & 1/4 \\ 1/4 & 3/4 \end{pmatrix}$$

As all elements in the $\text{RGA}(G(0))$ are positive all combinations are possible.
At the crossover frequency we get

$$\text{RGA}(G(10i)) = \begin{pmatrix} \frac{12-30i}{116} & \frac{104+30i}{116} \\ \frac{104+30i}{116} & \frac{12-30i}{116} \end{pmatrix}.$$

We have elements close to 1 if u_1 controls y_2 and u_2 controls y_1 .

[Go back](#)

8.3

(a)

$$\text{RGA}(G(s)) = \begin{pmatrix} \frac{s-1}{s+1} & \frac{2}{s+1} \\ \frac{2}{s+1} & \frac{s-1}{s+1} \end{pmatrix}$$

yields

$$\text{RGA}(G(0)) = \begin{pmatrix} -1 & 2 \\ 2 & -1 \end{pmatrix}.$$

As we want to avoid pairing corresponding to negative elements in the $\text{RGA}(0)$ we have to choose $u_1 \leftrightarrow y_2$ and $u_2 \leftrightarrow y_1$.

(b) As

$$G(0) = \begin{pmatrix} 1 & -2 \\ 1 & -1 \end{pmatrix}$$

we choose $W_1 = G^{-1}(0)$ and $W_2 = I$. A controller that decouples the system in steady state is thus in the form

$$F(s) = W_1 F^{\text{diag}}(s) W_2 = \begin{pmatrix} -F_{11}(s) & 2F_{22}(s) \\ -F_{11}(s) & F_{22}(s) \end{pmatrix}.$$

Let us verify that the closed-loop system is decoupled statically

```

G = [(s-1)/(s+1) 2/(s+1); 2/(s+1) (s-1)/(s+1)];
G0 = freqresp(G,0)
W1 = inv(G0);

Fd = [2 0; 0 0.1];
F = W1*Fd;
bodemag(G*F)
freqresp(G*F,0)
step(feedback(G*F,eye(2)))

```

[Go back](#)

8.4

Ideally, we want $Q(s)$ to be $Q(s) = G^{-1}(s)$. A realizable choice is:

$$Q(s) = \frac{\tau s + 1}{K(\lambda s + 1)}.$$

This results in

$$F_y(s) = \frac{Q(s)}{1 - Q(s)G(s)} = \frac{\tau}{K\lambda} \left(1 + \frac{1}{\tau s} \right).$$

This is a PI controller with gain $K_{PI} = \frac{\tau}{K\lambda}$ and integration time $T_I = \tau$. The sensitivity function is

$$S(s) = 1 - G(s)Q(s) = \frac{\lambda s}{\lambda s + 1}$$

and the complementary sensitivity function is

$$T(s) = G(s)Q(s) = \frac{1}{\lambda s + 1}.$$

Thus, we have

$$|S(i\omega)| \leq 1 \quad \forall \omega$$

which seems to disagree with Bode's integral theorem. However, as the loop gain GF_y decreases as $1/\omega$ Bode's integral theorem is not applicable.

[Go back](#)

8.5

The system is a nonminimum phase system. We chose to replace the right-half plane zero with a zero mirrored in the imaginary axis.

$$Q(s) = \frac{s^2 + 5s + 6}{(6 + 3s)(\lambda s + 1)}.$$

This results in the controller

$$F(s) = \frac{s^2 + 5s + 6}{s(3\lambda s + 6(\lambda + 1))},$$

which can be rewritten as

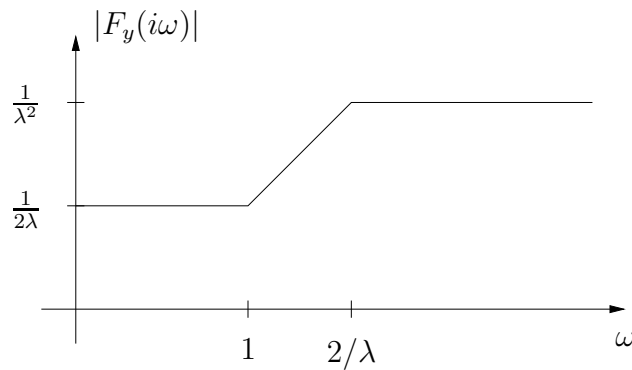
$$F(s) = \frac{5}{6(1 + \lambda)} \left(1 + \frac{6}{5s} + \frac{s}{5} \right) \frac{1}{\frac{3\lambda}{6(\lambda+1)}s + 1}.$$

This is a PID controller with a filter added to make it realizable.

[Go back](#)

8.6

$$\begin{aligned} Q(p) &= G^{-1}(p) \frac{1}{(\lambda p + 1)^2} = \frac{p(p+1)}{(\lambda p + 1)^2} \\ F_y(p) &= (1 - Q(p)G(p))^{-1} Q(p) = \frac{p+1}{\lambda^2 p + 2\lambda} = \frac{1}{2\lambda} \cdot \frac{1+p}{1 + \frac{p}{2/\lambda}} \\ \Rightarrow \quad u &= -\frac{1}{2\lambda} \cdot \frac{1+p}{1 + \frac{p}{2/\lambda}} y \end{aligned}$$



High bandwidth $\Rightarrow \lambda$ small $\Rightarrow F_y(p) \approx \frac{1+p}{2\lambda} \Rightarrow$ PD controller

Go back

8.7

(a) We have the transfer function matrix

$$G(s) = \frac{1}{s/20 + 1} \begin{pmatrix} \frac{9}{s+1} & 2 \\ 6 & 4 \end{pmatrix}.$$

The poles are the least common denominator of the minors

$$\begin{aligned} g_{11}(s) &= \frac{9}{(s/20+1)(s+1)} & g_{12}(s) &= \frac{2}{s/20+1} \\ g_{21}(s) &= \frac{6}{s/20+1} & g_{22}(s) &= \frac{4}{s/20+1} \end{aligned}$$

and the minor

$$\det G(s) = \frac{1}{(s/20 + 1)^2} \left(\frac{36}{s+1} - 12 \right) = \frac{24(1 - s/2)}{(s/20 + 1)^2(s+1)}.$$

This results in the poles -20 , -20 and -1 . The zeros are given by $\det G(s)$ normalized with the pole polynomial. This yields a zero located at $s = 2$. We have to take proper care of the right-half plane zero in the IMC design.

(b)

$$\begin{aligned} G^{-1}(s) &= (s/20 + 1) \begin{pmatrix} \frac{9}{s+1} & 2 \\ 6 & 4 \end{pmatrix}^{-1} \\ &= \frac{(s/20 + 1)(s+1)}{24(-s/2 + 1)} \begin{pmatrix} 4 & -2 \\ -6 & \frac{9}{s+1} \end{pmatrix} \end{aligned}$$

Mirror the right-half plane zero of $G(s)$ in the imaginary axis and add the factor $(\lambda s + 1)$.

$$Q(s) = \frac{(s/20 + 1)(s+1)}{24(\lambda s + 1)(s/2 + 1)} \begin{pmatrix} 4 & -2 \\ -6 & \frac{9}{s+1} \end{pmatrix}$$

This results in the controller

$$F_y(s) = (I - Q(s)G(s))^{-1}Q(s) = \frac{(s/20 + 1)(s + 1)}{24s(\lambda s/2 + \lambda + 1)} \begin{pmatrix} 4 & -2 \\ -6 & \frac{9}{s+1} \end{pmatrix}$$

Go back

8.8

The inverse of the system is:

$$G^{-1}(s) = \frac{1}{-s + 1} \begin{pmatrix} (s + 1)(s + 2) & -3(s + 1)^2 \\ -(s + 1)(s + 2) & 2(s + 1)(s + 2) \end{pmatrix}$$

Mirror the right-half plane zero in the imaginary axis and form $Q(s)$

$$Q(s) = \frac{1}{\lambda s + 1} \frac{-s + 1}{s + 1} G^{-1}(s) = \frac{1}{\lambda s + 1} \begin{pmatrix} s + 2 & -3(s + 1) \\ -(s + 2) & 2(s + 2) \end{pmatrix}$$

The controller is given by $F_y = (I - QG)^{-1}Q$. The corresponding sensitivity function is

$$S = I - GQ = I - \frac{1}{\lambda s + 1} \frac{-s + 1}{s + 1} I = \begin{pmatrix} \frac{s(\lambda s + 2 + \lambda)}{(s + 1)(\lambda s + 1)} & 0 \\ 0 & \frac{s(\lambda s + 2 + \lambda)}{(s + 1)(\lambda s + 1)} \end{pmatrix}$$

$|S(i\omega)| \rightarrow 0$ as $\omega \rightarrow 0 \Rightarrow$ The controller has integral action.

Go back

8.9

- (a) Since system is quadratic, i.e the number of inputs equals the number of outputs, the zeros can be determined as the poles of $G^{-1}(s)$. This gives

$$G^{-1}(s) = \frac{1}{\det G(s)} \begin{pmatrix} \frac{1}{s + 1} & \frac{-3}{s + 2} \\ \frac{-\alpha}{s + 1} & \frac{2}{s + 1} \end{pmatrix} =$$

$$= \frac{1}{s(2-3\alpha)+4-3\alpha} \begin{pmatrix} (s+1)(s+2) & -3(s+1)^2 \\ -\alpha(s+1)(s+2) & 2(s+1)(s+2) \end{pmatrix}$$

The pole polynomial becomes

$$s(2-3\alpha)+4-3\alpha$$

and hence the zero polynomial is given by

$$n(\alpha) = \frac{3\alpha - 4}{2 - 3\alpha}.$$

The location of the zero is shown in the figure below.

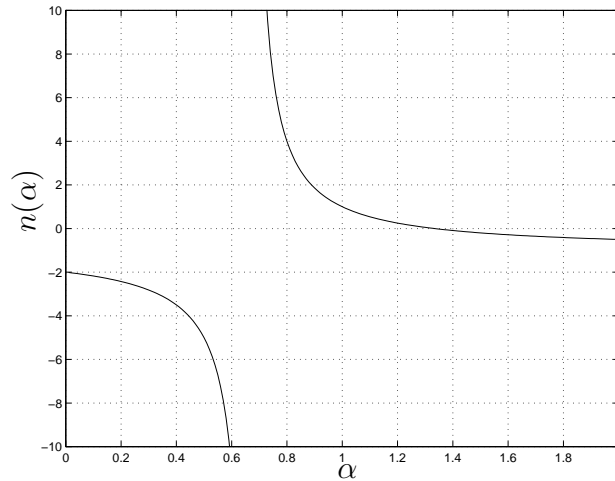


Figure 7: Location of the zero as function of α .

- (b) In order to obtain that $F(s)G(s)$ is diagonal, $F(s)$ has to contain the inverse of $G(s)$. For $2/3 < \alpha < 4/3$ we have $n(\alpha) > 0$, i.e. the zero is located in the right half plane. This means that $F(s)$ is unstable, which should be avoided if possible.
- (c) Using static decoupling F is a constant matrix containing $G(0)^{-1}$. For $\alpha = 4/3$ the inverse does not exist, since the rows of $G(0)$ becomes linear dependent.

Go back

8.10

- (a) The RGA can be computed as follows.

```
>> s = tf('s');
G = [1/(s+2) 2/(s+4); 1/(s+1) 1/(s+2)];
G0 = freqresp(G,0);
RGA = G0.*inv(G0).'
```

```
RGA =
```

```
    -1     2
     2    -1
```

Negative elements in the diagonal of the RGA indicate that it will not be possible the control system using a diagonal regulator.

- (b) The transfer function matrix of the closed loop system is given by

$$G_c(s) = (I + G(s)F(s))^{-1}G(s)F(s)$$

and using Matlab it can be computed as

```
>> F = diag([5 5]);
Gc = feedback(G*F,eye(2));
pole(Gc)
```

```
ans =
```

```
-14.4265
-2.3060 + 1.3712i
-2.3060 - 1.3712i
 0.0385
-14.4265
-2.3060 + 1.3712i
-2.3060 - 1.3712i
 0.0385
```

which means that the poles of the closed loop system are -14.4 , $-2.3 \pm 1.37i$ and 0.04 . Since there is a pole in the right half plane the system is unstable.

Note: $G(s)$ has 4 poles and a constant regulator does not add any poles. The number of poles are doubled by the functions `feedback` and `tf`. The extra poles are removed by using state space representation and the command `minreal`.

```
>> pole(minreal(ss(Gc)))
```

```
ans =
```

```
-14.4265
 0.0385
-2.3060 + 1.3712i
-2.3060 - 1.3712i
```

- (c) The RGA analysis has told us that we should use input 2 to control the first output, and input 1 to control the second output. In other words $U_1(s) = F_1(s)(R_2(s) - Y_2(s))$ and $U_2(s) = F_1(s)(R_1(s) - Y_1(s))$. Consequently, with proportional controls $F_1(s) = F_2(s) = K$ we must

have $U(s) = \begin{pmatrix} 0 & K \\ K & 0 \end{pmatrix}$

```
>> F = [0 5;5 0];
Gc = feedback(G*F,eye(2));
pole(Gc)
```

```
-16.2449 + 0.0000i
-4.8756 + 0.0000i
-1.4398 + 0.9522i
-1.4398 - 0.9522i
```

The closed-loop is thus stable. There are still strong couplings, both dynamically and statically

```
step(Gc)
figure;bodemag(Gc);grid on
```

The problem can also be modified by renumbering the input signals, i.e. switch the columns in $G(s)$. This gives the modified transfer function matrix

$$\bar{G}(s) = \begin{pmatrix} \frac{2}{s+4} & \frac{1}{s+2} \\ \frac{1}{s+1} & \frac{1}{s+2} \end{pmatrix}$$

and the possibility to use the original definition of controls $U_1(s) = K(R_1(s) - Y_1(s))$ and $U_2(s) = K(R_2(s) - Y_2(s))$

```
G = [2/(s+4) 1/(s+2); 1/(s+2) 1/(s+1)];  
F = [5 0;0 5];  
Gc = feedback(G*F,eye(2));  
step(Gc)
```

[Go back](#)

9 Minimization of Quadratic Criteria: LQG

9.1

- (a) We have $A = B = C = N = M = 1$. Thus, the Riccati equation for the Kalman filter is

$$2P + R_1 - \frac{P^2}{R_2} = 0,$$

The positive semidefinite solution is $P = R_2 + R_2\sqrt{1 + \frac{R_1}{R_2}}$. Hence, the Kalman gain is

$$K = \frac{1}{R_2}P = 1 + \sqrt{1 + \frac{R_1}{R_2}} = 1 + \sqrt{1 + \beta}.$$

Analogously we get for the state feedback

$$L = \frac{1}{Q_2}S = 1 + \sqrt{1 + \frac{Q_1}{Q_2}} = 1 + \sqrt{1 + \alpha}.$$

This results in the controller

$$F_y(p) = L(p - 1 + L + K)^{-1}K = \frac{(1 + \sqrt{1 + \alpha})(1 + \sqrt{1 + \beta})}{p + 1 + \sqrt{1 + \alpha} + \sqrt{1 + \beta}}.$$

- (b) The poles of the transfer functions of the closed loop system are given by the eigenvalues of $A - BL$ and $A - KC$ respectively. Thus, the poles are

$$-\sqrt{1 + \alpha} \quad \text{and} \quad -\sqrt{1 + \beta}.$$

A small penalty on u (α large) results in a pole far from the origin in the left-half plane. A large penalty on u results in a pole close to -1, i.e. the pole of the original system mirrored in the imaginary axis. The pole located in $-\sqrt{1 + \beta}$ can be affected in a similar way using the Kalman filter design parameters.

Go back

9.2

(a) Introduce the state $x = z$. Then the system can be written as

$$\begin{aligned}\dot{x} &= -x + u + v, \\ y &= x + e, \\ z &= x,\end{aligned}$$

i.e. $A = -1$, $B = 1$, $M = 1$, $Q_1 = q_1$ and $Q_2 = 1$. $Q_{12} = 0$. We cannot measure the state but we measure y . According to the separation theorem V is minimized if we

- (i) Estimate the state $\hat{x}(t)$ using a Kalman filter.
- (ii) Use feedback according to $u(t) = -L\hat{x}(t)$, where L is computed using LQ theory.

Thus,

$$\dot{\hat{x}} = A\hat{x} + Bu + K(y - C\hat{x})$$

where $K = PC^T R_2^{-1}$ and P is the positive semidefinite solution to

$$AP + PA^T + NR_1 N^T - PC^T R_2^{-1} CP = 0.$$

In our case this is a scalar equation

$$P^2 + 2P - r_1 = 0$$

with the solution

$$P = -1 + \sqrt{1 + r_1},$$

i.e.

$$K = -1 + \sqrt{1 + r_1}.$$

Use the feedback $u = -L\hat{x}$ with $L = Q_2^{-1} B^T S$, where S is the solution to

$$A^T S + SA + M^T Q_1 M - SBQ_2^{-1} B^T S = 0$$

As $M = 1$, $Q_1 = q_1$ and $Q_2 = 1$ we get

$$L = S = -1 + \sqrt{1 + q_1}.$$

The loop gain is

$$\begin{aligned}G(s)F_y(s) &= \frac{1}{s+1} L \frac{1}{1+s+L+K} K = \\ &= \frac{(-1 + \sqrt{1+r_1})(-1 + \sqrt{1+q_1})}{(s+1)(s-1 + \sqrt{1+r_1} + \sqrt{1+q_1})}\end{aligned}$$

- (b) The parameters r_1 and q_1 influence the loop gain in the same way due to symmetry.

(c)

$$G(s)F_y(s) = \frac{(-1 + \sqrt{1+r_1})(-1 + \sqrt{1+q_1})}{(s+1)(s-1 + \sqrt{1+r_1} + \sqrt{1+q_1})}$$

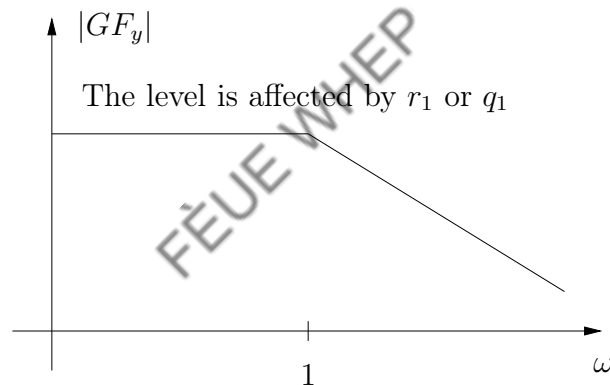
What happens when r_1 or $q_1 \rightarrow \infty$?

$$r_1 \rightarrow \infty \Rightarrow G(s)F_y(s) = \frac{(-1 + \sqrt{1+q_1})}{(s+1) \frac{(s-1 + \sqrt{1+r_1} + \sqrt{1+q_1})}{-1 + \sqrt{1+r_1}}} \rightarrow \frac{-1 + \sqrt{1+q_1}}{s+1}.$$

Analogously we get

$$\lim_{q_1 \rightarrow \infty} G(s)F_y(s) = \frac{-1 + \sqrt{1+r_1}}{s+1}$$

By varying q_1 and/or r_1 we can shape the loop gain according to the sketch below:



[Go back](#)

9.3

State-space form:

$$\dot{x} = \begin{pmatrix} 0 & 1 \\ 0 & 0 \end{pmatrix} x + \begin{pmatrix} 0 \\ 1 \end{pmatrix} u$$

$$z = \begin{pmatrix} 1 & 0 \end{pmatrix} x$$

$$y = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} x$$

The weighting matrices are $Q_1 = 1$ and $Q_2 = \eta$. The Riccati equation is:

$$A^T S + SA + M^T Q_1 M - SBQ_2^{-1}B^T S = 0$$

Define

$$S = \begin{pmatrix} s_1 & s_2 \\ s_2 & s_3 \end{pmatrix},$$

This results in

$$\begin{pmatrix} 0 & 0 \\ s_1 & s_2 \end{pmatrix} + \begin{pmatrix} 0 & s_1 \\ 0 & s_2 \end{pmatrix} + \begin{pmatrix} 1 & 0 \\ 0 & 0 \end{pmatrix} - \frac{1}{\eta} \cdot \begin{pmatrix} s_2^2 & s_2 s_3 \\ s_2 s_3 & s_3^2 \end{pmatrix} = 0$$

The positive definite solution is

$$\begin{aligned} s_1 &= \sqrt{2} \cdot \eta^{1/4} \\ s_2 &= \eta^{1/2} \\ s_3 &= \sqrt{2} \cdot \eta^{3/4} \end{aligned}$$

This yields the optimal feedback

$$\begin{aligned} L &= Q_2^{-1} B^T S = \frac{1}{\eta} \cdot \begin{pmatrix} 0 & 1 \end{pmatrix} \begin{pmatrix} \sqrt{2}\eta^{1/4} & \eta^{1/2} \\ \eta^{1/2} & \sqrt{2} \cdot \eta^{3/4} \end{pmatrix} \\ &= \frac{1}{\eta} \cdot (\eta^{1/2} \quad \sqrt{2}\eta^{3/4}) = (\eta^{-1/2} \quad \sqrt{2} \cdot \eta^{-1/4}) \end{aligned}$$

The poles are the eigenvalues of $A - BL$. Define $\mu = \eta^{-1/4} \Rightarrow L = (\mu^2 \quad \sqrt{2} \cdot \mu)$. This results in

$$0 = \det \begin{pmatrix} s & -1 \\ \mu^2 & s + \sqrt{2} \cdot \mu \end{pmatrix} = s^2 + \sqrt{2}\mu s + \mu^2,$$

i.e.

$$\begin{aligned} s &= -\frac{\mu}{\sqrt{2}} \pm \sqrt{\frac{\mu^2}{2} - \mu^2} = -\frac{\mu}{\sqrt{2}} \pm i \cdot \frac{\mu}{\sqrt{2}} = \\ &= -\frac{\mu}{\sqrt{2}} \cdot (1 \pm i) = -\frac{1}{\sqrt{2} \cdot \eta^{1/4}} \cdot (1 \pm i) \end{aligned}$$

If η decreases the poles will be placed further away from the origin. This results in an increased input signal $u(t)$. Compare this result to the criterion.

Go back

9.4

Description of the system:

$$\begin{aligned}\dot{x}(t) &= \begin{pmatrix} 0 & 1 \\ 0 & -\alpha \end{pmatrix} x(t) + \begin{pmatrix} 0 \\ H \end{pmatrix} \mu(t) + \begin{pmatrix} 0 \\ \nu \end{pmatrix} \tau_d(t) \\ y(t) &= (1 \quad 0) x(t) + v_m(t)\end{aligned}$$

The steady state Riccati equation is

$$0 = S \begin{pmatrix} 0 & 1 \\ 0 & -\alpha \end{pmatrix} + \begin{pmatrix} 0 & 0 \\ 1 & -\alpha \end{pmatrix} S + \begin{pmatrix} 1 & 0 \\ 0 & 0 \end{pmatrix} + S \begin{pmatrix} 0 \\ H \end{pmatrix} \frac{1}{\rho} (0 \quad H) S$$

Component by component we get

$$\begin{aligned}0 &= 1 - \frac{H^2}{\rho} s_{12}^2 \\ 0 &= -\frac{H^2}{\rho} s_{12} s_{22} + s_{11} - \alpha s_{12} \\ 0 &= -\frac{H^2}{\rho} s_{22}^2 + 2s_{12} - 2\alpha s_{22}\end{aligned}$$

The positive definite solution is

$$\begin{aligned}s_{11} &= \frac{\sqrt{\rho}}{H} \sqrt{\alpha^2 + \frac{2H}{\sqrt{\rho}}} \\ s_{12} &= \frac{\sqrt{\rho}}{H} \\ s_{22} &= -\frac{\rho}{H^2} \left(\alpha - \sqrt{\alpha^2 + \frac{2H}{\sqrt{\rho}}} \right)\end{aligned}$$

why the feedback gain L is given by

$$\begin{aligned}L &= Q_2^{-1} B^T S = \frac{1}{\rho} (0 \quad H) \begin{pmatrix} \sqrt{\frac{\rho}{H}} \sqrt{\alpha^2 + \frac{2H}{\sqrt{\rho}}} & \frac{\sqrt{\rho}}{H} \\ \frac{\sqrt{\rho}}{H} & \frac{\rho}{H^2} (-\alpha + \sqrt{\alpha^2 + \frac{2H}{\sqrt{\rho}}}) \end{pmatrix} \\ &= \left(\frac{1}{\sqrt{\rho}} \quad \frac{1}{H} \left(-\alpha + \sqrt{\alpha^2 + \frac{2H}{\sqrt{\rho}}} \right) \right)\end{aligned}$$

The separation theorem states that it is optimal to use the estimated states in the feedback. Thus, the optimal feedback is

$$\mu(t) = - \left(\frac{1}{\sqrt{\rho}} \quad \frac{1}{H} \left(-\alpha + \sqrt{\alpha^2 + \frac{2H}{\sqrt{\rho}}} \right) \right) \hat{x}$$

[Go back](#)

9.5

First find a state-space realization of the system

$$G(s) = \frac{1}{s(s+1)} \quad \Leftrightarrow \quad \ddot{y} + \dot{y} = u$$

Let $x_1 = y$, $x_2 = \dot{y} \Rightarrow$

$$\begin{aligned} \dot{x}(t) &= \begin{pmatrix} 0 & 1 \\ 0 & -1 \end{pmatrix} x(t) + \begin{pmatrix} 0 \\ 1 \end{pmatrix} u(t) \\ y(t) &= \begin{pmatrix} 1 & 0 \end{pmatrix} x(t) + n(t) \end{aligned}$$

We want good robustness properties around the frequency $\omega = 0.5$ rad/s, i.e. we want the magnitude of the complementary sensitivity function $T(s)$ to be small at this frequency. As $T(s)$ is the transfer function from the measurement noise $n(t)$ to the output signal $y(t)$ we can proceed as follows:

If we estimate $\hat{x}(t)$ using the Kalman filter we will minimize the covariance matrix of the estimation error. The model we use for $n(t)$ will tell us for which frequencies the measurements of $y(t)$ are inaccurate. The Kalman filter will suppress measurements at those frequencies, i.e. $|T(i\omega)|$ will be small.

As the Kalman gain K does not influence the closed-loop system $G_c(s)$, we can choose Q_1 and Q_2 to get a desired $G_c(s)$.

If we study the transfer function of the Kalman filter, i.e. the transfer function from $y(t)$ to $\hat{y}(t)$, we get an indication of how the measurement noise affects the suppression.

We want the noise model to have much energy around the frequency $\omega = 0.5$ rad/s. One such model is $n(t) = H(p)w(t)$ where $w(t)$ is white noise, the poles of $H(s)$ are located at $s = -0.01 \pm 0.5i$ and there is a zero at $s = 0$, i.e.

$$H(s) = \frac{K_n s}{s^2 + 0.02s + 0.2501}.$$

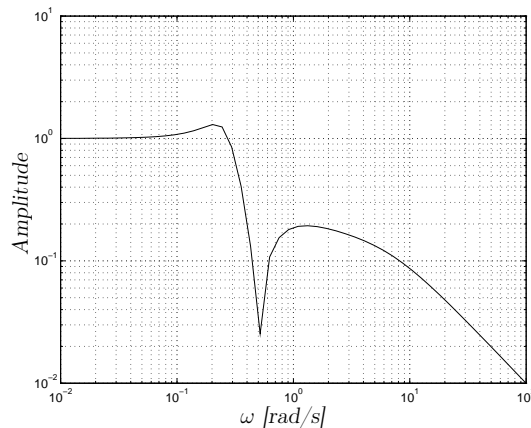
Using controllable canonical form we get

$$\begin{aligned}\dot{x}_n(t) &= \begin{pmatrix} -0.02 & -0.2501 \\ 1 & 0 \end{pmatrix} x_n(t) + \begin{pmatrix} 1 \\ 0 \end{pmatrix} w(t) \\ n(t) &= (K_n \ 0) x_n(t)\end{aligned}$$

Extending the original state-space form with the noise model yields

$$\begin{aligned}\dot{\bar{x}}(t) &= \begin{pmatrix} 0 & 1 & 0 & 0 \\ 0 & -1 & 0 & 0 \\ 0 & 0 & -0.02 & -0.2501 \\ 0 & 0 & 1 & 0 \end{pmatrix} \bar{x}(t) + \begin{pmatrix} 0 \\ 1 \\ 0 \\ 0 \end{pmatrix} u(t) + \begin{pmatrix} 0 \\ 0 \\ 1 \\ 0 \end{pmatrix} w(t) \\ y(t) &= (1 \ 0 \ K_n \ 0) \bar{x}(t).\end{aligned}$$

If this model, with an appropriate value of K_n , is used to compute the Kalman gain K , the magnitude curve of the transfer function from $y(t)$ to $\hat{y}(t) = \hat{x}_1(t)$ will look as in the figure below. Signals at frequencies around $\omega = 0.5$ rad/s are heavily attenuated.



[Go back](#)

9.6

Let G be the system, F the controller, y the output signal and v the disturbance. This results in

$$y = \frac{1}{1 + GF} v = Sv$$

$$u = \frac{F}{1 + GF}v$$

where S is the sensitivity function.

F is chosen such as the criterion

$$J(F) = E\{y(t)^2 + \alpha u(t)^2\}, \quad \alpha > 0$$

is minimized given that $\Phi_v(\omega) = \delta(\omega)$.

The criterion can be written as

$$J(F) = \int_{-\infty}^{\infty} \Phi_y(\omega) + \alpha \Phi_u(\omega) d\omega = \frac{1}{(1 + G_0 F_0)^2} + \alpha \frac{F_0^2}{(1 + G_0 F_0)^2}.$$

where F_0 and G_0 are the stationary gains of the controller and the system respectively. As F minimizes $J(F)$ we have

$$\frac{\partial V}{\partial F} = 0 = \frac{2\alpha F_0 - 2G_0}{(1 + G_0 F_0)^3}$$

which yields $F_0 = G_0/\alpha$. Thus, the sensitivity function at $\omega = 0$ is

$$S_0 = \frac{1}{1 + G_0 F_0} = \frac{1}{1 + G_0^2/\alpha}.$$

Go back

9.7

(a) How to solve the problem:

- (i) Compute the Kalman filter: $\dot{\hat{x}} = A\hat{x} + Bu + K(y - C\hat{x})$
- (ii) Use feedback: $u = -L\hat{x}$, where L is computed using LQ theory.
- (i) + (ii) results in the controller $F_y = L(sI - A + BL + KC - KDL)^{-1}K$.

- (i) State-space description: Let $x_1 = z$, $x_2 = \nu$, $v_1 = v$, $v_2 = e$ and $x = (x_1, x_2)^T$. This results in

$$\begin{aligned}\dot{x} &= \underbrace{\begin{pmatrix} -1 & 1 \\ 0 & -\epsilon \end{pmatrix}}_A x + \underbrace{\begin{pmatrix} 1 \\ 0 \end{pmatrix}}_B u + \underbrace{\begin{pmatrix} 0 \\ 1 \end{pmatrix}}_N v_1 \\ y &= \underbrace{\begin{pmatrix} 1 & 0 \end{pmatrix}}_C x + v_2\end{aligned}$$

Furthermore, we have

$$\begin{aligned}R_1 &= \Phi_{v_1}(\omega) = \Phi_v(\omega) = 1 \\ R_2 &= \Phi_{v_2}(\omega) = \Phi_e(\omega) = 1 \\ R_{12} &= \Phi_{v_1 v_2} = 0\end{aligned}$$

The Kalman filter is given by: $K = PC^T R_2^{-1}$ with P according to

$$AP + PA^T + NR_1 N^T - PC^T R_2^{-1} CP = 0$$

Define $P = \begin{pmatrix} p_1 & p_2 \\ p_2 & p_3 \end{pmatrix}$.

This results in $\lim_{\epsilon \rightarrow 0} P = \begin{pmatrix} \sqrt{3}-1 & 1 \\ 1 & \sqrt{3} \end{pmatrix}$ and

$$K = PC^T R_2^{-1} = \begin{pmatrix} \sqrt{3}-1 \\ 1 \end{pmatrix}$$

- (ii) Compute L such that

$$\min_L \int_0^\infty x_1^2(t) + u^2(t) dt = \min_L \int_0^\infty y^T Q_1 y + u^T Q_2 u dt$$

where $Q_1 = Q_2 = 1$.

The optimal L is given by: $L = Q_2^{-1} B^T S$ where S is the positive semidefinite solution of

$$A^T S + SA + C^T Q_1 C - SBQ_2^{-1} B^T S = 0.$$

Define $S = \begin{pmatrix} s_1 & s_2 \\ s_2 & s_3 \end{pmatrix}$.

This results in $\lim_{\epsilon \rightarrow 0} S = \begin{pmatrix} \sqrt{2}-1 & 1-\frac{1}{\sqrt{2}} \\ 1-\frac{1}{\sqrt{2}} & * \end{pmatrix}$ and

$$L = Q_2^{-1} B^T S = \begin{pmatrix} \sqrt{2}-1 & 1-\frac{1}{\sqrt{2}} \end{pmatrix}$$

The LQG controller is

$$\begin{cases} \dot{\hat{x}} = A\hat{x} + Bu + K(y - C\hat{x}) \\ u = -L\hat{x} \end{cases}$$

with K and L as above.

The static gain of the sensitivity function ($\epsilon = 0$):

(i) and (ii) yields

$$F_y = L(sI - A + BL + KC - KDL)^{-1}K = \begin{Bmatrix} D=0 \\ s=0 \end{Bmatrix} = 1$$

$$S(0) = \frac{1}{1 + F_y(0)G(0)} = \frac{1}{1+1} = \frac{1}{2}.$$

(b) Compute L using LTR(y):

$$\begin{aligned} L_{ltr} &= Q_2^{-1}B^T S \\ S : A^T S + SA + C^T \rho Q_2 C - SBQ_2^{-1}B^T S &= 0 \\ \Rightarrow S &= \begin{pmatrix} \sqrt{1+\rho} - 1 & \frac{\sqrt{1+\rho}-1}{\sqrt{1+\rho+\epsilon}} \\ \frac{\sqrt{1+\rho}-1}{\sqrt{1+\rho+\epsilon}} & * \end{pmatrix} \Rightarrow L_{ltr} = \begin{pmatrix} \sqrt{1+\rho} - 1 & \frac{\sqrt{1+\rho}-1}{\sqrt{1+\rho+\epsilon}} \end{pmatrix} \end{aligned}$$

The static gain of the sensitivity function:

$$S(0) \rightarrow \frac{\epsilon}{\sqrt{3}\epsilon + 1}$$

when $\rho \rightarrow \infty$. It is necessary to let $\epsilon \rightarrow 0$ to get $S(0) \rightarrow 0$, i.e. introducing an integrator into the system.

Go back

9.8

Define the matrices

$$\begin{aligned} \dot{x} &= \overbrace{\begin{pmatrix} 0 & 1 & -1 \\ -\frac{1}{2}w_0^2 & -0.01 & 0.01 \\ \frac{1}{2}w_0^2 & 0.01 & -0.01 \end{pmatrix}}^A x + \overbrace{\begin{pmatrix} 0 \\ w_0 \\ 0 \end{pmatrix}}^B u \\ z &= \underbrace{\begin{pmatrix} 0 & 0 & 1 \end{pmatrix}}_M x, \end{aligned}$$

where

$$w_0^2 = \frac{k}{50}.$$

The Bode plot for $k = 1$ is given. There is a resonance peak at $w_0 \approx 0.14$.

Introduce measurement noise: Let $y = z + v_2$, where v_2 is colored measurement noise. Using the feedback $u = -L\hat{x} + \tilde{p}r$ we can write z as

$$z = G_c r - T v_2 + \tilde{s} v_1.$$

The robustness criterion implies that $|T(i\omega_0)|$ should be small to handle large errors in k . A large spectrum for v_2 at $w = w_0$ will force T to be small at this frequency. Let v_2 be colored noise with a peak in the spectrum at w_0 . This can be achieved by choosing poles in $-0.01 \pm 0.14i$ and a zero in 0, i.e.

$$v_2 = \frac{k_2 p}{p^2 + 0.021p + 0.02} w,$$

where w is white noise.

State-space representation of v_2 :

$$\begin{aligned} \dot{x}_v &= \underbrace{\begin{pmatrix} -0.02 & -0.02 \\ 1 & 0 \end{pmatrix}}_{A_v} x_v + \underbrace{\begin{pmatrix} 1 \\ 0 \end{pmatrix}}_{B_v} w \\ v_2 &= \underbrace{(k_2 \quad 0)}_{C_v} x_v \end{aligned}$$

The extended model is

$$\begin{aligned} \dot{\bar{x}} &= \begin{pmatrix} A & 0 \\ 0 & A_v \end{pmatrix} \bar{x} + \begin{pmatrix} B \\ 0 \end{pmatrix} u + \begin{pmatrix} 0 \\ B_v \end{pmatrix} w, & \bar{x} &= \begin{pmatrix} x \\ x_v \end{pmatrix} \\ y &= (M \quad C_v) \bar{x} \\ z &= (M \quad 0) \bar{x} \\ v_2 &= (0 \quad C_v) \bar{x} \end{aligned}$$

with A , B , M , A_v , B_v , C_v as above.

Go back

9.9

The loop gain is

$$L(sI - A)^{-1}B = \frac{18}{(s-1)(s+2)}$$

The Nyquist curve will approach the origin with the angle -180° . An LQ controller always approaches the origin with -90° .

[Go back](#)

9.10

The system has the following controllability and observability matrices

$$\mathcal{C} = \begin{pmatrix} -4 & -12 \\ 8 & 24 \end{pmatrix}, \quad \mathcal{O} = \begin{pmatrix} 1 & 1 \\ 3 & 3 \end{pmatrix},$$

respectively. Thus, the system is neither controllable nor observable. Since $V(T)$ tends to the quadratic norm of the LQG problem as $T \rightarrow \infty$, we must have $V(\infty) = \infty$.

[Go back](#)

9.11

[Go back](#)

9.12

[Go back](#)

9.13

[Go back](#)

9.14

(a) According to Equation (9.7a) in the text book

$$L = Q_2^{-1} B^T S$$

where

$$0 = A^T S + SA + M^T Q_1 M - SBQ_2^{-1} B^T S$$

Here $A = \alpha$, $B = 1$, $M = 1$, $Q_1 = 1$ and $Q_2 = \rho$. This gives

$$0 = 2\alpha S + 1 - \frac{S^2}{\rho}$$

which implies

$$S^2 - 2\rho\alpha S - \rho = 0$$

with solution

$$S = \rho\alpha \left(\frac{+}{-} \right) \sqrt{(\alpha\rho)^2 + \rho}$$

This gives

$$L = \alpha + \sqrt{\alpha^2 + 1/\rho}$$

(b) Using the result from above gives for the case $\alpha = 1$

$$L = 1 + \sqrt{1 + 1/\rho}$$

i.e. $L \rightarrow 2$, while the case $\alpha = -1$ gives

$$L = -1 + \sqrt{1 + 1/\rho}$$

i.e. $L \rightarrow 0$. For $\alpha = 1$ the open loop system is unstable and has to be stabilized using feedback, and hence $u = 0$ will not work.

[Go back](#)

9.15

We see that the lower of the two curves is positive when the upper is increasing, and turns negative when the upper curves starts to decrease. This is consistent with the lower curve being the derivative of the upper, i.e.. $x_1 = y$ is the upper and $x_2 = \dot{y}$ is the lower.

The choices (iii) and (iv) give the same gain vector L since the weight matrices are the except for a common scaling. Since tuning of weight matrices is about the relative cost of signals being non-zero, simply scaling all costs will not change the optimal way to control the system. The figures (A) and (C) show the same simulation results. The matrices in (i) put less weight on the input u which should imply a faster response, i.e. (B). The choice (ii) puts a high weight on the velocity which should lead to a slower response, i.e. (D).

Answer: (i) – B, (ii) – D, (iii) – A and C, (iv) – A and C

[Go back](#)

9.16

- (a) The system has the poles

$$0 \quad 0 \quad -0.0850 + 0.7435i \quad -0.0850 - 0.7435i$$

Since the system has two poles in the origin the system is not asymptotically stable.

- (b) With

```
Q1 = diag([0 0 0 1]);
Q2 = diag(1);
L = lqr(A,B,Q1,Q2)
```

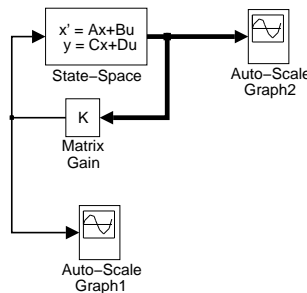
the feedback gain becomes

$$L = (-7.0973 \quad 2.0419 \quad 8.1531 \quad 1)$$

and the poles of the closed loop system (i.e. the eigenvalues of $A - BL$) becomes

$$\begin{aligned} & -0.1809 + 0.8271i \\ & -0.1809 - 0.8271i \\ & -0.4368 + 0.2915i \\ & -0.4368 - 0.2915i \end{aligned}$$

The closed loop system can be simulated using the model below (note, the LTI system is a system with the dynamics $\dot{x} = Ax + Bu$ with all states as output since they are required for the feedback, hence a system defined by the matrices $(A, B, \text{eye}(4), \text{zeros}(4, 1))$. If one wants to look at a specific state, such as the controlled state, one would have to place a matrix gain outside the LTI block to create this signal. Alternatively, we

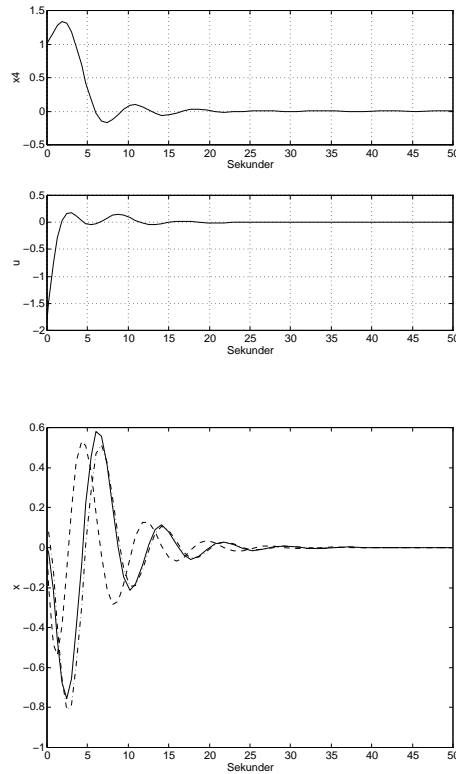


can create and simulate the system, and output the signals of interest, directly in MATLAB

```
Gc_r_to_x = ss(A-B*L,B,eye(4),zeros(4,1))
[x,t] = initial(Gc_r_to_x,x0);
plot(t,x)
% [h;u] = [[0 0 0 1]*x;-L*x + r] = [0 0 0 1;-L]*x + [0;1]*r
Gc_r_to_height_and_u = ss(A-B*L,B,[0 0 0 1;-L],[0;1])
[height_and_u,t] = initial(Gc_r_to_height_and_u,x0);
plot(t,height_and_u)
```

The controlled state x_4 and u are given by the figure and it can be seen that all signals tend to zero. The other states are given in the figure

below, where $x_1 \leftrightarrow$ solid line
 $x_2 \leftrightarrow$ dashed line
 $x_3 \leftrightarrow$ dash-dotted line



- (c) Increasing Q_2 reduces u and decreasing Q_2 increases u .
- (d) An example of matrices that gives a feedback such that the conditions are fulfilled is given by

$$Q_1 = \begin{pmatrix} 250 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 \end{pmatrix}, \quad Q_2 = 1$$

The feedback gain becomes

$$L = (-6.8105 \quad 7.5015 \quad 22.0860 \quad 1)$$

- (e) We are primarily controlling the height which is the fourth state. Hence $z = (0 \ 0 \ 0 \ 1)x = Mx$. Consequently a penalty on $z^T Q_1 z$ (which simply will be $z^2 q_{11}$ for some positive scalar q_{11} as z is scalar) will

$$\text{expand into } (Mx)^T Q_1 Mx = x^T \begin{pmatrix} 0 \\ 0 \\ 0 \\ 1 \end{pmatrix} q_{11} (0 \ 0 \ 0 \ 1) x = x^T \begin{pmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & q_{11} \end{pmatrix} x$$

[Go back](#)

9.17

- (a) The requirements are fulfilled by the state feedback

$$u(t) = -Lx(t)$$

where

$$L = (-4.4721 \quad -4.9405 \quad 19.1028 \quad 6.1811)$$

which is achieved from LQ-minimization using

$$Q_1 = \begin{pmatrix} 20 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{pmatrix} \quad Q_2 = 1$$

The poles of the closed loop system, i.e the eigenvalues of $A - BL$, all have the absolute value 2.36.

- (b) The closed loop system has the characteristic equation

$$\lambda^4 + l_4\lambda^3 + l_3\lambda^2 - 7l_2\lambda - 7l_1 = 0$$

The closed loop system is stable if all roots are located (strictly) in the left half of the complex plane. A necessary condition (but not sufficient) is that all coefficients in the polynomial are strictly positive. A loss of a measurement of state variable can be interpreted as $l_i = 0$ for some i , and this violates the condition.

[Go back](#)

10 Loop Shaping

10.1

The criterion to minimize is the \mathcal{H}_2 norm of G_{ec} . The system on state-space form:

$$\begin{aligned}\dot{x}_1 &= -x_1 + u \\ y &= x_1\end{aligned}$$

Weighting functions

$$W_u(s) = 5, \quad W_T(s) = 0.5, \quad W_S(s) = \frac{1}{s}$$

Form the extended system G_0 :

$$\begin{aligned}z_1 &= W_u u = 5u \\ z_2 &= W_T G u = 0.5x_1 \\ z_3 &= W_S(Gu + w) = x_2\end{aligned}$$

where x_2 is a new state defined as

$$x_2 = \frac{1}{p}(Gu + w) \quad \Leftrightarrow \quad \dot{x}_2 = x_1 + w$$

This yields

$$\begin{aligned}\dot{x} &= \begin{pmatrix} -1 & 0 \\ 1 & 0 \end{pmatrix} x + \begin{pmatrix} 1 \\ 0 \end{pmatrix} u + \begin{pmatrix} 0 \\ 1 \end{pmatrix} w \\ z &= \begin{pmatrix} 0 & 0 \\ 0.5 & 0 \\ 0 & 1 \end{pmatrix} x + \begin{pmatrix} 5 \\ 0 \\ 0 \end{pmatrix} u \\ y &= (1 \ 0) x + w\end{aligned}$$

Check M and D

$$D^T (M \ D) = \begin{pmatrix} 5 & 0 & 0 \end{pmatrix} \begin{pmatrix} 0 & 0 & 5 \\ 0.5 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix} = \begin{pmatrix} 0 & 0 & 25 \end{pmatrix} \neq \begin{pmatrix} 0 & 0 & 1 \end{pmatrix}$$

Hence, define a new input signal \tilde{u} as

$$\tilde{u} = (D^T D)^{1/2} u + (D^T D)^{-1/2} D^T M x = 5u \quad \Leftrightarrow \quad u = \frac{1}{5} \tilde{u}$$

This is just a scaling of the original input signal. Thus we get a new B matrix

$$\tilde{B} = \frac{1}{5} B$$

Solve the Riccati equation: $A^T S + S A + M^T M - S \tilde{B} \tilde{B}^T S = 0$

Define

$$S = \begin{pmatrix} s_1 & s_2 \\ s_2 & s_3 \end{pmatrix}$$

which yields

$$\begin{pmatrix} -s_1 + s_2 & -s_2 + s_3 \\ 0 & 0 \end{pmatrix} + \begin{pmatrix} -s_1 + s_2 & 0 \\ -s_2 + s_3 & 0 \end{pmatrix} + \begin{pmatrix} 0.25 & 0 \\ 0 & 1 \end{pmatrix} - \frac{1}{25} \begin{pmatrix} s_1^2 & s_1 s_2 \\ s_1 s_2 & s_2^2 \end{pmatrix} = 0$$

Hence,

$$\begin{cases} -2s_1 + 2s_2 + 0.25 - \frac{1}{25}s_1^2 = 0 \\ -s_2 + s_3 - \frac{1}{25}s_1 s_2 = 0 \\ 1 - \frac{1}{25}s_2^2 = 0 \end{cases}$$

which has the positive semidefinite solution $s_1 = 4.686$, $s_2 = 5$ and $s_3 \geq \frac{5^2}{4.686}$. Thus, the state feedback for the scaled system is

$$\tilde{L} = \tilde{B}^T S = \left(\frac{1}{5}s_1 \quad \frac{1}{5}s_2 \right) = (0.937 \quad 1)$$

For the original system we get

$$L = \frac{1}{5} \tilde{L} = (0.187 \quad 0.2)$$

The controller is

$$\begin{aligned} \dot{\hat{x}} &= A\hat{x} + Bu + N(y - C\hat{x}) \\ u &= -L\hat{x} \end{aligned}$$

Go back

10.2

The criterion to minimize is the \mathcal{H}_∞ norm of G_{ec} . The extended system is the same as in the previous exercise. The controller is $L = B^T S$, for the smallest value of γ for which

$$A^T S + SA + M^T M + S(\gamma^{-2} NN^T - BB^T)S = 0$$

has a positive semidefinite solution. If we solve this numerically using MATLAB we see that $\gamma \geq 5.12$ produce positive definite solutions. For $\gamma = 5.2$ we get

$$L = (2.6873 \quad 2.7632)$$

10.3

- (a) The frequency weights $W_S = \frac{1}{s}$ and $W_T = W_u = 1$ result in

$$z_1 = W_u u = u$$

$$z_2 = W_T G u = Cx$$

$$z_3 = W_S(Gu + w) \quad \Leftrightarrow \quad \dot{z}_3 = Cx + w = y$$

and

$$\begin{pmatrix} \dot{x} \\ \dot{z}_3 \end{pmatrix} = \begin{pmatrix} A & 0 \\ C & 0 \end{pmatrix} \begin{pmatrix} x \\ z_3 \end{pmatrix} + \begin{pmatrix} B \\ 0 \end{pmatrix} u + \begin{pmatrix} 0 \\ 1 \end{pmatrix} w$$

$$y = (C \quad 0) \begin{pmatrix} x \\ z_3 \end{pmatrix} + w$$

Controllers for the \mathcal{H}_2 and \mathcal{H}_∞ cases can be computed using the expressions in the textbook.

- (b) The observer is given by

$$\dot{\hat{x}} = A\hat{x} + Bu \quad \Leftrightarrow \quad \hat{x} = (pI - A)^{-1}Bu$$

$$\dot{\hat{z}}_3 = C\hat{x} + (y - C\hat{x}) = y \quad \Leftrightarrow \quad \hat{z}_3 = \int y d\tau$$

The state feedback is

$$u = -(L \quad -\alpha) \begin{pmatrix} \hat{x} \\ \hat{z}_3 \end{pmatrix} = -L\hat{x} + \alpha\hat{z}_3 = -L(pI - A)^{-1}Bu + \alpha \int y d\tau$$

By solving for u we get the desired controller structure

$$u = \frac{\alpha}{1 + L(pI - A)^{-1}B} \int y d\tau$$

- (c) If the system contains an integrator we have $\det(pI - A) = p \cdot \xi(p)$ which implies that

$$u = \frac{\alpha}{1 + \frac{1}{p\xi(p)}L(pI - A)^a B} \cdot \frac{1}{p}y = \frac{\alpha\xi(p)}{p\xi(p) + L(pI - A)^a B}y$$

i.e. the integral part of the controller is cancelled.

Go back

10.4

Go back

10.5

A normal requirement is to have a small sensitivity function for low frequencies, which means that C is excluded. Bode's integral theorem states that it is impossible to achieve that $|S(i\omega)| < 1 \quad \forall \omega$, which excludes B. Therefore A is the best alternative.

Go back

10.6

- (a) Straightforward calculations give

$$S(s) = \frac{1}{1 + G(s)F(s)} = \frac{1}{1 + \frac{1}{s+1}K} = \frac{s+1}{s+1+K},$$

$$T(s) = \frac{G(s)F(s)}{1 + G(s)F(s)} = \frac{\frac{1}{s+1}K}{1 + \frac{1}{s+1}K} = \frac{K}{s+1+K} \quad \text{and}$$

$$G_{ru}(s) = \frac{F(s)}{1 + G(s)F(s)} = \frac{K}{1 + \frac{1}{s+1}K} = \frac{K(s+1)}{s+1+K}$$

- (b)
 - In Alternative II it is required that both S and T are small for low frequencies. Since $S + T = 1$ always holds this requirement is not realistic.
 - In alternative III the requirement on T is the opposite to a correct specification for the complementary sensitivity function. Normally one requires, for robustness and measurement noise reasons, that T is small for high frequencies.
 - Alternative I is realistically specified with small S for low frequencies and small T for high frequencies.
- (c) The requirements can be fulfilled by e.g choosing $K = 9$.

[Go back](#)

10.7

[Go back](#)

12 Stability of Nonlinear Systems

12.1

The state variables $x_1 = y$ and $x_2 = \dot{y}$ yield

$$\dot{x} = \begin{pmatrix} x_2 \\ -0.2(1 + x_2^2)x_2 - x_1 \end{pmatrix} = f(x),$$

and

$$\dot{V} = V_x f(x) = x_1 \dot{x}_1 + x_2 \dot{x}_2 = -0.2x_2^2(1 + x_2^2).$$

Hence, $\dot{V} < 0$ except when $x_2 = 0$. If $x_2 \equiv 0$ we have that $x_1 = \text{constant} = 0$. Thus the zero solution is asymptotically stable.

[Go back](#)

12.2

No, $V(x) \geq 0, \forall x$ is not fulfilled.

[Go back](#)

12.3

The slopes

$$\begin{cases} k_1 = 0.5 \\ k_2 = 3 \end{cases} \Rightarrow$$

result in a circle going through the points $-1/3$ and -2 .

[Go back](#)

12.4

The nonlinearity can be bounded by two lines with slope $k_1 = 0$ and $k_2 = 1$ respectively. This leads to a degenerate (infinitely large) circle intersecting the real line at $-\infty$ and -1 . According to the circle criterion the closed loop system is thus asymptotically stable if the Nyquist curve of the open loop gain $G(i\omega)$ lies to the right of the line $s = -1$.

[Go back](#)

12.5

The circle criteria applies to a model with one nonlinear static block, and one linear dynamic block. Here, we have two linear blocks, so our first step is to convert it to the standard case.

To begin with, we note that we cannot simply move the constant K straight through the nonlinear block and put it next to the linear dynamics, since $Kf(u_1) \neq f(Ku_1)$. Nevertheless, we can move it to the linear dynamics block, but we have to go the other way around.

Let z denote the output from the linear dynamics and call the linear dynamics $G(s) = \frac{1}{s(s+1)}$. We have $Z(s) = G(s)U_2(s)$, $u_1 = K(r - z)$, $u_2 = f(u_1)$. This can be written as $u_1 = Kr - Kz$, i.e., $u_1 = \tilde{r} - y$ if we define $\tilde{r} = Kr$ and $Y(s) = KZ(s) = (KG(s))U_2(s) = G_o(s)U_2(s)$. In this form, we have $u_2 = f(\tilde{r} - y)$, $Y(s) = G_o(s)U_2(s)$ which is exactly the type of loop analyzed using the circle criteria.

This was just a long-winded way of proving that you can pull K along the loop backwards and multiply it with the linear dynamics, remembering that when you pull it through the loop, you are redefining the external variable.

The nonlinearity is bounded by two lines with slope $k_1 = 0$ and $k_2 = \infty$ respectively. The circle will thus cover the entire left half plane, and to satisfy the circle criterion the linear dynamics must thus have a strictly positive real part. The linear loop gain is $G_o(s) = KG(s)$ and after simplifying the expression $KG(i\omega) = \frac{K}{(i\omega)(i\omega+1)} = \frac{K(-i\omega)(-i\omega+1)}{(i\omega)(i\omega+1)(-i\omega)(-i\omega+1)}$ to find the real part,

we arrive at

$$\operatorname{Re} G_o(i\omega) = -\frac{K}{\omega^2 + 1} < 0, \quad \forall \omega$$

Hence, stability cannot be guaranteed for any $K > 0$.

Notice that there is really no need to simplify as far as done here, as you know the denominator is a positive real number as it is an absolute value $zz^* = |z|^2$. All you have to do is to study the real part of the numerator after having multiplied with the conjugate of the denominator, since all you need is the sign of this term.

Go back

12.6

The state variables $x_1 = \Phi, x_2 = \dot{\Phi}$ yield

$$\begin{aligned} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= -\frac{mg}{J} \sin x_1 l \end{aligned}$$

The controller

$$l = l_0 + \varepsilon \Phi \dot{\Phi}$$

results in

$$\begin{aligned} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= -\frac{mg}{J} \sin x_1 (l_0 + \varepsilon \Phi \dot{\Phi}) \end{aligned}$$

As a candidate Lyapunov function we use

$$V(x) = \frac{1}{2} J x_2^2 + mgl_0(1 - \cos x_1)$$

which corresponds to the energy of the system. We get

$$\dot{V} = Jx_2\dot{x}_2 + mgl_0 \sin x_1 \dot{x}_1 = -\varepsilon mgx_2^2 x_1 \sin x_1 \leq 0 \quad (-\pi/2 < x_1 \pi/2)$$

with $\dot{V} = 0$ only when $x_1 \equiv 0$ or $x_2 \equiv 0$. $x_1 \equiv 0 \Rightarrow x_2 = 0$ and $x_2 \equiv 0 \Rightarrow x_1 = 0$.

Answer: $\dot{V}(x) < 0$ for all $x_1 \neq 0, x_2 \neq 0$ implies that $x \rightarrow 0$.

[Go back](#)

12.7

The states $x_1 = y, x_2 = \dot{y}$ and the controller

$$u = -\operatorname{sgn}(ax_1 + bx_2)$$

yield

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= -2x_1 - 3x_2 - \operatorname{sgn}(ax_1 + bx_2)\end{aligned}$$

The Lyapunov function candidate

$$V(x) = \left(\frac{\alpha}{2}x_1^2 + \frac{\beta}{2}x_2^2\right)$$

results in

$$\dot{V} = (\alpha - 2\beta)x_1x_2 - 3\beta x_2^2 - \beta x_2 \operatorname{sgn}(ax_1 + bx_2)$$

Take, for example, $\alpha = 2, \beta = 1, a = 0, b = 1$, which result in

$$\dot{V} = -3x_2^2 - |x_2| \leq 0$$

[Go back](#)

12.8

The nonlinearity (which is symmetric around the origin) is

$$f(u) = u + \arctan(u)$$

The derivative of $f(u)$ is

$$f'(u) = 1 + \frac{1}{1+u^2}$$

and has its maximum $f'(0) = 2$ for $u = 0$, while $f'(u) \rightarrow 1$ as $|u| \rightarrow \infty$. This implies that the most critical part of the curve to bound from above is in the origin where the slope is 2, and the critical part to bound from below is when the function tends to infinity with a slope of 1 (draw!).

```
u = -5:0.01:5;
plot(u,u+atan(u),u,1*u,u,2*u)
```

We can write the linear bounds as

$$1 \leq \frac{u + \arctan(u)}{u} \leq 2$$

The conclusion is that the Nyquist curve of the linear part of the system must lie outside and not encircle the circle passing through -1 and $-1/2$.

[Go back](#)

12.9

According to the circle criterion the system is stable if the Nyquist curve lies outside and does not encircle the circle passing through $-4/3$ and $-4/7$.

According to the textbook the loop gain for an LQ controller lies outside and does not encircle the circle passing through 0 and -2 . As this circle encompasses the above smaller circle the system is stable.

[Go back](#)

12.10

- (a) A stationary point requires all derivatives to be 0. Hence the first equation tells us $y = x$ in all stationary points. Using this in the second equation leads to $x(\rho - z) - x = 0$ which means $x = 0$ or $\rho - z = 1$. If

$x = 0$ we have $y = 0$ and the last equation yields $z = 0$. If $z = \rho - 1$ the last equation yields $x = \pm\sqrt{\beta(\rho - 1)}$ and $y = \pm\sqrt{\beta(\rho - 1)}$ (assuming $\rho \geq 1$). Hence, we have three possible solutions, $(0, 0, 0)$, $(\sqrt{\beta(\rho - 1)}, \sqrt{\beta(\rho - 1)}, \rho - 1)$ and $(-\sqrt{\beta(\rho - 1)}, -\sqrt{\beta(\rho - 1)}, \rho - 1)$.

- (b) The Jacobian is given by
$$\begin{pmatrix} -\sigma & \sigma & 0 \\ \rho - z & -1 & -x \\ y & x & -\beta \end{pmatrix}$$
- (c) Numerical computation of eigenvalues of the linearized system in the three stationary points reveals that the system is locally asymptotically unstable in one of the stationary points, and stable in two. Read more about the system here https://en.wikipedia.org/wiki/Lorenz_system.

```
sigma = 1;
rho = 2;
beta = 3;

x0 = [0;0;0];
A = [-sigma sigma 0;
rho-x0(3) -1 -x0(1);
x0(2) x0(1) -beta];
eig(A) % Unstable
x0 = [sqrt(beta*(rho-1));sqrt(beta*(rho-1));rho-1]
A = [-sigma sigma 0;
rho-x0(3) -1 -x0(1);
x0(2) x0(1) -beta];
eig(A) % Stable
x0 = [-sqrt(beta*(rho-1));-sqrt(beta*(rho-1));rho-1]
A = [-sigma sigma 0;
rho-x0(3) -1 -x0(1);
x0(2) x0(1) -beta];
eig(A) % Stable
```

[Go back](#)

12.11

The nonlinearity can be enclosed in a sector with $k_1 = 0$ and $k_2 = 1/2$, meaning the forbidden circle will be the region to the left of -2 . By drawing the nyquist curve of $G(s)$ we note that it never moves to the left of $-1/8$ meaning the closed-loop is guaranteed to be stable for $K = 1$. For this to hold also when we use larger K , and thus enlarge the nyquist curve of the loop gain, we must have $K \cdot \frac{-1}{8} > -2$ which simplifies to $K < 16$.

To find the value $-1/8$ analytically, we first expand $G(i\omega) = \frac{1}{(i\omega+1)^2} = \frac{(1-i\omega)^2}{(\omega^2+1)^2} = \frac{1-2i\omega-\omega^2}{(\omega^2+1)^2}$. The real part is $R(\omega) = \frac{1-\omega^2}{(\omega^2+1)^2} = \frac{n(\omega)}{d(\omega)}$, and we must find the minimum of this function, i.e., a point where the derivative is 0. Using the hint, this means we must solve $(\frac{d}{d\omega}n(\omega))d(\omega) - n(\omega)(\frac{d}{d\omega}d(\omega)) = 0$. This expression expands to $(-2\omega)(1+\omega^2)^2 - (1-\omega^2)(2(1+\omega^2)2\omega) = 0$ which factors to $-2\omega(1+\omega^2)((1+\omega^2)+2(1-\omega^2)) = 0$ and simplifies to $-2\omega(1+\omega^2)(3-\omega^2) = 0$. Hence, the solution of interest is $\omega = \sqrt{3}$. Plug this into the $R(\omega)$ and the final value is $-1/8$!

Go back

13 Phase Plane Analysis

13.1

a) From

$$\ddot{y} - (0.1 - \frac{10}{3}\dot{y}^2)\dot{y} + y + y^2 = 0$$

a reasonable guess is to try the states $x_1 = y$ and $x_2 = \dot{y}$. With these, we can write the system on state-space form

$$\begin{aligned}\dot{x}_1 &= x_2 = f_1(x_1, x_2) \\ \dot{x}_2 &= -x_1(1 + x_1) + x_2(0.1 - 10x_2^2/3) = f_2(x_1, x_2)\end{aligned}$$

b) To find the stationary points, we must find solutions to $\dot{x}(t)$. We call the stationary points \bar{x} here.

$$f(\bar{x}) = 0 \quad \Rightarrow \quad \bar{x}_2 = 0 \quad \text{and} \quad \bar{x}_1(1 + \bar{x}_1) = 0.$$

$$\text{SP I: } \begin{cases} \bar{x}_1 = 0 \\ \bar{x}_2 = 0 \end{cases}, \quad \text{SP II: } \begin{cases} \bar{x}_1 = -1 \\ \bar{x}_2 = 0 \end{cases}$$

c) To linearize the system around the stationary point we use

$$f(x) \approx f(\bar{x}) + f_x(\bar{x})(x - \bar{x}) = f_x(\bar{x})(x - \bar{x})$$

as $f(\bar{x}) = 0$. The matrix $f_x(x)$ is the Jacobian of f . The ij entry is $\frac{\partial f_i}{\partial x_j}(x)$:

$$\begin{aligned}\frac{\partial f_1}{\partial x_1} &= 0, & \frac{\partial f_1}{\partial x_2} &= 1, \\ \frac{\partial f_2}{\partial x_1} &= -1 - 2x_1, & \frac{\partial f_2}{\partial x_2} &= 0.1 - 10x_2^2.\end{aligned}$$

Make the change of variables $z = x - \bar{x}$.

SP I:

Linear approximation $\dot{z} = Az$, with

$$A = \begin{pmatrix} 0 & 1 \\ -1 & 0.1 \end{pmatrix}$$

The eigenvalues of the matrix A are given by

$$0 = \det(\lambda I - A) = \lambda(\lambda - 0.1) + 1,$$

i.e. we have unstable dynamics since we have positive real part on both eigenvalues.

$$\lambda = 0.05 \pm \sqrt{0.05^2 - 1}$$

SP II:

Linear approximation $\dot{z} = Bz$, where

$$B = \begin{pmatrix} 0 & 1 \\ 1 & 0.1 \end{pmatrix}$$

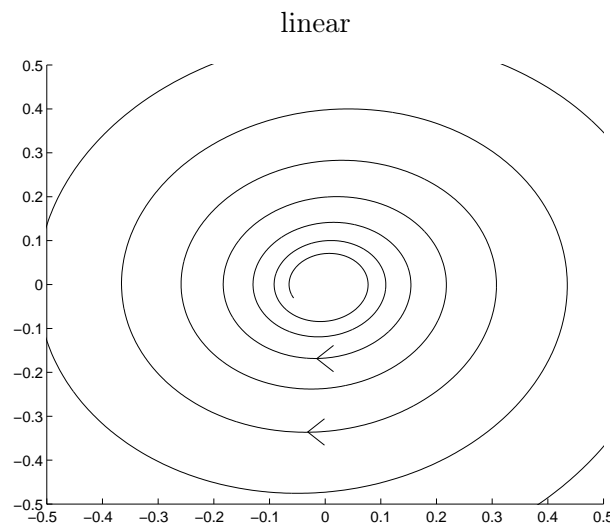
The eigenvalues of B ,

$$\begin{aligned} 0 &= \det(\lambda I - B) = \lambda(\lambda - 0.1) - 1 \\ \lambda &= 0.05 \pm \sqrt{0.05^2 + 1}, \quad \lambda_1 = -0.95, \quad \lambda_2 = 1.05 \end{aligned}$$

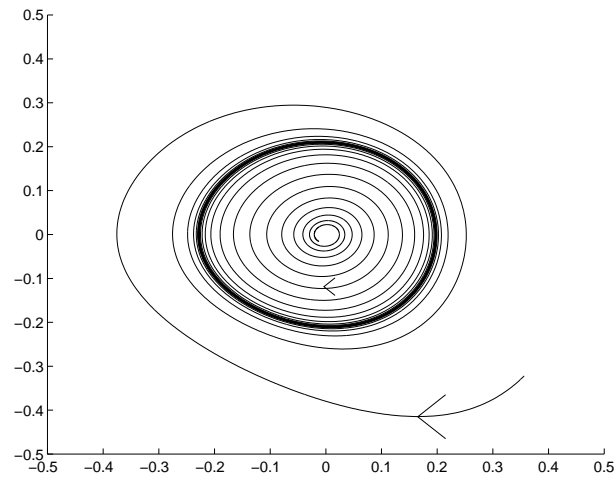
Once again unstable, since one of the eigenvalues is positive.

d) **SP I :**

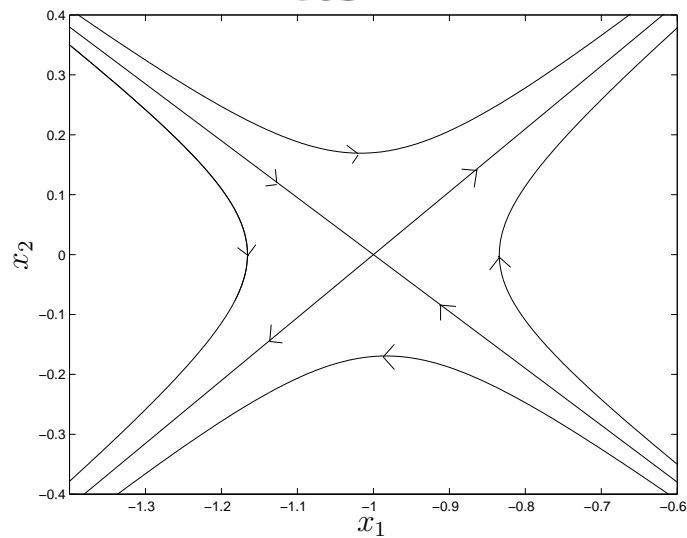
The linear approximation has an unstable focus at $(0,0)$. For unstable nodes the the nonlinear differential equation has a stationary of the same type as the linear approximation. Note that the linear approximation is only valid *close* to the stationary point.



nonlinear

**SP II:**

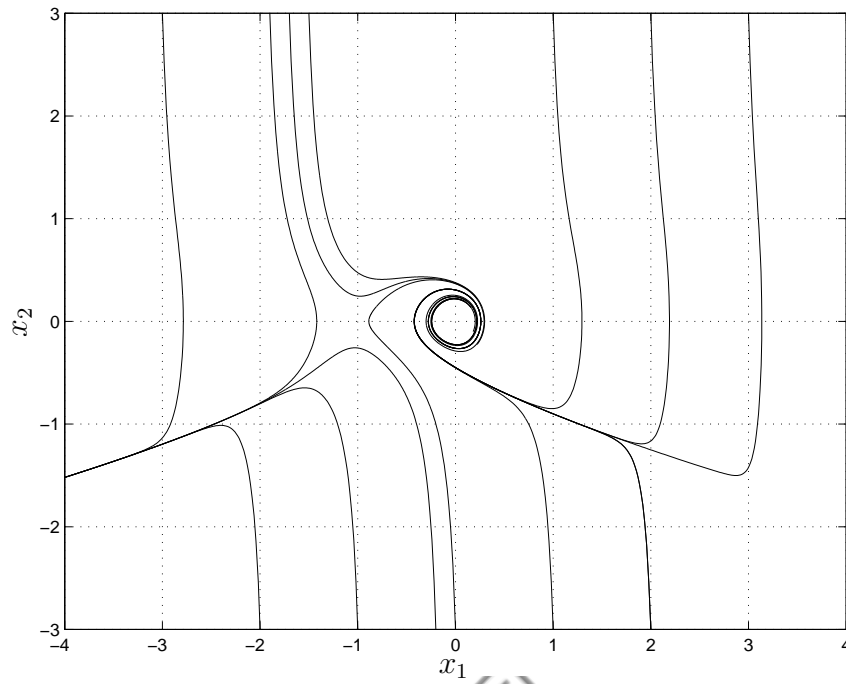
The linearized approximation has a saddle point at $(-1,0)$. This is true for the nonlinear differential equation too. The eigenvector corresponding to the stable eigenvalue is $(1, -0.95)$, and the eigenvalue corresponding to the unstable eigenvalue is $(1, 1.05)$.

**Far away from the stationary points:**

How do the trajectories behave far from the origin? Form the derivative,

$$\frac{dx_2}{dx_1} = \frac{\dot{x}_2}{\dot{x}_1} = \frac{-x_1(1+x_1) + x_2(0.1 - 10x_2^2/3)}{x_2}$$

When x_1 is bounded and $x_2 \rightarrow \pm\infty$, we have that $\dot{x}_2/\dot{x}_1 \rightarrow \infty$. Hence, the trajectories are vertical when $|x_2|$ grows and also when $x_2 \rightarrow 0$.



[Go back](#)

13.2

The nonlinearity is described by

$$f(x) = \begin{cases} x + a, & x < -a \\ 0, & -a \leq x \leq a \\ x - a, & x > a \end{cases}$$

The relationship between x and e is given by

$$p(p + B)x(t) = Ke(t)$$

i.e.

$$\ddot{x} + B\dot{x} = Ke$$

In addition we have that

$$e = u - f(x) = -f(x)$$

which yields

$$\ddot{x} + B\dot{x} + Kf(x) = 0$$

Introduce the states $x_1 = x$ and $x_2 = \dot{x}$. The state-space form is

$$\dot{x}_1 = x_2 \quad (13)$$

$$\dot{x}_2 = -Kf(x_1) - Bx_2 \quad (14)$$

Partition into regions where $f(x)$ is linear.

1. The region $x_1 < -a$:

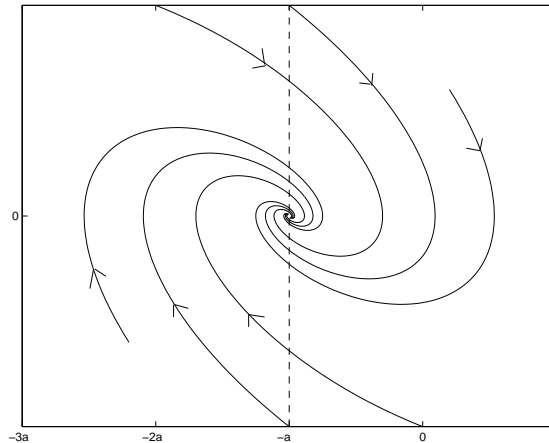
The stationary point is $x_1 = -a, x_2 = 0$. The change of variables $z_1 = x_1 + a, z_2 = x_2$ results in linear state-space equations, $\dot{z} = Az$, where

$$A = \begin{pmatrix} 0 & 1 \\ -K & -B \end{pmatrix}$$

The eigenvalues of A are

$$\lambda = -\frac{B}{2} \pm \sqrt{\frac{B^2}{4} - K}$$

Thus, the point $x = (-a, 0)$ is a stable node or a stable focus. Sketching the phase portrait when $4K > B^2$, results in a stable focus.



2. The region $-a \leq x_1 \leq a$:

The stationary points are line segments: $-a \leq x_1 \leq a$ and $x_2 = 0$. The dynamic equations are

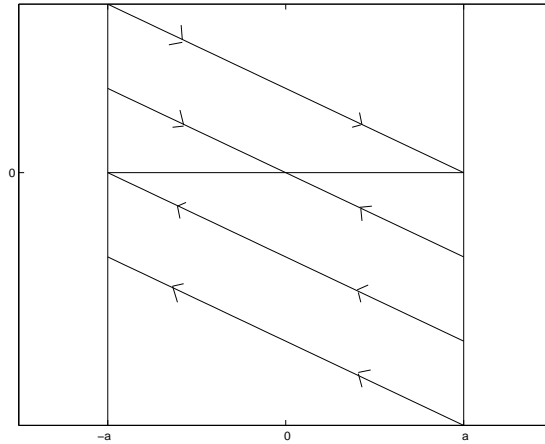
$$\dot{x}_1 = x_2$$

$$\dot{x}_2 = -Bx_2$$

Form the derivative

$$\frac{dx_2}{dx_1} = \frac{\dot{x}_2}{\dot{x}_1} = -B$$

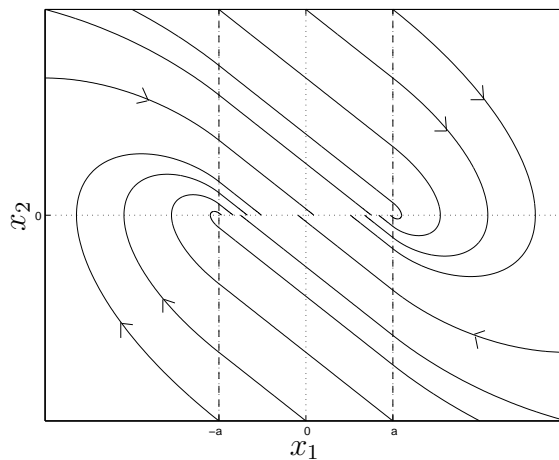
In the entire region the trajectories have the slope $-B$.



3. The region $x_1 > a$:

stationary point $x_1 = a, x_2 = 0$. Make the change of variables $z_1 = x_1 - a, z_2 = x_2$. This results in the same state-space equation for z as for the case $x_1 < -a$, $\dot{z} = Az$. If $4K > B^2$, we have the unstable focus $x = (a, 0)$ analogously to the case $x_1 < -a$.

To get the phase portrait we need to join the three partial solutions found in 1, 2 and 3.



[Go back](#)

13.3

(a) Introduce $x_1 = y, x_2 = \dot{y}$, which yields

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= -\operatorname{sgn} x_1\end{aligned}$$

There are no stationary points. The phase portrait is constructed in two steps depending on $\operatorname{sgn} x_1$. When $x_1 > 0$ we have

$$\frac{dx_2}{dx_1} = -\frac{1}{x_2}$$

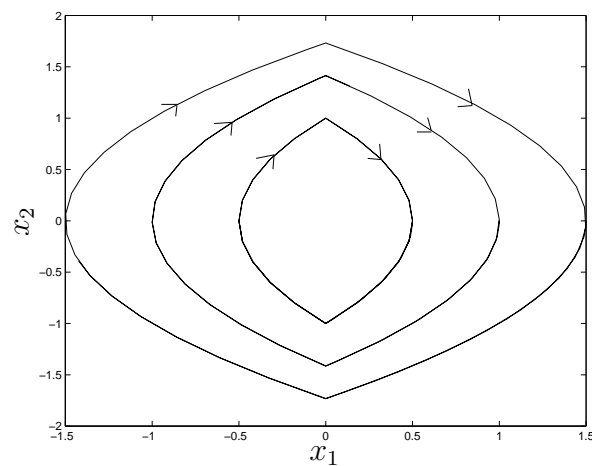
This separable differential equation has the solution

$$\frac{1}{2}x_2^2 + x_1 = \text{constant}$$

i.e. x_1 as a function of x_2 is a number of parabolas. When $x_1 < 0$ we get analogously

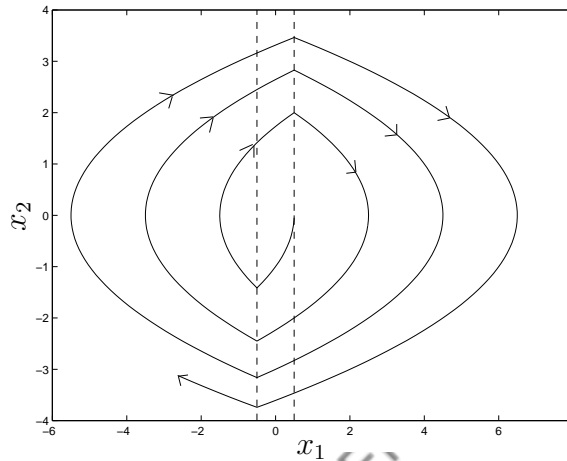
$$\frac{1}{2}x_2^2 - x_1 = \text{constant}$$

The phase portrait:



- (b) For $x_1 > a$ we have $\frac{1}{2}x_2^2 + x_1 = \text{konst}$ and for $x_1 < -a$ we have $\frac{1}{2}x_2^2 - x_1 = \text{konst}$. For the case $|x_1| \leq a$ the relay will have the same output as before it entered the region, i.e. the parabola is continuing.

The phase portrait when $a = 0.5$:



[Go back](#)

13.4

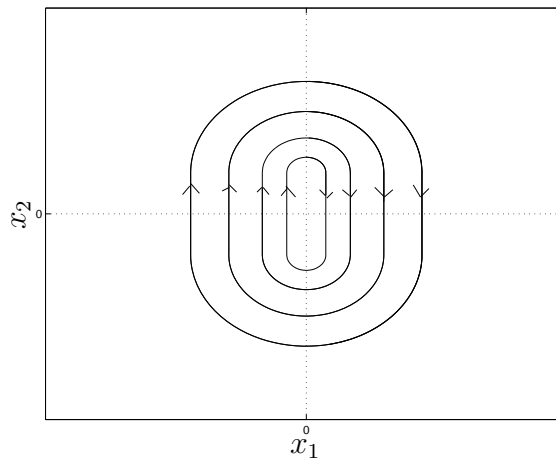
- (a) Let $x_1 = y$ and let x_2 be the input to the nonlinearity. This results in the state-space form

$$\begin{aligned}\dot{x}_1 &= f(x_2) \\ \dot{x}_2 &= -x_1\end{aligned}$$

where

$$f(x) = \begin{cases} x + 1, & x < -1 \\ 0, & -1 \leq x \leq 1 \\ x - 1, & x > 1 \end{cases}$$

We get centers in the stationary points $x = (0, -1)$ (for $x_2 \leq -1$) and $x = (0, 1)$ (for $x_2 \geq 1$). When $-1 < x_2 < 1$ we have $x_1 = \text{constant}$. This results in the phase portrait

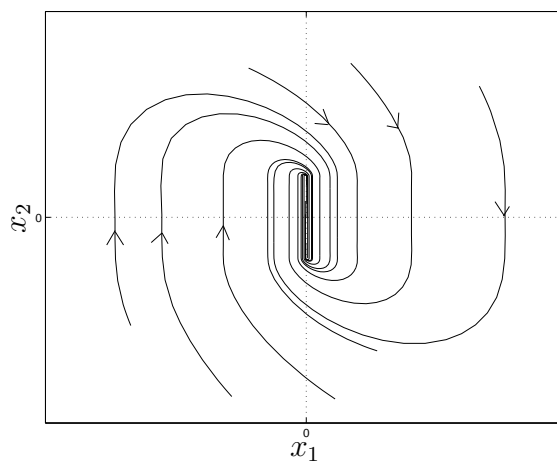


The car will not return to the desired position with proportional control.

(b) Now, we have the state-space form

$$\begin{aligned}\dot{x}_1 &= f(x_2) \\ \dot{x}_2 &= -x_1 - f(x_2)\end{aligned}$$

The difference from (a) is that the stationary points are stable focuses. Hence, we get the phase portrait



The car returns to the desired position.

Go back

13.5

$$\dot{x}_2 = 0 \quad \Rightarrow \quad 0 = -3x_2\left(1 + \frac{1}{6}x_1\right) + x_1x_2 = \frac{1}{2}(x_1 - 6)x_2$$

i.e.

$$x_1 = 6 \quad \text{or} \quad x_2 = 0.$$

Two cases:

$$\begin{aligned} \text{(i) } x_2 = 0 \text{ and } \dot{x}_1 = 0 &\Rightarrow 0 = 2x_1 - 0.2x_1^2 = 0.2(10 - x_1)x_1, \\ \text{i.e. } x_1 = 0 \text{ or } x_1 = 10. \end{aligned}$$

$$\text{(ii) } x_1 = 6 \text{ and } \dot{x}_1 = 0 \quad \Rightarrow$$

$$0 = 2 \cdot 6\left(1 + \frac{1}{6} \cdot 6\right) - 6 \cdot x_2 - 0.2 \cdot 6^2\left(1 + \frac{1}{6} \cdot 6\right) = 24 - 6x_2 - 14.4$$

Hence, the stationary points (SP) are

$$\text{SP I: } \begin{cases} x_1 = 0 \\ x_2 = 0 \end{cases}, \quad \text{SP II: } \begin{cases} x_1 = 10 \\ x_2 = 0 \end{cases}, \quad \text{SP III: } \begin{cases} x_1 = 6 \\ x_2 = 1.6 \end{cases}$$

The Jacobian is

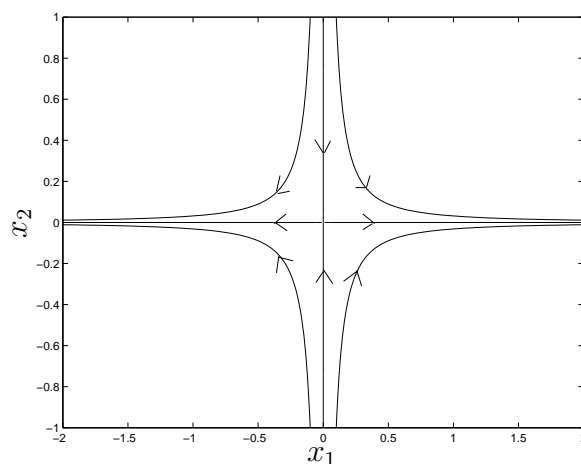
$$H(x) = f_x(x) = \begin{pmatrix} 2 - 0.4x_1 - x_2/(1 + x_1/6)^2 & -x_1/(1 + x_1/6) \\ x_2/(1 + x_1/6)^2 & -3 + x_1/(1 + x_1/6) \end{pmatrix}$$

SP I:

$x_1 = x_2 = 0$ yield

$$H_1 = \begin{pmatrix} 2 & 0 \\ 0 & -3 \end{pmatrix}$$

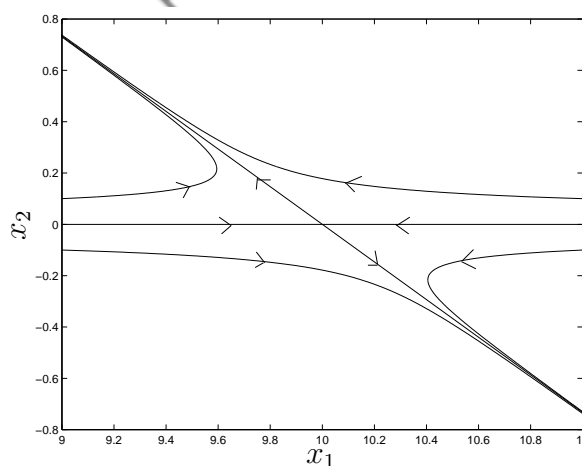
The origin is a saddle point with trajectories according to

**SP II:**

When $x_1 = 10$ and $x_2 = 0$ the Jacobian is

$$H_2 = \begin{pmatrix} -2 & -3.75 \\ 0 & 0.75 \end{pmatrix}$$

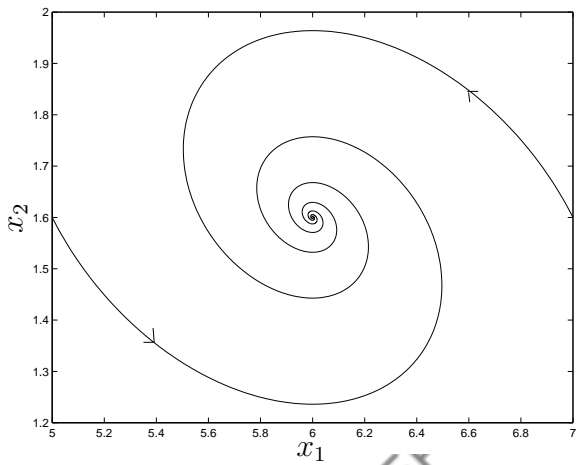
This is also a saddle point. The eigenvector corresponding to the unstable eigenvalue is $(3.75, -2.75)$, and the eigenvector corresponding to the stable eigenvalue is $(1, 0)$. The phase portrait is

**SP III:**

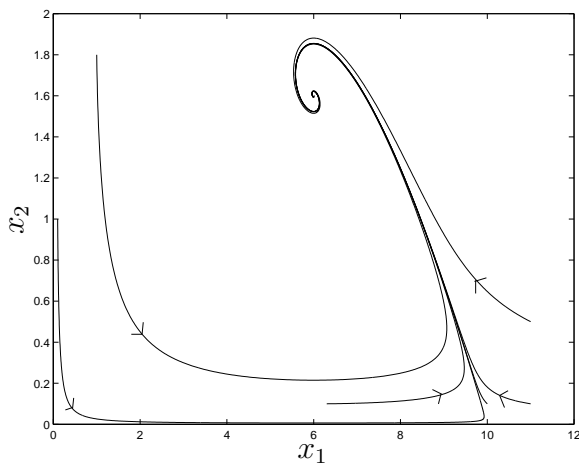
$x_1 = 6$, $x_2 = 1.6$ yield

$$H_3 = \begin{pmatrix} -0.8 & -3 \\ 0.4 & 0 \end{pmatrix}$$

The eigenvalues are $-0.4 \pm 1.02i$. We have a stable focus with the phase portrait



IF we join these phase portraits together we get:



[Go back](#)

13.6

Introduce the states $x_1 = y$ and $x_2 = \dot{y} \Rightarrow$.

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= -x_1 + f(x_2)\end{aligned}$$

For $x_2 > 0$ we have

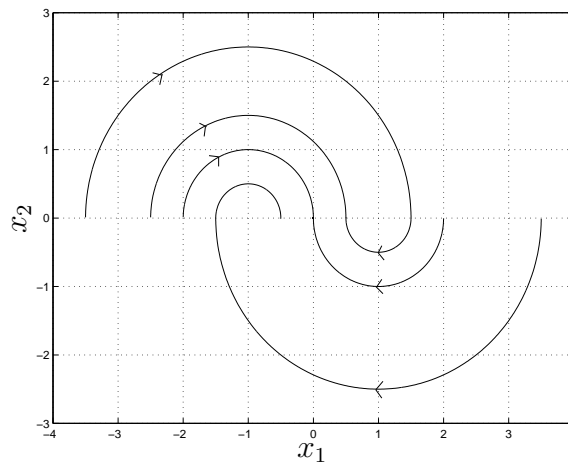
$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -x_1 - 1 \end{cases} \Rightarrow \text{stationary point } (-1, 0)$$

The Jacobian is

$$\begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}$$

As the eigenvalues are $\pm i \Rightarrow x = (-1, 0)$ is a center.

For $x_2 < 0$ we get analogously that $x = (1, 0)$ is a center. If we join the phase portraits together we get

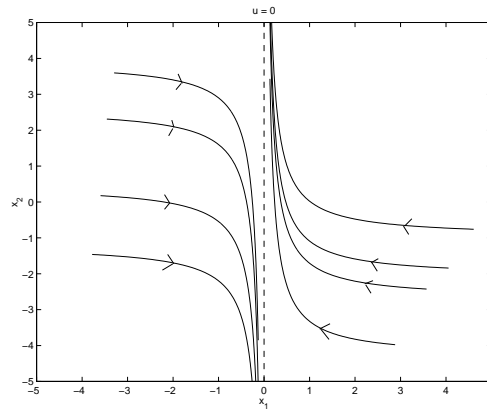


The system will tend to a point on the x_1 -axis, i.e. \dot{y} will tend to zero.

13.7

- (a) We have stationary points for $x_1 = 0$, i.e. the entire x_2 -axis, when $u = 0$. The trajectories are described by

$$\frac{dx_2}{dx_1} = -\frac{1}{x_1^2} \quad \Leftrightarrow \quad x_2 = \frac{1}{x_1} + C.$$



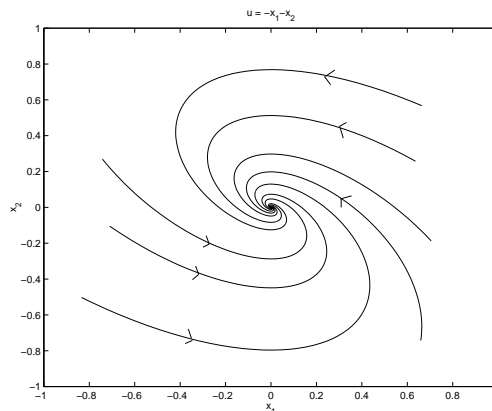
(b)

$$\dot{V} = V_x \dot{x} = -2x_1^4 + 2x_1u + 2x_1x_2 \quad \Rightarrow \quad \text{Use } u = -x_1 - x_2$$

This results in $\dot{V} = -2x_1^4 - 2x_1^2 < 0$ and the stationary point $x_1 = x_2 = 0$. The corresponding linearized system is

$$\dot{x} = \begin{pmatrix} -1 & -1 \\ 1 & 0 \end{pmatrix} x, \quad \text{with the eigenvalues } \lambda = -\frac{1}{2} \pm i\frac{\sqrt{3}}{2}.$$

The stationary point is a stable focus.



Go back