Appendix

This appendix presents the continuous-time designs of model reference adaptive control using state feedback for state tracking, state feedback for output tracking, or output feedback for output tracking (and its discrete-time version), and multivariable design, as well as that of adaptive pole placement control. Key issues such as *a prior* system knowledge, controller structure, plant-model matching, adaptive laws, and stability are addressed.

A.1 Model Reference Adaptive Control

There are three types of model reference adaptive control (MRAC) designs: state feedback for state tracking; state feedback for output tracking; and output feedback for output tracking, all described in this section.

A.1.1 MRAC: State Feedback for State Tracking

Consider a linear time-invariant plant described by

$$\dot{x}(t) = Ax(t) + bu(t),\ x(t) \in R^n,\ u(t) \in R \tag{A.1}$$

where $A \in R^{n\times n}$, $b \in R^n$ are unknown constant parameter matrices, and the state vector $x(t)$ is available for measurement.

The control objective is to design a *state feedback* control signal $u(t)$ that ensures that all signals in the closed-loop system are bounded and the plant state vector $x(t)$ asymptotically tracks a given reference state vector $x_m(t)$ generated from the reference model system

$$\dot{x}_m(t) = A_m x_m(t) + b_m r(t),\ x_m(t) \in R^n,\ r(t) \in R \tag{A.2}$$

where $A_m \in R^{n\times n}$, $b_m \in R^n$ are known constant matrices such that all eigenvalues of A_m are in the open left-half complex plane, and $r(t)$ is bounded.

To meet this control objective, we assume

(A.1-1) There exist a constant vector $k_1^* \in R^n$ and a constant scalar $k_2^* \in R$ such that

$$A + bk_1^{*T} = A_m,\ bk_2^* = b_m \tag{A.3}$$

(A.1-2) sign$[k_2^*]$, the sign of the parameter k_2^*, is known.

While Assumption (A.1-2) is needed for implementing an adaptive law, Assumption (A.1-1) is the so-called matching condition such that if the parameters of A and b are known and (A.3) is satisfied, then the control law

$$u(t) = k_1^{*T}x(t) + k_2^* r(t) \tag{A.4}$$

achieves the desired control objective that the closed-loop system becomes

$$\dot{x}(t) = Ax(t) + b(k_1^{*T}x(t) + k_2^* r(t)) = A_m x(t) + b_m r(t) \tag{A.5}$$

whose state vector $x(t)$ is bounded, and so is the control $u(t)$ in (A.4), and that the tracking error $e(t) = x(t) - x_m(t)$ satisfies

$$\dot{e}(t) = A_m e(t),\ e(0) = x(0) - x_m(0) \tag{A.6}$$

which indicates that $\lim_{t\to\infty} e(t) = 0$ exponentially. It is clear that Condition (A.3) is also necessary for the control law (A.3) to achieve the stated control objective, even if the plant parameters A and b are known.

In the adaptive control problem, the parameters of A and b are unknown so that the control law (A.4) cannot be used for control. In this case, the following adaptive controller structure is used:

$$u(t) = k_1^T(t)x(t) + k_2(t)r(t) \tag{A.7}$$

where $k_1(t)$ and $k_2(t)$ are the estimates of k_1^* and k_2^*, respectively. The adaptive control design task now is to choose adaptive laws to update these estimates so that the stated control objective is still achievable.

Defining the parameter errors as

$$\tilde{k}_1(t) = k_1(t) - k_1^*,\ \tilde{k}_2(t) = k_2(t) - k_2^* \tag{A.8}$$

and using (A.1), (A.3), and (A.7), we obtain

$$\begin{aligned} \dot{x}(t) &= Ax(t) + b\left(k_1^T(t)x(t) + k_2(t)r(t)\right) \\ &= A_m x(t) + b_m r(t) + b_m\left(\frac{1}{k_2^*}\tilde{k}_1^T(t)x(t) + \frac{1}{k_2^*}\tilde{k}_2(t)r(t)\right). \end{aligned} \tag{A.9}$$

Substituting (A.2) in (A.9), we have the tracking error equation

$$\dot{e}(t) = A_m e(t) + b_m \left(\frac{1}{k_2^*} \tilde{k}_1^T(t) x(t) + \frac{1}{k_2^*} \tilde{k}_2(t) r(t) \right). \tag{A.10}$$

Consider the positive definite function

$$V(e, \tilde{k}_1, \tilde{k}_2) = e^T P e + \frac{1}{|k_2^*|} \tilde{k}_1^T \Gamma^{-1} \tilde{k}_1 + \frac{1}{|k_2^*|} \tilde{k}_2^2 \gamma^{-1} \tag{A.11}$$

as a measure of the system error signals $e(t)$, $\tilde{k}_1(t)$, and $\tilde{k}_2(t)$, where $P \in R^{n \times n}$ is constant, $P = P^T > 0$ and satisfies

$$P A_m + A_m^T P = -Q \tag{A.12}$$

for some constant $Q \in R^{n \times n}$ such that $Q = Q^T > 0$, $\Gamma \in R^{n \times n}$ is constant and $\Gamma = \Gamma^T > 0$, and $\gamma > 0$ is a constant scalar.

The time derivative of $V(e, \tilde{k}_1, \tilde{k}_2)$ is

$$\begin{aligned} \dot{V} &= \frac{d}{dt} V = \left(\frac{\partial V}{\partial e} \right)^T \dot{e}(t) + \left(\frac{\partial V}{\partial \tilde{k}_1} \right)^T \dot{\tilde{k}}_1(t) + \frac{\partial V}{\partial \tilde{k}_2} \dot{\tilde{k}}_2(t) \\ &= 2e^T(t) P \dot{e}(t) + \frac{2}{|k_2^*|} \tilde{k}_1^T(t) \Gamma^{-1} \dot{\tilde{k}}_1(t) + \frac{2}{|k_2^*|} \tilde{k}_2(t) \gamma^{-1} \dot{\tilde{k}}_2(t). \end{aligned} \tag{A.13}$$

Substituting (A.10) and (A.12) in (A.13), we have

$$\begin{aligned} \dot{V} &= -e^T(t) Q e(t) + e^T(t) P b_m \frac{2}{k_2^*} \tilde{k}_1^T(t) x(t) + e^T(t) P b_m \\ &\quad + \frac{2}{k_2^*} \tilde{k}_2(t) r(t) + \frac{2}{|k_2^*|} \tilde{k}_1^T(t) \Gamma^{-1} \dot{\tilde{k}}_1(t) + \frac{2}{|k_2^*|} \tilde{k}_2(t) \gamma^{-1} \dot{\tilde{k}}_2(t). \end{aligned} \tag{A.14}$$

To make $\dot{V} \le 0$, we choose the adaptive laws for $k_1(t)$ and $k_2(t)$ as

$$\dot{k}_1(t) = -\text{sign}[k_2^*] \Gamma x(t) e^T(t) P b_m \tag{A.15}$$

$$\dot{k}_2(t) = -\text{sign}[k_2^*] \gamma r(t) e^T(t) P b_m \tag{A.16}$$

with $\Gamma = \Gamma^T > 0$, $\gamma > 0$, $k_1(0)$ and $k_2(0)$ being arbitrary.

Indeed, with this choice of $\dot{k}_1(t)$ and $\dot{k}_2(t)$, (A.14) becomes

$$\dot{V} = -e^T(t) Q e(t) \le 0. \tag{A.17}$$

From (A.17), we conclude that $x(t)$, $k_1(t)$, and $k_2(t)$ are all bounded, and so are $u(t)$ in (A.7) and $\dot{e}(t)$ in (A.10), and that $e(t) \in L^2$.[1]

[1] A *continuous-time* vector signal $x(t) \in R^n$ belongs to L^2 if

$$\sqrt{\int_0^\infty (x_1^2(t) + \cdots + x_n^2(t))\, dt} < \infty$$

To show that $\lim_{t\to\infty} e(t) = 0$, we see that for $e = [e_1, \cdots, e_n]^T$, $e_i(t) \in L^2$ and $\dot{e}_i(t) \in L^\infty$, $i = 1, 2, \ldots, n$. It follows that

$$\int_0^t e_i^2(\tau)|\dot{e}_i(\tau)|d\tau \leq \sup_{t\geq 0} |\dot{e}_i(t)| \int_0^\infty e_i^2(\tau)d\tau < \infty \tag{A.18}$$

for any $t \geq 0$. This implies that $\lim_{t\to\infty} \int_0^t e_i^2(\tau)|\dot{e}_i(\tau)|d\tau$ exists and is finite, and, therefore, $\lim_{t\to\infty} \int_0^t e_i^2(\tau)\dot{e}_i(\tau)d\tau$ exists and is finite. From the identity

$$e_i^2(t) = |e_i^3(t)|^{\frac{2}{3}} = |3\int_0^t e_i^2(\tau)\dot{e}_i(\tau)d\tau + e_i^3(0)|^{\frac{2}{3}} \tag{A.19}$$

we have that $\lim_{t\to\infty} e_i^2(t)$ exists and is zero as $e_i(t) \in L^2$. This proves that $\lim_{t\to\infty} e_i(t) = 0$, $i = 1, 2, \ldots, n$, so that $\lim_{t\to\infty} e(t) = 0$.

In summary, we have the following result.

Theorem A.1. *The adaptive controller (A.7), with the adaptive laws (A.15) and (A.16), and applied to the plant (A.1), ensures that all closed-loop signals are bounded and the tracking error* $e(t) = x(t) - x_m(t)$ *goes to zero as* $t \to \infty$.

A.1.2 MRAC: State Feedback for Output Tracking

Consider a linear time-invariant plant in the state variable form

$$\begin{aligned} \dot{x}(t) &= Ax(t) + bu(t) \\ y(t) &= Cx(t) \end{aligned} \tag{A.20}$$

for some unknown constant matrices $A \in R^{n\times n}$, $b \in R^{n\times 1}$, and $C \in R^{1\times n}$, with $n > 0$. The input–output description of this plant is

$$y(s) = C(sI - A)^{-1}bu(s) = \frac{Z(s)}{P(s)}u(s) \tag{A.21}$$

where $P(s) = \det(sI - A)$ and

$$Z(s) = z_m s^m + \cdots + z_1 s + z_0,\ z_m \neq 0,\ m < n \tag{A.22}$$

with s being the Laplace transform variable or the time differentiation operator: $s[x](t) = \dot{x}(t)$, $t \in [0, \infty)$, as the case may be.

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and it belongs to L^∞ if

$$\sup_{t\geq 0} \|x(t)\|_\infty = \sup_{t\geq 0} \max_{1\leq i\leq n} |x_i(t)| < \infty.$$

A signal $x(t)$ is bounded if it belongs to L^∞.

To design an adaptive *state feedback* model reference controller for generating the plant input $u(t)$, which ensures closed-loop signal boundedness and asymptotic tracking of a given reference signal $y_m(t)$ by the plant output $y(t)$, we need the following assumptions:

(A.2-1) (A, b, C) is stabilizable and detectable,
(A.2-2) $Z(s)$ is a stable polynomial,
(A.2-3) The degree m of $Z(s)$ is known, and
(A.2-4) $\mathrm{sign}[z_m]$, the sign of z_m, is known.

Assumption (A.2-1) is needed for output matching, Assumption (A.2-2) is needed for model reference control, and Assumption (A.2-3) is needed for constructing a reference model system, while Assumption (A.2-4) is used for designing an adaptive parameter update law.

The reference model, independent of the dynamics of (A.20), is chosen as

$$y_m(t) = W_m(s)[r](t), \; W_m(s) = \frac{1}{P_m(s)} \tag{A.23}$$

where $P_m(s)$ is a desired stable polynomial of degree $n - m$, and $r(t)$ is a bounded reference input signal.

The adaptive model reference controller structure is

$$u(t) = k_1^T(t)x(t) + k_2(t)r(t) \tag{A.24}$$

where $k_1(t) = [k_{11}(t), k_{12}(t), \ldots, k_{1n}(t)]^T \in R^n$ and $k_2(t) \in R$ are the adaptive estimates of the unknown parameters $k_1^* = [k_{11}^*, k_{12}^*, \ldots, k_{1n}^*]^T \in R^n$ and k_2^*, which satisfy the matching eqaution

$$\det(sI - A - bk_1^{*T}) = P_m(s)Z(s)\frac{1}{z_m}, \; k_2^* = \frac{1}{z_m} \tag{A.25}$$

that is, all zeros of $\det(sI - A - bk_1^{*T})$ are stable [118].

With this definition of k_1^* and k_2^*, the fixed version of (A.24) is

$$u(t) = k_1^{*T}x(t) + k_2^*r(t) \tag{A.26}$$

which would lead to the desired closed-loop system: $y(s) = W_m(s)r(s)$.

In the adaptive control problem when A, b, and z_m are all unknown parameters, we need to develop adaptive laws to update the parameter estimates $k_1(t)$ and $k_2(t)$. With the controller (A.24), the closed-loop system is

$$\begin{aligned} \dot{x}(t) &= (A + bk_1^{*T})x(t) + bk_2^*r(t) + b((k_1(t) - k_1^*)^T x(t) + (k_2(t) - k_2^*)r(t)) \\ y(t) &= Cx(t). \end{aligned} \tag{A.27}$$

From (A.25) it follows that

$$C(sI - A - bk_1^{*T})^{-1}bk_2^* = \frac{Z(s)k_2^*}{\det(sI - A - bk_1^{*T})} = \frac{1}{P_m(s)} = W_m(s). \quad \text{(A.28)}$$

In view of (A.23), (A.27), and (A.28), the tracking error equation is

$$\begin{aligned} e(t) &= y(t) - y_m(t) \\ &= \rho^* W_m(s)[(k_1 - k_1^*)^T x + (k_2 - k_2^*)r](t) + Ce^{(A+bk_1^{*T})t}x(0) \quad \text{(A.29)} \end{aligned}$$

where $\rho^* = z_m$, and $\lim_{t\to\infty} Ce^{(A+bk_1^{*T})t}x(0) = 0$ exponentially.

To derive an estimation error equation, we define

$$\theta(t) = \left[k_1^T(t), k_2(t)\right]^T \quad \text{(A.30)}$$

$$\theta^* = \left[k_1^{*T}, k_2^*\right]^T \quad \text{(A.31)}$$

$$\omega(t) = \left[x^T(t), r(t)\right]^T \quad \text{(A.32)}$$

$$\zeta(t) = W_m(s)[\omega](t) \quad \text{(A.33)}$$

$$\xi(t) = \theta^T(t)\zeta(t) - W_m(s)[\theta^T\omega](t) \quad \text{(A.34)}$$

$$\epsilon(t) = e(t) + \rho(t)\xi(t) \quad \text{(A.35)}$$

where $\rho(t)$ is an estimate of $\rho^* = z_m$. Then, from (A.29)–(A.35), ignoring the exponentially decaying term $Ce^{(A+bk_1^{*T})t}x(0)$, we have

$$\epsilon(t) = \rho^*(\theta(t) - \theta^*)^T\zeta(t) + (\rho(t) - \rho^*)\xi(t) \quad \text{(A.36)}$$

which is linear in the parameters errors $\theta(t) - \theta^*$ and $\rho(t) - \rho^*$.

We then choose the adaptive laws for $\theta(t)$ and $\rho(t)$ as

$$\dot\theta(t) = -\frac{\Gamma \mathrm{sign}[z_m]\zeta(t)\epsilon(t)}{1 + \zeta^T(t)\zeta(t) + \xi^2(t)} \quad \text{(A.37)}$$

$$\dot\rho(t) = -\frac{\gamma\xi(t)\epsilon(t)}{1 + \zeta^T(t)\zeta(t) + \xi^2(t)} \quad \text{(A.38)}$$

where $\Gamma = \Gamma^T > 0$ and $\gamma > 0$ are adaptation gains.

This adaptive control scheme has the following properties [55], [95], [118].

Lemma A.1. *The adaptive laws (A.37) and (A.38) ensures that* $\theta(t) \in L^\infty$, $\rho(t) \in L^\infty$, $\epsilon(t)/m(t) \in L^2 \cap L^\infty$, $\dot\theta(t) \in L^2 \cap L^\infty$, *and* $\dot\rho(t) \in L^2 \cap L^\infty$, *where*

$$m(t) = \sqrt{1 + \zeta^T(t)\zeta(t) + \xi^2(t)}. \quad \text{(A.39)}$$

Proof: Consider the positive definite function

$$V(\tilde{\theta}, \tilde{\rho}) = \frac{1}{2}(|\rho^*|\tilde{\theta}^T \Gamma^{-1}\tilde{\theta} + \gamma^{-1}\tilde{\rho}^2) \tag{A.40}$$

where

$$\tilde{\theta}(t) = \theta(t) - \theta^*,\ \tilde{\rho}(t) = \rho(t) - \rho^*. \tag{A.41}$$

The time derivative of $V(\tilde{\theta}, \tilde{\rho})$, along the trajectories of (A.37) and (A.38), is

$$\dot{V} = -\frac{\epsilon^2(t)}{m^2(t)}. \tag{A.42}$$

This, together with (A.36)–(A.40), leads to the results of the lemma. ∇

Theorem A.2. *All signals in the closed-loop control system, with the plant (A.20), the reference model (A.23), and the controller (A.24) updated by the adaptive law (A.37) and (A.38), are bounded, and the tracking error* $e(t) = y(t) - y_m(t)$ *satisfies*

$$\lim_{t\to\infty} (y(t) - y_m(t)) = 0 \tag{A.43}$$

$$\int_0^\infty (y(t) - y_m(t))^2 dt < \infty. \tag{A.44}$$

The proof of Theorem A.2 can be found in [118].

A.1.3 MRAC: Output Feedback for Output Tracking

Consider the linear time-invariant plant

$$y(t) = G(s)[u](t) \tag{A.45}$$

where $G(s) = k_p(Z(s)/P(s))$, k_p is a constant high frequency gain, and $Z(s)$ and $P(s)$ are monic polynomials of degrees n and m, respectively.

Given a reference model system

$$y_m(t) = W_m(s)[r](t) \tag{A.46}$$

where $W_m(s)$ is stable and $r(t)$ is bounded, the control objective is to find an *output feedback* control $u(t)$ such that all closed-loop signals are bounded and the plant output y tracks the reference output y_m asymptotically.

In this case, the plant state variables are not needed for control.

To meet this objective, we make the following assumptions:

(A.3-1) $Z(s)$ is a stable polynomial,
(A.3-2) The degree n of $P(s)$ is known,
(A.3-3) The relative degree $n^* = n - m$ of $G(s)$ is known,
(A.3-4) The sign of k_p is known, and
(A.3-5) $W_m(s)=1/P_m(s)$ for a stable polynomial $P_m(s)$ of degree n^*.

The desired model reference controller structure is

$$u(t) = \theta_1^T \omega_1(t) + \theta_2^T \omega_2(t) + \theta_{20} y(t) + \theta_3 r(t) \tag{A.47}$$

where

$$\omega_1(t) = \frac{a(s)}{\Lambda(s)}[u](t),\ \omega_2(t) = \frac{a(s)}{\Lambda(s)}[y](t) \tag{A.48}$$

with $a(s) = [1, s, \cdots, s^{n-2}]^T$, $\theta_1,\ \theta_2 \in R^{n-1},\ \theta_{20}$, and $\theta_3 \in R$, and $\Lambda(s)$ being a monic stable polynomial of degree $n-1$.

It can be shown [55], [118] that with $\theta_3^* = k_p^{-1}$, there exist constant parameters θ_1^*, $\theta_2^* \in R^{n-1}$, and $\theta_{20}^* \in R$ such that

$$\begin{aligned} &\theta_1^{*T} a(s)P(s) + (\theta_2^{*T} a(s) + \theta_{20}^* \Lambda(s))k_p Z(s) \\ = \;& \Lambda(s)(P(s) - k_p \theta_3^* Z(s) P_m(s)) \end{aligned} \tag{A.49}$$

and that the controller (A.47) implemented with θ_1^*, θ_2^*, θ_{20}^*, and θ_3^*, that is,

$$u(t) = \theta_1^{*T} \omega_1(t) + \theta_2^{*T} \omega_2(t) + \theta_{20}^* y(t) + \theta_3^* r(t) \tag{A.50}$$

ensures that all the closed-loop signals are bounded and the output tracking is achieved: $y(t) = y_m(t) + \eta_0(t)$, for some exponentially decaying $\epsilon_0(t)$ that depends on system initial conditions.

For the adaptive control problem when $G(s)$ is unknown, an adaptive version of the controller (A.47) is implemented with $\theta_1 = \theta_1(t)$, $\theta_2 = \theta_2(t)$, $\theta_{20} = \theta_{20}(t)$, and $\theta_3 = \theta_3(t)$, where $\theta_1(t)$, $\theta_2(t)$, $\theta_{20}(t)$, and $\theta_3(t)$ are the estimates of θ_1^*, θ_2^*, θ_{20}^*, and θ_3^*, to be updated from an adaptive law.

To derive an error equation, operating both sides of (A.49) on $y(t)$ and using (A.45): $P(s)[y](t) = k_p Z(s)[u](t)$, we have

$$\begin{aligned} &\theta_1^{*T} a(s) k_p Z(s)[u](t) + (\theta_2^{*T} a(s) + \theta_{20}^* \Lambda(s)) k_p Z(s)[y](t) \\ = \;& \Lambda(s) k_p Z(s)[u](t) - \Lambda(s) Z(s) P_m(s)[y](t). \end{aligned} \tag{A.51}$$

This equality, with $Z(s)$ and $\Lambda(s)$ stable and the effect of initial conditions ignored, leads to the parametrized plant model

$$u(t) = \theta_1^{*T} \frac{a(s)}{\Lambda(s)}[u](t) + \theta_2^{*T} \frac{a(s)}{\Lambda(s)}[y](t) + \theta_{20}^* y(t) + \theta_3^* W_m^{-1}(s)[y](t). \tag{A.52}$$

Introducing $\rho^* = 1/\theta_3^* = k_p$ and

$$\omega(t) = \left[\omega_1^T(t), \omega_2^T(t), y(t), r(t)\right]^T \tag{A.53}$$

$$e(t) = y(t) - y_m(t),\ \tilde{\theta}(t) = \theta(t) - \theta^* \tag{A.54}$$

where

$$\theta(t) = \left[\theta_1^T(t), \theta_2^T(t), \theta_{20}(t), \theta_3(t)\right]^T \tag{A.55}$$

$$\theta^* = \left[\theta_1^{*T}, \theta_2^{*T}, \theta_{20}^*, \theta_3^*\right]^T \tag{A.56}$$

substituting (A.52) in (A.47) with adaptive parameter estimates, ignoring the effect of the initial conditions, and using (A.55) and (A.56), we obtain the tracking error equation

$$\begin{aligned} e(t) &= \rho^* W_m(s)[\tilde{\theta}^T \omega](t) \\ &= -\rho^* \left(\theta^{*T} W_m(s)[\omega](t) - W_m(s)[\theta^T \omega](t)\right). \end{aligned} \tag{A.57}$$

Since both θ^* and ρ^* are unknown, the second equality of (A.57) suggests that we define the estimation error

$$\epsilon(t) = e(t) + \rho(t)\xi(t) \tag{A.58}$$

where $\rho(t)$ is the estimate of ρ^*, and

$$\xi(t) = \theta^T(t)\zeta(t) - W_m(s)[\theta^T \omega](t) \tag{A.59}$$

$$\zeta(t) = W_m(s)[\omega](t). \tag{A.60}$$

For $\tilde{\rho}(t) = \rho(t) - \rho^*$, using (A.57), (A.59), and (A.60), we express $\epsilon(t)$ as

$$\epsilon(t) = \rho^* \tilde{\theta}^T(t)\zeta(t) + \tilde{\rho}(t)\xi(t) \tag{A.61}$$

which is linear in the parameter errors $\tilde{\theta}(t)$ and $\tilde{\rho}(t)$.

Finally, we choose the adaptive laws for $\theta(t)$ and $\rho(t)$:

$$\dot{\theta}(t) = \frac{-\mathrm{sign}[k_p]\Gamma\epsilon(t)\zeta(t)}{m^2(t)} \tag{A.62}$$

$$\dot{\rho}(t) = \frac{-\gamma\epsilon(t)\xi(t)}{m^2(t)} \tag{A.63}$$

where $\Gamma = \Gamma^T > 0$ and $\gamma > 0$ are constant adaptation gains, and

$$m(t) = \sqrt{1 + \zeta^T(t)\zeta(t) + \xi^2(t)}. \tag{A.64}$$

This adaptive controller has the following desired properties
[55], [95], [118].

Lemma A.2. *The adaptive update law (A.62) and (A.63) ensures that* $\theta(t) \in L^\infty$, $\rho(t) \in L^\infty$, $\epsilon(t)/m(t) \in L^2 \cap L^\infty$, *and* $\dot{\theta}(t) \in L^2 \cap L^\infty$.

Theorem A.3. *All signals in the closed-loop control system with the plant (A.45), reference model (A.46), and controller (A.47) updated by the adaptive laws (A.62) and (A.63) are bounded, and*

$$y(t) - y_m(t) \in L^2, \ \lim_{t\to\infty} (y(t) - y_m(t)) = 0. \tag{A.65}$$

The proof of Lemma A.2 is similar to that of Lemma A.1, and the proof of Theorem A.2 can be found in [118], [125].

Discrete-Time Design. While both state feedback for output tracking and output feedback for output tracking designs can be developed in discrete time, here we only illustrate how to develop a discrete-time adaptive output feedback control design for output tracking. Such a design can be obtained in a way similar to that for the continuous-time case: the same controller structure (A.47), the same parametrization (A.49)–(A.57), and the same estimation error (A.58), but in (A.45)–(A.61), we need to replace s by z which, for the discrete-time case, is the z-transform variable or the time advance operator: $z[x](k) = x(k+1)$, $k \in \{0,1,2,3,\ldots\}$. In particular, a stable polynomial of degree n_p can be chosen as z^{n_p}, for example, $\Lambda(z) = z^{n-1}$ for (A.47) and $W_m(z) = z^{-n^*}$ for (A.46). However, the adaptive law for updating the controller parameter vector $\theta(k)$ in (A.47) is different:

$$\theta(k+1) = \theta(k) - \frac{\mathrm{sign}[k_p]\Gamma\epsilon(k)\omega(k-n^*)}{m^2(k)} \tag{A.66}$$

$$\rho(k+1) = \rho(k) - \frac{\gamma\epsilon(k)\xi(k)}{m^2(k)} \tag{A.67}$$

where the adaptation gains Γ and γ satisfy

$$0 < \Gamma = \Gamma^T < \frac{2}{k_p^0} I_{2n}, \ 0 < \gamma < 2 \tag{A.68}$$

for a known constant $k_p^0 \geq |k_p|$.

The adaptive laws (A.66) and (A.67) ensures that $\theta(k) \in L^\infty$, $\rho(k) \in L^\infty$, $\epsilon(k)/m(k) \in L^2 \cap L^\infty$, and $\theta(k+i_0) - \theta(k) \in L^2$ for any finite integer $i_0 > 0$.[2]

[2] A *discrete-time* vector signal $x(k) \in R^n$ belongs to L^2 if

$$\sqrt{\sum_{k=0}^{\infty} (x_1^2(k) + \cdots + x_n^2(k))} < \infty$$

The proof of this result is similar to that for the continuous-time case. The time increment of the positive definite function

$$V(\tilde{\theta}, \tilde{\rho}) = |\rho^*|\tilde{\theta}^T \Gamma^{-1}\tilde{\theta} + \gamma^{-1}\tilde{\rho}^2 \tag{A.69}$$

along the trajectories of (A.66) and (A.67) is

$$\begin{aligned} & V(\tilde{\theta}(k+1), \tilde{\rho}(k+1)) - V(\tilde{\theta}(k), \tilde{\rho}(k)) \\ = & -\left(2 - \frac{|k_p|\omega^T(k-n^*)\Gamma\omega(k-n^*) + \gamma\xi^2(k)}{m^2(k)}\right)\frac{\epsilon^2(k)}{m^2(k)} \\ \leq & -\alpha_1 \frac{\epsilon^2(k)}{m^2(k)} \end{aligned} \tag{A.70}$$

for some constant $\alpha_1 > 0$. This implies that $\theta(k) \in L^\infty$, $\rho(k) \in L^\infty$, and $\epsilon(k)/m(k) \in L^2$. From (A.61), we have $\epsilon(k)/m(k) \in L^\infty$, and from (A.66), we have $\theta(k+1) - \theta(k) \in L^2$. Finally, using the inequality

$$\|\theta(k+i_0) - \theta(k)\|_2 \leq \sum_{i=0}^{i_0-1} \|\theta(k+i+1) - \theta(k+i)\|_2 \tag{A.71}$$

we have that $\theta(k+i_0) - \theta(k) \in L^2$ for any finite integer i_0.

Based on this result, we can also establish the result of Theorem A.3 for the discrete-time adaptive control scheme [118].

A.2 Multivariable MRAC

Consider the linear time-invariant plant

$$y(t) = G(s)[u](t) \tag{A.72}$$

where $y(t) \in R^M$, $u(t) \in R^M$, $G(s) = Z(s)P^{-1}(s)$ is strictly proper and has full rank, and $Z(s)$ and $P(s)$ are $M \times M$ right co-prime polynomial matrices with $P(s)$ column proper [95]. The symbol s is used as the Laplace transform variable or the time differentiation operator: $s[x](t) = \dot{x}(t)$, $t \in [0, \infty)$. The following development of multivariable MRAC is applicable to the discrete-time case, with different system stability conditions (which lead to the special choice of adaptation gains in parameter adaptive laws).

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and it belongs to L^∞ (i.e., it is bounded) if

$$\sup_{k\geq 0} \|x(k)\|_\infty = \sup_{k\geq 0} \max_{1\leq i\leq n} |x_i(k)| < \infty.$$

The control objective is to find a feedback control $u(t)$ for the plant (A.72) with unknown $G(s)$ such that all signals in the closed-loop system are bounded and $y(t)$ asymptotically tracks the reference output

$$y_m(t) = W_m(s)[r](t),\ W_m(s) = \xi_m^{-1}(s) \tag{A.73}$$

where $\xi_m(s)$ is a modified interactor matrix of $G(s)$ [32], [95], [123], and $r(t)$ is a bounded signal. The modified interactor matrix $\xi_m(s)$ has a stable inverse and represents the zero structure at infinity of $G(s)$. The high frequency gain matrix K_p of $G(s)$ is defined as $K_p = \lim_{D\to\infty} \xi_m(s)G(s)$, finite and nonsingular. To design MRAC schemes, we assume

(A.4-1) All zeros of $G(s)$ (that is, zeros of $\det[Z(s)]$) are stable,
(A.4-2) $\bar{\nu} \geq \nu$, the observability index of $G(s)$, is known,
(A.4-3) For some known $S_p \in R^{M\times M}$, $\Gamma_p = K_p^T S_p^{-1} = \Gamma_p^T > 0$, and
(A.4-4) A modified interactor matrix $\xi_m(s)$ of $G(s)$ is known.

Note that Assumption (A.4-3) may be relaxed, as all leading principal minors of K_p are nonzero and their signs are known (see Section 5.3.4).

Controller Structure. The MRAC controller structure for (A.72) is

$$u(t) = \Theta_1^T\omega_1(t) + \Theta_2^T\omega_2(t) + \Theta_{20}y(t) + \Theta_3 r(t) \tag{A.74}$$

where $\Theta_1 = [\Theta_{11}, \ldots, \Theta_{1\bar{\nu}-1}]^T$, $\Theta_2 = [\Theta_{21}, \ldots, \Theta_{2\bar{\nu}-1}]^T$, Θ_{20}, Θ_3, $\Theta_{ij} \in R^{M\times M}$, $i = 1, 2$, $j = 1, \ldots, \bar{\nu} - 1$, and $\omega_1(t) = F(s)[u](t)$, $\omega_2(t) = F(s)[y](t)$, $F(s) = A(s)/\Lambda(s)$, with $A(s) = [I, DI, \ldots, D^{\bar{\nu}-2}I]^T$, and $\Lambda(s)$ being a monic stable polynomial of degree $\bar{\nu} - 1$.

Error Model. With the specification of $\Lambda(s)$, $\xi_m(s)$, $P(s)$, and $Z(s)$, there exist Θ_1^*, Θ_2^*, $\Theta_3^* = K_p^{-1}$ such that $I - \Theta_1^{*T}F(s) - \Theta_2^{*T}F(s)G(s) - \Theta_{20}^*G(s) = \Theta_3^* W_m^{-1}(s)G(s)$ [32], from which, for any $u(t)$, we have

$$\begin{aligned} &K_p\left(u(t) - \Theta_1^{*T}\omega_1(t) - \Theta_2^{*T}\omega_2(t) - \Theta_{20}^*y(t) - \Theta_3^*r(t)\right) \\ &= \xi_m(s)[y - y_m](t). \end{aligned} \tag{A.75}$$

With the controller (A.74), we write (A.75) as

$$\xi_m(s)[y - y_m](t) = K_p\tilde{\Theta}^T(t)\omega(t) \tag{A.76}$$

where $\tilde{\Theta}(t) = \Theta(t) - \Theta^*$ with $\Theta(t)$ the estimate of $\Theta^* = [\Theta_1^{*T}, \Theta_2^{*T}, \Theta_{20}^*, \Theta_3^*]^T$, and $\omega(t) = [\omega_1^T(t), \omega_2^T(t), y^T(t), r^T(t)]^T$.

Let $e(t) = y(t) - y_m(t)$ and introduce $\bar{e}(t) = h(s)\xi_m(s)[e](t)$ for $h(s) = 1/f(s)$ with $f(s)$ being a monic and stable polynomial such that $h(s)\xi_m(s)$ is proper and stable. Define the estimation error

$$\epsilon(t) = \bar{e}(t) + \Psi(t)\xi(t) \tag{A.77}$$

where $\Psi(t)$ is the estimate of K_p, and

$$\xi(t) = \Theta^T(t)\zeta(t) - h(s)[\Theta^T\omega](t) \tag{A.78}$$

$$\zeta(t) = h(s)[\omega](t). \tag{A.79}$$

It then follows from (A.76) (filtered by $h(s)\xi_m(s)$) and (A.77)–(A.79) that

$$\epsilon(t) = K_p\tilde{\Theta}^T(t)\zeta(t) + \tilde{\Psi}(t)\xi(t) \tag{A.80}$$

where $\tilde{\Psi}(t) = \Psi(t) - \Psi^*$. The development containing (A.73)–(A.80) can be done for the discrete-time case, using discrete-time transfer functions.

Adaptive Laws. We choose the *continuous-time* adaptive laws as

$$\dot{\Theta}^T(t) = -\frac{S_p\epsilon(t)\zeta^T(t)}{m^2(t)} \tag{A.81}$$

$$\dot{\Psi}(t) = -\frac{\Gamma\epsilon(t)\xi^T(t)}{m^2(t)} \tag{A.82}$$

where $K_pS_p = (K_pS_p)^T > 0$ (see Assumption (A.4-3)), $\Gamma = \Gamma^T > 0$, and

$$m(t) = \sqrt{1 + \zeta^T(t)\zeta(t) + \xi^T(t)\xi(t)}. \tag{A.83}$$

Then the time derivative of the positive definite function

$$V = \mathrm{tr}[\tilde{\Theta}(t)\Gamma_p\tilde{\Theta}^T(t)] + \mathrm{tr}[\tilde{\Psi}^T(t)\Gamma^{-1}\tilde{\Psi}(t)] \tag{A.84}$$

with $\Gamma_p = \Gamma_p^T > 0$ (see Assumption (A.4-3)), along (A.81) and (A.82), is

$$\dot{V} = -\frac{2\epsilon^T(t)\epsilon(t)}{m^2(t)} \le 0. \tag{A.85}$$

Similarly, we choose the *discrete-time* adaptive laws as

$$\Theta^T(t+1) = \Theta^T(t) - S_p\epsilon(t)\zeta^T(t) \tag{A.86}$$

$$\Psi(t+1) = \Psi(t) - \Gamma\epsilon(t)\xi^T(t) \tag{A.87}$$

where $2I > K_pS_p = (K_pS_p)^T > 0$ and $2I > \Gamma = \Gamma^T > 0$.

Then, the time increment of

$$V = \mathrm{tr}[\tilde{\Theta}(k)\Gamma_p\tilde{\Theta}^T(k)] + \mathrm{tr}[\tilde{\Psi}^T(k)\Gamma^{-1}\tilde{\Psi}(k)] \tag{A.88}$$

with $\Gamma_p = K_p^T S_p^{-1} = \Gamma_p^T > 0$, along (A.86) and (A.87), is

$$\begin{aligned} & V(\tilde{\Theta}(t+1), \tilde{\Psi}(t+1)) - V(\tilde{\Theta}(t), \tilde{\Psi}(t)) \\ = & -\frac{\epsilon^T(t)}{m^2(t)}\left(2I - \frac{\zeta^T(t)\zeta(t)S_p^T K_p^T + \xi^T(t)\xi(t)\Gamma}{m^2(t)}\right)\epsilon(t) \leq 0. \end{aligned} \tag{A.89}$$

From (A.85) or (A.89), the closed-loop system stability and asymptotic tracking properties can be proved [123].

A.3 Adaptive Pole Placement Control

Consider the linear time-invariant plant described by

$$y(t) = G(s)[u](t) \tag{A.90}$$

where $y(t) \in R$ and $u(t) \in R$ are the measured plant input and output, respectively, and $G(s) = Z(s)/P(s)$ with

$$P(s) = s^n + p_{n-1}s^{n-1} + \cdots + p_1 s + p_0 \tag{A.91}$$

$$Z(s) = z_{n-1}s^{n-1} + z_{n-2}s^{n-2} + \cdots + z_1 s + z_0 \tag{A.92}$$

for some unknown but constant parameters p_i, z_i, $i = 0, 1, \cdots, n-1$.

The control objective is to find an output feedback control signal $u(t)$ such that all closed-loop signals are bounded and $y(t)$ asymptotically tracks a reference output $y_m(t)$ satisfying

$$Q(s)[y_m](t) = 0 \tag{A.93}$$

where $Q(s)$ is a monic polynomial of degree n_q with no zeros in $Re[s] > 0$ and only nonrepeated zeros on the $j\omega$-axis so that $y_m(t)$ is bounded.

For examples, $Q(s) = 1$ for $y_m(t) = 0$; $Q(s) = s$ for $y_m(t) = a \neq 0$; $Q(s) = s^2 + \sigma^2$ for $y_m(t) = a\sin(\sigma t) + b\cos(\sigma t)$ with $\sigma \neq 0$ and not both a, b being zero; $Q(s) = (s^2 + \sigma_1^2)(s^2 + \sigma_2^2)$ for $y_m(t) = a_1\sin(\sigma_1 t) + a_2\sin(\sigma_2 t) + b_1\cos(\sigma_1 t) + b_2\cos(\sigma_2 t)$ with $\sigma_1 \neq 0$, $\sigma_2 \neq 0$, $\sigma_1 \neq \sigma_2$, not both a_1, b_1 being zero, and not both a_2, b_2 being zero; and $Q(s) = s(s^2 + \sigma^2)$ for $y_m(t) = a\sin(\sigma t) + b\cos(\sigma t) + c$ with $c \neq 0$, $\sigma \neq 0$ and not both a, b being zero.

The following assumptions are needed for this control problem:

(A.5-1) $Q(s)P(s)$ and $Z(s)$ are co-prime, and
(A.5-2) The order n of $P(s)$ is known.

Design Procedure. We use the following indirect approach to design an adaptive controller: First estimate the plant parameters in

$$\theta_p^* = [p_0, p_1, \cdots, p_{n-1}]^T \in R^n, \ \theta_z^* = [z_0, z_1, \cdots, z_{n-1}]^T \in R^n \tag{A.94}$$

and then use a design equation to calculate the parameters of an output feedback controller from the adaptive estimates of θ_p^* and θ_z^*.

Parameter estimation. Let the estimates of θ_p^* and θ_z^* be

$$\theta_p = [\hat{p}_0, \hat{p}_1, \cdots, \hat{p}_{n-1}]^T \in R^n, \ \theta_z = [\hat{z}_0, \hat{z}_1, \cdots, \hat{z}_{n-1}]^T \in R^n \tag{A.95}$$

and choose a monic stable polynomial $\Lambda(s)$ as

$$\Lambda(s) = s^n + \lambda_{n-1}s^{n-1} + \cdots + \lambda_1 s + \lambda_0. \tag{A.96}$$

Then, operating both sides of (A.90) by the stable filter $1/\Lambda(s)$ and using

$$\theta^* = [\theta_z^{*T}, (\theta_\lambda - \theta_p^*)^T]^T, \ \theta_\lambda = [\lambda_0, \cdots, \lambda_{n-1}]^T \in R^n \tag{A.97}$$

$$\begin{aligned}\phi(t) = \Bigg[& \frac{1}{\Lambda(s)}[u](t), \frac{s}{\Lambda(s)}[u](t), \cdots, \frac{s^{n-2}}{\Lambda(s)}[u](t), \frac{s^{n-1}}{\Lambda(s)}[u](t), \\ & \frac{1}{\Lambda(s)}[y](t), \frac{s}{\Lambda(s)}[y](t), \cdots, \frac{s^{n-2}}{\Lambda(s)}[y](t), \frac{s^{n-1}}{\Lambda(s)}[y](t)\Bigg]^T \in R^{2n}\end{aligned} \tag{A.98}$$

we express the plant (A.90) as

$$y(t) = \frac{Z(s)}{\Lambda(s)}[u](t) + \frac{\Lambda(s) - P(s)}{\Lambda(s)}[y](t) = \theta^{*T}\phi(t). \tag{A.99}$$

Denote $\theta(t)$ as the estimate of θ^*:

$$\theta(t) = [\theta_z^T(t), (\theta_\lambda - \theta_p(t))^T]^T \tag{A.100}$$

and define the estimation error

$$\epsilon(t) = \theta^T(t)\phi(t) - y(t). \tag{A.101}$$

Using (A.99) and (A.101), we have

$$\epsilon(t) = \tilde{\theta}^T(t)\phi(t), \ \tilde{\theta}(t) = \theta(t) - \theta^*. \tag{A.102}$$

Based on the error model (A.102), we can use some adaptive algorithms to update $\theta(t)$. For example, the normalized gradient algorithm is

$$\dot{\theta}(t) = -\frac{\Gamma\phi(t)\epsilon(t)}{m^2(t)}, \ \theta(0) = \theta_0, \ \Gamma = \Gamma^T > 0 \tag{A.103}$$

where $m(t) = \sqrt{1 + \alpha\phi^T(t)\phi(t)}$ with $\alpha > 0$.

This algorithm guarantees that $\theta(t)$, $\dot{\theta}(t)$, and $\epsilon(t)/m(t)$ are bounded, and $\dot{\theta}(t)$ and $\epsilon(t)/m(t)$ belong to L^2, as verified using the positive definite function $V(\tilde{\theta}) = \tilde{\theta}^T \Gamma^{-1} \tilde{\theta}$ whose time derivative is $\dot{V} = -2\epsilon^2(t)/m^2(t)$.

Control design. Let the estimates of $P(\lambda)$ and $Z(\lambda)$ be

$$\hat{P}(\lambda, \theta_p) = \lambda^n + \hat{p}_{n-1}\lambda^{n-1} + \cdots + \hat{p}_1\lambda + \hat{p}_0 \tag{A.104}$$

$$\hat{Z}(\lambda, \theta_z) = \hat{z}_{n-1}\lambda^{n-1} + \hat{z}_{n-2}\lambda^{n-2} + \cdots + \hat{z}_1\lambda + \hat{z}_0 \tag{A.105}$$

such that $Q(\lambda)\hat{P}(\lambda, \theta_p)$ and $\hat{Z}(\lambda, \theta_z)$ are co-prime polynomials.[3]

Choose a monic desired closed-loop characteristic polynomial $A^*(s)$ of degree $2n + n_q - 1$ which has all its zeros in $Re[s] < 0$.

Find the polynomials $C(\lambda, \psi_c)$ and $D(\lambda, \psi_d)$:

$$C(\lambda, \psi_c) = \lambda^{n-1} + c_{n-2}\lambda^{n-2} + \cdots + c_1\lambda + c_0 \tag{A.106}$$

$$D(\lambda, \psi_d) = d_{n_q+n-1}\lambda^{n_q+n-1} + \cdots + d_1\lambda + d_0 \tag{A.107}$$

where $\psi_c = [c_0, c_1, \cdots, c_{n-2}]^T \in R^{n-1}$ and $\psi_d = [d_0, d_1, \cdots, d_{n+n_q-1}]^T \in R^{n+n_q}$, by solving the Diophantine equation[4]

$$C(\lambda, \psi_c)Q(\lambda)\hat{P}(\lambda, \theta_p) + D(\lambda, \psi_d)\hat{Z}(\lambda, \theta_z) = A^*(\lambda) \tag{A.108}$$

and obtain $C(s, \psi_c)$ and $D(s, \psi_d)$ by the substitution

$$C(s, \psi_c) = C(\lambda, \psi_c)|_{\lambda=s} = s^{n-1} + \cdots + c_1 s + c_0 \tag{A.109}$$

$$D(s, \psi_d) = D(\lambda, \psi_d)|_{\lambda=s} = d_{n_q+n-1}s^{n_q+n-1} + \cdots + d_1 s + d_0. \tag{A.110}$$

Note that as an operator, $c_i s^i$, $i = 1, \ldots, n-1$, is different from $s^i c_i$ if $c_i = c_i(t)$ is a time-varying signal, and so is $d_j s^j$, $j = 1, \ldots, n_q + n - 1$, different from $s^j d_j$ if $d_j = d_j(t)$ is a time-varying signal.

Then the pole placement controller structure is

$$u(t) = (\Lambda_1(s) - C(s, \psi_c)Q(s))\frac{1}{\Lambda_1(s)}[u](t) + D(s, \psi_d)\frac{1}{\Lambda_1(s)}[y_m - y](t) \tag{A.111}$$

where $\Lambda_1(s) = \Lambda_0(s)\Lambda_c(s)$, with $\Lambda_0(s)$ and $\Lambda_c(s)$ being some monic and stable polynomials of degrees $n_q - 1$ and n, respectively.

Under the condition that $Q(\lambda)\hat{P}(\lambda, \theta_p)$ and $\hat{Z}(\lambda, \theta_z)$ are co-prime, this adaptive control scheme ensures that all signals in the closed-loop system are bounded and that $\lim_{t\to\infty}(y(t) - y_m(t)) = 0$.

[3] The adaptive estimate $\theta(t)$ from (A.103) does not automatically have this property. Special modifications of such $\theta(t)$ are needed for this property [55].

[4] If $Q(\lambda)\hat{P}(\lambda, \theta_p)$ and $\hat{Z}(\lambda, \theta_z)$ are co-prime, then for any $A^*(\lambda)$ of degree n_q+n-1, this equation has a unique solution $(C(\lambda, \phi_c),\ D(\lambda, \psi_d))$.

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Index