Adaptive Control Design and Analysis

(Supplemental Notes)

Gang Tao
Department of Electrical and Computer Engineering
University of Virginia
Charlottesville, VA 22903

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Contents

Matrix Theory 1
Radial Unboundedness Condition on $V(x)$ for Asymptotic Stability 2
Passivity of A Mass-Damper-Spring System 3
Positive Real Functions 4
Parameter Projection Properties 6
Explicit Swapping Lemma 7
Discrete-Time Swapping Lemma 9
Convergence of Discrete-Time Gradient Algorithm 10
Plant Zeros and Controller Order for MRAC 11
Tracking Performance in MRAC 14
Analysis of “Bursting Phenomenon” in Robust Adaptive Control 19
Robust MRAC with Switching $\sigma$-Modification 21
Discrete-time MRAC System Example 22
Indirect Observer Based Adaptive Control 24
Adaptive Control of $\dot{x}(t) = Ax(t) + B\Lambda u(t) + B\Theta^* \phi(x)$ 27
Parametrization of (9.102) 29
<table>
<thead>
<tr>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plant Signal Identities for MIMO Cases</td>
<td>30</td>
</tr>
<tr>
<td><strong>Proof of</strong> $\lim_{t \to \infty} e(t) = 0$ in (9.187C)</td>
<td>33</td>
</tr>
<tr>
<td>Adaptive Robot Control for Time-Varying Parameters</td>
<td>34</td>
</tr>
<tr>
<td>Derivation of (10.152)</td>
<td>40</td>
</tr>
<tr>
<td>Backstepping Design with Nonsmooth Inverse Signal</td>
<td>41</td>
</tr>
</tbody>
</table>
Matrix Theory

Symmetric Matrices

For a matrix \( M = M^T \), we have \( M = \sum_{i=1}^{n} \lambda_i e_i e_i^T \) where \( \lambda_i \) and \( e_i \) are the eigenvalues and eigenvectors of \( M \) such that \( e_i^T e_i = 1 \) and \( e_i^T e_j = 0 \) with \( i \neq j \). With \( P = [e_1, e_2, \ldots, e_n] \) and \( \Lambda = \text{diag}\{\lambda_1, \lambda_2, \ldots, \lambda_n\} \), it follows that \( M = P \Lambda P^T \), where \( PP^T = P^T P = I \), and, in addition, for \( M \) nonsingular, that \( M^{-1} = P \Lambda^{-1} P^T \).

For \( M = M^T \geq 0 \), we define \( M^{1/2} = \sum_{i=1}^{n} \sqrt{\lambda_i} e_i e_i^T = P \Lambda^{1/2} P^T \) and express \( M = M^{1/2} M^{1/2} \), where \( (M^{1/2})^T = M^{1/2} \). It also follows that for \( M \) nonsingular, \( (M^{1/2})^{-1} = \sum_{i=1}^{n} \frac{1}{\sqrt{\lambda_i}} e_i e_i^T = P \Lambda^{-1/2} P^T \). On the other hand, for \( Q = P \Lambda^{1/2} \), we have \( M = QQ^T \) (as compared with \( M = M^{1/2} M^{1/2} \) for \( (M^{1/2})^T = M^{1/2} = P \Lambda^{1/2} P^T \)).
Radial Unboundedness Condition on $V(x)$ for Asymptotic Stability

One can draw the surface plots of $V(x) = c$, for different values of $c$, with

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

For this $V(x)$, such surface plots are closed curves for $c < 1$ but are open curves for $c > 1$, as $V(x)$ is not radially unbounded.

One can draw the phase-plane plot of $\dot{x} = f(x)$, by obtaining its numerical solutions for some typical initial conditions, for

$$\dot{x}_1 = -\frac{6x_1}{(1 + x_1^2)^2} + 2x_2, \quad \dot{x}_2 = -\frac{2(x_1 + x_2)}{(1 + x_1^2)^2}. \hfill (2)$$

For some initial conditions, the solution trajectories do not converge to the origin. (Are there any trajectories going to $\infty$?)

One can draw the vector field of $\dot{x}_1 = -\frac{6x_1}{(1 + x_1^2)^2} + 2x_2, \dot{x}_2 = -\frac{2(x_1 + x_2)}{(1 + x_1^2)^2}$.

The solution $x = [x_1, x_2]^{T}$ of this system on the hyperbola $x_2 = \frac{2}{x_1 - \sqrt{2}}$ satisfies

$$g_1(x) = \frac{\dot{x}_2}{\dot{x}_1} = -\frac{1}{1 + 2\sqrt{2}x_1 + 2x_1^2}, \hfill (3)$$

while the slope of this hyperbola is

$$g_2(x) = \frac{dx_2}{dx_1} = -\frac{1}{1 - 2\sqrt{2}x_1 + \frac{x_1^2}{2}}. \hfill (4)$$

It follows that $0 > g_1(x) > g_2(x)$ for $x_1 > \sqrt{2}$ so that the trajectories to the right of the hyperbola branch in the first quadrant cannot cross the branch.

This example shows that for asymptotic stability, the radial unboundedness of $V(x)$ is a crucial condition [179], [351], [426].
Passivity of A Mass-Damper-Spring System

Consider a mass-damper-spring mechanical system with equation of motion

$$M\ddot{x} + D\dot{x} + Kx = F,$$

(5)

where $M$ is the mass, $D$ is the damping constant, $K$ is the spring constant, $x$ is the mass position, and $F$ is the force acting on the mass.

The system energy is

$$V(x, \dot{x}) = \frac{1}{2} M\dot{x}^2 + \frac{1}{2} Kx^2$$

(6)

whose time derivative is

$$\frac{d}{dt}V(x, \dot{x}) = F\dot{x} - Dx^2.$$  

(7)

Over any time interval $[0, T]$, it follows that

$$V(x(T), \dot{x}(T)) = V(x(0), \dot{x}(0)) + \int_0^T F(t)\dot{x}(t)\, dt - \int_0^TDx^2(t)\, dt.$$  

(8)

Since $D \geq 0$, we have

$$-\int_0^T F(t)\dot{x}(t)\, dt \leq V(x(0), \dot{x}(0)),$$

(9)

which means that the energy extracted from the system is less than or equal to the initial system energy. From (7), the term $F\dot{x}$ clearly represents the system absorbed power from the input force $F$. For passivity analysis, the product of system input and output is defined as such power. In this sense we consider the velocity $v = \dot{x}$ as system output. Then, the system admittance (the reciprocal of impedance) is

$$G(s) = \frac{V(s)}{F(s)} = \frac{s}{Ms^2 + Ds + K},$$

(10)

which is positively real (the mechanical-electric analogy pairs are “force vs. voltage” and “velocity vs. current”).

If we consider $x$ as system output, the term $Fx$ does not represents system power (i.e., $\int_0^T F(t)x(t)\, dt$ does not represents system energy) so that the passivity analysis is not applicable. In other words, the system transfer function for passivity analysis in terms of positive realness is defined in terms of system impedance (or admittance) relating current (velocity) to voltage (force) (or voltage (force) to current (velocity)).

Without the passivity property, the transfer function from the input force to the output position, $\frac{1}{Ms^2 + Ds + K}$, cannot be positive real.
Positive Real Functions

A popular definition of positive real (PR) functions is that a function $h(s)$ of the complex variable $s = \sigma + j\omega$ is positive real if (i) $h(s)$ is real for real $s$, and (ii) $\text{Re}[h(s)] \geq 0$ for all $s$ such that $\text{Re}[s] > 0$.

One may “induce” a definition of strictly positive real (SPR) functions as: a function $h(s)$ of the complex variable $s = \sigma + j\omega$ is strictly positive real if (i) $h(s)$ is real for real $s$, and (ii) $\text{Re}[h(s)] > 0$ for all $s$ such that $\text{Re}[s] > 0$. This definition was once used in the early literature for SPR functions. It turns out that, like some other definitions of SPR functions, this definition does not capture the physical meaning of strictly positive realness, as indicated by the following example.

It is well-understood that a proper definition of strictly positive functions, which captures the physical meaning of strictly positive realness, is that $h(s)$ is strictly positive real if $h(s - \varepsilon)$ is positive real for some $\varepsilon > 0$. Based on this definition,

$$h(s) = \frac{s + 1}{s^2 + s + 1}$$  \hspace{1cm}(11)$$

is only positive real but not strictly positive real. From the expressions

$$h(s) = \frac{\sigma + j\omega + 1}{(\sigma + j\omega)^2 + \sigma + j\omega + 1} = \frac{\sigma + j\omega}{\sigma^2 - \omega^2 + \sigma + 1 + j(\omega + 2\sigma\omega)}$$

$$= \frac{\sigma + 1 + j\omega)(\sigma^2 - \omega^2 + \sigma + 1 - j(\omega + 2\sigma\omega))}{(\sigma^2 - \omega^2 + \sigma + 1)^2 + (\omega + 2\sigma\omega)^2}$$  \hspace{1cm}(12)$$

$$\text{Re}[h(s)] = \frac{\sigma + 1)(\sigma^2 - \omega^2 + \sigma + 1) + \omega(\omega + 2\sigma\omega)}{(\sigma^2 - \omega^2 + \sigma + 1)^2 + (\omega + 2\sigma\omega)^2}$$

$$= \frac{\sigma^3 + 2\sigma^2 + \sigma\omega^2 + 2\sigma + 1}{(\sigma^2 - \omega^2 + \sigma + 1)^2 + (\omega + 2\sigma\omega)^2},$$  \hspace{1cm}(13)$$

we see that $h(s)$ satisfies the above “induced” definition of SPR functions: (i) $h(s)$ is real for real $s$, and (ii) $\text{Re}[h(s)] > 0$ for all $s$ such that $\text{Re}[s] = \sigma > 0$. Hence, the conclusion is that this “induced” definition for SPR functions is not proper.

Note that this $h(s)$ is not SPR, because

$$\text{Re}[h(j\omega)] = \frac{1}{(1 - \omega^2)^2 + \omega^2}$$  \hspace{1cm}(14)$$

does not satisfy the necessary condition for SPRness: $\lim_{\omega^2 \to \infty} \omega^2 \text{Re}[h(j\omega)] > 0$, or because for any chosen $\varepsilon > 0$,

$$\text{Re}[h(j\omega - \varepsilon)] = \frac{-\varepsilon^3 + 2\varepsilon^2 - \varepsilon\omega^2 - 2\varepsilon + 1}{(\varepsilon^2 - \omega^2 - \varepsilon + 1)^2 + (\omega - 2\varepsilon\omega)^2} < 0$$  \hspace{1cm}(15)$$
whenever $\omega^2 > (-\epsilon^3 + 2\epsilon^2 - 2\epsilon + 1)/\epsilon$, that is, $h(s - \epsilon)$ cannot be positive real for any $\epsilon > 0$. $(P(\epsilon) = -\epsilon^3 + 2\epsilon^2 - 2\epsilon + 1 = (\epsilon - 1)(-\epsilon^2 + \epsilon - 1) > 0$ for $\epsilon \in (0,1)$ and $P(\epsilon) < 0$ for $\epsilon > 1$. For $h(s - \epsilon)$ to be stable, $\epsilon \in [0,0.5)$ is needed as $(s - \epsilon)^2 + s - \epsilon + 1 = s^2 + (1 - 2\epsilon)s + 1 - \epsilon + \epsilon^2$ and $1 - \epsilon + \epsilon^2 > 0$ for any $\epsilon$.}
Parameter Projection Properties

The property (3.183) follows from the observation: if $\theta_j(t) = \theta^a_j$ and $g_j(t) < 0$, then $f_j(t) = -g_j(t) > 0$ and $\theta_j(t) - \theta^*_j = \theta^a_j - \theta^*_j \leq 0$, so that $(\theta_j(t) - \theta^*_j)f_j(t) \leq 0$; and if $\theta_j(t) = \theta^b_j$ and $g_j(t) > 0$, then $f_j(t) = -g_j(t) < 0$ and $\theta_j(t) - \theta^*_j = \theta^b_j - \theta^*_j \geq 0$, so that $(\theta_j(t) - \theta^*_j)f_j(t) \leq 0$.

Similarly, the property (3.214) follows from the observation: if $\theta_j(t) + g_j(t) > \theta^b_j$, then $f_j(t) = \theta^b_j - \theta_j(t) - g_j(t) < 0$ and $\theta_j(t) - \theta^*_j + g_j(t) + f_j(t) = \theta^b_j - \theta^*_j \geq 0$, so that $f_j(t)(\theta_j(t) - \theta^*_j + g_j(t) + f_j(t)) \leq 0$; and if $\theta_j(t) + g_j(t) < \theta^b_j$, then $f_j(t) = \theta^b_j - \theta_j(t) - g_j(t) > 0$ and $\theta_j(t) - \theta^*_j + g_j(t) + f_j(t) = \theta^b_j - \theta^*_j \leq 0$, so that $f_j(t)(\theta_j(t) - \theta^*_j + g_j(t) + f_j(t)) \leq 0$. 
Explicit Swapping Lemma

The swapping lemma (5.331) was an important lemma in the development of a stable model reference adaptive control system. It states that for a stable and proper rational function \( h(s) \) with a minimal realization \( h(s) = c(sI - A)^{-1}b + d \) and two vector signals \( \Theta(t) \) and \( \omega(t) \), it follows that

\[
\Theta^T(t)h(s)[\omega](t) - h(s)[\Theta^T \omega](t) = h_c(s)[h_b(s)[\omega^T \dot{\Theta}](t),
\]

where \( h_c(s) = c(sI - A)^{-1} \) and \( h_b(s) = (sI - A)^{-1}b \). Here we derive an alternative form of this lemma, explicitly in terms of the parameters of the function \( h(s) \).

Denoting \( P_m(s) = s^{n^*} + an_{n^*-1}s^{n^*-1} + \cdots + a_1s + a_0 \), for vector signals \( \Theta(t) \) and \( \omega(t) \), from (5.138) and with \( a_{n^*} = 1 \), we have

\[
\Theta^T(t) \left( \frac{1}{P_m(s)} \right)[\omega](t) - \frac{1}{P_m(s)}[\Theta^T \omega](t) = \sum_{i=1}^{n^*} \left( \frac{\sum_{j=0}^{n^*-i} a_{n^*-j}s^{n^*-i-j}}{P_m(s)} \right) \left[ \Theta^T \frac{s^{j-1}}{P_m(s)}[\omega] \right](t). \tag{17}
\]

Introducing \( F(s) = f_{n^*-1}s^{n^*-1} + \cdots + f_1s + f_0 \), we express

\[
F(s) \left[ \Theta^T(t) \frac{1}{P_m(s)}[\omega] \right](t) = \left( f_{n^*-1}s^{n^*-2} + \cdots + f_1 \right) \left[ \Theta^T \frac{1}{P_m(s)}[\omega] + \Theta^T \frac{s}{P_m(s)}[\omega] \right](t) + \Theta^T(t) \frac{f_0}{P_m(s)}[\omega](t)
\]

\[
= \left( f_{n^*-1}s^{n^*-2} + \cdots + f_1 \right) \left[ \Theta^T \frac{1}{P_m(s)}[\omega] \right](t)
\]

\[
+ \left( f_{n^*-1}s^{n^*-2} + \cdots + f_2 \right) \left[ \Theta^T \frac{s}{P_m(s)}[\omega] + \Theta^T \frac{s^2}{P_m(s)}[\omega] \right](t)
\]

\[
+ \Theta^T(t) \frac{f_1s + f_0}{P_m(s)}[\omega](t) = \cdots
\]

\[
= \sum_{i=1}^{n^*-1} \left( \sum_{j=1}^{n^*-i} \right) f_{n^*-j}s^{n^*-i-j} \left[ \Theta^T \frac{s^{j-1}}{P_m(s)}[\omega] \right](t) + \Theta^T(t) \frac{F(s)}{P_m(s)}[\omega](t) \tag{18}
\]

and use it to derive

\[
\frac{F(s)}{P_m(s)}[\Theta^T \omega](t) - \Theta^T(t) \frac{F(s)}{P_m(s)}[\omega](t) = \sum_{i=1}^{n^*-1} \left( \sum_{j=1}^{n^*-i} \right) f_{n^*-j}s^{n^*-i-j} \left[ \Theta^T \frac{s^{j-1}}{P_m(s)}[\omega] \right](t)
\]
\[
-F(s) \left[ \sum_{i=1}^{n^*} \alpha_i(s) \left( \sum_{j=0}^{n^*-i} a_{n^*-j}s^{n^*-i-j} \right) \right] \left( \dot{\theta}^T \frac{s^{i-1}}{P_m(s)} [\omega] \right)(t)
\]

- \[ \sum_{i=1}^{n^*-1} \left( \sum_{j=1}^{n^*-i} f_{n^*-j}s^{n^*-i-j} - \frac{F(s) \sum_{j=0}^{n^*-i} a_{n^*-j}s^{n^*-i-j}}{P_m(s)} \right) \left[ \dot{\theta}^T \frac{s^{i-1}}{P_m(s)} [\omega] \right](t)
\]

- \[ \left[ \frac{F(s)}{P_m(s)} \left( \dot{\theta}^T \frac{s^{n^*-1}}{P_m(s)} [\omega] \right) \right](t)
\]

\[
= \sum_{i=1}^{n^*-1} \frac{\alpha_i(s)}{P_m(s)} \left[ \dot{\theta}^T \frac{s^{i-1}}{P_m(s)} [\omega] \right](t) - \frac{F(s)}{P_m(s)} \left[ \dot{\theta}^T \frac{s^{n^*-1}}{P_m(s)} [\omega] \right](t).
\] (19)

Hence, we obtain the explicit swapping lemma:

\[
\frac{F(s)}{P_m(s)} [\theta^T \omega](t) - \theta^T(t) \left[ \frac{F(s)}{P_m(s)} [\omega] \right](t) = \sum_{i=1}^{n^*} \frac{\alpha_i(s)}{P_m(s)} \left[ \dot{\theta}^T \frac{s^{i-1}}{P_m(s)} [\omega] \right](t),
\] (20)

where

\[
\alpha_i(s) = P_m(s) \sum_{j=1}^{n^*-i} f_{n^*-j}s^{n^*-i-j} - F(s) \sum_{j=0}^{n^*-i} a_{n^*-j}s^{n^*-i-j}, \quad i = 1, 2, \ldots, n^* - 1,
\] (21)

are polynomials of degrees \( n^* - 1 \) or less, and \( \alpha_{n^*}(s) \triangleq -F(s) \).

For a proper transfer function \( h(s) = f_{n^*} + \frac{F(s)}{P_m(s)} \), it follows that

\[
h(s) [\theta^T \omega](t) - \theta^T(t) h(s)[\omega](t) = \sum_{i=1}^{n^*} \frac{\alpha_i(s)}{P_m(s)} \left[ \dot{\theta}^T \frac{s^{i-1}}{P_m(s)} [\omega] \right](t).
\] (22)
Discrete-Time Swapping Lemma

Let a stable and proper rational function \( h(z) \) have a minimal realization \( h(z) = c(zI - A)^{-1}b + d \) and \( \Theta(t) \) and \( \omega(t) \) be two vector signals, and denote \( h_c(z) = c(zI - A)^{-1} \) and \( h_b(z) = (zI - A)^{-1}b \). Then,

\[
\theta^T(t)h(z)[\omega](t) - h(z)[\theta^T \omega](t) = h_c(z)[(h_b(z)z)[\omega^T](z-1)[\theta]](t). \tag{23}
\]

**Proof:** Using the discrete-time convolution: \( y(t) = C\sum_{i=0}^{t-1}A^{t-i-1}Bu(i) \) for \( y(t) = C(zI - A)^{-1}B[u](t) \), and its modification: \( w(t) = C\sum_{i=0}^{t-1}A^{t-i-1}Bu(i) \) for \( w(t) = C(zI - A)^{-1}Bz[u](t) = C(zI - A)^{-1}B[u](t+1) \), we express

\[
h_c(z)[(h_b(z)z)[\omega^T](z-1)[\theta]](t)
= c \sum_{\tau=0}^{t-1} A^{t-\tau-1} \sum_{i=0}^{\tau} A^{\tau-i}b \omega^T(i) (\theta(\tau + 1) - \theta(\tau))
\]

\[
= c \sum_{i=0}^{t-1} A^{t-i-1}b \omega^T(i) \theta(t)
\]

\[
+ c \sum_{\tau=0}^{t-2} A^{t-\tau-1} \sum_{i=0}^{\tau} A^{\tau-i}b \omega^T(i) (\theta(\tau + 1) - c \sum_{\tau=0}^{t-1} A^{t-\tau-1} \sum_{i=0}^{\tau} A^{\tau-i}b \omega^T(i) \theta(\tau))
\]

\[
= \theta^T(t)h(z)[\omega](t) + c \sum_{\sigma=1}^{t-1} A^{t-\sigma-1} \sum_{i=0}^{\sigma-1} A^\sigma b \omega^T(i) \theta(\sigma)
\]

\[
- c \sum_{\tau=0}^{t-1} A^{t-\tau-1}b \omega^T(\tau) \theta(\tau) - c \sum_{\tau=1}^{t-1} A^{t-\tau-1} \sum_{i=0}^{\tau-1} A^{\tau-i}b \omega^T(i + 1) \theta(\tau)
\]

\[
= \theta^T(t)h(z)[\omega](t) - h(z)[\theta^T \omega](t), \tag{24}
\]

where, by definition,

\[
c \sum_{i=0}^{t-1} A^{t-i-1}b \omega(i) = h(z)[\omega](t) \tag{25}
\]

\[
c \sum_{\tau=0}^{t-1} A^{t-\tau-1}b \omega^T(\tau) \theta(\tau) = h(z)[\theta^T \omega](t). \tag{26}
\]

This is the discrete-time version of the swapping lemma (5.331), whose explicit version can also be derived, similar to the continuous-time case above.
Convergence of Discrete-Time Gradient Algorithm

This section gives the proof of Theorem 3.2 which specifies the conditions for parameter convergence of the discrete-time gradient algorithm (3.83).

Consider the linear time-invariant system described by the difference equation

\[ P(z)[y](t) = Z(z)[u](t), \quad (27) \]

where \( y(t) \in R \) and \( u(t) \in R \) are the measured system input and output, and

\[ P(z) = z^n + p_{n-1}z^{n-1} + \cdots + p_1z + p_0, \quad (28) \]

\[ Z(z) = z_mz^n + z_{m-1}z^{m-1} + \cdots + z_1z + z_0. \quad (29) \]

Choose a stable polynomial \( \Lambda(z) = z^n + \hat{\lambda}_n z^{n-1} + \cdots + \hat{\lambda}_1z + \hat{\lambda}_0 \) whose zeros are in \( |z| < 1 \), and express (27) as

\[ y(t) = \theta^T \phi(t), \quad (30) \]

where

\[ \theta^* = [z_0, z_1, \ldots, z_{m-1}, z_m, \lambda_0 - p_0, \lambda_1 - p_1, \ldots, \lambda_{n-2} - p_{n-2}, \lambda_{n-1} - p_{n-1}]^T \in \mathbb{R}^{n+m+1}, \quad (31) \]

\[ \phi(t) = \left[ \frac{1}{\Lambda(z)}[u](t), \frac{z}{\Lambda(z)}[u](t), \ldots, \frac{z^{m-1}}{\Lambda(z)}[u](t), \frac{z^m}{\Lambda(z)}[u](t), \frac{1}{\Lambda(z)}[y](t), \frac{z}{\Lambda(z)}[y](t), \ldots, \frac{z^{n-2}}{\Lambda(z)}[y](t), \frac{z^{n-1}}{\Lambda(z)}[y](t) \right]^T \in \mathbb{R}^{n+m+1}. \quad (32) \]

Assume that the system input \( u(t) \) has \( n+m+1 \) frequencies, that is,

\[ u(t) = \alpha_0 + \sum_{i=1}^{N} (\alpha_i \sin \omega_i t + \alpha_{N+i} \cos \omega_i t), \quad (33) \]

where \( \omega_j \neq \omega_k + 2\pi l \) for \( j \neq k \) and an integer \( l \), \( \alpha_0 = 0 \) and \( N = \frac{n+m+1}{2} \) for an even \( n+m+1 \) (or \( \alpha_0 \neq 0 \) and \( N = \frac{n+m}{2} \) for an odd \( n+m+1 \)), and \( \alpha_i \neq 0, i = 1, 2, \ldots, 2N. \)

**Property 1** There exist an integer \( \delta_1 > 0 \) and a constant \( \alpha_1 > 0 \) such that

\[ \sum_{t=\sigma}^{\sigma+\delta_1} U(t, t+n+m)U^T(t, t+n+m) \geq \alpha_1 \mathbb{I}, \quad \forall \sigma \geq 0, \quad (34) \]

where \( U(t, t+n+m) = [u(t), u(t+1), \ldots, u(t+n+m)]^T \in \mathbb{R}^{n+m+1}. \)

**Proof:**
Plant Zeros and Controller Order for MRAC

Consider a LTI system
\[ \dot{x} = Ax + Bu, \quad y = Cx \]
where \( A \in \mathbb{R}^{n \times n}, B \in \mathbb{R}^{n \times M}, C \in \mathbb{R}^{M \times n} \), whose transfer matrix \( G(s) = C(sI - A)^{-1}B \) is an \( M \times M \) strictly proper rational matrix.

For SISO systems with \( M = 1 \), we have \( G(s) = \frac{Z(s)}{P(s)} \) for some nominal polynomials \( Z(s) \) and \( P(s) \) with \( \partial P(s) = n_p \leq n \) (if \( P(s) \) is taken as \( P(s) = \det[sI - A] \), then \( n_p = n \)). If we take \( n \) as the system order for MRAC design, that is, the controller order is \( n - 1 = \partial \Lambda(s) \), then we need to assume that all zeros of \( Z(s) \) (which are the system zeros) are stable, for a stable MRAC system. With such a full-order representation \( P(s) = \det[sI - A] \), the condition that all zeros of \( Z(s) \) are stable implies that the realization \((A, B, C)\) is stabilizable and detectable. The choice of the controller order \( n - 1 \) has another interpretation in the SISO case: the output feedback controller structure is a reparametrization of a state feedback control law \( u = k_1^T x + k_2^T r \), using a reduced-order (with order \( n - 1 \)) state observer (see Section 4.4.2).

For SISO systems, if, due to pole-zero cancellations caused by a nonminimal realization \((A, B, C)\), the degree \( n_p \) of \( P(s) \) is less than \( n \): \( n_p < n \), and the controller order is \( n_p - 1 \), we need to assume that, in addition to all zeros of \( Z(s) \) being stable, the pole-zero cancellations leading \( G(s) = C(sI - A)^{-1}B \) to \( \frac{Z(s)}{P(s)} \) should be also stable ones, that is, the nonminimal realization \((A, B, C)\) should be stabilizable and detectable. In this case, the system zeros include the zeros of \( Z(s) \) and those cancelled zeros in \( G(s) \), while the transfer function zeros are those \( s_z \) which make \( \frac{Z(s_z)}{P(s_z)} \) zero and they can be a subset of the zeros of \( Z(s) \), as there may be some more zero-pole cancellations between \( Z(s) \) and \( P(s) \) in \( \frac{Z(s_z)}{P(s_z)} \).

In the SISO case, a state feedback control design \( u(t) = K_1^T x(t) + K_2 r(t) \) requires that all zeros of \( Z(s) \) are stable (in this case, \( P(s) = \det[sI - A] \) which is necessary and sufficient for stable plant-model matching: \( \det[sI - A - BK_1^T] = P_n(s)Z(s)/z_m \) (not needed for just pole placement control). An output feedback control design is based on a nominal transfer function \( \frac{Z(s)}{P(s)} \) with order \( n_p = \partial P(s) \), and all zeros of \( Z(s) \) need to be stable for stable plant-model matching based on \( \frac{Z(s)}{P(s)} \). If \( n \leq n_p \), a stable \( Z(s) \) implies that \((A, B, C)\) is stabilizable and detectable. If \( n > n_p \), then there are zero-pole cancellations in \( C(sI - A)^{-1}B = \frac{Z(s)}{P(s)} \), and \((A, C)\) needs to be detectable, otherwise those undetectable modes (eigenvalues of \( A \)) cannot be stabilized (output feedback does not change unobservable modes); \((A, B)\) needs to be stabilizable, otherwise those unstabilizable modes (eigenvalues of \( A \)) cannot be stabilized (output or state feedback does not change uncontrollable modes).
For MIMO systems, there are also different ways to express the system transfer matrix $G(s) = C(sI - A)^{-1}B$ (we first assume that $G(s)$ has full rank $m$ for an $M \times M$ $G(s)$, that is, $\text{rank}[G(s)] < M$ only for a finite number of values of $s$).

The first (simple) form of $G(s)$ is

$$G(s) = C(sI - A)^{-1}B = \frac{N(s)}{d(s)}, \quad d(s) = \det[sI - A], \quad \partial d(s) = n. \quad (36)$$

If we use a controller of order $n - 1 = \partial \Lambda(s)$, we need to assume that all zeros of $\det[N(s)]$ are stable, which may be thought as the system zeros (note that the zeros of $G(s)$, as defined in the literature, may be only a subset of those zeros if $(A, B, C)$ is not a minimal realization). In fact, different from the SISO case, the controller order can be reduced to $n - M$ in the MIMO case, because a reduced-order observer has order $n - M$, so that the state feedback control law $u = K_1^Tx + K_2^Tr$ can be reparametrized using the input and output signals, leading to an $(n - M)$th order output feedback controller.

Given that the observability index of a minimal realization of $G(s)$ is $v \leq n - M + 1$, that is, $v - 1 \leq n - M$, the order of the usual MRAC structure, which has been chosen as $v - 1$, is the minimal order to meet the desired plant-model matching. If for a system, the order of a reduced-order observer can be chosen as $v - 1$, the parametrization of the output feedback controller with order $v - 1$ can then be seen as a reparametrization of a state feedback controller, using input and output signals. If not, the controller parametrization may be considered as being one used for plant-model matching. In MRAC for multivariable systems, we call $v$ the observability index of the system transfer matrix $G(s)$, in the sense that $v$ is the observability index of a minimal realization of the system transfer matrix $G(s)$.

The second form of $G(s)$ is a left matrix-fraction description:

$$G(s) = C(sI - A)^{-1}B = \bar{P}_1^{-1}(s)\bar{Z}_l(s) \quad (37)$$

where $\bar{P}_1(s)$ and $\bar{Z}_l(s)$ are $M \times M$ polynomial matrices with $\bar{P}_1(s)$ row reduced (a polynomial matrix $\bar{P}_1(s)$ is row reduced if the elements in the $i$th row of $\bar{P}_1(s)$ have a largest degree $\bar{v}_i$ and the matrix $\Gamma_r = \lim_{s \to 0} \text{diag}\{s^{-\bar{v}_1}, s^{-\bar{v}_2}, \ldots, s^{-\bar{v}_M}\}\bar{P}_1(D)$ is nonsingular), and the $i$th row degree of $\bar{Z}_l(s)$ is less than the $i$th row degree of $\bar{P}_1(s)$, $i = 1, 2, \ldots, q$ (it is denoted that $\partial \bar{P}_1(s) = \bar{v}$, that is, $\bar{v} = \max\{\bar{v}_i\}$). A controller can be designed with $\partial \Lambda(s) = \bar{v} - 1$. In this case, for stable MRAC, we need to assume that the zeros of $\det[\bar{Z}_l(s)]$ are stable and also that $(A, B, C)$ is stabilizable and detectable (as some pole-zero cancellations may occur when obtaining the system model $\bar{P}_1^{-1}(s)\bar{Z}_l(s)$ from $(A, B, C)$).

Note that a left matrix-fraction description $G(s) = \bar{P}_1^{-1}(s)\bar{Z}_l(s)$, can always be made to have a row reduced $\bar{P}_1(s)$, by using elementary row operations represented
by a unimodular matrix \( M(s) \) (such a matrix is defined to have a non-zero constant determinant) on a non-reduced \( \tilde{P}_l(s) \):

\[
G(s) = \tilde{P}_l^{-1}(s)\tilde{Z}_l(s) = (M(s)\tilde{P}_l(s))^{-1}M(s)\tilde{Z}_l(s) = \tilde{P}_l^{-1}(s)\tilde{Z}_l(s), \quad (38)
\]

for \( \tilde{P}_l(s) = M(s)\tilde{P}_l(s) \) and \( \tilde{Z}_l(s) = M(s)\tilde{Z}_l(s) \).

The third form of \( G(s) \) is a left co-prime matrix-fraction description:

\[
G(s) = C(sI-A)^{-1}B = P_l^{-1}(s)Z_l(s) \quad (39)
\]

where \( P_l(s) \) and \( Z_l(s) \) are left co-prime \( M \times M \) polynomial matrices (\( P_l(s) \) and \( Z_l(s) \) are left co-prime if any \( M \times M \) polynomial matrix \( W(s) \) such that \( Z_l(s) = W(s)Z(s) \) and \( P_l(s) = W(s)P(s) \) for some polynomial matrices \( Z(s) \) and \( P(s) \)—such a \( W(s) \) is called a common left divisor of \( Z_l(s) \) and \( P_l(s) \), is a unimodular matrix, that is, \( \text{det}[W(s)] \) is a non-zero constant), with \( P_l(s) \) row reduced and \( \partial P_l(s) = \nu \leq \tilde{\nu} \). A controller can be designed with \( \partial A(s) = \nu - 1 \). In this case, we also need to assume that the zeros of \( \text{det}[Z_l(s)] \) are stable and that \( (A,B,C) \) is stabilizable and detectable.

The situation is similar for \( G(s) \) in a right matrix-fraction description:

\[
G(s) = C(sI-A)^{-1}B = Z_r(s)P_r^{-1} \quad (40)
\]

with \( P_r(s) \) being column reduced (that is, \( P_r^T(s) \) being row reduced) and \( Z_r(s) \) and \( P_r(s) \) being right co-prime (that is, \( Z_r^T(s) \) and \( P_r^T(s) \) being left co-prime). In particular, the column degrees of \( P_r(s) \) are denoted as \( \mu_i \) (or \( \tilde{\mu}_i \) for \( \tilde{P}_r(s) \) if \( G(s) = C(sI-A)^{-1}B = \tilde{Z}_r(s)\tilde{P}_r^{-1} \) with \( \tilde{Z}_r(s) \) and \( \tilde{P}_r(s) \) not right co-prime), \( i = 1, 2, \ldots, m \), and the column degrees of \( Z_r(s) \) (\( \tilde{Z}_r(s) \)) are less than that of \( P_r(s) \) (\( \tilde{P}_r(s) \)). If \( (A,B,C) \) is a minimal realization, then \( \mu_i, i = 1, 2, \ldots, m \), are the controllability indexes of \( (A,B) \). If \( (A,B,C) \) is not a minimal realization, then one can a controllable realization \( (\tilde{A}_c, \tilde{B}_c, \tilde{C}_c) \) whose controllability indexes are \( \tilde{\mu}_i \), for \( \tilde{Z}_r(s)\tilde{P}_r^{-1}(s) \) (or \( (A_c, B_c, C_c) \)) whose controllability indexes are \( \mu_i \), for \( Z_r(s)P_r^{-1} \).

Note that if \( (A,B,C) \) is a minimal realization, then \( \nu \) is the observability index of \( (A,C) \) (we may also call \( \nu \) the observability index of \( G(s) \)). When \( (A,B,C) \) is not minimal, for the second case (or the third case), we can find an observable realization \( (\tilde{A}_o, \tilde{B}_o, \tilde{C}_o) \) whose observability indexes are \( \tilde{\nu}_i \), for \( \tilde{P}_l^{-1}(s)\tilde{Z}_l(s) \) (or \( (A_o, B_o, C_o) \)) whose observability indexes are \( \nu_i \), for \( P_l^{-1}(s)Z_l(s) \). In this sense, we may also call \( \nu \) (or \( \nu \)) the observability of \( G(s) = \tilde{P}_l^{-1}(s)\tilde{Z}_l(s) \) (or \( G(s) = P_l^{-1}(s)Z_l(s) \)).

In summary, for MRAC of a system \( (A,B,C) \), a basic assumption is that \( (A,B,C) \) is stabilizable and detectable, in addition to the assumption that all zeros of the system transfer matrix \( G(s) = C(sI-A)^{-1}B \) are stable; in other words, all system zeros are required to be stable.

For more about the definitions of the zeros and poles of \( G(s) \), see Kailath (1980) and Rugh (1996), and also see more notes.
Tracking Performance in MRAC

It has been shown that for a linear time-invariant plant \( y(t) = G(s)[u](t) \), where \( G(s) = k_p \frac{Z(s)}{P(s)} \) with all zeros of \( Z(s) \) stable and with relative degree \( n^* = \) degree of \( P(s) - \) degree of \( Z(s) \), a model reference adaptive controller to generate the plant input \( u(t) \) ensures that the tracking error \( e(t) = y(t) - y_m(t) \) has the desired asymptotic convergence property: \( \lim_{t \to \infty} e(t) = 0 \), where \( y_m(t) = W_m(s)[r](t) \) is the reference output to be tracked by the plant output \( y(t) \), for a bounded reference input signal \( r(t) \) and a stable transfer function \( W_m(s) = \frac{1}{P_m(s)} \) with a stable polynomial \( P_m(s) \) of degree \( n^* \). In this work, we show that the tracking error \( e(t) \) in a model reference adaptive control (MRAC) system has some stronger convergence properties as stated in the following theorem.

**Theorem 1** For a model reference adaptive control system with the plant relative degree \( n^* > 0 \), the tracking error \( e(t) = y(t) - y_m(t) \) has the convergence property: \( \lim_{t \to \infty} \frac{d^i e(t)}{dt^i} = 0 \), for \( i = 0, 1, \ldots, n^* - 1 \).

**Proof:** The property: \( \lim_{t \to \infty} e(t) = 0 \) follows directly from the established properties: \( \int_0^\infty e^2(t)dt < \infty \) and \( \frac{de(t)}{dt} \) is bounded. However, it is not clear whether \( \frac{d^i e(t)}{dt^i} \) is square-integrable or not for \( i > 0 \). To prove the convergence of \( \frac{d^i e(t)}{dt^i} \), for \( i = 1, \ldots, n^* - 1 \), we recall that for a function \( f(t) \) defined on \( [t_0, \infty) \), \( \lim_{t \to \infty} f(t) = 0 \) if for every \( \eta > 0 \), there exists a \( T = T(\eta) > 0 \) such that \( |f(t)| < \eta \), \( \forall t > T \). Hence, our goal now is to show that in each case of \( i = 1, \ldots, n^* - 1 \), for any given \( \eta \), there exists a \( T > 0 \) such that \( |\frac{d^i e(t)}{dt^i}| < \eta \), for all \( t > T \). To reach this goal, we use a method to decompose the signal \( \frac{d^i e(t)}{dt^i} \) into two fictitious parts: one being small enough and one converging to zero asymptotically with time going to infinity. To this end, we first introduce two fictitious filters \( K(s) \) and \( H(s) \) from

\[
K(s) = \frac{a^{n^*}}{(s+a)^{n^*}}, \quad sH(s) = 1 - K(s) \quad (41)
\]

where \( a > 0 \) is a generic constant to be specified. The filter \( H(s) \) is given as

\[
H(s) = \frac{1}{s}(1 - K(s)) = \frac{1}{s} \frac{(s+a)^{n^*} - a^{n^*}}{(s+a)^{n^*}} \quad (42)
\]

which is strictly proper (with relative degree 1) and stable and whose impulse response function is

\[
h(t) = L^{-1}[H(s)] = e^{-at} \sum_{i=1}^{n^*} \frac{a^{n^*-i}}{(n^*-i)!} t^{n^*-i}. \quad (43)
\]
It can be verified that the $L^1$ signal norm of $h(t)$ is

$$\|h(\cdot)\|_1 = \int_0^\infty |h(t)| \, dt = \frac{n^*}{a}. \quad (44)$$

It has been established that the tracking error $e(t) = y(t) - y_m(t)$ satisfies

$$e(t) = \frac{k_p}{P_m(s)} [\tilde{\theta}^T \omega](t), \quad (45)$$

where $P_m(s)$ is a stable polynomial of degree $n^*$, and $\tilde{\theta}(t)$ is the parameter error vector and $\omega(t)$ is the controller regressor vector, and both are bounded. Using (41): $1 = sH(s) + K(s)$, we express $\dot{e}(t) = \frac{de(t)}{dt}$ as

$$\dot{e}(t) = \frac{k_p s}{P_m(s)} [\tilde{\theta}^T \omega](t)$$

$$= H(s) \frac{k_p s^2}{P_m(s)} [\tilde{\theta}^T \omega](t) + sK(s) \frac{k_p}{P_m(s)} [\tilde{\theta}^T \omega](t)$$

$$= H(s) \frac{k_p s^2}{P_m(s)} [\tilde{\theta}^T \omega](t) + sK(s)[e](t). \quad (46)$$

To demonstrate the new technique for proving the new result in Theorem 1: $\lim_{t \to \infty} \frac{d e(t)}{dt} = 0$, for $i = 1, \ldots, n^* - 1$, under the previously proved properties of MRAC: all closed-loop signals are bounded and the tracking error $e(t) = y(t) - y_m(t)$ satisfies $\lim_{t \to \infty} e(t) = 0$, we first consider the case of $n^* = 2$, where

$$K(s) = \frac{a^2}{(s+a)^2}, \quad H(s) = \frac{s+2a}{(s+a)^2}. \quad (47)$$

Since $\tilde{\theta}^T(t)\omega(t)$ is bounded and $\frac{k_p s^2}{P_m(s)}$ is stable and proper, $\frac{k_p s^2}{P_m(s)} [\tilde{\theta}^T \omega](t)$ is bounded.

It follows from the above $L^1$ signal norm expression of $H(s)$ that

$$|H(s) \frac{k_p s^2}{P_m(s)} [\tilde{\theta}^T \omega](t)| \leq \frac{c_0}{a}. \quad (48)$$

for any $t \geq 0$ and some constant $c_0 > 0$ independent of $a > 0$. Since $\lim_{t \to \infty} e(t) = 0$ as established and $sK(s)$ is stable and strictly proper (with relative degree $n^* - 1 = 1$ in this case of $n^* = 2$), it follows, for any $a > 0$ in $K(s)$, that

$$\lim_{t \to \infty} sK(s)[e](t) = 0. \quad (49)$$

\footnote{For a stable and strictly proper transfer function $W(s)$ and a system $z(t) = W(s)v(t)$, if $\lim_{t \to \infty} v(t) = 0$, then $\lim_{t \to \infty} z(t) = 0$ (see p. 68 of Naranda and Annaswamy (1989) and and p. 263 of Tao (2003)).}
To show that, for every \( \eta > 0 \), there exists a \( T > 0 \) such that \( |\dot{e}(t)| < \frac{\eta}{2} \), \( \forall t > T \), we set \( a > a(\eta) = \frac{2\epsilon_0}{\eta} \) for the fictitious filter \( H(s) \) in (42) so that \( \frac{\eta_0}{a} < \frac{3}{2} \) in (48), and let \( T = T_d(a(\eta), \eta) \geq T(\eta) > 0 \) such that \( |sK(s)e(t)| < \frac{\eta}{2} \) for all \( t \geq T \) (which is ensured by the property that \( \lim_{t \to \infty} sK(s)e(t) = 0 \) for any finite \( a > 0 \) in the other related fictitious filter \( K(s) \) in (41)) (the peak value of \( |sK(s)e(t)| \) depends on the parameter \( a \), so that the above time instant \( T = T_d(a(\eta), \eta) \) also depends on \( a \) too). Then, it follows from (46) and (48) that

\[
|\dot{e}(t)| < \frac{\eta}{2} + \frac{\eta}{2} = \eta, \quad \forall t > T,
\]

(50)

which implies that \( \lim_{t \to \infty} \dot{e}(t) = 0 \). In other words, if \( \lim_{t \to \infty} \dot{e}(t) \neq 0 \), then there exist an \( \eta_0 > 0 \) and a sequence of time instants \( t_i \) with \( \lim_{i \to \infty} t_i = \infty \) such that \( |\dot{e}(t_i)| > \eta_0 \) for all \( i = 1, 2, \ldots \), which, from the above analysis, is impossible.

Similarly, for the case when \( n^* = 3 \) with

\[
K(s) = \frac{a^3}{(s+a)^3}, \quad H(s) = \frac{s^2 + 3as + 3a^2}{(s+a)^3}, \quad (51)
\]

we see that in (46), \( \frac{k_p s^3}{P_m(s)} \) is stable and strictly proper, and so is \( sK(s) \), so that \( \lim_{t \to \infty} \dot{e}(t) = 0 \). To show \( \lim_{t \to \infty} \frac{d^2 e(t)}{dt^2} = 0 \), from (46), we express \( \frac{d^2 e(t)}{dt^2} \) as

\[
\frac{d^2 e(t)}{dt^2} = H(s) \frac{k_p s^3}{P_m(s)} [\ddot{\theta}^T \omega](t) + s^2 K(s)[e](t), \quad (52)
\]

in which \( \frac{k_p s^3}{P_m(s)} [\ddot{\theta}^T \omega](t) \) is bounded (as \( \frac{s^3}{P_m(s)} \) is stable and proper and \( \ddot{\theta}^T(t)\omega(t) \) is bounded) and \( H(s) \) satisfies (44), and \( \lim_{t \to \infty} s^2 K(s)[e](t) = 0 \). Hence, similar to (50), now with \( \frac{d^2 e(t)}{dt^2} \) replacing \( \dot{e}(t) \) in (50), it also follows that \( \lim_{t \to \infty} \frac{d^2 e(t)}{dt^2} = 0 \).

In general, for a MRAC system with an arbitrary relative degree \( n^* > 0 \), from (46), we express the \( i \)th-order time derivative \( \frac{d^i e(t)}{dt^i} \) of \( e(t) \) as

\[
\frac{d^i e(t)}{dt^i} = H(s) \frac{k_p s^{i+1}}{P_m(s)} [\ddot{\theta}^T \omega](t) + s^i K(s)[e](t). \quad (53)
\]

Since \( \ddot{\theta}^T(t)\omega(t) \) is bounded, and for each \( i = 1, 2, \ldots, n^* - 1 \), \( \frac{k_p s^{i+1}}{P_m(s)} \) is stable and strictly proper (proper for \( i = n^* - 1 \)), we have

\[
|H(s) \frac{k_p s^{i+1}}{P_m(s)} [\ddot{\theta}^T \omega](t)| \leq \frac{c_i}{a} \quad (54)
\]
Hence, with the use of the fictitious parameter $a > 0$ in $H(s)$ and $K(s)$, similar to (50), it can be shown that for every $\eta > 0$, there exists a $T = T(\eta) > 0$ such that $|\frac{d^i e(t)}{dt^i}| < \eta$ for all $t > T$, so that $\lim_{t \to \infty} \frac{d^i e(t)}{dt^i} = 0$, for $i = 1, 2, \ldots, n^* - 1$.

For a general MRAC system, the reference input signal $r(t)$ is only required to be bounded (which is sufficient for the result of Theorem 1). If the time-derivative of $r(t)$ is also bounded, we have the following additional property.

**Corollary 1**  For a MRAC system with relative degree $n^* > 0$, if both $r(t)$ and $\dot{r}(t)$ are bounded, then the tracking error $e(t) = y(t) - y_m(t)$ has the convergence property: $\lim_{t \to \infty} \frac{d^i e(t)}{dt^i} = 0$, for $i = 0, 1, \ldots, n^*$.

**Proof:** We just need to show the additional convergence property: $\lim_{t \to \infty} \frac{d^{n^*} e(t)}{dt^{n^*}} = 0$. To this end, we consider (53) with $i = n^*$:

$$\frac{d^{n^*} e(t)}{dt^{n^*}} = H(s) \frac{k_p s^{n^*}}{P_m(s)} s[\hat{\theta}^T \omega](t) + s^{n^*} K(s) [e](t).$$

In this case, $s[\hat{\theta}^T \omega](t) = \frac{d}{dt} (\hat{\theta}^T (t) \omega(t)) = \hat{\theta}^T (t) \omega(t) + \tilde{\theta}^T \dot{\omega}(t)$ is bounded, because $\dot{\theta}(t)$ and $\dot{\omega}(t)$ (whose last component is $\dot{r}(t)$ which is bounded by assumption) are bounded. Hence $k_p s^{n^*} \frac{P_m(s)}{s^{n^*}} s[\hat{\theta}^T \omega](t)$ is bounded. Since $s^{n^*} K(s)$ in (56) is stable and proper and $H(s)$ satisfies (44), we have, with the property: $\lim_{t \to \infty} e(t) = 0$ and an inequality similar to (50) with $\frac{d^n e(t)}{dt^n}$ replacing $\dot{e}(t)$, that $\lim_{t \to \infty} \frac{d^{n^*} e(t)}{dt^{n^*}} = 0$.

**Remark 1**  As a comparison, in the ideal case of nominal model reference control when $\theta(t) = \theta^*$ in the controller structure, it can be shown that

$$P_m(s)[y - y_m](t) = -k_p \varepsilon_1(t)$$

for some exponentially decaying and initial condition related term $\varepsilon_1(t)$, where $P_m(s)$ is a stable polynomial of degree $n^*$. It follows from this equation that the tracking error $e(t) = y(t) - y_m(t)$ has the convergence property: $\lim_{t \to \infty} \frac{d^i e(t)}{dt^i} = 0$, for $i = 0, 1, \ldots, n^*$ (under the condition that $r(t)$ is bounded). The difference between the nominal control and adaptive control cases is that in the adaptive control case the additional condition that $\dot{r}(t)$ is bounded was used to show $\lim_{t \to \infty} \frac{d^{n^*} e(t)}{dt^{n^*}} = 0$, which is not needed for the nominal control case.
Remark 2 For the case when \( y_m(t) = 0 \), it follows that \( \lim_{t \to \infty} d^i y(t) = 0 \), for \( i = 0, 1, \ldots, n^* - 1 \), an output regulation result of MRAC.

For a state-space system model: \( \dot{x} = Ax + Bu, y = Cx \) with \( x(t) \in \mathbb{R}^n \), if \( n^* = n \), then \( \lim_{t \to \infty} d^i y(t) = 0 \), for \( i = 0, 1, \ldots, n - 1 \), as established above, and \( \lim_{t \to \infty} x(t) = 0 \), as \( (A, C) \) is observable implied by the condition that \( n^* = n \) (the system transfer function does not have any finite zeros as its numerator is just a constant \( k_p \)). This is the adaptive asymptotic state regulation result, guaranteed by a MRAC design.

In particular, for an \( n \)th order system with \( (A, B) \) in the controllable canonical form and \( y(t) = x_1(t) \), the system transfer function explicitly has no finite zeros as its numerator is a constant, leading to \( n^* = n \) and an observable \( (A, C) \), so that \( \lim_{t \to \infty} x(t) = 0 \), an inherent property of a MRAC system. \( \square \)

This new tracking error convergence property: \( \lim_{t \to \infty} d^i e(t) = 0 \), for \( i = 1, \ldots, n^* \), of adaptive control, is extendable to adaptive nonlinear tracking control systems in which the controlled nonlinear plant has a defined relative degree \( n^* \).
Analysis of “Bursting Phenomenon” in Robust Adaptive Control

As illustrated in [134] that adaptive control systems with a fixed $\sigma$-modification may show some bursting phenomenon. Consider the adaptive system analyzed on page 232 (for Theorem 5.8) with $\sigma_1(t) = \sigma_0$ being a constant in (5.198):

$$\dot{\theta}(t) = -\text{sign}[k_p] \Gamma \omega(t)e(t) - \Gamma \sigma_1(t)\theta(t), \ t \geq 0$$

(that is, with a fixed $\sigma$-modification). For $V$ given in (5.43): $V = e^2 + |k_p|\tilde{\theta}^T \Gamma^{-1} \tilde{\theta}$, we have from (5.204) that

$$\dot{V} \leq -a_m e^2(t) + \frac{d^2(t)}{a_m} - 2|k_p| \sigma_0 \tilde{\theta}^T(t) \theta(t)$$

$$= -a_m e^2(t) + \frac{d^2(t)}{a_m} - 2|k_p| \sigma_0 \tilde{\theta}^T(t) \tilde{\theta}(t) + 2|k_p| \sigma_0 \tilde{\theta}^T(t) \theta^*$$

$$\leq -a_m e^2(t) + \frac{d^2(t)}{a_m} - |k_p| \sigma_0 \tilde{\theta}^T(t) \tilde{\theta}(t) + |k_p| \sigma_0 \theta^* \theta^*.$$  (58)

For $\Gamma = \gamma I$ and $a_m > \gamma \sigma_0$ (that is, with a small $\sigma_0$), we have

$$\dot{V} \leq -\gamma \sigma_0 V + \frac{d^2(t)}{a_m} + |k_p| \sigma_0 \theta^* \theta^*$$  (59)

which, for $|d(t)| \leq d_0$, leads to

$$\lim_{t \to \infty} V(t) \leq \frac{d_0^2}{a_m \gamma \sigma_0} + \frac{|k_p| \theta^* \theta^*}{\gamma}.$$  (60)

This implies that the upper bound for the tracking error $e(t) = y(t) - y_m(t)$, for $d(t) = d_0 = 0$ (in the absence of disturbances $d(t)$), may be as large as $\frac{|k_p| \theta^* \theta^*}{\gamma}$, independent of $\sigma_0$. We already knew that for $d(t) = 0$ and $\sigma_1(t) = \sigma_0 = 0$, the adaptive control system ensures that $\lim_{t \to \infty} e(t) = 0$. Now from the above analysis we know that for a small $\sigma_0 \neq 0$, the error bound on $|e(t)|$ can be up to $\frac{|k_p| \theta^* \theta^*}{\gamma}$. On the other hand, from

$$\dot{V} \leq -a_m e^2(t) + \frac{d^2(t)}{a_m} - 2|k_p| \sigma_0 \tilde{\theta}^T(t) \theta(t)$$  (61)

and the adaptive control system signal boundedness, similar to (5.204), we have

$$\int_{t_1}^{t_2} e^2(t) \, dt \leq \gamma_0 + k_0(t_2 - t_1)d_0^2 + c_0(t_2 - t_1)\sigma_0.$$  (62)
for some constant \( c_0 > 0 \), where the \( \sigma_0 \) related term is due to using a fixed \( \sigma \)-modification \( \sigma_1(t) = \sigma_0 \), instead of the switching \( \sigma \)-modification which leads to (5.205) and in turn to (5.204). The inequality (62) implies that, when \( d_0 = 0 \) (in the absence of disturbances), we have the mean error

\[
\frac{1}{t_2 - t_1} \int_{t_1}^{t_2} e^2(t) \, dt \leq \frac{\gamma_0}{t_2 - t_1} + c_0 \sigma_0
\]

which is of the magnitude of \( c_0 \sigma_0 \), but the absolute error \( |e(t)| \), as from (60), could be as large as \( \frac{|k_p|}{T} \theta^T \theta^* \), independent of \( \sigma_0 \). This is the so-called “bursting phenomenon” of robust adaptive control with a fixed \( \sigma \)-modification: the tracking error \( e(t) \) may go to a large value independent of \( \sigma_0 \) for a small interval of time but in the mean sense the error \( e(t) \) is of the order of \( \sigma_0 \). This analytically explains what was observed in the simulation results [134].
Robust MRAC with A Switching σ-Modification

To derive (5.224), for \( V(\tilde{\theta}, \tilde{\rho}) = |\rho^*|\tilde{\theta}^T \Gamma^{-1} \tilde{\theta} + \gamma^{-1} \tilde{\rho}^2 \), using
\[
\epsilon(t) = \rho^* \tilde{\theta}^T(t) \xi(t) + \tilde{\rho}(t) \xi(t) + \mu \eta(t), \quad \eta(t) = \Delta(s)[u](t)
\] (64)
we have
\[
\dot{V} = -2 \frac{\epsilon^2(t)}{m^2(t)} + 2 \frac{\epsilon(t) \eta(t)}{m^2(t)} - 2 \sigma_1(t) |\rho^*| \tilde{\theta}^T(t) \theta(t) - 2 \sigma_2(t) \tilde{\rho}(t) \rho(t)
\]
\[
= -2 \frac{\epsilon^2(t)}{m^2(t)} - \left( \frac{\epsilon(t)}{m(t)} - \frac{\mu \eta(t)}{m(t)} \right)^2 + \mu^2 \frac{\eta^2(t)}{m^2(t)}
\]
\[
- 2 \sigma_1(t) |\rho^*| \tilde{\theta}^T(t) \theta(t) - 2 \sigma_2(t) \tilde{\rho}(t) \rho(t)
\]
\[
\leq -2 \frac{\epsilon^2(t)}{m^2(t)} + \mu^2 \frac{\eta^2(t)}{m^2(t)} - 2 \sigma_1(t) |\rho^*| \tilde{\theta}^T(t) \theta(t) - 2 \sigma_2(t) \tilde{\rho}(t) \rho(t)
\] (65)
which is (5.224). Since \( \left[ \frac{\eta(t)}{m(t)} \right] \leq b_0 \) for some constant \( b_0 > 0 \), we have \( \dot{V} < 0 \) if
\[
2 \sigma_1(t) |\rho^*| \tilde{\theta}^T(t) \theta(t) + 2 \sigma_2(t) \tilde{\rho}(t) \rho(t) > \mu^2 b_0^2.
\] (66)

For \( \tilde{\theta} = \theta - \theta^* \) and \( \tilde{\rho} = \rho - \rho^* \), by definition, we have
\[
\sigma_1(t) |\rho^*| \tilde{\theta}^T(t) \theta(t) \geq 0, \quad \sigma_2(t) \tilde{\rho}(t) \rho(t) \geq 0, \quad t \geq 0
\] (67)
\[
\lim_{\|\theta\|_2 \to \infty} \tilde{\theta}^T \theta = \infty, \quad \lim_{|\rho| \to \infty} \tilde{\rho} \rho = \infty.
\] (68)
This implies that there exist constants \( \theta^0 > 0 \) and \( \rho^0 > 0 \) such that \( \|\theta(t)\|_2 \geq \theta^0 \) or/and \( |\rho(t)| \geq \rho^0 \) implies that \( \dot{V} < 0 \). Hence, the boundedness of \( \theta(t) \) and \( \rho(t) \) is ensured. One choice of such \( (\theta^0, \rho^0) \) is
\[
\theta^0 = \max \{ 2M_1, \sqrt{ \frac{b_0^2 + \|\theta^*\|_2^2 + \rho^*^2}{4\sigma_{10}} } + \frac{1}{2} \|\theta^*\| \} \}
\[
\rho^0 = \max \{ 2M_2, \sqrt{ \frac{b_0^2 + \|\theta^*\|_2^2 + \rho^*^2}{4\sigma_{20}} } + \frac{1}{2} |\rho^*| \}.
\] (69)
Discrete-time MRAC System Example

Consider the first-order plant

\[ y(t + 1) = a_p y(t) + b_p u(t) \] (70)

with two unknown parameters \( a_p \) and \( b_p \) and choose the reference model

\[ y_m(t + 1) = -a_m y_m(t) + b_m r(t) \] (71)

with \( |a_m| < 1 \) for stability. We use the adaptive controller structure

\[ u(t) = k_1(t) y(t) + k_2(t) r(t), \] (72)

where \( k_1(t) \) and \( k_2(t) \) are estimates of the unknown parameters

\[ k_1^* = \frac{-a_p - a_m}{b_p}, \quad k_2^* = \frac{b_m}{b_p}. \] (73)

In this case the tracking error equation becomes

\[ e(t + 1) = -a_m e(t) + b_p \tilde{k}_1(t) y(t) + b_p \tilde{k}_2(t) r(t), \] (74)

where \( \tilde{k}_1(t) = k_1(t) - k_1^* \) and \( \tilde{k}_2(t) = k_2(t) - k_2^* \).

Defining \( \rho^* = b_p \) and

\[ \theta^* = [k_1^*, k_2^*]^T, \quad \theta = [k_1, k_2]^T, \] (75)

\[ \omega(t) = [y(t), r(t)]^T \] (76)

and introducing the filtered vector signal

\[ \zeta(t) = \frac{1}{z + a_m} [\omega](t) = \left[ \frac{1}{z + a_m} [y](t), \frac{1}{z + a_m} [r](t) \right]^T = [\zeta_1(t), \zeta_2(t)]^T, \] (77)

where, as a notation, \( \zeta_1(t) = \frac{1}{z + a_m} [y](t) \) denotes the output of the system with transfer function \( \frac{1}{z + a_m} \) and input \( y(t) \) (it satisfies the equation: \( \zeta_1(t + 1) = -a_m \zeta_1(t) + y(t) \), for generating \( \zeta_1(t) \) from \( y(t) \)), we rewrite (74) as

\[ e(t) = \rho^* \frac{1}{z + a_m} [\theta^T \omega - \theta^{*T} \omega](t) \]

\[ = \rho^* \left( \frac{1}{z + a_m} [\theta^T \omega](t) - \theta^{*T} \zeta(t) \right). \] (78)
The design task is to find adaptive laws to update the parameter estimates \( \theta(t) \) and \( \rho(t) \) (which is an estimate of \( \rho^* \)) such that the estimation error

\[
\varepsilon(t) = e(t) - \rho(t) \left( \frac{1}{z + a_m} [\theta^T \omega](t) - \theta^T(t) \zeta(t) \right)
\] (79)

is small in some sense. With (78), this error can be expressed as

\[
\varepsilon(t) = \rho^* \tilde{\theta}^T(t) \zeta(t) + \tilde{\rho}(t) \xi(t),
\] (80)

where

\[
\xi(t) = \theta^T(t) \zeta(t) - \frac{1}{z + a_m} [\theta^T \omega](t)
\] (81)

\[
\tilde{\theta}(t) = \theta(t) - \theta^*, \quad \tilde{\rho}(t) = \rho(t) - \rho^*.
\] (82)

We choose the gradient adaptive laws for \( \theta(t) \) and \( \rho(t) \):

\[
\theta(t+1) = \theta(t) - \text{sign}[b_p] \Gamma \varepsilon(t) \zeta(t) m_2(t), \quad 0 < \Gamma = \Gamma^T < \frac{2}{b_p^0} I_2,
\] (83)

\[
\rho(t+1) = \rho(t) - \gamma \varepsilon(t) \xi(t) m_2(t), \quad 0 < \gamma < 2,
\] (84)

where \( \text{sign}[b_p] \) is the sign of \( b_p \), \( b_p^0 \geq |b_p| \) is a known upper bound on \( |b_p| \), and

\[
m(t) = \sqrt{1 + \zeta^T(t) \zeta(t) + \xi^2(t)}.
\] (85)

The stability analysis of this MRAC system is given in Section 6.3.1.
Indirect Observer Based Adaptive Control

Consider a continuous-time linear multivariable time-invariant plant

\[ y(t) = G(s)[u](t), \quad u(t) \in \mathbb{R}^M, \quad y(t) \in \mathbb{R}^M, \quad M \geq 1, \tag{86} \]

where \( G(s) \) is an \( M \times M \) strictly proper rational transfer matrix which can be expressed as

\[ G(s) = D^{-1}(s)N(s) = C_g(sI - A_g)^{-1}B_g, \tag{87} \]

where, for the observability index \( \nu \) of \( G(s) \) (its minimal realization),

\[ D(s) = s^\nu I_M + A_{\nu - 1}s^{\nu - 1} + \cdots + A_1s + A_0, \tag{88} \]

\[ N(s) = B_ms^\nu + \cdots + B_1s + B_0, \tag{89} \]

\[ A_g = \begin{bmatrix} -A_{\nu - 1} & I_M & 0 & \cdots & \cdots & 0 \\ -A_{\nu - 2} & 0 & I_M & \cdots & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ -A_1 & 0 & \cdots & \cdots & 0 & I_M \\ -A_0 & 0 & \cdots & \cdots & 0 & 0 \end{bmatrix} \in \mathbb{R}^{M\nu \times M\nu}, \tag{90} \]

\[ B_g = \begin{bmatrix} 0 \\ \vdots \\ 0 \\ B_p \end{bmatrix} \in \mathbb{R}^{M\nu \times M}, \quad B_p = \begin{bmatrix} B_m \\ \vdots \\ B_1 \\ B_0 \end{bmatrix} \in \mathbb{R}^{M(m + 1) \times M}, \tag{91} \]

\[ C_g = \begin{bmatrix} I_M & 0 & \cdots & 0 & 0 \end{bmatrix} \in \mathbb{R}^{M \times M\nu} \tag{92} \]

for the \( M \times M \) identity matrix \( I_M \) and some \( M \times M \) parameter matrices \( A_i, \quad i = 0, 1, \ldots, \nu - 1 \), and \( B_j, \quad j = 0, 1, \ldots, m, \quad m \leq \nu - 1 \).

In view of (87), the plant (86) can be expressed as

\[ \dot{x}(t) = Ax(t) + A_p y(t) + \begin{bmatrix} 0 \\ B_p \end{bmatrix} u(t), \quad y(t) = C_g x(t), \tag{93} \]

where \( A \in \mathbb{R}^{M\nu \times M\nu} \) and \( A_p \in \mathbb{R}^{M\nu \times M} \) are

\[ A = \begin{bmatrix} 0 & I_M & 0 & \cdots & \cdots & 0 \\ 0 & 0 & I_M & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \cdots & \cdots & 0 & I_M \\ 0 & 0 & \cdots & \cdots & 0 & 0 \end{bmatrix}, \quad A_p = \begin{bmatrix} -A_{\nu - 1} \\ -A_{\nu - 2} \\ \vdots \\ -A_1 \\ -A_0 \end{bmatrix}. \tag{94} \]
**Control objective.** The control objective is to design an adaptive feedback control signal \( u(t) \) to make the plant output \( y(t) \) tracks the output \( y_m(t) \) of the reference model system

\[
y_m(t) = W_m(s) [r](t), \quad W_m(s) = \xi_m^{-1}(s),
\]

where \( \xi_m(s) \) is the modified left interactor matrix of \( G(s) \): \( \lim_{s \to \infty} \xi_m(s) G(s) = K_p \) is finite and non-singular.

**Adaptive parameter estimation.** Based on the input-output model of the plant: \( D(s)[y](t) = N(s)[u](t) \), we can design an adaptive parameter estimation scheme to generate adaptive estimates \( \hat{A}_i \) and \( \hat{B}_j \) of the coefficient matrices \( A_i \) and \( B_j \) of \( D(s) \) and \( N(s) \). From the parameter estimates \( \hat{A}_i \) and \( \hat{B}_j \), we can obtain the adaptive estimates \( \hat{A}_p \) and \( \hat{B}_p \) of the parameter matrices \( A_p \) and \( B_p \), and the adaptive estimates \( \hat{A}_g \) and \( \hat{B}_g \) of the parameter matrices \( A_g \) and \( B_g \).

**Adaptive state observer.** With adaptive estimates \( \hat{A}_p \) and \( \hat{B}_p \), we construct an adaptive state observer for the plant (93) as

\[
\dot{x}(t) = \hat{A} \hat{x}(t) + \hat{A}_p y(t) + \begin{bmatrix} 0 \\ \hat{B}_p(t) \end{bmatrix} u(t) + L(y(t) - \hat{y}(t)), \quad \hat{y}(t) = C_g \hat{x}(t),
\]

where \( L \in R^{M_y \times M} \) is such that \( A - LC_g \) is a desired stable matrix. For the state estimation error \( \hat{x}(t) = x(t) - \hat{x}(t) \), it follows that

\[
\dot{\hat{x}}(t) = (A - LC_g) \hat{x}(t) + \hat{A}_p(t) y(t) + \begin{bmatrix} 0 \\ \hat{B}_p(t) \end{bmatrix} u(t), \quad \hat{y}(t) = C_g \hat{x}(t)
\]

where \( \hat{y}(t) = \hat{y}(t) - y(t), \hat{A}_p(t) = \hat{A}_p(t) - A_p \) and \( \hat{B}_p(t) = \hat{B}_p(t) - B_p \).

**Adaptive control law.** The adaptive control law is an indirect-design (that is, its parameters are calculated from plant parameter estimates) and explicit-observer (that is, it uses the state estimate \( \hat{x}(t) \) for feedback control) based control law:

\[
u(t) = K_1^T(t) \hat{x}(t) + K_2(t) r(t),
\]

where \( K_1(t) \in R^{M_y \times M} \) and \( K_2(t) \in R^{M \times M} \) are parameter matrices which satisfy

\[
C_g (\lambda I - \hat{A}_g - \hat{B}_g K_1^T) \hat{B}_g K_2 = W_m(\lambda),
\]

point-wise in the time variable \( t \), where \( \hat{A}_g(t) \) and \( \hat{B}_g(t) \) are the on-line estimates of \( A_g \) and \( B_g \). It can be shown that such solutions \( K_1 \) and \( K_2 \) exist if the estimated plant \( \langle \hat{A}_g, \hat{B}_g, C_g \rangle \) has \( \xi_m(s) \) as its modified left interactor matrix point-wise, that is, \( \lim_{s \to \infty} \xi_m(s) C_g (\lambda I - \hat{A}_g)^{-1} \hat{B}_g = K_p \) is finite and non-singular, for each time \( t \) as \( \hat{A}_g = \hat{A}_p(t) \) and \( \hat{B}_g = \hat{B}_p(t) \) (note that \( K_p \) may be a function of time \( t \)).
In the single-input single-output case with $M = 1$, such a condition becomes: $\hat{B}_m(t) = \hat{b}_m(t) \neq 0$ for any $t$, which can be easily ensured by using a parameter projection based adaptive parameter estimator. In this case, $\xi_m(\lambda) = \lambda^{\nu-m}$ and $\hat{K}_p = \hat{b}_m(t)$, for $\nu = n$ being the plant order.
Adaptive Control of $\dot{x}(t) = Ax(t) + B\Lambda u(t) + B\Theta^* \phi(x)$

Consider an $M$-input linear time-invariant plant

$$\dot{x}(t) = Ax(t) + B\Lambda u(t) + B\Theta^* \phi(x), \quad x(t) \in \mathbb{R}^n, \quad u(t) \in \mathbb{R}^M,$$

(100)

where $(A, B)$ is a known and controllable pair, $\Lambda = \text{diag}\{\lambda_1, \lambda_2, \ldots, \lambda_M\}$ is an unknown and diagonal matrix with $\lambda_i \neq 0$ for $i = 1, 2, \ldots, M$, $\Theta^*$ is an unknown matrix, and $\phi(x)$ is a known vector.

With $(A, B)$ known and controllable, we can find known constant matrices $K_{10} \in \mathbb{R}^{n \times M}$ and $K_{20} \in \mathbb{R}^{M \times M}$ such that

$$A_m = A + BK_{10}^T, \quad B_m = BK_{20}$$

(101)

are known with $A_m$ being stable, for constructing a good reference model system

$$\dot{x}_m(t) = A_m x_m(t) + B_m r(t), \quad x_m(t) \in \mathbb{R}^n, \quad r(t) \in \mathbb{R}^M,$$

(102)

where $r(t)$ is a bounded and piecewise continuous reference input.

We first introduce the parameter matrices

$$K_1^* = \Lambda^{-1} K_{10}^T, \quad K_2^* = \Lambda^{-1} K_{20}, \quad K_3^* = -\Lambda^{-1}\Theta^*.$$  

(103)

Then, for $K_1, K_2$ and $K_3$ being the estimates of $K_1^*, K_2^*$ and $K_3^*$, we use the adaptive controller

$$u(t) = K_1^T(t)x(t) + K_2(t)r(t) + K_3(t)\phi(x).$$

(104)

From the definitions of $K_1^*, K_2^*$ and $K_3^*$, we have

$$BA(K_1^*T x(t) + K_2^* r(t) + K_3^* \phi(x)) = B(K_{10}^T x(t) + K_{20} r(t) - \Theta^* \phi(x)).$$

(105)

Introducing the parameter errors

$$\tilde{K}_1 = K_1 - K_1^*, \quad \tilde{K}_2 = K_2 - K_2^*, \quad \tilde{K}_3 = K_3 - K_3^*$$

(106)

we express the control signal from (104) as

$$u(t) = \tilde{K}_1^T(t)x(t) + \tilde{K}_2(t)r(t) + \tilde{K}_3(t)\phi(x) + K_1^*T x(t) + K_2^* r(t) + K_3^* \phi(x)$$

(107)

and, in view of (101) and (105), the closed-loop system as

$$\dot{x}(t) = A_m x(t) + B\Lambda(\tilde{K}_1^T(t)x(t) + \tilde{K}_2(t)r(t) + \tilde{K}_3(t)\phi(x))$$

$$+ B\Lambda(K_1^*T x(t) + K_2^* r(t) + K_3^* \phi(x)) + B\Theta^* \phi(x)$$

$$= A_m x(t) + B_m r(t) + B\Lambda(\tilde{K}_1^T(t)x(t) + \tilde{K}_2(t)r(t) + \tilde{K}_3(t)\phi(x)).$$

(108)
Then, in view of this equation and (102), the tracking error $e(t) = x(t) - x_m(t)$ satisfies

$$\dot{e}(t) = A_m e(t) + B\lambda (\tilde{K}_1(t)x(t) + \tilde{K}_2(t)r(t) + \tilde{K}_3(t)\phi(x)).$$

(109)

Introducing $\theta_i^*$ such that $\theta_i^{*T}$ is the $i$th row of $[K_1^T, K_2^T, K_3^T]$, $i = 1, 2, \ldots, M$, letting $\hat{\theta}_i(t) = \theta_i(t) - \theta_i^*$, $i = 1, 2, \ldots, M$, and defining

$$\omega(t) = [x^T(t), r^T(t), \phi^T(x)]^T,$$

we express (109) as

$$\dot{e}(t) = A_m e(t) + B\Lambda \begin{bmatrix} \tilde{\theta}_1^T(t) \omega(t) \\ \tilde{\theta}_2^T(t) \omega(t) \\ \vdots \\ \tilde{\theta}_{M-1}^T(t) \omega(t) \\ \tilde{\theta}_M^T(t) \omega(t) \end{bmatrix}.$$  

(111)

Letting $P = P^T > 0$ satisfy

$$PA_m + A_m^T P = -Q$$

(112)

for a chosen constant matrix $Q \in \mathbb{R}^{n \times n}$ such that $Q = Q^T > 0$, and $\tilde{e}_i(t)$ be the $i$th component of $e^T(t)PB$, $i = 1, 2, \ldots, M$, we design the adaptive law for $\theta_i(t)$ as

$$\dot{\hat{\theta}}_i(t) = -\text{sign}[\lambda_i] \Gamma_i \tilde{e}_i(t) \omega(t),$$

(113)

where $\Gamma_i = \Gamma_i^T > 0$ is a chosen constant adaptation gain matrix, and $\text{sign}[\lambda_i]$ is the sign of $\lambda_i$, $i = 1, 2, \ldots, M$.

To analyze the adaptive control system performance, we consider the positive definite function

$$V(e, \hat{\theta}, i = 1, 2, \ldots, M) = e^T P e + \sum_{i=1}^M|\lambda_i| \tilde{\theta}_i^T \Gamma_i^{-1} \tilde{\theta}_i$$

(114)

and derive its time derivative along the trajectory of (111) and (113) as

$$\dot{V} = 2e^T(t)P \dot{e}(t) + 2\sum_{i=1}^M|\lambda_i| \tilde{\theta}_i^T(t) \Gamma_i^{-1} \dot{\hat{\theta}}_i(t)$$

$$= 2e^T(t)PA_m e(t) + 2\sum_{i=1}^M \tilde{e}_i(t) \lambda_i \tilde{\theta}_i^T(t) \omega(t) + 2\sum_{i=1}^M |\lambda_i| \tilde{\theta}_i^T(t) \Gamma_i^{-1} \hat{\theta}_i(t)$$

$$= -e^T(t) Q e(t).$$

(115)

From this result, it follows that $x(t)$ and $\theta_i(t)$ are bounded and $e(t) \in L^2$, and in turn, that $u(t)$ is bounded, and so is $\dot{e}(t)$, so that $\lim_{t \to \infty} e(t) = 0$. 
Parametrization of (9.102)

To define the parameters Θ_i^+, i = 1, 2, 20, 3, to satisfy (9.105):

\[ \Theta_1^{+T} A(D) P_0(D) + (\Theta_2^{+T} A(D) + \Theta_20^{+} \Lambda(D)) Z_0(D) = \Lambda(D) (P_0(D) - \Theta_3^{+} \xi_m(D) Z_0(D)), \]

for the nominal version of the multivariable MRAC controller (9.102), we consider an equivalent version (see (117) below) of (9.105) by dividing it from the right by \( P_0^{-1}(D) \) and use the left matrix-fraction description of \( G_0(D) = P_0^{-1}(D) Z_{\ell}(D) \) such that \( G_0(D) = P_0^{-1}(D) Z_{\ell}(D) = Z_0(D) P_0^{-1}(D) \) (as used in the proof of Lemma 9.3), to obtain

\[ \Theta_1^{+T} A(D) + (\Theta_2^{+T} A(D) + \Theta_20^{+} \Lambda(D)) P_0^{-1}(D) Z_{\ell}(D) = \Lambda(D) (1 - \Theta_3^{+} \xi_m(D) P_0^{-1}(D) Z_{\ell}(D)). \]

(117)

Expressing \( \Lambda(D) \Theta_3^{+} \xi_m(D) P_0^{-1}(D) \) with \( \Theta_3^{+} = K_p^{-1} \) as

\[ \Lambda(D) \Theta_3^{+} \xi_m(D) P_0^{-1}(D) = Q_{\ell}(D) + R_{\ell}(D) P_0^{-1}(D) \]

(118)

for some \( M \times M \) polynomial matrices \( Q_{\ell}(D) \) and \( R_{\ell}(D) \) such that \( \partial_{ci}[R_{\ell}(D)] \leq v \). Then, we define \( \Theta_1^{+}, \Theta_2^{+} \) and \( \Theta_20^{+} \) from

\[ \Theta_2^{+T} A(D) + \Theta_20^{+} \Lambda(D) = -R_{\ell}(D), \]

(119)

\[ \Theta_1^{+T} A(D) = \Lambda(D) I_M - Q_{\ell}(D) Z_{\ell}(D) \]

(120)

(recall that \( \partial[\Lambda(D)] = v - 1 \) and \( \partial[A(D)] = v - 2 \) for the controller structure (9.102)). With this definition of \( \Theta_2^{+}, \Theta_20^{+} \) and \( \Theta_1^{+} \), (117) is satisfied, and so is (9.105).

From (9.105) (that is, (117)), we have the plant–model matching equation

\[ I_M - \Theta_1^{+T} F(D) - (\Theta_2^{+T} F(D) + \Theta_20^{+}) G_0(D) = \Theta_3^{+} W_m^{-1}(D) G_0(D) \]

(121)

where \( W_m(D) = \xi_m^{-1}(D) \). From this equation with \( \lim_{D \to \infty} \Theta_3^{+} W_m^{-1}(D) G_0(D) = I_M \) as from the definition of \( \xi_m(D) \), we have

\[ \lim_{D \to \infty} \Theta_1^{+T} F(D) = 0, \]

(122)

which implies that \( \partial[\Theta_1^{+T} A(D)] \leq v - 2 \), that is, (120) is solvable.
Plant Signal Identities for MIMO Cases

It has been shown that there exist $\Theta_1^*, \Theta_2^*$ and $\Theta_3^*$ to satisfy the matching equation

$$\Theta_1^T A(D) P_0(D) + \Theta_2^T A(D) Z_0(D) = \Lambda(D) (P_0(D) - \Theta_3^* \xi_m(D) Z_0(D)), \quad (123)$$

which is nominally (mathematically) equivalent to

$$I_M - \Theta_1^T F(D) - \Theta_2^T F(D) G_0(D) = \Theta_3^* W_m^{-1}(D) G_0(D). \quad (124)$$

Operating this identity on $u(t)$, we may get the plant signal identity:

$$u(t) - \Theta_1^T \omega_1(t) - \Theta_2^T \omega_2(t) = \Theta_3^* W_m^{-1}(D)[y](t). \quad (125)$$

(This identity was used in deriving (9.119), and its similar version with $\Theta_2^*$ was used in deriving (9.365).)

As an alternative procedure to obtain the plant parametrized signal identity (similar to that in (5.30) for the SISO case with $M = 1$), we start by considering (9.84) and (9.85):

$$\Theta_2^T A(D) = Q_l(D) P_1(D) - \Lambda(D) K_p^{-1} \xi_m(D) \quad (126)$$

$$\Theta_1^T A(D) = \Lambda(D) I_M - Q_l(D) Z_l(D). \quad (127)$$

and obtain the signal identities:

$$\Theta_2^T A(D) \frac{\Lambda(D)}{\Lambda(D)} [y](t) = \frac{1}{\Lambda(D)} Q_l(D) P_1(D) [y](t) - K_p^{-1} \xi_m(D) [y](t) \quad (128)$$

$$\Theta_1^T A(D) \frac{\Lambda(D)}{\Lambda(D)} [u](t) = u(t) - \frac{1}{\Lambda(D)} Q_l(D) Z_l(D) [u](t). \quad (129)$$

Using the open-loop plant signal identity: $P_l(D)[y](t) = Z_l(D)[u](t)$, and from (128)–(129), we finally have the parametrized plant signal identity (125):

$$u(t) = \Theta_1^T A(D) \frac{\Lambda(D)}{\Lambda(D)} [u](t) + \Theta_2^T A(D) \frac{\Lambda(D)}{\Lambda(D)} [y](t) + K_p^{-1} \xi_m(D) [y](t). \quad (130)$$

This identity holds for any input signal $u(t)$, in a feedback structure.

Similarly, for the controller structure (9.102) with nominal parameters:

$$u(t) = \Theta_1^T \omega_1(t) + \Theta_2^T \omega_2(t) + \Theta_2^* v(t) + \Theta_3^* r(t), \quad (131)$$

in which $\hat{\Lambda}[\Lambda(D)] = \nu - 1$, we have the polynomial matching equation

$$\Theta_1^T A(D) P_0(D) + (\Theta_2^T A(D) + \Theta_2^* \Lambda(D)) Z_0(D) = \Lambda(D) (P_0(D) - \Theta_3^* \xi_m(D) Z_0(D)) \quad (132)$$
and the transfer matrix matching equation

\[ I_M - \Theta_1^T F(D) - (\Theta_2^T F(D) + \Theta_{20}^*) G_0(D) = \Theta_3^T W_m^{-1}(D) G_0(D). \]  

(133)

The nominal parameters \( \Theta_1^*, \Theta_2^*, \Theta_{20}^* \) and \( \Theta_3^* \) are defined from

\[(\Theta_2^T A(D) + \Theta_{20}^* \Lambda(D)) = -R_l(D) = Q_l(D) P_l(D) - \Lambda(D) K_p^{-1} \xi_m(D) \]  

(134)

and

\[ \Theta_1^T A(D) = \Lambda(D) I_M - Q_l(D) Z_l(D), \]  

(135)

that is, dividing \( \Lambda(D) K_p^{-1} \xi_m(D) \) on the right by \( P_l(D) \) to get \( R_l(D) \) and \( Q_l(D) \). From these equations, we obtain the signal identities:

\[(\Theta_2^T A(D)/\Lambda(D) + \Theta_{20}^*) [y](t) = \frac{1}{\Lambda(D)} Q_l(D) P_l(D)[y](t) - K_p^{-1} \xi_m(D)[y](t) \]  

(136)

\[ \Theta_1^T A(D)/\Lambda(D) [u](t) = u(t) - \frac{1}{\Lambda(D)} Q_l(D) Z_l(D)[u](t). \]  

(137)

Using the open-loop plant signal identity: \( P_l(D)[y](t) = Z_l(D)[u](t) \), we have the plant signal identity in a feedback form:

\[ u(t) = \Theta_1^T A(D)/\Lambda(D) [u](t) + \Theta_2^T A(D)/\Lambda(D) [y](t) + \Theta_{20}^* y(t) + K_p^{-1} \xi_m(D)[y](t) \]  

(138)

which also holds for any input signal \( u(t) \), similar to that in (130).

Both parametrized plant signal identities (130) and (138) are useful for adaptive control: either verify the nominal controller structure

\[ u(t) = \Theta_1^T A(D)/\Lambda(D) [u](t) + \Theta_2^T A(D)/\Lambda(D) [y](t) + \Theta_{20}^* y(t) + \Theta_3^* r(t) \]  

(139)

for the matching equation (123), or

\[ u(t) = \Theta_1^T A(D)/\Lambda(D) [u](t) + \Theta_2^T A(D)/\Lambda(D) [y](t) + \Theta_{20}^* y(t) + \Theta_3^* r(t) \]  

(140)

for the matching equation (132), where \( \Theta_3^* r(t) = K_p^{-1} \xi_m(D)[y_m](t) \) for the reference output \( y_m(t) = \xi_m^{-1}(D)[r](t) \), leading to \( \xi_m(D) [y - y_m](t) = 0 \) exponentially, or they motivate the adaptive controller structure

\[ u(t) = \Theta_1^T A(D)/\Lambda(D) [u](t) + \Theta_2^T A(D)/\Lambda(D) [y](t) + \Theta_3 r(t) \]  

(141)

for the matching equation (123), or

\[ u(t) = \Theta_1^T A(D)/\Lambda(D) [u](t) + \Theta_2^T A(D)/\Lambda(D) [y](t) + \Theta_{20} y(t) + \Theta_3 r(t) \]  

(142)
for the matching equation (132), leading to the desired tracking error equation:

\[ e(t) = y(t) - y_m(t) = \xi_m^{-1}(D)K_p[\tilde{\Theta}^T \omega](t) \]  

(143)

for \( \tilde{\Theta}(t) = \Theta(t) - \Theta^* \) with

\[ \Theta(t) = [\Theta_1^T(t), \Theta_2^T(t), \Theta_3(t)]^T, \quad \Theta^* = [\Theta_1^{*T}, \Theta_2^{*T}, \Theta_3^*]^T \]  

(144)

\[ \omega(t) = [\omega_1^T(t), \omega_2^T(t), r^T(t)]^T \]  

(145)

for the matching equation (123), or

\[ \Theta(t) = [\Theta_1^T(t), \Theta_2^T(t), \Theta_20(t), \Theta_3(t)]^T, \quad \Theta^* = [\Theta_1^{*T}, \Theta_2^{*T}, \Theta_20^*, \Theta_3^*]^T \]  

(146)

\[ \omega(t) = [\omega_1^T(t), \omega_2^T(t), y^T(t), r^T(t)]^T \]  

(147)

for the matching equation (132). Both tracking error equations can be used to derive some desired estimation errors for designing stable adaptive laws to update the controller parameters \( \Theta(t) \).
Proof of $\lim_{t \to \infty} e(t) = 0$ in (9.187C)

To show $\lim_{t \to \infty} e(t) = 0$, we just need to prove that for any given $\delta > 0$, there exists a $T = T(\delta) > 0$ such that $\|e(t)\| < \delta$, for all $t > T$. For such a proof, we only require the existence of $T$ for each chosen $\delta$ which may be set to be arbitrarily small but do not consider the limit of $\delta$ going to zero, that is, $\delta > 0$ such that $\frac{1}{\delta}$ is finite.

In view of (9.187C) (for the continuous-time case), for any given $\delta > 0$, we can choose (think of) a pair of $K_3(s)$ and $H_3(s)$ to make $\frac{c_4}{a_3} < \frac{\delta}{2}$ (that is, $a_3 > \frac{2c_4}{\delta}$ is finite) (recall that $K_3(s)$ and $H_3(s)$ are not real but ficticious filters, and they are not employed in the control design but only used in the expression (decomposition) of $e(t)$ in (9.185C)), and let $T > 0$ such that $|z_1(t)| < \frac{\delta}{2}$ for any $t > T$ (such a $T = T(\delta)$ exists, because $\lim_{t \to \infty} z_1(t) = 0$; also note that $z_1(t)$ depends on $K_3(s)$). It follows that $\|e(t)\| < \delta$ for any $t > T$, which implies that $\lim_{t \to \infty} e(t) = 0$. 
Adaptive Robot Control for Time-Varying Parameters

Consider the manipulator dynamic equation (9.586):

$$D(q,t)\ddot{q} + \frac{dD(q,t)}{dt}\dot{q} + C(q,\dot{q},t)\dot{q} + \phi(q,t) = u.$$  \hspace{1cm} (148)$$

With (9.590): v = \dot{q}_d - \Lambda_0(q - q_d), s = \dot{q} - v, e = q - q_d, we have (9.592):

$$D(q,t)s + C(q,\dot{q},t)s$$

$$= u - D(q,t)\dot{v} - C(q,\dot{q},t)v - \phi(q,t) - \frac{dD(q,t)}{dt}\dot{q}$$

$$\triangleq u - Y(q,\dot{q}_d,\dot{q},\dot{q}_d,\dot{q}_d,t)\theta^*(t) - \frac{dD(q,t)}{dt}\dot{q},$$  \hspace{1cm} (149)$$

for some known function matrix $Y(q,\dot{q}_d,\dot{q},\dot{q}_d,\dot{q}_d,t) \in \mathbb{R}^{n \times n_\theta}$ and unknown parameter vector $\theta^*(t) \in \mathbb{R}^{n_\theta}$ which may be time-varying.

In this study, we consider the case when

$$\theta^*(t) = \theta_0^* + \delta\theta^*(t)$$  \hspace{1cm} (150)$$

where $\theta_0^*$ is a constant vector and $\delta\theta^*(t)$ is the variation of $\theta^*(t)$ with respect to $\theta_0^*$ (note that both $\theta_0^*$ and $\delta\theta^*(t)$ are unknown). We will develop and analyze some alternative adaptive control schemes to that presented in (9.607)–(9.608).

Adaptive Control Scheme I

As an alternative scheme to (9.607)–(9.608), we use the control law

$$u(t) = Y(q,\dot{q}_d,\dot{q},\dot{q}_d,\dot{q}_d,t)\theta_0(t) - m(t)\psi(t) - m_1(t)\psi_1(t) - K_Ds(t),$$  \hspace{1cm} (151)$$

$$m(t) = k_0\|\dot{q}(t) + v(t)\|f(q), \quad k_0 > 0, \quad \psi(t) = m(t)s(t),$$  \hspace{1cm} (152)$$

$$m_1(t) = k_1\|Y(q,\dot{q}_d,\dot{q},\dot{q}_d,\dot{q}_d,t)\|, \quad k_1 > 0, \quad \psi_1(t) = m_1(t)s(t),$$  \hspace{1cm} (153)$$

and the update law for the estimate $\hat{\theta}_0(t)$ of $\theta_0^*$:

$$\dot{\theta}_0(t) = -\Gamma(Y^T(q,\dot{q}_d,\dot{q},\dot{q}_d,\dot{q}_d,t)s + \sigma(t)\theta_0(t)), \quad \Gamma = \Gamma^T > 0$$  \hspace{1cm} (154)$$

where $\sigma(t)$ is a switching signal similar to that in (9.610), using a design parameter $\sigma_0 > 0$ and the knowledge of the upper bound $M_0$ on $\|\theta_0^*\|$:

$$\sigma(t) = \begin{cases} 
0 & \text{if } \|\theta_0(t)\| < M_0, \\
\sigma_0\left(\frac{\|\theta_0(t)\|}{M_0} - 1\right) & \text{if } M_0 \leq \|\theta_0(t)\| < 2M_0, \\
\sigma_0 & \text{if } \|\theta_0(t)\| \geq 2M_0.
\end{cases}$$  \hspace{1cm} (155)$$
This adaptive control scheme has the properties: all signals in the closed-loop system are bounded, and the tracking error \( e(t) = q(t) - q_d(t) \) satisfies
\[
\int_{t_1}^{t_2} \|e(t)\|^2 dt \leq \alpha_0 \left( \frac{\gamma^2}{k_0^2} + \frac{\gamma_1^2}{k_1^2} \right) (t_2 - t_1) + \beta_0
\]
(156)
for some constants \( \alpha_0 > 0, \beta_0 > 0 \) and any \( t_2 > t_1 \geq 0 \), where \( \gamma_1 > 0 \) is the upper bound on \( \sup_{t \geq 0} \| \delta \theta^*(t) \| \). Moreover, \( e(t) \in L^2 \) and \( \lim_{t \to \infty} e(t) = 0 \) in the absence of parameter time variations, that is, when \( \delta \theta^*(t) = 0 \) and \( \frac{\partial D(q,t)}{\partial t} = 0 \).

The proof of these properties is based on the positive definite function
\[
V_0(s, \tilde{\theta}_0) = \frac{1}{2} (s^T D_0 s + \tilde{\theta}_0^T \Gamma^{-1} \tilde{\theta}_0), \quad \tilde{\theta}_0(t) = \theta_0(t) - \theta_0^*, \quad D = D(q(t), t),
\]
(157)
which has the following time derivation:
\[
\dot{V}_0 = -s^T(t)K_D s(t) - m^2(t)s^T(t)s(t) - m_1(t)s^T(t)s(t) - \sigma(t) \tilde{\theta}_0^T(t) \theta_0(t) - \frac{1}{2} s^T(t) \frac{\partial D(q,t)}{\partial t} (q(t) + v(t)) - s^T(t)Y(q, q_d, \dot{q}, \dot{q}_d, \dot{q}_d, t) \delta \theta^*(t)
\]
\[
\leq -s^T(t)K_D s(t) - \left( m(t) \|s(t)\| \right) - \frac{\gamma}{4k_0} + \frac{\gamma_1^2}{16k_0^2} + \frac{\gamma^2}{16k_0^2} - \sigma(t) \tilde{\theta}_0^T(t) \theta_0(t).
\]
(158)

With this adaptive control scheme, as indicated by (156), the tracking performance can be influenced by the design parameters \( k_0 \) and \( k_1 \) in the feedback control law (151)–(153) (one may increase \( k_0 \) and \( k_1 \) to reduce the tracking error \( e(t) \)).

**Adaptive Control Scheme II**

A different adaptive control scheme can be developed, employing a switching control law which uses adaptive estimates of parameter variation uncertainty bounds, to improve system tracking performance.

To derive such a scheme, we denote the parameter variation uncertainties as
\[
g(q, \dot{q}, q_d, \dot{q}_d, t) = \frac{1}{2} \frac{\partial D(q,t)}{\partial t} (\dot{q} + v) = [g_1, g_2, \ldots, g_n]^T
\]
(159)
\[
h(q, \dot{q}, q_d, \dot{q}_d, \dot{q}_d, t) = Y(q, q_d, \dot{q}, \dot{q}_d, \dot{q}_d, t) \delta \theta^*(t) = [h_1, h_2, \ldots, h_n]^T
\]
(160)
and make use of the bounding relationship
\[
|g_i(q, \dot{q}, q_d, \dot{q}_d, t)| \leq a_i^* \alpha_i(q, \dot{q}, q_d, \dot{q}_d, t), \quad i = 1, 2, \ldots, n
\]
(161)
\[
|h_i(q, \dot{q}, q_d, \dot{q}_d, \dot{q}_d, t)| \leq b_i^* \beta_i(q, \dot{q}, q_d, \dot{q}_d, \dot{q}_d, t), \quad i = 1, 2, \ldots, n
\]
(162)
for some unknown constants \( a_i^* \) and \( b_i^* \), and known functions \( \alpha_i(q, \dot{q}, q_d, \dot{q}_d, t) \) and \( \beta_i(q, \dot{q}, q_d, \dot{q}_d, t), \ i = 1, 2, \ldots, n. \)

If the parameters \( a_i^* \) and \( b_i^* \) were known, one could use the control law

\[
 u(t) = Y(q, q_d, \dot{q}, \dot{q}_d, t)\phi(t) - \phi_1(t) - K_Ds(t),
\]

\[
 \phi(t) = [\text{sgn}[s_1(t)]a_1^*\alpha_1, \text{sgn}[s_2(t)]a_2^*\alpha_2, \ldots, \text{sgn}[s_n(t)]a_n^*\alpha_n]^T,
\]

\[
 \phi_1(t) = [\text{sgn}[s_1(t)]b_1^*\beta_1, \text{sgn}[s_2(t)]b_2^*\beta_2, \ldots, \text{sgn}[s_n(t)]b_n^*\beta_n]^T,
\]

where \( \theta_0(t) \) is updated from (154), and the sgn function is defined as

\[
 \text{sgn}[w] = \begin{cases} 
 1 & \text{if } w > 0, \\
 0 & \text{if } w = 0, \\
 -1 & \text{if } w < 0. 
\end{cases}
\]

Form \( V_0 \) defined in (157), this control law leads to

\[
 V_0 = -s^T(t)K_Ds(t) - s^T(t)\phi(t) - s^T(t)\phi_1(t) - \sigma(t)\tilde{\theta}_0^T(t)\theta_0(t) \\
 - \frac{1}{2}s^T(t)\frac{\partial \bar{D}(q, t)}{\partial t}(\dot{q} + v(t)) - s^T(t)Y(q, q_d, \dot{q}, \dot{q}_d, t)\tilde{\theta}^T(t) \\
 = -s^T(t)K_Ds(t) - \sum_{i=1}^n s_i(t)|a_i^*\alpha_i - \sum_{i=1}^n s_i(t)|b_i^*\beta_i - \sigma(t)\tilde{\theta}_0^T(t)\theta_0(t) \\
 - \sum_{i=1}^n s_i(t)g_i - \sum_{i=1}^n s_i(t)h_i \\
 \leq -s^T(t)K_Ds(t).
\]

The last equality follows from the facts that \( \sum_{i=1}^n s_i(t)|a_i^*\alpha_i - \sum_{i=1}^n s_i(t)g_i \geq 0, \)

\( \sum_{i=1}^n s_i(t)|b_i^*\beta_i - \sum_{i=1}^n s_i(t)h_i \geq 0 \)

and \( \sigma(t)\tilde{\theta}_0^T(t)\theta_0(t) \geq 0. \) From (167), one may conclude that all signals in the closed-loop system are bounded, and the tracking error \( e(t) = q(t) - q_d(t) \) converges to zero as \( t \) goes to \( \infty. \)

When the parameters \( a_i^* \) and \( b_i^* \) are unknown, one can use the control law

\[
 u(t) = Y(q, q_d, \dot{q}, \dot{q}_d, t)\phi(t) - \phi(t) - \phi_1(t) - K_Ds(t),
\]

\[
 \phi(t) = [\text{sgn}[s_1(t)]a_1(t)\alpha_1, \text{sgn}[s_2(t)]a_2(t)\alpha_2, \ldots, \text{sgn}[s_n(t)]a_n(t)\alpha_n]^T,
\]

\[
 \phi_1(t) = [\text{sgn}[s_1(t)]b_1(t)\beta_1, \text{sgn}[s_2(t)]b_2(t)\beta_2, \ldots, \text{sgn}[s_n(t)]b_n(t)\beta_n]^T,
\]

where \( \theta_0(t) \) is updated from (154), and the parameters \( a_i(t) \) and \( b_i(t) \) are estimates of \( a_i^* \) and \( b_i^* \) and updated from the adaptive laws:

\[
 \dot{a}_i(t) = \kappa_{ai}|s_i(t)|\alpha_i(q, \dot{q}, q_d, \dot{q}_d, t), \ \kappa_{ai} > 0, \ i = 1, 2, \ldots, n,
\]

\[
 \dot{b}_i(t) = \kappa_{bi}|s_i(t)|\beta_i(q, \dot{q}, q_d, \dot{q}_d, t), \ \kappa_{bi} > 0, \ i = 1, 2, \ldots, n.
\]
Consider the positive definite function

\[ V(s, \tilde{\theta}_0, \bar{a}_i, \bar{b}_i) = \frac{1}{2} (s^T D s + \tilde{\theta}_0^T \Gamma^{-1} \tilde{\theta}_0 + \sum_{i=1}^{n} \kappa_{ai}^{-1} \dot{a}_i^2 + \sum_{i=1}^{n} \kappa_{bi}^{-1} \dot{b}_i^2), \]  

(173)

where \( \tilde{\theta}_0(t) = \theta_0(t) - \theta_0^* \), \( D = D(q(t), t) \), \( \bar{a}_i(t) = a_i(t) - a_i^* \), \( \bar{b}_i(t) = b_i(t) - b_i^*, i = 1, 2, \ldots, n \). Using (149), (168), (154), (171) and (172), we have

\[
\dot{V} = -s^T(t)K_D s(t) - s^T(t) \phi(t) - s^T(t) \phi_1(t) - \sigma(t) \tilde{\theta}_0^T(t) \theta_0(t) \\
- \frac{1}{2} s^T(t) \frac{\partial D(q, t)}{\partial t} (\dot{q}(t) + v(t)) - s^T(t) Y(q, q_d, \dot{q}, \dot{q}_d, \ddot{q}_d, t) \delta \theta^*(t) \\
+ \sum_{i=1}^{n} \bar{a}_i \kappa_{ai}^{-1} \dot{a}_i + \sum_{i=1}^{n} \bar{b}_i \kappa_{bi}^{-1} \dot{b}_i
\]

\[
= -s^T(t)K_D s(t) - \sum_{i=1}^{n} |s_i(t)| a_i(t) \alpha_i - \sum_{i=1}^{n} |s_i(t)| b_i(t) \beta_i - \sigma(t) \tilde{\theta}_0^T(t) \theta_0(t) \\
- \sum_{i=1}^{n} s_i(t) g_i - \sum_{i=1}^{n} s_i(t) h_i + \sum_{i=1}^{n} \bar{a}_i \kappa_{ai}^{-1} \dot{a}_i + \sum_{i=1}^{n} \bar{b}_i \kappa_{bi}^{-1} \dot{b}_i
\]

\[
\leq -s^T(t)K_D s(t) - \sum_{i=1}^{n} |s_i(t)| a_i(t) \alpha_i - \sum_{i=1}^{n} |s_i(t)| b_i(t) \beta_i - \sigma(t) \tilde{\theta}_0^T(t) \theta_0(t) \\
+ \sum_{i=1}^{n} |s_i(t)| a_i^* \alpha_i + \sum_{i=1}^{n} |s_i(t)| b_i^* \beta_i + \sum_{i=1}^{n} \bar{a}_i \kappa_{ai}^{-1} \dot{a}_i + \sum_{i=1}^{n} \bar{b}_i \kappa_{bi}^{-1} \dot{b}_i
\]

\[
= -s^T(t)K_D s(t) - \sigma(t) \tilde{\theta}_0^T(t) \theta_0(t) \\
\leq -s^T(t)K_D s(t).
\]

(174)

This result also implies that all signals in the closed-loop system are bounded, and the tracking error \( e(t) = q(t) - q_d(t) \) converges to zero as \( t \) goes to \( \infty \).

However, since the adaptive control scheme (168) uses the sign functions in \( \phi(t) \) and \( \phi_1(t) \) and such switching signals are discontinuous when \( s_i(t) \) passes through zero, it may lead to chattering of system response.

**Remark 3** The adaptive control scheme (168) may have certain advantage for performance even if when the parameters \( a_i^* \) and \( b_i^* \) are known. This is because the parameters \( a_i^* \) and \( b_i^* \) are only the upper bounds for the parameter variation uncertainties \( g_i \) and \( h_i \) in (159) and (160), and some smaller (and unknown) bounds may exist and can be estimated by the adaptive laws (171) and (172). The use of smaller bounds is desirable because it leads to smaller control signals. In this case, the adaptive laws (171) and (172) can be modified by setting

\[
\dot{a}_i(t) = 0, \ t \geq \tau \text{ if } a_i(\tau) = a_i^*, \]

(175)

\[
\dot{b}_i(t) = 0, \ t \geq \tau \text{ if } b_i(\tau) = b_i^* \]

(176)
With this modification, we also have \( \dot{V} \leq -s^T(t)K_Ds(t) \), as desired. \( \square \)

**Adaptive Control Scheme III**

As mentioned above, the use of the discontinuous sgn functions in \( \phi(t) \) and \( \phi_1(t) \) in the adaptive control scheme (168) may cause chattering of system response. To overcome possible chatterings, we can modify the control law (168) as

\[
\begin{align*}
    u(t) &= Y(q, q_d, \dot{q}, \dot{q}_d, t)\theta_0(t) - \hat{\phi}(t) - \dot{\phi}_1(t) - K_Ds(t), \\
    \hat{\phi}(t) &= [\text{sat}[s_1(t); \varepsilon_1]a_1(t)\alpha_1, \ldots, \text{sat}[s_n(t); \varepsilon_n]a_n(t)\alpha_n]^T, \\
    \dot{\phi}_1(t) &= [\text{sat}[s_1(t); \eta_1]b_1(t)\beta_1, \ldots, \text{sat}[s_n(t); \eta_n]b_n(t)\beta_n]^T,
\end{align*}
\]

where the \text{sat} function is defined as

\[
\text{sat}[s_i; x_i] = \begin{cases} 
1 & \text{if } s_i > x_i, \\
\frac{s_i}{x_i} & \text{if } |s_i| \leq x_i, \\
-1 & \text{if } s_i < -x_i
\end{cases}
\]

for some chosen \( x_i > 0 \) (for \( x_i = \varepsilon_i \) or \( x_i = \eta_i \) in (178) and (179)), \( i = 1, 2, \ldots, n \), with the associated indicator functions

\[
\chi[s_i; x_i] = \begin{cases} 
1 & \text{if } |s_i| > x_i, \\
0 & \text{if } |s_i| \leq x_i.
\end{cases}
\]

Such functions have the property: \( \chi[s_i; x_i](1 - \chi[s_i; x_i]) = 0 \) (that is, \( \chi[s_i; x_i] = 0 \) whenever \( 1 - \chi[s_i; x_i] = 1 \), and \( \chi[s_i; x_i] = 1 \) whenever \( 1 - \chi[s_i; x_i] = 0 \).

The adaptive laws are also modified as

\[
\begin{align*}
    \dot{a}_i(t) &= \chi[s_i; \varepsilon_i]\kappa_{ai}[s_i(t)\alpha_i(q, \dot{q}, q_d, \dot{q}_d, t), \kappa_{ai} > 0, i = 1, 2, \ldots, n, \\
    \dot{b}_i(t) &= \chi[s_i; \eta_i]\kappa_{bi}[s_i(t)\beta_i(q, \dot{q}, q_d, \dot{q}_d, t), \kappa_{bi} > 0, i = 1, 2, \ldots, n.
\end{align*}
\]

With this modification, we have

\[
\begin{align*}
    \dot{V} &= -s^T(t)K_Ds(t) - s^T(t)\hat{\phi}(t) - s^T(t)\dot{\phi}_1(t) - \sigma(t)\hat{\theta}^T_0(t)\theta_0(t) \\
    &\quad - \frac{1}{2}s^T(t)\frac{\partial D(q, t)}{\partial t}(\dot{q}(t) + v(t)) - s^T(t)Y(q, q_d, \dot{q}, \dot{q}_d, t)\delta \theta^*(t) \\
    &\quad + \sum_{i=1}^n \hat{a}_i\kappa_{ai}^{-1}\dot{a}_i + \sum_{i=1}^n \hat{b}_i\kappa_{bi}^{-1}\dot{b}_i \\
    &= -s^T(t)K_Ds(t) - \sum_{i=1}^n s_i(t)\text{sat}[s_i; \varepsilon_i]a_i(t)\alpha_i - \sum_{i=1}^n s_i(t)\text{sat}[s_i; \eta_i]b_i(t)\beta_i
\end{align*}
\]
\[- \sum_{i=1}^{n} s_i(t) g_i - \sum_{i=1}^{n} s_i(t) h_i + \sum_{i=1}^{n} a_{ii}^{-1} a_i + \sum_{i=1}^{n} b_{ii}^{-1} b_i - \sigma(t) \hat{\theta}_0^T(t) \theta_0(t) \leq -s^T(t) K_D s(t) - \sum_{i=1}^{n} s_i(t) \text{sat}[s_i; \varepsilon_i] a_i(t) \alpha_i - \sum_{i=1}^{n} s_i(t) \text{sat}[s_i; \eta_i] b_i(t) \beta_i \]
\[+ \sum_{i=1}^{n} s_i(t) a_i^* \alpha_i + \sum_{i=1}^{n} s_i(t) b_i^* \beta_i + \sum_{i=1}^{n} a_{ii}^{-1} a_i + \sum_{i=1}^{n} b_{ii}^{-1} b_i - \sigma(t) \hat{\theta}_0^T(t) \theta_0(t) \]
\[= -s^T(t) K_D s(t) - \sum_{i=1}^{n} (1 - \chi[s_i; \varepsilon_i]) s_i(t) \text{sat}[s_i; \varepsilon_i] a_i(t) \alpha_i \]
\[- \sum_{i=1}^{n} (1 - \chi[s_i; \eta_i]) s_i(t) \text{sat}[s_i; \eta_i] b_i(t) \beta_i + \sum_{i=1}^{n} (1 - \chi[s_i; \varepsilon_i]) |s_i(t)| a_i^* \alpha_i \]
\[+ \sum_{i=1}^{n} (1 - \chi[s_i; \eta_i]) |s_i(t)| b_i^* \beta_i - \sigma(t) \hat{\theta}_0^T(t) \theta_0(t). \quad (184) \]

This modified scheme would ensure the closed-loop signal boundedness but not the asymptotic convergence of the tracking error \(e(t) = q(t) - q_d(t)\) to zero (only a bounded tracking error \(e(t) = q(t) - q_d(t)\) of the order \(\varepsilon_i\) and \(\eta_i\)).
**Derivation of (10.152)**

In this case, the expression (5.30) also holds

\[
    u(t) = \phi_1^* \frac{a(s)}{\Lambda(s)} [u(t)] + \phi_2^* \frac{a(s)}{\Lambda(s)} [y](t) + \phi_3^* P_m(s)[y](t) \tag{185}
\]

(with \(\theta_i^*\) replaced by \(\phi_i^*\) as the new notation and the exponentially decaying term \(\varepsilon_1(t)\) ignored). Recall (10.39):

\[
    u(t) = u_d(t) + (\theta - \theta^*)^T \omega(t) + d_N(t) \tag{186}
\]

and (10.15) (with \(a_s(t) = 0\) for simplicity):

\[
    u_d(t) = -\theta^T(t) \omega(t) \tag{187}
\]

from which we have

\[
    u(t) = -\theta^* \omega(t) + d_N(t). \tag{188}
\]

Using (186) for \(u(t)\) in the left side of (185) and (188) for \(u(t)\) in the right side of (185), we obtain

\[
    u_d(t) + (\theta - \theta^*)^T \omega(t) + d_N(t)
    = \phi_1^* \frac{a(s)}{\Lambda(s)} [-\theta^* \omega(t) + d_N(t)] + \phi_2^* \frac{a(s)}{\Lambda(s)} [y](t) + \phi_3^* y(t) + \phi_3^* P_m(s)[y](t). \tag{189}
\]

Subtracting (189) from (10.151) and recalling (10.148), we have (10.152).
Backstepping Design with Nonsmooth Inverse Signal

For an adaptive inverse control scheme using a backstepping feedback design for the output-feedback nonlinear systems (see Section 10.6.3), the nonlinearity inverse signal \( \omega(t) \) in (10.280), defined in (10.21) or (10.35) for the case of an adaptive dead-zone inverse \( \hat{D}J(\cdot) \) or an adaptive backlash inverse \( \hat{B}J(\cdot) \), is not smooth (it is not even continuous; but it is well-defined). A backstepping-based feedback control design needs certain smoothness condition. Will the nonsmoothness of \( \omega(t) \) cause any problem? The answer is “no”. It turns out that the use of filters in the state observer helps to avoid the potential difficulty with nonsmoothness of such signals in a backstepping design.

Starting from (10.277), (10.283) and (10.284), we see that the nonsmooth signal \( \omega(t) \) passes the filters \( p_{ij}(s) \), where \( p_{ij}(s) \) are the components of \( (sI - A_0)^{-1} \) for \( A_0 = A - kc \) with \( k = [k_1, k_2, \ldots, k_n]^T \) and \( (A, c) \) in (10.272). This means that only the signals \( \omega_{ij}(t) \) in (10.285) are associated with the nonsmooth signal \( \omega(t) \). In fact, it is the signal \( \hat{x}_2(t) \) defined in (10.128) that plays a key role in the backstepping design procedure, that is, only those of the signals \( \omega_{ij}(t) \), for \( i = 0, 1, \ldots, m \) and \( j = 2 \), are crucial in the backstepping design.

From the canonical form of \( A \) and \( c \), we can obtain

\[
\begin{align*}
p_{02}(s) &= s + k_1, \quad p_{12}(s) = s(s + k_1), \quad p_{22}(s) = s^2(s + k_1), \ldots, \\
p_{n-2,2}(s) &= s^{n-2}(s + k_1), \quad p_{n-1,2}(s) = -k_2s^{n-2} - k_3s^{n-3} - \cdots - k_{n-1}s - k_n.
\end{align*}
\]

This calculation can be verified by using the symbolic algebra operations in Matlab. For example, for \( n = 5 \), we can use

\[
a = \text{sym('[s+k1, -1, 0, 0, 0; k2, s, -1, 0, 0; k3, 0, s, -1, 0; k4, 0, 0, s, -1; k5, 0, 0, 0, s]')} \\
b = \text{inv(a)} \\
c = \text{symmul(det(a), b)}
\]

to get the numerator matrix of \( (sI - A_0)^{-1} \) as

\[
\begin{bmatrix}
s^4, & s^3, & s^2, & s, & 1 \\
-s^3k2-k3s^2-k4s-k5, & s^3(s+k1), & s^2(s+k1), & (s+k1)s, & s+k1 \\
-ss^2+k3^2+4s+k5^2, & -k3s^2-k4s-k5, & s^2(k2+s^2+s*k1), & k2s^2+s*k1, & k2s^2+s*k1 \\
-s^2(k4s+k5), & -s(k4s+k5), & -k4s-k5, & s(k2s+k3+s^3+s^2*k1), & k2s+k3+s^3+s^2*k1
\end{bmatrix}
\]
In the backstepping design procedure, the derivatives $\omega^{(j)}(t)$, $i = 0,1,\ldots,m$, $j = 0,1,\ldots,\rho-1$, are needed, where $\rho = n - m$. For the case of $\rho = n - m = 1$, no derivative of $\omega(t)$ is used. For $\rho = n - m \geq 2$, the highest degree of $s^{\rho-1}p_{i2}(s)$, $i = 0,1,\ldots,m - 1$, is $\rho - 1 + m = n - 1$, and the degree of $s^{\rho-1}p_{m2}(s)$ is $n$, so that $\frac{s^{\rho-1}p_{i2}(s)}{\Lambda(s)}$, $i = 0,1,\ldots,m - 1$, are strictly proper, and $\frac{s^{\rho-1}p_{m2}(s)}{\Lambda(s)}$ is proper. Therefore, the signals $\omega^{(j)}(t)$, $i = 0,1,\ldots,m$, $j = 0,1,\ldots,\rho - 1$, are well-defined, and so all other signals related the inverse signal $\omega(t)$ in the backstepping design procedure, that is, the nonsmoothness $\omega(t)$ does not cause any problem for the backstepping-based adaptive inverse control scheme.