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Model Reference Adaptive Pole-Placement Controller of Nonminimum Phase Systems

(비최소 위상 시스템에 대한 기준 모델 적응 폴-플레이스먼트 제어기)

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要 約

단일 입·출력 시불변 비최소 위상 공정에 대한 폴-플레이스먼트 제어기를 기준 모델을 사용하여 설계하였다. 제안된 폴-플레이스먼트 제어기는 변수 다항식 $S(q^{-1})$ 만을 사용하여 파라미터 형으로 나타내어 기준모델의 전달함수와 제어기의 전달함수를 같게 하는 방법을 사용하였다.

또한, 적응 폴-플레이스먼트 제어기는 위의 제어기에 적응 알고리즘을 적용하여 제어기의 파라미터를 추정하도록 설계되었다.

Abstract

A pole-placement control of discrete, deterministic, single-input single-output nonminimum phase systems is considered using a model reference type approach.

The proposed pole-placement controller is designed in the parameter form to make the transfer function of the controller equal to that of the reference model with only single variable polynomial $S(q^{-1})$.

The proposed adaptive pole-placement controllr is designed with the true system parameters by applying the adaptation method to the proposed pole-placement controller.

I. Introduction

The pole-placement control technique is a useful approach to the control of the dynamic system with known parameters.[1,2]

Such a scheme generally requires simple computations, so that the pole-placement controller is easily implemented with a digital computer.

From this standpoint, the pole-placement control system is designed in the discrete form using reference model.

The schemes by Åström and Wittenmark[1] are remarkable in the sense that the pole—placement controller is designed in the general form based on input-output models.

This paper proposes a novel scheme for designing the general pole-placement controller using reference model.

Since we can choose the reference model

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arbitrarily, with only single variable polynomial $S(q^{-1})$ we can make the transfer function of the pole-placement controller equal to that of the reference model.

It is important to point out that the desired response is obtainable only under the ideal condition that the cancellation of poles and zeros as required by the design is exact.

And if the parameters of the plant are changed by the load disturbances or a little non-linearity of the plant exists, the control objectives cannot be achieved easily.

So we introduce the adaptive algorithm to the proposed pole-placement controller. In the non-minimum phase system case the only situation in which global convergence does not follow is when a pole-zero cancellation is a limit point of the algorithm.

For these pathological cases [3] the adaptive scheme is extended to control non-minimum phase system by introduction of a suitable criterion function which includes weighting on the control inputs.

This modification in terms of zero shifting[4] is useful in showing the transformation of a non-minimum phase system into an augmented minimum phase plant.

These augmentations allow the proposed pole-placement controller to be used with non-minimum phase plant.

Computer simulations to illustrate the pole-placement control and the adaptive pole-placement control of a realistic non-minimum phase plant are also included.

II. Design of a Pole-Placement Controller Using Reference Model.

II-1. Statement of the problem in the linear Case.

Consider a single-input single-output, discrete, time-invariant plant described by

$$A(q^{-1})y(k) = q^{-d}B(q^{-1})\,U(k)\ ;\ d>0\ ;\quad y(0)\neq 0 \tag{1}$$

where

$$A(q^{-1}) = 1+a_1q^{-1} + \cdots + a_{na}q^{-na} \tag{2}$$

$$B(q^{-1}) = b_0+ b_1q^{-1}+ \cdots + b_{nb}q^{-nb}\ ;\quad b_0\neq 0 \tag{3}$$

$\{d\}$ represents the plant time delay, $\{u(k)\}$ and $\{y(k)\}$ are the plant input and out, respectively. $A(q^{-1})$ and $B(q^{-1})$ are relatively prime polynomials in the delay operator q^{-1}, and $B(q^{-1})$ is factored as $B(q^{-1}) = B^+(q^{-1})\,B^-(q^{-1})$ where the zeros of $B^+(q^{-1})$ are all in $|q|<1$ and the zeros of $B^-(q^{-1})$ are all in $|q|>1$, and

$$B^+(q^{-1}) = 1 + b_{1+}q^{-1} + \cdots + b_{nb+}q^{-nb^+} \tag{4}$$

$$B^-(q^{-1}) = b_0+b_{1-}q^{-1} + \cdots + b_{nb-}q^{-nb^-} \tag{5}$$

A reference model is given by

$$A_m(q^{-1})Y_m(k) = q^{-d}B_m(q^{-1})U_m(k) \tag{6}$$

where

$$A_m(q^{-1}) = a_{mo}+ a_{ml}q^{-1}+ \cdots + a_{ma}q^{-ma} \tag{7}$$

$$B_m(q^{-1}) = b_{m0}+b_{ml}q^{-1}+ \cdots + b_{mb}q^{-mb} \tag{8}$$

$\{U_m(k)\}$ and $\{Y_m(k)\}$ are the reference model input and output, respectively, and $A_m(q^{-1})$ is an asymptotically stable polynomial, and $B_m(q^{-1})$ is factored as $B_m(q^{-1}) = B^-(q^{-1})\,B'_m(q^{-1})$

where

$$B'_m(q^{-1})=b_{m0}'+ b_{ml}'q^{-1} + \cdots + b_{mb}\,q^{-mb'} \tag{9}$$

The problem to be considered is to design a controller for the system (1) in such a way that the closed loop satisfies certain requirements on both tracking and regulation. The tracking (or reference following) and regulation objectives are specified seperately as follows.

i) The control should be such that in tracking, the output of the plant satisfies the equation

$$A_m(q^{-1})Y(k) = q^{-d}B_m(q^{-1})U_m(k) \tag{10}$$

ii) The regulation objectives are specified in terms of the desired response to initial errors, assuming that the command U_m is zero. Compute a control law such that the plant-model error vanishes with the dynamics defined by a monic asymptotically stable polynominal $C(q^{-1})$, i.e.

$$C(q^{-1})\,[Y(k+d) - Y_m(k+d)] = 0 \;\forall k > 0 \qquad (11)$$

where

$$C(q^{-1}) = 1 + C_1 q^{-1} + \cdots + C_{nc} q^{-nc} \qquad (12)$$

II-2 Design procedures.

The requirements above on tracking and regulation can be fulfilled with a simple design procedure, basically a pole-placement design.

A first requirement is obtained by using an explicit reference model (6) and by applying the following lemma.[5]

Lemma II-1

Consider the plant (1), the reference model (6) and the plant control input given by (Fig. 1)

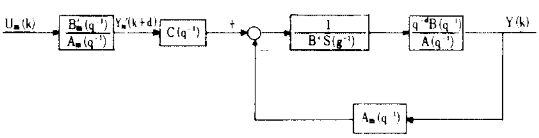

Fig. 1. Linear control.

$$U(k) = \frac{1}{B^+(q^{-1})S(q^{-1})}\,[\,C(q^{-1})Y'_m(k+d) - A_m(q^{-1})Y(k)\,] \qquad (13)$$

or

$$U(k) = C(q^{-1})Y'_m(k+d) - A_m(q^{-1})Y(k) + B_s(q^{-1})\,U(k) \qquad (13\text{-a})$$

with

$$B_s(q^{-1}) = 1 - B^+(q^{-1})\,S(q^{-1})$$

where

$$S(q^{-1}) = 1 + S_1 q^{-1} + \cdots + S_{ns} q^{-ns} \qquad (14)$$

$$Y'_m(k) = q^{-d}\,\frac{B'_m(q^{-1})}{A_m(q^{-1})}\,U_m(k) \qquad (15)$$

verify the identity

$$C(q^{-1}) = A(q^{-1})S(q^{-1}) + q^{-d}B^-(q^{-1})A_m(q^{-1}) \qquad (16)$$

Then

(1) An initial plant-model error or an initial output disturbance converges to zero with the dynamics of the C - polynomial, i.e.

$$C(q^{-1})e(k+d) = 0 \;\forall k \geqslant 0 \Rightarrow \lim_{k\to\infty} e(k) = 0 \qquad (17)$$

where

$$e(k) = Y(k) - Y_m(k) \qquad (18)$$

(2) The transfer function from $U_m(k)$ to $Y(k)$ is given by eqn. (10).

(3) If one chooses

$$n_s \geqslant 0\;;\; n_c = \max\,(n_a + n_s,\, d + n_b^- m_a)$$

then we can obtain a monic asymptotically polynomial $C(q^{-1})$ by using a polynomial $S(q^{-1})$.

Proof.

Using eqns. (1), (6), (13) and (15), eqn. (17) becomes

$$\begin{aligned} C(q^{-1})e(k+d) &= C(q^{-1})\,[Y(k+d) - Y_m(k+d)] \\ &= [A(q^{-1})S(q^{-1}) + q^{-d}B^-(q^{-1})A_m(q^{-1})]\,Y(k+d) - C(q^{-1})Y_m(k+d) \\ &= B(q^{-1})S(q^{-1})U(k) + B^-(q^{-1})A_m(q^{-1})Y(k) - C(q^{-1})Y_m(k+d) \\ &= 0 \end{aligned} \qquad (19)$$

From eqn. (1), (6) and (13)

$$Y(k) = \frac{C(q^{-1})}{A(q^{-1})\,S(q^{-1}) + q^{-d}B^-(q^{-1})\,A_m(q^{-1})}\;\frac{q^{-d}B^-(q^{-1})B'm(q^{-1})}{A_m(q^{-1})}\,U_m(k) \qquad (20)$$

and using the identity (16) one concludes that (20) is identical to (6). ■

Let us rewrite (13) and (19) in the parameter form

$$U(k) = [C(q^{-1})Y'm(k+d) - Am(q^{-1})Y(k) - -P^T\phi(k)] \quad (21)$$

$$C(q^{-1})\, e(k+d) = B^-(q^{-1})[U(k) + Am(q^{-1})Y(k) - C(q^{-1})Y'm(k+d) + P^T\phi(k)] \quad (22)$$

where

$$\phi^T(k) = [U(k-1)U(k-2) - U(k-n_{b+} - n_s)] \quad (23)$$

$$P^T = [S_1+b_{1+} \quad s_2+b_{1+}\; s_1+b_{2+} \cdots b_{nb+}\; s_{ns}]$$

Thus we can control a general non-minimum phase system using reference model.

Note that in some cases, it would be desirable to use $B'_m(q^{-1})$ instead of $A_m(q^{-1})$ to satisfy the identity (16).

And also in the minimum phase system case it is shown for the algorithms in [6] that $\{Y(k)\}$ tends to $\{Y_m(k)\}$. We do not make this claim for the general case. In fact it is clear that this is always impossible when the system does not have a stable inverse. Thus, special consideration has to be given to the question of steady-state errors in the output sequence.

If required, integral action can be incorporated into the pole-placement controller in a straight forward fashion.

This may be achieved by requiring that the polynomial $S(q^{-1})$ have the form

$$S(q^{-1}) = (1-q^{-1})^{\ell} S_1(q^{-1}). \quad (24)$$

III. Design of an Adaptive Pole-Placement Controller

In the previous section, we have calculated the coefficients of $S(q^{-1})$ to obtain a monic asymptotically stable polynomial $C(q^{-1})$, but in this section, we employ adaptive alrotithm under the obtained monic asymptotically stable polynomical $C(q^{-1})$ to adjust the parameters of the controller.

For the control of non-minimum phase systems we have to make J (the criterion function) arbitrarily small and yet U finite.

This can be achieved by incorporating a criterion function that allows the control input to be weighted.

With such a facility non-minimum phase systems can be easily controlled by the appropriate weighting on U to give a finite J and a bounded U.

Let us define the output, Y', of an augmented plant as

$$Y'(k) = Y(k) - M(q^{-1})\, U(k-d) \quad (25)$$

where $M(q^{-1})$ is a constant polynomial in q^{-1}. $M(q^{-1})$ is a chosen such that all the zeros of the augmented plant lie inside the unit circle in the Z-plane.

To eliminate steady-state offset the $M(q^{-1})$ polynomial has the form

$$M(q^{-1}) = M'\,(1-q^{-1})^m \quad (26)$$

Then we can use the controller of Fig. 1 in section II.

It is natural to replace the vector P in (23) by adjustable parameters $\hat{p}(k)$ which will be updated by the adaptation mechanism.

Therefore the control U(k) in the adaptive case is given by

$$U(k) = [C(q^{-1})\, Y_m(k+d) - A_m(q^{-1})Y'(k) - \hat{P}^T(k)\,\phi(k)] \quad (27)$$

which can also be written as

$$C(q^{-1})\, Y_m(k+d) = A_m(q^{-1})Y'(k) + P_o^T(k)\phi_o(k) \quad (28)$$

where

$$\hat{P}_o^T(k) = [\hat{b}_o(k) : \hat{P}^T(k)]$$

$$\phi_o^T(k) = [U(k) : \phi^T(k)]$$

Introducing (28) into (22) and defining the filtered error as

$$e^f(k+d) = [A_m(q^{-1})Y'(k) - C(q^{-1})Y_m(k+d) + P_o^T\phi(k)] \quad (29)$$

one obtains

$$e^f(k) = [P_o - \hat{P}_o(k-d)]^T\,\phi_o(k-d) \quad (30)$$

where $P^T = [b_o : P^T]$

$\phi_o^T(k) = [U(k) : \phi^T(k)]$

Let us define the auxiliary error as

$$\bar{e}(k) = [\hat{P}_o(k-d)-\hat{P}_o(k)]^T\phi_o(k\text{-}d) \qquad (31)$$

and the augmented error as

$$e^*(k) = e^f(k)+\bar{e}(k) = [P_o-\hat{P}_o(k)]^T \phi_o(k-d) \qquad (32)$$

To evaluate the deviation between the augmented plant and the reference model, we introduce the following criterion function:

$$J(k) = \sum_{j=d}^{k} \lambda_2\lambda_1^{k\text{-}j}[\bar{Y}(j)-\hat{P}_o^T(k)\,\phi_o(j-d)]^2 \qquad (33)$$

where λ is a weighting coefficient given

as $0<\lambda_1<1$, $\lambda_2 = 1 - \lambda_1$ and $\bar{Y}(k) = C(q^{-1})Y'(k) - A_m(q^{-1})Y(k\text{-}d)$

The estimate $\hat{P}_o(k)$ is determined so that the criterion function $J(k)$ becomes minimum at each k.

Letting the gradient of $J(k)$ with respect to $\hat{P}_o(k)$ be zero, and employing the matrix inversion lemma, we can obtain the following recursive equations.

$$\hat{P}_o(k+1) = \hat{P}_o(k)+L(k+1)[\bar{Y}(k+1)-\hat{P}_o(k)^T\phi_o(k-d+1)] \qquad (34)$$

$$\Gamma(k+1) = \frac{1}{\lambda_1}[I-L(k+1)\phi_o(k-d+1)^T]\,\Gamma(k) \qquad (35)$$

$$L(k+1) = \frac{\dfrac{\Gamma(k)\,\phi_o(k-d+1)}{\lambda_1}}{\dfrac{1}{\lambda_2} + \phi_o(k-d+1)^T\dfrac{\Gamma(k)}{\lambda_1}\phi_o(k-d+1)} \qquad (36)$$

We can avoid unnecessary storage and improve both accuracy and computational efficiency by applying UDU^T factorization methods.[7]

To obtain gain L(k), let us define following algorithm.

Let

$$f = U^T(k-1)\,\phi_o(k-d) \qquad (37)$$

$$V = D(k-1)\,f/\lambda_1 \qquad (38)$$

where

$$f^T = [f_1\ f_2 - f_n]$$

$$V^T = [V_1\ V_2 - V_n]$$

$$V_i = d_i\,(k-1)\,f_i/\lambda_1 \quad (i=1,2,-n)$$

$$r_1 = \frac{1}{\lambda_2} + V_1 f_1 \qquad (39)$$

$$d_1(k) = d_1(k-1)/\lambda_1\lambda_2 r_1 \qquad (40)$$

$$K_2^T = [V_1\ 0 \overset{n-1}{\text{---}}\ 0] \qquad (41)$$

For j = 2,3, -- , n cycle from (42) to (46)

$$r_j = r_{j-1} + V_i f_i \qquad (42)$$

$$d_j(k) = \frac{d_j(k-1)}{\lambda_1}\,\frac{r_j\text{-}\hat{z}}{r_j} \qquad (43)$$

$$U_j(k) = U_j(k-1)+w_jK_j \qquad (44)$$

$$\text{where} \quad w_j = -\frac{f_j}{r_{j-1}} \qquad (45)$$

$$K_{j+1} = K_j+V_jU_j(k-1) \qquad (46)$$

where $U_j(k-1)$ is the j-th column of matrix $U(k-1)$

$$L(k) = \frac{K_{n+1}}{r_n} \qquad (47)$$

IV. Sensitivity to Modelling Errors

It is highly unrealistic to assume that the mathematical model of the plant used in a control design is accurate. Therefore, it is important to understand how modelling errors will influnce the closed-loop properties, which is discussed in this section.

It is assumed that the control design is based on the mathematical model $H = \dfrac{q^{-d}B(q^{-1})}{A(q^{-1})}$,

while the true model is $H^o = \dfrac{q^{-d}B^o(q^{-1})}{A^o\,(q^{-1})}$.

The control law of (13) represents a combination of a feedforward from the command signal U_m with the pulse-transfer function

$$H_{ff}(q^{-1}) = \frac{C(q^{-1})\,B'_m(q^{-1})}{B^+(q^{-1})S(q^{-1})A_m(q^{-1})} \qquad (48)$$

and a feedback from the measured output Y with the pulse-transfer function

$$H_{fb}(q^{-1}) = \frac{A_m(q^{-1})}{B^+(q^{-1})S(q^{-1})} \qquad (49)$$

The following result [1] describes the influence of modelling errors on the stability of the closed-loop system.

Theorem IV. 1.

Consider a pole-placement design based on an approximate model

$$H = \frac{q^{-d}B(q^{-1})}{A(q^{-1})}$$

Let H be the true pulse-transfer function of the plant to be controlled. Assume that H and H^o have the same number of poles outside the unit disc and that H_m is stable. Then the closed-loop system related to $H^o(q^{-1})$ is stable if

$$|H(q^{-1})-H^o(q^{-1})| < \left|\frac{H(q^{-1})}{H_m(q^{-1})}\right|\left|\frac{H_{ff}(q^{-1})}{H_{fb}(q^{-1})}\right| \qquad (50)$$

for $|q^{-1}| = 1$, where H_{ff} and H_{fb} are defined by (48) and (49), respectively.

Proof.

Loop gain is

$$H_{1g} = \frac{q^{-d}B^-(q^{-1})A_m(q^{-1})}{A(q^{-1})S(q^{-1})} \qquad (51)$$

It follows from (16) that

$$1 + H_{1g} = \frac{C(q^{-1})}{A(q^{-1})S(q^{-1})} = \frac{H}{H_m}H_{ff} \qquad (52)$$

After multiplication by $\dfrac{A_m(q^{-1})}{B^+(q^{-1})S(q^{-1})}$,

the condition given by (50) can be written as

$$\left|\frac{A_m(q^{-1})}{B^+(q^{-1})S(q^{-1})}H - \frac{A_m(q^{-1})}{B^+(q^{-1})S(q^{-1})}H^o\right| < \left|\frac{H}{H_m}H_{ff}\right| = |1 + H_{1g}|$$

or $\quad |H_{1g} - H_{1g}^o| < |1+H_{1g}|$

Hence,

$$|H(q^{-1}) - H^o(q^{-1})| < \left|\frac{H(q^{-1})}{H_m(q^{-1})}\right|\left|\frac{H_{ff}(q^{-1})}{H_{fb}(q^{-1})}\right|$$

V. Computer Simulations

We present in this section simulation results illustrating the controller designs in section II and III acting in tracking, parameter disturbances and regulation.

The non-minimum phase plant before a parameter change occurs ($k < l_1$) is represented by:

$$Y(k) = \frac{q^{-1}(0.51+1.21q^{-1})}{1-0.44\,q^{-1}}U(k)\ ;Y(0)=0.05 \qquad (53)$$

This plant corresponds to an open-loop description of a condensation polymer process.[4]

At time $k=l_1$ a change of the plant parameters is made.[8] The plant is then ($k \geqslant l_1$) characterized by

$$Y(k) = \frac{q^{-1}(0.4+1.2\,q^{-1})}{1-0.1\,q^{-1}}U(k) \qquad (54)$$

At time $k=l_2$, the plant is made to the original state as equation (53) ($k \geqslant l_2$).

At time $k=l_3$, the input of the reference model is made to zero. ($k \geqslant l_3$).

In this simulation, $l_1=100$, $l_2=200$ and $l_3=300$ are chosen.

For the simulation of the controller in section II the reference model is chosen, arbitrarily, as ;

$$Y_m(k)=q^{-1}\frac{0.2(0.51+1.21q^{-1})}{1-0.312q^{-1}}U_m(k) \qquad (55)$$

hence

$$Y_m'(k) = q^{-1}\frac{0.2}{1-0.312\,q^{-1}}U_m(k) \qquad (56)$$

where Um(k) represents the setpoint.
We choose $S(q^{-1})$ as $(1-q^{-1})(1+0.5q^{-1})$.
Then $P^T = [-0.5 \quad -0.5]$
and $C(q^{-1}) = 1\text{-}0.43q^{-1}+0.77088q^{-2} - 0.15752q^{-3}$

For the simulation of the controller in section III the augmented plant is chosen as :

$$Y'(k) = q^{-1}\,\frac{2.51-1.67q^{-1}+0.88q^{-2}}{1-0.44q^{-1}}\,U(k) \qquad (57)$$

where $M(q^{-1})$ is chosen as $-2(1-q^{-1})$ to eliminate steady-state offset.
And the reference model is chosen as:

$$Ym(k) = q^{-1}\,\frac{0.25}{1-0.5q^{-1}}\,Um(k) \qquad (58)$$

We choose $S(q^{-1})$ as $(1-q^{-1})$.
Then $C(q^{-1}) = 1\text{-}0.44q^{-1}-0.06q^{-2}$.
And $\lambda_1=0.91$, $\lambda_2=0.09$,
$Po^T(o) = [-2.5 \quad -4.0 \quad -2.0 \quad 0.0]$

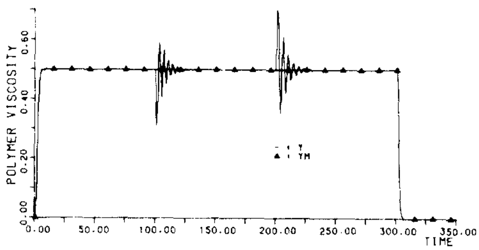

Fig. 2. Output response of the section 2.

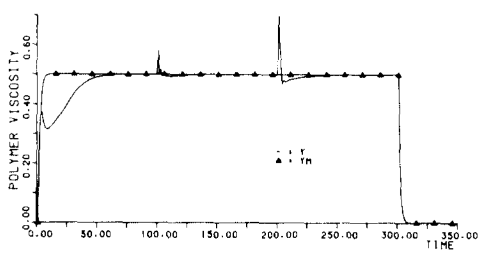

Fig. 3. Output response of the section 3.

VI. Conclusions

We have proposed a novel scheme for designing the pole-placement controller and its

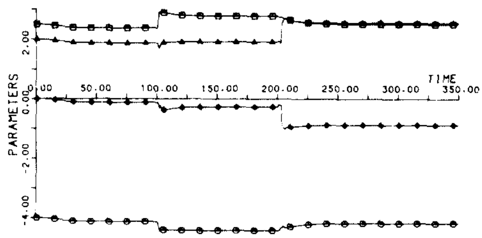

Fig. 4. Controller parameters of the Fig. 3.

application to an adaptive pole-placement controller using reference model.

The design procedure of our scheme is simple and systematic, and the control algorithms can be easily implemented with a microcomputer.

The plant to be controlled has not been assumed to be either stable or stably invertible.

Computer simulations illustrating the performance of the algorithms have also been given for the realistic non-minimum phase plant.

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