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적분동작이 포함된 적응제어기

Adaptive Controllers with Integral Action

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요 약

부하외란에 의한 잔류편차를 제거할 수 있는 적분동작이 포함된 적응제어기가 제안된다. 잔류편차에 대한 효과적인 적분동작 및 강인한 인식은 영이득예측자[2]에 의해서 달성될 수 있다. 본 논문에서 이 시스템은 좀더 일반화된 구조의 시스템으로 개선되고, 비최소위상 시스템을 취급할 수 있게하는 detuning control 가중치는 온라인으로 동조된다. 이 개선된 시스템의 이산시간 시스템이 개발되어 설계 파라미터의 선정에 유연성을 준다. 본 논문 제어 시스템은 모형화되지 않은 동적특성에 대하여도 강인한 것으로 보여진다.

Abstract—A class of adaptive controllers with integral action is proposed, which may reject the offset due to any load disturbance on the plant. Effective integral action and robust identification against the offset can be achieved via the zero-gain predictor[2]. The system is improved, in this paper, to be of more generalized structure, and the detuning control weight which can cope with nonminimum-phase systems is tuned on-line. Discrete-time versions of the improved system are developed, which may be more flexible for the choice of the design parameters. The resulting control systems may also be shown to be robust to the unmodelled dynamics.

1. Introduction

In this paper, adaptive control systems with integral action are developed based on the GMV (Generalized Minimum Variance) technique. The steady-state error or the offset caused by any load disturbance or nonzero mean setpoint can be eliminated via a variety of approaches, including adaptive parameter adjustment, but the usual method is to incorporate integral action. This can be achieved by physically cascading an integrator into the control loop [1, 4, 5], or modelling the system to include an offset term[2]. These system can be derived by resorting to CARIMA (Controlled Auto-Regressive and Integrated Moving Average) model of the plant[3], which appears to be the most appropriate representation of disturbed real process.

This paper improves the continuous-time control systems proposed by Gawthrop[2] to be of more generalized structure as follows;

1) The system is described as a CARIMA model, which may be more natural. For example, polynomials $A(s)$ and $B(s)$ which define the system dynamics are assumed such that $A(0)\neq 0$ and $B(0)\neq 0$ instead of $A(0)=B(0)=0$ in the system[2], but the resulting controllers are similar.

2) The control system may be of more generalized structure, which results in that choice of the polyno-

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mial C(s) determining the disturbance dynamics is flexible, i.e., C(s) is chosen such that $C(0)\neq 0$ instead of $C(0)=1$.

3) The detuning term is tuned on-line: A detuned version of the control systems can cope with non-minimum-phase systems. But how to choose the detuning weight is still not clear. A method is proposed here, by which the weight could be tuned on-line. This controller may also be shown to be robust against the unmodelled dynamics.

In addition to the improvements, discrete-time versions of the systems are developed as follows:

1) Discretized or digitalized control systems may be required with developements of microprocessors or digital computers.

2) The choice of the order of the polynomial C(s) may be not restricted, since the "proper" condition for obtaining finite gain at infinite frequency in the continuous-time approach is not required. This fact contributes to establishment of more generalized structure.

The control systems are described as discrete-time systems. The paper is organized as follows. The system is modelled and the control objective is discussed in section 2. In section 3, the controller is derived. A detuned version of the controller is presented in section 4. Simulation results and conclusion are given in section 5 and 6 respectively.

2. Problem statements

In the continuous-time approach, the system has been modelled to include an offset term for obtaining a zero-gain predictor. This leads to that the polynomials A(s) and B(s) which define the system dynamics are such that $A(0)=B(0)=0$. In this paper, the system is described as a CARIMA model, and the polynomials $A(0)\neq 0$ and $B(0)\neq 0$, which may be shown to be more natural than the former.

Let the system be modelled as a CARIMA model

$$A(q^{-1})Y(t)=B(q^{-1})\,U(t-k)+Z(t) \qquad (2.1)$$

where Y(t), U(t), Z(t) are the output, control, and disturbance respectively. $A(q^{-1})$ and $B(q^{-1})$ are polynomials in the unit backward-shift operator q^{-1}. Assumptions are made that the system time delay k is to be known, $A(0)=1$ and $B(0)\neq 0$, and the disturbance has stationary increments, i.e., it is modelled as

$$D(q^{-1})Z(t)=C(q^{-1})\xi(t) \qquad (2.2)$$

where $\xi(t)$ is a random process with zero-mean, $C(q^{-1})$ and $D(q^{-1})$ are polynomials such that $C(0)\neq 0$ and

$$D(q^{-1})=1-q^{-1} \qquad (2.3)$$

The system (2.1) can be expressed as

$$\bar{A}(q^{-1})Y(t)=\bar{B}(q^{-1})U(t-k)+C(q^{-1})\xi(t) \qquad (2.4)$$

where $\bar{A}(q^{-1})$ denotes $D(q^{-1})A(q^{-1})$, and this notation is used throughout.

Let the reference model be given by

$$P(q^{-1})Ym(t)=R(q^{-1})W(t-k) \qquad (2.5)$$

where Ym(t) is the output, W(t) the setpoint, and $P(q^{-1})$ and $R(q^{-1})$ are polynomials such that $P(0)\neq 0$ and $R(0)\neq 0$. In the sequel, it is assumed $R(q^{-1})=1$ for convenience.

The controller is to be chosen to minimize the performance index

$$J=E[V(t)^2|t] \qquad (2.6)$$

where the expectation is conditional on data acquired up to time t, and the signal V(t) is given by

$$\begin{aligned} V(t)&=P(q^{-1})[Y(t+k)-Ym(t+k)] \\ &=P(q^{-1})\,Y(t+k)-W(t) \\ &=\phi(t+k)-W(t) \end{aligned} \qquad (2.7)$$

The control law minimizing the performance index (2.6) is, under the assumption that $\xi(t)$ and W(t) are independent with each other, given by

$$\phi^*(t+k)=W(t) \qquad (2.8)$$

where $\phi^*(t+k)$ is an optimal prediction of $\phi(t+k)=P(q^{-1})Y(t+k)$.

3. A controller with integral action

3.1. A zero-gain predictor

It is a well-known fact that satisfactory estimation in the adaptive framework requires that the prediction error and data are uncorrelated. Without imposing a stochastic framework, it is not possible to use this fact directly. Therefore, the requirement is that the predictor structure should be such that both the prediction error and data have small average values.

Zero-mean data could be obtained by incorporating a $(1-q^{-1})$ term into a predictor. The predictor with such a property is called a zero-gain predictor, which blocks any constant component in the signals and the average values are removed. The prediction error

$\eta(t)=\phi(t)-\phi^*(t)$

also has a zero-average value, according to zero-mean $\xi(t)$. The predictor may be obtained by restricting the polynomials to be of a certain form as follows:

$$\gamma\varphi=1 \tag{3.1}$$

where γ and φ are definded by

$$C(q^{-1})=\gamma+\bar{C}o(q^{-1}), \qquad \gamma=C(1) \tag{3.2}$$

and

$$P(q^{-1})=\varphi+\bar{P}o(q^{-1}), \qquad \varphi=P(1) \tag{3.3}$$

respectively, and

$$Co(q^{-1})=\gamma_0+\gamma_1 q^{-1}\cdots,\ Po(q^{-1})=\varphi_0+\varphi_1 q^{-1}+\cdots.$$

It seems to be flexible for the choice of the polynomials $C(q^{-1})$ and $P(q^{-1})$ compared with the case of fixing $C(s)=P(s)=1$ at $s=0$ in Gawthrop[2]. Thus, this leads to a control system of more generalized structure. Moreover, because the "proper" condition is not required in the discrete-time system, the order of the polynomial $C(q^{-1})$ can be chosen to be of arbitrary one and this also contributes to establishment of more generalized structure.

Under the assumption, the auxiliary signal $\phi(t+k)$ in the signal(2.7) can be expressed as

$$\phi(t+k)=Y(t)/C(q^{-1})+\psi(t+k) \tag{3.4}$$

where the predictive form of $\psi(t+k)$ can be given by

$$\psi(t+k)=[\bar{G}_0(q^{-1})U(t)+\bar{F}_0(q^{-1})Y(t)]/C(q^{-1})+E_0(q^{-1})\xi(t+k) \tag{3.5}$$

from the polynomial identity

$$P(q^{-1})C(q^{-1})=\bar{A}(q^{-1})Eo(q^{-1})+q^{-k}F(q^{-1}) \tag{3.6}$$

and $Go(q^{-1})=Eo(q^{-1})B(q^{-1})$. Due to the observation of (3.6) at the steady-state, the polynomial $F(q^{-1})$ can be decomposed as

$$F(q^{-1})=1+\bar{F}o(q^{-1}) \tag{3.7}$$

With this relation, the identity (3.6) can be rewritten as

$$P(q^{-1})C(q^{-1})=q^{-k}+\bar{A}(q^{-1})Eo(q^{-1})+q^{-k}\bar{F}o(q^{-1}) \tag{3.8}$$

The predictor is now given by

$$\psi^*(t+k)=[\bar{G}o(q^{-1})U(t)+\bar{F}o(q^{-1})Y(t)]/C(q^{-1}) \tag{3.9}$$

and thus becomes a zero-gain predictor. The order of the polynomials $Go(q^{-1})$ and $Fo(q^{-1})$ are

$$\rho Go(q^{-1})=\rho B(q^{-1})+k-1$$
$$\rho Fo(q^{-1})=\max\{[\rho A(q^{-1})-1],\ [\rho P(q^{-1})+\rho C(q^{-1})-k-1]\}$$

respectively, where $\rho(\cdot)$ denotes the order of the polynomial $(\cdot)$.

3.2 A Controller with Integral action

The resulting control law from the predictor(3.9) can be given by

$$\phi^*(t+k)=Y(t)/C(q^{-1})+\psi^*(t+k)=W(t) \tag{3.10}$$

or

$$U(t)=\{[C(q^{-1})W(t)-Y(t)]-\bar{F}o(q^{-1})Y(t)\}/\bar{G}o(q^{-1}) \tag{3.11}$$

This control law can be rewritten as

$$U(t)=\frac{1}{Go(q^{-1})}\left[\frac{\gamma W(t)-Y(t)}{D(q^{-1})}+Co(q^{-1})W(t)-Fo(q^{-1})Y(t)\right] \tag{3.12}$$

where the polynomial $D(q^{-1})=(1-q^{-1})$ as in (2.3). Since the choice of $C(q^{-1})$ is arbitrary, i.e., $C(0)\neq 0$ and its order not restricted, all the gain factors γ and γ_i, $i=0,\cdots$, representing the setpoint integral, proportional, and derivative terms can be set to any values. This fact leads to the system of generalized structure, which may be important in typical industrial applications.

The closed-loop response to the setpoint with the control is described by

$$Y(t)=\frac{1}{P(q^{-1})}W(t-k)+\frac{Eo(q^{-1})}{P(q^{-1})}\xi(t) \tag{3.13}$$

and the model following property might be achieved. For the response to track the setpoint, instead of the reference output, the polynomial $R(q^{-1})$ in (2.5) should be specified by $R(1)=\varphi$. Due to direct pole-zero cancellation the control law can not deal with nonminimum-phase systems. This difficulty could be overcome by detuning the control law.

The controller discussed above is to be optimal. In an adaptive framework, the standard RLS algorithm[5] may be used to estimate the control parameters.

4. A detuned control law

As mentioned in the previous section, a model-reference method with the control(3.2) is not applicable to nonminimum-phase systems due to direct pole-zero cancellation. Furthermore, the method is sensitive to unmodelled dynamics or model order-mismatch. These problems could be alleviated by using a detuned version of the control law. Here, the detuned controller is introduced and a method of choosing the weights of the detuned term is also developed.

The cost function (2.6) is now given in terms of the auxiliary signal

$$V(t)=P(q^{-1})Y(t+k)+Q(q^{-1})U(t)-W(t) \tag{4.1}$$

where $Q(q^{-1})$ is given by

$$Q(q^{-1})=\overline{M}(q^{-1})/C(q^{-1}) \tag{4.2}$$

and

$$M(q^{-1})=m_0+m_1 q^{-1}$$

The polynomial $M(q^{-1})$ may be specified as inverse form of PI (Proportional+Integral) controller for smoothing the excessive control action. In the sequel, the polynomial is assumed to be constant m for convenience.

The cost function could be minimized by the control law.

$$\Phi^*(t+k)=\phi^*(t+k)+Q(q^{-1})U(t)=W(t) \tag{4.3}$$

A detuned version of the controller (3.12) is now expressed as

$$U(t)=\frac{1}{(Go(q^{-1})+m)}\left[\frac{\gamma W(t)-Y(t)}{D(q^{-1})}+Co(q^{-1})W(t)-Fo(q^{-1})Y(t)\right] \tag{4.4}$$

The closed-loop response to the setpoints with this control law may be written by

$$Y(t)=\frac{B(q^{-1})C(q^{-1})W(t-k)}{P(q^{-1})B(q^{-1})C(q^{-1})+\overline{M}(q^{-1})A(q^{-1})}+\frac{C(q^{-1})[B(q^{-1})Eo(q^{-1})+m]\xi(t)}{P(q^{-1})B(q^{-1})C(q^{-1})+\overline{M}(q^{-1})A(q^{-1})} \tag{4.5}$$

As expected, direct pole-zero cancellation might not arise here. And the clossed-loop characteristics of the system could be described by

$$P(q^{-1})B(q^{-1})C(q^{-1})+\overline{M}(q^{-1})A(q^{-1})=0$$

which is then shown to improve the stability of the closed-loop system. Therefore, a model-reference following property might be achieved.

It is still not clear how to choose the detuning weight, although the detuned control system has such advantages. A method which is capable of tuning the weight on-line is to be discussed in the sequel.

Now consider a transfer funtion

$$T(q^{-1})=[1/Q(q^{-1})][q^{-k}B(q^{-1})/A(q^{-1})] \tag{4.6}$$

which is the open-loop transfer function from $W(t)$ to $Y(t)$. Define

$$T'(q^{-1})=C(q^{-1})B(q^{-1})/mA(q^{-1}) \tag{4.7}$$

then it has a scaler gain, at the steady-state,

$$T'(1)=1/h \tag{4.8}$$

Thus it can also be written as

$$m = h\ C(1)B(1)/A(1) \quad (4.9)$$

Remark 1 : It is proposed that, in the sence of adaptation of m,

$$m(t) = \alpha(t)h\ |C(1)\hat{B}(1)/\hat{A}(1)| \quad (4.10)$$

where $\hat{A}(q^{-1})$ denotes the estimate of $A(q^{-1})$, and this notation is used throughout.

Remark 2 : In direct adaptive schemes, m(t) is chosen such that

$$m(t) = \alpha(t)h\ |\hat{G}(1)/\hat{F}(1)| \quad (4.11)$$

which is obtained from the relationship of the polynomial identities and the equation (4.10).

Remark 3 : The open-loop gain h is usually specified as unity and

$$\alpha(t) = |Uo(t)-U(t-1)|\ /\ \{1+|Uo(t)-U(t-1)|\} \quad (4.12)$$

where Uo(t) is an exact model-following control. $\alpha(t)$ and h determine the control activities, i. e., the large one leads to the small control, and viceversa.

5. Simulation results

To verify the performance of this control system, a system was simulated and compared with a GMV controller[5] under the same conditions.

The simulation scheme was given as follows:
A series of runs of 550 samples was executed. The noise $\xi(t)$ is assumed to be Gaussian. Step-wise setpoints W(t) with nonzero-mean were chosen to illustrate any possible offsets. The simulated system was chosen as

$$(1.0-q^{-1})Z(t) = \xi(t)$$

$$(1.0-1.2q^{-1}-0.11q^{-2})Y(t) = (0.5+1.0q^{-1})U(t-1)+Z(t),$$

and

$$(1.6-0.6\ q^{-1})Ym(t) = W(t-1).$$

Where the disturbances Z(t) acting on the plant were assumed to be zero at the first 250 samples. The plant was modelled as

$$(1.0+a_1\ q^{-1})Y(t) = (b_0+b_1\ q^{-1})U(t-1)+Z(t)$$

to examine the robustness of the control system against the unmodelled dynamics. Estimation of the control parameters was performed by the standard RLS estimator with an asymptotic sample length of 100 samples and zero-initial estimates except $g_0 = f_0 = 1$ for the numerical stability. Where the number of the control parameters to be estimated is three for the system in this peper and four for the GMV system. The detuning term with open-loop gain h=1 was included into the control law. In the GMV control law, the detuning term was specified to be of the form $D(q^{-1})m(t)$, where

$$m(t) = m\sum_{i=0}^{ng} g_i$$

where ng is the order of the polynomial $G(q^{-1})$, and g_i coefficients. Since the best choice of m might be dependent on the system to be examined, a trial and error approach should be required for obtaining satisfactory results of the given system. In this simulation study, m was chosen to be 0.5

Fig.1 is the simulation results of the GMV control method and Fig.2 is those of the control method in this paper. As shown in the figures, the GMV control algorithm, in the presence of nonzero-mean disturbances, leads to offsets or overshoots at the output of the system. Estimates of these control parameters may be unsettled when the disturance levels are changed significantly, In this case, the control system does not have effective intergral action tin the control loop and, thus, offsets may occur. Moreover, a small value of the detuning weight m(t) leads to overshoots at both the output and the input. On the other hand, the output of the control system in this paper is shown to trace well. This seems to be a result of satisfactory estimation of the control parameters nased on the zero-gain prediction approach and suitable choice of the detuning weight

6. Conclusion

To manipulate the unstable and inverse unstable systems with unmodelled dynamics in the presence of nonzero-mean load disturbances and setpoints, a class of adaptive control systems with integral action are proposed in this paper. In particular, concentrations are placed on the improvement of the control system proposed by Gawthrop[2] in continuous-time case to more general system. And the detuning weight is tuned recursively. Then the resulting systems are developed to be discrete-time versions, where specifications of the design parameters may be more flexible and, thus, the systems have generalized structure. Its performances have been verified on the simulation studies. A number of simulations, however, have shown that the method may fail for a certain system.

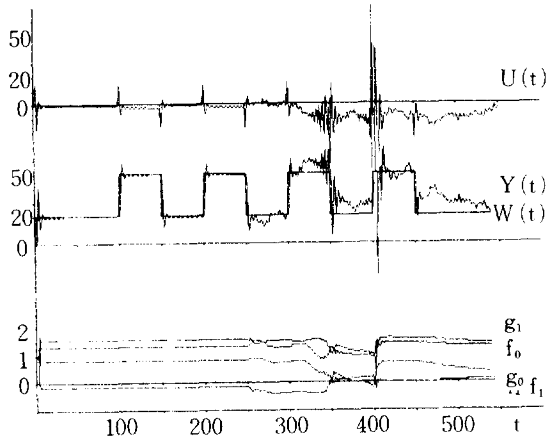

Fig. 1 Simulation results of the GMV control system

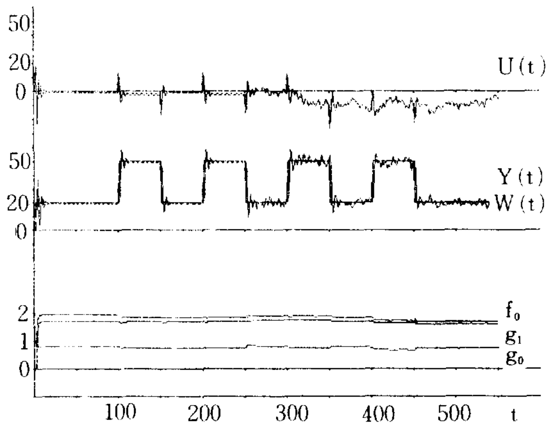

Fig. 2 Simulation results of the system in this paper

REFERENCES

1) Allidina, A.Y. and F.M.Hughes, "Self-tuning controller with integral action", Opt. Contr. Appl. Methods, Vol.3, pp. 335-362, 1982.
2) Gawthrop, P.J., "Self-tuning PID controllers : algorithms and implementation", IEEE Trans. Automat. Contr., Vol. AC-31, No.3, pp.201-209, 1986.
3) Scattolini, R., "A multivariable self-tuning controller with integral action", Automatica, Vol. 22, No,6, pp.619-627, 1986.
4) Goodwin, G.C., and D.Q.Mayne, "A continuous-time adaptive controller with integral action", IEEE Trans. Automat.Contr., Vol. AC-30, No. 8, pp. 793-799, 1985.
5) Clarke, D.W., "Self-tuning control of nonminimum-phase systems", Automatica, Vol. 20, No. 5, pp. 501-517, 1984.