15th Triennial World Congress, Barcelona, Spain

# NEW ROBUST CONTROL OF PWM POWER AMPLIFIER

**Kohji HIGUCHI* , Kazushi NAKANO*,**
**Kuniya ARAKI** and Fumiho CHINO****

**Dept.of Electronic Engineering,The University of Electro-Communications*
*1-5-1 Chofu-ga-oka,Chofu,Tokyo 182-8585,Japan*
***NF ELECTRONIC INSTRUMENTS*
*6-3-20 Tunashimahigashi,Kohoku-ku,Yokohama 223-8508,Japan*

Abstract :A method for designing a digital robust controller for PWM power amplifier is proposed.The present controller design is based on approximate 2-degree-of-freedom digital integral-type control systems. A digital signal processor(DSP) is implemented to the digital controller obtained here. It is shown from an experiment that a sufficiently robust digital control systems are realizable. *Copyright © 2002 IFAC*

Keywords: PWM power amplifier, Digital control, Approximate 2-degree-of-Freedom system, Robust control, Integral control.

## 1. INTRODUCTION

A Pulse Width Modulation (PWM) switching circuit is used for an electric-power conversion circuit, a LC low path filter is inserted between the conversion circuit and the load for noise removal, and a PWM power amplifier which constitutes feedback control systems so that the output voltage supplied to load might be proportional to an reference input is used as an amplifier itself or as a power supply (Sugimoto, 1988; Komatuzaki, 1991; Fukuda, 1992; Fukuda, 1993). If the characteristics and parameters of load are decided and there is little change in those, satisfactory performance is obtained in general. In many applications, however, load cannot be specified, i.e., its characteristics are changeable from capacitive to inductive. In addition, its amplitude is also sharply changed from the zero to the maximum rating. Usually, design conditions are changed for each load and then each controller is re-designed. Then, a so-called robust PWM power amplifier which can cover such an extensive load change and also direct-current power supply voltage change with one controller is needed. Authors proposed (Higuchi, 2000) previously the method for designing an analog controller for PWM power amplifiers which can attain such a demand, different from other methods (Aida, 1991). In this paper, the methodology of this design is applied to digital control systems, and a method of designing the digital controller for obtaining robust control systems with almost the same characteristics as the analog controller is proposed. A fundamental procedure of this design method is stated as follows: First, a controlled objective is described by discrete time systems with consideration of the input dead time, secondally the state feedback system which attains the ginen reference characteristics given is constituted, finally after approximating this state feedback system, a robust compensator is added and 2-degree-of-freedom digital robust systems are constituted. The existence of input dead time is mainly due to the conversion times of AD/DA and the computing time of DSP. The digital controller obtained here is actually realized by using the DSP. An experiment demonstrates that the proposed digital controller can satisfy all the specifications.

## 2. POWER AMPLIFIER

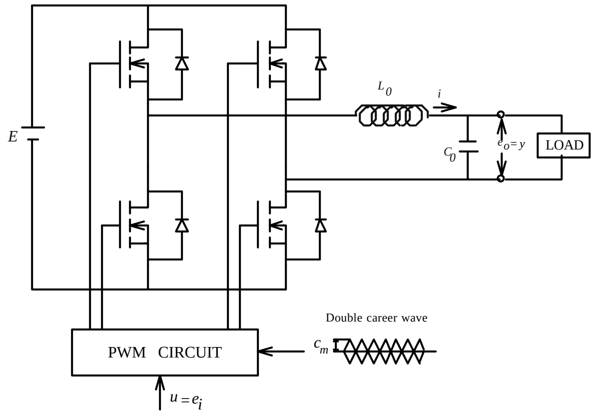


Fig. 1 PWM power amplifier

The power amplifier of composition as shown in Fig. 1 was manufactured. The triangular wave double career system was adopted as PWM switching signal generating part. Career frequency of a triangular wave is $100[kHz]$, and the amplitude $c_m$ is $10[V]$. The power amplification part is a full bridge type chopper circuit , and the voltage of direct-current power supply $E$ is $150[V]$. LC circuit is a filter for removing a career and a switching noise. This Values $L_0$ and $C_0$ of LC circuit are determined that a control system will become low sensitivity at the same time they reduce those noise.

If the frequency of input $u$ is smaller than the frequency of the career enough, the state equation at no load of PWM power amplifier of Fig. 1 can be expressed as follows from the state equalizing method (Fukuda,1993).

$$\begin{aligned} \dot{x} &= A_c x + B_c u \\ y &= Cx \end{aligned} \tag{1}$$

where

$$x = \begin{bmatrix} e_o \\ i \end{bmatrix} \quad A_c = \begin{bmatrix} 0 & 1/C_0 \\ -1/L_0 & -R_0/L_0 \end{bmatrix}$$

$$B_c = \begin{bmatrix} 0 \\ K_p/L_0 \end{bmatrix} \quad C = \begin{bmatrix} 1 & 0 \end{bmatrix}$$

$$u = e_i \quad y = e_0 \quad K_p = -E/c_m$$

When realizing a digital controller by DSP, the delay time exists until the output point of control input from the start point of sampling operation for the computing time and AD and DA conversion time. Sampling period is defined as $T$ and delay time is defined as $L(\leq T)$. This delay time $L$ is considered that it is equivalent to the input dead time which exists in the controlled object as shown in Fig. 2.

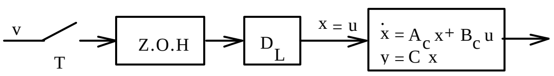


Fig.2 The controlled object with input dead time.

Then the state equation of the system of Fig. 2 is expressed as follows when one state is introduced.

$$\begin{aligned} x_d(k+1) &= A_d x_d(k) + B_d v(k) \\ y(k) &= C_d x_d(k) \end{aligned} \tag{2}$$

where

$$x_d = \begin{bmatrix} x(k) \\ x(k) \end{bmatrix} \qquad x(k) = u(k)$$

$$A_d = \begin{bmatrix} e^{A_c T} & e^{A_c (T-L)} \int_0^L e^{A_c t} B_c dt \\ 0 & 0 \end{bmatrix}$$

$$\approx \begin{bmatrix} 1 & \dfrac{T}{C_0} & 0 \\ -\dfrac{T}{L_0} & 1 - \dfrac{T \cdot R_0}{L_0} & \dfrac{K_p \cdot L}{L_0} \\ 0 & 0 & 0 \end{bmatrix}$$

$$B_d = \begin{bmatrix} \int_0^{T-L} e^{A_c t} B_c dt \\ 1 \end{bmatrix} \approx \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix} \quad C_d = \begin{bmatrix} C & 0 \end{bmatrix}$$

Now, the power amplifier with which it is satisfied of the following specifications is designed and manufactured by constituting a digital control system to PWM power amplifier (object) at no load. This specification is almost the same as that of the case where an analog controller is used (Higuchi,2000). For example, such specification is required of the power supply for a low frequency immunity test.

(1) The band width of the control system is about $2[KHz]$ to each load ,i.e., no-load, resistance load $(8.8 \leq RL < \infty[\Omega])$,capacity load $(0 \leq CL \leq 50[mF])$ and parallel load with resistance $(8.8 \leq RL < \infty[\Omega])$ and capacity $(0 \leq CL \leq 50[mF])$.

(2) With all the loads of (1), an over-shoots do not appear in a step responses.

(3) The specification of (1) and (2) is satisfied regardless of change of a large direct-current power supply.

The load change for controlled object and direct-current power supply change are considered as parameter changes for controlled object(4). Such parameter changes for controlled object can be replaced to equivalent disturbances as shown in Fig. 3 as also in a discrete time system. Moreover, if saturation arises to an input or the frequency of an input is not so small compared with the frequency of a career, the controlled object will change to a nonlinear system. Such characteristics changes can also be replaced to equivalent disturbances as shown in Fig. 3. In Fig.3 $B(2,1)$ is an element of $B = \int_0^T e^{A_c T} B_c dt$.

Therefore, what is necessary is just to constitute the control system which the pulse transfer functions from equivalent disturbance $q_u, q_{\bar{y}}$ and $q_y$ to output $y$ become as small as possible, in order to press down the influence of these parameter change, i.e. , load change, and direct-current power-supply change, i.e., to make it robust. Next, the easy designing method which can suppress the influence of such disturbance, with the target characteristic held is shown.

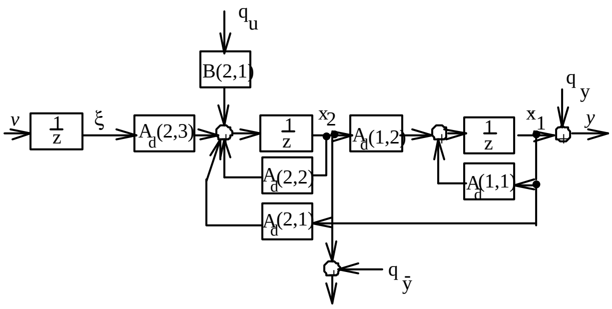

Fig.3 Load changes (parameter changes) and equivalent disturbances

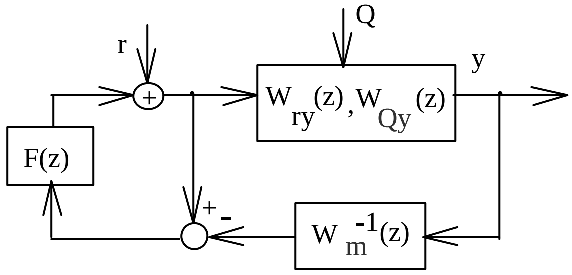


Fig.5 System which added the inverse system and filter.

3. THE DESIGN METHD OF APPROXIMATE 2-DEGREE-OF-FREEDOM DIGITAL INTEGRAL-TYPE CONTROL SYSTEM

First, the pulse transfer function between the reference input $r$ and the output $y$ is specified as follows:

$$W_{ry}(z)=\frac{(1+H_1)(1+H_2)(1+H_3)(z-nwd_0(1,1))(z-nwd_0(2,1))}{(1-nwd_0(1,1))(1-nwd_0(2,1))(z+H_1)(z+H_2)(z+H_3)} \tag{3}$$

Here it shall be specified that the relation $H_2$ and $H_1, H_3$ becomes $H_2 >> H_1, H_3$. Then $W_{ry}(z)$ can approximate by the following $W_m(z)$.

$$W_{ry}(z)\approx W_m(z)=\frac{1+H_2}{z+H_2} \tag{4}$$

Constitute the system shown in Fig.4 with the application of a state feedback to the controlled object of eq.(2), and we decide $F=[F(1,.1) \quad F(1,2) \quad F(1,3)]$ and $G$ so that $W_{ry}(z)$ becomes eq.(3).

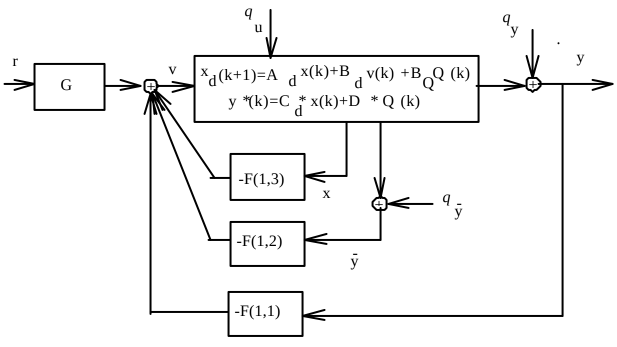

Fig. 4 State feedback system.

$Q$ is defined as $Q=\begin{bmatrix} q_u & q_{\bar{y}} & q_y \end{bmatrix}^T$ and the pulse transfer function between this equivalent disturbance $Q$ and the output $y$ is defined as $W_{Qy}(z)$. The system which added the inverse system and filter as shown in Fig.5 to the system of Fig.4 is constituted.

In Fig.5 the pulse transfer function of $F(s)$ is as follows.

$$F(z)=\frac{k_z}{z-1+k_z} \tag{5}$$

Here the transfer functions of between $r$ and $y$, $Q$ and $y$ of the system of Fig. 5 are as follows.

$$y=\frac{(1+H_1)(1+H_2)(1+H_3)(z-nwd_0(1,1))(z-nwd_0(2,1))}{(1-nwd_0(1,1))(1-nwd_0(2,1))(z+H_1)(z+H_2)(z+H_3)}$$
$$\times\begin{pmatrix} 1+\dfrac{k_z}{z-1+k_z} \\ \left(-1+\dfrac{(z-nwd_0(1,1))(z-nwd_0(2,1))(1+H_1)(1+H_3)}{(1-nwd_0(1,1))(1-nwd_0(2,1))(z+H_1)(z+H_3)}\right)\end{pmatrix}^{-1} r \tag{6}$$

$$y=\frac{z-1}{z-1+k_z}$$
$$\times\begin{pmatrix} \dfrac{z-1}{z-1+k_z}-\dfrac{k_z}{z-1+k_z} \\ \times\left(\dfrac{(z-nwd_0(1,1))(z-nwd_0(2,1))(1+H_1)(1+H_3)}{(1-nwd_0(1,1))(1-nwd_0(2,1))(z+H_1)(z+H_3)}\right)\end{pmatrix}^{-1} W_{Qy}(s)Q \tag{7}$$

Here since $(1+H_1)/(z+H_1)\approx 1$, $(1+H_1)/(z+H_1)\approx 1$, the equations (6) and (7) are as follows .:

$$y\approx\frac{1+H_2}{z+H_2}r \tag{8}$$

$$y\approx\frac{z-1}{z-1+k_z}W_{Qy}(z)Q \tag{9}$$

From the equation (8) and (9) it turns out that the characteristic from $r$ to $y$ can be specified with $H_2$ and the characteristic from $Q$ to $y$ can be independently specified with $k_z$. That is, the system of Fig.5 is of an approximate 2-degree-of-freedom, and its sensitivity against disturbance, i.e., load change becomes lower with the increase of $k_z$. Now, if equivalent conversion of the controller of Fig.5 is carried out, the approximate 2-degree-of-freedom digital integral type control system as shown in Fig.6 will be obtained. In Fig.6 $k_1, k_2, k_3$ and $k_i$ are as follows:

$$k_1=k_z(1/(1+H_2))G+F(1,1)$$

$$k_2 = F(1,2) \qquad k_3 = F(1,3)$$
$$k_i = k_z G \tag{10}$$

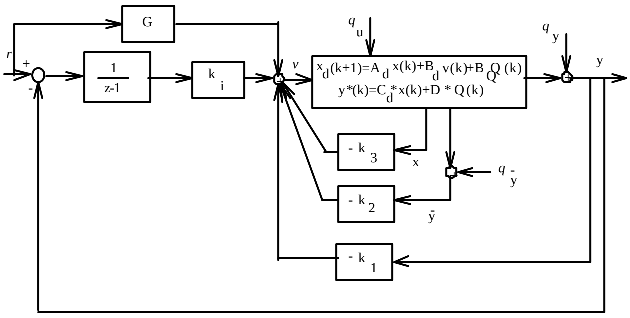


Fig.6 Approximate 2-degree-of-freedom digital integral type control system

## 4. DESIGN OF CONTROLLER AND EXPERIMENTS

$F\ (1,1)$ is the function of $C_0$. Therefore $k_z$ becomes the function of $C_0$ when the parameter $k_1$ of eq.(10) is beforehand set up so that it may become as small as possible in order to decrease voltage feedback noise. Moreover, the approximate value of the gain crossover frequency $w_c$ is the function of $L_0$, $C_0$, $T$, and $R_0$, if these are set up, $w_c$ will be decided and a phase margine $P_m$ becomes the function of $k_z$. Design procedure is as follows from these.

1. $L_0$ and $R_0$ value is set as the same value as the case of an analog controller(Higuchi,2000).
2. Sample period $T$ is set up suitably.
3. $H_1$, $H_2$, and $H_3$ of eq.(3) are specified so that it satisfies specification (1).
4. It sets up with $k_1 \approx -1$ in eq.(10), and $k_z - C_0$ curve is drawn.
5. $C_0$ is set up suitably, the approximate value of the gain crossover frequency $w_c$ is calculated, and $P_m - k_z$ curve is drawn.
6. $k_z$ is calculated from $k_z - C_0$ curve.
7. $P_m$ is calculated from $P_m - k_z$ curve and it checks whether $P_m$ is large enough.
8. It checks whether all specifications are fulfilled by the simulation by deciding the parameter (10) of a controller from these $C_0$ and $k_z$.
9. When not satisfying all specifications, it returns to 2. and even 8. is repeated.

The transfer function of the controlled object is as follows.:

$$G_p(s) = \frac{K_p}{L_0 C_0 s^2 + R_0 C_o s + 1} \tag{11}$$

$L_0$ and $R_0$ are set up as $L_0 = 180[mH]$ and $R_0 = 1.24\,[\Omega]$. The input delay time by computing time etc. presupposes that it is almost equal to sampling period T. Let the sampling period $T$ be a $12[ms]$. $H_1$, $H_2$ and $H_3$ are specified as follows:

$$H_1 = -0.48 \quad H_2 = -0.891 \quad H_3 = -0.2 \tag{12}$$

The $k_z - C_0$ curve is derived from the following equation.

$$k_1 = k_z(1/(1+H_2))G + F(1,1) \approx -1 \tag{13}$$

The draw is shown in Fig.7.

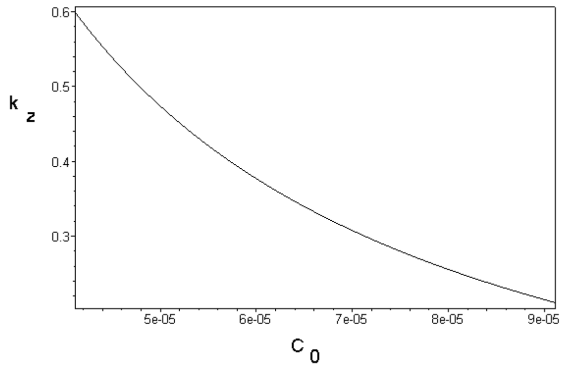


Fig.7 $k_z - C_0$ curve

$C_0$ is set up as $C_0 = 51.175[mF]$. Then $w_c$ is decided as $w_c = 41677[rad/s]$ and $P_m - k_z$ curve is drawn as shown in Fig.8.

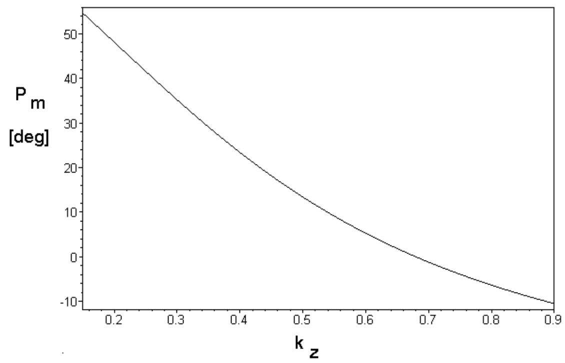


Fig.8 $P_m - k_z$ curve

From $k_z - C_0$ curve, when $C_0$ is $51.175[mF]$, $k_z$ is $0.48$. Then $P_m$ is $17.2[\deg]$ from $P_m - k_z$ curve. Then $F$ and $G$ are decided as

$$F = [-0.113 \ \ -0.42435 \ \ 0.33114]$$
$$G = -0.20174 \tag{14}$$

The parameters of the digital controller are decided as

$$k_1 = -0.96437 \quad k_2 = -0.42435$$
$$k_3 = 0.33114 \quad k_i = -1.0486 \tag{15}$$

The simulation results of the output voltage $y = e_o$, the input voltage $u = e_i$ and the current $i$ at no load, resistance load $(RL)$, capacity load $(CL)$ are shown in Fig.9, and it turns out that all specifications are satisfied.

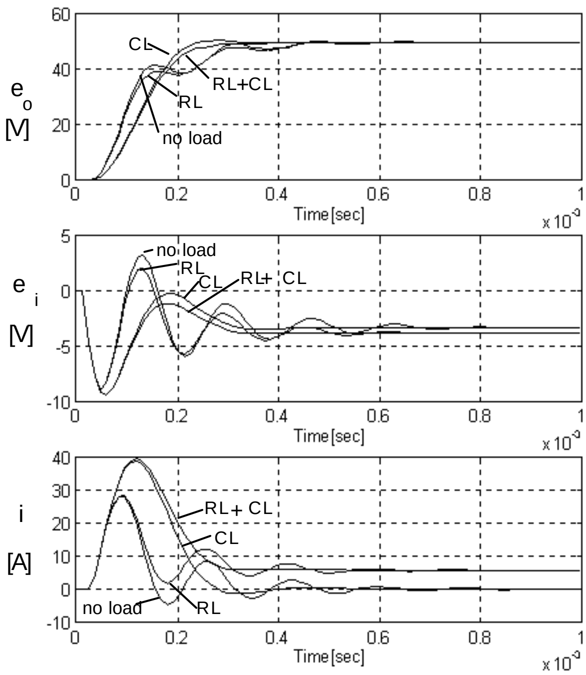


Fig.9 Simulation result at various loads, where $RL = 8.8[\Omega]$ and $CL = 50[mF]$.

TMS320C31 of TI is used for DSP which realizes the digital controller of Fig. 6. ADS7800 and DAC813 of BB are used for AD and a DA converter , respectively. A voltage amplifier with steady-state gain $g$ is realized. Then $k_i$ in Fig.6 becomes $k_i = k_z Gg$. That is, the gain $1/g$ will go into a major feedback loop. The experimental results when realizing the digital controller with the parameter of eq.(14) by DSP, and connecting with the controlled object of Fig. 1 are shown in Figs. 10, 11, and 12. It turns out that the almost same responses as the simulation result of Fig. 8 is carried out, and specifications are satisfied from these.

The gain characteristics between the reference input $r$ and the output $y = e_o$ at the experiment at no-load is shown in Fig.13. It turns out that the band width is actually about $2[kHz]$.

In the low frequency immunity test etc. of the apparatus which uses AC sine wave as an input power supply, the power supply which can simulate 90 degrees phase instant-cutting-off and a 90 degrees phase injection correctly is required. The experiment result of the response of output $y = e_o$ at no-load when giving such target instruction $r$ is shown in Fig. 14. When the band width of a control system is about 2 [KHz] and there is no over-shoot, it turns out that the output follows target instructions correctly.

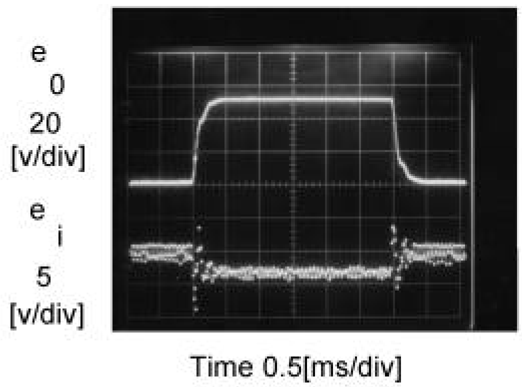


Fig.10 Experimental result of step response at no-load

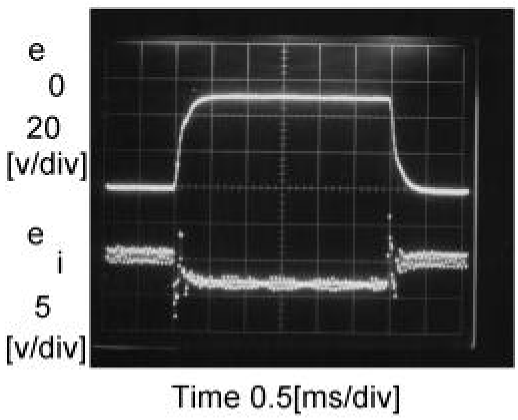


Fig.11 Experimental result of step response at resistance load ($RL = 8.8[\Omega]$)

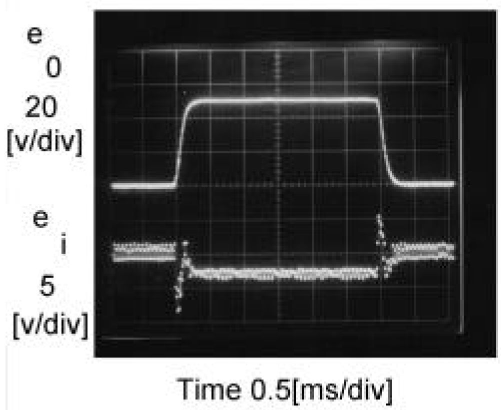


Fig.12 Experimental result of step response at capacity load ($CL = 50[mF]$)

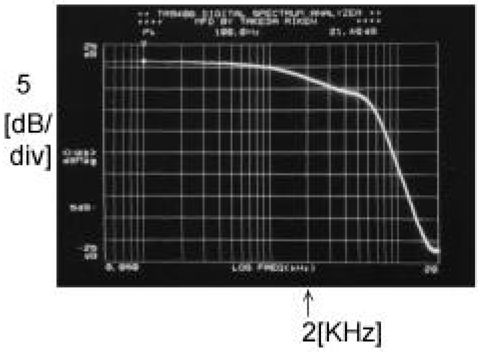


Fig.13 The gain characteristics between $r$ and $y = e_o$ at the experiment at no-load

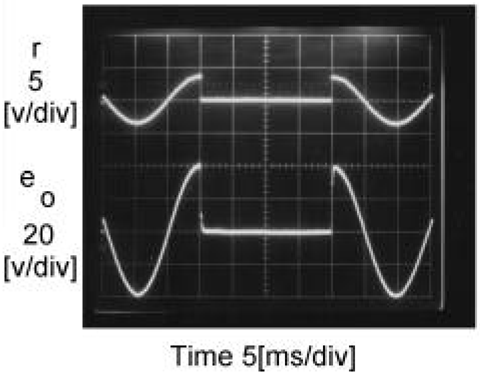


Fig.14 The experiment result of the response of $y = e_o$ at no-load when giving target instruction $r$

## 5. CONCLUSION

In this paper, the method was proposed for designing the controller with consideration of the input dead time of PWM power amplifier so as to attain the robustness against an extensive load change and a large direct-current power-supply change. The digital controller obtained by this design method is realized by using a DSP which was implemented to the controlled objective (PWM power amplifier which consists of a PWM signal generating part, an electric power conversion part, and a LC filter). It is shown from an experiment that sufficiently robust digital control systems are realizable. When a digital controller was designed by the simulation without taking the input dead time into consideration and implemented by a DSP, the output voltage has oscillated. This fact demonstrated the usefulness of our method which the input dead time was taken into consideration.

## REFERENCE

Sugimoto, H., S. Morimoto and M. Yano (1988). A HIGH PERFORMANCE CONTROL METHOD OF A VOLTAGE-TYPE PWM CONVERTER, *IEEE PESC'88 RECORD*, 360/368.

Komatuzaki, Y.(1991). A design for Switching Power Supply Voltage Controller with Robust Transient Response against Wide Load Variation, *TECHNICAL REPORT OF IEICE*, PE 91-12, 37/44 (*in Japanese*).

Fukuda, H., H. Yonemori , and M. Nakaoka (1992). Analysis and Design of High-Frequency Carrier PWM Power amplified Conversion System for Magnetic-field Current Tracking Control, *TECHNICAL REPORT OF IEICE*, PE 92-37, 37/44 ( *in Japanese)*.

Fukuda, H. and M. Nakaoka (1993). State-Vector Feedback Controlled-based 100kHz Carrier PWM Power Conditioning Amplifier and Its High-Precision Current-Tracking Scheme, *IEEE IECON'93*, 1105/1110.

Higuchi, K., K. Nakano, K. Araki and F. Chino (2000). The Robust Design of PWM Power Amplifier by the Approximate 2-Degree-of-Freedom Integral Type Servo System', *IEEE IECON-2000*, 2297/2302 .

Aida, K., and T. Kitamori (1991). Design of a PI-Type State Feedback Optimal Servo System, *Int. J. of Control*, 52-3, 613/624.