Robust Digital Control for an LLC Current-Resonant DC-DC Converter

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ABSTRACT

If a pulse frequency and a load resistance of an LLC current-resonant DC-DC converter are changed, the dynamic characteristics are varied greatly, that is, the LLC current-resonant DC-DC converter has non-linear characteristics. In many applications of DC-DC converters, loads cannot be specified in advance, and they will be changed suddenly from no loads to full loads. A DC-DC converter system used a conventional single controller cannot be adapted to change dynamics and it occurs large output voltage variation. In this paper, a robust digital controller for suppress the change of step response characteristics and variation of output voltage in the load sudden change is proposed. Experimental studies using a micro-processor for the controller demonstrate that this type of digital controller is effective to suppress the variation.

Keywords: LLC Current-Resonant, DC-DC Converter, Approximate 2DOF, Digital Robust Control, Micro-Processor

1. INTRODUCTION

In many applications of DC-DC converters, loads cannot be specified in advance, i.e., their amplitudes are suddenly changed from the zero to the maximum rating. In an LLC current-resonant DC-DC converter, if a pulse frequency and a load resistance are changed, the dynamic characteristics are varied greatly, that is, the LLC current-resonant DC-DC converter has non-linear characteristics [1-5]. Usually, a controller of the DC-DC converter is designed to an approximated linear controlled object at one operating point. In a non-linear DC-DC converter system, it is not enough for design of controller considering only one operation point. As a technique to improve dynamic performance, a gain-scheduled control method will be considered. However, this method needs to switch many controllers designed at many operation points. So it requires a complicated control routine when controllers are implemented on a micro-processor. Then, the controller which can cover sudden load changes and dynamic characteristics changes with only one controller is needed. A conventional control is performed in the analog control [6-7]. However, it is difficult to retain sufficient robustness of the DC-DC converters by these technique. The robust control method using approximate 2-degree-of-freedom (2DOF) for improving start-up characteristics and load sudden change characteristics of a boost DC-DC converters has been proposed [8-9]. In this paper, the method using approximate 2DOF is applied to the current-resonant DC-DC converter. The DC-DC converter is a non-linear system and the models are changed at each operation point. The design method of the approximate 2DOF controller which can cope with the non-linear system or changing of model with one controller is proposed. It is shown that the procedure of design becomes easy, the controller topology also becomes simple from the conventional one?and the control system becomes more robust than the conventional methods. This controller is actually implemented on the micro-processor and is connected to the LLC current resonant DC-DC converter. Experimental studies demonstrate that the digital controller designed by proposed method attains the good performance and is useful.

2. LLC CURRENT-RESONANT DC-DC CONVERTER

2.1 State-space model of the DC-DC converter

The LLC current-resonant DC-DC converter is shown in Fig. 1. In Fig. 1, V_{in} =24[V], C_1=0.33[μF], L_r =0.44[μH], L_m =3.5[μH], n=1.11, C_f=480[μF] and R_L =25[Ω]. I_{out} is an output current, and v_o is an output voltage. RX62T is a micro-computer. V_{ref} is a reference input voltage, u_{freq} is a control input, and P_{freq} is a switching pulse frequency. A small signal

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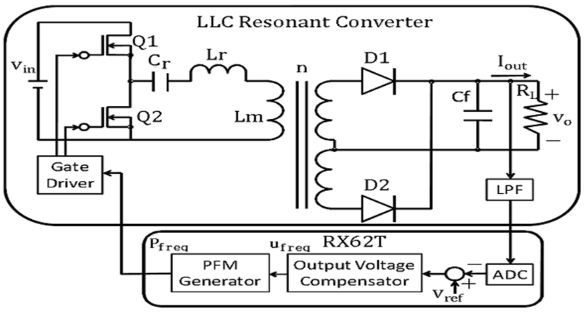

Fig.1: *LLC current-resonant DC-DC converter.*

model of the LLC current-resonant DC-DC converter is described by the third order model as follows:

$$\frac{\Delta V_o(s)}{\Delta U_{freq}} = G_{PFM}\frac{G_{DC}}{\left(1+\frac{s}{k_c\omega_f}\right)\left(1+s\left(\frac{kQ_0}{\omega_0}\right)+\left(\frac{1}{\omega_0}\right)s^2\right)} \quad (1)$$

Here, $\Delta V_o = V_o - V_s$ and $\Delta U_{freq} = U_{freq} - U_{sfreq}$, where V_s is V_o, and U_{sfreq} is U_{freq} at some operating point. G_{DC} is a rate of change of DC gain at a switching operating point, that is, $G_{DC} = k_pV_{in}$ and k_p is changed depending on the switching operating point. G_{PFM} is a conversion factor between u_{freq} and $1/P_{freq}$, k is a correction factor, ω_f is a reciprocal of time constant by R_L and C_f, Q_0 is the resonant frequency by L_r and C_r, and Q0 is a quality factor of the resonant circuit. As ω_0 is very high frequency and the gain value is very small as shown in the blue line of Fig. 2, this model is approximated by the first order model as follows:

$$\frac{\Delta V_o(s)}{U_{freq}(s)} \approx G_{PFM}\frac{G_{DC}}{\left(1+\frac{s}{k_c\omega_f}\right)} \quad (2)$$

This frequency characteristics is shown in the green line of Fig.2.

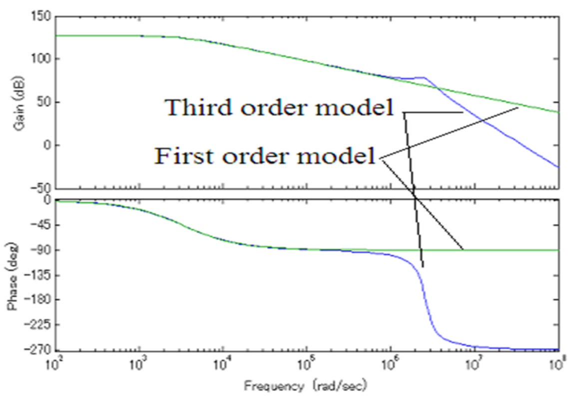

Fig.2: *The frequency characteristics of the third order model and the first order model.*

The state-space model of the DC-DC converter is derived from eq. (2) as follows:

$$\begin{aligned}\dot{x} &= Ax + Bu\\ y &= Cx\end{aligned} \quad (3)$$

where,

$$A = [-k\omega_f] \quad B = [G_{DC}G_{PFC}k\omega_f]$$
$$C = [1] \quad y = x = \Delta V_o \quad u = \Delta U_{freq}$$

2.2 Discretization

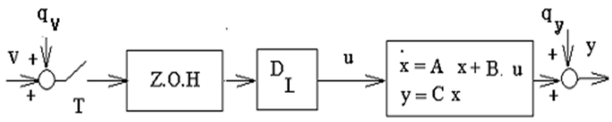

Fig.3: *Controlled object with input delay time L< T.*

When realizing a digital controller by the microcomputer, a delay time exists between the starting time of the sampling operation and the outputting time of the control signal due to the calculation and AD/DA conversion time and the computing time. This delay time L(< T : sampling time) is considered to be equivalent to the input dead time which exists in the controlled object as shown in Fig. 2. Then the discrete-time model of the system in Fig. 3 will be obtained as follows:

$$\begin{aligned}x_d(k+1) &= A_dx_d(k) + B_dv(k) + B_dq_v(k)\\ y(k) &= C_dx_d(k) + q_y(k)\end{aligned} \quad (4)$$

where,

$$x_d(k) = \begin{bmatrix} x(k)\\ z(k)\end{bmatrix} \quad B_d = \begin{bmatrix} \int_0^{T-L} Bd\tau\\ 1\end{bmatrix}$$
$$A_d = \begin{bmatrix} e^{AT} & \int_{T-L}^{T} e^{A\tau} Bd\tau\\ 0 & 0\end{bmatrix} \quad C_d = [1 \ \ 0]$$

From eqs. (3) and (4), each parameter of the converter depends on R_L and G_{DC}. Therefore, the small signal model of the converter at the operating point will be changed depending on R_L and G_{DC}. The changes of R_L and G_{DC} are considered as parameter changes in eq. (2) and (3). Such parameter changes can be transformed to equivalent disturbances q_v and q_y as shown in Fig. 3. Therefore, what is necessary is just to constitute a control system whose pulse transfer functions from equivalent disturbances q_v and q_y to the output y become as small as possible in their amplitudes, in order to robustize or suppress the influence of these parameter changes. The approximate 2DOF digital controller is designed for robust control.

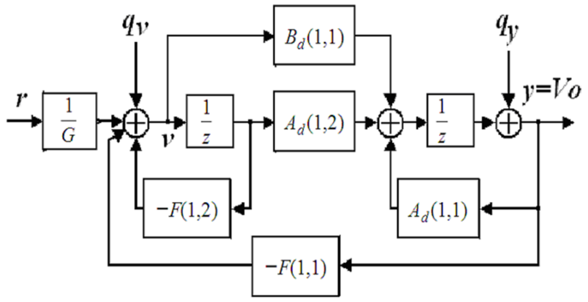

Fig.4: *Model matching system using state feedback and feedforward.*

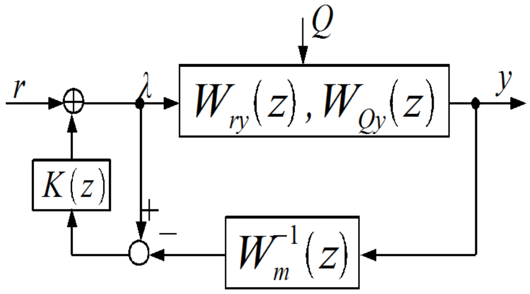

Fig.5: *System reconstituted with inverse system and filter.*

3. DESIGN METHOD OF APPROXIMATE 2DOF DIGITAL CONTROLLER

To the system of eq. (4), the model matching control system is constituted using the state feedback and feed forward as shown in Fig. 4.

In Fig. 4, $F = [F(1,1)F(1,2)]$ and G are the state feedback and feed forward parameters, respectively. In the system of Fig. 4, the transfer function from the reference input r to the output y is specified as follows:

$$W_{ry}(z) = \frac{(1+H_1)(1+H_2)(1-n_1)}{(z+H_1)(z+H_2)(z-n_1)} \tag{5}$$

n_1 is a zeros for the discrete-time controlled object. It shall be specified that the relation of H_1 and H_2 becomes $|H_1| >> |H_2|$. Then W_{ry} can be approximated to the following first-order discrete model:

$$W_{ry}(z) = W_m(z) = \frac{1+H_1}{z+H_1} \tag{6}$$

A system added an inverse system and a filter to the system in Fig. 4 is constituted as shown in Fig. 5. In Fig. 5, the transfer function $K(z)$ is as follows:

$$K(z) = \frac{k_z}{z-1+k} \tag{7}$$

And, the transfer function $W_{Qy}(z)$ between the equivalent disturbance $Q = [q_v \;\; q_y]^T$ to y of the system in Fig.4 is defined as

$$W_{Qy}(z) = \lfloor W_{q_v y}(z) \;\; W_{q_y y(z)} \rfloor \tag{8}$$

If $|H_1| >> |H_2|$, the transfer functions between $r-y$, $q_v - y$ and $q_y - y4ofthesysteminFig.5aregivenby$

$$y \approx \frac{1+H_1}{z+H_1} r \tag{9}$$

$$y \approx \frac{z-1}{z-1+k_z} W_{Qy}(z) Q \tag{10}$$

From eqs. (8), (9), it turns out that the characteristics from r to y can be specified with H1 and the characteristics from q_v and q_y to y can be independently specified with k_z. That is the system in Fig. 5 is an approximate two-degrees-of-freedom system. If the gain characteristics of the transfer function between Q and y is small, even if the disturbance enters to system, the characteristics of the transfer function between r and y does not change. Then the approximate 2DOF system is a robust system. What is necessary is just to enlarge k_z as much as possible in the range of $0 < kz < 1$ for making the gain characteristics of the transfer function between Q and y small in the wide frequency range.

If the equivalent conversion of the controller in Fig. 5, we obtain the approximate 2DOF digital integral type control system shown in Fig. 6. The topology of the controller is very simple.

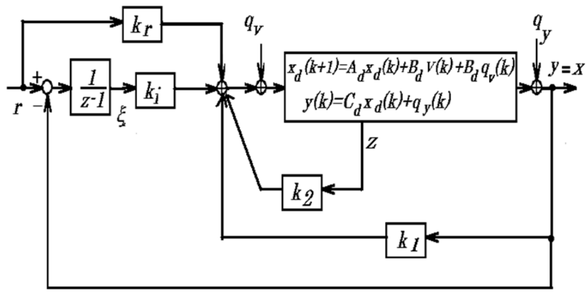

Fig.6: *Approximate 2DOF digital integral type control system.*

In Fig. 6, the parameters of the controller are as follows:

$$\begin{aligned} k_1 = -F(1,1) \quad & -K_z G/(1+H_1) \quad k_2 = -F(1,2) \\ k_1 = k_2 G \quad & \quad k_r = G \end{aligned} \tag{11}$$

4. EXPERIMENTAL STUDIES

The DC gain characteristics between $U_{freq}-V_o$ are shown in Fig. 7. It turns out that it becomes nonlinear depending on load resistances. The nominal model is decided by making the point Op (V_o=12[V], U_{freq}=190[kHz], R_L=6[Ω]) in Fig. 7 to the operating point. Then $G_{DC}G_{PFM}$ of eq. (1) is determined as 0.055. Moreover, the small step responses when

ΔU_{freq} was changed from 190 (the operating point in Fig. 7) to 209 are shown Fig. 8.The response is changing depending on load resistances. Moreover, it turns out that the real responses differ from the responses of the first-order models of eq. (1) considerably when R_L is more than 25[Ω]. From the response at R_L=6[Ω] in Fig. 8, $1/k\omega_f$ was decided as 260e-6.Converting eq. (1) to the discrete-time system (3) with T=10 [μs] and L=0.999T, the digital controller is designed.

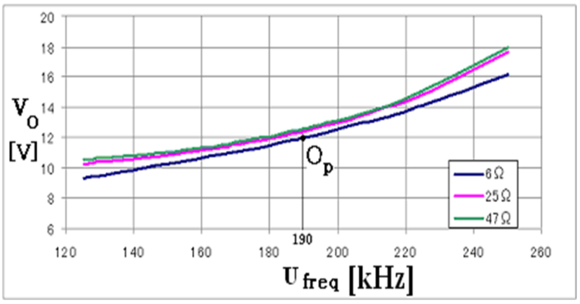

Fig.7: *DC gain between U_{freq} and V_o.*

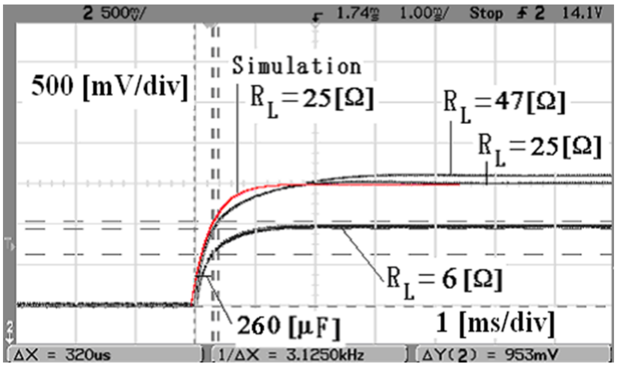

Fig.8: *Small step responses at the operating point Op.*

First of all, H_1 is determined as follows so that the step response of the closed loop system becomes quicker than that of the controlled object.

$$H_1 = -0.94 \tag{12}$$

Next, from $|H_1| >> |H_2|$, H_2 are determined as follows:

$$H_2 = -0.1 \tag{13}$$

Then F and G become as

$$F = [10.15 \ \ -0.1206], \quad G = 30.40 \tag{14}$$

Next, in order to make the approximate 2DOF system more robust, it is better to set up k_z as large as possible. However, k_z must be decided that the moving poles p_1 and p_2 do not approach near H_1. Then, from the root locus of Fig.9, k_z is determined to be 0.3.

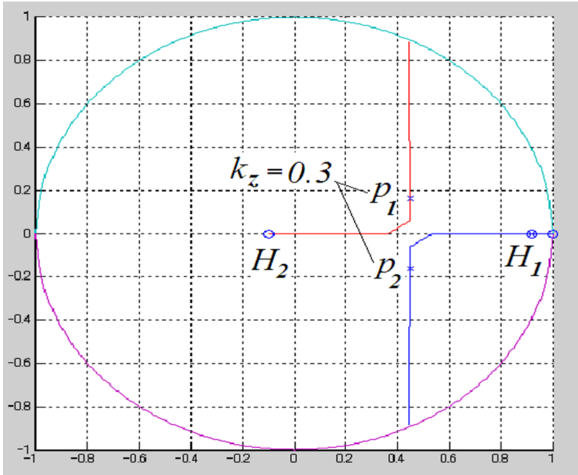

Fig.9: *Root locus.*

Then, from eq. (10), the parameters of the proposed controller become as

$$\begin{array}{ll} k_1 = -162.1, & k_2 = 0.1206 \\ k_i = 9.108 & k_r = 30.40 \end{array} \tag{15}$$

By the way, the transfer function of the PI controller is as follows:

$$G_{PI}(z) = K_p + \frac{K_I}{z-1} \tag{16}$$

The parameters of the PI controller are determined as

$$K_p = 50 \qquad K_I = 1.5 \tag{17}$$

We manufactured the current-resonant DC-DC converter. Experimental setup is shown in Fig. 10. In this experiment, the micro-processor RX62T by Renesas Electronics Corp. is used. The Renesas RX62T is a high-performance microcontroller with a maximum operating frequency of 100MHz and a operation performance of 165[MIPS]. They are equipped with PWM timers, high-speed 12-bit A/D converter, and 10-bit A/D converter.

The simulation result of step response at load R_L =6[Ω] using the proposed controller is shown in Fig. 11. This response is almost the same as the response of the first-order delay system with the dominant pole H1. The experimental results of step responses at load RL =6[Ω] and R_L =25[Ω] using the proposed controllers are shown in Fig. 12 and Fig.13, respectively. From Fig.11 and Fig.12, it turns out that the experimental step responses are almost the same as the simulation one. And comparing Fig.12 with Fig.13, it turns out that the experimental step responses are almost the same. Even if the loads are changed, the step responses are almost the same and are maintaining the responses by the dominant pole H_1=0.94. This shows that the approximate 2DOF system is robust enough by setting up of k_z=0.3.

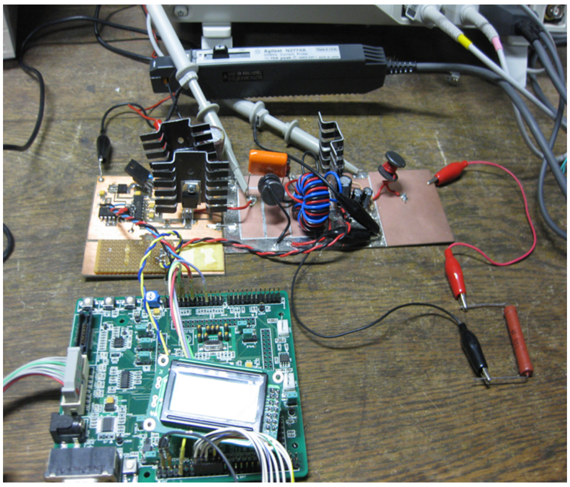

Fig.10: *Experimental setup.*

The simulation result at load sudden change (R_L:6↔25[Ω]) is shown in Fig. 14. The experiment result at load sudden change (R_L:6↔ 25[Ω]) is shown in Fig. 15. It turns out that the experimental step response is almost the same as the simulation one and the output voltage regulation is suppressed to about 50 [mV]. This voltage regulation is smaller than the reference [6-7]. The experiment result at load sudden change (R_L:6↔25[Ω]) when the input voltage is increased by 10 [%] (V_{in} = 26.4 [V]) is shown in Fig. 16. It turns out that the experimental step response is almost the same as the experimental one with V_{in} = 24 [V]. V_{in} is included in G_{DC} of eq.(3). So the change of V_{in} is the one of the parameter change of the controlled object, that is, the parameter change is included in the equivalent disturbance Q. Since the influence by change of the input voltage is suppressed like the change of load resistance, the output voltage regulation is hardly different from the one at the change of only load resistance. That is, the proposed control system is robust enough. In reference [7], when the input voltage is changed, the output voltage regulations differ greatly. So the control system of reference [7] is not robust.

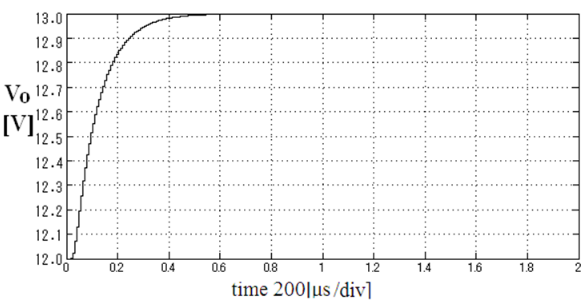

Fig.11: *Simulation result of step response at R_L= 6(Ω)using the proposed controller, V_{in}=24[V].*

The simulation result of step response at load R_L

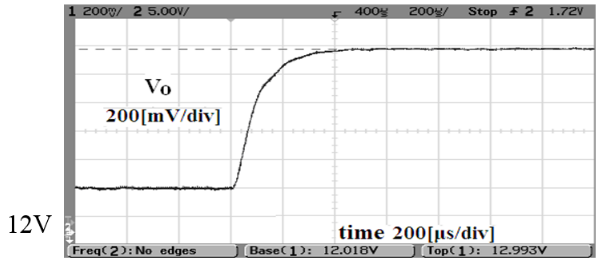

Fig.12: *Experimental result of step response at R_L=6 (Ω)using the proposed controller, V_{in}=24[V].*

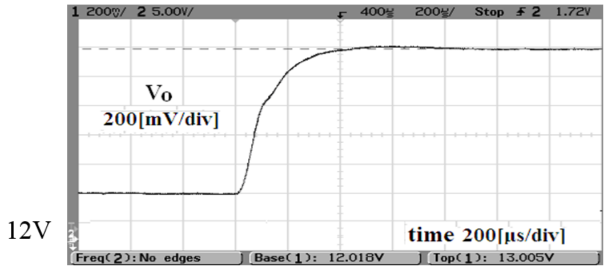

Fig.13: *Experimental result of step response R_L=25(Ω) using the proposed controller, V_{in}=24[V].*

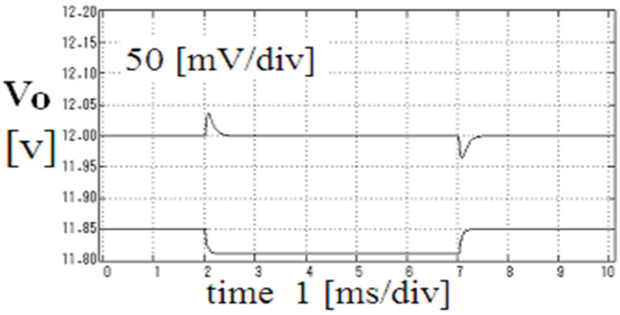

Fig.14: *Simulation result of sudden load change R_L : 6 ↔ 25(Ω)using the proposed controller, V_{in}=24[V].*

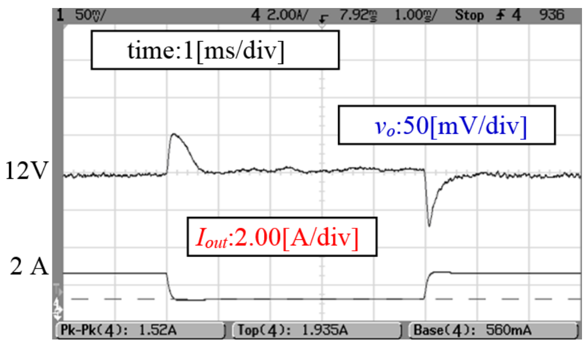

Fig.15: *Experimental results of sudden load change R_L : 6 ↔ 25(Ω) using the proposed controller, V_{in}=24[V].*

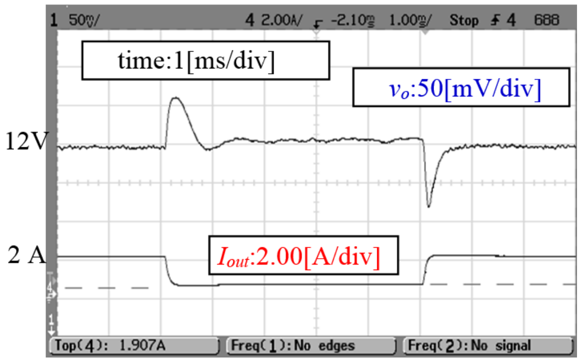

Fig.16: *Experimental results of sudden load change* R_L : 6 $\leftrightarrow$ 25(Ω) *using the proposed controller,* V_{in}=26.4[V].

=6[Ω] using the PI controller is shown in Fig. 17. The experimental results of step responses at load R_L =6[Ω] and R_L =25[Ω] using the PI controllers are shown in Fig. 18 and Fig.19, respectively. From Fig.17 and Fig.18, it turns out that the experimental step responses are different. The PI controller cannot fully eliminate the influence of nonlinearity. And comparing Fig.18 with Fig.19, it turns out that when the loads are changed, the experimental step responses are different. That is, the system using the PI controller is not robust.

The experimental result at load sudden change used a usual PI controller is shown in Fig. 12. The output voltage variation in sudden load change is about 100[mV], and the recovery time is longer than the one in Fig. 15. This result is almost the same as the result of the conventional method of reference [6-7]. From these results, the system using the PI controller or the conventional methods cannot attain good regulations. As a result, it turns out that the proposed controller is effective practically.

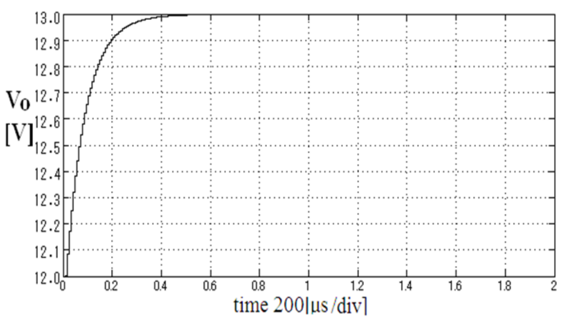

Fig.17: *Simulation result of step response at* R_L= 6(Ω)*using the PI controller,* V_{in}=24[V].

5. CONCLUSIONS

In this paper, the concept of the controller of the LLC current-resonant DC-DC converter to attain good robustness was given. The proposed digital controller was implemented on the micro-processor. The LLC current resonant DC-DC converter built-in this micro-processor was manufactured. It was shown from experiments that the proposed approximate 2DOF digital controller can suppress the variations of the output voltages in sudden load changes even if the input voltage was changed. This fact demonstrates the usefulness and practicality of our proposed method.

A future subject is to realize the robust control of the system in which PFC and LLC converter were combined.

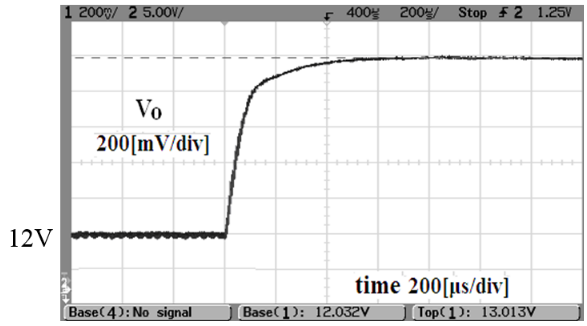

Fig.18: *Experimental result of step response at* R_L= 6(Ω) *using the PI controller,* V_{in}=24[V].

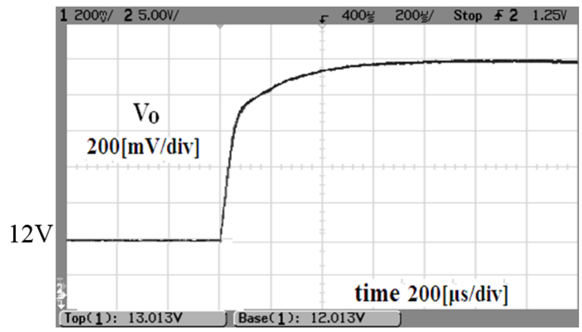

Fig.19: *Experimental result of step response at* R_L=25(Ω) *using the PI controller,* V_{in}=24[V].

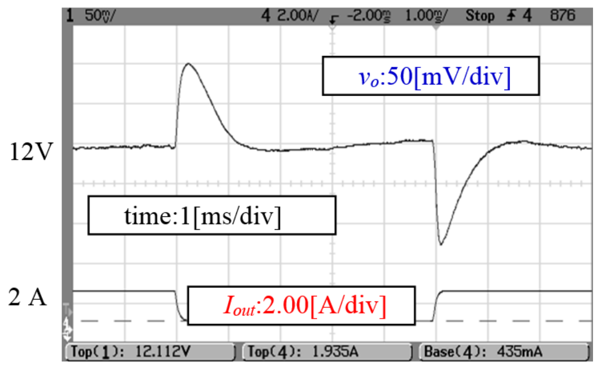

Fig.20: *Experimental results of sudden load change* R_L:6 $\leftrightarrow$ 25(Ω), *where using PI controller,* V_{in}=24 [V].

References

[1] J. Lazar and R. Martinelli, "Steady-State Analysis of the LLC Series Resonant Converter", *Applied Power Electronics Conference and Exposition 2002, APEC 2001, 16th ANNUAL IEEE*, pp.728-735 (2001)

[2] B. Yang, F. Lee, A. Zhang and H. Guisong, "LLC Resonant Converter for Front End DC/DC Conversion," *Applied Power Electronics Conference and Exposition 2002, APEC 2001, 17th ANNUAL IEEE*, pp.1108-1112 (2002)

[3] G. Hsieh, C. Tsai, S. Hsieh, "Design Considerations for LLC Series-Resonant Converter in Two-Resonant Regions," *Power Electronics Specialist Conference, PESC2007*, pp.731-736 (2007)

[4] J. Jung and J. Kwon, "Theoretical Analysis and Optimal Design of LLC Resonant Converter," *European Conference on Power Electronics and Applications*, pp.1-10 (2007)

[5] C. Lai, R. Lee, T. Wang and K. Shyu, "Design and Implementation of a Single-Stage LLC Resonant Converter with High Power Factor," *IEEE International Symposium on Industrial Electronics, ISIE 2007*, pp.455-460.

[6] C. Jin, T. Ninomiya and S. Tomioka, "Dynamic Characteristics of Current -Resonant DC-DC Converter," *IEIC Technical Report, EE2008-35*, pp.1-6 (2008)

[7] J. Jang, M. Joung, B. Choi, and H.-G. Kim, "Dynamic Analysis and Control Design of Optocoupler-Isolated LLC Series Resonant Converters with Wide Input and Load Variations," *IEEE Energy Conversion Congress and Exposition,. ECCE 2009. IEEE*, pp. 758 - 765 (2009)

[8] K. Higuchi, K. Nakano, T. Kajikawa, E. Takegami, S. Tomioka, K. Watanabe, "A New Design of Robust Digital Controller for DC-DC Converters," *IFAC 16th Triennial World Congress*, (CD-ROM), 2005.

[9] K. Higuchi, E. Takegami, K. Nakano, T. Kajikawa, S. Tomioka, "Digital Robust Control for DC-DC Converter with Second-Order Characteristics," *ECTI-CON'2009*, pp.161-169, 2009.

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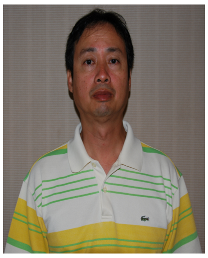

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