IMPLEMENTATION OF HIGH PRECISION MAGNET POWER SUPPLY USING THE DSP*

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Abstract

This paper presents an implementation of a precision magnet power supply (MPS) for the Pohang Light Source using the digitally controlled pulse width modulation method. The maximum output current of the power supply was 600 A at a precision of ~60 ppm. The digital control circuit of the power supply was implemented using two high speed 16-bit analog-to-digital converters and the TMS320F2808 digital signal processor. Three IGBTs are used at MPS with phase shifted parallel operation to increase the power rating and operating frequency. The duty ratio for IGBT control was determined using the PI control method. To reduce the output current ripple, the damped L-C filter was fabricated at both the DC link and output sides. Various experimental results, such as stability, drift, and controllability, are given to verify the characteristics of the DSP based magnet power supply.

INTRODUCTION

Many IGBT switching mode power supplies were implemented using a pulse width modulation controller such as UC1825 from Texas Instrument Co. This type of power supply can be a high precision MPS if the resolution and stability of the digital-to-analog converter, which generates an analog reference signal, is high. However, it requires many hardware components to implement the controller, and its output current is susceptible to variation of operating conditions.

A fully digital-controlled MPS has many advantages over the analog one. Its output characteristics are less sensitive to noise and less susceptible to parameter variations from thermal and aging effects. Also, it has a flexible and re-configurable control system [1]. We implemented and tested a fully digital-controlled MPS for the Pohang Light Source. The MPS used the phase shifted parallel operation [2] of three IGBTs to increase the power rating and operating frequency. The IGBTs were operated at a same duty ratio but 120^{o} interleaved control phases. Because of this, the system operating frequency increased to three times higher than the one for a normal switching mode MPS, and the ripple content in output current could be reduced easily. The output of each IGBT was connected to a small separate inductor to reduce the size of inductor in the filter circuit. Digital control for MPS system was built on the DSP TMS320F2808 from TI Co. In this paper, we present the design details and measured results of the fully digital-controlled high precision MPS for the Pohang Light Source.

AVERAGE CURRENT MODE CONTROL

The control of a switching mode power supply can be analysed using either the average current or peak current mode operations. For the peak current mode control, the instantaneous peak current of the inductor, which regulates the output current, is sensed and used to determine the duty ratio of the PWM. The duty ratio determined from the measured instantaneous peak current is quite fluctuating, and the peak current mode control is not suitable for a precision MPS for accelerator. For the average current mode (ACM) control [3], the load current is sensed and averaged for a given control loop duration, compared with the reference value to obtain the current error, and the current error is fed into the compensator in the control loop. Because the load current is averaged over the control loop duration, the noise immunity of the control loop is very high. A schematic diagram of buck mode MPS using the ACM control is shown in Fig.1. Here, D is the steady state duty ratio of the IGBT control signal, $H(s)$ is the gain of current sensor, and $G_c(s)$ is the proportional and integral (PI) compensator for load current regulation.

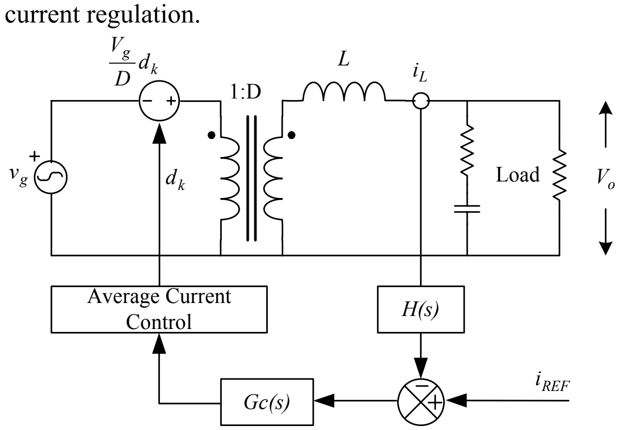

Figure 1: A schematic diagram of bucked mode MPS using the ACM control.

For an analysis of the MPS, we assume that the current through the filter capacitor is much smaller than the load current, and the converter operated at a continuous conduction mode. Then the inductor current is given by the following equations:

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$$L\frac{di_L}{dt} = V_g - V_o \quad \text{for} \quad t_k \le t < t_k + d_k T_s \tag{1}$$

$$L\frac{di_L}{dt} = -V_o \quad \text{for} \quad t_k + d_k T_s \le t < t_{k+1} \tag{2}$$

Thus the initial inductor current at the (k+1)*th* switching cycle can be written as

$$i_L(k+1) = i_L(k) + \frac{V_g(k) - V_o(k)}{L} d(k)T_s - \frac{V_o(k)}{L}(1 - d(k))T_s \tag{3}$$

where d_k is the duty ratio function of the PWM and T_s is the switching period. In a steady state operation, the inductor current should follow the reference current $i_{REF}(k)$, i.e., $i_{REF}(k) = i_L(k)$.

From (3), the duty ratio in k th switching period is given by

$$d(k) = \frac{[i_{REF}(k+1) - i_{REF}(k)]L/T_s}{V_g(k)} + \frac{V_o(k)}{V_g(k)} \tag{4}$$

Then, the time averaged load current $< i_L >$ becomes

$$< i_L > = \frac{1}{T_s}\int_t i_L dt \quad \text{for} \quad t_k \le t < t_k + T_s . \tag{5}$$

During each switching period, the averaged load current $< i_L >$ was calculated in DSP using the digitized load current and the error from $i_{REF}(k)$ was fed into the PI compensator.

SYSTEM CONFIGURATION

The system configuration of the designed MPS is shown in Fig. 2. It used three IGBTs as switching elements. The IGBTs were operated at a same duty ratio but 120^o interleaved control phases. Accordingly, the duty cycle of each IGBT was 1/3 of the converter duty cycle. This parallel operation mode also contributes to deal with the thermal problems of the switching components and inductors. Each IGBT output terminal was connected to a small separate inductor in the output filter circuits with reduced size of inductor.

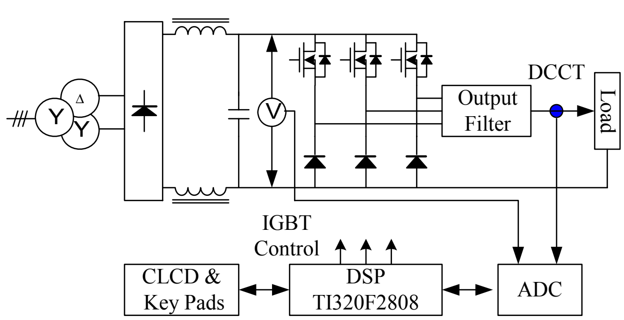

Figure 2: System configuration of MPS.

After each current of three IGBT were added up to output current of 600 A, the second stage output filter circuits were implemented to eliminate higher harmonics.

The output filter and load magnet were modelled as shown Fig. 3, where the load magnet was modelled with R_1 and L_1, and the filter circuit consisted of C_2, R_2 and L_3. The implemented output filter had more components in shown Fig. 3, and the values, which was not appeared components, are so small that the effects of these components could be ignored.

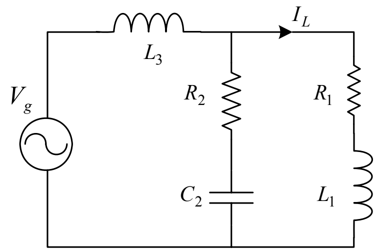

Figure 3: Equivalent circuit of output filter and load magnet.

The transfer function $G_p(s)$ of the circuit in Fig. 3 is

$$G_p(s) \equiv \frac{I_L(s)}{V_g(s)} = \frac{b_1 S + 1}{a_3 S^3 + a_2 S^2 + a_1 S^1 + R_1} \tag{6}$$

where $a_1 \equiv L_1 + L_3 + R_1 R_2 C_2$, $a_2 \equiv R_2 C_2 L_1 + L_3 R_1 C_2 + L_3 R_2 C_2$, $a_3 \equiv C_2 L_1 L_3$, and $b_1 = C_2 R_2$. The control loop for the designed MPS is given in Fig. 4. It does not contain any voltage feedback or reference feed-forward. The coefficients of PI compensator $G_c(s)$ were determined directly using the characteristic equation of the control loop.

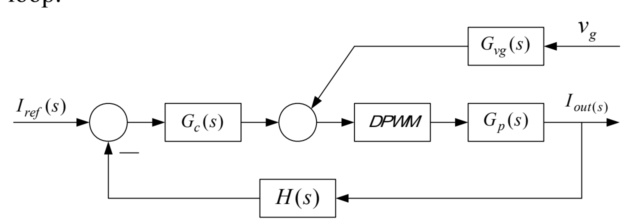

Figure 4: Block diagram of control loop of MPS.

For the diagram in Fig. 4, the output current is given by

$$I_{out}(s) = I_{ref}\frac{G_c(s)G_p(s)}{1 + T(s)} + v_g \frac{G_{vg}(s)G_p(s)}{1 + T(s)} \tag{7}$$

where $T(s) \equiv G_c(s)G_p(s)H(s)$ and $H(s)$ is the DC current transducer (DCCT) gain. Here the *DPWM* does not affect the loop equation. The step response of the control system is shown in Fig. 5, which shows a critical damp response and the rising time up to 600 A takes about 500 ms owing to the relatively large time constant of the magnet load.

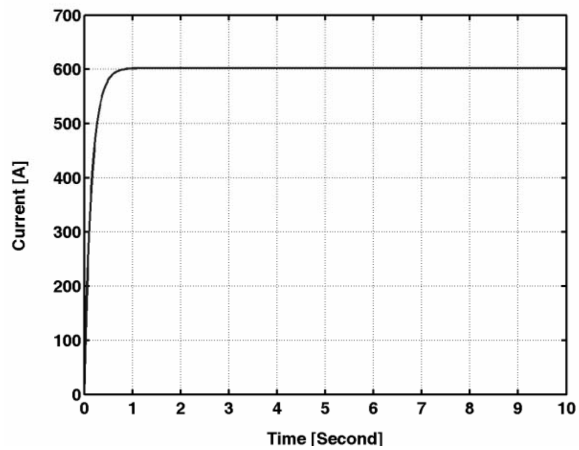

Figure 5: Step response of the overall control system.

We used the DSP TMS320F2808 to control the overall power supply system and the BSM300GB60DLC IGBTs from EUPEC Co. which can drive 300A at 600V link voltage. The controllability of the TMS320F2808 with the micro-edge positioning technology [4] was less than 10 ppm, which satisfied the required accuracy for the MPS. Since three IGBTs were connected in parallel with 120 ^{o} interleaved control phases, the system operating frequency was tripled to 60 kHz and the MPS could drive up to 600 A.

For the control loop, the output current was measured using a DDCT and the output of DDCT was converted into a digital signal using two AD977A 16-bit analog-to-digital converters (ADCs) from Analog Devices. For control loop duration, each ADC digitised the output current ten times and the outputs of both ADCs were summed and averaged to reduce the system noise. The ADCs and EPWM duty control were executed at 20kHz using a real time interrupt routine. The overall control loop was updated for every 500 µs using the Timer Interrupt of the TMS320F2808. Both key scan and the LCD display operations were carried out at the base routine to avoid any serious control timer interrupt.

EXPERIMENTAL RESULTS

A 60 sec. short term current stability at different load current was measured using the HP3458A digital voltmeter from Agilent Co, and the results are given in Table 1 and Fig. 5. The stability of load current was less than 60 ppm at 550 A output current, and it decreased gradually to the 10 ppm at 50 A, as given in Table 1.

Table 1: The measured fluctuation of output current for 60 sec.

Load Current [A]	50	150	250	350	450	550
Stability [ppm]	10	20	30	40	50	60

The long term stability for two hours was also measured at 350A to check the system drift. The DC drift of output current was -10 ppm approximately and the peak-to-peak fluctuation increased to ~50 ppm. We confirmed that the current unbalance between IGBTs at a load current of 600 A was less than 1 A.

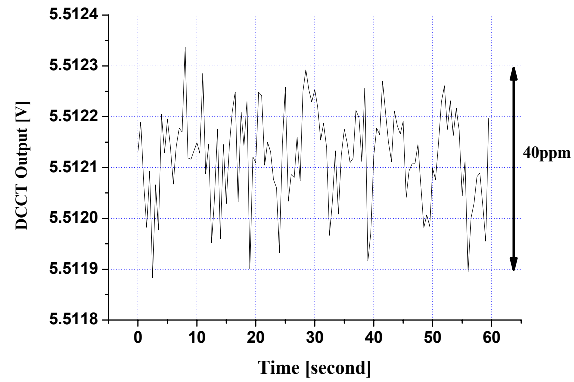

Figure 6: Current stability at load current of 550 A.

CONCLUSIONS

A fully digital-controlled high precision MPS for the Pohang Light Source was designed and tested. The maximum output current of the power supply was 600 A at a precision of ~60 ppm for a load inductance of 5 mH. Three IGBTs were connected in parallel with 120 ^{o} interleaved control phases to increase the output current and system operation frequency. The duty factor for IGBT control was determined using the PI control method. The digital controller of MPS was implemented using two high speed 16-bit ADCs, the DSP TMS320F2808 and the other many components. The current unbalance between IGBTs at a load current of 600 A was less than 1 A. The temperature difference between the IGBTs was ~ 10 °C at the heat sink which was water-cooled to 45 °C for a load current of 580 A.

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