Digital Controller for Rapid Cycling Synchrotron Magnet Power Supply with Very High Tracking Precision

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Abstract -- In Rapid Cycling Synchrotron, very high tracking precision is required for the magnets power supply, while the load is high order resonant network. In this paper, a digital PID controller combined with repetitive control algorithm is studied for such an application. Feedforward scheme is also adopted to reduce the tracking error under the periodical reference. A digital controller based on DSP is built and tested on a prototype power supply with emulated load. Experimental results show that the proposed control system achieves nearly 0.1% reference tracking precision.

Index Terms—Rapid Cycling Synchrotron, Magnet Power Supply, Very High Tracking Precision, Digital Controller.

I. INTRODUCTION

Synchrotron is used not only in high energy physics, but also in the other science fields such as medicine. Magnet power supplies for Synchrotron are specialized power supplies which have been used in accelerators for particle beam excitation and control. Due to the high precision requirement of magnetic field, magnet current has to be controlled with little reference tracking error. A typical waveform of Rapid Cycling Synchrotron (RCS) magnet current is shown in Fig. 1. The magnet current is dc biased sinusoidal, and the repetition rate is several tens Hz. Proton beam is injected at the valley and then accelerated. Magnet current should be reset after the proton beam is extracted at the peak. In the accelerate process, reference tracking error specification is very strict. Furthermore, in order to avoid huge throughout on utility, White Circuit, shown in Fig. 2 is widely used in RCS magnet power supply. As is shown, power supply only provides active power, while reactive power flows between magnet and resonant elements.

Fig. 1. A typical waveform of magnet current in RCS

The Rapid Cycling Synchrotron (RCS) for China Spallation Neutron Source (CSNS) is a Rapid Cycling Bevatron, the repetitive frequency of which is 25 Hz. Dynamic magnetic field generated by the electromagnet of RCS deflects and focalizes the proton beam. In order to avoid departure from normal trajectory of proton beam, the magnet current provided by magnet power supply demands extremely high tracking precision which is 0.1% for RCS magnet power supply. Therefore, it is significant to study on the control strategies of high tracking precision magnet power supply.

In [1], a combined feedback and feed forward control scheme is proposed for fast response switch mode magnet power supply and the simulation results are provided. However, in real applications, the circuit parameters are not ideal and even the inductance varies with the inductor current. Thus feed forward path which is parameter dependent may not deliver the desired results. In the paper [2], repetitive controller is used in the proton synchrotron magnet power supply and high accuracy magnet current is obtained. Unfortunately, the power stage is thyristor converter, whose small signal model differs from that of switch mode power supply. Also, a combined PI and feed forward control scheme is applied to a dynamic switch mode magnet power supply [3], but the load circuit is not White Circuit. In general, there is seldom literature which studies on the digital control scheme of switch mode dynamic magnet power supply used in RCS.

The paper is mainly devoted to the control schemes for high-tracking-precision dynamic magnet power supply on the condition of large signal reference. Theoretical analysis of availability of the two-loop control scheme is given and PI controller for the inner inductor current loop is provided. Modified PID is applied to outer magnet current loop. In order to enhance the tracking precision, the feed forward scheme is also adopted. Also, to further reduce the tracking error, a feed forward scheme is also adopted.
error, the repetitive controller is utilized in parallel with PID controller. This control scheme is proposed to achieve very high dc biased 25Hz reference tracking precision used in RCS. Also, the proposed control scheme is verified in the low power prototype.

II. TOPOLOGY OF MAGNET POWER SUPPLY AND ITS SMALL SIGNAL MODEL

The topology of magnet power supply for RCS is shown in Fig. 3. It mainly includes three stages: the two quadrant chopper for magnet current regulation, the\( LC \) low pass filter for high frequency harmonics reduction and the load. As shown in Fig. 3, \( U_{in} \) is the dc link voltage produced by a boost converter; \( L \) is the inductor of low pass filter and \( R_L \) is its ESR; \( C \) is the capacitor of low pass filter and \( R_C \) is its ESR; \( C_{ch} \) is the resonant capacitor of White Circuit and \( R_{ch} \) is its ESR; \( L_m \) is the magnet inductor and \( R_m \) is its ESR; \( C_{ch} \) is the resonant inductor of White Circuit and \( R_{ch} \) is its ESR. The circuit parameters are listed in Table I.

The impedance characteristic of load circuit (White circuit) is shown in Fig. 4. As is shown, the series resonant frequency is 25Hz. Thus, the load impedance in 25Hz is extremely small. This means that we can obtain the large magnet current with very low dc link voltage. In addition, with this load, the two quadrant chopper just provides the active power to the load and the passive power is near to zero. This property is beneficial to the design of the power stage. This circuit is widely used due to these merits although it increases the power loss because of the equivalent series resistors of the resonant elements.

Using the traditional modeling method [4], the small signal model can be obtained and the duty cycle to magnet current transfer function and duty cycle to filter inductor current transfer function are shown in (1) and (2).

\[
G_{id}(s) = \frac{\dot{i}_d(s)}{d(s)} = k_1\left(\frac{s^2 + a_1s + a_2^2}{s^2 + a_3s + a_4^2}\right)
\]

\[
G_{iod}(s) = \frac{\dot{i}_{od}(s)}{d(s)} = k_2\left(\frac{s^2 + R_m + R_{ch}}{L_m + L_{ch}}(s^2 + a_5s + a_6^2)(s^2 + a_7s + a_8^2)\right)
\]

where \( G_{id}(s) \) is the duty cycle to filter inductor current transfer function; \( G_{iod}(s) \) is the duty cycle to magnet current transfer function; coefficient \( a_1, a_2, a_3, a_4, a_5, a_6, a_7, a_8, k_1 \) and \( k_2 \) are related to circuit parameters.

By substituting parameters shown in table I for (1) and (2), bode diagram of (1) and (2) can be obtained in Fig. 5.

![Fig. 4. Impedance characteristic of load circuit.](image)

![Fig. 5. Bode diagram of (1) and (2).](image)
III. DIGITAL CONTROLLER DESIGN FOR HIGH TRACKING PRECISION MAGNET POWER SUPPLY

In this magnet power supply, the magnet current reference is a dc biased 25Hz sinusoidal wave as like the Fig.1 and the tracking precision is required better than 0.1%. In a linear closed loop control system, the tracking precision can be represented in s domain as follows

\[ TP = \frac{e(s)}{r(s)} = \frac{1}{1 + T(s)} \]  \hspace{1cm} (3)

where \( TP \) is the tracking precision; \( r(s) \) is the reference; \( e(s) \) is the error; \( T(s) \) is the open loop transfer function.

According to (3), in order to meet the requirement of tracking precision, the loop gain in DC should be as high as possible, which can be easily implemented by using the integrator. In addition, the open loop gain at the frequency of 25Hz should be higher than 60dB, which is a big challenge for the design of the compensator, especially when the switching frequency is low.

As shown in Fig. 5, phase lag of the duty cycle to magnet current transfer function in the mid-frequency is near to 270 degrees. If we try to directly compensate the system with one closed loop, the final compensator should include three poles and three zeros at least to guarantee both the stability and high bandwidth. This compensator is too complex. Furthermore, this control scheme is difficult to realize the current sharing between multiple two quadrant choppers, which is required in the final product in the future.

In order to solve these problems, we adopt two loop control scheme. The inner loop is the filter inductor current feedback and the outer loop is the magnet current feedback. There are three advantages of applying the inner loop: (1) suppress the low frequency disturbance from the DC link voltage; (2) reconstruct the property of the plant, which benefits the design of outer loop controller; (3) realize the current sharing between multiple two quadrant choppers. In order to meet the requirement of the tracking precision, the purpose of the outer-loop compensator design should be to boost the loop gain in the DC and in 25Hz. In this paper, we adopt the combined PID and repetitive controller as well as the reference feed forward method to compensate the outer loop.

The detailed analysis is presented as follows.

A. Design of PID controller

The diagram of the control system using dual feedback is shown in Fig. 6.

As shown in Fig. 6, \( R(s) \) is the reference transfer function; \( G_a(s) \) and \( G_d(s) \) are the compensators for filter inductor current and magnet current loop; \( K_{PWM} \) is the gain of PWM module; \( G_{id}(s) \) is the duty cycle to filter inductor current transfer function; \( H_1 \) and \( H_2 \) are the feedback coefficient of filter inductor current and magnet current; \( K_{ADC1} \) and \( K_{ADC2} \) are the gain of ADCs; \( G_{id}(s) \) is the inductor current to magnet current transfer function. These parameter values are listed as following: \( K_{PWM}=12500 \), \( H_1=H_2=0.4 \), \( K_{ADC1}=K_{ADC2}=64000 \) and the switching frequency is 20 kHz. The open inner loop transfer functions before compensation and after compensation are presented in (4) and (5).

\[ T(s) = K_{PWM}G_{id}(s)K_{ADC1}H_1 \]  \hspace{1cm} (4)

\[ T(s) = G_{a}(s)K_{PWM}G_{id}(s)K_{ADC1}H_1 \]  \hspace{1cm} (5)

A PI controller is used as the compensator \( G_a(s) \) and its discrete form in z-domain is derived in (6). The bode diagrams of (4) and (5) are shown in Fig. 7 and Fig. 8 respectively.

\[ G_a(z) = \frac{1.326z - 1.275}{z - 1} \]  \hspace{1cm} (6)

Fig. 7. Bode diagram of (4)

Fig. 8. Bode diagram of (5)
As shown in fig. 8, the open loop gain in low frequency is very high, which improves the ability to resist the disturbance from the dc link voltage. In addition, after compensating for the inner loop, in low frequency, the filter inductor current is similar to the property of the current source. Therefore, it decreases the plant order in low frequency, which is beneficial to the outer loop design.

The open outer loop transfer functions before compensation and after compensation are presented in (7) and (8).

\[
T_i(s) = G_m(s)G_{iH}(s)H_zK_{\text{DC2}}
\]

\[
T_i(s) = G_m(s)G_{iH}(s)H_zK_{\text{DC2}}
\]

where \( G_m(s) \) is the close loop transfer function of inner current loop.

A modified PID controller is used as the compensator \( G_{co}(s) \) and its discrete form in z-domain is derived in (9). The bode diagrams of (7) and (8) are shown in Fig. 9 and Fig. 10.\[
G_{co}(z) = \frac{26.9z^2 - 53.29z + 26.45}{z^2 - 1.756z + 0.7559}
\]

Fig. 9. Bode diagram of (7)

Fig. 10. Bode diagram of (8)

As shown in Fig. 10, the loop gain at 25Hz is only 30dB. According to (3), the tracking precision is only 3.16% for 25Hz sinusoidal reference component, which cannot meet the high tracking precision requirements of magnet power supply.

B. Combined PID and Repetitive Controller with Reference Feedforward

In order to further enhance the tracking precision, a combined PID and repetitive controller with reference feedforward control scheme is applied to the magnet current loop. The schematic diagram of the proposed digital controller is shown in Fig. 11.

Supposing there is no repetitive controller in parallel with the PID controller, the \( R(z) \) to \( E(z) \) impulse transfer function can be derived as (10)

\[
\frac{E(z)}{R(z)} = \frac{1 - T_i(z)G_f(z)}{1 + T_i(z)}
\]

where \( T_i(z) = G_m(z)G_{iH}(z)H_zK_{\text{DC2}} ; \)

\( T_i(z) = G_m(z)G_{iH}(z)G_{co}(z)K_{\text{DC2}}H_z ; \)

\( G_f(z) \) is the reference feedforward transfer function.

Because \( T(z) \) is nearly equal to one at the frequency of 25Hz, the \( G_f(z) \) should be one to eliminate tracking error according to (10). Also, to undermine the periodic perturbation, the repetitive controller is added in parallel with PID controller. Thus, the \( R(z) \) to \( E(z) \) impulse transfer function with repetitive controller can be obtained

\[
\frac{E(z)}{R(z)} = \frac{(z^N - Q(z))(1 - T(z))}{(1 + T_i(z))[z^N - (Q(z) - z^NS(z)T(z))^N]}
\]

In order to ensure the stability of the system and improve the tracking ability, the parameters are selected as following [5]-[7].

\[
N = 800; Q(z) = 0.95; k = 15;
S(z) = G_m(z)G_p(z);
G_p(z) = \frac{0.062832z}{1.062832z - 1}.
\]
IV. EXPERIMENTAL VERIFICATION

The proposed digital controller based on DSP TMS320F28335 is built and tested on the low power prototype, and the circuit parameters are shown in table I. The load, control board and power stage is shown in Fig. 12.

(a) load            (b) control board       (c) power stage

Fig. 12. Experimental hardware

The magnet current is 3+2\sin(50\pi t) A. Reference and error data are output through a two channel DAC. The reference and error waveforms using PID controller is shown in Fig. 13. The reference and error waveforms using PID controller with reference feedforward is shown in Fig. 14. The reference and error waveforms using proposed digital controller is shown in Fig. 15.

The peak-peak value of error signal in Fig. 13 is 0.342V, while the peak value of reference signal is 2V. The error data is enlarged to 6 times in Fig. 13. Thus, we can obtain the tracking precision $TP_1$ using PID controller

$$TP_1 = \frac{0.342}{6 \times 2} \times 100\% = 2.85\% \quad (13)$$

The peak-peak value of error signal in Fig. 14 is 0.176V, while the peak value of reference signal is 2V. The error data is enlarged to 64 times in Fig. 14. Thus, we can obtain the tracking precision $TP_2$ using PID controller with reference feedforward

$$TP_2 = \frac{0.176}{64 \times 2} \times 100\% = 0.1375\% \quad (14)$$

The peak-peak value of error signal in Fig. 15 is 0.136V, while the peak value of reference signal is 2V. The error data is enlarged to 64 times in Fig. 15. Thus, we can obtain the tracking precision $TP_3$ using proposed digital controller

$$TP_3 = \frac{0.136}{64 \times 2} \times 100\% = 0.10625\% \quad (15)$$

The waveform of dc link voltage is shown in Fig. 16. The main component of disturbance from dc link voltage is 25 Hz and its harmonics, because output power of magnet power supply is not constant. Due to low value of filter capacitor, there is a large disturbance from dc link voltage. RMS value of the dc link voltage is about 20V. In Fig. 15, we can see that the peak-peak value of dc link voltage is 7V. Thus, the disturbance $DB$ from dc link voltage can be obtained

$$DB = \frac{7}{20} \times 100\% = 35\% \quad (16)$$

Although disturbance from the dc link voltage is large, the proposed controller still meets the very high tracking precision required for RCS. This further verify the ability of the proposed controller to resist the disturbance from dc link voltage,
V. CONCLUSION

In the paper, three control schemes have been applied to the RCS magnet power supply. PI controller for inner inductor current loop is used to resist dc link voltage disturbance and reconstruct the plant. PID controller is also applied to magnet current to ensure the high tracking precision of magnet current. However, a single PID controller can only achieve the 2.85% tracking precision when the reference is dc biased 25Hz. By adding reference feedforward to PID controller, the tracking precision is increased to 0.1375%, which still cannot meet the tracking 0.1% tracking precision required for RCS. Finally, the repetitive controller is added in parallel with PID controller to undermine the periodical disturbance, which further enhances the tracking precision to 0.10625%. Experimental results show that the proposed digital controller can achieve nearly 0.1% tracking precision when the reference is dc biased 25Hz and has strong ability to resist disturbance from the dc link voltage.

REFERENCES