Study on a Highly Stabilized Power Supply for Hybrid-Magnet Superconducting Outsert

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Abstract The superconducting outsert of the 40 T hybrid-magnet in High Magnetic Field Laboratory (HMFL) of Chinese Academy of Sciences (CAS) requires a highly stabilized power supply. In this paper, two kinds of power supply design are briefly presented and both advantages and disadvantages are analyzed. In order to overcome the drawbacks of switching power supply, a series regulated active filter is adopted and a new design is proposed which ensures cooperative relationship between the feedback control loops of the switching converter and the series regulated active filter. Besides, unlike the traditional switching power supply, which can generate positive voltage only, this new design can also generate negative voltage which is needed in the quench protection for the superconducting magnet. In order to demonstrate the effectiveness of the methodology, a low-power prototype has been accomplished. The simulation and experiment results show that the power supply achieves high precision under the combined action of two feedback control loops. The peak-to-peak amplitude of the output ripple voltage of the prototype is 0.063\%, while the peak-to-peak amplitude of the output ripple current is 120 ppm.

Keywords: superconducting magnet, switching power supply, series regulating, feedback control, quench protection

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(Some figures may appear in colour only in the online journal)

1 Introduction

A highly magnetic field is an extreme environment that can be widely used for nuclear science, chemistry, biology, etc. Compared with the conventional magnets, superconducting magnets have much smaller volume, much better flexibility and much higher efficiency. These features result in superconducting magnets being used in various scientific equipment, especially in magnetic confinement fusion devices such as International Thermonuclear Experimental Reactor (ITER) \cite{1} and Experimental Advanced Superconducting Tokamak (EAST) \cite{2−4}.

The hybrid-magnet superconducting outsert in HMFL requires an 8 V/14 kA power supply with high current and voltage precision. The power supply is planned to adopt two kinds of schemes. One is the silicon-controlled rectifier (SCR) which is similar to the scheme used in EAST \cite{5} and the other is the switching power supply.

This paper gives a brief account of both schemes and describes an improved scheme based on the switching power technology. The experimental results have illustrated the feasibility of the improved scheme, which can thus provide a simple reference for the power supply systems of other magnetic confinement fusion devices.

2 Original schemes

2.1 Dual reverse star-shaped SCR scheme

Fig. 1 shows the topology of a dual reverse star-shaped SCR. The secondary side of the transformer consists of a pair of windings with equivalent turns and opposite polarity for each phase. The windings are connected to two three-phase half-wave rectifiers as shown in Fig. 1. SM is the superconducting magnet.

Lp1 and Lp2 in Fig. 1 are balance inductors which also function as filters. The balance inductors can ensure that both the three-phase half-wave rectifiers operate at the same time so that this rectifier can output double current, like a three-phase full-bridge rectifier, to meet the demand of high current.

With the help of voltage signals produced by synchronous transformer and feedback control loops, the phase shifter and thyristor trigger can produce pulses with an appropriate delay angle which determines the average DC output voltage of the dual reverse star-
shaped SCR. Therefore, the power supply can respond rapidly to the changes of load currents or realize constant output.

The advantages of this solution are matured technique, high reliability and high precision. However, the disadvantages are low power factor, bulky and heavy equipment, etc.

2.2 Switching power supply scheme

Fig. 2 shows the topology of switching power supply. In this figure, $V_{\text{grid}}$ is the three-phase grid input voltage, D1-D6 are rectifier diodes, $L_{\text{in}}$ is the input filter inductance, $C_{\text{in}}$ is the input filter capacitor, QA-QF are power MOSFETs, $L_s$ is the freewheel inductance, $L_f$ is the input filter inductance, $C_f$ is the input filter capacitor.

The input portion of the power supply is a three-phase full-bridge uncontrolled rectifier. Then there is a soft switching inverter. A saturated inductor is connected to the transformer primary coils in series. Meanwhile, synchronous rectifiers are connected to transformer secondary coils. In order to obtain high-power output performance, the output portion of the power supply adopts multiple rectifying circuits.

The high-frequency switch technology makes the power supply smaller and more efficient, which also conforms to the development trend of magnetic confinement fusion devices. However, because of the high ripple in its output current/voltage, the original switching power supply scheme cannot be directly used for the superconducting magnet.

3 Improved scheme based on switching power technology

In order to reduce the output ripple, the passive filter, if used alone, requires much larger output inductance and capacitance, which will increase the cost and volume of the device and exacerbate dynamic characteristics. Moreover, the low-frequency output ripple cannot be depressed effectively. Accordingly, an active filter is usually added to the output of the switching power supply.

The improved scheme adopts a series regulator, which takes advantage of the on-resistance variability of adjusting transistors to depress the input voltage ripple and obtain extremely low output ripple voltage. At present, the HMFL power supplies in USA, Holland and Japan have adopted this kind of active filter.

3.1 Selection of adjusting transistor of series regulator

In this paper, the output voltage of the power supply is 8 V. If the voltage across adjusting transistor is high, the power loss of the series regulator will definitely impair the supply’s efficiency.

With the development of the power MOSFET, the minimum on-resistance ($R_{ds}$) is measured to be of the order of mΩ, which makes the drain-to-source voltage ($V_{ds}$) remain low when the through-current of power MOSFET is high.

Because of the positive temperature coefficient (PTC) characteristic of the MOSFET in steady conducting state, a rise in the temperature of one of the shunt MOSFETs will result in an increase in the resistance, which leads to the decrease of the current and temperature of the MOSFET. Thereby, it is easy to implement automatic current-sharing control and achieve current equilibrium when the adjusting transistor adopts MOSFETs in parallel connection.

3.2 Control solution in the improved scheme

Fig. 3 shows a functional block diagram of the routine series regulated switching power supply.

Because the two feedback control loops in the figure are mutually independent, only the voltage across the adjusting transistor ($V_{ds}$) can be regulated when the load changes suddenly. However, if the load change is great, $V_{ds}$ will change dramatically, which may damage the adjusting transistor.

To solve the problem mentioned above, the improved scheme designates $V_{ds}$ as a feedback parameter to control the output of switching converter as shown in
Fig. 4. The feedback loop 1 adopts the single-loop voltage control strategy while the feedback loop 2 adopts the double closed-loop control strategy, which consists of an outer current loop and an inner voltage loop.

When the output current increases, the feedback loop 2 will motivate \( V_{GS} \) to decrease, which leads to an increase in \( R_{ds} \) and \( V_{ds} \).

On one hand, the last output voltage can be acquired as

\[
V_o = V_o' - V_{ds}. \tag{1}
\]

Therefore, an increase in \( V_{ds} \) can result in a decrease in \( V_o \).

On the other hand, the feedback loop 2, as \( V_{ds} \) increases, will narrow the PWM pulse width, which leads to a decrease in \( V_o' \) and \( V_o \) as shown in Eq. (1).

As \( V_o \) decreases, the output current will decrease too, which makes the power supply approach to a steady state.

The above analysis results show that the feedback loop 1 and loop 2 act synergistically to stabilize the output.

4 The effect of improved scheme in quench protection

During a quench, the tremendous energy stored in the superconducting magnet can be converted into heat. If there are no discharge circuits, the thermal stress caused by the local over-heating may damage the magnet. Moreover, the high voltage rise caused by the quench may lead to dielectric breakdown. Therefore, it is extremely important to introduce a quench protection circuit [6].

Fig. 5 shows a basic quench protection circuit. In this figure, DS represents the fast DC circuit breaker, \( R_e \) the dump resistor, \( R_p \) the grounding resistor for the magnet, and SM the superconducting magnet.

If the current passing through the coils of the superconducting magnet (\( I_{sm} \)) is very high when the magnet quenches, no matter what is the design of the power supply, the DS contact must be interrupted, and the energy stored in the superconducting magnet should be consumed on the dump resistor.

If a dual reverse star-shaped SCR is adopted, a negative voltage (\( -U_o \)) can be generated by the inverter. Meanwhile, \( -U_o \) can be expressed as:

\[
-U_o = L_{sm} \frac{di}{dt}, \tag{2}
\]

where \( L_{sm} \) represents magnet inductance.

The period of time that \( I_{sm} \) drops from \( I \) to 0 A during a quench is:

\[
t = \frac{I}{(-\frac{di}{dt})}. \tag{3}
\]

In case of a quench, \( I^2t \) must be smaller than \( M \) to protect the magnet from damage. Therefore:

\[
\int_0^t (I + \frac{di}{dt})^2 dt = \int_0^t \frac{I_{sm}}{L_{sm}} (I - \frac{U}{E_{sm}} t)^2 dt < M. \tag{4}
\]

We obtain:

\[
I < \sqrt[3]{\frac{3U_oM}{L_{sm}}}. \tag{5}
\]

Thus, if \( I_{sm} \) accords with formula (5) when the magnet quenches, the negative voltage \( -U_o \) can force \( I_{sm} \) to...
decrease linearly to realize a quench protection without the operation of protective switches.

However, if the original switching power supply, which cannot generate a negative voltage, is adopted, it is impossible for \( I_{sm} \) to decrease safely and controllably without the operation of protective switches even if \( I_{sm} \) accords with formula (5).

In the improved scheme, if \( I_{sm} \) is very low when the magnet quenches, we can set the feedback control loop 2 as Fig. 6 shows. The feedback voltage is the voltage across the adjusting transistor and the reference voltage can be calculated according to the demand of quench protection.

\[
V_{ds} = \frac{4L_{lk}f_o}{K^2}, \quad (8)
\]

where \( d_{\text{eff}} \) is the modulation of effective duty cycle, \( d \) is the modulation of duty cycle of the primary voltage, \( d_i \) is the duty cycle modulation due to the change of the filter inductor current (\( I_{f}^2 \), \( d_o \) is the duty cycle modulation due to the change of the input voltage (\( V_{in} \)), \( D \) is the duty cycle of the primary voltage set by the control, \( D' = 1 - D \), and \( I_s \) is the saturated current of the saturated inductor.

Based on the small signal model of the buck converter, a small signal model of the actual switching power supply, which is the main component of the experimental prototype, is presented in Fig. 8.

### 5 Development of prototype

Table 1 shows the circuit parameters’ values of the low-power experimental prototype.

#### 5.1 Design of feedback control loop 1

It is assumed that the feedback control loop 2 functions perfectly when feedback control loop 1 is taken into account. Thus, the adjusting transistor could be modelled as a fixed resistor \( R_{ds} \).

Ref. [7] introduced a small-signal model of a conventional full-bridge phase-shifted soft switching power supply. Compared with the conventional scheme, a saturated inductor was connected to the transformer primary coils in series and the load consisted of a resistor in series with an inductor. Through a similar analysis as performed in Ref. [7], some results can be concluded that:

\[
d_{\text{eff}} = d + d_i + d_o, \quad (6)
\]

\[
d_i = -\frac{K \cdot R_o}{V_{in}} I_{f}, \quad (7)
\]

\[
R_{ds} = 4L_{lk}f_o, \quad (8)
\]

\[
d_o = \left[ \frac{L_{lk}}{K} \left( \frac{D'V_o T_o}{2L_s} + 2L_o I_s \right) \right] \frac{f_o}{V_{in}}, \quad (9)
\]

Table 1. Circuit parameters’ values

<table>
<thead>
<tr>
<th>Circuit parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage (three phase AC), ( V_{in} )</td>
<td>380 V</td>
</tr>
<tr>
<td>Output voltage (DC), ( V_o )</td>
<td>8 V</td>
</tr>
<tr>
<td>Output current (DC), ( I_o )</td>
<td>50 A</td>
</tr>
<tr>
<td>Switching frequency, ( f_s )</td>
<td>200 kHz</td>
</tr>
<tr>
<td>Peak value of sawtooth wave, ( V_M )</td>
<td>2.8 V</td>
</tr>
<tr>
<td>Transformer winding turns ratio, ( K )</td>
<td>38</td>
</tr>
<tr>
<td>Transformer leakage inductance, ( L_{lk} )</td>
<td>7.5 ( \mu )H</td>
</tr>
<tr>
<td>Saturated inductance, ( L_o )</td>
<td>150 ( \mu )H</td>
</tr>
<tr>
<td>Output filtering inductance, ( L_t )</td>
<td>2 ( \mu )H</td>
</tr>
<tr>
<td>Output filtering capacitance, ( C_t )</td>
<td>7.5 mF</td>
</tr>
<tr>
<td>Equivalent series resistance of ( C_t ), ESR</td>
<td>1 m( \Omega )</td>
</tr>
<tr>
<td>Load inductance, ( L_i )</td>
<td>10 ( \mu )H</td>
</tr>
<tr>
<td>DC load resistance, ( R_o )</td>
<td>0.15 ( \Omega )</td>
</tr>
<tr>
<td>Reference voltage across adjusting transistor, ( V_{ref} )</td>
<td>1 V</td>
</tr>
<tr>
<td>The ratio of sampling of voltage across adjusting transistor</td>
<td>1:1</td>
</tr>
<tr>
<td>The ratio of output voltage sampling</td>
<td>1:4</td>
</tr>
<tr>
<td>The ratio of output current sampling</td>
<td>1:4</td>
</tr>
</tbody>
</table>

\[
V_{ds} = V_o R_{ds} / [R_{ds} + R_o]. \quad (12)
\]

Thus, the control-to-output transfer function can be expressed as
\[ G_{vd} = \frac{V_{ds}^2}{d} \bigg|_{V_{in}=0} = \frac{V_{in} \cdot R_{ds}}{K \cdot [s^2 L_tC_t + s \left( \frac{L_t}{R_0 + R_{in}} + R_d C_t \right) + \frac{R_d}{R_0 + R_{in}} + 1] \cdot (R_{ds} + R_o)}. \] **(13)**

Taking the equivalent series resistance (ESR) zero of the output filtering capacitance into account, the open-loop transfer function of the feedback control loop 1 without compensation can be expressed as

\[ A_{p0}(s) = G_{vd}G_M H_a = \frac{V_{in} \cdot R_{ds}(1 + s \cdot ESR \cdot C_t)}{K \cdot [s^2 L_tC_t + s \left( \frac{L_t}{R_0 + R_{in}} + R_d L_t \right) + \frac{R_d}{R_0 + R_{in}} + 1] \cdot (R_{ds} + R_o) \cdot V_M}. \] **(14)**

where \( G_M = 1/V_M \) is the pulse width modulation (PWM) transfer function, \( H_a \) is the transfer function of the sampling system for the voltage across adjusting transistor.

Transform Eqs. (14) into (15):

\[ A_{p0}(s) = \frac{\frac{V_{in} \cdot R_{ds}}{K \cdot (R_d + R_o + R_{ds})} \cdot \left( 1 + s \cdot ESR \cdot C_t \right)}{\frac{s^2 L_tC_t}{R_0 + R_{in}} + 1} \cdot \left( 1 + s \cdot \frac{ESR \cdot C_t}{V_M} \right) \cdot \left( \frac{V_{in}}{\frac{R_{ds}}{L_tC_t}} \right) = \frac{V_{in} \cdot R_{ds}}{V_M \left( \frac{1}{\omega_0^2} + \frac{1}{\omega_0} + 1 \right)}, \] **(15)**

where

\[ V_{in} = \frac{V_{in}}{K} \cdot \frac{R_{ds}}{R_d + R_o + R_{ds}}; \] **(16)**

\[ \omega_0 = \frac{1}{\sqrt{LC'}}; \] **(17)**

\[ Q' = R_{ds} \sqrt{\left( \frac{C'}{L} \right)} = \sqrt{\frac{L_tC_t(R_o + R_{ds})(R_d + R_o + R_{ds})}{L_tC_t(R_o + R_{ds})}}. \] **(18)**

\[ ESR \cdot C_t = ESR' \cdot C'. \] **(19)**

By solving Eqs. (16)-(19), we obtain

\[ L' = \frac{[L_t + R_tC_t(R_o + R_{ds})]R_{ds}}{R_d + R_o + R_{ds}}, \] **(20)**

\[ C' = \frac{L_t + C_t(R_o + R_{ds})}{[L_tC_t(R_o + R_{ds})]R_{ds}}, \] **(21)**

\[ ESR' = \frac{ESR \cdot [L_t + R_tC_t(R_o + R_{ds})]R_{ds}}{L_tC_t(R_o + R_{ds})}. \] **(22)**

According to Eq. (15), the feedback control loop 1 can be equivalent to the feedback control loop of the buck converter shown in Fig. 9, where \( V_{in} \) is the input DC voltage, \( f_s \) is the switching frequency, \( L' \) is the filtering inductance, \( C' \) is the filtering capacitance, \( ESR' \) is the equivalent series resistance of \( C' \), \( R_{ds} \) is the load resistance. Besides, the ratio of sampling of voltage across adjusting transistor is 1:1, the peak value of sawtooth wave is \( V_M \). Then we can design a typical 3-order PID compensation network for this buck converter [8].

Fig. 10 presents the Bode plots of the compensated open-loop transfer function of the feedback control loop 1. According to the Bode plots, the compensated system has a phase margin of 45.4 degrees, the open-loop gain crosses the 0 dB axis with a −1 slope and the crossover frequency is 36.3 kHz. These results completely satisfy the stability criterion [8, 9].

### 5.2 Design of feedback control loop 2

Fig. 11 illustrates the functional block diagram of the double closed-loop control strategy of feedback loop 2.

The design of feedback loop 2 is mainly based on the frequency domain simulation in PSPICE. The inner loop adopts proportion control strategy, while the outer loop adopts single-pole single-zero PI compensation, as shown in Fig. 12.
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The experimental results of the prototype are given in Fig. 15. In order to verify the automatic regulation ability during a sudden load change, the load resistance is set to change from 0.16 Ω to 0.12 Ω at the moment of 10 ms.

As shown in Fig. 15, when the load changes suddenly, $V_o$ decreases from 8 V to 6 V and $I_o$ reverts to 50 A rapidly after a small-scope fluctuation. Then it takes about 0.3 ms for $V_{ds}$ to revert to 1 V. These results show that the two feedback loops act synergistically to ensure a good performance of the feedback control system.

Fig. 16 shows the voltage waveform of the prototype in detail. According to the figure, the peak-to-peak amplitude of the ripple voltage is about 5 mV, i.e., 0.063% of the output voltage (8 V), which is much smaller than the permissible voltage ripple factor (1%).

Fig. 14 presents the parametric curves of $I_o$, $V_o$, $V'_o$ (the output voltage of the switching converter), and $V_{ds}$ in PSPICE.

The current measurement is based on the high precision DCCT. In order to get the maximum of the peak-to-peak amplitude of ripple current, the current is measured successively under the rated load during an hour. According to the measurement result, the maximum value is 50.004 A while the minimum value is 49.998 A, so the peak-to-peak amplitude of the ripple current is about 6 mA, i.e., 0.012% of the output current (50 A), which is also smaller than the permissible current ripple factor (0.1%).

7 Conclusion

Compared with the traditional SCR power supply, the high-power high-frequency switching power supply
has smaller volume, lower cost but worse output ripple. In addition, it cannot output negative voltage to realize controllable current decrease for the superconducting magnet in some cases.

In this paper, an improved scheme is proposed to solve the above-mentioned problem. The experiments on the prototype, which has high output precision and good dynamic response characteristics, have validated the effectiveness of the improved scheme. The improved scheme can be applied to the power supply systems of other magnetic confinement fusion devices after certain modifications.

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