A Multimodule Hybrid Converter for High-Temperature Superconducting Magnetic Energy Storage Systems (HT-SMES)

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***Abstract*—A multimodule hybrid converter mainly designed for use in high-temperature superconducting magnetic energy storage systems (HT-SMES) is investigated in this paper. The converter consists of a four-module current-source converter and one module of voltage-source converter to reach the high power rating and low harmonics requirement of the HT-SMES. Circuit topology, control method, and simulation results are given.**

***Index Terms*—Energy storage system, high-temperature superconductor, hybrid converter, multimodule current-source converter (CSC), superconducting magnetic energy storage system.**

I. Introduction

SUPERCONDUCTING magnetic energy storage systems (SMES) have been proposed for use in flexible ac transmission systems (FACTS) because of their high efficiency and rapid response to power demand [1], [2]. This paper proposes a novel control of a hybrid converter used in a high power high-temperature superconducting magnetic energy storage system (HT-SMES), consisting of a high-temperature superconducting coil as the energy storage device and two power converters to provide the power-conditioning system capability (PCS) when the SMES is connected to the utility system.

Low-temperature SMES requires liquid hydrogen for its operation, which makes it expensive to operate. With the availability of high-temperature superconducting coil, only liquid nitrogen is required, which is readily available and much cheaper than liquid hydrogen. In our experiment, the high-temperature superconducting coil is made using Ag-clad $(Bi, Pb)_2Sr_2Ca_2Cu_3O_{10+x}$ wire manufactured by Australian Superconductors Ltd., using the oxide powder-in-tube technique. The coil is placed in a tank containing liquid nitrogen.

The PCS can be either a current source converter (CSC) or a VSC connected to a two-quadrant boost chopper. Since the superconducting coil is inherently a current source, the CSC has more advantages to be a PCS for SMES than the VSC [3], [4].

High-power SMES system needs a converter with a high power rating. Current-source converters (CSCs) can be paralleled to directly constitute a multimodule converter to meet the high-power requirement. Because of the high-power requirement, the switching frequency of the high-power devices is restricted to a relatively low switching rate. On the other hand, low contents of harmonics can be obtained in a multimodule CSC based on phased-shifted sinusoidal pulse width modulation (SPWM) control [4], [5]. In [4], an almost harmonic-free ac current is obtained by employing 48 CSC modules while the switching frequency is 2 ON-OFF per cycle.

Though a multimodule CSC consisting of a large number of modules gives us the advantages of low-harmonic ac current and high total power rate, if too many modules are used, this will increase the cost of the converter and complicate the construction and maintenance of the system. It is, therefore, desirable that a small number of modules are used, while maintaining good waveform and high-power features of the multimodule converter. In this paper, we propose the use of a hybrid converter to achieve the above objective.

The hybrid converter can be made up of two VSCs–one of them usually uses a low-frequency switching device to control most of the high-power output of the SMES while the other uses relatively low power but is operated at a much higher frequency to control the remaining part of the power such as the harmonics [6]–[8]. Thus low harmonics can be obtained while the total switching loss can be kept small. They are called hybrid because they make use of two voltage source converters with a different power level and switching frequency.

Another type of hybrid converter is composed of a CSC and a VSC [9].

In this paper, we propose a hybrid converter consisting of a multimodule CSC combined with a VSC for use as the PCS of the HT-SMES. The name "hybrid" comes from the mixing of these two different circuit topologies. The current-source part consists of four CSC modules controlled by low switching frequency phase-shifted SPWM. The voltage-source part is a normal VSC working on high switching frequency and current tracing control mode to control the harmonics generated by the CSC.

II. Topology of the Hybrid Converter

The main circuit of the proposed hybrid four-module CSC is shown in Fig. 1. Each current converter module is a six-valve CSC. The chosen switching device is the gate turn-off thyristor

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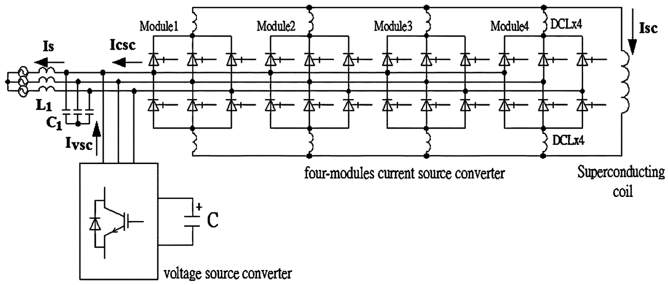

Fig. 1. Main circuit of the hybrid converter for SMES.

(GTO), or insulated gated controlled thyristor (IGCT), which has a high power rating and low working frequency. The ac side of each module is connected together directly. The dc sides of the converter modules are paralleled through dc inductors. Then, these four converter modules are connected to the superconducting coil. The ac side of the VSC is connected to the ac sides of the CSC directly. The VSC uses insulated-gate bipolar transistors (IGBTs), which allows it to work at a much higher switching frequency.

For a CSC, the large output capacitor is indispensable for filtering and for commutating the switches. However, the large capacitor can resonate with the inductance of the power system, which can cause amplification of the low-order harmonics. In the proposed topology of Fig. 1, we avoid this potential problem by replacing the large output capacitors with the dc capacitor used in the voltage source converters (VSC) to provide the necessary reactive power for the commutation of the switches. Hence, only a much smaller capacitor is required for the output filter and, therefore, the resonance problem can be avoided.

The VSC absorbs the harmonics generated by the four-module CSC and looks like an embedded active power filter. The power rating of the VSC is much smaller than the CSC and the total switching loss is low. Using such a topology, the harmonic generated by the SMES is low even when a few current source modules are used.

III. Control of the CSC

A. SPWM Control of the CSC

Phase-shifted SPWM control is applied to this four-module CSC to reduce the harmonics while the switching frequency is low (it can be as low as 150 or 300 Hz). Fig. 2 gives the control logic diagram. The SPWM for the CSC needs the trilogic signal [11].

The signals are produced as follows:

$$X_a = \mathrm{sign}[v_{ma}(wt) - v_c(w_s t + \phi_s)]$$
$$X_b = \mathrm{sign}[v_{mb}(wt) - v_c(w_s t + \phi_s)]$$
$$X_c = \mathrm{sign}[v_{mc}(wt) - v_c(w_s t + \phi_s)]$$
$$\mathrm{sign}(x) = \begin{bmatrix} 1 & x \geq 0 \\ -1 & x < 0 \end{bmatrix}$$
$$Y_a = \frac{1}{2}(X_a - X_b)$$
$$Y_b = \frac{1}{2}(X_b - X_c)$$
$$Y_c = \frac{1}{2}(X_c - X_a).$$

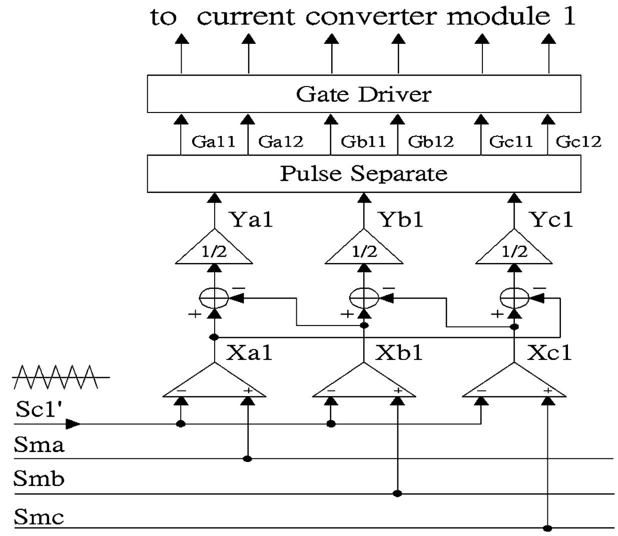

Fig. 2. Control logic of the CSC modules.

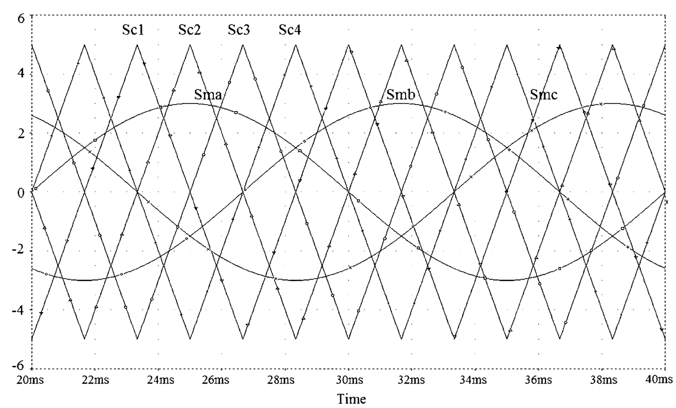

Fig. 3. Triangle carrier and modulation signal of each current module.

As shown in Figs. 2 and 3, the modulation signals of each module are the same, the amplitude of the carrier of each module is equal, and the phase angle of the carrier of each module is evenly shifted. In a four-module converter, at three-times carrier frequency, the carrier phase-angle difference between adjacent modules is

$$(360^\circ/3)/4 = 30^\circ.$$

Under such a control method, the average dc current of each module differs from each other largely because of the different carrier angle when the frequency of the carrier is low. Fig. 4 gives the dc current of each of the four modules. When the modulation index increases, the current unbalance between each module can be quite serious. Such unbalance can make this direct paralleled multimodule converter almost infeasible.

B. Current Balance Control of the Four-Module Converter

A carrier-swapping method is applied to this phase-shifted SPWM controller to equalize the dc current-share of these four modules, as shown in Fig. 5. The phase-shifted triangle carriers of the four modules are swapped one at a time at every positive

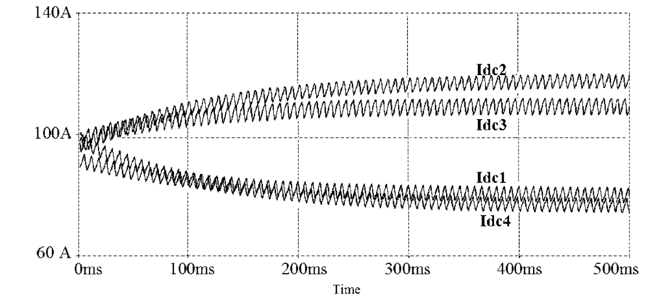

Fig. 4. Simulation waveform of dc current of each module with phase-shifted SPWM without current balance control $\mathrm{M} = 0.6, \alpha = 30°$.

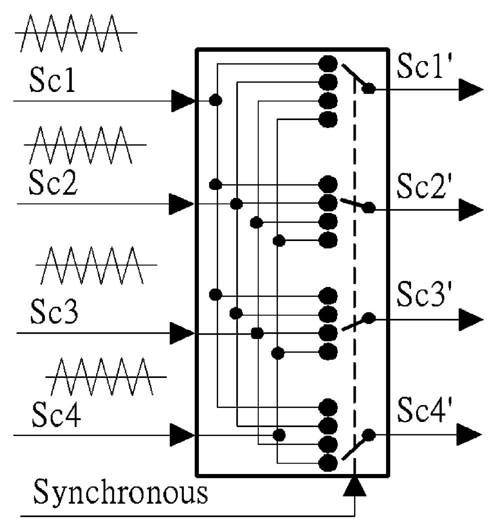

Fig. 5. Triangle carrier swapping logic.

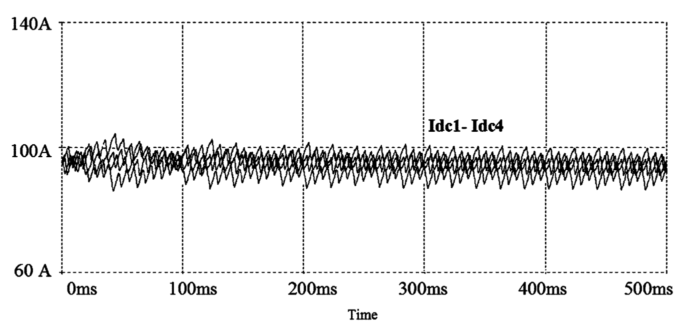

Fig. 6. Simulation waveform of dc current of each module, phase-shifted SPWM with carrier-swapping $\mathrm{M} = 0.6, \alpha = 30°$.

edge of the synchronous signal. For example, Sc1' (triangle carrier signal of module 1) is swapped as follows:

$$\rightarrow \mathrm{Sc1} \rightarrow \mathrm{Sc2} \rightarrow \mathrm{Sc3} \rightarrow \mathrm{Sc4} \rightarrow \mathrm{Sc1}.$$

Current unbalance caused by the carrier phase-shifting can then be eliminated. Fig. 6 shows the dc current of each module with carrier swapping. The total ac current of the four modules before the output filter is shown in Fig. 7.

C. *Reduce the Swapping Period*

The carrier-swapping period is the same as the period of the synchronous signal–it equals to the period of the modulation

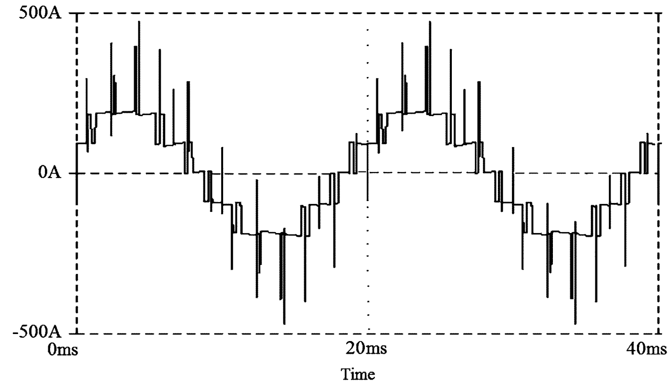

Fig. 7. Simulation waveform of total ac current of four modules.

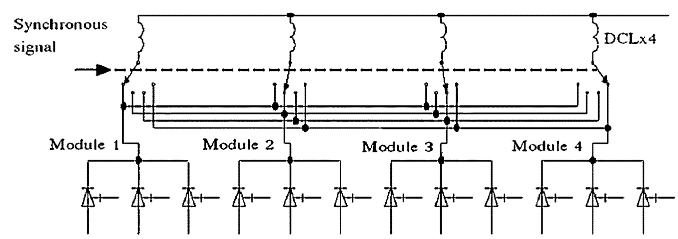

Fig. 8. Switch the DC inductors to average the DC current.

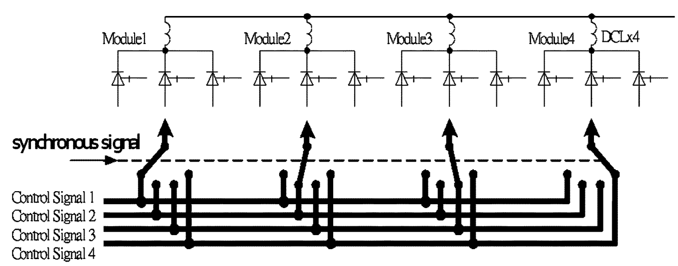

Fig. 9. Switch the gate control signal to average the dc current.

signal. Considering the additional switching action needed by the swapping operation, this is the shortest period found while using carrier swapping to balance current sharing. At such a period, the value of the dc inductors in each module should be about 100 mH. Too small of a value results in a larger ripple in the dc current of each module and serious distortion in the total ac current of the four modules. Smaller ripple in the dc current and less distorted total ac current can be obtained using larger dc inductors, but this method is not economical.

In Fig. 8, the carrier signal is not swapped but the dc current of each module is averaged by switching the dc inductors. Each dc inductor is swapped between the four modules as follows:

$$\rightarrow \text{Module } 1 \rightarrow \text{Module } 2 \rightarrow \text{Module } 3 \rightarrow \text{Module } 4 \rightarrow \text{Module } 1.$$

The swapping frequency can be very high and, hence, the dc current ripple can be small even if the dc inductors are small. But a lot of auxiliary switches are needed in this method; it is still not an economical and feasible way.

Since the main circuit of each module is completely the same, the difference is only in the control signal. If we switch the control signal between the four modules, the effectiveness of current averaging is the same as switching the dc inductors. Fig. 9

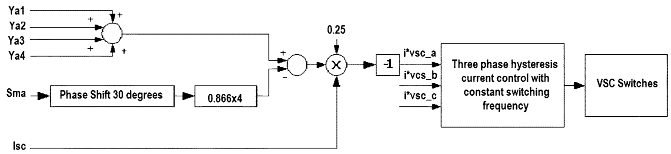

Fig. 10. Block diagram for phase A of the controller of the VSC.

shows the control logic of this method. The gate control signals of the upper three switches and the lower three switches in each module can be swapped separately (Figs. 8 and 9 only show the upper side switch), then a higher swapping frequency can be used without increasing the additional switching operation required by the swapping operation. This is an improved method compared with the carrier swapping method. The disadvantage of this method is that it needs a more complex logic circuit to switch the control signal, but logic circuit is relatively cheap and, hence, it is an insignificant disadvantage.

IV. Control of the VSC

By reducing the swapping period using the way it was mentioned above, the ripple of dc current can be kept low even when the dc inductor is small, the total ac current of the four modules can trace the control signal $(Y_1+Y_2+Y_3+Y_4)$ with less distortion, since the SPWM-controlled converter is a linear amplifier [4].

The ac current of phase A of the four-module current-source converter is

$$I_a = (Y_{a1} + Y_{a2} + Y_{a3} + Y_{a4}) \cdot K$$

where K is the amplification coefficient of the four-module current converter

$$K = I_{\rm sc}/4$$

where $I_{\rm sc}$ is the current of the super-conducting coil, If the modulation signal of phase A is

$$S_{ma} >= M \cdot \sin(\omega t)$$

where M is the modulation index of SPWM. The fundamental part of I_a is

$$I_{a1} = 0.866 \cdot 4 \cdot M \cdot \sin(\omega t + 30^\circ) \cdot K.$$

Thus, the harmonic part of the I_a is as follows:

$$I_{ah} = ((Y_{a1}+Y_{a2}+Y_{a3}+Y_{a4}) - 0.866 \cdot 4 \cdot M \cdot \sin(\omega t + 30^\circ)) \cdot K.$$

The current signal, which the VSC should trace, is

$$I_{\rm vsc} = (-1) \cdot I_{ah}.$$

Fig. 10 gives the strategy for phase A of the VSC controller. The hysteresis control described in [12] is adopted in the simulation. The simulation waveforms of I_a, I_{a1}, and I_{ah} are shown in Fig. 11.

Under such control, the VSC absorbs the harmonics of the ac current of the four-module CSC and, hence, only the fundamental part of the total ac current of four modules is injected into the utility power network. The VSC, therefore, acts as an active filter absorbing the harmonic current generated by the four-module CSC. In this case, the current error signal can be obtained easily because I_a is foreknown. This is one of the advantages of this embedded active power filter compared with the common separate active power filter.

To obtain an ac source current (I_s) with the lowest harmonics, the ac current of the VSC $(I_{\rm vsc})$ should approach I_{ah} as close as possible and the total ac current of the four-module CSC $(I_{\rm csc})$ should approach $(Y_{a1}+Y_{a2}+Y_{a3}+Y_{a4}) \cdot K$ as close as possible. By adopting proper current tracing control and using high-speed switches, the former requirement can be satisfied. To satisfy the latter one, the dc current unbalancing of the four-module CSC and the dc current ripple of each module should be reduced as much as possible, so the current balance control and the way to reduce the swapping period mentioned in the former section are necessary.

V. Simulation

Fig. 12 shows the simulation waveform of $I_{\rm csc}$, $I_{\rm vsc}$, and I_s. The VSC is under hysteresis control to trace the harmonic current signal which is $(-1) \cdot I_{ah}$. The value of the output filter components is $L_1 = 150$ μH, $C_1 = 10$ μF, and the current of superconducting coil is 140 A. We can see that the current of the ac source (I_s) is nearly harmonic free even if the output capacity is very small. Fig. 13 gives the waveform of I_s when the VSC is not connected to the four-module CSC and the output capacity is 850 μF; a significant value of the seventh harmonic can be found in this condition.

VI. Conclusion

A multimodule current-source-based hybrid converter is investigated in this paper. The properties of the CSC, such as fast response and bidirectional power control, can be obtained and are very suitable for an SMES system. Because of the directly paralleled VSC, the value of the output capacitors can be reduced to a much smaller value, the resonance in the output filter of the CSC is avoided, and the harmonics of the utility source current are kept low. The VSC in this hybrid converter is an embedded active power filter. Its control algorithm is simplified compared with the common separate active power filter because the current tracing signal is foreknown. The adoption of the four current modules not only improves the total power rating but also reduces the harmonics to a known value while phase-shifted SPWM control is employed; thus, the power rating of the VSC can be reduced to a much smaller value compared with the CSC. In many applications, parallel operation is preferred because it offered some degree of fault tolerance compared with other single-module systems. The SPWM control of the multi-module current converter also gives us more flexibility of the control of SMES when compared with the square-wave control mode.

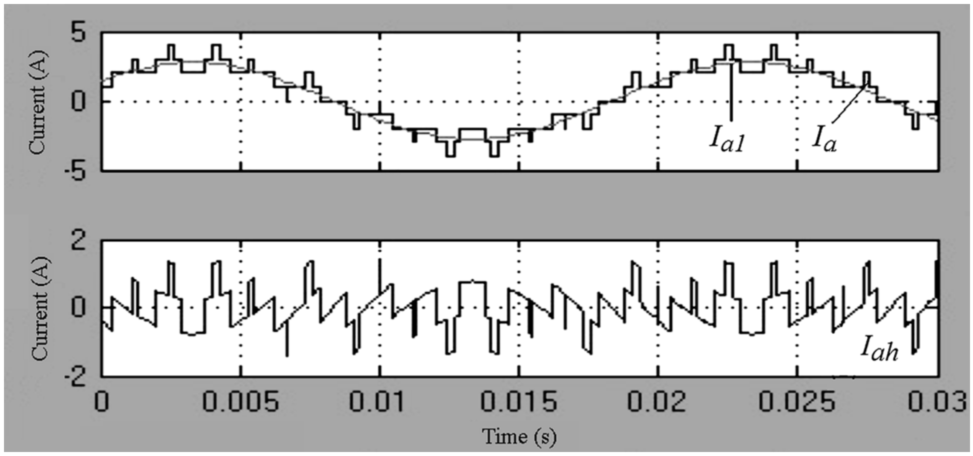

Fig. 11. Simulation waveform of the total current of the four-module converter (I_a), its fundamental part (I_{a1}), and its harmonic part (I_{ah}).

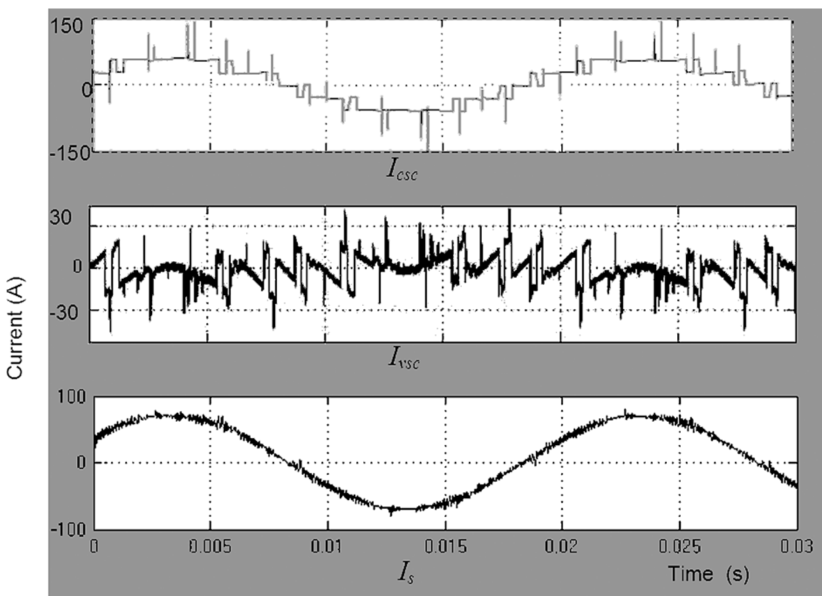

Fig. 12. Simulation waveform of the ac current of the four-module CSC $(I_{\rm csc})$, VSC $(I_{\rm vsc})$, and the utility power source (I_s).

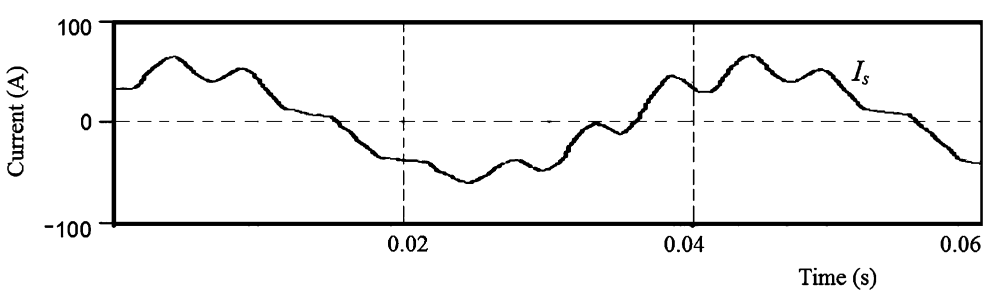

Fig. 13. AC source current of a pure four-module CSC (without the VSC).

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