Utilization of the Series Resonant DC Link Converter as a Conditioning System for SMES

Karl W. Marschke, *Member, IEEE*, Paulo P. A. Caldeira, *Member, IEEE*, and Thomas A. Lipo, *Fellow, IEEE*

***Abstract*—In this paper a new superconductive magnetic energy storage (SMES) system utilizing a high-frequency series resonant dc link power converter of high efficiency as the conditioning converter is presented. This system generates a high-frequency (20 kHz or more) resonant current in a series link and switching is done at zero current instants, reducing switching losses to a minimal value. Through the utilization of an adequate control strategy, the input power factor can be fully adjusted during the charging, storing, and discharging modes of the SMES, improving the overall system efficiency. Different semiconductor devices are employed as the switching elements of the resonant converter and switching losses are established for each case. Experimental results from a monophase and three-phase system verified the results obtained from digital simulation.**

I. Introduction

A superconductive magnetic energy storage (SMES) system consists basically of a large superconducting storage coil, its cooling system, and an ac-to-dc power converter system, typically called the power conditioning system. SMES has recently drawn attention not only as a potential means for load leveling but also as a power system stabilizer [1], [2].

For utility load leveling a SMES system has two operational modes: 1) during low demand periods it absorbs surplus ac power as stored energy in the dc SMES coil and 2) during high-demand periods the energy stored into the superconducting coil is then delivered back to the utility network through the ac–dc converter.

The basic requirements for the power conditioning system that interfaces the utility system and the SMES coil are as follows:

1) Ability to operate at very high currents or ability to operate with parallel devices or parallel bridges
2) Power transfer in both directions due to load-leveling purposes
3) High efficiency
4) Fast changes in power level or direction
5) Minimum reactive power consumption under normal operation and supplying or taking reactive power from the utility network as necessary at the same level of real power
6) Ability to independently control the real and reactive power

In [1] and [2], different power conditioning circuits have been studied for this purpose utilizing GTO converters. In contrast to these hard-switched converters, high-power density ac/ac converters utilizing resonant link schemes have recently been developed. These converters not only have high power density but also possess very low switching losses, since the switching of the devices is made at zero voltage or zero current instants.

In general, switching schemes for resonant converters can be classified according to their resonant ac link and resonant dc link modes of operation. The resonant ac circuits utilize a parallel or series resonant link, impressing both polarities of ac voltage and current on the link thus requiring bidirectional switches in the input and output converters [3]–[5]. The resonant dc circuits can also utilize a parallel [5] or series resonant link.

The series dc link circuits realize pulsating dc currents in the link by adding dc offsets to the ac resonant currents. A high-frequency series resonant dc link, ac-to-ac power converter is proposed in [6] utilizing only 12 unidirectional switches. As shown in Fig. 1(a) the capacitor C_0 and L_0 cause a resonant high-frequency current i_s to flow from the input ac source to the load while the inductance L_d provides a dc bias current (I_d) to the resonant current i_s. The resonant pulses can then be distributed in the input phases by a pulse density modulation (PDM) control strategy as shown in Fig. 1(b). The four thyristors conducting in the two bridges turn on and off at zero current instants, reducing switching losses significantly. In [7] a design methodology and a control strategy are established for the complete ac/ac drive system.

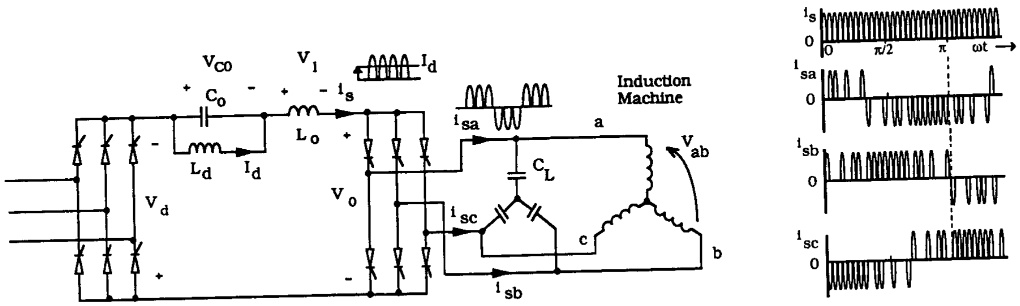
This paper proposes the utilization of a series resonant dc link power converter as the conditioning system for SMES. The superconducting coil L_d provides the bias dc current for the resonant pulses. The system provides bidirectional power flow enabling three operational modes, which are charging, storing, and discharging. Operation with extremely low switching losses (both at the instants of turn on and turn off of the devices), continuous operation from start-up with unity fundamental input power factor (if desired), independent control of real and reactive power in all modes of operation, and low harmonic

Manuscript received May 31, 1990; revised February 21, 1992.

K. W. Marschke is with Marquip, Inc., Madison, WI 53706.
P. P. A. Caldeira is with Philips Research Laboratories, Briarcliff Manor, NY 10510.
T. A. Lipo is with the Department of Electrical and Computer Engineering, University of Wisconsin, Madison, WI 53706.

IEEE Log Number 9200724.

0885-8993/92$03.00



(a) (b)

Fig. 1. AC/AC drive system utilizing the series resonant dc link.

content make the system suitable to be utilized in high-power applications such as a power-conditioning circuit for SMES with excellent overall efficiency.

Different semiconductor device structures, like gate turn-off thyristors (GTO) in the gate assisted turn-off mode [8]–[10] and silicon-controlled rectifiers (SCR), are utilized as the switching elements of the resonant converter. Switching losses are established and compared for each case. Resonant frequencies up to 32 kHz were obtained in the experimental monophase and three-phase model.

A control strategy utilizing sigma–delta modulators for continuous control of the active and reactive input/output power during the charging/discharging modes is proposed in this paper. Complete and continuous control of the input power factor and low harmonic distortion can be cited as advantages of this control scheme.

II. System Analysis

Fig. 2 shows the three-phase resonant converter utilized as a power conditioning drive in a superconductive energy storage system. The bias inductor L_d in the original system is substituted by the superconducting element. The resonant frequency of the system is determined by the capacitor C_0 and the inductor L_0. In the generation of the first resonant pulse the capacitor C_0 and the inductors L_0 and L_d are completely discharged. Through an adequate choice of the initial output of the switching matrix a positive voltage is applied to the link resulting in the generation of the first resonant pulse. When the resonant current i_s returns to zero, one or two switches of the three-phase bridge are turned off. During this interval the bias current I_d flows into C_0 causing a linear discharge of this voltage, resulting in the application of a controllable dV/dt over the switches. When an adequate positive voltage (V_{swt}) is observed over the switches they are triggered and a new pulse is generated. This positive bias voltage over the devices increases the system stability ensuring that the resonant pulses reach zero in each cycle and the application of a negative voltage over the devices equals V_{swt} at turn-off. It can be observed that no special operating mode or circuit is needed during start-up, since from the instant of the first resonant pulse, the system is already in the charging mode.

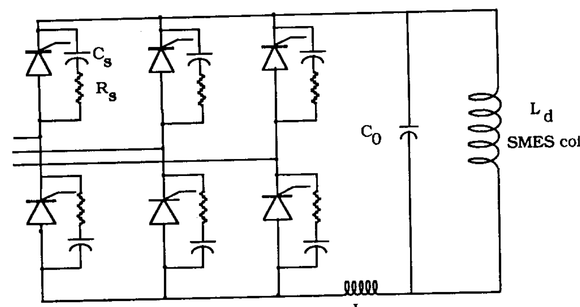


Fig. 2. Series resonant dc link converter utilized as a power conditioning system for an SMES.

The resonant current and resonant capacitor voltage can be expressed as

$$i_s(t) = I_d - I_d \cos \omega_0 t + \left(\frac{V_{swt}}{Z_0}\right) \sin \omega_0 t \tag{1}$$

$$v_c(t) = (V_d) - I_d Z_0 \sin \omega_0 t - (V_{swt}) \cos \omega_0 t \tag{2}$$

where

$$Z_0 = \sqrt{\frac{L_0}{C_0}} \qquad \omega_0 = \frac{1}{\sqrt{L_0 C_0}} \qquad V_{swt} = V_d - V_{c0}. \tag{3}$$

V_d is the output voltage of the resonant converter and V_{c0} represents the initial voltage over the resonant capacitor at the beginning of the current pulse. During the interval between pulses the offset bias current I_d circulates through the resonant capacitor C_0, which discharges linearly from $V_d + V_{swt}$ to $V_d - V_{swt}$ during a time interval equal to

$$2t_p = 2\,\frac{V_{swt} C_0}{I_d}. \tag{4}$$

The resonant period (T_r) can be defined as

$$T_r = T_0 + 2t_p \tag{5}$$

where

$$T_0 = \frac{1}{f_0} = \frac{2\pi}{w_0}. \tag{6}$$

III. System Control

It can be observed that the average value of the resonant pulses is approximately equal to the value of the bias current I_d. If I_d becomes greater than the average value of i_s, the resonant pulses will not reach the zero crossing point e, and an unstable operational condition would occur. Regulation of this bias current is then necessary. Through an adequate switching sequence, regulation of I_d at the desired level is obtained during the three operational modes.

The power control scheme utilizes two commands:

1) Real power command, which determines the value of I_d current to be maintained in the superconductive coil
2) Reactive power comman, which determines the phase shift between the input current and input voltage

For the determination of the proper switching matrix of the resonant converter the bias current I_d is initially compared with a reference current $I_{d\text{ref}}$, which is determined from the real power command, which generates an error signal (ϵ). A proportional-integral regulator and a three-phase sine multiplier are then utilized to produce current references for each phase, which are at a certain angle (Ω) with respect to the input voltages determined by the reactive power command. The comparison of these three current references with the actual input currents through a sigma-delta modulator, generates three errors that are used to determine the phase to be triggered to generate the next resonant pulse. The sum of these three errors must satisfy the relation

$$\epsilon_{a\text{in}} + \epsilon_{b\text{in}} + \epsilon_{c\text{in}} = 0 \tag{7}$$

which establishes the triggering principle based on the fact that the three errors can never have the same polarity. The triggering principle is then the following:

1) The thyristor in the phase having the larger error out of the two phases of the same polarity is chosen to be triggered.
2) The phase corresponding to the error with the opposite polarity error is selected as the other triggering phase.

Actual values for the real and reactive powers are then compared with the real and reactive commands, generating errors that are utilized in an external compensation feedback loop with the purpose to correct the reference current I_d and the phase shift Ω to obtain the desired power levels.

The control of the fundamental input power factor is obtained through adequate distribution of the resonant pulses in the three phases (PDM). Regulation of the current I_d at the desired level in the SMES coil is guaranteed in the three operational modes through the application of the appropriate voltage to the resonant link. Consequently, the real and reactive powers of the SMES are fully controlled. During the energy-storing process the superconducting element should be shorted, which can be accomplished by turning-on two switches connected to the same phase. No extra switches are then necessary for this mode. The complete control diagram is shown in Fig. 3.

This control system represents a very flexible alternative for controlling a SMES system, since it can be adjusted to any requirement in terms of input power factor or input power. If a situation of maximum input power is desired during the charging process, the bias current reference ($I_{d\text{ref}}$) would be set in its total final value. In this mode, the input phase selection would be determined by the most positive and negative phases in each instant. If an operational condition requires control of power factor and input power, $i_{d\text{ref}}$ should then slowly change during the charging process, determining the charging speed and consequently limiting the i_d variation. The displacement angle (Ω) should then be zeroed in the control scheme.

If unity displacement power factor and low distortion factor without input power control is the desired feature for a SMES system, then a geometric PDM control strategy can be utilized. In this simple open-loop control structure, the pulses are distributed in each phase accordingly to the input voltage amplitude in a geometric standard manner. This control strategy can be easily implemented, since no sigma–delta modulators or proportional-integral regulators are required.

Figs. 4 and 5 show simulated results for the proposed system. Fig. 4(a) shows the input current and voltage in phase a during the start up and charging process in the case of maximum input power requirement. Fig. 5(a) shows the input current and voltage in phase a during the start-up charging process in the case of the unity power factor requirement. Figs. 4(b) and 5(b) present the spectral component for each case. Fig. 5 shows the resonant current i_s and the charging current I_d during the storing process.

IV. Reduced Switching Frequency Strategy

With the objective of reducing the switching frequency when GTO's are utilized as the main switching devices in the resonant converter without altering the resonant frequency, an improved switching strategy can be utilized. When the end of the resonant pulse is reached, the new switching requirement obtained from the control scheme is compared with respect to the previous one. If both switches are required to be turned on in the next pulse, only one switch is turned off and then back on during the beginning of the following pulse. If both switches need to be on again in the following pulse the device that did not suffer commutation in the previous pulse is turned off then back on. If two different switches needed to be utilized in the next pulse then both switches are turned off. In this case the maximum switching frequency that a device could be subjected to is always half of the resonant frequency.

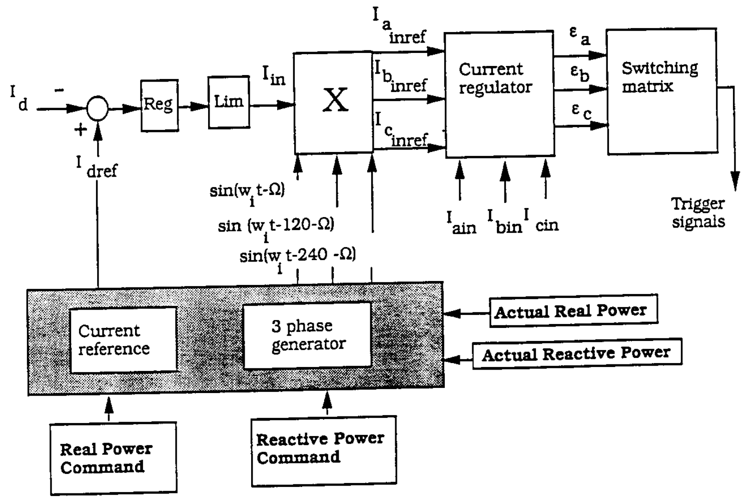


Fig. 3. Real and reactive power control in the SMES system.

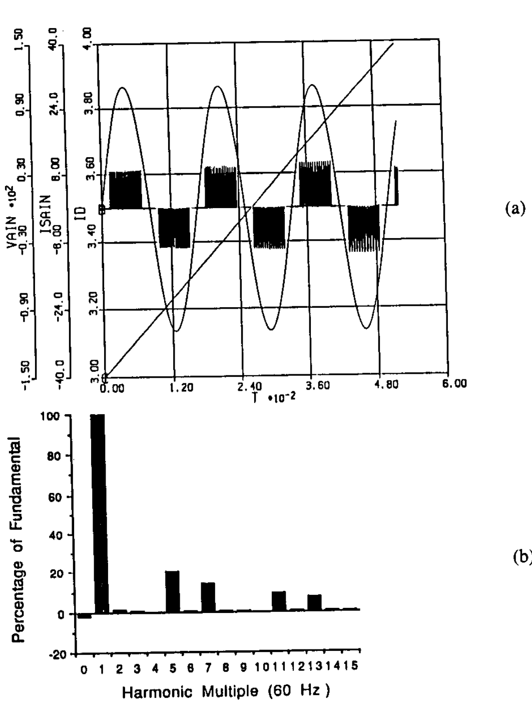


Fig. 4. (a) Input voltage, current, charging current. (b) Spectrum of the input current for the maximum power control mode.

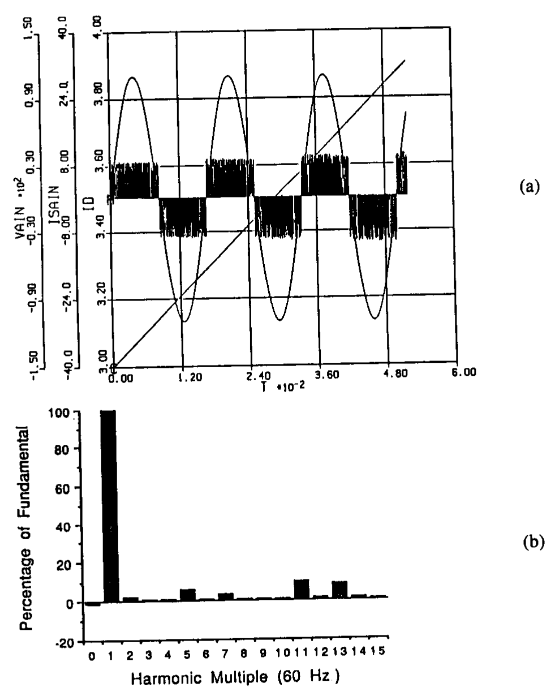


Fig. 5. (a) Input voltage, current, charging current. (b) Spectrum of the input current for the Geometric PDM control.

V. Design Methodology

A complete design methodology and total loss calculation for a resonant dc current link system has been presented in [7]. The same design methodology can be utilized for the resonant converter utilized as the conditioning system for the SMES element. Since L_d can be considered an ideal lossless element, the reduction of the initial voltage over the resonant capacitor at the beginning of each resonant pulse caused by the equivalent series resistance (ESR) of the bias inductor will not be presented herein. Consequently, no external exciters circuits or compensators are necessary for this purpose, even in high-power applications, and a negative voltage naturally occurs across the switches at the end of the resonant pulse.

The same selection criteria for L_0, C_0, and V_{swt}, as in [7], can be applied considering a resonant frequency of approximately 30 kHz and a bias current of 15 A, resulting in

$$Z_0 = \frac{V_{swt}}{I_d} = 0 \tag{8}$$

$$V_{swt} \geq \left[\frac{I_d t_{rr}}{C_0} + \frac{2\sqrt{3}\pi}{f_r} (V_{inpk} f_i \right] \frac{1}{K_r} = 35\ V \tag{9}$$

where K_r represents the resonant pulse peak reduction due to the ESR of the resonant inductor and t_{rr} the reverse recovery time of the switching device. The resonant elements can be chosen as

$$C_0 = 0.5\ \mu F \quad \text{and} \quad L_0 = 60\ \mu H.$$

The peak resonant current and voltage can be calculated utilizing the results in [7], resulting in

$$i_{speak} = I_d + \sqrt{I_d^2 + \left(\frac{V_{swt}}{Z_0}\right)^2} = 36.21\ A \tag{10}$$

$$v_{cpeak} = V_d + Z_0 \sqrt{I_d^2 + \left(\frac{V_{swt}}{Z_0}\right)^2} = 330\ V \tag{11}$$

$$\Delta v_{cmax} = 2V_d + 2Z_0 \sqrt{I_d^2 + \left(\frac{V_{swt}}{Z_0}\right)^2} = 660\ V. \tag{12}$$

The reverse recovery time available is $t_p = 5\ \mu$s.

Device Selection

The dc current link resonant system requires the utilization of unidirectional current switches naturally commutated, since the current pulse drops naturally to zero due to the resonant scheme utilized. Since switching losses are extremely reduced, commutation delays, and, most importantly, turn-off time, are parameters to be considered in the selection of the devices.

Initially, GTO's in the gate assisted turn-off mode (GATT) were utilized as the main switches of the resonant converter. The utilization of the turn-off ability of the gate circuit (GATT), acting as an additional recovery mechanism, improves the commutation process efficiency and time, requiring the utilization of small snubbers.

In an attempt to improve the switching characteristics of a gate-assisted turn-off GTO, series and parallel diodes were incorporated to the basic device in order to avoid the reverse and forward recovery processes reducing significantly the turn-off losses and time. Reverse voltage stress over the GTO's are reduced to practically zero in this structure.

Fig. 6 shows the GTO being utilized in the gate assisted turn-off mode, employing a series and an antiparallel diode. In the turn-on process a gate current consisting of a large front porch and a continuous back porch is required. An improved turn-on gate pulse, and a very small energy discharge of the snubber capacitor reduce the on-time restriction of the device. Since controllable and reduced current slopes are a natural characteristic of the resonant current pulse, reduced turn-on and turn-off times are obtained.

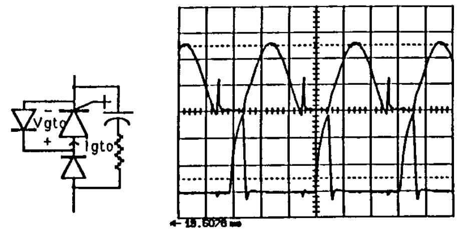


Fig. 6. Current and voltage over the GTO in the resonant link converter 10 A/div, 50 V/div, 10 μs/div.

The fast-recovery diode and the antiparallel diode in the GATT structure play an important role on the device turn-off process. The series diode avoids the reverse anode current and counterpolarization eliminating the GTO recovery process. In order to provide a path for the gate current to perform the depletion of the GTO central junction, reducing the forward recovery current, the antiparallel diode is utilized. Turn-off losses due to reverse recovery and forward recovery losses are substantially reduced.

A perfect synchronization of the gate turn-off pulse with the current zero crossing is of extreme importance in order to decrease the turn-off losses and turn-off time. After the anode voltage starts to rise during the turn-off process a small anode current (tail current) is still maintained as carriers are removed from the central region of the device.

The dv/dt of the reapplied anode voltage is basically controlled by the resonant capacitor and the dc bias current level during the turn-off process. The reapplied voltage slope is very low, consequently snubber action for its limitation is not required. The snubber circuit is only utilized with the purpose to produce an alternative discharge path for the reverse recovery current of the fast recovery series diode and a very small snubber resistor and capacitor are needed. Typical hard-switching minimum off time (approximately 35 μs) is reduced due to the zero current soft-switching process to approximately 5–8%. Tail times are one of the most limiting factors for an increase in switching frequencies.

Switching frequencies of the order of 32 kHz for peak currents of 45 A were obtained utilizing this switching structure. Typical waveforms with turn-on and turn-off times are shown in Fig. 6. The total switching losses (considering turn-on, conducting, and turn-off losses) for the GTO were 24 W (considering a 6.40-kW system). Turn-off times of 3–5 μs were measured.

Since the resonant pulse drops naturally to zero with a controllable di/dt, SCR's would constitute a natural choice for the switching devices. Switching frequencies of the order of 27 kHz for peak currents of 45 A were obtained utilizing this switching structure. Current and voltage over the SCR during the charging mode and during the turn-on and turn-off instants are shown in Fig. 7. Since the switching losses were extremely small, as can

(a)

(b)

(c)

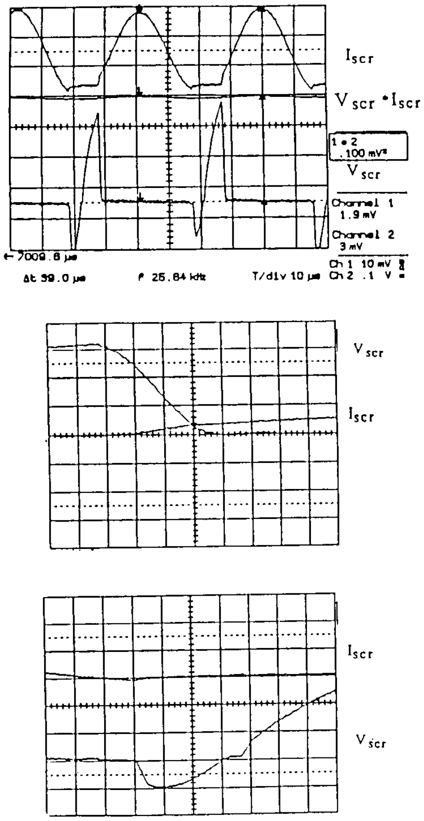


Fig. 7. Current and voltage over the SCR in the resonant link converter, 10 A/div, 50 V/div. (a) During the charging, 10 μs/div. (b) During the turn-on, 0.2 μs/div. (c) During the turn-off, 1 μs/div.

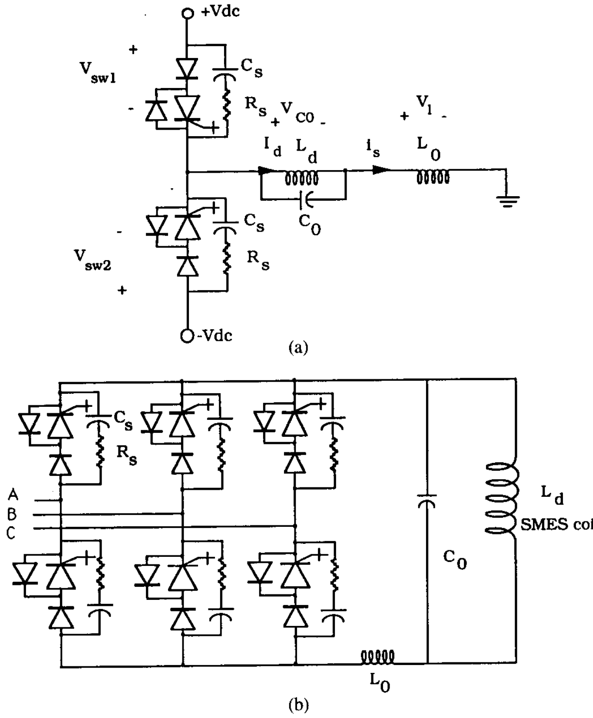


Fig. 8. Experimental monophase and three-phase models.

be seen in Fig. 7(a) (the result of the multiplication of V_{scr} by I_{scr} is shown in the middle line), they could not be measured accurately. The reduction in the switching losses is mainly due to a much faster turn-on process, reduction on the conduction losses and tail current. Turn-off times of 3–4 μs were observed for this operating condition. If a larger reverse recovery time is necessary at higher currents, V_{swt} could be increased accordingly. Consequently, SCR's constitute the best and simplest alternative for the switches in the resonant link converter and enables very high power ratings.

VI. Experimental Results

A monophase and three-phase model were assembled, where different switch alternatives were investigated, as shown in Fig. 8(a) and (b). Fig. 9(a) shows the complete charging, storing, and discharging process of a SMES element and Fig. 9(b) shows the dc current through the superconducting coil.

Fig. 10 shows the three-phase system where maximum input power is required. The geometric PDM was currently being implemented. A DSP56000 microprocessor was utilized as the controlling element for the SMES system. Driver circuits to be utilized with GTO's and SCR's

(a)

(b)

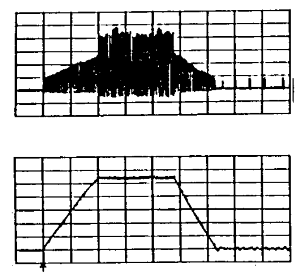


Fig. 9. (a) Charging, storing, and discharging process of an SMES system, 10 A/div, 5 ms/div. (b) Current in the superconducting element: 2 A/div, 5 ms/div.

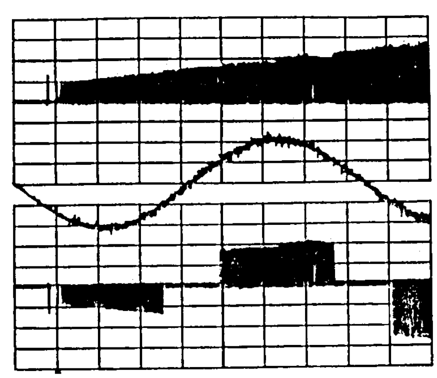


Fig. 10. Resonant current, input voltage, and input current in the three-phase experimental model utilizing maximum power mode control, 10 A/div, 50 V/div, 2 ms/div.

at switching frequencies of 35 kHz were developed for this specific application. For the resonant elements, ceramic capacitors and litz wire wound inductors were utilized in an effort to reduce ESR losses. The switching devices utilized in the experimental model are the following:

GTO	IR 81RDT
Series diode	IR R18CGF10A
Antiparallel diode	IR 40HFL100S05
SCR	IR 81RM80

Considering that the switching losses are dramatically reduced, the capacitor C_0 plays a minor role in the overall losses, and L_d can be considered as an ideal inductor with losses for the overall system consisting of

1) Conduction losses which can be calculated as

$$P_{\text{cond}} = V_{\text{on}}\left[I_d T_0 + \frac{4}{w_0}\left(\frac{V_{swt}}{Z_0} - I_d\right)\right]\frac{1}{Tr} = 17.55\ W. \tag{13}$$

2) Losses due to the snubber elements, which are very small, considering their function is only to provide a path for the discharge of the reverse recovery current of the SCR, and at turn-on the di/dt of the resonant current is relatively small ($R_s = 100\ \Omega$, $C_s = 0.022\ \mu$F).
3) Losses due to the resistive element of the input ac capacitors.
4) Loss caused by the ESR element (R_{L0}) of the resonant inductor, which can be calculated as

$$P_{RL0} = \frac{R_{L0}}{2}\left(3I_d^2 + \frac{V_{swt}^2}{Z_0^2}\right)\frac{T_0}{T_r} = 175.51\ W. \tag{14}$$

The estimated efficiency of the overall three-phase system is approximately 95%.

VII. Conclusion

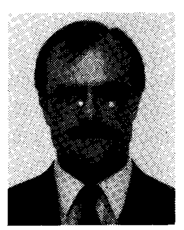
A series resonant dc current link converter to be utilized as a power conditioning rectifier in a SMES system has been proposed. A control strategy allowing complete control of the input active and reactive power, and continuous adjustment of the input power factor has been presented. Since the switches are commutated at zero current, extremely reduced switching losses, near elimination of snubber losses, improved harmonic content, bidirectional operation, high efficiency, and the natural ability of shortening the storage element are cited as advantages of the resonant converter in this application. Since the resonant current naturally decays to zero and a negative reverse voltage is present, SCR's constitute the ideal element to be utilized as the switching devices enabling very high power ratings. Resonant frequencies of 27 kHz with current peaks of 45 A, and efficiencies near 95% were obtained for the system.

Acknowledgment

The authors would like to acknowledge the support of the Wisconsin Electric Machines and Power Electronics Consortium (WEMPEC), the Wisconsin Electric Utilities Research Foundation, and International Rectifier for donating the GTO devices.

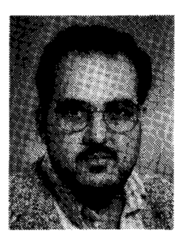
References

[1] J. Wang, J. Skiles, and R. Kustom, T. Ise, F. Tsang, and J. Cleary, "Studies of power conditioning circuits for superconductive magnetic energy storage," in *Proc. Power Specialist Conf.*, 1988, pp. 307–312.
[2] T. Ise, J. Skiles, R. Kustom, and J. Wang, "Circuit configuration of the GTO converter for superconducting magnetic energy storage," in *Proc. Power Specialist Conf.*, 1988, pp. 108–115.
[3] P. Sood and T. A. Lipo, "Power conversion distribution system using a resonant high-frequency ac link," in *Proc. IEEE-IAS Annu. Meet. Conf. Rec.*, 1986, pp. 533–541.
[4] H. K. Lauw, J. B. Klaassens, N. G. Butler, and D. B. Seely, "Variable-speed generation with the series-resonant converter," in *Proc. IEEE-PES Winter Meet. Conf. Rec.*, 1987/1988.
[5] D. M. Divan, "The resonant dc link converter—A new concept in static power conversion," in *Proc. IEEE IAS Annu. Meet. Conf. Rec.*, 1986, pp. 648–656.
[6] Y. Murai and T. A. Lipo, "High frequency series resonant dc link power conversion," in *Proc. IEEE IAS Annu. Meet. Conf. Rec.*, 1988, pp. 772–779.
[7] P. Caldeira, T. A. Lipo, Y. Murai, and S. Mochizuki, "Design and control of a series resonant dc link power converter drive," in *Proc. IPEC Conf. Rec.*, 1990, pp. 397–404.
[8] S. M. Tenconi and M. Zambelli, "The reverse blocking GTO as a very fast turn-off thyristor," in *Proc. IEEE IAS Conf. Rec. Annu. Meet.*, 1986, pp. 377–383.
[9] A. Mertens and H. C. Skudelny, "Switching losses in a GTO inverter for induction heating," in *Proc. IEEE Power Electronics Spec. Conf. Rec.*, 1989, pp. 91–98.
[10] L. Melasani, L. Rosette, and P. Tenti, "50 kW, 30 kHz GTO inverter for induction heating applications," *UEI Workshop on Induction Heating and Melting*, Hannover, Germany, 1988, pp. 105–109.



Karl W. Marschke (M'90) received the B.S. and M.S. degrees from the Milwaukee School of Engineering in 1988 and the University of Wisconsin—Madison in 1990 respectively, both in electrical engineering.

In 1991 he assumed the position of Senior Associate Engineer at Marquip Inc., Madison, WI. He is currently Assistant Operations Manager, Europe, for Marquip International Inc. His research areas are electrical machine drives and power electronics.



Paulo P. A. Caldeira (M'90) received the B.S. degree in 1979 from the Universidade Federal of Rio Grande do Sul, Brazil and the M.S. degree in 1985 from the Universidade Federal do Rio de Janeiro, Brazil, both in electrical engineering. He received the Ph.D. degree from the University of Wisconsin, Madison, in 1990.

From 1980 to 1986 he was employed by Petrobras S.A. as an Equipment Engineer and was involved in maintenance, specification, and design of drive systems for the oil drilling industry. He also worked as a Project Engineer in the area of drive systems for paper industry in Marquip Inc., Madison, WI, and he is currently a Senior Member of the Research Staff at Philips Research Laboratories, Briarcliff Manor, NY. His research interests are in switching and resonant converter technology, electromagnetic interference, and electric machine drives.

Dr. Caldeira is a member of the Tau Beta Pi Engineering Honor Society and the IEEE Power Electronics Society.

Thomas A. Lipo (M'64–SM'71–F'87) is a native of Milwaukee, WI. He received the B.E.E. and M.S.E.E. degrees from Marquette University, Milwaukee, WI, in 1962 and 1964 and the Ph.D. degree in electrical engineering from the University of Wisconsin in 1968.

From 1969 to 1979 he was an Electrical Engineer in the Power Electronics Laboratory of Corporate Research and Development of the General Electric Company, Schenectady NY. He became Professor of Electrical Engineering at Purdue University, West Lafayette, IN, in 1979, and in 1981 he joined the University of Wisconsin in the same capacity where he is presently the Grainger Professor for Power Electronics and Electrical Machines.

Dr. Lipo has maintained a deep research interest in power electronics and ac drives for over 25 years. He has received eleven patents and has 13 IEEE prize paper awards for his work including corecipient of the Best Paper Award in the IEEE Industry Applications Society Transactions for 1984. In 1986 he received the Outstanding Achievement Award from the IEEE Industry Applications Society for his contributions to the field of ac drives and in 1990 he received the William E. Newell Award of the IEEE Power Electronics Society for contributions to the field of power electronics.