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Controlling DC–DC Converter for Energy Cache Classification with Dynamic Desolation

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***Abstract*-this paper addresses a bidirectional dc-dc converter suitable for an energy storage system with an additional function of galvanic isolation. An energy storage device such as an electric double layer capacitor is directly connected to a dc side of the dc-dc converter without any chopper circuit. Nevertheless, the dc-dc converter can continue operating when the voltage across the energy storage device drops along with its discharge. Theoretical calculation and experimental measurement reveal that power loss and peak current impose limitations on a permissible dc-voltage range. This information may be useful in design of the dc-dc converter. Experimental results verify proper charging and discharging operation obtained from a 200-V, 2.6-kJ laboratory model of the energy storage system. Moreover, the dc-dc converter can charge the capacitor bank from zero to the rated voltage without any external recharging circuit.*Index Terms*-Bidirectional isolated dc-dc converter, electricdouble layer capacitors, energy storage system, loss analysis,starting procedure.**

I. INTRODUCTION

The Electric power generated by renewable energysources is unstable in nature, thus producing a bad effecton the utility grid. This fact spurs research on energy storagesystems to smooth out active-power flow on the utility grid [1],[2]. Fig. 1 shows a simplified existing energy storage systememploying a line-frequency (50- or 60-Hz) transformer, a PWMconverter, a bidirectional chopper, and an energy storage devicesuch as electric double layer capacitors (EDLCs) or lithium-ionbatteries. The transformer is indispensable for some applicationsthat require voltage matching and/or galvanic isolation betweenthe utility grid and the energy storage device. Replacingthe line-frequency transformer with a high-frequency isolateddc-dc converter would make the energy storage system morecompact and flexible.Various bidirectional isolated dc-dc converters have been proposedas the interface to energy storage devices with focus onautomotive or fuel cell applications. Most of the presented dc-dcconverters have asymmetrical circuit configurations to couplethe two dc links having largely different voltages, several tensvolts and several hundred volts [3]–[10 Fig. 2 depicts a bidirectional isolated dc-dc converter presentedin 1991 [11], [12]. It had two symmetrical single-phasevoltage-source full-bridge converters. It suffered from a lowefficiency because the first-generation IGBTs were used asswitching power devices at that time [11]. However, advancementin power device technology over the last decade hasenabled the dc-dc converter to operate at an efficiency as highas 97% by using the latest trench-gate IGBTs [13].

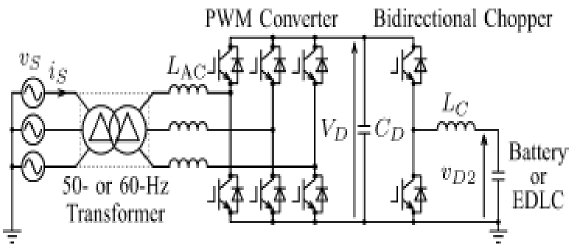

Fig. 1.Existing energy storage system employing a 50- or 60-Hz transformer for voltage matching and/or galvanic isolation.

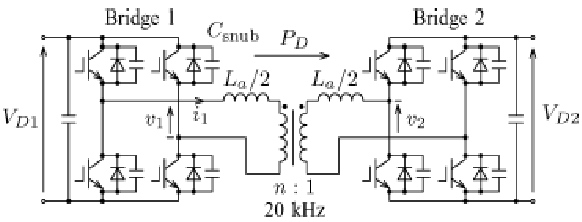

Fig.2 Bidirectional isolated dc-dc converter.

A similardc-dc converter in [14] has also achieved an efficiency of 97%.In addition, the use of silicon-carbide power devices in the nearfuture will raise it to 99%. Therefore, the dc-dc converter inFig. 2 has become a promising candidate as a power electronicinterface for an energy storage system. A bidirectional converterhas been discussed to exchange electric power betweena fuel cell, a battery, and a load [15], based on a three-portextension of the circuit presented in [11].Appropriately choosingthe transformer turn ratio enables to design the voltage ratingof the energy storage device, independent of the utility voltage.The energy storage device is directly connected to one of the dclinks of the dc-dc converter without any chopper circuit. Nevertheless,the dc-dc converter continues operating even when thevoltage across the energy storage device, drops along withits discharge.However, no paper has addressed the

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permissible voltagerange of in terms of power loss and peak current andno experimental verification has been confirmed, concerningFig. 3. This paper analyzes the relationshipsloss, the peak current, and in a dc-dc converter rated at10 kW and 20 kHz with fixed to 320 V. Then, the dc-dcconverter is designed, constructed, and tested to verify theanalysis. A 2.6-kJ energy storage system using an electrolyticcapacitor bank, together with the dc-dc converter, demonstrates stable charging and discharging operation. Besides, the dc-dcconverter can charge the capacitor bank from zero to the ratedvoltage without any external recharging or starting-up circuit.

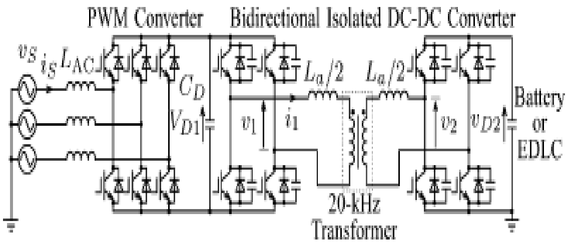

Fig.3 Alternative energy storage system based on the bidirectional isolateddc-dc converter.

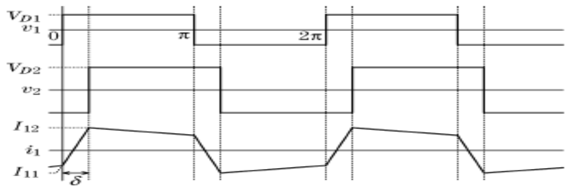

Fig.4 Simplified theoretical waveforms used to analyze the power losses whenVD_1< VD_2.

II.THE BIDIRECTIONAL ISOLATED DC–DC CONVERTER

A. Operation Principle and Simplified Theoretical Waveforms:

Fig.4 illustrates simplified theoretical waveforms of thedc-dc converter where. The two single-phasevoltage-source full-bridge converters produce square voltagesand the power transfer can simply be controlled byadjusting the phase shift between and as expressed by[11]

$$P_D = \frac{V_{D1}V_{D2}}{\omega L}\left(\delta - \frac{\delta^2}{\pi}\right) \qquad (1)$$

Where, is the switching angular frequency of the twosingle-phase voltage-source full-bridge converters, and is thesum of the transformer leakage inductance and that of theAuxiliary inductors.This paper defines a set of two instantaneousvalues of the current as "switching currents," and which are calculated asand is the instantaneousvalues of when

$$I_{11} = -\frac{(V_{D1} + V_{D2})\delta + (V_{D1} - V_{D2})(\pi - \delta)}{2\omega L} \qquad (2)$$

$$I_{12} = \frac{(V_{D1} + V_{D2})\delta - (V_{D1} - V_{D2})(\pi - \delta)}{2\omega L}. \qquad (3)$$

and,respectively, change its polarity from negative to positive.This paper refers a single-phase voltage-source full-bridgeconverter as a "bridge." In the following experiments, the transformerturn ratio is unity for the sake of simplicity.

TABLE I
CIRCUIT PARAMETERS OF THE DC–DC CONVERTER

Rated power		10 kW
Rated DC voltage	V_{D1}, V_{D2}	360 V
DC capacitor	C_D	7,100 µF
Unit capacitance constant	H	46 ms
Transformer core material		Finemet FT-3M
Transformer turn ratio	n	1 : 1
Transformer leakage inductance	L_{trans}	1.6 µH (1.6%)
Transformer winding resistance	R_{trans}	17 mΩ (0.13%)
Auxiliary inductor	$L_a/2$	20 µH (19%)
Auxiliary inductor core material		Ferrite (PC44)
Inductor winding resistance	$R_a/2$	20 mΩ (0.15%)
Snubber Capacitor	C_{snub}	0.01 µF (1.6%)
Switching Frequency	f	20 kHz

Based on single-phase 360 V, 10 kW, and 20 kHz.

*B. Experimental Circuit of the DC–DC Converter:*Table I summarizes the circuit parameters of the dc-dc converter. Four auxiliary inductors, totally having are connected in series with the transformer to obtain an inductanceof together with the leakage inductance ofthe transformer, .1 the inductance of 41.6is sufficient to maintain a control resolution of power transferaround 120 W because a time resolution of the controller operatedat 20 MHz is 50 ns, corresponding to 0.36 at 20 kHz.The following sections analyze relationships between powertransfer and power losses in the dc-dc converter. The powerlosses depend not only on the power transfer, but also the dc
Voltage. When drops along with discharge of theenergy storage device, power loss increases at a given powertransfer.

III. SNUBBER LOSS

A. Operating Points and ZVS Conditions

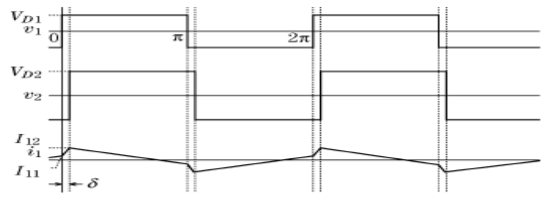

Fig.5 Waveforms when a positive I forces bridge 1 to operate in hard switchingmanner.

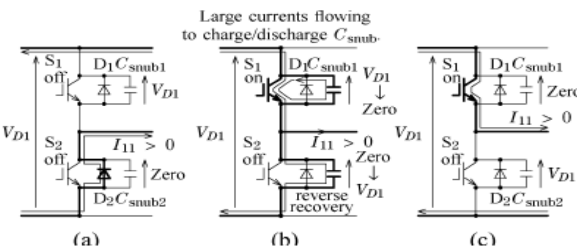

Fig.6 hard switching on a leg in Bridge 1: (a) just before the dead time ends(b) rapid charging/discharging of C, and C, (c) after commutation.

In Fig. 2, a snubber capacitor is connected in parallelwith each IGBT both to reduce switching loss

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and to damp outovervoltage. If the IGBT is turned on with its snubber capacitorcharged, the IGBT shorts out the snubber capacitor and dissipatesthe energy stored in the capacitor. This paper refers to thispower loss as "snubber loss."Each IGBT can be turned on in zero-voltage switching (ZVS)manner to generate no snubber loss when both dc voltages areequal and the power transfer is sufficient to ensurethe ZVS operation. However, when the IGBTis not necessarily turned on in ZVS manner,1Each inductor has an inductance value of L =4 = 10 _H. Two inductorswere connected in series to exhibit L =2 = 20 _H. Each of two sets of thetwo series-connected inductors was connected to each side of the transformer.The use of four separated inductors resulted from the authors' previous researchwhere the total inductance had to be adjusted in a range from 10 _H to 40 _H.

Fig.5shows simplified theoretical waveforms when theIGBTs in bridge 1 are turned on in hard-switching manner.The power transfer is less than that in Fig. 4 although the dcvoltages and are the same as those in Fig. 4. Theso-called "reverse recovery" occurs in the free-wheeling diodesin bridge 1 because the switching current is positive as canbe seen in Fig. 5. However, the four IGBTs in bridge 2 areturned on in ZVS manner. One can classify the turn-on processesof the IGBTs in bridges 1 and 2 into the following three:1) hard switching operation; 2) incomplete ZVS operationand 3) ZVS operation, depending on the power transfer,the phase shift , the dc voltages and , and the deadtime. The hard-switching operation and the incomplete ZVSoperation can take place only in one bridge, whose dc voltage islower than the other.

Switching Operation:

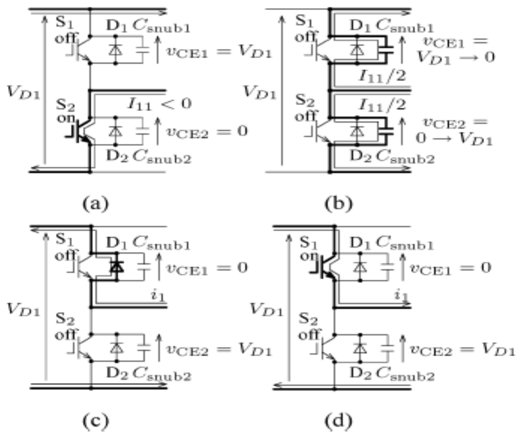

Fig.7 ZVS on a leg in Bridge1: (a) just before thedead time starts, (b) justafter the dead time starts, (c) diode freewheeling, and (d) current polarity alternates after the dead time.

Fig. 6 shows circuit modeswhen a leg (for example, consisting of and operates inhard-switching manner. The IGBTs in bridge 1 are turned on inhard-switching manner if the dc voltage is lower thanand the following equation is satisfied [11]

$$\delta \leq \frac{V_{D2} - V_{D1}}{2V_{D2}}\pi. \qquad (4)$$

The snubber capacitor of or has been charged at[see Fig. 6(a)] before the end of the dead time. Just afteris turned on, experiences reverse recovery[See Fig.6(b)]. Only an equivalent resistance of limitsthe charging/discharging currents, resulting in a joule loss ofNote that represents an amount ofenergy lost at one switching per leg whereThen, the snubber loss in bridge 1 is calculatedas

$$P_{\text{snub1}} = 4W_{\text{snub1}}f = 4C_{\text{snub}}V_{D1}^2 f. \qquad (5)$$

2) ZVS Operation: Fig. 7 shows circuit modes when alegin bridge 1 operates in ZVS manner. Before the dead time, thecurrent of is flowing in [see Fig. 7(a)]. Turning offstarts the dead time. The current flowing in is commutated to end A resonance begins between the inductance(see Fig. 2), and discharges fromto zero while charges from zero to. Oncedischarges down to zero, the current is commutated to [seeFig. 7(c)]. Providing a gating signal during conduction ofmakes ready to conduct the current. Actually starts to conductthe current in ZVS manner after the current in decaysto zero and alternates its polarity [see Fig. 7(d)]. This operationresults in no snubber loss.

3) Incomplete ZVS Operation: The IGBTs in bridge 1can not necessarily be turned on in ZVS manner even if theswitching current is negative. Unlike in the ZVS operation,does not discharge down to zero, and does notcharge up to, if the magnitude of, or is smallerthan, where

$$I_{\min} = \frac{2\sqrt{V_{D1}V_{D2}}}{Z_r} \qquad (6)$$

$$Z_r = \sqrt{\frac{L}{C_{\text{snub}}}} \qquad (7)$$

as stated in [12]. In this case, the operation of the leg makes adirect transition from Fig. 7(b) to (d), not through (c). Turningon with the charged snubber capacitor results in anamount of snubber loss. This paper refers to this as "incompleteZVS operation". The following is the calculation of the snubber loss caused bythe incomplete ZVS operation. The collector-emitter voltage of, in Fig. 7(b) can be expressed as

$$v_{\text{CE1}}(t) = \frac{(V_{D1} + V_{D2}) + (V_{D1} - V_{D2})\cos\omega_r t}{2} - \frac{Z_r|I_{11}|\sin\omega_r t}{2} \qquad (8)$$

where is the time after the dead time starts, andis the resonant angular frequency of and the collector-emitter voltage is not zero at the endof the dead time . As a result, the IGBT dissipates energy of when it is turned on.Therefore, the snubber loss in bridge 1 is calculated as

$$P_{\text{snub1}} = 4fC_{\text{snub}}\{v_{\text{CE1}}(T_d)\}^2. \qquad (9)$$

The snubber loss is proportionalto the capacitance of the snubber capacitors. Minimizingthe parasitic inductance leads to the use of small snubber capacitorswithout an excessive overvoltage appearing across anIGBT, thus resulting in reducing the snubber loss.

IV. CURRENT AND RELATED LOSSES

A. Conducting Loss in the IGBTs:

This paper approximates both the on-state voltage across theIGBT and the forward voltage drop across the freewheelingdiode to be 1.5V that is independent of the current flowing in them [13]. The conductingloss in the IGBTs and diodes can be calculated from theaverage of the absolute value of the current or when either bridge 1 or bridge 2 is operated in hard-switchingmanner, calculation on Fig. 5 yields:

$$\langle |i_1| \rangle = \frac{1}{\omega L}\left\{\frac{V_{D1}V_{D2}}{|V_{D1}-V_{D2}|}\frac{\delta^2}{\pi} + |V_{D1}-V_{D2}|\frac{\pi}{4}\right\}. \quad (10)$$

On the other hand, when both bridge 1 and bridge 2 are operatedin either ZVS or incomplete ZVS manner can be calculatedfrom Fig. 4:

$$\langle i_1 \rangle = \frac{V_{D1}V_{D2}}{\omega L(V_{D1}+V_{D2})}\left\{-\frac{\delta^2}{\pi} + 2\delta + \frac{(V_{D1}-V_{D2})^2}{V_{D1}V_{D2}}\frac{\pi}{4}\right\}. \quad (11)$$

Both and have to be obtained first to calculateand then either (10) or (11) should be applied, depending on theswitching manner.

B. Copper Loss in the Transformer and the Auxiliary Inductors:

$$I_1 = \frac{\sqrt{V_{D1}V_{D2}}}{\omega L}\sqrt{-\frac{4}{3\pi}\delta^3 + \delta^2 + \frac{\pi^2}{12}\frac{(V_{D1}-V_{D2})^2}{V_{D1}V_{D2}}} \quad (12)$$

Regardless of the switching manner, the copper loss in the transformerand the inductors is obtained as

$$P_{copp} = (R_{trans} + R_a)\cdot I_1^2 \quad (13)$$

Where, is the winding resistance of the transformerand is that of the auxiliary inductors

C. Core Loss in the Auxiliary Inductors.

The four auxiliary inductors were constructed using ferrite(TDK PC44) cores. The effective cross-sectional area of eachcore was the effective volume was and the turn number was. An air gap ofwas introduced in the magnetic path. Thus, the Instantaneous magnetic flux density is approximately expressedas

$$b_{ind} \simeq \frac{\mu_0}{g} N i_1 \quad (14)$$

where the permeability of vacuum is. The datasheet ofPC44 indicates that its core loss per volume iswhen the maximum flux density is 200 MT at afrequency of 100 kHz at a temperature of 25.In a power electronic context, the Steinmetz equation

Helps well to calculate an amount ofcore loss in a magnetic material where is a material constant,is a frequency of magnetization, and is the maximum flux Density. The exponents and are not constant values, butdepend on and the waveform of the magnetic flux. Whena ferrite core is excited at 20 kHz, the eddy-current loss can bemuch smaller than the hysteresis loss. Thus, is assumed tobe unity. The other exponent usually ranges from 2.6 to 2.8however this paper takes an approximation of so as tokeep the analysis simple. This approximation allows treating thecore loss in the same way as the copper loss.If the core loss per volume in PC44 can be approximatedby, the coefficient is given as 0.15. Thispaper assumes that a sinusoidal 20-kHz current having an rmsvalue as large as is responsible for the core loss in the auxiliary Inductors. The core loss in the four auxiliary inductors canbe calculated as

$$P_{core}[\mathrm{W}] = 4mf\left(\frac{\mu_0}{g}N\sqrt{2}I_1\right)^2 V_e = \frac{8mf\mu_0^2N^2V_e}{g^2}I_1^2 \quad (15)$$

where is the coefficient to transform an rms value into anamplitude. Therefore, the core loss in the four auxiliary inductorscan be treated as an equivalent winding resistance of

$$R_{core(ind)} = \frac{8mf\mu_0^2N^2V_e}{g^2} = 23\ \mathrm{m\Omega}. \quad (16)$$

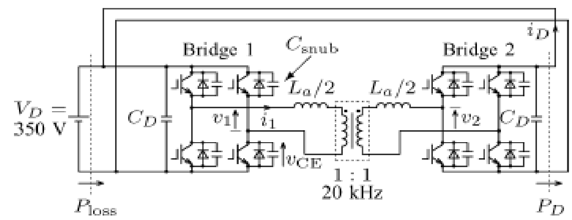

Fig. 8.Experimental circuit to measure the power loss of the dc-dc converter.

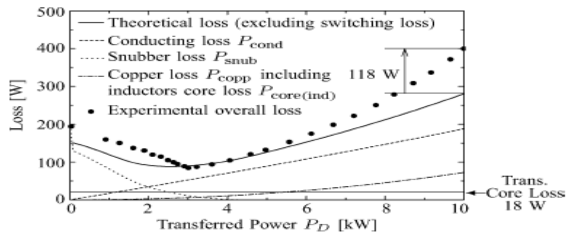

Fig.9 Comparison between calculated and experimental power losses atV = V = 350 V.

This equivalent resistance enables to calculate the core lossin the auxiliary inductors as a part of copper loss. Further detailedanalysis, for example, as found in [16], should bring animprovement of accuracy to the core-loss calculation althoughit is out of the scope of this paper.

V. POWER LOSSES AND LOWER LIMIT OF

A Comparison between Theoretical and Experimental Losses:

Theoretical losses described in the previous section are comparedto measured results on the basis of an experimental dc-dcconverter rated at 10 kW and 20 kHz. Fig. 8 shows the experimentalcircuit to measure the overall loss of the dc-dc converter.The circuit parameters in Fig. 8 are the same as those in Table I.Both theoretical calculation and experimental measurement arecarried out under.Note that in Fig. 8 is a dc voltage source. A connectionbetween the two dc links allows the power to be regeneratedback to the

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dc voltage source. Thus, the power coming fromequals , that is the overall loss in the dc-dc converter.Fig. 9 shows comparisons between the theoretical and experimentallosses. The solid line corresponds to the theoreticaloverall loss, although it excludes the switching loss inthe IGBTs or Even in the ZVS operation, the switchingloss is not zero due to the so-called “tail current” in the IGBTs. Itrequires modeling of the IGBT switching behavior in this dc-dcconverter to theoretically predict the switching loss. Howeverit is beyond the scope of this paper.When the theoretical losses were obtained asfollows. The conducting loss was. The snubberloss was. The copper loss both in the transformerand the inductors including the core loss inthe inductors.The core lossin the transformer wasAlmost independent of the power transfer.Thus, the theoretical overall loss is 282 W.The experimental results, on the other hand, was400 W. Thus, the difference between the theoretical and measuredresults were 118 W. It would include the switching lossin the IGBTs that was excluded from the theoretical overallloss. In [13], the switching loss in the IGBTs was 90 W whena power of 10 kW was transferred. Although the difference ofcannot be identified, the theoreticalcalculations can be valid because the error of 28Wcorrespondsto 0.28% of the power transfer of 10 kW, and 7% of the measuredoverall loss of 400 W.A 250-V, 5-kW bidirectional isolated dc-dc converter wasconstructed and tested in

$$I_{1pp} = \frac{\pi V_{D1}}{\omega L}\frac{d}{2}. \qquad (17)$$

The dc-dc converter employedthe third-generation planar-gate IGBTs rated at 600 V and100 A. The overall loss of the dc-dc converter was measuredat six switching frequencies from 10 to 20 kHz with a step of2 kHz. This measurement enabled to extract the switching lossfrom the overall loss because the switching loss is proportionalto the switching frequency. As a result, the switching losswas approximately 75 W at a power transfer of 5 kW and aswitching frequency of 20 kHz. The switching loss of 90 Win the 350-V, 10-kW dc-dc converter in this paper can be reasonable value, considering the raised voltage from 250 V to350 V, the increased power from 5 to 10 kW, and the se of thelatest trench-gate IGBTs.

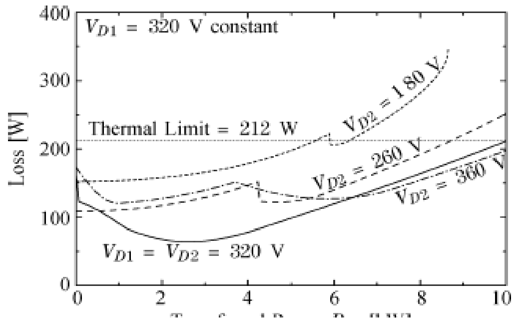

Fig.10 Sum of theoretical conducting and snubber losses (P+P)when P is positive.

B. Thermal Limit andV_{D2}

Fig.10shows calculated results of conducting and snubberlosses in the IGBTs when the power transferis positive. One dc voltage was kept constant at 320V,whilethe other dc voltage was changed as a parameter. Achievingthe ZVS operation becomes difficult with, comparedto, resulting in an increased snubberloss around.Fig. 10 defines atas a “thermal limit.” The losses in the IGBTs, whichare the most dominant in the overall loss, may make the IGBTmodules mounted on a heat sink suffer from the highest temperature.The temperature of the IGBT modules, more precisely thesemiconductor chips in the modules, determines the maximumpower transfer. Therefore, this paper considers only the lossesin the IGBT modules as the thermal limit.

VI.PEAK CURRENT IN THE AUXILIARY INDUCTORS

The ferrite cores in the auxiliary inductors would bemagneticallysaturated if the current exceeds 60 A because themagnetic flux density reaches 0.3 T as calculated by (14). Thedc-dc converter has to be operated considering the limitation onthe peak value of ore. The peak current imposes limitationson the dc voltage whenthe peak currentequals.Both thepower loss and the peak current impose limitations on the powertransfer and the dc voltage. Operation of the dc-dc converterhas to satisfy both limitations.Fig. 12 shows the observed waveforms when one dc voltageis 320V while the other is 360V at from bridges 1to 2. Fig. 13 shows another example of the observed waveformswhich was taken when while atfrom bridges 2 to1. The power transfer waslimited below 5 kW, as shown in Fig. 11 when the dc-dc converterhad a set of dc voltages of and the peak current was 60 A in Fig. 13, as calculatedin this section.

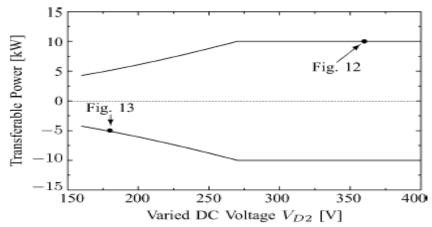

Fig.11 Transferable powers when the peak value of currentis limited to 60A and two operating points in experiments.

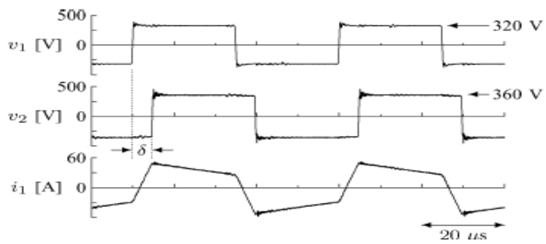

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Fig.12 Experimental waveforms when V = 320V, V = 360 V, _ =35 and P = 10 kW.

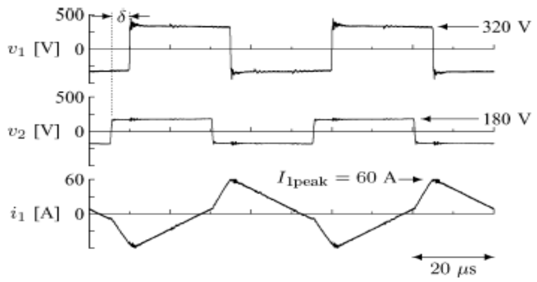

Fig. 13 Experimental waveforms when V = 320 V, V= 180 V, _ =□41 and P = □5 kW.

VII. APPLICATION TO AN ENERGY STORAGE SYSTEM

A.The 200-V, 10-kW, 2.6-kJ Laboratory Model

Fig.14 depicts the experimental energy storage system ratedat 200 V, 10 kW, and 2.6 kJ. Circuit parameters in the dc-dcconverter were the same as those in Table I. An electrolytic capacitorbank of 60,000 was used to simulate an EDLCbank. The capacitor bank is charged up to 350 V and dischargeddown to 190 V. Thus, energy of 2.6 kJ is stored into, and releasedout of, the capacitor bank. It corresponds to 70% of theenergy stored in at. The carrier frequency ofthe PWM converter used as the front end was 10 kHz.

B. Charging and Discharging of the Capacitor Bank

Fig.15 shows the observed waveforms when the energystorage capacitor bank was repetitively charged up to350 V, and then discharged down to 190 V. The waveform ofwas observed via a low-pass filter with a cut-off frequencyof 800 Hz. The maximal power transfer was 9.3 kW. In thisexperiment, the phase shift had a square waveform with amplitude of to make the controller simple. In actual energystorage systems, however, the power transfer should begiven by power demand, or a higher level controller regulatingthe voltage on the utility grid.

VIII. STARTING PROCEDURE

At the starting of the system, an inrush current wouldFlow into the auxiliary inductors and the transformer ifbridge 1 produced a square voltage of artand the inrush current would result in magneticsaturation in the cores of leading to an even larger inrushcurrent. Hence, this paper presents a special operation modecalled “recharging operation” to charge the capacitor bankfrom zero to the rated voltage of 320 V, preventing such aninrush current from flowing. The recharging operation allowsthe energy storage system to require no external starting-up orrecharging circuit. The recharging operation presented in thispaper can be considered as a well-known “soft-start” procedureFig.14 A 200-V, 10-kW experimental circuit with energy-storage electrolytic capacitors of 60000-F tosimulate an EDLC bank.However, experimental verification ofthe recharging operation makes a technical contribution to theenergy storage system with the large dc capacitor bank. In this experiment, hadbeen already charged up to 320 V before the dc-dc converter started the recharging operation. Bridge 1 produced a voltagewith a duty ratio of rather than 100% while bridge2 was operated as a diode rectifier with all the IGBTs kept off.The first pulse of had a half duty ratio of to suppressdc magnetization of the transformer. The starting procedure took approximately8 s to charge the capacitor bank from zero to 320 V. No excessiveinrush current flowed into the capacitor bank. When thevoltage across the capacitor bank, reached 275 V, the dc-dcconverter changed its operation mode from the recharging operationto the normal operation with the square voltages ofand having no phase shift (and) thepower transfer in the normal operation at has anegative-feedback effect to balance the two dc voltagesand due to the existence of the dead time. Thereforewas charged up to the same voltage as naturally afterreached 275 V.

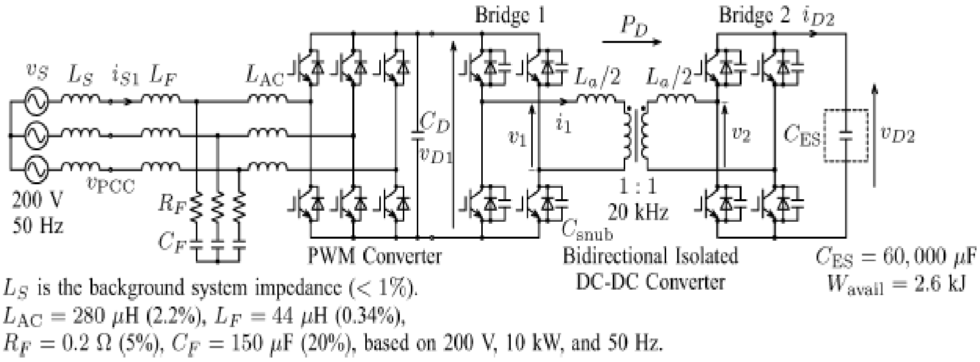

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Fig. 14.A 200-V, 10-kW experimental circuit with energy-storage electrolytic capacitors of 60000-μF to simulate an EDLC bank.

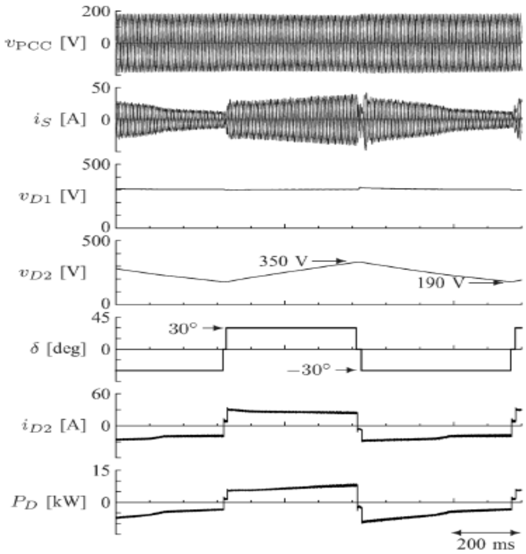

Fig.15 Experimental waveforms of charging and discharging of the capacitor bank (W = 2:6 KJ)

IX. CONCLUSION

This paper has addressed a bidirectional isolated dc-dcconverter suitable for an energy storage system. Theoreticalcalculations of power losses and peak current have clarified thedc-voltage limitations in the energy storage system. Experimentalresults have revealed that the dc-dc converter can chargeand discharge the capacitor bank properly. Moreover, the dc-dcconverter can charge the capacitor bank from zero to the ratedvoltage without any external recharging circuit.

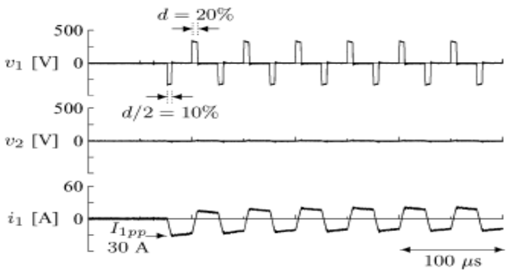

Fig.16 Transient waveforms when the dc-dc converter starts recharging the capacitor bank.

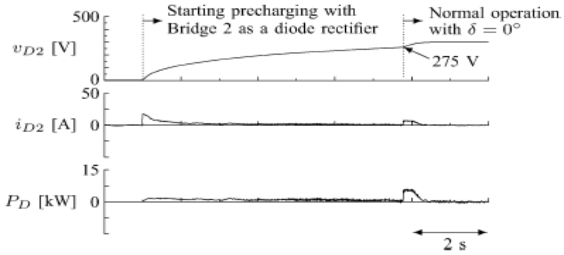

Fig.17 Experimental waveforms of dc voltage, current, and power in the recharging period.

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