A Comparison of High-Power DC–DC Soft-Switched Converter Topologies

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Abstract—**The purpose of this paper is to compare the properties of several soft-switching converter topologies when used to achieve dc–dc conversion at high-power and high-voltage levels. As an example, a 100-kW transformer isolated converter with 700–1400 Vdc input is designed with an estimated mass in the 90-kg range and an energy efficiency of 95%.**

I. Introduction

MANY industrial and military applications are arising that require high-power dc–dc conversion. These applications include shipboard, spaceborne, and transportation power systems. By employing new high-voltage high-power IGBT's, along with modern soft-switching techniques, the switching frequency can be significantly higher than that obtainable using gated turnoff (GTO) devices which in turn can lead to smaller, lighter mass, and potentially more cost-effective equipment. This paper summarizes a circuit topology trade-off study and the conceptual design of a 100 kW 700 V to 155 V dc–dc converter based on modern high-frequency, soft-switching techniques. First, new high-voltage insulated-gate bipolar transistor (IGBT) power device characteristics are considered and their switching losses estimated as a function of stress-reducing turn-off capacitors to determine the practical upper limit on switching frequency. Based on these results, five candidate topologies are evaluated. Finally, a conceptual design is done to get detailed mass and volume projections. As part of this conceptual design several magnetic core materials were evaluated for use in both transformers and resonant inductors.

II. Component Selection

A. High-Voltage IGBT's

To operate at high frequency and high voltage the IGBT is the best candidate power switching device. For the power levels considered here (100-kW range), power MOSFET's are not a practical choice due to their voltage and on-resistance limitations. IGBT's have considerably higher speed switching capability than GTO's as well as a relatively high impedance MOS gate. However, to operate reliably from a high-voltage dc bus with its inevitable transients, it is felt that IGBT's with breakdown voltage ratings equal to or exceeding 1400 V would be desirable for some applications while voltages exceeding 2000 Vdc would be needed for others. These types of devices are now being aggressively pursued by manufacturers being driven mainly by applications for high-power ac drives. Based on a survey of these new devices as well as preliminary data supplied by the manufacturers, designs were undertaken based on these devices.

It is desirable to operate at as high a switching frequency as possible to minimize mass of transformers, capacitors, and inductors. As the voltage capability of devices increases, their switching speeds generally decrease so that their switching losses can be expected to rise. Most manufacturers provide losses assuming no snubber (which is useful for conventional low-frequency hard-switched circuits). The higher-frequency soft-switched circuits under consideration in this study generally have a stress-reducing snubber capacitance across the device to reduce switching losses. Therefore, the switching losses of these new high-voltage IGBT's were estimated as a function of snubber capacitance using a simple turnoff model which includes the turnoff "tailing" current [1]. Some sample results, (based on a 1400-V, 400-A IGBT) are shown in Fig. 1 as a function of snubber capacitance for a switched current of 300 A (the current needed for most topologies for a 100-kW converter). Results for other devices were similar. Based on these results it was judged that a switching frequency around 20 kHz was achievable with a snubber capacitor per device of 0.6 μF. This gives approximately 166 W of switching loss per device—664 W switching losses for four devices in a bridge (i.e., 0.66% of the power delivered). The topologies were all evaluated at the 20-kHz switching frequency.

B. High-Frequency Magnetic Core Materials

Based on a switching frequency of 20 kHz, several available core materials were investigated for use in the transformers and inductors [2]. Table I shows some core material comparisons. Based on the results of this table it was concluded that ferrite core material is the most suitable choice. It is the least expensive (it is assumed that the needed core can be built up by stacking existing C- or E-cores) and in general will result in the lightest mass due to its low loss characteristics at 20 kHz. Several sample designs were performed using the other materials to arrive at this conclusion. In addition,

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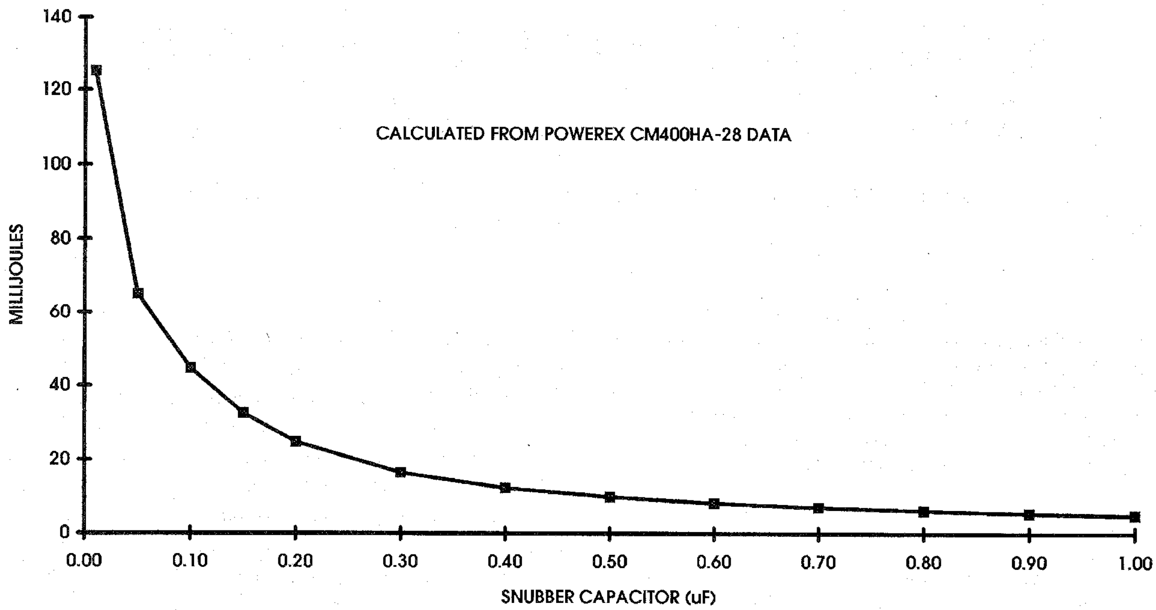


Fig. 1. IGBT turnoff energy as a function of snubber capacitance.

TABLE I
COMPARISON OF CORE MATERIALS AT 20 kHz, 50 kVA.

Core Material	Bm* (Gauss)	Relative Size**	~ Cost ($/kg)	Relative Core Costs
Amorphous Metal (2605S-3)	2,379	1.2	33	67
(2605SC)	1,000	2.4	33	128
(2714A)	4,200	0.8	176	232
Ferrite (3B7)	1,600	1.3	2.2	2.9
(H7C4)	2,160	1.0	2.2	2.3
(3C80)	1,700	1.2	2.2	2.8
Orthonol (0.051mm)	280	6.2	24	241
(0.025mm)	650	3.3	33	176
(0.013mm)	950	2.5	42	169
48 Alloy (0.051mm)	600	3.5	24	136
Supermalloy (0.051mm)	2,000	1.4	33	76
(0.025mm)	2,500	1.2	44	86
(0.013mm)	2,900	1.1	55	96
Square Permalloy 80 (0.051mm)	2,400	1.2	33	66
(0.025mm)	2,000	1.4	44	101
(0.013mm)	2,400	1.2	55	110
Magnesil	low	large		
Supermendur	low	large		

* Flux density that results in 5 W/lb. or 80 mW/cubic cm.
** Size proportional to (1/kBm)^^.75 where k is core stacking factor
k = .75 for tape wound cores
k = 1 for Ferrites

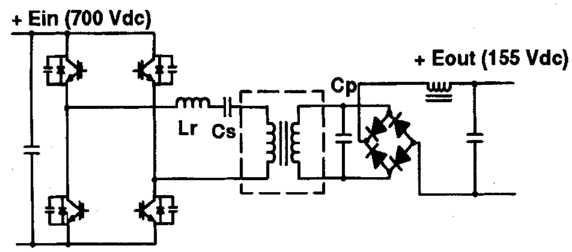
modern powdered iron cores were considered for the resonant inductors; but, here again, it was found that ferrite offered the lowest mass and loss designs. Therefore, in the next section ferrite material was assumed in the sample designs performed for the candidate topologies.

III. CANDIDATE TOPOLOGIES

Five candidate topologies were identified for the 100-kW converter as shown in Fig. 2. A hard-switched pulsewidth modulated (PWM) bridge converter serves as the baseline [shown in Fig. 2(e)]. Because the hard-switched converter would have no (or very little) snubber capacitance, it is not possible to operate the circuit at 20 kHz. As seen in Fig. 1, the switching losses with a snubber capacitance of zero are very high (130 mJ) being more than ten times as high as that resulting with a snubber capacitor of 0.6 μF. Therefore, for the hard-switched case, a switching frequency of 5 kHz was assumed (a factor of 4 less than the 20 kHz assumed for all the other soft-switched schemes). Note that conventional lossy snubbers are impractical at high frequency due to the excessive losses that would result in discharging the snubber each cycle. The other four schemes in Fig. 2 (a)–(d) are soft switched and circulate the snubber energy back to the main (or load) bus in a lossless manner prior to turn on of the main switches (hence soft-switching).

The phase-shifted bridge topology, as shown in Fig. 2(c), is as simple as the conventional PWM circuit but, due to the circuit operation and a properly designed transformer, soft switching is maintained [3], [4]. Voltage is controlled by phase shifting one converter leg relative to the other converter leg to form a PWM-type output signal. By proper design of the transformer leakage and magnetizing inductance, the proper amount of energy is stored each cycle such that when a power switch turns off, this inductive energy is interchanged with the snubber capacitors across each device to "soft switch" the converter pole. In essence, the snubber capacitors resonate with the transformer leakage inductance and magnetizing inductance when each device turns off (which results in soft switching).

The dual active bridge (DAB) converter was previously developed to demonstrate a high-power (50-kW), high-voltage (2000-Vdc), high-power density (0.2 kg/kW) dc–dc converter [5]. The circuit operation will be described around the chosen configuration shown in Fig. 2(b). It consists of two voltage-



SERIES/PARALLEL RESONANT

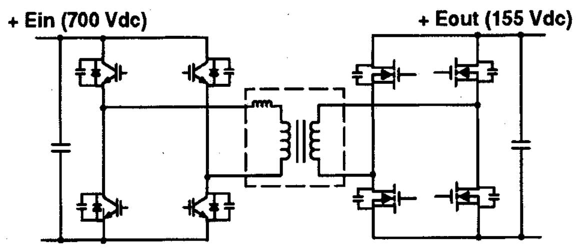
ADVANTAGES

- Sinusoidal currents
- Soft switching, primary & secondary
- Small output Lo
- Low ripple current in Co
- Xmer leakage is no problem

DISADVANTAGES

- Large resonant inductor
- High–voltage, high–current resonant Capacitors
- Variable Frequency Control

(a)



DUAL ACTIVE BRIDGE

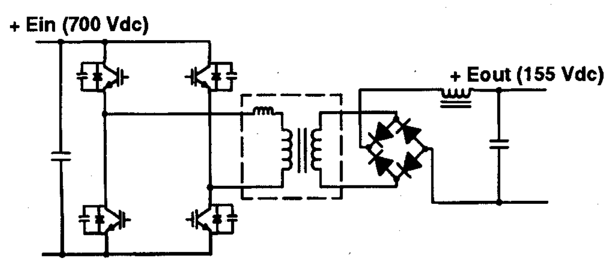
ADVANTAGES

- Constant frequency
- No output filter inductor
- Soft switching primary & secondary
- Resonant inductance built into transformer

DISADVANTAGES

- Soft switching may be lost at light load
- Eight active power switches
- High ripple current in Cout
- Relatively large Xmer

(b)



PHASE–SHIFTED BRIDGE

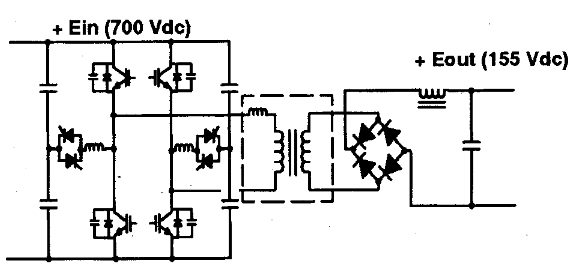
ADVANTAGES

- Simple Control
- Constant frequency
- Resonant inductor built into Xmer

DISADVANTAGES

- Output diodes hard switched
- Circulating current during "off–time"

(c)



AUX. RESONANT COMMUTATED BRIDGE

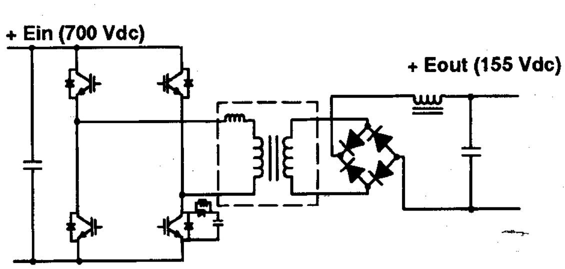
ADVANTAGES

- Smallest Xmer
- Lowest IGBT stresses

DISADVANTAGES

- Auxiliary active devices needed
- Output diodes hard switched
- Two small resonant inductors
- Auxiliary device type unclear

(d)



HARD–SWITCHED PWM (BASELINE)

ADVANTAGES

- Smallest transformer
- Simple control

DISADVANTAGES

- High IGBT and diode switching stresses
- High–speed feedback diodes needed
- Frequency limited

(e)

Fig. 2. (a)-(e). Candidate topologies for 100 kW dc-to-dc converter.

TABLE II
TOPOLOGY COMPARISONS BASED ON SAMPLE 100-kW DESIGNS

Characteristic	Hard Switched PWM	Phase Shifted Bridge	ARCP	Series/Parallel/ Resonant	DAB
Switching Frequency (kHz)	5	20	20	20	20
Control Complexity	simple	simple	complex	moderate	moderate*
Constant Frequency	yes	yes	yes	no	yes
Circulating Current (A)	no	yes	no	yes	yes
Peak IGBT Current (A) (Vin=665 Vdc)	211	335	211	317	240 (input) 978 (output)
IGBT Stresses	high	moderate	low	moderate	moderate
No. of Active Devices	4	4	8	4	8
Resonant Inductor	no	in Transformer	yes	biggest	in Transformer
Resonant Capacitor	no	no	no	largest	no
Output Rectifier Stresses	high	moderate	moderate	lowest	IGBT low
Ripple Current in Co	low	low	low	low	high
Transformer Mass (kg)	23	15	13	13	20
Output Inductor Mass (kg)	20	11	11	7	0
Resonant Inductor Mass (kg)	0	0	0	13	0
Total Magnetics Mass (kg)	43	26	24	33	20
Rel. Magnetics Mass	1.00	0.60	0.56	0.77	0.47

*Restricted load range 40 - 100 %

sourced active bridges, one operating in the inversion mode and the other in the rectification mode, interfaced through a high-frequency isolation transformer. Each bridge is controlled to generate a high-frequency resonant-transition (edge resonant) square wave voltage at its transformer terminals ($\pm V_{\text{in}}, \pm V_{\text{out}}$). By incorporating a controlled amount of leakage inductance into the transformer, the two square waves can be appropriately phase-shifted to control the amount of power flow from one dc source to the other. Active bridges on either side of the transformer allows bidirectional power transfer. Power is delivered from the bridge generating the leading square wave. Maximum power transfer is achieved at a phase-shift of 90°. High efficiency is obtained since all the devices operate under zero voltage source (ZVS) conditions. The circuit can achieve both step-up or step-down voltage conversion depending upon the control phase-shift and the load. As seen, the control is simple, requiring a fixed-frequency phase-shifting strategy under a restricted load range. However, at lighter loads additional inductive energy is needed to complete the resonant transition from rail to rail. This can be achieved by one of several techniques: 1) By increasing the magnetizing current (reducing the magnetizing inductance of the transformer). This also leads to increased conduction losses; 2) by switching in a resonant inductor at lighter loads as done in the ARCP converter (discussed below); 3) by incorporating freewheeling states at the input converter, increasing control complexity; 4) by decreasing the switching frequency proportionately as the load decreases, thereby increasing the circulating current.

The auxiliary resonant commutated (ARC) bridge converter shown in Fig. 2(d) comprises two auxiliary resonant commutated poles (ARCP) [6], [7]. The ARCP combines the features of a conventional hard-switching PWM converter with the resonant transition switching of the phase-shifted bridge topology. At light load conditions, a pair of zero current switching auxiliary devices is used to commutate the resonant phase legs from rail to rail. At full load, the auxiliary devices are not triggered (avoiding high conduction losses) and the converter resonant capacitors commutate against the leakage inductance of the transformer (just as in the phase-shifted bridge). The auxiliary devices are switching against half the input bus voltage keeping the turn-on losses low. However, for 20-kHz applications a fast zero current turnoff thyristor-type device (having little reverse recovery time) needs to be selected. Gate-assisted turnoff thyristors (GATO's), GTO's, and MOS-controlled thyristor (MCT) devices are the only viable solutions to date to implement the ARCP auxiliary switch.

Unlike the above topologies, the series–parallel resonant converter shown in Fig. 2(a) [8], [9] is a load resonant converter with relatively heavy magnetics mass which is dominated by a large resonant inductor that must process all the power as well as circulate additional kVA (the kVA rating of the resonant inductor is typically 3 times the power delivered).

Sample designs of all these converters were done for a 100-kW power level and an input voltage of 700 Vdc (±5% steady-state and ±16% transient) and an output voltage of 155 Vdc (±5%). Table II shows the topology comparisons based on the sample designs for these five converters, and Table III shows the loss calculations for the converter components at full load for the soft switched approaches.

The device switching losses were calculated as a function of snubber value using the data from curves such as that in Fig. 1 (there is a separate curve for each turnoff current). The "tail current" characteristic as well as the forward voltage drop data was taken from manufacturers data for the device of Fig. 1. The forward drop could be modeled with sufficient accuracy by assuming an ideal diode drop of 1.2 V in series with a 5 mΩ resistor. The magnetics components were designed. The numbers in the table are for a C core of amorphous metal designed for a 50°C hot-spot rise and an approximate 50% copper loss/50% core loss split. The resonant inductors were designed using 3C80 ferrite C cores. The snubber capacitors are different for the different circuits since they were selected for the actual peak currents turned off. In addition, the tradeoff of switching losses vs. conduction losses for the different topologies was selected as that which were judged to be the best tradeoff for a practical design for the given circuit.

As seen in Table II, the hard-switched converter has the heaviest magnetics because of the lower switching frequency necessary due to the higher IGBT and diode-switching energy loss. Also the stresses are highest due to the hard switching. All of the circuits except the DAB have an output filter inductance that greatly reduces the required ripple current carrying requirements of the output capacitance. Thus, the DAB (which does not have an output filter inductor) has the highest ripple current requirement for the output filter capacitance. The series–parallel converter has the disadvantage of a relatively large resonant inductor and resonant capacitor. However, the converter results in the lowest output diode stresses due to the fact the parallel resonant capacitor is placed effectively in parallel with the rectifier devices and therefore acts as a lossless snubber. The control of the ARCP is conceptually simple but is rated as complex due to the extra gate drivers and sensing circuits required. The ARCP and the series–parallel converter have the smallest transformers due to the fact that they carry no circulating current as the other soft-switched circuits do.

The phase-shifted bridge transformer is somewhat larger due to some additional "circulating" current. However, the overall magnetics mass is comparable to the other low-mass schemes. In addition, because the circuit uses the transformer leakage inductance as a circuit element, the transformer primary and secondary windings are not extremely tightly coupled. This allows the primary and secondary windings to be separated for good voltage isolation (and low capacitance for reduced common-mode electromagnetic interference) between primary and secondary.

Table III illustrates that a 95% efficiency should be attainable with several approaches. Table III presents the DAB loss estimates under full load conditions at minimum input voltage, 665 Vdc (worst case). Moreover, for these calculations, the converter was designed to operate under ZVS conditions for a restricted load range of 40–100% with practically no magnetizing current. With the ARCP it is possible to operate the converter similar to the phase-shifted bridge converter without the need to select a transformer that has sufficient low magnetizing inductance. This eliminates the circulating currents of the phase-shifted bridge that otherwise add to conduction losses. The reduced condution loss characteristic of the ARCP converter is illustrated in Table III.

TABLE III
CONVERTER LOSS CALCULATIONS AT 100 kW

	LOSS (WATTS)				
	HARD SWITCHED PWM (BASELINE) (@ 5kHz)	PHASE-SHIFTED BRIDGE (@ 20kHz)	SER./PAR. RESONANT (@ 20kHz)	ARCP (@ 20kHz)	DAB[f] (@ 20kHz)
IGBTs					
conduction	900	1072	1000	900	786
switching	844[a]	824[b]	587[c]	472[d]	739
TRANSFORMER					
copper	150	150	150	150	150
core	148	148	148	148	148
RESONANT INDUCTOR			1875 (Q=200)	20	
OUTPUT RECTIFIERs / IGBTs	1548	1748	1748	1748	2194
OUTPUT INDUCTOR					
copper	83	83	83	83	
core	54	54	54	54	
AUX. DEVICES				264[e]	
CAPACITORS	100	100	150	150	250
GATE DRIVE + LOGIC	10	10	10	10	20
MISC.	250	250	250	250	250
TOTAL	4087	4239	6055	4149	4537
EFFICIENCY	96.1%	95.9%	94.3%	96%	95.7%

[a]no snubber
[b]Csn = 0.3 uF
[c]Csn = 0.15 uF, switching losses increase with lighter loads (higher frequency)
[d]Csn = 0.2 uF
[e]losses incurred only at lightly loaded conditions
[f]Restricted load range 40 - 100 %, Cp = 0.3uF, Cs = 2uF

IV. EXAMPLE DESIGN OF PHASE-SHIFTED BRIDGE

As an example, Fig. 3 shows a practical design of the phase-shifted bridge. The phase-shifted bridge was judged to be the most suitable for a typical dc–dc converter application due to its relative simplicity, low magnetics mass, constant frequency operation, lack of auxiliary circuits, and lack of a relatively large resonant inductor and resonant capacitor. The other circuits are more suitable for specialized applications. For example, the DAB for bidirectional power flow, the series–parallel for high-voltage output where large transformer leakage inductance will result (which can be part of the resonant inductor), and the ARCP where extremely low magnetics mass and losses are required. The 100-kW phase-shifted converter was designed using two 50-kW bridges operating in parallel but with their switching periods operating 90° out of phase. This arrangement has the following advantages.

- By operating the two phase-shifted bridges 90° out of phase, the ripple current at both the input and the output is greatly reduced (due to harmonic cancellations and reductions) and the lowest ripple frequency is doubled. These effects result in smaller input and output filters with less stress.
- The power for each component converter is reduced to 50 kW. This brings the power ratings of components down to more manageable levels consistent with available hardware. At the 50-kW level, ferrite cores near the size needed for the magnetics are more readily available and more manageable wire sizes can be used. In addition, the lower power is more amenable to the desired high switching frequency.
- In the event of the failure of one 50-kW module, there is the possibility of continued operation (at 50 kW) from the other module.

Such an arrangement using series–parallel resonant converters for a lower power application has already been demonstrated [10]. Based on the arrangement of Fig. 3 the converter has been simulated using PSPICE to aid in design and to determine component stress. Fig. 4 shows output from a sample PSPICE simulation. This case shows a single 50-kW module operating at full power and illustrates the effect of

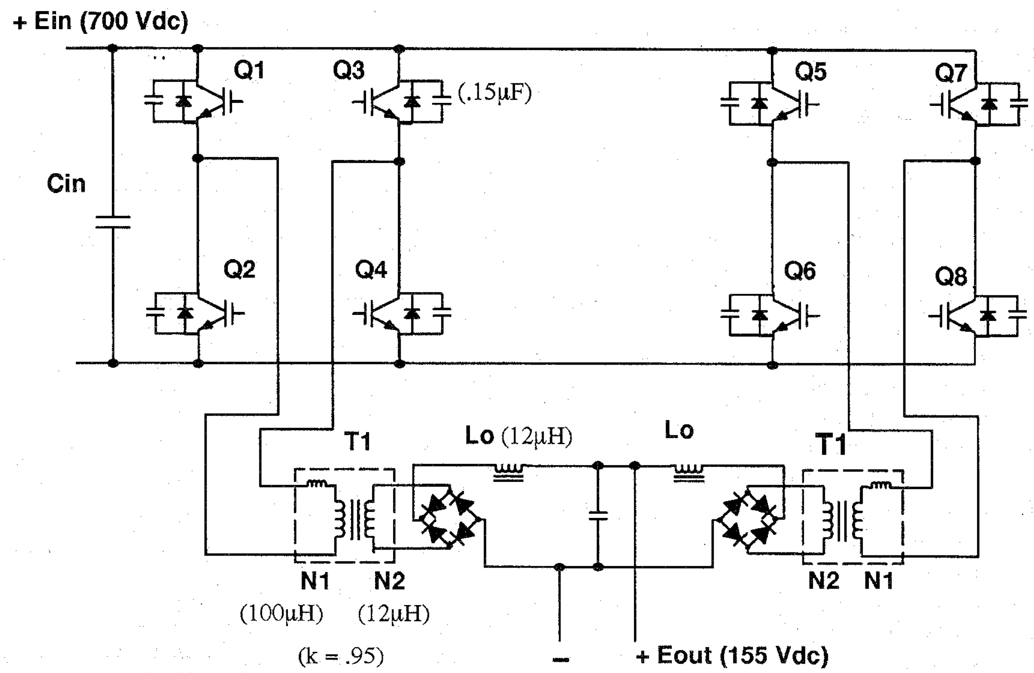


Fig. 3. Practical implementation of resonant phase-shifted bridge.

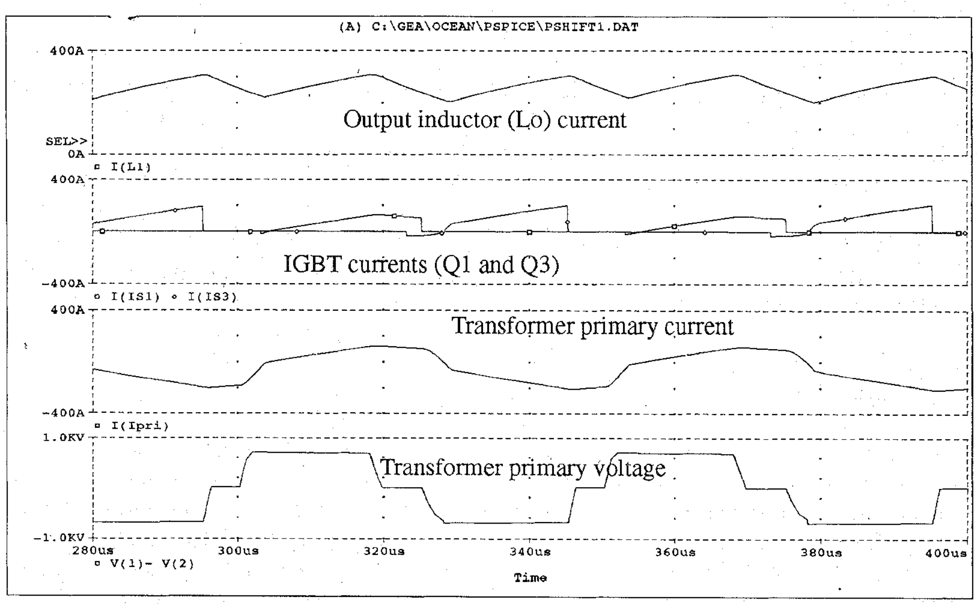


Fig. 4. Simulation results at full load.

both leakage and magnetizing inductance. The transformer primary current is comprised of two components—a sloped component is due to transformer magnetizing current (which is needed to supply soft switching at light load), and a rectangular component is the reflected load current (which stores energy in the transformer leakage inductance to maintain soft switching at heavy loads). The sloped edges of the transformer voltage illustrate the relatively slow rate of voltage rise across a power IGBT as it turns off (hence soft switching). Using these simulation results and other analysis, a detailed design of the phase-shifted bridge was performed to obtain size and mass estimates.

Based on the circuit and component designs discussed above, a conceptual package, using conventional packaging techniques, was designed and mass estimates for all components estimated for the phase-shifted resonant bridge. Based on forced air cooling within the chassis and conduction cooling through one side of the chassis and assuming a 6.35-mm-thick aluminum chassis the following was estimated:

- output power $= 100$ kW;
- input voltage $= 700$ Vdc ($\pm 16\%$ trans., $\pm 5\%$ steady state);
- output voltage $= 155$ Vdc ($\pm 5\%$);
- converter dimensions $= 0.25 \times 0.38 \times 0.76$ m $= 0.072$ m^3;
- converter mass $= 89$ kg;
- efficiency $= 95\%$.

It is felt that a more compact and lower mass design could be realized with internal liquid cooling.

V. Summary and Conclusion

Converter conceptual designs and comparisons at the 100-kW level to convert high voltage directly to a lower dc utilization voltage have been performed. The recent development of high-voltage, high-power IGBT's has made high-frequency dc–dc conversion at high-power levels feasible. A circuit trade-off of five approaches led to the selection of the resonant phase-shifted bridge soft-switched topology operating at 20-kHz switching frequency for an example application that required simple unidirectional power flow. A specific design to convert 700 Vdc to 155 Vdc (with transformer isolation) was completed and simulated. An efficiency of 95%, a mass of 89 kg, and a volume of 0.072 m^3 were determined using conventional packaging techniques.

If bidirectional power flow is needed (e.g., a battery charger/discharger application) then the DAB converter offers an excellent alternative. If very high efficiency must be maintained from no load to full load, then the additional complexity of the ARCP phase-leg may be justified, since this approach has the highest efficiency. It appears that load resonant converters (e.g., series/parallel resonant converter) are not attractive at such high power levels due to the difficulty of implementing a compact, high efficiency resonant inductor that must process the entire load power.

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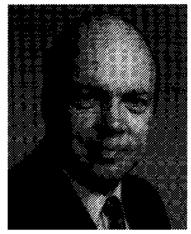

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Since joining the research staff of the GE Corporate Research and Development Center in 1968, he has conducted research and advanced development of power converters for dc and ac motor drives, induction heating, high-frequency lamp ballasts, uninterruptable power supplies, utility interactive photovoltaics, and both low-voltage and high-voltage power supplies. His present interests include the analysis and development of high-frequency, high-density, soft-switching power converters and their components. Prior to joining the Corporate Research and Development, he was employed at the GE Semiconductor Products Department where he developed high-current diodes and SCR's. He has published 37 papers and has 74 patents.

Dr. Steigerwald received the 1993 William E. Newell Award from the IEEE Power Electronics Society for outstanding achievement in power electronics. He has been awarded three prize paper awards. He has served on the program committees for numerous IEEE APEC and PESC Conferences and was an at-large member of the IEEE Power Electronics Society Administrative Committee (1990–92). He is currently an Associate Editor of the IEEE Transactions on Power Electronics and a member of Eta Kappa Nu.

Rik W. De Doncker (M'87) received the Dr. Eng. degree from the Katholieke Universiteit Leuven, Leuven, Belgium, in 1986.

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