# A THREE-PHASE SOFT-SWITCHED HIGH POWER DENSITY DC/DC CONVERTER FOR HIGH POWER APPLICATIONS.

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## Abstract

This paper presents three dc/dc converter topologies suitable for high power density high-power applications. All three circuits operate in a soft switched manner making possible a reduction in device switching losses and an increase in switching frequency. The three-phase dual-bridge converter proposed is seen to have the most favorable characteristics. This converter consists of two three phase inverter stages operating in a high frequency six-step mode. In contrast to existing single phase ac link dc/dc converters, low r.m.s. current ratings are obtained for both the input and output filter capacitors. This is in addition to smaller filter element values due to the higher frequency content of the input and output waveforms. Furthermore, the use of a three-phase symmetrical transformer instead of single phase transformers and a better utilization of the available apparent power of the transformer (as a consequence of the controlled output inverter) significantly increase the power density attainable.

#### Introduction

The area of high power density dc /dc converters has been an important research topic, especially for switched mode power supply applications rated at up to 500 watts. The needs of the next generation of aerospace applications require extremely high power densities at power levels in the multi-kilowatt to megawatt range. The implications of realizing high power density and low weight systems at these power levels have rarely been addressed. This paper examines considerations for the selection of topologies which can realize the low weight constraints that are of primary importance.

Recognizing that higher switching frequencies are the key to reducing the size of the transformer and filter elements, it is apparent that some form of soft switching converter with zero switching loss is required, if system efficiencies and heat sink size are to be maintained at a reasonable level. By far the most attractive circuit so far has been the series resonant converter (SRC) [1]. Using thyristors with a single LC circuit for device commutation and energy transfer, the topology is extremely simple in realization and offers the possibility of power densities in the 0.9-1.0 Kg/KW range at power levels up to 100 KW.

The following problems can be identified with the SRC. Thyristor commutation requirements demand higher current ratings from the devices and higher VA ratings from the LC components. Thyristor recovery times significantly slow down the maximum switching frequencies attainable. Snubber inductors and RC networks are needed to effect current transfer without encountering a diode recovery problem. Capacitive input and output filters have to handle ripple currents at least as large as the load current. Although switching frequencies in the 10 KHz range yield dramatic reduction in converter size when compared to conventional hard switching circuits, it is clear that systems operating at similar frequencies and with lower component ratings are potentially capable of even higher power densities. This paper proposes a new soft-switching dc/dc converter topology suitable for high power applications. Soft switched converters are characterized by intrinsic modes of operation which allow an automatic and lossless resetting of the snubber elements through an appropriate recirculation of trapped energy. The capability to eliminate losses associated with the snubber now permit the use of oversized snubbers resulting in dramatically lower device switching losses, even at substantially higher frequencies. Examples of soft switched dc/dc converters are the parallel output SRC operated above resonance [9], the resonant pole, the pseudoresonant converter [2,3] and all quasi-resonant converters [4,5]. For multi-quadrant operation and for dc/ac inverter applications, typical examples of soft-switched topologies are the resonant dc link inverter and the quasi-resonant current mode or resonant pole inverter [5]. The proposed circuit utililizes the resonant pole as the basic switching element for both input and output devices and yields substantial benefits in power density and operating characteristics.

#### Soft Switched DC/DC Converters

The preferred dc/dc converter topology for high power applications has been the full bridge circuit operated at constant frequency under a pulse width control strategy. The topology features minimal voltage and current stresses in the devices, minimum VA rating of the high frequency transformer, as well as low ripple current levels in the output filter capacitor. The power density levels that can be reached are limited by peak and average device switching losses, transformer leakage inductances and output rectifier reverse recovery. The factors above constrain the maximum frequency attainable, and thus the smallest size possible, given the state of the art in component technology. Most of the soft switching converters reported in literature attempt to tackle one or more of the problems listed above, typically at the expense of substantially higher component stresses [4,5]. For high power operation, that is unacceptable. Soft switching variations of the full bridge converter are thus the most favoured topologies.



Figure1: Pseudo resonant full bridge dc/dc converter [2].

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The pseudo-resonant dc/dc converter proposed in reference 2 is shown in Figure 1. It uses capacitive snubbers and can be designed with device stresses approaching that of the conventional full bridge. However, the circuit uses the transformer as a voltage transfer element and the interactions of its leakage inductance ( $L_1$ ) and the output rectifer are unresolved. The maximum switching frequency limit is reached when the energy lost due to  $L_1$  and the peak diode reverse recovery current become unacceptable.

Transferring the output filter inductance to the ac side completely changes the operating characteristics of the circuit [2]. As shown in Figure 2a, the inductance  $L_r$  is now in series with the transformer leakage inductance, L1. Further, energy trapped in  $(L_l+L_r)$  during reverse recovery is now transferred to the load in a lossless manner. This implies that a further increase in switching frequency may be possible. It can also be shown that the additional inductor L required in the previous circuit (Figure 1) is no longer absolutely essential, as adequate control range can be obtained at constant frequency while maintaining soft-switching for all devices. Operating waveforms for this single-phase phase shifted dc/dc converter are shown in Figure 2b. It can be seen that soft switching is obtained by ensuring that device turn-on only occurs when its anti-parallel diode is conducting. The biggest problem with this circuit seems to be the high ripple current in the output filter capacitor. For low voltage power supplies, this is a very important consideration. For high power high voltage aerospace applications, it is not seen to be a major problem because power densities obtainable with multi-layer ceramic capacitors are at least as good as with inductors.

It has been proposed in Reference [7] that the diode recovery process is akin to the existence of an active device in anti-parallel with it. Observing that the circuit in Figure 2a naturally handles the diode reverse recovery process, it is proposed that the diodes be replaced by active devices as shown in Figure 3a. Many high frequency converters already use synchronous rectifiers in essentially the same location. The converter can now be operated



Figure 2: a) Topology A, single phase shifted dc/dc converter b) Waveforms for topology A



a) single phase dual bridge dc/dc converter b) waveforms for topology B

with a simpler control strategy in which the input and output bridges generate square waves which are phase shifted from each other. In keeping with our philosophy, regions of operation can be identified which permit soft-switching of <u>all</u> devices in <u>both</u> bridges. Figure 3b shows operating waveforms for the single phase dual bridge dc/dc converter topology. The circuit in Figure 3a can be extended to a polyphase

The circuit in Figure 3a can be extended to a polyphase version. The three-phase dual bridge circuit is shown in Figure 4a. Again, examining the modes of operation for the converter, it is possible to identify regions where <u>both</u> sets of switches experience soft-switching. As in the case of the single phase dual bridge converter, both bridges generate quasi-square waves phase shifted from each other. It should be noted that the soft-switching transition is actually resonant in nature [2,6] but is assumed to be almost instantaneous for the derivation of first-order operating characteristics. The three-phase dual bridge converter has substantially lower filter ratings when compared to its single phase counterpart. Consequently, it has the potential of realizing the highest power density.

It should be noted that all three converters, denoted converter A, B, and C for circuits in Figure 2, 3, and 4 respectively, exhibit desirable properties with regard to parasitics such as device storage time, transformer leakage inductance and diode reverse recovery. It is shown in the paper that transformers which use the leakage impedance as an energy transfer element have the potential of reaching higher power densities. While this technique has been extensively used at lower power levels, it has been felt that the higher VA rating of the composite transformer was an unacceptable penalty at higher power levels. It will be shown that the resulting increase in switching frequency possible, more than compensates for the increased VA rating, allowing a substantial reduction in the overall size of the converter. The use of dual bridges also yields unexpected gains in power flow. Analysis in paper is restricted to unidirectional power flow only.





Figure 4: Topology C, a) Circuit schematic of the proposed 3 phase dc/dc converter,

b) Operating waveforms:
 V<sub>AP</sub>- Primary line-neutral voltage,
 V<sub>AS</sub>- Secondary line- neutral voltage,
 I<sub>AP</sub>- Primary line current,
 I<sub>AS</sub>- Secondary line current,
 I<sub>S</sub>- Source side dc link current,
 I<sub>O</sub>- Load side dc link current.

# Analysis Of Single Phase Converters

## Converter A:

In order to derive the operating charateristics of the three dc/dc converters, it is assumed that the transfer of current from device to diode on turn-off is instantaneous. The actual switching locus depends on the value of capacitance, C, used and the current level. For a typical device such as a BJT with a current fall time,  $t_f$ , and a turn-off current of  $I_p$ , the device energy loss per switching cycle can be found approximately to be [8]

$$E_{sw} = \frac{I_{p}^{2} t_{f}^{2}}{24 C}$$
(1)

As there are no turn-on or snubber dump losses, C can be made fairly small while retaining a fast switching characteristic and low device losses, simultaneously. This justifies the assumption of a fast, almost instantaneous switching transition for analysis over a full cycle.



Figure 5: Equivalent circuits of three dc/dc converter topologies:
a) Single-phase phase-shifted converter, topology A,
b) Single-phase dual bridge converter, topology B,
c) Three phase dual bridge converter, topology C.

The equivalent circuits for the three converters are shown in Figure 5. Replacing the transformer with an equivalent inductance, L, simplifies circuit analysis. For converter A, three operating modes can be identified as shown in Figure 2b. The phase shift between the two resonant poles is  $\phi = \omega t_p$ , where  $\omega$  is the switching frequency. The equivalent voltage applied across the load is Vab and has pulse width of  $\beta = \omega t_b$ . The inductor current i as a function of  $\theta = \omega t$  is given below. In Mode 1,

$$i(\theta) = \left(\frac{V_s + V_0}{\omega L}\right)\theta + i(0)$$
(2)

where  $V_s$ ,  $V_o$  are input and output dc voltages and i(0) is the initial current at  $\theta = 0$ . Mode 1 ends at  $\theta = \phi$ . Given the output rectifier it is clear that  $0 \le \phi \le \beta$ . In Mode 2,

$$i(\theta) = \left(\frac{V_{s} - V_{0}}{\omega L}\right)(\theta - \phi) + i(\phi)$$
(3)

similarly, the current in Mode 3 can be found to be

$$i(\theta) = \left(\frac{-V_0}{\omega L}\right)(\theta - \beta) + i(\beta)$$
(4)

At the end of the half cycle  $i(\pi) = -i(0)$ . Solving for i(0), we can obtain the complete current waveform.

The soft-switching constraints require that the device be conducting at turn-off. From Figure 2b this implies that  $i(\pi) > 0$ , Further, given the diode bridge on the output,  $i(\phi)=0$ . Using the above relationships,

$$\phi = \frac{1}{2} \left(\beta - d\pi\right) \tag{5}$$

where

$$d = \frac{V_0}{V_s}$$
(6)

From  $i(\theta)$  and the converter switching functions, the supply and output average currents, (I<sub>s</sub> and I<sub>0</sub>) can be found to yield power transfer at a given  $\beta$ , d and  $\omega$ . This is found to be

$$P_0 = V_s I_s = \frac{dV_s^2}{4\omega L} \left[ 2\beta - \pi d^2 - \frac{\beta^2}{\pi} \right]$$
(7)

Input and output filter capacitor rms current ratings can also be calculated from  $i(\theta)$ . Peak device voltage and current ratings are also easily found.

also easily found. Estimating the size/power density of the transformer is not as straightforward as it might seem. As the leakage reactance is used for the main energy transfer element, the size associated with it is fairly significant. It is presumed that the VA rating of the 'composite' transformer is smaller than that obtained for the combined VA ratings of an ideal transformer and series inductances corresponding to the leakage inductance. Figure 6 shows the equivalent circuit model of a single phase transformer. The KVA rating of the transformer is then calculated to be (neglecting magnetizing current),

$$(KVA)_{T} \leq 2(V_{m}, I) + 2\left(I^{2}, \frac{X_{I}}{2}\right)$$
(8)  
Ideal Leakage  
Transformer

where  $V_m$  and I are rms quantities and  $X_I = \omega L$ . For given values of  $V_s$  and  $P_o$ , it is also assumed that transformer size is proportional to its KVA rating and inversely proportional to the frequency.



Figure6: Equivalent circuit model of a single phase transformer.

Operating characteristics for converter A are calculated based on the principles listed above and are shown in Figure 7. Figure 7a depicts the range of control possible with variation of  $\beta$  while maintaining soft-switching operation for all four devices. It is seen that maximum power transfer occurs at  $\beta = 180^{\circ}$  and d = 0.58. The range of control obtainable at various values of d is limited by the requirement of soft-switching. Figure 7b shows a variation of the transformer KVA versus the power output P<sub>0</sub>. It is seen that at the design point, the KW/(KVA)<sub>T</sub> ratio is approximately 0.3 pu.



Figure7: Topology A



### Converter B:

A similar analysis can be carried out for converter B. This converter has only two modes of operation. The two bridges are presumed to operate with phase shift  $\phi$ . The current  $i(\theta)$  is once again given by equations 2 and 3 for modes 1 and 2 respectively. The boundary conditions now dictate that  $i(0) = -i(\pi)$  at the end of Mode 2. Solving for  $i(\theta)$ , the output power P<sub>0</sub>, is:

$$P_0 = \frac{V_s^2}{\omega L} d\phi \left[ 1 \frac{\phi}{\pi} \right]$$
(9)

1-PHASE DC-DC CONVERTER



1-PHASE DC-DC CONVERTER



Figure8: Topology B,

- a) Variation of output power versus \$\u03c6 for various values
   of d, upper boudary natural commutation locus for
   input bridge, lower boundary diode operation for
   output bridge,
- b) (KVA)<sub>T</sub> versus P<sub>0</sub> as a function of d. Boundary corresponds to diode operation for output bridge.

The constraints which define soft-switching boundaries can now be specified for the input and output bridges to be  $i(0) \le 0$  and  $i(\phi) \ge 0$  respectively. These constraints enclose the desired operating region for the converter. Exceeding the first constraint results in natural commutation of the input bridge devices and gives snubber dump. For the output converter, the constraint equation corresponds to diode bridge operation.

Figure 8a shows the variation of P<sub>0</sub> as a function of  $\phi$  for different values of d. The upper boundary corresponds to the input converter, while the lower curve represents diode operation. For d=1 it can be seen that  $\phi$  can vary over the entire range of 0 to  $\pi/2$ giving control from zero to full power. Figure 8b plots the transformer KVA against the output power for various values of d. The boundary, corresponding to an output diode bridge, is identical to that in Figure 7b with  $\beta = \pi$ .

Examination of Figure 8b shows that a transformer KVA of 1.356 pu is required with output diodes if d is to be varied over the range 0 to 1. With dual bridges, for the same transformer KVA, it is now possible to transfer  $P_0=0.59$  pu, as against a previous maximum of 0.3 pu. Taking into consideration the phase shift control possible in converter A, the KW/(KVA)<sub>T</sub> ratio increases from 0.3 to 0.435 an astounding improvement of 45% in the power density.

Compared to normal hard switched converters with  $KW/(KVA)_T$  ratios approching unity, this may seem to be very poor transformer utilization. However, if the switching frequency for the proposed converter can be made substantially higher, actual size/weight could be much lower. Transformer sizing will be examined in greater detail later in the paper. It is apparent that along with further gains in transformer power density, a significant reduction in input/output filter size and ripple current rating will result from selecting the three phase dc/dc dual bridge converter.

## Analysis Of Three Phase DC/DC Converter

The circuit schematic of the new three phase dual bridge soft-switching ac link dc/dc converter is shown in Figure 4. The proposed converter consists of two three-phase inverter stages, each operating in a six-step mode with controlled phase shift. Using two active bridges not only permits bidirectional power flow, but also allows control at a fixed frequency. The ac link transformer is Y-Y connected and is three phase symmetric with the leakage inductances used as energy transfer elements.



Figure9: Schematic of a symmetric 3 phase transformer.

In the following analysis, it is assumed that the primary and secondary resistances of the transformer can be neglected and the turns ratio is 1:1. Figure 9 shows the schematic of the 3 phase symmetric transformer. Using the relationships  $\sum i_{ip} = 0$  and  $\sum i_{is}$ 

= 0 for Y connected transformers, the transformer equations can be derived to be

$$V_{ip}(t) = L_{pl} \frac{di_{ip}}{dt} + \left( L_{pp}^{ii} + L_{pp}^{ij} \right) \frac{di_{ip}}{dt} + \left( L_{ps}^{ii} + L_{ps}^{ij} \right) \frac{di_{is}}{dt}$$
(10)

and

$$V_{is}(t) = L_{sl} \frac{di_{is}}{dt} + \left(L_{ss}^{ii} + L_{ss}^{ij}\right) \frac{di_{is}}{dt} + \left(L_{sp}^{ii} + L_{sp}^{ij}\right) \frac{di_{ip}}{dt}$$
(11)

where  $V_{ip}$  and  $V_{is}$  are the primary and secondary voltages for the i<sup>th</sup> phase,  $L_{pl}$  and  $L_{sl}$  are the primary and secondary leakage inductances,  $L_{pp}$  and  $L_{ss}$  are self inductances and  $L_{ps}$ ,  $L_{sp}$  are the mutual inductances between the appropriate phase windings given by the superscript notation used.

Using properties of a symmetric transformer and defining

$$L_{m} = L_{ss}^{ii} + L_{ss}^{ij} = L_{pp}^{ii} + L_{pp}^{ij}$$
(12)

one can derive

$$\sigma(L_{m} + L_{sl}) \frac{di_{is}}{dt} = V_{is}(t) - \frac{L_{m}}{L_{m} + L_{pl}} V_{ip}(t)$$
(13)

$$\sigma(L_{m} + L_{pl}) \frac{di_{p}}{dt} = V_{ip}(t) - \frac{L_{m}}{L_{m} + L_{sl}} V_{is}(t)$$
(14)

where  $\sigma$  is a leakage factor given by

$$\sigma = \frac{(L_m + L_{sl})(L_m + L_{pl}) - L_m^2}{(L_m + L_{pl})(L_m + L_{sl})}$$
(15)

The value of  $\sigma$  is typically a small number around the ratio of the leakage to the magnetizing inductance.

Equations 13 and 14 are the basic equations which govern the current in the circuit. Further assuming that  $L_{sl} = L_{pl} \ll L_m$  (for 1:1 turns ratio), then equations 13 and 14 reduce to

$$L_{\sigma} \frac{di_{is}}{dt} = V_{is}(t) - V_{ip}(t)$$
(16)

$$L_{\sigma} \frac{di_{p}}{dt} = V_{ip}(t) - V_{is}(t)$$
(17)

where

$$L_{\sigma} = \sigma(L_m + L_{sl}) = \sigma(L_m + L_{pl}) \cong (L_{sl} + L_{pl})$$
(18)

The simplified single equivalent circuit reduces to that shown in Figure 6 with  $L_{\sigma} = L$ . It can be seen that with the assumptions above,  $i_{is} = i_{ip} = i(\theta)$ .

In order to calculate the three line currents, the classic six step line to neutral voltage waveform is assumed for both the primary and secondary windings. The difference between the two voltages is applied across  $L_{\sigma}$ . Six modes, corresponding to different driving voltages, can be identified over a  $180^{\circ}$  conduction cycle. Using the property of a balanced 3 phase set and  $\sum i_{ip} = 0$ , one can obtain full information by calculating two currents over 1/3 of a period.

Solving for  $i(\theta)$  over a half period  $0 \rightarrow \pi$ ,

for  $0 \le \theta \le \phi$ 

$$i(\theta) = i(0) + \frac{V_s}{3} \frac{(1+d)}{\omega L_g} \theta$$
(19)

for  $\phi \le \theta \le \pi/3$ 

$$i(\theta) = i(\phi) + \frac{V_s}{3} \frac{(1 - d)}{\omega L_{\sigma}} (\theta - \phi)$$
(20)

for  $\pi/3 \le \theta \le \phi + \pi/3$ 

$$i(\theta) = i(\frac{\pi}{3}) + \frac{V_s}{3} \frac{(2-d)}{\omega L_\sigma} (\theta - \frac{\pi}{3})$$
(21)

for  $\phi + \pi/3 \le \theta \le 2\pi/3$ 

$$i(\theta) = i(\phi + \frac{\pi}{3}) + \frac{V_s}{3} \frac{(2 - 2d)}{\omega L_{\sigma}} (\theta - \phi - \frac{\pi}{3})$$
(22)

for  $2\pi/3 \le \theta \le 2\pi/3 + \phi$ 

$$i(\theta) = i(\frac{2\pi}{3}) + \frac{V_s}{3} \frac{(1-2d)}{\omega L_{\sigma}} \left(\theta - \frac{2\pi}{3}\right)$$
(23)

for  $\phi + 2\pi/3 \le \theta \le \pi$ 

$$i(\theta) = i(\phi + \frac{2\pi}{3}) + \frac{V_s}{3} \frac{(1-d)}{\omega L_\sigma} (\theta - \frac{2\pi}{3} - \phi)$$
(24)

Equating  $i(0) = -i(\pi)$  and solving, we obtain

$$i(0) = \frac{V_s}{3\omega L_\sigma} \left[ \frac{2\pi d}{3} - d\phi - \frac{2\pi}{3} \right]$$
(25)

Using  $i(\theta)$ , the supply side dc link current can be reconstructed using the input bridge switching function, and is shown in Figure 4b. From this, the average output power is calculated to be

$$P_{0} = \frac{V_{s}^{2}}{\omega L_{\sigma}} d\phi \left[\frac{2}{3} - \frac{\phi}{2\pi}\right]$$
(26)

The above results apply for  $0 \le \phi \le \pi/3$ .

For  $\pi/3 \le \phi \le 2\pi/3$ , a similar set of equations can be derived. The average output power for  $\phi$  in this range can then be found to be

$$P_0 = \frac{V_s^2}{\omega L_\sigma} d \left( \phi - \frac{\phi^2}{\pi} - \frac{\pi}{18} \right)$$
(27)

Based on  $i(\theta)$ , the important parameters such as the transformer KVA, input and output capacitor ripple current and peak device stresses can be found.

Figure 10a is similar to Figure 8a and shows the variation of P<sub>0</sub> as a function of  $\phi$  for different values of d. Once again, the curve for d=1 shows wide range of control, i.e. from zero power for  $\phi = 0$  to maximum power for  $\phi = \pi/2$ . The lower boundary corresponds to the soft-switching locus for a diode output bridge. This locus is derived by finding the value of  $\phi = \phi_c$  such that  $i(\phi_c) = 0$ . For  $0 \le \phi \le \pi/3$  this yields the lower boundary d<sub>1</sub>.

$$\mathbf{d}_{\mathbf{l}} = 1 - \frac{3\phi}{2\pi} \tag{28}$$

The upper boundary governs the transition for the input bridge between natural commutation and soft-switching. This corresponds to the relationship i(0) = 0, which yields  $d_u$ ,





Figure10: Topology C,

- a) Variation of output power versus φ for d=0.25 to d=2.5 in steps of 0.25. Upper boundary - natural commutation for input bridge, lower boundary diode operation for output bridge,
- diode operation for output bridge, lower boundary diode operation for output bridge, b)  $(KVA)_T$  versus P<sub>0</sub> for d=0.1 to 1.0, in steps of 0.1, boundary indicates diode operation,
- c) R.M.S current in input capacitor filter.
- d) R.M.S current in output capacitor filter.

$$d_{u} = \frac{1}{1 - \frac{3\phi}{2\pi}}$$
(29)

For  $\pi/3 \le \phi \le 2\pi/3$ , the lower and upper boundaries are obtained to be

$$d_{l} = \frac{2}{3} - \frac{3\phi}{\pi}$$
(30)

$$d_{u} = \frac{1}{\frac{3}{2} - \frac{3\phi}{\pi}}$$
(31)

Figure 10b shows the KVA rating of the transformer for  $0 \le d \le 1$  as a function of the output power. The locus corresponding to output diode bridge operation is also plotted. Once again, it can be seen that maximum output power with the dual bridge is 0.46 pu with a KW/(KVA)<sub>T</sub> ratio of 0.48. For the diode bridge, P<sub>0</sub>=0.265 pu and KW/(KVA)<sub>T</sub>=0.38, a substantial difference. Figures 10c and 10d show the ripple current in the input and output capacitor filters as a function of d and  $\phi$ . It will be seen that for d=0.5 and  $\phi$ 

=  $60^{\circ}$ , the output current ripple goes to zero. At the maximum

power transfer point of  $P_0=0.46$  pu,  $\phi = 50.1^{\circ}$  and the output current ripple is 0.095 pu while the input current ripple is 0.0925 pu. For lower values of d, the output current ripple increases. However, under all conditions, the ripple is substantially smaller than for either of the single phase converters. Clearly, given the operating range of the converter, an optimization is possible which yields the smallest total filter size.

The analysis of the two dual bridge converter topologies has yielded interesting and fairly counter-intuitive results in terms of overall system power density. It is not obvious why, given a transformer, one is able to obtain more power with a dual bridge than with a single bridge. Also, it is not clear whether the resulting system, using transformer leakage inductances gives power density than a conventional hard switched dc/dc converter in which the leakage elements are parasitics. In order to examine these issues better, a fundamental component model of the system is invoked and analyzed next. Fundamental Model Analysis



Figure 11: Fundamental model of the dual bridge dc/dc converter

Once again ignoring winding resistances and core losses, the transformer model in Figure 6 can be invoked for both the single phase and three phase converters. Figure 11 shows the fundamental equivalent circuit for the two converters.  $V_S$  of angle of 0° and  $V_0$  of angle of  $\phi^\circ$  are the two voltage sources applied across the effective leakage impedence term  $X_1 = j\omega L$ . This model is identified to the synchronous machine equivalent circuit and may be expected to demonstrate similar properties. The current vector i can be calculated to be

$$\mathbf{i} = \frac{\mathbf{V}_{s}}{\mathbf{X}_{1}} \left( 1 - 2d\cos\phi + d^{2} \right)^{1/2} \angle \left[ \tan^{-1} \left( \frac{d\cos\phi - 1}{d\sin\phi} \right) \right]$$
(32)

The power output can then be calculated to be

$$P_0 = \frac{V_s^2}{X_1} d\sin\phi$$
(33)

which is identical to that for a synchronous machine. Using equation 8, the KVA of the composite transformer can be found to be

$$(KVA)_{T} = \frac{V_{s}^{2}}{X_{I}} \left[ \left( 1 - 2d^{2}\cos 2\phi + d^{4} \right)^{1/2} + \left( 1 - 2d\cos \phi + d^{2} \right) \right]$$
(34)  
Ideal Transformer Leakage

Once again, this includes leakage and ideal transformer terms with the assumption that the equivalent composite transformer will be at least as small as the two components taken separately.

As expected, maximum power is transfered at  $\phi = \pi/2$ , a result corroborated by Figures 8a and 10a. Once again, if the KW/(KVA)<sub>T</sub> is plotted against  $\phi$  for different d, d=1 shows the possibility of maximizing that ratio. Plotting P<sub>0</sub> versus (KVA)<sub>T</sub> also shows similar shapes with less than 10% error due to harmonic components being neglected. Consequently, the fundamental model is assumed to be sound and gives a good first estimate on system design.

Considering the issue of transformer size may also be easier in the fundamental model. Assume that for a given  $(KVA)_T$ , transformer size is inversely proportional to the frequency, given the core material. Given the task of designing a transformer for a conventional hard switched dc/dc converter, one chooses d = 1 and L small for maximizing transformer utilization. This implies a small value for  $\phi = \phi_h$ . If this transformer is designed for frequency  $\omega = \omega_h$ , then from equation 33 and 34, we get

$$\omega_{\rm h}L = X_{\rm l} = \frac{V_{\rm s}^2}{P_0}\sin\phi_{\rm h} \tag{35}$$

$$(KVA)_{T} = 2P_{0}\left(1 + \tan\frac{\phi_{h}}{2}\right)$$
(36)

This is approximately equal to  $2P_0$  as  $\phi_h$  is assumed small. Thus the transformer size for hard-switching will be proportional to  $S_h$ , where

$$S_{h} = \frac{2P_{0}}{\omega_{h}} \left( 1 + \tan \frac{\phi_{h}}{2} \right)$$
(37)

Similarly, examining the soft-switched converter, we can see that maximum power transfer occurs at  $\phi = \phi_s = \pi/2$  and d = 1. Under these conditions and at a frequency  $\omega_s$ , we find for the same ratings

$$_{s}L = X_{1} = \frac{V_{s}^{2}}{P_{0}}$$
 (38)

$$(KVA)_{\rm T} = 4P_0 \tag{39}$$

The size is now obtained by dividing  $(KVA)_T$  by  $\omega_s$ . Comparing the sizes of the two transformers, we can see that

ω

$$\frac{S_s}{S_h} = \frac{2\sin\phi_h}{1+\tan\frac{\phi_h}{2}}$$
(40)

where S<sub>s</sub> is the sizing constant for the soft switched converter.

As typical values for  $\phi_h$  are in the 2-10° range, it can be seen that significant reduction in transformer size are possible by switching to a scheme where the leakage inductances are the current transfer elements. This comparison is further strengthened when the losses resulting from interaction of diode reverse recovery and leakage inductance are considered for a current source output dc/dc converter.

# Comparison of Proposed Converters

The three converters proposed have been presented in detail including sufficient information for the development of operating characteristics. In the interests of maintaining reasonable paper length, only a small fraction of the converter curves are presented. In order to better compare the three topologies, Table 1 presents the detailed specifications for various components based on the equations and curves presented in the paper. The design is denormalized so as to conform to a 100 KW specification at 500 Volts / 200 Amps dc.

Examining peak device stresses, the 3 phase dual bridge offers the lowest ( $V_{ce} \cdot I_c$ ) stress at (1.265  $\cdot$  load KW) as opposed to a factor of 3.45 for converter A. Converter B is reasonably good at 1.35. Similar conclusions are seen from the transformer (KVA)<sub>T</sub> ratings. Converter C needs a transformer which is 30% smaller than for converter A and 10% smaller than for converter B. Both dual bridge converters compare reasonably well until we come to input and output filters. The total capacitive KVA needed for converter C is 40.7 KVA as opposed to 141.6 KVA for converter B and 226 KVA for converter C, a truly substantial reduction. The r.m.s current ratings calculated for capacitors are well within the capability of multilayer ceramic capacitors available commercially, provided proper packaging is done.

A value of  $\omega L$  has been specified in order to attain the

desired specifications. The actual choice of  $\omega$  and L will depend on the core material and detailed transformer design. A dual bridge 3 phase converter is presently being fabricated at the 50 KW level and detailed experimental results will be presented in a later paper. It is anticipated that limitations on transformer size minimization will be imposed by the weak scaling factors that govern how the leakage inductances reduce with size as operating frequency is increased. Although it has been shown that the dual bridge converters are inherently capable of higher power density, converter A has many

### **TABLE 1**

	Converter A single phase phase shift	Converter B single phase dual bridge	Converter C three phase dual bridge
1. <u>Converter Specs.</u> (Vs) a. Power rating (Po) b. Output voltage (Vo) c. Output current (Io) d. $\phi$ e. $\beta$	862.07V 100KW 500V (d=0.58) 200A 37.8° 0 - 180°	500V 100KW 500V (d=1) 200A 45.75°	500V 100KW 500V (d=1) 200A 50.1°
<ol> <li>Device Specs.         <ol> <li>Type of device</li> <li>No of devices</li> <li>Peak voltage</li> <li>Peak current</li> <li>(Vpk * Ipk)/KW</li> </ol> </li> </ol>	MCT	MCT	MCT
	4+4 (diodes)	4+4 = 8	6+6 = 12
	Vs = 862.07V	Vs = 500V	Vs = 500V
	401.02A	270.68A	253.45A
	3.45	1.35	1.265
3. <u>Transformer Specs.</u> ωL	2.24Ω	1.475Ω	1.15Ω
a. Max. primary volts	862.07V	500V	2Vs/3 = 333V
b. Peak primary current	401.02A	270.68A	253.45A
c. Max. sec. volts	500V = Vo	500V	2Vo/3 = 333V
d. Max. sec. current	401.02A	270.68A	253.45A
e. KVA	298.59	229.83	207.83
f. RMS current	217.06A	244.07A	162.78A
<ol> <li><u>Reactive Elements</u> <ol> <li>Input filter                 <ul></ul></li></ol></li></ol>	Vs = 862.07V	500V	500V
	196.27A	141.02A	40.17A
	169.2	70.51	20.1
<ul> <li>b. Output filter         <ol> <li>Capacitor voltage</li> <li>Capacitor rms current</li> <li>Capacitor KVA</li> </ol> </li> </ul>	Vo = 500V	500V	500V
	113.53A	142.37A	41.3A
	56.77	71.19	20.65

desirable characteristics. For higher output voltages, where active devices may be unable to operate, diode rectifiers may be the only alternative along with converter A. Further, for applications where the reliability or cost penalties of additional gate turn-off devices is unacceptable, again converter A may be the only viable option.

#### **Conclusions**

Three new dc/dc converter topologies suitable for high power-density high power applications have been presented in this paper. All three circuits operate in a soft switched manner making possible a reduction of device switching losses and an increase in switching frequency. Along with soft-switching, all proposed circuits utilize the leakage reactance of the ac link transformers as active current transfer elements and eliminate problems of interaction between these leakage inductances and diode reverse recovery. The dual bridge topologies are also capable of buck-boost operation and bi-directional power flow, although that aspect has not been analyzed in detail in this paper.

The current transfer mode of operation makes it easier to parallel multiple modules for extending the power capacity of the system. The use of a three phase ac link system dramatically reduces the capacitor ripple currents making it possible to use high power density multi-layer ceramic capacitors. The dual bridge converters are also seen to offer an unexpected gain in the power density attainable as a result of the controlled action of the two bridges. As the snubbers used are purely capacitive, these would supplement the internal device capacitance, giving a clean power structure. The total number of system components is also seen to be minimal - the input and output filter capacitors, two bridges and one transformer. All device and component parasitics are seen to be used favorably. The three phase dual bridge converter proposed has the most favorable characteristics including:

- small number of components
- low device and component stresses
- zero (or low) switching losses for all devices
- small filter components
- high efficiency (no trapped energy)
- high power density using a symmetrical 3 phase
- transformer
- bidirectional power flow
- buck-boost operation possible
- low sensitvity to system parasitics
- parallel operation possible as a result of current transfer.

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