SWITCH TRANSITIONS IN THE SOFT-SWITCHING FULL-BRIDGE PWM PHASE-SHIFT DC/DC CONVERTER : ANALYSIS AND IMPROVEMENTS

ANALYSE ET AMELIORATION DES ETATS TRANSITOIRES DANS LE CONVERTISSEUR CONTINU-CONTINU EN PONT, PWM A COMMUTATION DOUCE ET REGULATION PAR VARIATION DE PHASE

Richard Redl*

ELFI S.A.
Derrey-la-Cabuche
CH-1756 Onnens FR
Switzerland

Laszlo Balogh and David W. Edwards

Ascom Energy Systems
Murtenstrasse 133
CH-3000 Berne 5
Switzerland

(*Author for correspondence)

Résumé

Est présenté ici une analyse de la quantité de l'énergie disponible pour la commutation du bras *passif-actif* (synchronisé), dans un convertisseur en pont à commutation douce, utilisant un réseau d'aide à la commutation en fonction du courant de la charge. On démontre que le choix approprié des inductances du circuit permet d'obtenir la commutation douce dans une large plage du courant de charge, sans que les pertes de conduction deviennent importantes. Les courants dans les diodes d'écrêtage du réseau d'aide à la commutation dépendent de l'emplacement de la self de *commutation. Si la self se trouve entre le bras passif-actif* (modulé) et le transformateur, des courants excessifs parcourent les diodes d'écrêtage, résultant des pertes de commutation élevées et une défaillance éventuelle de ces diodes. Les courants de diodes ne sont pas excessifs si le réseau d'aide à la commutation est entre le bras *actif-passif* du pont et le transformateur. Les prédictions de l'analyse sont vérifiées par les expérimentations sur un convertisseur de 3 kW fonctionnant à 200 kHz.

Abstract

An analysis is presented for determining the commutating energy available for the clocked or *passive-to-active* leg of the full-bridge soft-switching converter with external commutating aid, in function of the load current. It is shown that by properly choosing the values of the circuit inductances, soft switching can be achieved over a wide range of load current without a significant penalty in conduction loss. The currents in the clamp diodes of the commutating aid depend on the location of the commutating inductor. If the inductor is connected between the *passive-to-active* leg and the transformer, excess current circulates in the clamp diodes, leading to switching losses and potential failure of the diodes. The excess current is not present if the commutating inductor is connected between the *active-to-passive* leg and the transformer. The predictions of analysis are verified by computer simulation and by experimental results taken from a 3-kW converter with 200-kHz clock frequency.

INTRODUCTION

The soft-switching full-bridge PWM phase shift dc/dc converter is the preferred choice of topology in many high-power telecom applications. The main reason is that the converter offers a unique combination of two seemingly mutually exclusive advantages of the hard-switching (or square-wave) and soft-switching (load-resonant, quasi-resonant, or multi-resonant) converters. Those advantages are:

- good exploitation of the power transistors and diodes (i.e., small conduction losses in the semiconductor devices) *and*
- small switching losses.

Additional benefits are:

- low EMI,
- full integration of the stray components (junction capacitances, leakage inductance, body diodes) in the power processor, and
- minimum amount of stored energy in the additional passive waveshaping components.

If all four switches of the converter operate with soft switching there are no switch turn-on losses. The switch turn-off losses can also be made negligible by using fast drive circuits. The dynamic losses of the output rectifier diodes remain significant, however (especially if relatively slow diodes are used, as, e.g. in medium-voltage or high-voltage applications). Also, at light load, the passive-to-active (or *P-A*) leg of the bridge loses soft switching. (The passive-to-active leg is the one which has only passive-to-active transitions. The meaning of the terms *active* and *passive* will be discussed in the next section, together with a review of the operation of the converter.)

A simple, but effective, solution to both problems is available [1], [2]. The solution is adding a commutating aid comprising a small inductor ("commutating" inductor) and two low-current diodes ("clamp" diodes) to the basic converter, as shown in Figure 1.

The presence of the commutating inductor reduces the rate of change of current in the rectifier diode at turn-off and leads to reduced voltage overshoot. An additional benefit of the commutating inductance is that it increases the amount of energy (the "commutating" energy) available for charging or discharging the capacitance which loads the junction of the two switches of the *P-A* leg. The increased commutating energy *extends the range of soft switching toward lighter loads.*

At first, we analyze the leg transitions of the converter. The commutating energy of the *P-A* leg is determined as function of the load current. We show that at heavy load the dominant source of energy are the commutating inductance and the leakage inductance. At light load the energy is supplied by the magnetizing inductance. At medium load all circuit inductances (including the output filter inductance) contribute. (Note that the contribution of the output filter inductance is negative, i.e. it reduces the commutating energy.) Our analysis reveals that the available commutating energy has a minimum between the borderline of continuous and discontinuous mode and the load current where the magnetizing current is equal to the valley of the reflected filter inductor current. By properly choosing the values of the magnetizing and commutating inductances, the minimum value of the commutating energy can be selected to ensure soft switching over a wide range of load current without a significant penalty in conduction loss.

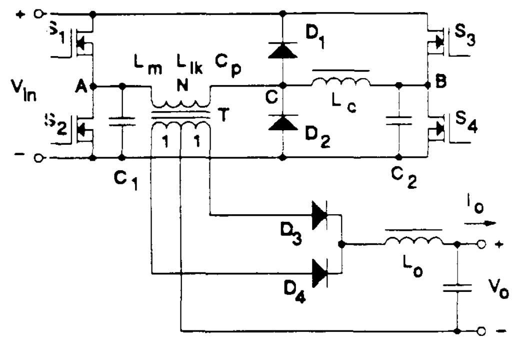

Figure 1. Soft-switching full-bridge PWM phase-shift dc/dc converter with commutating inductor and clamp diodes.

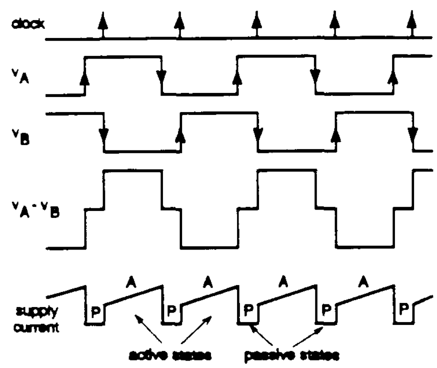

Figure 2. Waveforms and operating states.

The purpose of the two clamp diodes placed between the junction of the commutating inductor/power transformer and the positive and negative supply buses is to prevent the voltage overshoot across the output rectifier diodes in a simple lossless manner. We demonstrate that the two possible locations of the commutating inductor — between the *P-A* leg and the transformer and between the *A-P* leg and the transformer — are not *equivalent if clamp diodes are used.* The main difference between the two locations is in the currents of the clamp diodes. If the commutating inductor is connected between the *P-A* leg and the transformer, excess currents circulate in the clamp diodes. The excess diode currents cause switching and conduction losses and can lead to failure of the converter. The excess currents are not present if the commutating inductor is connected between the *A-P* leg and the transformer.

In addition to the analyses, we also give design considerations regarding (1) the choice of magnetizing and commutating inductances and (2) circuits which reduce the currents in the clamp diodes (in case the commutating inductor is between the *P-A* leg and the transformer).

The analyses are confirmed by extensive computer simulations and experimental measurements on a 3-kW telecom rectifier.

OPERATING STATES AND EQUIVALENT CIRCUITS DURING SWITCH TRANSITIONS OF THE CONVERTER

The operation of the basic soft-switching full-bridge PWM phase-shift converter has been described in detail in many references [3]-[9]. In this section we discuss only the operating states and the equivalent circuits during the switch transitions. The addition of the commutating aid causes only minor changes in the operation. The effects are mostly quantitative; they will be discussed where appropriate.

Operating states

Each switch of the bridge is driven with approximately 50% duty ratio. (Note that in order to avoid shoot-through, a small dead time must be inserted between the drive signals of a leg.) Regulation is achieved by varying the position, or phase, of the switching instants of the two legs. Figure 2 shows the various waveforms and states of the converter.

The converter has four main operating states, determined by the four main allowed on/off combinations of the switch states. When two diagonally placed switches are conducting, energy is absorbed from the input voltage source. That state is called *active*. When two switches on the same side of the power bus are conducting, there is no energy absorption. That state is called *passive*.

As can be seen, the *clocked* leg of the converter (in our case the leg with the center point marked A) switches only from the passive state to the active state. For that reason, it is called the *passive-to-active* (or *P-A*) leg. The *modulated* leg (with the center point marked B) always switches from the active state to the passive state and is called *active-to-passive* (or *A-P*) leg. As will be shown, there is a significant difference between the switching processes of the two legs.

Besides the active and passive states during which two of the switches are always in conduction, the converter has four more states when none of the switches (or their anti-parallel diodes) are in conduction. Those states are the *transition* states. Again, there are two different types of transition states (or transitions, for short). During the *P-A* transition, the *P-A* leg switches over from the passive to the active state. During the *A-P* transition, the *A-P* leg switches over from the active to the passive state.

In addition to the two types of main states (active and passive) and two types of transition states (*A-P* and *P-A*), other converter states can be also defined. Those states correspond to the conduction (or the lack of it) of the various diodes (anti-parallel, output, clamp) in the converter. We shall discuss some of those states in the relevant parts of the paper.

Active-to-passive transition

Figure 3 shows an equivalent circuit of the converter during the *A-P* transition.

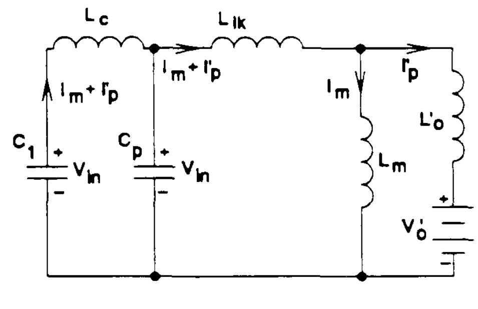

Figure 3. Equivalent circuit during the *A-P* transition.

In the equivalent circuit the transformer is represented by the magnetizing inductance L_m, the leakage inductance L_{lk}, and an equivalent parasitic capacitance C_p (all transformed to the primary side). The location of the single equivalent parasitic capacitance is a function of the transformer geometry and winding configuration and is somewhat arbitrary. C_1 is the sum of the equivalent capacitances of the two switches of the leg and the external stray and snubber capacitances loading the leg center point. (The equivalent capacitance of a switch is the linear capacitance which stores the same amount of energy as does the nonlinear junction capacitance when the voltage across it is V_{in}.) In Figure 3, the voltages across the capacitances and the currents in the inductances are the initial conditions of the transition. I_m is the peak value of the magnetizing current, I_p' is the peak value of the current in the output inductor transformed to the input side, V_o' is the output voltage transformed to the input side.

Above a very light load, the stored energy of the output inductor is much larger than the other stored energies. This fact makes the introduction of the much simplified equivalent circuit of Figure 4 possible. That equivalent circuit yields a linear capacitor-voltage vs. time function. The simplified equivalent circuit is usually valid down to a few percent of the full load current.

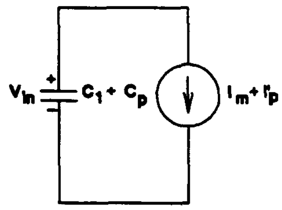

Figure 4. Simplified equivalent circuit during *A-P* transition.

The *A-P* transition is completed when either the anti-parallel diode of the other, second, switch in the leg (the body diode of the MOSFET switch or an external diode) comes into conduction, or the second switch is turned on, whichever happens earlier. A delay time must be inserted between the turn-off of the first switch and the turn-on of the second switch. In order to avoid lossy charge or discharge of the capacitance loading the junction of the two switches, the delay time must be above a certain limit. Assuming that the minimum current in the primary winding is large enough that the linear approximation of the leg-voltage transition is valid, the limit is

$$t_{d(A-P)} > \frac{V_{in(max)}(C_1 + C_p)}{I_{pr(min)}} \qquad (1)$$

Theoretically the upper limit for the delay time is the minimum duration of the passive state. In practice, it is advisable to turn on the second switch well before the *P-A* transition begins, so that the minority charge carries are removed from the body diode before reverse voltage is forced across that diode.

Passive-to-active transition

Depending on the operating mode (continuous or discontinuous) and circuit parameters, the converter can be characterized by three different equivalent circuits during the *P-A* transition.

Figure 5 shows the equivalent circuit when the converter is in DICM. (DICM is short for discontinuous inductor current mode.)

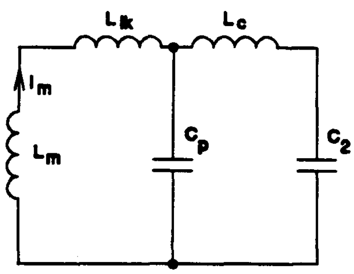

Figure 5. Equivalent circuit during the *P-A* transition, DICM.

In DICM, the magnetizing, leakage and commutating inductances carry the same current (I_m) when the transition begins. The equivalent circuit remains valid throughout the transition.

In CICM , the transition begins with the equivalent circuit shown in Figure 6. (CICM is short for continuous inductor current mode.)

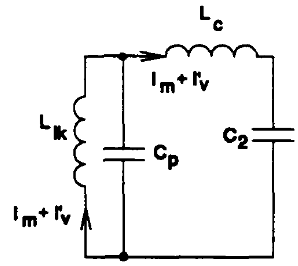

Figure 6. Equivalent circuit at the beginning of the *P-A* transition in CICM.

If the magnetizing current is smaller than the reflected valley current, the transformer remains in shorted condition until the transition is over. (The transition ends either because the voltage across C_1 swings all the way to the other bus, or because the other switch is turned on by the drive signal.) If the magnetizing current is larger than the reflected valley current, two things can happen: (1) The transition is over before the current in the leakage inductance drops below the difference between the magnetizing current and the reflected inductor current. The equivalent circuit of Figure 6 remains valid throughout the transition. (2) The current in the leakage inductance drops below the difference between the magnetizing current and the reflected inductor current. When that happens the transformer comes out from the shorted condition and the equivalent circuit of Figure 7 becomes applicable.

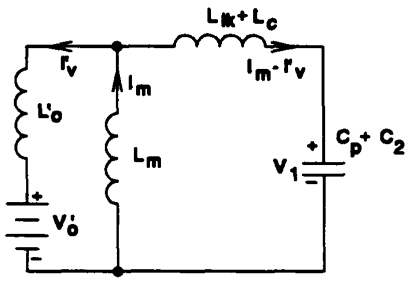

Figure 7. Equivalent circuit during the *P-A* transition in CICM when the transformer comes out from the shorted condition.

In the figure, for simplicity, we use the following assumptions: 1. The filter inductor current has not changed during the first part of the transition, i.e. its value is I_v (the valley current). 2. The ladder network of L_{lk}, C_p, L_c, and C_2 can be replaced by the network comprising only one inductor with an inductance of $L_{lk} + L_c$ and one capacitor with a capacitance of $C_p + C_2$. The initial capacitor voltage V_1 is calculated from the law of conservation of energy; the result is

$$V_1 = 2\sqrt{\frac{L_c + L_{lk}}{C_p + C_2} I_m I_v'} \qquad (2)$$

COMMUTATION ENERGY FOR THE *P-A* LEG

As the equivalent circuits demonstrate, the energy stored in the filter inductor does not help the *P-A* transition. This is why the *P-A* leg can lose soft switching at a much lower load current than the *A-P* leg. It is also evident that by storing energy in the magnetizing inductance, the range of soft switching of the *P-A* leg can be extended.

For an optimization of the design, it is useful to determine the energy available for commutation of the *P-A* leg as function of the load current, with the magnetizing inductance and the leakage plus commutating inductances as parameters.

In DICM, from the equivalent circuit of Figure 5, the commutating energy is

$$E_c = \frac{1}{2}(L_m + L_{lk} + L_c)I_m^2 \tag{3}$$

In DICM the magnetizing current is function of the load current because the duty ratio also depend on the load current. A simple derivation yields

$$I_m = \frac{1}{2L_m}\sqrt{\frac{2I_oL_oNV_oV_mT}{\left(\frac{V_m}{N} - V_o\right)}} \tag{4}$$

The components and voltages are those defined in Figure 1; *T* is the period of the clock frequency (i.e. the output frequency).

The load current at the borderline of DICM and CICM is

$$I_o = \frac{V_oT}{2L_o}\left(1 - \frac{V_o}{V_m}N\right) \tag{5}$$

In CICM, when the magnetizing current is smaller than the reflected valley current, the commutating energy is

$$E_c = \frac{1}{2}(L_{lk} + L_c)(I_m + I_v')^2 \tag{6}$$

In CICM the magnetizing current is

$$I_m = \frac{V_oNT}{2L_m} \tag{7}$$

The reflected valley current is

$$I_v' = \frac{I_o}{N} - \frac{V_oT}{2L_oN}\left(1 - \frac{V_o}{V_m}N\right) \tag{8}$$

For the case when the magnetizing current is larger than the valley current, the commutation energy is

$$E_c = \frac{1}{2}L_m(I_m - I_v')^2 + \frac{1}{2}(L_{lk} + L_c)(I_m + I_v')^2 \tag{9}$$

Figure 8 shows the calculated commutating energy in function of the load current for a converter with the following parameters:

V_{in}	=	380	V
V_o	=	52	V
T	=	5	µs
N	=	14/3	
L_o	=	18	µH
$I_{o(max)}$	=	50	A

The minimum commutating energy needed for ensuring soft switching of the *P-A* leg of the example converter is

$E_{c(min)}$ = 200 µJ

which corresponds to an equivalent loading capacitance of

$C_p + C_t$ = 2.77 nF

The sum of the leakage and commutating inductances is

$L_{lk} + L_c$ = 11 µH

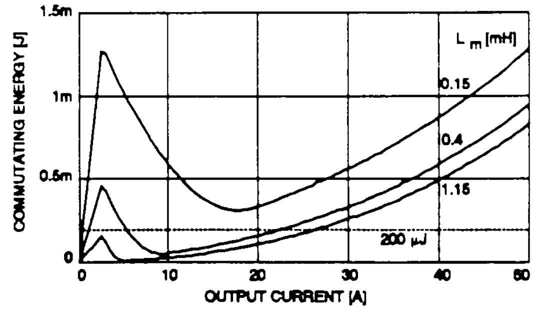

Figure 8. Commutating energy vs. output current (parameter: magnetizing inductance).

In Figure 8 the parameter is the magnetizing inductance. As can be seen, with 150 µH magnetizing inductance, soft switching is maintained from full load down to practically zero load.

In Figure 9 the parameter is the sum of the leakage and commutating inductances. The magnetizing inductance is

L_m = 1.15 mH

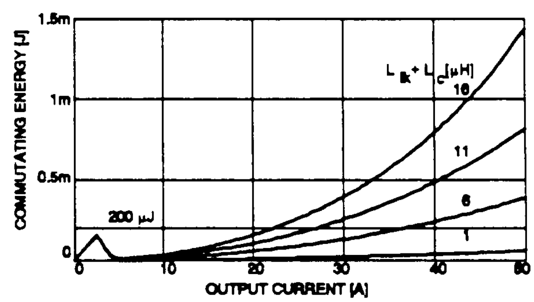

Figure 9. Commutating energy vs. output current (parameter: sum of leakage and commutating inductances).

By comparing Figures 8 and 9, it becomes obvious that reducing the magnetizing inductance is a more effective means of maintaining soft switching at light load than increasing the commutating inductance. Although a complete loss analysis is outside the scope of this paper, it is clear that the penalty of using low magnetizing inductance is increased conduction losses in the switches and in the windings of the transformer and the commutating inductor. Increasing the commutating inductance also increases the conduction losses, but much more gradually.

CURRENTS IN THE CLAMP DIODES

A task of the commutating inductance added in series with the primary winding of the transformer is to increase the energy available for the transition of the *P-A* leg. Without parasitic capacitances around the transformer and the output rectifier diodes, the inductor itself would be sufficient to accomplish that task. Unfortunately, the parasitic capacitances cause ringing across the transformer windings and lead to excessive voltage overshoot across the rectifier diodes. To reduce the ringing and overshoot, dissipative [5] and nondissipative [10] clamps have been recommended. Another, less expensive but equally effective, solution was proposed in [1]. The idea is to add clamp diodes between the positive or negative supply bus and the junction of the commutating inductor and the transformer terminal. The clamp diodes prevent the ringing across the primary winding of the transformer, so now the ringing will be excited only by the energy stored in the leakage inductance. That energy is usually only about 10% of the total energy stored in the commutating inductance as is, therefore, much easier to absorb. The excess energy stored in the commutating inductance is returned either to the supply through the clamp diodes or to the output through the transformer.

Due to the presence of the clamp diodes, the two possible locations of the commutating inductor — between the *P-A* leg and the transformer and between the *A-P* leg and the transformer — are not equivalent. The main difference between the two locations is in the currents of the clamp diodes. Figure 10 shows the current in the clamp diode D_2 (in Figure 1), together with the bridge voltage v_{AB} and transformer voltage v_{AC}, for the case when the commutating inductor is at the *P-A* leg. (Note that in Figure 1, the leg marked with B is the *P-A* leg.)

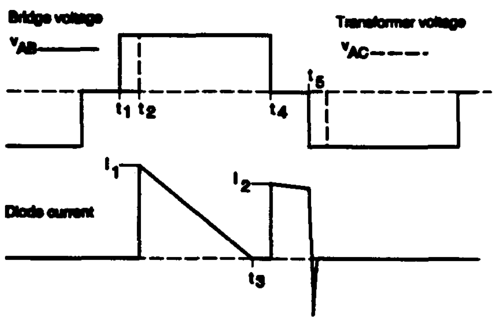

Figure 10. Bridge and transformer voltages and current in the clamp diode D_2 when the commutating inductor is at the *P-A* leg.

Figure 11 shows the current in the clamp diode D_1, together with the bridge voltage v_{AB} and transformer voltage v_{CB}, for the case when the commutating inductor is moved to the *A-P* leg. (Note that in Figure 1, the leg marked with A is the *A-P* leg.)

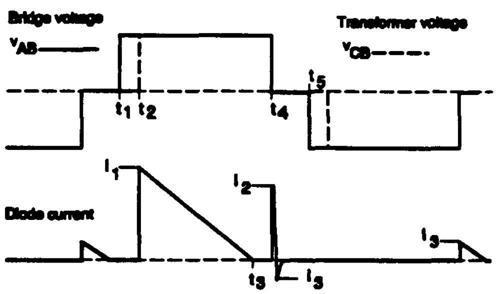

Figure 11. Bridge and transformer voltages and current in the clamp diode when the commutating inductor is at the *A-P* leg.

As can be seen, during the active state a current ramp flows in the clamp diode. The ramp begins after the passive-to-active transition, when the voltage across the transformer reaches V_{in}. Note that the voltage build-up across the transformer is delayed by the presence of the commutating inductor. The delay time $t_1 - t_2$ is

$$t_1 - t_2 = \frac{I_v' L_c}{V_{in}}$$

The peak I_1 of the current ramp is approximately

$$I_1 = \frac{V_{in}}{\sqrt{\frac{L_{lk} + L_c}{C_1 + C_p}}} \qquad (11)$$

The ramp decays with a slope which is approximately equal to the sum of the slopes of the magnetizing current and the reflected filter inductor current plus the diode forward voltage divided by the commutating inductance. If the time constant $(L_{lk} + L_c)/R_{on}$ is small, an exponential decay term also appears. In the active state the location of the commutating inductance has no effect on the current waveform of the clamp diode.

Due to various transformer parasitics, current flows in the clamp diode at the active-to-passive transition, too. (The peak value I_2 of that current cannot be as simply calculated as that of the current during the active state.) The location of the commutating inductance has a very distinct effect on the waveform. If the commutating inductor is at the *P-A* leg, the current will be nearly constant during the whole duration of the passive state. The current circulates in the commutating inductor, the clamp diode, and the switch which is connected to the commutating inductor on the same side of the supply bus as is the clamp diode (S_4 in the case of D_2). The decay is influenced by several factors, including the slope of the reflected filter inductor current and the effect of the forward voltage drop of the clamp diode. The reflected slope increases the current, while the diode forward voltage drop decreases it. Computer simulations and observations of the operation of the actual converter show that the resulting slope is around zero during normal operation and is positive (i.e. the current increases) at small duty ratios.

At the end of the passive state, the clamp diode turns off with a di/dt of V_{in}/L_c. That di/dt value is high enough to develop a large reverse current pulse in the clamp diode and to cause significant switching losses.

If the commutating inductor is at the *A-P* leg, current flows only very briefly in the clamp diode at the active-to-passive transition. The peak current I_2 of the diode will be approximately the same as before, but the duration of the current pulse will be very short. The reason is that the excess current now flows through the clamp diode which is connected to the opposite bus. This produces a high decay rate (V_{in}/L_c) and quick termination of the current. (Observed current pulse durations are in the 50 to 100 ns range.) The high decay rate also causes switching losses in the diode but the total loss is still much less than before.

It is interesting to note that in the second location of the commutating inductor, a third current pulse appears in the clamp diode at the moment when the active-to-passive transition takes place in the other direction. The peak of the third current pulse I_3 is the same as the reverse peak current of the other clamp diode. The current pulse decays with the sum of the slope of the reflected filter inductor current and the forward voltage drop of the clamp diode divided by L_c.

DESIGN CONSIDERATIONS AND EXPERIMENTAL RESULTS

Choice of magnetizing and commutating inductances

The choice of the magnetizing and commutating inductances depends on the operating conditions of the converter. If the converter operates close to full load most of the time, it is sufficient to use only the commutating inductance for maintaining soft switching. If the load varies over a wide range, it might become necessary to reduce the magnetizing inductance of the transformer to ensure soft switching in the whole load-current range.

As was shown in the section on commutation energy, it is possible to extend the range of soft switching practically all the way to zero load current by reducing the magnetizing inductance. The excess magnetizing current, however, leads to excess conduction losses. We clearly have an optimization problem. To achieve the highest average efficiency, we must also take into account the probability distribution function of the load. A discussion of the optimization of the converter for efficiency is the subject of a planned future paper.

Circuits to reduce currents in the clamp diodes

When the commutating inductor is at the *P-A* leg, the dc and rms current and also the switching losses in the clamp diodes are much larger than when the inductor is at the *A-P* leg. At small duty ratios where there is not enough time for the current to decay to zero during the active state, the switching losses can easily cause thermal runaway and, eventually catastrophic failure, of the diodes.

Although there seems to be no practical reason which would justify keeping the commutating inductor at the *P-A* leg, it is easy to prevent current runaway in the clamp diodes even in that position. Figure 12 shows a simple circuit which works satisfactorily. The resistor *R* speeds up the decay of the current. In another circuit (Figure 13), the same is achieved by using two back-to-back Zener diodes.

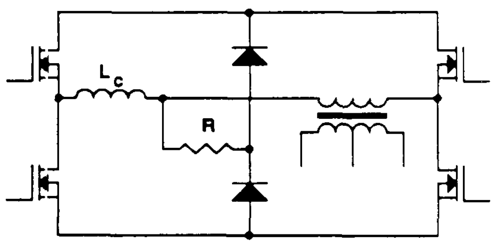

Figure 12. Resistive damping of the currents in the clamp diodes.

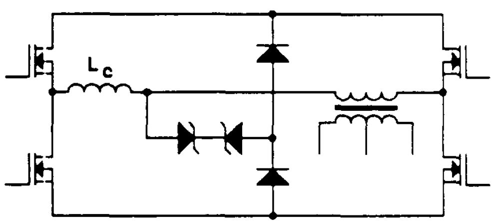

Figure 13. Damping of the currents with Zener diodes.

The extra voltage drops introduced by the additional components (resistor or Zener diodes) appear across the output rectifiers. During the design of the converter, those extra voltage drops must be taken into account. The other design parameter is the dissipation. Because the currents in the clamp diodes are not easily predictable, computer simulations and/or laboratory tests are recommended to refine the paper designs.

Although not essential, a small amount of resistive damping might be also beneficial in the case when the commutating inductor is at the *A-P* leg. The damping ensures that the clamp diodes are not conducting when a transition takes place in the converter and so it helps further reducing the switching losses.

Experimental results

We recorded the current waveforms in the clamp diodes of a 3-kW converter under various conditions. Figures 14 through 16 show those waveforms. The parameters of the converter are given in a previous section discussing the commutating energy.

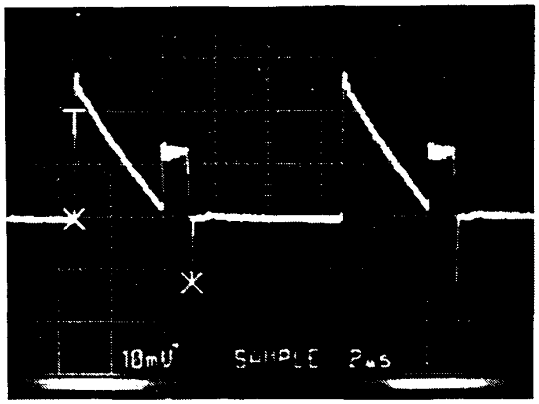

Figure 14. Current in a clamp diode without damping resistor at full load. The commutating inductor is at the *P-A* leg. Scales: 1 A/div., 2 μs/div.

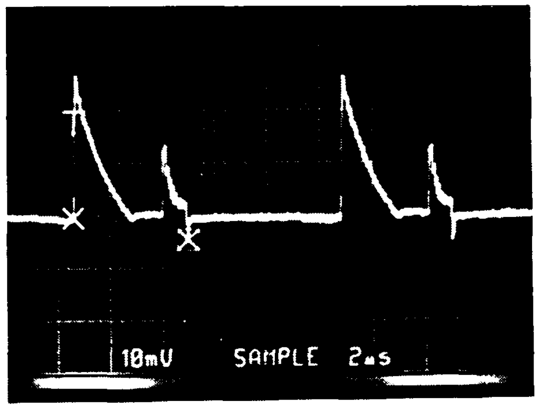

Figure 15. Current in a clamp diode with a 4.2 ohm damping resistor at full load. The commutating inductor is at the *P-A* leg. Scales: 1 A/div., 2 μs/div.

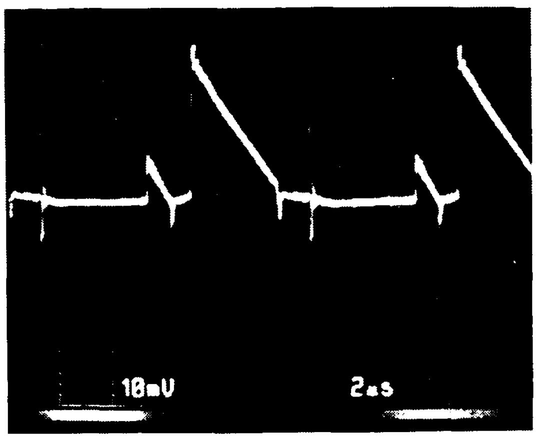

Figure 16. Current in a clamp diode without damping resistor at full load. The commutating inductor is at the *A-P* leg. Scales: 1 A/div., 2 μs/div.

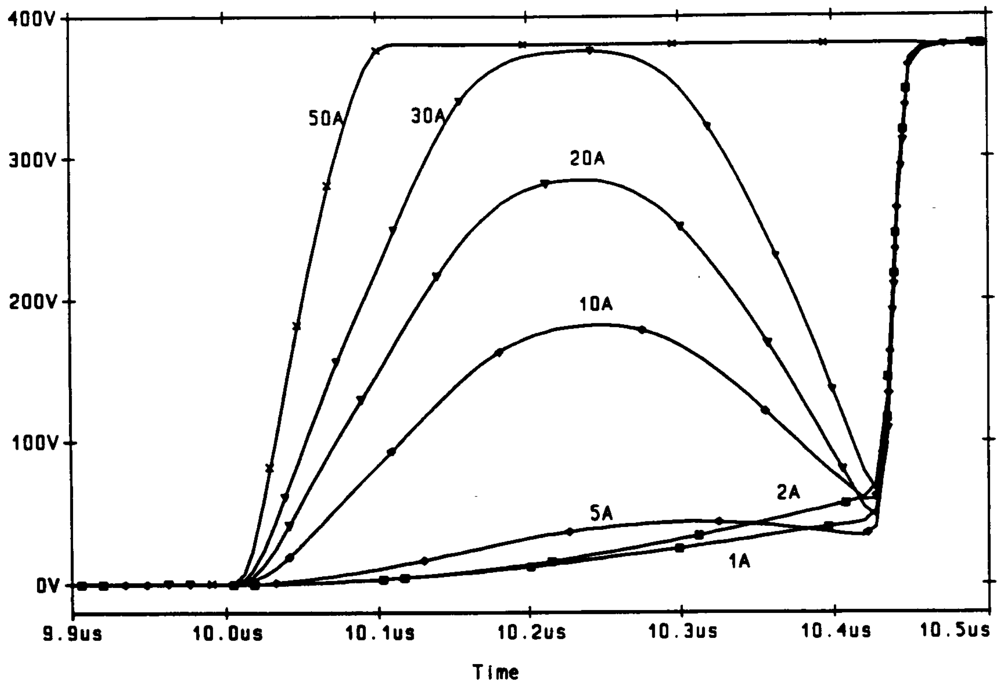

Figure 17. Transition of the *P-A* leg. L_m = 1.2 mH, L_c = 10 μH, L_g = 1.7 μH. Parameter: output current.

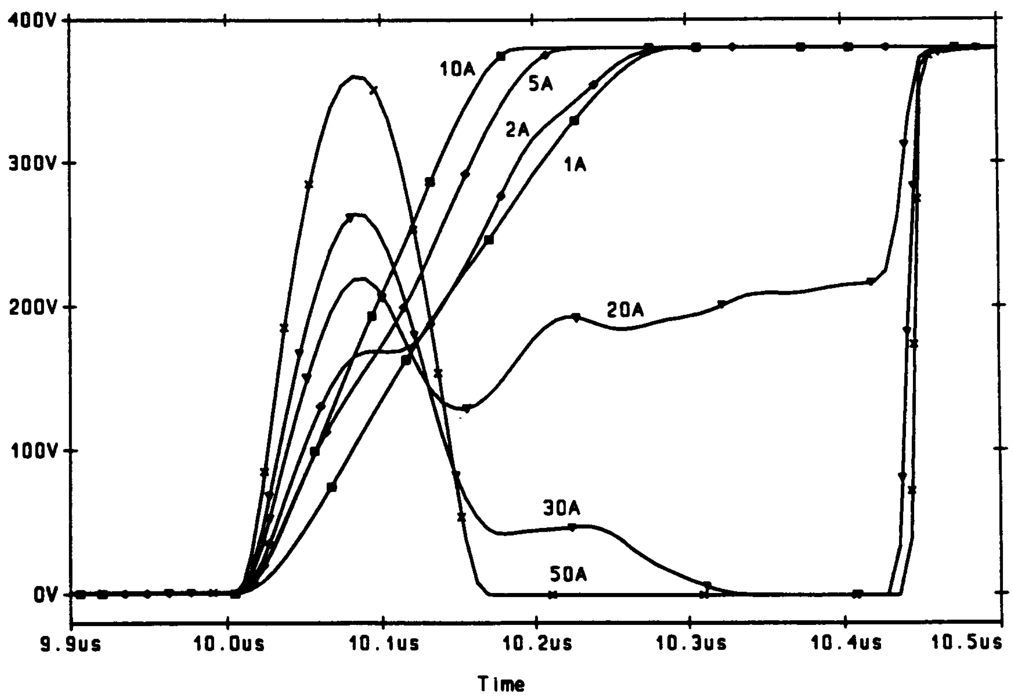

Figure 18. Transition of the *P-A* leg. L_m = 150 μH, L_c = 0, L_g = 1.7 μH. Parameter: output current.

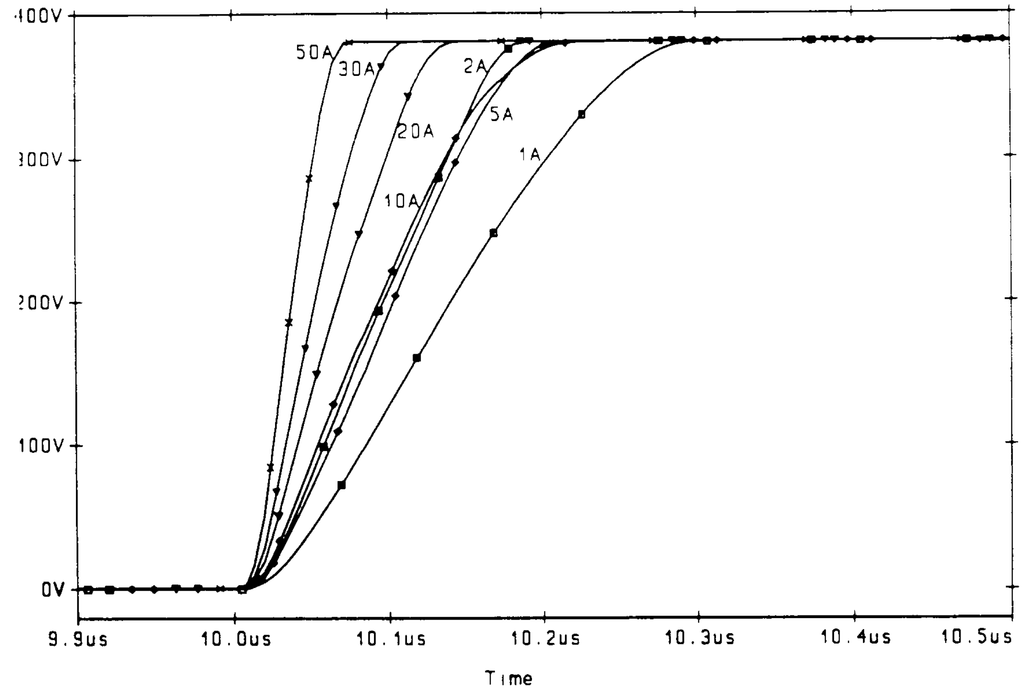

Figure 19. Transition of the *P-A* leg. L_m = 150 μH, L_c = 10 μH, L_k = 1.7 μH. Parameter: output current.

RESULTS OF COMPUTER SIMULATIONS

We carried out SPICE simulation of the 3-kW converter, to verify the validity of the simplified models used for determining the transitions and calculating the commutation energy available for the *P-A* leg. Figures 17 through 19 show some of the results of the simulation.

Figure 17 shows the voltage transition of the *P-A* leg with the parameter values L_m = 1.2 mH and $L_c + L_k$ = 11.7 μH . Figure 18 shows the voltage transition of the *P-A* leg with the parameter values L_m = 150 μH and $L_c + L_k$ = 1.7 μH . Figure 19 shows the voltage transition of the *P-A* leg with the parameter values L_m = 150 μH and $L_c + L_k$ = 11.7 μH . On all three figures the parameter is the output current which varies between 1 A and 50 A.

The waveforms on Figure 17 show that with high magnetizing and commutating inductances soft switching is achieved only above 30 A, i.e. 60% of the full load. This is expected from the data in Figure 8. Those data indicate that the commutating energy becomes insufficient for soft switching at a load current below about 27 A.

The waveforms on Figure 18 show that with low magnetizing inductance and without externally added commutating inductance soft switching is achieved only below 10 A. The waveforms on Figure 19 show that with low magnetizing inductance and high commutating inductance soft switching is achieved from almost zero load to full load. This is also expected from the data in Figure 8.

SUMMARY

We presented equivalent circuits for the switch transitions of the full-bridge soft-switching converter with an external commutating inductor and clamp diodes. We also determined the commutating energy available for the *passive-to-active* leg, taking into account both the magnetizing inductance and the commutating inductance. The commutating energy shows a minimum in function of the load current.

In addition to the switch transitions, we investigated the currents in the clamp diodes. Both the currents and the switching losses of the clamp diodes are smaller when the commutating inductor is located between the *active-to-passive* leg and the transformer. If needed, further reduction of the currents can be achieved by adding a small resistor or two Zener diodes to the converter.

The predictions based on analyses using simple equivalent circuits were verified by experimental date taken from a 3-kW converter with 200-kHz clock frequency and by extensive SPICE simulations.

REFERENCES

[1] R. Redl, N. O. Sokal, L. Balogh, "A novel soft-switching full-bridge DC/DC converter: analysis, design considerations, and experimental results at 1.5 kW, 100 kHz," *PESC '90 Record*, pp. 162–172.

[2] R. Redl and L. Balogh, "Soft-Switching Full-Bridge DC/DC Converting," U.S. Patent 5,198,969, March 30, 1993.

[3] R. A. Fisher, et al, "A 500 kHz, 250 W dc-dc converter with multiple outputs controlled by phase-shifted PWM and magnetic amplifiers," *HFPC • May 1988 Proceedings*, pp. 100–110.

[4] M. M. Walters and W. M. Polivka, "A high-density modular power processor for distributed military power systems," *Proceedings of APEC '89*, pp. 403–412.

[5] L. H. Mweene, et al, "A 1 kW, 500 kHz front-end converter for a distributed power supply system," *Proceedings of APEC '89*, pp. 423–432.

[6] D. M. Sable and F. C. Lee, "The operation of a full bridge, zero-voltage-switched, PWM converter," *Proceedings of VPEC Seminar '89*, pp. 92–97.

[7] D. B. Dalal, "A 500-kHz multi-output converter with zero voltage switching," *Proceedings of APEC '90*, pp. 265–274.

[8] J. A. Sabate, et al, "Design considerations for high-voltage high-power full-bridge zero-voltage-switching PWM converter," *Proceedings of APEC '90*, pp. 275–284.

[9] Q. Chen, et al, "Optimization and design issues of low output voltage, off-line, zero-voltage-switched PWM converters," *Proceedings of APEC '92*, pp. 73–80.

[10] J. A. Sabate, et al, "High-voltage, high-power, ZVS, full-bridge PWM converter employing an active snubber," *Proceedings of APEC '91*, pp. 158–163.