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Leakage Inductance (Part 2): Overcoming Power Losses And EMI

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Part 1 of this article series focused on the science and math of leakage inductance and described methods for calculating leakage inductance and related quantities. Here, part 2 will show how leakage inductance can be our foe. This part describes some of the problems that leakage inductance creates and how to deal with those problems.

Leakage inductance is our foe when it creates problems such as power losses and electromagnetic interference (EMI). For unregulated outputs in transformer-isolated converters, load regulation is made worse by the existence of leakage inductance. In most isolated converters, leakage inductance contributes to both power losses and EMI, but there are ways in which power losses and EMI can be avoided by design.

The first course of action is to design the transformer for low leakage inductance. Part 1 of this article series suggested how lowering leakage inductance could be accomplished. But sometimes lowering leakage inductance is too costly or requires more space than what's available, so other methods are used to deal with the power losses and EMI that result from leakage inductance.

In this part 2, the various clamp and snubber options are discussed including dissipative (RCD clamp and RC snubber) and non-dissipative (LCD clamp) solutions as well as the more-complex active-clamp solutions. The pros and cons and varying requirements of the different approaches are discussed mainly within the context of the flyback topology. But there is also discussion of the LCD clamp in the single-ended forward converter, and active clamps in the coupled-boost converter. Finally, this part looks at techniques for improving load regulation that has been degraded by the effects of leakage inductance.

Leakage Inductance-Induced Ringing

Fig. 1 illustrates one of the problems of an unclamped flyback converter. Ringing will occur in an LC network whenever there is either a voltage applied to the inductor or a current in the inductor.

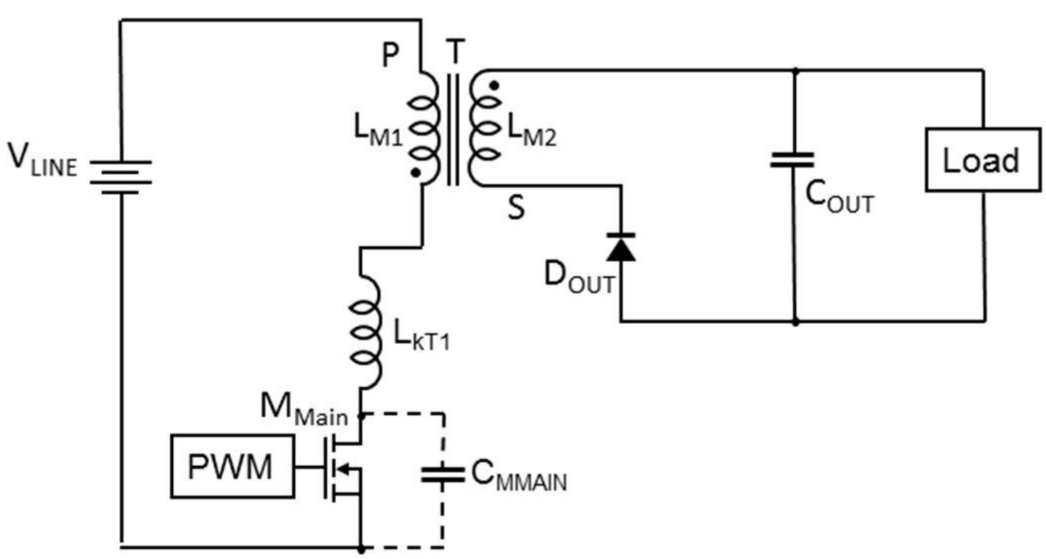

Fig. 1. An unclamped flyback converter illustrating circuit elements that contribute to ringing during the off-state of the main switch.

During the on-state of the main switch, M_{MAIN}, energy in the flyback transformer builds linearly in both the primary magnetizing inductance, L_{M1}, and in the leakage inductance, L_{kT1}, referred to the primary winding. Recall from part 1 of this article series that leakage inductance in the transformer model can be referred to the primary, secondary, or both. The transformer behaves the same regardless of where the leakage inductance, or any uncoupled series inductance, is placed.

During the on-state of the main switch there can be no ringing in the primary circuit because both ends of the primary winding are connected to relatively fixed voltages, which are referred to as ac grounds. An ac ground can be a ground node, a low-impedance dc voltage-source node, or a node connected through a relatively large-value capacitor to another ac ground. The dotted terminal of the primary winding is connected through the main switch to ground and the undotted terminal is connected to an ac ground at the line input, which likely also contains an input capacitor (not shown). Therefore, the primary circuit is effectively clamped during the on-state of the main switch.

Strictly speaking, there is ringing when a winding is connected to a capacitor that is an ac ground. But if the capacitor is large enough to be considered an ac ground, then the ringing frequency is very likely much less than the switching frequency of the converter. Consequently, the ringing effects, which include changes in capacitor voltages, are insignificant during a switching cycle.

On the other hand, when the main switch turns off there is energy, E, stored in the leakage inductance,

$$E = \frac{1}{2} L_{kT1} {i_{peak}}^2$$

where L_{kT1} is the leakage inductance referred to the primary winding and i_{peak} is the peak primary winding current. The energy, E, in the leakage inductance results in ringing with capacitive elements connected to the drain of the main switch, which include the intrinsic capacitance of the main switch, C_{MMAIN}, intrawinding and interwinding capacitances of the transformer primary winding, and printed circuit board (PCB) parasitic capacitance.

In many, if not most cases, if the ringing is not eliminated, it will cause an electromagnetic compliance (EMC) failure. In the Fig. 1 circuit, the primary circuit is unclamped during the off-state of the main switch and ringing will exist at the drain terminal of the main switch. A solution to this ringing problem is to clamp the primary winding during the off-state.

During the off-time of the output rectifier, D_{OUT}, the secondary winding is unclamped, as can be seen from Fig. 2, which illustrates leakage inductance, L_{kT2}, referred to the secondary winding. During the on-time of the output rectifier, the secondary winding is clamped and cannot ring because both terminals of the secondary winding are connected to ac grounds. The dotted terminal of the secondary winding is connected to the positive terminal of the output capacitor, C_{OUT}, and the undotted terminal is connected through the on rectifier, D_{OUT}, to the negative terminal of C_{OUT}. During the on-time of the output rectifier, current flows in the secondary winding and there is magnetic energy stored in both the magnetizing inductance of the transformer and in the leakage inductance, L_{kT2}, referred to the secondary winding.

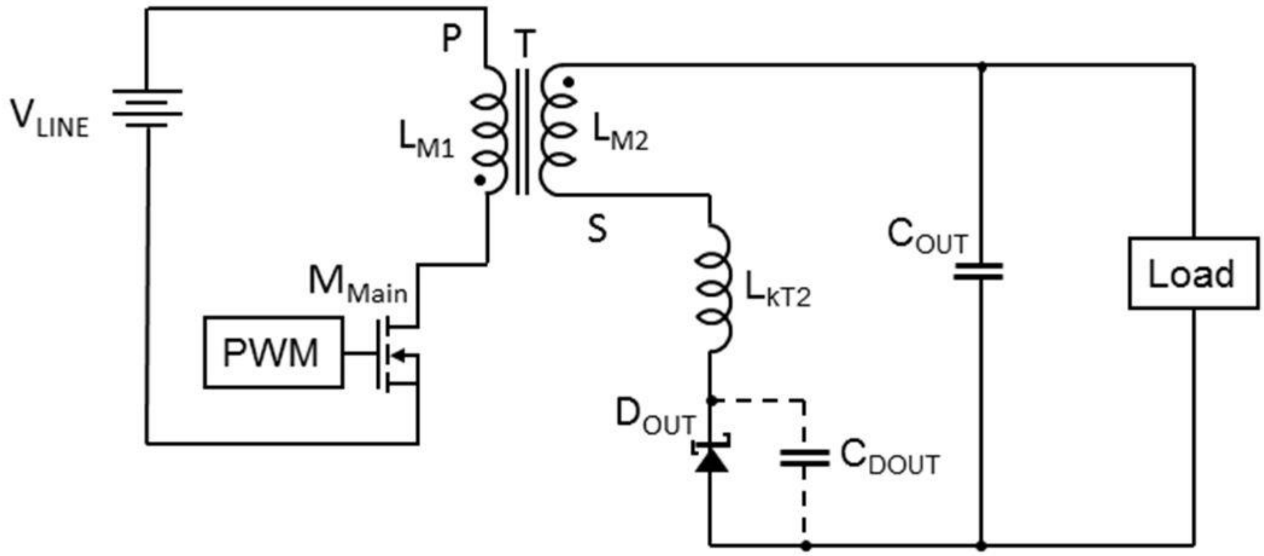

Fig. 2. An unclamped flyback converter illustrating circuit elements that contribute to ringing during the off-state of the rectifier switch.

In a discontinuous conduction mode (DCM) flyback converter, the output rectifier, D_{OUT}, turns off when the secondary winding current reaches zero. When the output rectifier turns off, there may be a small reverse recovery and associated reverse-recovery current, if the output rectifier is a junction rectifier. After the output rectifier turns off, the cathode voltage of D_{OUT} increases.

During the voltage transition of D_{OUT} there is current that charges the intrinsic capacitance of the output rectifier, C_{DOUT}, and other parasitic capacitances connected to the cathode terminal of D_{OUT}. This results in stored energy in the capacitances connected to the cathode terminal of D_{OUT} due to voltage applied to the capacitances.

Meanwhile, stored energy in the leakage inductance, L_{kT2}, is created due to capacitor charging currents flowing in L_{kT2}. Ringing will result from the exchanges of stored energy between the parasitic capacitances and the leakage inductance. This ringing may or may not be sufficient to cause an EMC failure depending on the magnitude of the energies.

If the flyback operates in continuous conduction mode (CCM) then the energies exchanged during the ringing are larger due to the fact that the rates of voltage and current slewing are much greater than in the DCM case. In the CCM case, the transition is driven by the turn-on of the main switch and results in higher currents in capacitive parasitics and higher voltages applied to leakage inductance. The likelihood of an EMC failure is more likely in the CCM case.

In the CCM case, when the main switch turns on, the transformer winding polarity reverses. During the turn-on transition, at the first instant that the main switch turns on, the voltage applied at the drain of the main switch is the line voltage, V_{Line}, plus the primary winding voltage, $V_{Reflected}$, which is equal to the load voltage reflected into the primary circuit by the transformer. At the instant that the main switch turns on, the main switch voltage is,

$$V_{MMAIN} = V_{Line} + V_{Reflected} \quad .$$

Current increases in the main switch as its gate voltage reaches and passes the gate threshold voltage. As the main switch turns on, capacitances at the drain of the main switch discharge and the main switch voltage collapses, reversing the winding polarity of the transformer. The reversal of winding voltage creates a large applied voltage to L_{kT2} causing the secondary winding current to ramp down very quickly.

While the secondary winding current ramps down, the primary winding current ramps up quickly. D_{OUT} remains on during the current transition but turns off quickly due to the high voltage applied to L_{kT2}. The high voltage applied to L_{kT2} remains but lessens after D_{OUT} turns off and reverse recovers as the cathode voltage of D_{OUT} rises, after the secondary winding current has dropped to zero. As the D_{OUT} cathode voltage rises, the (now reversed) current in L_{kT2} rises quickly creating stored energy in L_{kT2} during the reverse recovery and voltage transition of D_{OUT}. This stored energy creates ringing with the capacitances at the cathode of D_{OUT}.

Dissipative Clamps And Snubbers For Leakage Inductance-Induced Ringing

Most leakage inductance-induced ringing can be eliminated by adding a clamp. The ringing that remains after the addition of a clamp can be eliminated with an RC snubber. Clamping an inductor involves connecting both ends of the inductor to ac grounds. The most commonly used type of clamp for leakage inductance-induced ringing is the RCD clamp, shown in Fig. 3a.

The RCD clamp is a simple dissipative clamp. It is dissipative because it dissipates leakage inductance energy that would otherwise create ringing. However, the dissipated power is an undesirable power loss and reduces efficiency. At the end of the on-state, leakage inductance prevents the instantaneous decrease in primary current and instantaneous increase in secondary current desired for maximum efficiency. When the main switch turns off, the stored magnetic energy in the transformer causes the drain voltage of the main switch to rise until the clamp diode, D_{CLAMP1}, becomes forward biased.

An equivalent primary circuit during the time interval in which D_{CLAMP1} conducts is illustrated in Fig. 3b. D_{CLAMP1} remains forward biased as long as there is energy remaining in L_{kT1}. How long it takes for the leakage inductance energy to dissipate depends on how fast the current in the leakage inductance ramps down to zero. The rate at which the leakage inductance current, which is also the primary winding current, i_{PRI}, drops to zero depends on the voltage applied to the leakage inductance, as indicated in equation 1.

$$V_{LK} = -L_{kT1}\frac{di_{PRI}}{dt} \quad \text{and} \quad \frac{di_{PRI}}{dt} = -\frac{V_{LK}}{L_{kT1}} \qquad (1)$$

V_{LK} is the voltage applied to the leakage inductance, which is given by (neglecting the forward voltage of D_{CLAMP1}):

$$V_{LK} = V_{CCLAMP1} - V_{Reflected} \qquad (2)$$

where

$$V_{Reflected} = \frac{N_P}{N_S}V_{OUT} \quad . \qquad (3)$$

$V_{Reflected}$ is the output voltage reflected into the primary circuit and the voltage applied to the magnetizing inductance referred to the primary winding during the off-time of the main switch. If V_{LK} is relatively small then the voltage stress on M_{MAIN} will be less, but the time that it takes to ramp the current in the leakage inductance to zero will be increased. During the time that primary current continues to flow after the main switch is turned off, some of the energy stored in the magnetizing inductance of the transformer will flow to C_{CLAMP1}. The longer it takes for the primary current to ramp down to zero, the more energy will be transferred to C_{CLAMP1} and the less that will be deliverable to the load.

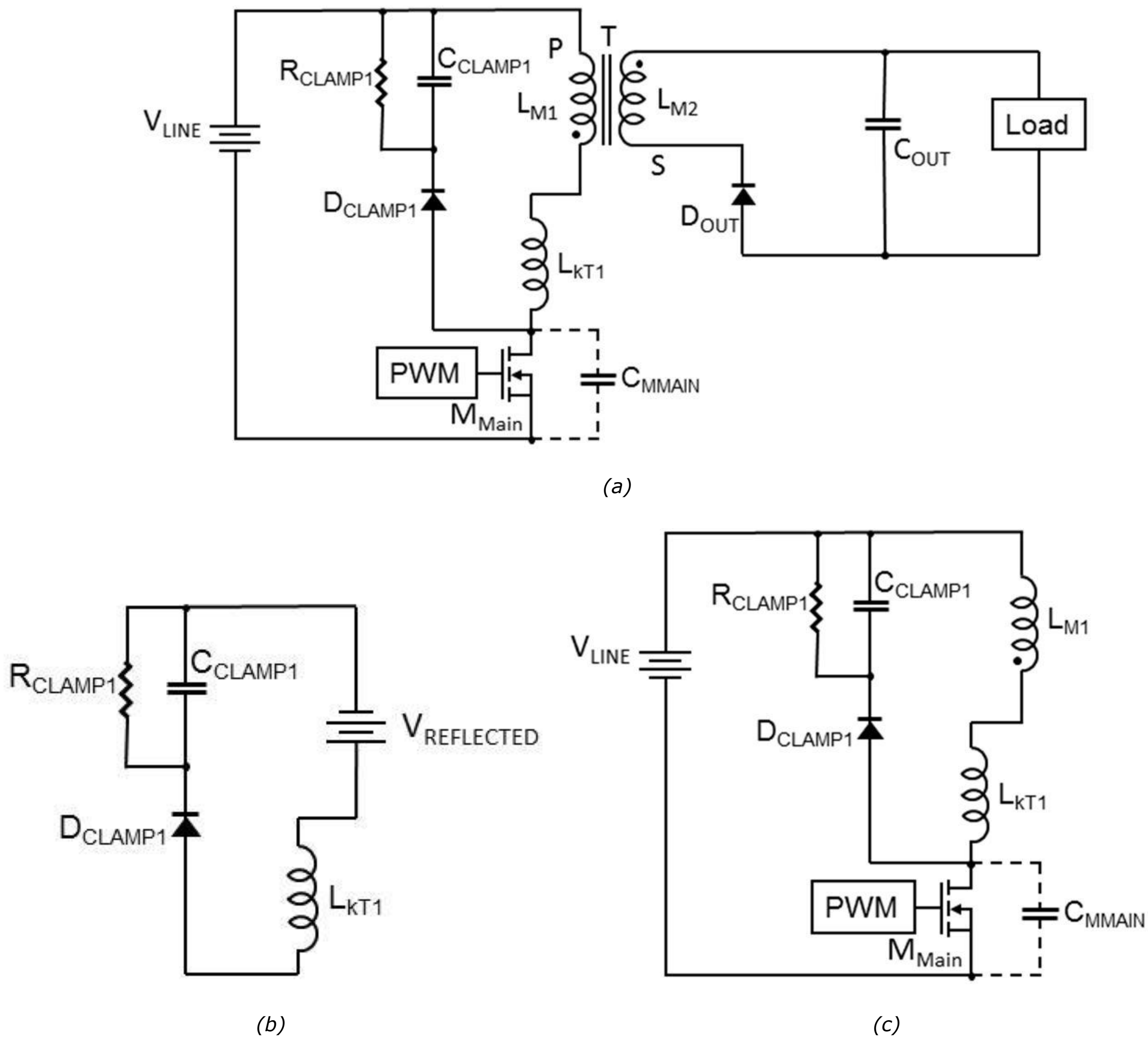

Fig. 3. A flyback converter with a primary dissipative RCD clamp for reducing leakage inductance induced ringing (a). The equivalent primary circuit when D_{CLAMP1} is forward biased (b). The primary circuit of (a) seen as a buck-boost converter with load, R_{CLAMP1}, referenced to positive line voltage (c).

Another way of visualizing energy transfer to C_{CLAMP1} is illustrated in Fig. 3c. The primary circuit with the secondary open is a non-isolated buck-boost converter with a load, R_{CLAMP1}, referenced to the positive line voltage. C_{CLAMP1} serves as the output capacitor for the buck-boost converter. For the isolated flyback converter, the intended load appears in the secondary circuit.

During the time that D_{CLAMP1} conducts, we have, in effect, two converters operating at the same time. One of these converters provides useful power to a desired load in the secondary circuit, while another in the primary circuit only contributes to power losses. In order to reduce the power dissipated in R_{CLAMP1} and reduce the time that it takes to ramp the primary current down to zero, we should operate C_{CLAMP1} at a voltage that is as high as possible without pushing M_{MAIN} to near voltage breakdown during an overload condition.

We could use a higher-voltage switch for M_{MAIN} and operate C_{CLAMP1} at a higher voltage to reduce the overlap time of the two converters. But that may require using a more expensive switch or a switch with higher $R_{DS(ON)}$ and higher M_{MAIN} conduction losses. Typically, small enough transition times, or overlap times, can be accomplished without resorting to a higher-voltage main switch.

Fig. 4a illustrates the drain voltage of a main switch with a clamp designed for minimal clamp power losses and small overlap time. For illustration purposes we will assume a fairly large (exaggerated) leakage inductance, which might correspond well to a real case in which there is no transformer interleaving and a very high isolation voltage requirement. The high clamp-capacitor voltage enables the current to transition quickly from primary circuit to secondary circuit, thereby minimizing overlap time. The cost of the fast transitioning is higher peak drain voltage for the main switch, which may dictate use of a main switch with a higher-voltage breakdown, leading to higher $R_{DS(ON)}$ and higher conduction losses.

Fig. 4b illustrates an alternative that uses a lower-value clamp resistor to lower the peak drain voltage and attempts to avoid using a higher-voltage main switch and its associated higher conduction losses while simultaneously achieving relatively low clamp-circuit losses. The overlap time for the Fig. 4b arrangement is about twice the overlap time for the Fig. 4a arrangement and the clamp resistor losses are considerably larger.

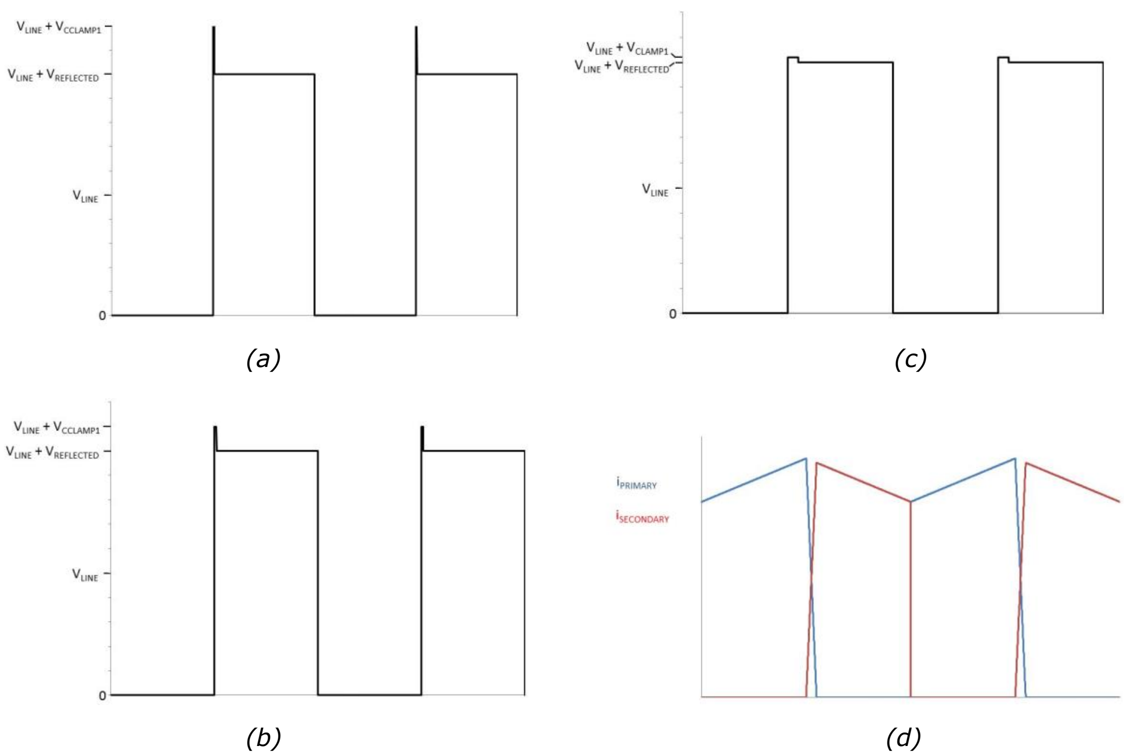

Fig. 4. Part (a) displays the drain voltage waveform obtained with a relatively large-value clamp resistor and relatively low clamp-circuit power losses. The high peak voltage may dictate a higher-voltage main switch. Changing to a typical-value clamp resistor with typical clamp-circuit power losses produces the drain voltage waveform shown in part (b). In this case, it is possible that the peak voltage may require a higher-voltage main switch. A third possibility is to use a relatively small-value clamp resistor with relatively high clamp-circuit power losses, leading to the drain voltage waveform in (c). In this case, it is unlikely that a higher-voltage main switch would be needed to accommodate the peak voltage. Part (d) shows the transformer winding current waveforms corresponding to part (c), illustrating the main switch turn-off transition overlap.

Fig. 4c illustrates an arrangement achievable with a very low-value clamp resistor to achieve a low peak drain voltage on the main switch. This approach avoids a higher-voltage main switch, but the clamp circuit losses may be unacceptable. Fig. 4d illustrates the long overlap time associated with the Fig. 4c arrangement.

At the end of the on-state of the main switch during the overlap time, stored energy in L_{kT1} and magnetizing inductance, L_{M1}, is transferred to the clamp capacitor causing the clamp capacitor voltage to rise. The wave forms in Fig. 4 assume a large-value clamp capacitor so that the voltage slope of C_{CLAMP1} is small. When the energy in L_{kT1} is completely dissipated, the current in L_{kT1} and in D_{CLAMP1} is zero.

At this point in time, current reverses in L_{kT1} and D_{CLAMP1}. The current in D_{CLAMP1} reverses briefly during its reverse recovery. After the reverse recovery, the leakage inductance rings with the parasitic capacitive elements connected to the drain of the main switch, which also includes the intrinsic junction capacitance of D_{CLAMP1}.

The RCD clamp does not completely eliminate leakage inductance-induced ringing. But the amplitude of the ringing that remains is much lower than it would be without the clamp because most of the energy in the leakage inductance has been transferred to the clamp capacitor. The remaining ringing can be mostly eliminated with an RC snubber (series RC network) connected in parallel with the primary winding.

On start-up, the clamp capacitor voltage would continue to rise in voltage and potentially cause a voltage breakdown of the main switch if there were no mechanism to discharge the clamp capacitor. However, there are two mechanisms for discharging the clamp capacitor. The discharge mechanism discussed above is the clamp resistor, R_{CLAMP1}. Another mechanism to be discussed shortly involves the rectifier reverse recovery of D_{CLAMP1}. But first, let's consider the clamp resistor.

The clamp resistor must be selected to accommodate the considerable power that it must dissipate. The clamp capacitor should be selected so that its voltage is relatively invariant over a switching cycle. Some change in capacitor voltage is acceptable, but if the capacitor value is too small, then the voltage excursions may be large enough to damage the main switch due to overvoltage when the clamp capacitor reaches its peak voltage.

For analytical purposes we will assume that the clamp capacitor is large so that its voltage is invariant and we will ignore the forward voltage of the clamp diode and assume that the forward voltage is zero. The voltage applied to L_{kT1} is,

$$V_{LK} = -L_{kT1}\frac{di_{PRI}}{dt}$$

from equation 1 above.

As the primary winding current ramps down during the transition time or overlap time, t_{OL}, the change in primary winding current is the negative of the peak primary winding current at the instant that the main switch turns off:

$$-i_{peak},$$

so that, substituting in equation 1,

$$V_{LK} = -L_{kT1}\frac{-i_{Peak}}{t_{OL}} \qquad (4)$$

and solving for t_{OL} from equation 4,

$$t_{OL} = \frac{L_{kT1} i_{Peak}}{V_{LK}} = \frac{L_{kT1} i_{Peak}}{\left(V_{CCLAMP1} - V_{Reflected}\right)} \quad . \tag{5}$$

During the overlap time, C_{CLAMP1} is charged. The amount of charge flowing into C_{CLAMP1} through D_{CLAMP1} is

$$Q_{CCLAMP1} = \frac{1}{2} i_{Peak} t_{OL} \quad . \tag{6}$$

The average current flowing into C_{CLAMP1}, using equations 5 and 6 is

$$i_{IN} = \frac{Q_{CCLAMP1}}{T_{SW}} = Q_{CCLAMP1} F_{SW} = \frac{1}{2} i_{Peak} t_{OL} F_{SW} = \frac{1}{2} \frac{L_{kT1} {i_{Peak}}^{2} F_{SW}}{\left(V_{CCLAMP1} - V_{Reflected}\right)} \tag{7}$$

where T_{SW} is the switching period and F_{SW} is the switching frequency.

The average current flowing out of C_{CLAMP1} is

$$i_{OUT} = \frac{V_{CCLAMP1}}{R_{CLAMP1}} \quad . \tag{8}$$

In the steady state

$$i_{OUT} = i_{IN} = i_{CLAMP1} = \frac{V_{CCLAMP1}}{R_{CLAMP1}} = \frac{1}{2} \frac{L_{kT1} {i_{Peak}}^{2} F_{SW}}{\left(V_{CCLAMP1} - V_{Reflected}\right)} \tag{9}$$

from equations 7 and 8.

Rearranging equation 9 to place it in quadratic form for solving for $V_{CCLAMP1}$ we get:

$$ {V_{CCLAMP1}}^{2} - V_{Reflected} V_{CCLAMP1} - \frac{1}{2} R_{CLAMP1} L_{kT1} F_{SW} {i_{peak}}^{2} = 0 \quad . \tag{10}$$

Using the quadratic formula and binomial theorem to solve for $V_{CCLAMP1}$ we get

$$V_{CCLAMP1} \approx V_{Reflected} + \frac{R_{CLAMP1} L_{kT1} F_{SW} {i_{peak}}^{2}}{2 V_{Reflected}} \tag{11}$$

From equations 2 and 11 we get, solving for V_{LK},

$$V_{Lk} \approx \frac{R_{CLAMP1} L_{kT1} F_{SW} {i_{peak}}^{2}}{2 V_{Reflected}} \quad . \tag{12}$$

Substituting equation 12 in equation 5 we get

$$t_{OL} \approx \frac{2V_{Reflected}}{R_{CLAMP1}F_{SW}i_{peak}} \quad . \tag{13}$$

The clamp circuit power losses are

$$P_{CLAMP1} = V_{CCLAMP1}i_{CLAMP1} \tag{14}$$

Then, combining equations 2, 3, 5, 7, 12, 13, and 14 we get:

$$P_{CLAMP1} \approx V_{CCLAMP1}\frac{1}{2}i_{peak}t_{OL}F_{SW} \approx \frac{{N_p}^2}{{N_s}^2}\frac{{V_{OUT}}^2}{R_{CLAMP1}} + \frac{1}{2}F_{SW}L_{kT1}{i_{peak}}^2 \quad . \tag{15}$$

From equation 15 we can see how the various circuit elements and operating parameters affect the clamp-circuit power losses. These power losses are increased by increasing switching frequency, increasing leakage inductance, increasing peak winding current, and decreasing clamp resistor value. In other words, to reduce clamp-circuit power losses one would need to increase the value of R_{CLAMP1} or reduce i_{peak}, L_{kT1}, or F_{SW}.

In addition, from equation 15 one would expect that a higher-power converter would have higher clamp-circuit losses and that raising the switching frequency will result in higher clamp-circuit losses thereby imposing a barrier to higher-frequency operation. As switching frequency rises, the clamp capacitor voltage rises and the number of transition overlaps per unit time also rises, so the amount of power lost to leakage inductance effects increases with increasing switching frequency. In Fig. 3a, circuit leakage inductance is one of the major barriers to higher-frequency operation, higher efficiency, and higher power density.

The second mechanism for discharging the clamp capacitor is the rectifier reverse recovery of D_{CLAMP1}. The energy that transfers from the clamp capacitor through the clamp diode to the transformer during reverse recovery is energy that is recovered and contributes to powering the load. The first (to my knowledge) publication about use of the clamp diode as a discharge mechanism can be credited to Power Integrations.

Wherever possible, it makes sense to increase the charge that flows during reverse recovery, which can be done by selecting a clamp diode that is as slow as possible. However, a clamp diode that is still reverse recovering when the main switch turns on is too slow and can potentially cause cross conduction and failure of the main switch and/or clamp diode. So, reverse-recovery time must be less than the off-time of the main switch at minimum line voltage in overload.

The best choice for a clamp diode is usually a fast recovery or standard recovery junction rectifier, rather than an ultrafast junction rectifier or schottky. Other considerations for clamp diode selection are thermal considerations. During reverse recovery, the magnitude of the diode reverse voltage is larger than the diode's forward voltage.

Note that more energy is dissipated in a slow recovery rectifier compared to an ultrafast rectifier so that a rectifier that can handle the additional power losses must be selected. However, the use of the slower clamp diode will reduce the clamp capacitor voltage and enable the designer to increase the value of the clamp resistor, R_{CLAMP1}. The higher value for R_{CLAMP1}, in turn, reduces power losses in this resistor, reduces the overlap time, and improves overall efficiency.

When clamp diode reverse recovery is complete, the primary winding is unclamped and leakage inductance will now ring with the parasitic capacitances connected to the drain of the main switch. The RCD clamp does not eliminate all primary circuit ringing, but it does reduce the energy that would be available to drive primary circuit ringing in the absence of the RCD clamp. The ringing that occurs after reverse recovery of the clamp diode can be eliminated with the use of an RC snubber, as illustrated in Fig. 5.

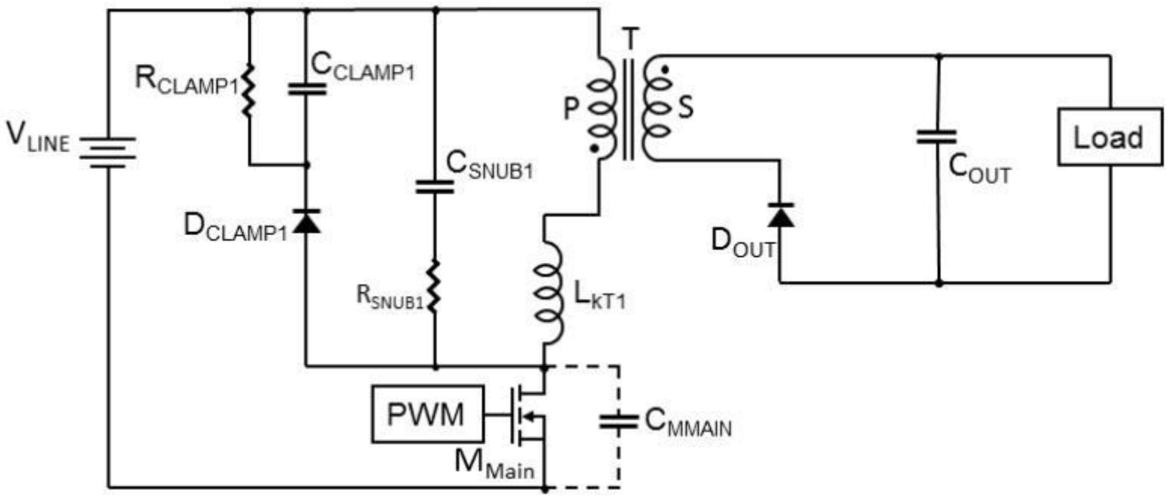

Fig. 5. A flyback converter with primary-side RCD clamp and RC snubber.

Fig. 5 illustrates the flyback circuit with both RCD clamp and RC snubber on the primary side. The RCD clamp eliminates most of the energy that was stored in the leakage inductance at the instant that the main switch turned off. The RC snubber damps the ringing that would otherwise occur after the reverse recovery of D_{CLAMP1}. You would like the ringing to be critically damped.

A technique for determining snubber component values intended to achieve close to critical damping (contributed by Bruce Carsten[1] and Phil Todd [2]) is as follows:

1. Measure the ringing frequency f_0, at the drain terminal of the main switch during the time interval when both clamp diode and main switch are off.

2. We want to find a source inductance value for the ringing that we have measured. If we can find a source capacitance value we can calculate the source inductance value from the formula for the LC resonant frequency that we measured in step 1.

We start by setting the snubber resistor value to 0 Ω. If we place a capacitor, C_0, whose value is three times the source capacitance value at the snubber capacitor position, then the total capacitance contributing to ringing will be four times the source capacitance value and the ringing frequency will be half the original ringing frequency, according to the LC resonance formula. The snubber capacitor with 0 Ω for a snubber resistor is in parallel with the source capacitance.

Since we don't know a value for C_0 we can find it by trial and error by starting with a very small value capacitor for C_0 at the snubber capacitor position and increasing the value of C_0 until the ringing frequency is half the original ringing frequency.

The leakage inductance will be a major contributor to the source inductance of the ringing but there may be other small contributors. Therefore, we should not assume that the source inductance and leakage inductance are equal. But we can use the value of leakage inductance, if we have measured or calculated leakage, to calculate a value of capacitance that should be in the magnitude range of the source capacitance.

If we do that calculation, we can use the value of capacitance calculated as an initial capacitance value for a trial and error exercise to find a value for C_0 that results in one half the original ringing frequency..

3. Calculate the source inductance using the formula:

$$L_0 = \frac{3}{\left(4\pi^2 {f_0}^2 C_0\right)}$$

where C_0 is the value of snubber capacitor that yields half the initial ringing frequency.

4. The snubber resistor value is then

$$R_{SNUBBER} = \sqrt{{3L_0}/{C_0}} \quad .$$

The snubber capacitor value should be chosen to be equal to or larger than C_0.

5. The RC snubber is dissipative. The power dissipated by the snubber is given by

$$P_{SNUBBER} = C_{SNUBBER} \Delta V_{CSNUBBER}{}^2 f_{SWITCHING} \quad .$$

From this formula you can see that the power loss of the snubber depends on the value of the snubber capacitor, but not on the value of snubber resistor. In the RC snubber, power loss is due to charging and discharging of the snubber capacitor. The snubber resistor determines how fast the charge/discharge happens but not the energy transferred into or out of the snubber capacitor.

A flyback circuit complete with all the necessary clamps and snubbers to effectively eliminate leakage inductance-induced ringing is shown in Fig. 6.

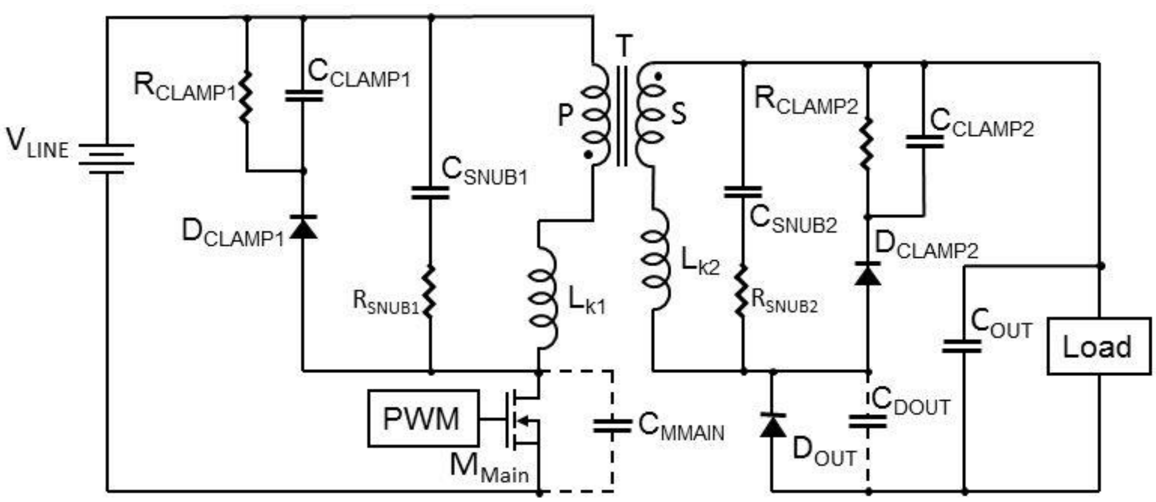

Fig. 6. A CCM flyback converter with primary and secondary clamps and snubbers.

The design of the clamp and snubber for the secondary circuit is the same as for the primary circuit except that the primary ringing occurs during the off-time of the main primary switch and the secondary ringing occurs during the off-time of the main secondary rectifier. The effect of leakage inductance is to delay the rise in current in the primary winding and maintain current flow in the secondary circuit during the turn-on transition of the main switch (and turn-off transition of D_{OUT}). The addition of clamps and snubbers can be used to effectively eliminate leakage inductance-induced ringing, but the additional clamps and snubbers are dissipative and contribute to power losses.

The existence of leakage inductance creates ringing in a typical flyback converter, which may contribute to a regulatory emissions failure. The typical remedies shown above can be effective at eliminating leakage inductance-induced ringing to bring the circuit into compliance for noise emissions, but these remedies come at the cost of significant power losses in dissipative clamps and snubbers that compromise efficiency and limit switching frequency range, thereby creating an upper bound to switching frequency and practically achievable power density.

Clearly, for a typical commercial flyback converter leakage inductance is a considerable foe. Unfortunately, leakage inductance is a foe in almost all isolated power converters, but there are some exceptions, which will be described in part 3.

A Non-Dissipative Clamp

A single-ended forward converter has a primary circuit much like a flyback converter with similar problems related to leakage inductance. For the primary circuit, an RCD clamp and an RC snubber will likely fix the EMI problem. For leakage inductance-induced ringing the situation is worse in a way in the single-ended forward converter because the primary winding is unclamped during the off-time of the main switch, but the secondary winding is unclamped during both the on-state and the off-state of the main switch. In some cases, three RCD clamps plus two RC snubbers may be necessary to eliminate leakage inductance-induced ringing to achieve EMC.

Fig.7 illustrates a non-dissipative LCD clamp (invented by Moshe Domb[3]) that can be used in either a single-ended forward or flyback converter to reduce clamp power losses. During the on-time of the main switch the primary winding is clamped, so no leakage inductance-induced ringing will occur. Also, during the on-time of the main switch, stored energy increases in the leakage inductance referred to the primary winding and energy increases in the magnetizing inductance of the transformer.

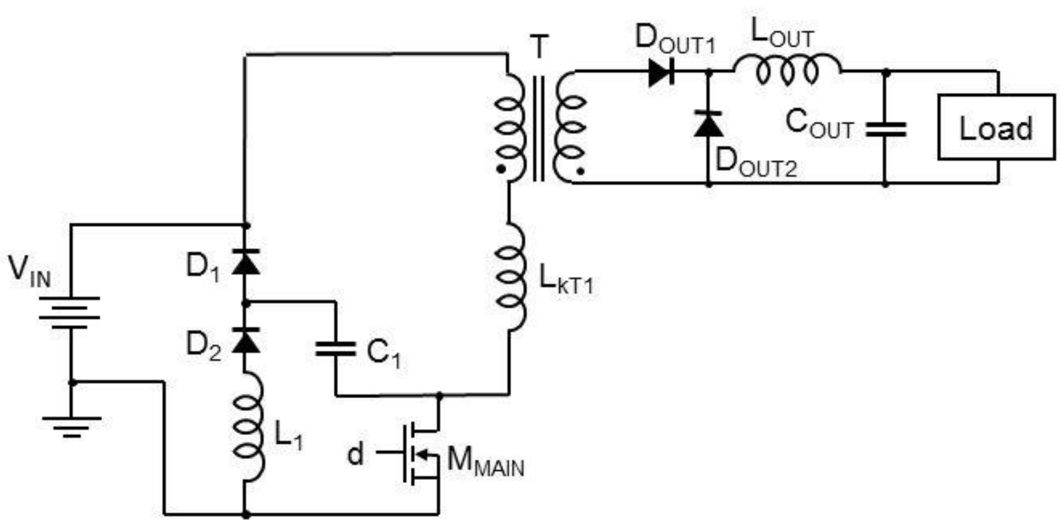

Fig. 7. A single-ended forward converter with a non-dissipative LCD clamp.

In a forward converter the stored energy in the magnetizing inductance is typically small. When the main switch turns off, the drain voltage of the main switch rises and the voltage at the anode of D_1 rises until diode D_1 becomes forward biased. D_1 becomes forward biased at a drain voltage much less than the line voltage since the upper plate of C_1 is initially positive with respect to the lower plate of C_1 during the turn-off transition of the main switch. The early turn-on of D_1 reduces turn-off switching losses, directing the primary winding current into C_1 and away from the channel of M_{MAIN}.

The clamp capacitor, C_1, discharges, reverses polarity, and charges up during the turn-off transition of the main switch until the current in L_{kT1} drops to zero. C_1 is typically a relatively small-value capacitor compared to the value of clamp capacitor used in an RCD clamp. When the leakage inductance energy has dropped to zero, D_1 turns off, the leakage inductance rings with the D_1 junction capacitance, the intrinsic output capacitance of the main switch, and other capacitive parasitics connected to the drain of the main switch. At the same time, the inductor L_1 rings with the junction capacitance of D_2. An RC snubber for the primary winding may be necessary to damp ringing after D_1 becomes reverse biased. A second RC snubber in parallel with L_1 may also be necessary.

When the main switch turns on, D_2 becomes forward biased and the stored energy in C_1 is transferred to L_1 as the C_1 voltage drops to zero. Then the stored energy in L_1 is transferred back to C_1, reversing the voltage applied to C_1. If the line voltage is near the low end of its range, diode D_1 may become forward biased briefly before the current in L_1 drops to zero. When D_2 becomes reverse biased, L_1 will ring with the junction capacitance of D_2. An RC snubber in parallel with L_1 may be necessary to damp this ringing.

In the LCD clamp, there are no dissipative elements and the leakage inductance energy is recirculated, but dissipative RC snubbers are often required in the primary. Also, the LCD clamp does not eliminate the problems related to leakage inductance-induced ringing in the secondary.

Nevertheless, there is another advantage of the LCD clamp in a forward converter as it allows the duty cycle range to be extended to values greater than 50% without an active reset circuit.[4] The LCD clamp also reduces turn-off switching losses of the main switch, which could be a significant benefit at low line voltages. The LCD clamp provides efficiency benefits for the primary circuit but does not eliminate all leakage inductance-induced ringing and it requires an additional inductor to achieve those benefits.

Active Clamps

An active clamp is the most effective method for eliminating leakage inductance-induced ringing. An active clamp circuit can eliminate power losses related to leakage inductance and enable extension of the duty cycle range in forward converters. It can also reduce the main switch peak voltage thereby potentially enabling the use of a lower-voltage main switch with lower conduction losses. An active clamp implementation of a flyback converter is shown in Fig. 8. The cost of an active clamp is an active switch to replace a clamp diode and, in many cases, a drive circuit for the active switch.

In the Fig. 8 circuit there are two active-clamp circuits that fully clamp all the leakage inductance all the time, eliminating the need for dissipative circuit elements. Both dissipative clamps and dissipative snubbers are obviated.

The circuit requires two active-clamp switches, but no clamp diodes. The two clamp diodes are replaced by two active switches. For lower-power circuits, EMC may be achievable without the secondary active clamp. When the main switch turns off, the drain voltage of the main switch rises until the body diode of the p-channel M_{CLAMP1} becomes forward biased. At this time, M_{CLAMP1} turns on.

Not shown, but assumed, are the gate networks that accomplish break-before-make switching so that M_{CLAMP1} and M_{MAIN} are never on simultaneously and cross conduction is avoided. M_{CLAMP1} remains on until the end of the switching cycle and is turned off just prior to the beginning of the on-state of M_{MAIN}. The undotted terminal of the primary winding is always connected to line voltage, which is an ac ground.

During the on-state of M_{MAIN}, the dotted terminal of the primary winding is connected through M_{MAIN} to ground. The primary winding cannot ring during the on-state of M_{MAIN}. During the off-state of M_{MAIN} the dotted terminal of the primary winding is connected through M_{CLAMP1} to C_{CLAMP1}, which is also an ac ground, so that during the off-state of the main switch the primary winding cannot ring. The primary winding is fully clamped and cannot ring at any time. Also, the clamp capacitor voltage is less than the clamp capacitor of a typical RCD clamp, which may enable the use of lower-voltage switches with lower $R_{DS(ON)}$ and lower conduction losses.

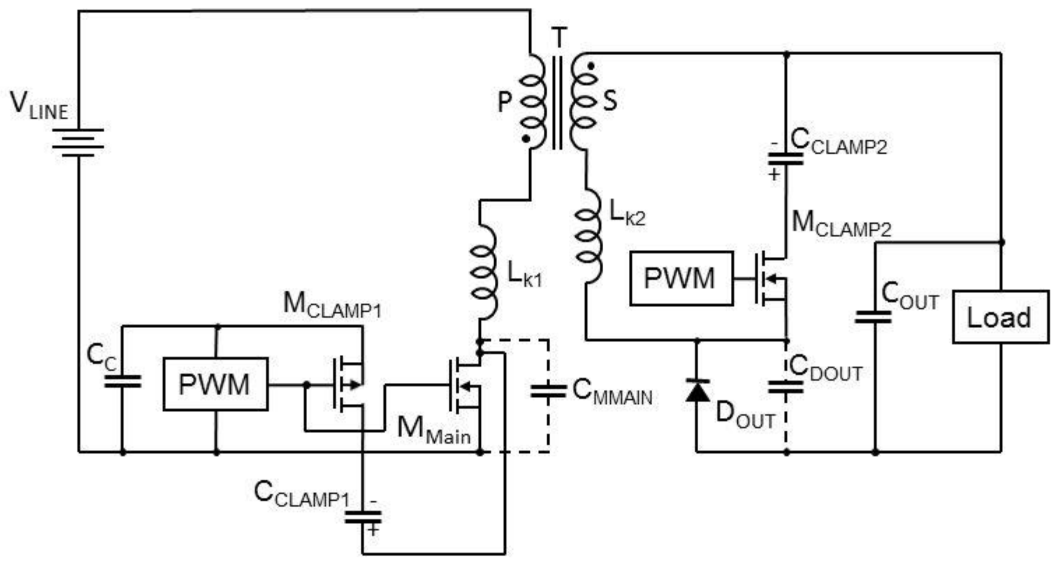

Fig. 8. A flyback converter with two active clamps.

The second active clamp in the Fig. 8 circuit appears in the secondary. The dotted terminal of the secondary winding is always connected to C_{OUT}, which is an ac ground, and the undotted terminal of the secondary winding is either connected through D_{OUT} to ground or through M_{CLAMP2} to C_{CLAMP2}, which is an ac ground. M_{CLAMP2} is off when D_{OUT} is on and M_{CLAMP2} is on when D_{OUT} is off. There is no time (except during the brief switching transitions) when the undotted terminal of the secondary winding is not connected to an ac ground. Both terminals of the secondary winding are connected to ac grounds in both operating states so the secondary winding is fully clamped and cannot ring.

With two active clamps the flyback converter is fully clamped and cannot ring. It uses no dissipative clamps or snubbers to completely eliminate leakage inductance-induced ringing. The two active-clamp solution is, by far, the best solution for both EMC and efficiency. There are additional advantages to the active-clamp flyback converter, which will be discussed in part 3 of this article series.

The down side of the two active-clamp approach is that it requires two active switches and often requires one or more high-side gate-drive circuits. Fig. 8 illustrates a primary-side active clamp circuit that uses a p-channel MOSFET for the active-clamp switch. With the p-channel switch, as shown, no level-shifting gate-drive mechanism is required. In situations where a p-channel switch is not workable or available, then an n-channel high-side switch with a level-shifting gate driver is required.

Another hurdle for fully clamping an isolated flyback converter is that the secondary active-clamp switch must be driven with a signal that is synchronized with the main switch and the timing signal for driving the gate of M_{CLAMP2} must also be isolated. In many cases, including some circuits now in high-volume production, the secondary-side active clamp may not be needed to achieve EMC.

In general, fully clamping every winding in a transformer is a way to completely and efficiently eliminate leakage inductance-induced ringing in any topology. But that job is more or less difficult than the flyback converter implementation depending on the topology and the number and location of active clamps needed to accomplish full clamping of every winding. In a single-ended forward converter with a single primary winding and a single secondary winding, three active clamps are required to accomplish full clamping, although in many cases EMC can be achieved with fewer than three active clamps.

The number of clamps and snubbers needed to eliminate leakage inductance-induced ringing varies. Some topologies require very little circuitry to eliminate this ringing while other topologies require a lot of circuitry. There is one relatively simple and widely applicable topology with a single primary winding and a single secondary winding in which full clamping can be accomplished with only one active clamp network on the primary side. And the topology requires no secondary clamps or snubbers, or a critical isolated switch timing circuit. If a p-channel clamp switch can be used, then no additional gate-drive ICs are required. This topology, which is depicted in Fig. 9, is the active-clamp coupled-boost converter.

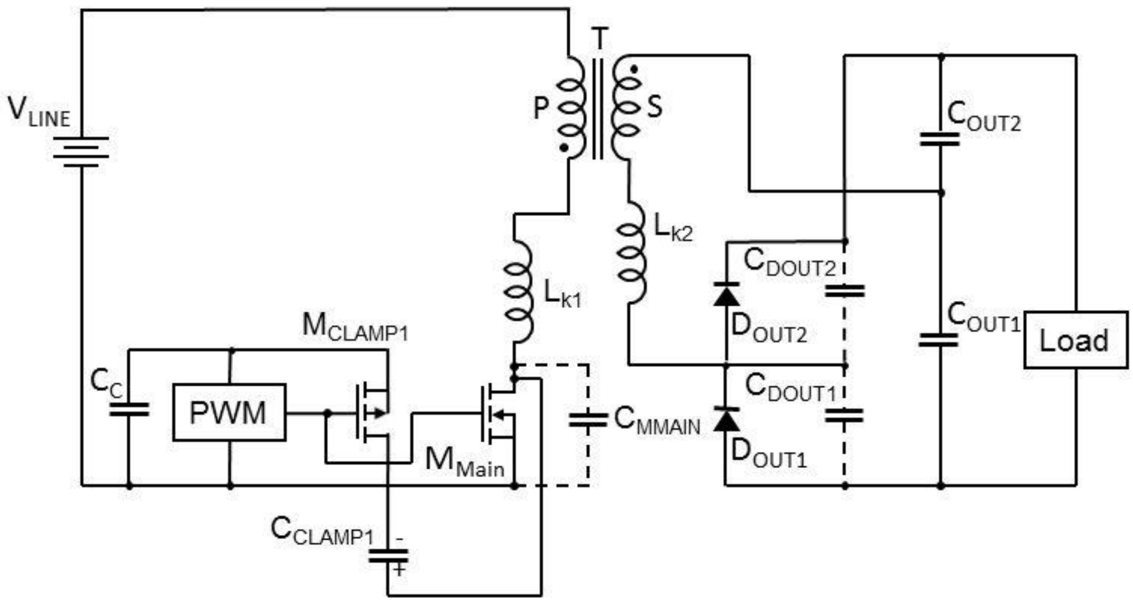

Fig. 9. An active-clamp coupled-boost converter. Only a single primary-side active clamp is required to eliminate all leakage inductance-induced ringing.

In the Fig. 9 circuit, the primary-side circuit structure is identical to the active-clamp flyback converter of Fig. 8 and the primary winding is fully clamped. The dotted terminal of the secondary winding is connected to the C_{OUT1} and C_{OUT2} capacitors at an ac ground. The undotted terminal of the secondary winding is alternately connected through rectifiers to secondary ground or the positive terminal of C_{OUT2}, another ac ground. Except for the brief switching transitions that occur at roughly the same time (slightly delayed) as the switching transitions in the primary circuit, the secondary winding is clamped. The secondary winding is fully clamped and experiences no leakage inductance-induced ringing.

The active-clamp coupled-boost converter requires only one active clamp to eliminate all leakage inductance-induced ringing and it has additional advantages over the active-clamp flyback converter. One of the advantages is that it has better transformer utilization since the transformer delivers energy to the output while the main switch is off. Although the flyback converter, does that too, the coupled-boost also delivers energy to the output while the main switch is on.

In addition, the number of secondary turns required is much smaller. The average secondary winding voltage for the coupled-boost converter is one half the load voltage. In contrast, the average secondary winding voltage for the flyback converter is dependent on the line voltage range, but always greater than the load voltage.

Switch conduction losses can be either more or less in the active-clamp coupled boost converter compared to the flyback converter. Worst-case primary-switch conduction losses are typically less in the boost converter but the two secondary rectifiers are connected in series so that, for low-voltage loads, the secondary-switch conduction losses may be larger in the coupled boost than in the flyback converter.

On the other hand, another advantage of the boost is that the maximum voltage stress of the secondary rectifiers is equal to the load voltage. Therefore, if synchronous rectifiers are used, lower-voltage, lower $R_{DS(ON)}$ switches can be used and the secondary-switch conduction losses for the boost will likely be less than the flyback with a synchronous rectifier.

The flyback converter excels for applications with very high line range (> ~7:1) because of its second-order transfer function. Even with diode rectifiers the secondary-switch losses may be as low or lower for the coupled boost because you may be able to use two low-voltage schottky rectifiers in the boost when a junction rectifier or high-voltage schottky is required for the flyback. The active-clamp coupled boost excels for fixed high-voltage applications and improves on the efficiency of a flyback for synchronous rectifier applications, except those with wide line range or very low load voltage.

Active clamps are a very effective way to fully clamp a transformer or coupled inductor, eliminate leakage inductance-induced ringing, eliminate clamp and snubber losses, reduce switch voltage stress and conduction losses, and increase efficiency. Additional information about fully clamped converters can be obtained by contacting the author (see "About The Author" below.)

Leakage Inductance And Load Regulation

As mentioned briefly above, one of the problems created by leakage inductance is the poor load regulation of unregulated outputs of a multi-output circuit. Leakage inductance appears as a frequency-dependent lossless resistor. There are a few ways to deal with the problem.

Obviously one can add a post regulator to the unregulated output. As discussed in part 1 of this article series, leakage inductance is a property of a winding pair. The problem of poor load regulation of unregulated outputs is more often seen in isolated circuits because of isolation voltage requirements that require space (insulation) between primary and secondary windings and because of EMC or safety requirements that may dictate the need for shields between primary and secondary windings to block common mode (CM) noise or provide a safety ground.

Leakage inductance between primary winding and secondary windings may be difficult to avoid and interleaving primary and secondary windings to reduce leakage inductance may be prohibitive in terms of cost and space. The leakage inductance between primary and secondary windings usually has little to do with poor regulation of the main output if the voltage control loop is closed around the main output. But the leakage inductance between different secondary windings has a lot more effect on the load regulation of unregulated outputs than primary-to-secondary pair leakage inductance, if at least one of the outputs is regulated with a voltage feedback loop.

One may be able to improve the load regulation dramatically for little added cost or space by interleaving secondary windings. Interleaving secondary windings reduces leakage inductance in the secondary winding pairs thereby improving load regulation. If the secondary windings share a common ground then leakage inductance can be reduced and load regulation improved by autotransformer winding the secondary windings. Another method that may provide some benefit mixes feedback signals from multiple outputs thereby improving load regulation of the previously unregulated output at the cost of poorer regulation of the output that previously provided all of the voltage feedback.

Summary

In this second part of the article series, we've discussed ways in which leakage inductance is our foe and creates design problems. These problems include EMI, increased voltage stress on switches and increased switch-conduction losses resulting from leakage inductance induced-ringing, power losses from clamps and/or snubbers used to reduce EMI caused by leakage inductance, and poorer load regulation. We've also discussed some ways to deal with these problems. One of the most-effective methods covered here is the use of active clamps to fully clamp the transformer. This approach eliminates leakage inductance-induced noise without incurring additional power losses.

The third and final part of this article series will discuss ways in which leakage inductance is our friend. In some designs, leakage inductance provides benefits in terms of converter size, cost, and efficiency that could not be accomplished easily (or at all) without leakage inductance.

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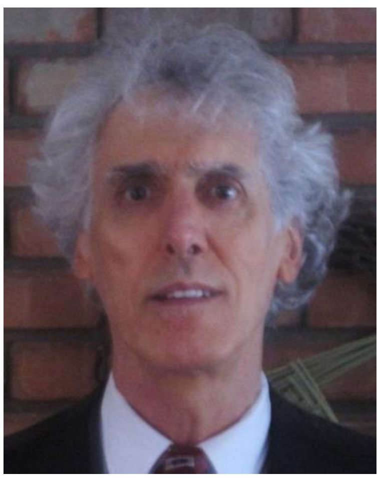
About The Author



Ernie Wittenbreder is widely recognized in the industry from his seminars, publications, and many patents. His areas of expertise include topology selection, design tradeoff decision making, high-efficiency design, soft switching, gate drivers, and design for EMC. His applications experience spans a broad range including high reliability and extremely low-noise converters for military and aerospace, telecommunications, solid-state lighting, very high density power supply in package (PSIP) converters, and a wide variety of commercial and industrial applications.

Wittenbreder has developed a power converter computer simulation product and SiC gate driver IC. He also has a great deal of litigation experience in patent infringement lawsuits. Wittenbreder began his career as a physics professor and holds MS Physics and MSEE degrees and BS, math and physics. He can be reached via this email.

For further reading on leakage inductance, see the How2Power Design Guide and enter "leakage inductance" in the keyword search. And for more on magnetics design in general, see the Design Guide, locate the "Design Area" category, and click on the "Magnetics" link.