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How to deal with Leakage Elements in FLYBACK Converters

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It is difficult to find a good tradeoff between cost and performance when designing offline flyback power supplies. The basic problem is the presence of the various leakage inductances of the transformer. Those leakage inductances cause voltage overshoot (which, in turn, calls for power switches with increased voltage rating), divert a portion of the primary current to a clamp, and degrade the voltage regulation in power supplies with primary regulation or with multiple outputs.

To evaluate the effect of the leakage inductances on the circuit performance in the design phase is a challenging task. Here a simulation program can be quite helpful. This paper reviews the magnetic and electric models of the two–winding and three–winding transformers, discusses how to extract the inductance values of the models by measurements and calculations, and how to feed a SPICE model with. Furthermore, the paper analyses the influence of the primary leakage inductance on the switch voltage waveform, provides guidelines on how to protect the switch from the voltage overshoot, and considers the effect of the secondary leakage inductances on the cross regulation of multiple–output flyback power supplies.

The Reluctance Concept

Since the transformer, or coupled inductances for a FLYBACK, is the key player in our setbacks, it is important to spend some time to understand the way the inductive leakage elements are spread to finally measure them and feed our simulator model with.

Several ways exist to depict how the magnetic variables are organized within the transformer. The reluctance method keeps the analogy with our electric circuits and therefore appears easier to us non–magnetic experts. Remember that the electrical resistance of a metallic piece tends to brake the current flow. That is to say, we impose a current variation dI (Amps) through the resistance R (Ohms) of the metallic part by applying a voltage variation dV (Volts) across it, or $dI = \dfrac{dV}{R}$. With magnetic materials, we could compare the current with the flux ϕ (Weber) which is constrained to circulate in the magnetic material by an external force F, the *magnetomotive force* (mmf) expressed in Amps–turns. This force is elaborated by a winding of N turns in which a current I flows. The resulting force F is equal to N x I. The obstacle which brakes the circulation of the flux is the *reluctance* $\Re$ of the magnetic medium and can be explained by: $\Re = \dfrac{dF}{d\phi} = N \bullet \dfrac{di}{d\phi}$. That is to say, how much of magnetomotive force F shall I apply to the magnetic material in order to obtain a change in the flux ϕ. The analogy between electricity and magnetism is depicted by **Figure 1**.
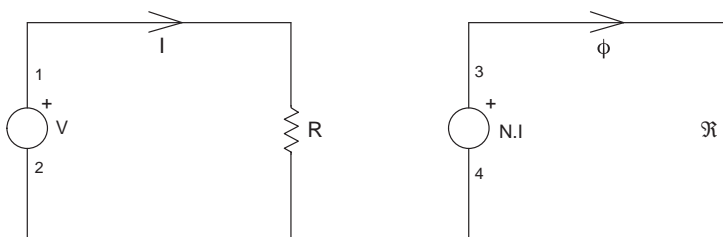

Figure 1. Analogy Between Electricity and Magnetism

To physically define $\Re$, let's start again with the definition of the resistance R of a metallic bar: $R = \dfrac{l_m}{\sigma \cdot A_e}$ with: l_m length of the material, σ the material conductivity and A_e the material area (also noted S for section). Applying the same analogy as before and replacing the electrical conductivity σ by μ, the *magnectic permeability*, we define the reluctance by: $\Re = \dfrac{l_m}{\mu \cdot A_e}$ with $\mu = \mu_o \cdot \mu_R$, the remaining symbols being the same as above. As one can see, the reluctance depends on the magnetic medium by its *relative* permeability μ_R. In the air, like in a gap or if flux lines were closing externally to the magnetic material (e.g. a leaky choke), μ_R drops to 1 and the reluctance simply becomes: $\Re_{air} = \dfrac{l_m}{\mu_o \cdot A_e}$, with μ_o the air magnetic permeability, $4 \cdot \pi \cdot 10^{-7}$ henry.meter in the International System (SI).

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Inductance and Permeance

Faraday's law of electromagnetic induction states that the voltage v appearing across an inductor L in which a current I circulates can be calculated by: $v = L \cdot \frac{di}{dt}$. This law can be expressed in a different way when applied to a single coil made of N turns of wire: $v = N \cdot \frac{d\phi}{dt}$. By equating both equations, we define the *self–inductance* of a single coil by $L = N \cdot \frac{d\phi}{di}$. We already defined the term $\frac{d\phi}{di}$ when we talked about reluctance: $\frac{d\phi}{di} = \frac{N}{\Re}$. By introducing this last equation in the definition of the self–inductance L, we come to: $L = \frac{N^2}{\Re}$ or by defining the *permeance* P as the reciprocal of reluctance $(P = \frac{1}{\Re})$, the inductance calculation simplifies to: $L = N^2 \cdot P$. We will see later on how powerful this concept is to easily make inductance appear.

The Two–winding Transformer Magnetic Model

Figure 2 shows a transformer wound across a square magnetic material. The primary side made of Np turns creates the necessary force F which gives birth to two components: ϕm who links both windings, but also ϕl1 which does not couple to the secondary and corresponds to a leakage path through the air. This leakage path could be responsible for ElectroMagnetic Interferences (EMI). Thanks to ϕm, a current Io circulates in the secondary, but this current also gives birth to another leakage flux ϕl2 whose polarity is opposite of that of ϕm. It is important to note that ϕm produces Io while ϕl2 is a consequence of it.
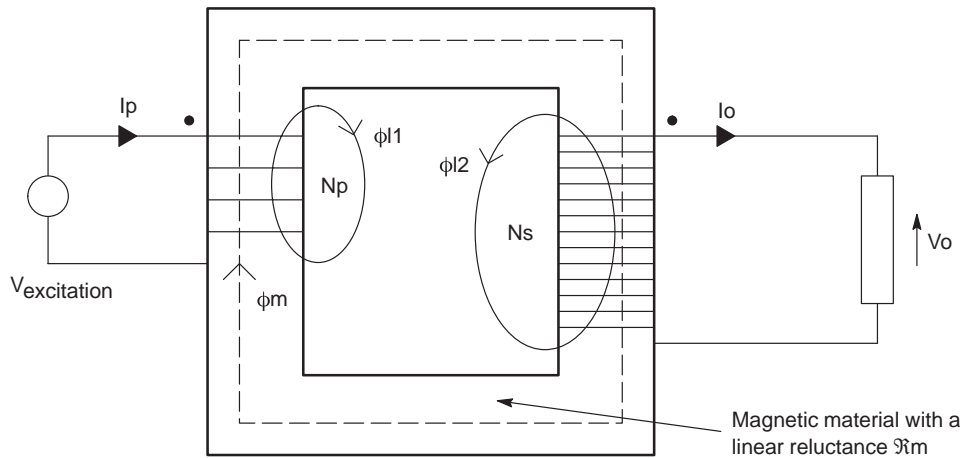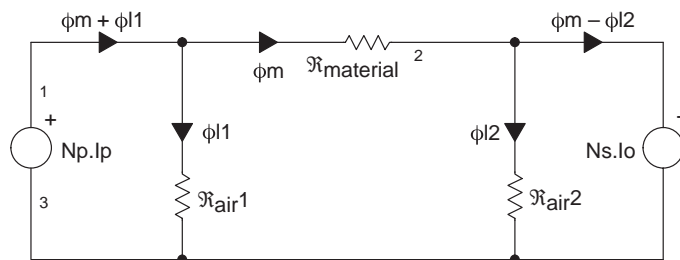

Figure 2.
A Simple Two–winding Transformer with its Associated Fluxes

Applying the technique depicted in Figure 1, we are now able to make a reluctance model immediately highlighting the way the various fluxes combine, as **Figure 2b** portraits.

$$\phi l1 = \frac{Np \cdot Ip}{\Re_{air}1}, \phi l2 = \frac{Ns \cdot Io}{\Re_{air}2}, \phi m = \frac{Np \cdot Ip - Ns \cdot Io}{\Re_{material}}$$

Figure 2b. The Reluctance Model of Figure 2's Transformer

The fluxes definitions are given beneath the drawing. You calculate the fluxes (ϕ) the same way as if you were evaluation a current by solving for a voltage (N.I) across a resistance ($\Re$). Now let's apply duality rules by which a series reluctance becomes a parallel permeance (and vice versa) and a voltage source turns into a current source. The result appears in **Figure 2c** where the current source is reversed to account for the flux circulation (ϕm–ϕl2).
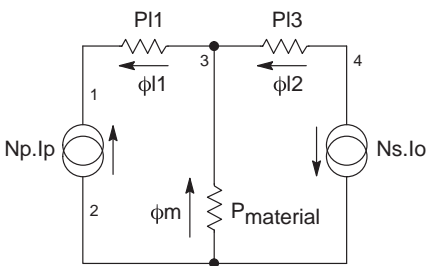

Figure 2c. Permeance Version of the Two–winding Transformer

The voltage induced across the left arm is $\phi m+\phi l1$, while the (output) voltage generated at the right arm is $\phi m-\phi l2$.

The goal of the next stages is to unveil the inductances. To do so, we will scale everything according to Np, our winding reference. If a current source N.I naturally forces a current flow into a resistance R, the voltage across R is obviously (N.I)xR. If we scale the current source by 1/N, we must multiply the resistance by N: V = Ix(N.R). As a consequence, the input source becomes Ip and every permeance is multiplied by Np. But we still need to keep the terms including Io (ϕm and $\phi l2$) constant to their previous values once the Np multiplication is done: $\phi l2$ = Ns.Io.Pl2 or, after multiplying Pl2 by N: $\phi l2 = \dfrac{Ns}{Np} \cdot Io \cdot (Pl2 \cdot Np)$, which is equivalent

to the previous equation. It is now pertinent to make the turn ratio appear, since $\dfrac{Ns}{Np} \cdot Io$ is the output current reflected to the primary. **Figure 2d** details this next step.
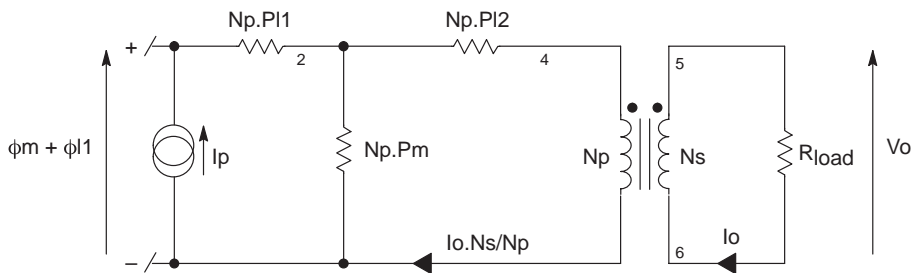

Figure 2d. First Stage Toward the Final Inductance Model

Each permeance in the network is once again scaled by the reference winding Np to come up with a full inductive network (remember $L = N^2 \cdot P$). According to Faraday's law, scaling ($\phi m+\phi l1$) by Np has the dimension of a voltage (Vp), since V.dt = N.dϕ.

As we previously stated, $\phi l2$ is the consequence of Io, the secondary current. Since this leakage inductance Ll2 "belongs" to the secondary, we keep it at the primary thanks to the reflection ratio $\dfrac{Np^2}{Ns^2}$. **Figure 2e** unveils the final drawing.
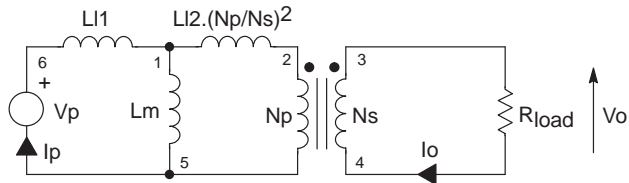

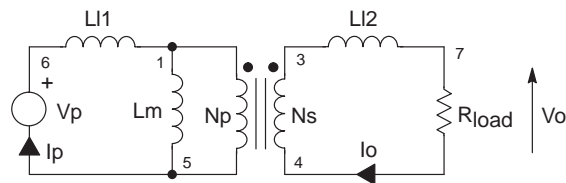**Figure 2e. The Final Electrical Model of the Two–winding Transformer**

Figure 2f. A More Practical Electrical Model

This model, called the π–model or the T–model, was introduced many years ago. Reference [1] and [2] offer a comprehensive description of the different methods used to model the magnetics in general.

Figure 2e configuration is not easy to implement in a simulator. Reflecting the secondary leakage back to its place facilitates the handling of this model, as **Figure 2f** depicts.

Determining the Physical Values of the Two–winding π Model

You often read: "short the secondary and measure the leakage inductance on the primary". This statement depends on the model you have adopted for simulation [3]. For instance, in figure 2f's model, shorting the secondary does not give you the final primary leakage value … Watch for the steps:

1. Inject a sinusoidal voltage Vp on the primary and measure the open–circuit voltage on the secondary Vs.

   Compute $N = \dfrac{Np}{Ns} = \dfrac{Vp}{Vs}$

2. Measure the primary inductance, the secondary being open. This gives you Lps_{open}

3. Repeat step 2 but the secondary is now shorted by a wire. You obtain Lps_{short}.

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4. Compute the coupling coefficient k with: $k = \sqrt{1 - \dfrac{Lps_{short}}{Lps_{open}}}$

5. Compute Ll1 with: $Ll1 = (1 - k) \cdot Lps_{open}$

6. Compute Ll2 with: $Ll2 = (1 - k) \cdot Lps_{open} \cdot \dfrac{1}{N^2}$

7. Compute Lm with: $Lm = k \cdot Lps_{open}$

8. Measure with an ohm–meter the primary and secondary DC resistances, respectively Rp and Rs.

Finally, your simulation could be implemented using **Figure 2g**'s schematic. It uses a perfect transformer, as described in [4], where the leakage and resistive elements have been added.
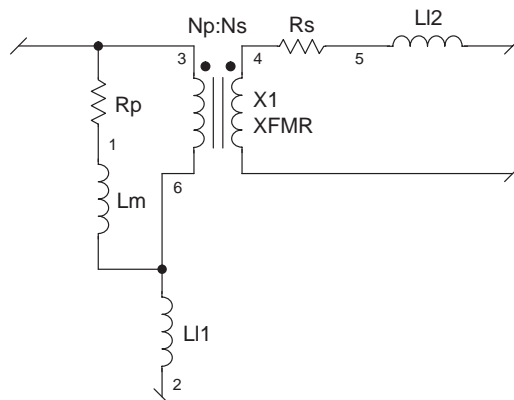
Figure 2g. The Complete Two–winding Transformer SPICE Model

The Three– winding Transformer

A three–winding transformer is often implemented in either dual–output SMPS or primary regulated supplies. As you imagine, the numerous fluxes coming into play greatly complicate the determination of the model. **Figure 3** shows the construction of the transformer and highlights the position of the fluxes.
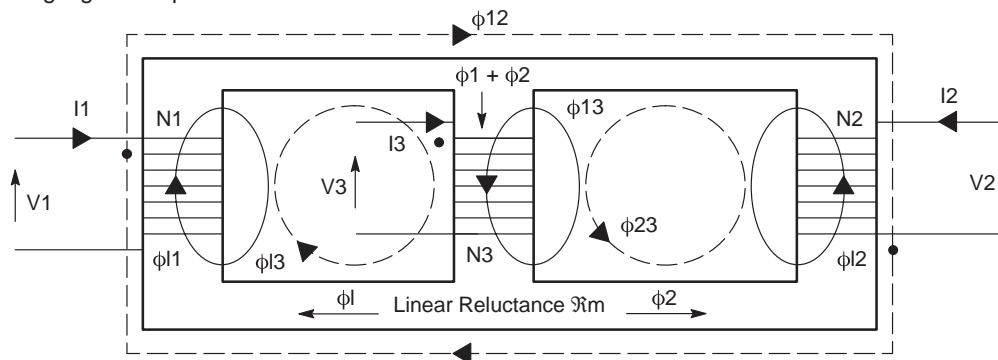

Figure 3. The Three–winding Transformer Complicates the Flux Graph

Figure 3 portraits the three leakage fluxes, $\phi l1$, $\phi l2$ and $\phi l3$ but also the linking fluxes between the three windings: $\phi12$, $\phi13$ and $\phi23$. As you can see on **Figure 3b**, the reluctance model differs in complexity from his **Figure 2** counterpart. For the sake of clarity, we will not go through the transformation steps as we did before but rather directly jump to the results given in [1].
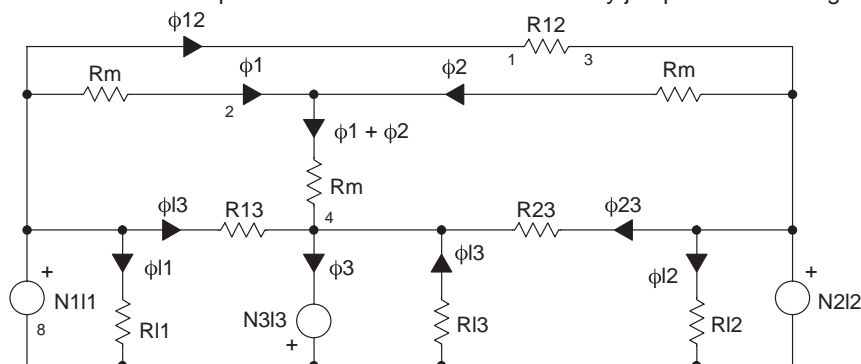

Figure 3b. Figure 3's Reluctance Model Becomes Difficult to Manipulate

These results are gathered upon **Figure 3c** drawing where the three leakage elements appear in series with each winding. The surprise comes from the primary leakage element which depends upon the leakage air path P1 but also from the permeance P23 between both secondary windings. As a matter of fact, if you improve the coupling between both secondaries (e.g. by twisting the wires) you _increase_ the primary leakage inductance. Reference [1] paper also demonstrated how the leakage elements stay practically independent of the air–gap length: the coupling coefficient diminishes as the air–gap increases (the magnetizing inductance becomes smaller) but leakage elements stay constant.
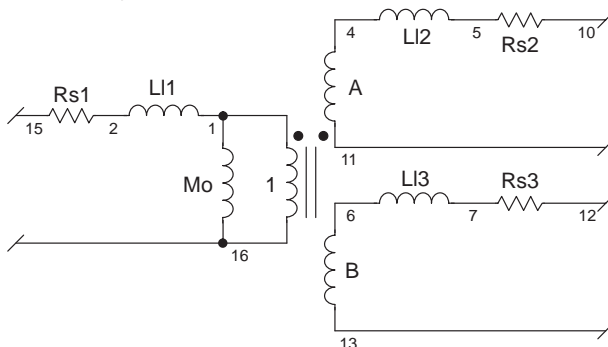
Figure 3c. Final Equivalent Model Showing Every Leakage Elements

Leakage elements definitions:
$Ll1 \approx n_1^2.(P1 + P23)$ Primary Winding
$Ll2 \approx n_2^2.(P2 + P13)$ Power Winding
$Ll3 \approx n_3^2.(P3 + P12)$ Auxiliary Winding
$Mo \approx n_1^2.(Pm)$

Determining the Physical Values of the Three–winding π–model

These measurements require a RLC bridge or a network analyzer operating at a frequency high enough to neglect the winding resistance (if you measure an impedance) but low enough to minimize the interwinding capacitances. In our numerical application, we selected 100 kHz which is the operating frequency of our converter. The network analyzer gives you a complete impedance–phase plot and ensures that you stay on the inductive portion during the measurement. With a standard LRC–meter, if changes in the operating frequency do not lead to big variations of the values you read, then you are well on the inductive portion.

The 4 W transformer used in our experiment was made by Thomson Television Components (Gray, France) on an small E core with a special insulation material [5]. As you will discover through the following steps, the results express how you combine the various leakage elements from the measurement data.

1. Inject a sinusoidal voltage Vp on the primary and measure the open–circuit voltages on the secondaries Vs_{power} and $Vs_{auxiliary}$.

   Compute $A = \dfrac{Vs_{power}}{Vp}$ and $B = \dfrac{Vs_{auxiliary}}{Vp}$. Measures gave A = 0.0817 and B = 0.156

2. Measure the inductance L1 seen from the Primary, the secondaries open: $L1 = Ll1 + Mo = 3.62$ mH

3. Measure the inductance L2 seen from Primary with the Power Winding open, the Auxiliary shorted:

   $$L2 = Ll1 + \frac{Mo \cdot \frac{Ll3}{B^2}}{Mo + \frac{Ll3}{B^2}} = 199 \ \mu H$$

4. Measure the inductance L3 seen from Primary with the Power Winding shorted, the Auxiliary open:

   $$L3 = Ll1 + \frac{Mo \cdot \frac{Ll2}{A^2}}{Mo + \frac{Ll2}{A^2}} = 127 \ \mu H$$

5. Measure the inductance L4 seen from the Power Winding, the Auxiliary shorted and the Primary open:

   $$L4 = Ll2 + A^2 \cdot \left[\frac{Mo \cdot \frac{Ll3}{B^2}}{Mo + \frac{Ll3}{B^2}} \right] = 1.405 \ \mu H$$

Please note that L1–L4 could also be impedances Z1–Z4. You should then divide each value by $2 \cdot \pi \cdot F_{measure}$ to obtain the inductance value. We now have a system of four equations with four unknowns. Feeding a math processor with these equations give the solutions in a snap–shot:

$$Ll1 = \sqrt{\left(L3 - L1 - \frac{L4}{A^2}\right) \cdot (L3 - L1)} + L1 = 58.5 \ \mu H$$

$$Ll2 = \frac{A^2 \cdot (Ll1 - L1) \cdot (L3 - Ll1)}{L3 - L1} = 466 \ nH$$

$$Ll3 = \frac{B^2 \cdot (Ll1 - L1) \cdot (L2 - Ll1)}{L2 - L1} = 3.558 \ \mu H$$

$Mo = L1 - Ll1 = 3.56$ mH

Series resistances are measured with a 4–wire ohm–meter and included in the Spice model.

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Some Precisions on the π–models

If you go through the equations that lead to the first two–winding transformer, you come–up with a set of three equations for four elements in the model. We say the model is *underdetermined* and we obviously need to give an arbitrary value to the primary/secondary turn ratios. It is however convenient to pick–up the real ratio to make the model close to its physical counterpart.

At the opposite, the three–winding model is unique. It does not need any arbitrary parameters to reveal the values of the elements its made of. The π–model cannot be used for transformers with more than three windings.

Three–winding SPICE Model

The SPICE of the above transformer mixes a pair of two–winding transformers as reference [4] details. For clarity reasons, we keep the leakage elements external to this ideal structure.

```
.SUBCKT XFMR–AUX 1 2 3 4 10 11 {RATIO_POW=??? RATIO_AUX=???}
*
* RATIO_POW = A
* RATIO_AUX = B
*
```
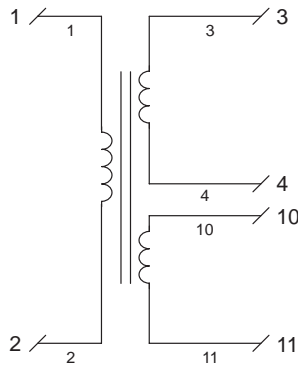


```
* 1:A
* 1:B
RP 1 2 1MEG
E1 5 4 1 2 {RATIO_POW}
F1 1 2 VM1 {RATIO_POW}
RS1 6 51 1U
VM1 5 6
E2 20 11 2 1 {RATIO_AUX}
F2 2 1 VM2 {RATIO_AUX}
RS2 21 61 1U
VM2 20 21
.ENDS
```

Positive ratios for A and B describe pin 3 and 10 as the positive outputs (e.g. for a FORWARD) by respect to 1. By entering negative values for A and B, you simulate separate or both windings in FLYBACK mode.

Floating nodes, as encountered in isolated supplies, often generate matrix errors in SPICE simulators. To avoid this pitfall, simply add a 100 Meg resistor between nodes 4 and 11 and the ground. INTUSOFT's IsSpice4 (San–Pedro, CA) elegantly implements .OPTIONS RSHUNT=100Meg which places a shunt element between each node and ground, thus easing the DC convergence.

The Switching Effect of the Leakage Inductance in FLYBACK Structures

As we can see upon figure 3d's sketch, the primary leakage element LI1 does not participate to the primary/secondary energy transfer. When the main switch SW1 closes, the primary current builds–up until the adequate threshold is reached and resets the internal latch (current mode supply). The current flows through Mo but also through LI1. As a matter of fact, when SW1 opens, both magnetic fields collapse and the voltage across the inductances reverses in an attempt to keep the amps–turn constant: Mo energy is thus coupled to the secondary and gives birth to the output current charge. Since LI1 cannot find a circulating path, you observe a large positive spike whose amplitude depends on the various parasitic elements the transformer is made of. If we stick

to **Figure 4** references, the peak voltage can be expressed by: $V_{peak} = Ip \cdot \sqrt{\dfrac{LI1}{C_{primary} + C_{oss}}} + V_{DC} + V_{FLYBACK}$, the

square root representing the characteristic impedance of the parasitic LC network. This network is also guilty of the oscillations which ring after the spike.

With Figure 4's numerical values, we come–up to a theoretical drain voltage of 1514 V (Vf_{DOUT}=800 mV) at the opening, confirmed by an IsSpice4 simulation.
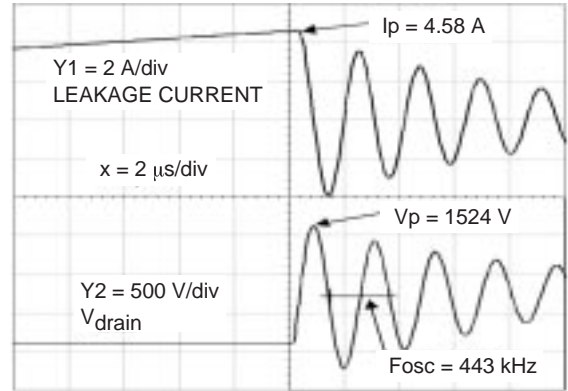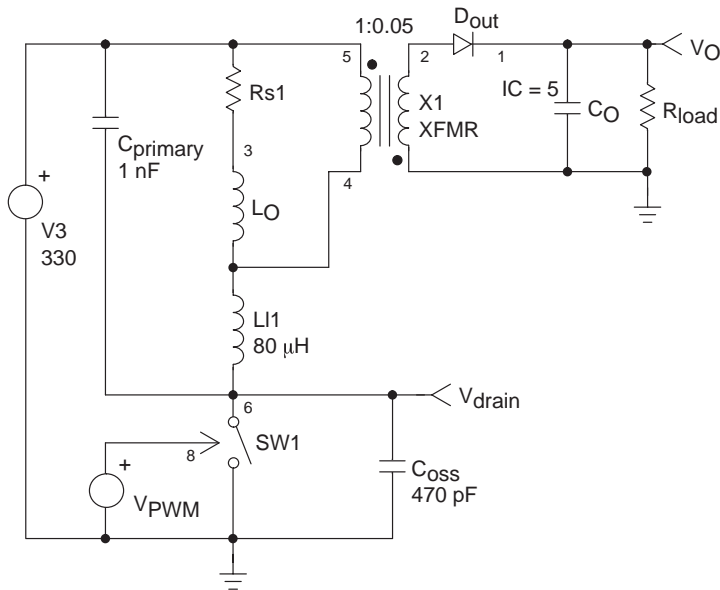
Figure 4. Simulating the Effects of the Primary Leakage Inductance LI1

If we have a lateral MOSFET in place of SW1, e.g. with a MC33363 or MC33370 recently introduced by MOTOROLA (Phoenix, AZ), the internal body–diode is weak and cannot accept any avalanche energy: the transistor is destroyed and a mean has to be found to protect it against this trouble.

Now, let's assume we implement a switching component which withstands an accidental avalanche thanks to its internal body–diode. It could be an MTD1N60E or even the new MMG05N60D IGBT, specifically tailored for low–power offline supplies. In that case, the leakage spike grows–up until the avalanche is reached, thus clipping the voltage (**Figure 4b**).
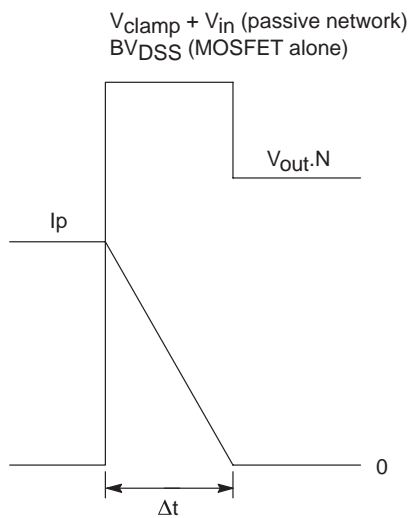
Figure 4b.
Turn–off Waveforms with a MOSFET Clipping the Peak
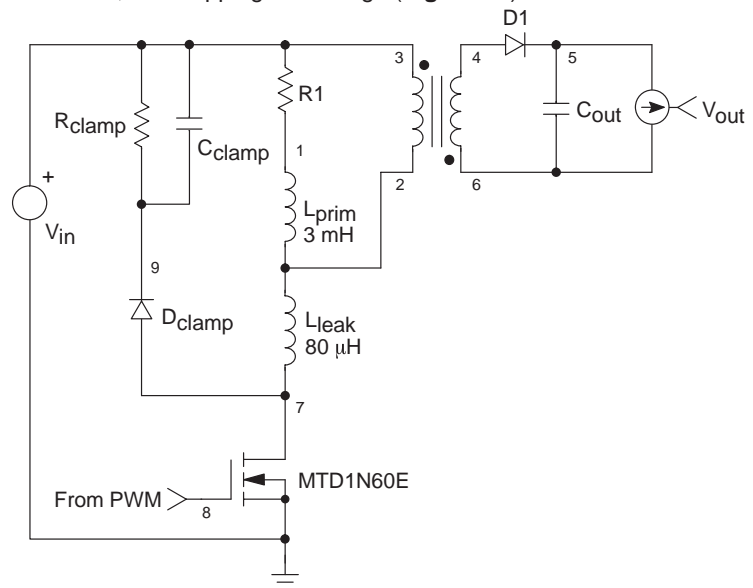
Figure 5a.
An RC Network Clamps the Drain Voltage at Turn–off

In lack of any external clipping network, we can calculate the amount of energy E_T dissipated in the transistor every time it opens. As we said, the leakage inductance tries to keep the current circulating at its level (Ip, when the transistor opens) during Δt and pushes the drain voltage up to BV_{DSS}. Ip(t) can be expressed by: $Ip(t) = Ip \cdot \dfrac{\Delta t - t}{\Delta t}$. You calculate the energy by integrating over time the cross–over area between current and voltage: $E_T = \displaystyle\int_0^{\Delta t} Id(t) \cdot V_{DS(t)} \cdot dt = \dfrac{1}{2} \cdot Ip \cdot BV_{DSS} \cdot \Delta t$. The time Δt is given by the *reset–voltage* applied across the leakage inductance: $BV_{DSS} - V_{out} \cdot N$, neglecting the output diode forward voltage. It is then easy to define $\Delta t = \dfrac{L_{leak} \cdot Ip}{BV_{DSS} - V_{out} \cdot N}$. By introducing this equation into the previous one, we get:

$P_T = \dfrac{1}{2} \cdot Ip^2 \cdot L_{leak} \cdot F \cdot \dfrac{BV_{DSS}}{BV_{DSS} - V_{out} \cdot N}$. This results depicts the average power the transistor would be the seat of if no mean were implemented to re–route the energy spike elsewhere. We shall then implement a clipping network to prevent the MOSFET from entering the avalanche.

Preventing the avalanche

By wiring a RC network from drain to V_{CC} via an ultra–fast diode (e.g. a MUR1100E), we will prevent the drain voltage to rise above $V_{in} + V_{clamp}$, the clipping voltage we want to impose (**figure 5a**). Let's us first consider that the voltage across the RC network is constant (ripple is low compared to average voltage) and equals V_{clamp}. The power dissipated at turn–off in the clamp is: $P_{clamp} = \dfrac{1}{2} \cdot V_{clamp} \cdot Ip \cdot \Delta t \cdot F$. Δt has already been defined, but this time, BV_{DSS} is replaced by V_{clamp}. Once introduced in the previous equation, we obtain the power dissipated in the clipping network:

$P_{clamp} = \dfrac{1}{2} \cdot Ip^2 \cdot L_{leak} \cdot F \cdot \dfrac{V_{clamp}}{V_{clamp} - V_{out} \cdot N}$. Since this power will mainly be dissipated by the resistor R_{clamp}, we can write the following equality: $\dfrac{V_{clamp}^2}{R_{clamp}} = \dfrac{1}{2} \cdot Ip^2 \cdot L_{leak} \cdot F \cdot \dfrac{V_{clamp}}{V_{clamp} - V_{out} \cdot N}$. By solving for R_{clamp} we calculate its value for a given level of clamping voltage: $R_{clamp} = \dfrac{2 \cdot V_{clamp} \cdot (V_{clamp} - V_{out} \cdot N)}{L_{leak} \cdot Ip^2 \cdot F}$.

It is important to minimize the ripple level V_{ripple} superimposed on V_{clamp}. A capacitor C_{clamp} will fulfill this function. If we agree that the amount of charges Q will equally split between R_{clamp} and C_{clamp} at turn–off we can write: $V_{ripple} \cdot C_{clamp} = I_{Rclamp} \cdot T$. Knowing that $I_{Rclamp} = V_{clamp} / R_{clamp}$ and then solving for C_{clamp} it comes:

$C_{clamp} = \dfrac{V_{clamp}}{V_{ripple} \cdot F \cdot R_{clamp}}$

Using a Zener Diode

A zener diode can be used to replace the RC network when the clamping level needs to be more precisely defined. However, the cost associated with this active solution is higher than the passive one. The electrical schematic is as follows:
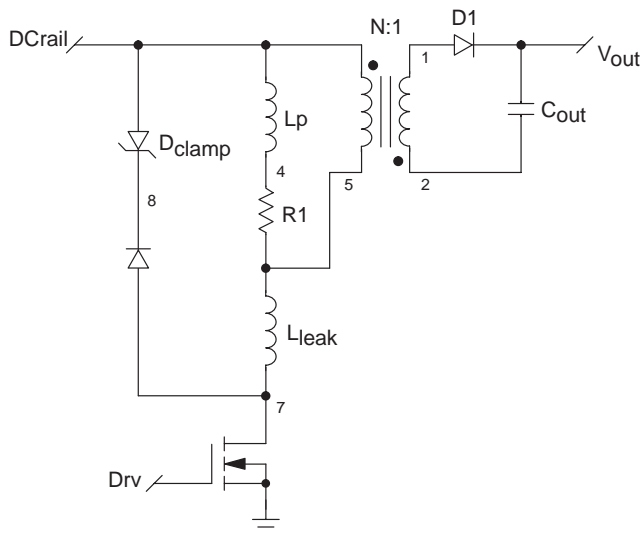

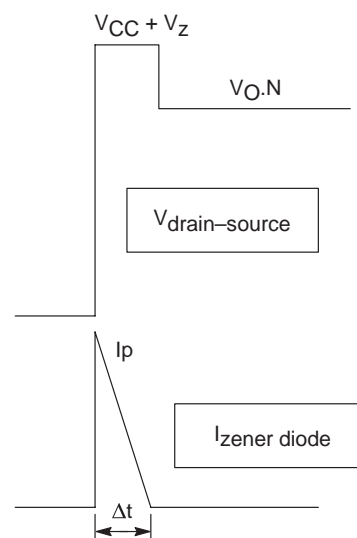**Figure 5b. Clipping with a Zener Defines a More Precise Protection Level**

Figure 5c. Clipping Waveforms with a Zener Diode

As previously written, the power dissipated at turn–off in the transistor can be expressed by:

$P_T = \frac{1}{2} \cdot Ip^2 \cdot L_{leak} \cdot F \cdot \dfrac{V_{CC} + V_z}{V_z - V_{out} \cdot N}$. The average power in the zener diode can easily be found from **Figure 5c's** simplified

drawing. After integration, it is found that $E_{zener} = \frac{1}{2} V_z \cdot Ip \cdot \Delta t$, where Δt is imposed by the reset–voltage across L_{leak}:

$\Delta t = \dfrac{L_{leak} \cdot Ip}{V_z - V_{out} \cdot N}$. If we introduce this last equation into the previous one, the power dissipated in the zener diode is:

$P_z = \dfrac{V_z \cdot Ip^2 \cdot L_{leak} \cdot F}{2 \cdot (V_z - V_o \cdot N)}$

When the Leakage Inductance Brakes the Energy Transfer...

When figure 5b's transistor is closed, input energy from the bulk capacitor is stored into the core's gap. The total involved energy is equal to $E_{stored} = \frac{1}{2} \cdot Ip^2 \cdot Lp$. When the transistor opens, the voltage across the primary inductance reverses and the primary/ secondary energy transfer should immediately take place. However, the primary current also circulated in the leakage inductance and causes an additional reversal voltage which sums–up to the previous one: the clipping diode is forward biased and routes some current to the clamp network, transforming the leakage energy into heat. Unfortunately, this leakage path causes some of the energy stored in the gap to be also transformed into heat: any current flowing in the leakage inductance forces the same current to flow through the mutual inductance. As a matter of fact, the leakage inductance delays the primary to secondary energy transfer (by Δt) until its current has dropped to zero. This parasitic inductance also diverts a substantial amount of the stored energy thus hurting the open–loop gain. To account for it, the primary current will be higher, forcing you to design a larger core transformer. You can diminish the above Δt either by decreasing L_{leak} or by increasing the reset voltage upon it, $(V_{clamp} - V_{out}.N)$. **Figure 6** shows the behavior of the various currents at the opening and details the timing involved with a 1:1 transformer.
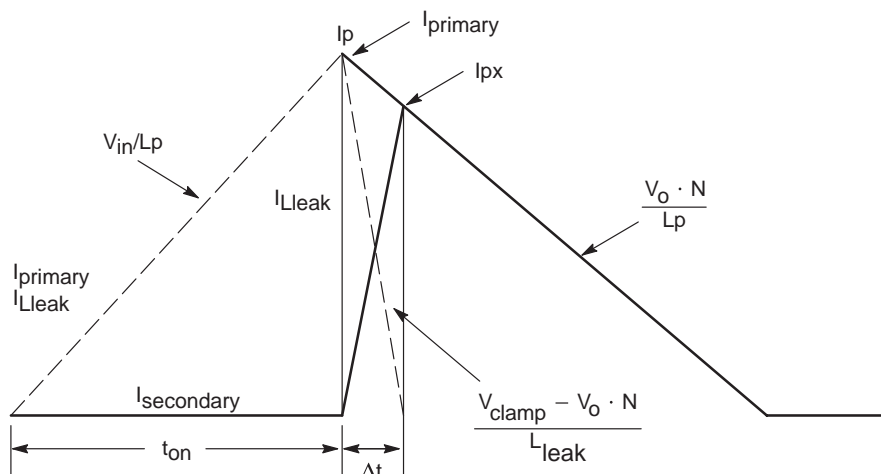

Figure 6. The Primary Leakage Effect Delays the Energy Transfer by Δt

To calculate the percentage of Ip who flows into the clamp, you can calculate the point at which $I_{secondary}$ catches–up with $I_{primary}$.

This point is called Ipx on Figure 6. As we previously said, $\Delta t = \dfrac{L_{leak} \cdot Ip}{V_{clamp} - V_{out} \cdot N}$. The slope of the decreasing primary current

is simply $\dfrac{N \cdot V_{out}}{Lp}$, but this equation can also be written as: $\dfrac{Ip - Ipx}{\Delta t} = \dfrac{N \cdot V_{out}}{Lp}$. Replacing Δt and solving for Ipx gives:

$\dfrac{Ipx}{Ip} = 1 - \dfrac{L_{leak}}{Lp \cdot \left(\dfrac{V_{clamp}}{V_o \cdot N} - 1\right)}$. This last equation gives you the effective percentage of primary current stolen by the leakage

inductance. Reference [6] details an interesting study of the primary leakage effects in a dual–transistor FLYBACK converter.

The Secondary Inductance Degrades the Cross–regulation

When you design primary regulated power supplies, a good cross–regulation is needed to avoid the output to run out of specs. Unfortunately, the leakage elements will, once again, play in your disfavor. **Figure 7** explains how the secondary leakage inductance, lumped in series with the power winding, generates a parasitic voltage which subtracts from the power voltage. This voltage simply equals: $L_{lumped} \cdot \dfrac{dI_{secondary}}{\Delta t}$.
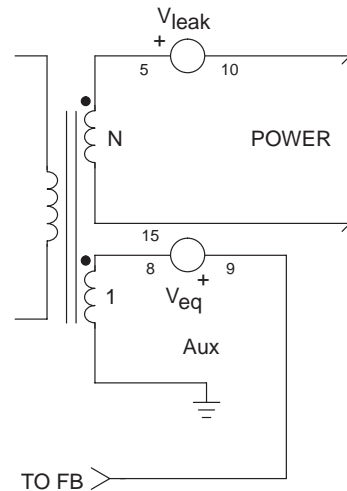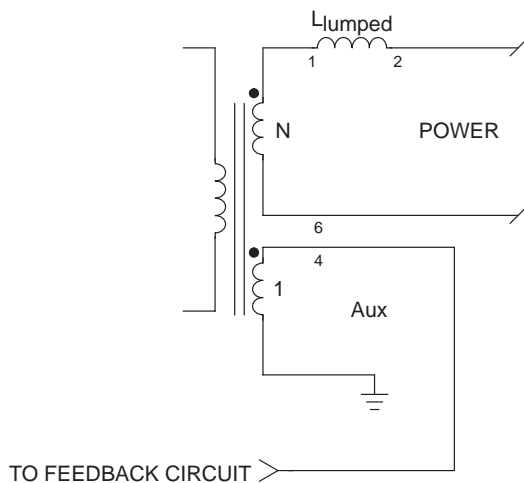
Figure 7. The Parasitic Voltage Subtracts from the Power...   **Figure 7b ...but Adds to the Auxiliary Level**

To calculate how this parasitic voltage V_{leak} reflects to the auxiliary winding, we can state that $V_{out} = V_{aux} \cdot N - V_{leak}$ should equal $(V_{aux} + V_{eq}) \cdot N$. When solved, it comes that $V_{eq} = -\dfrac{1}{N} \cdot V_{leak}$. **Figure 7b** properly represents how this source now reflected to the secondary adds to the auxiliary winding and creates the well known cross–regulation error: the feedback senses a higher auxiliary winding, due to V_{eq}, and decreases the duty–cycle forcing V_{out} to diminish.

Several methods exist to counteract the effects of the secondary leakage inductance. The simplest one consists in adding a resistance in series with the auxiliary winding. This resistance will be the seat of a voltage whose polarity fights V_{leak} and cancels the previous effect. However, the results are load dependent and does never give an optimum regulation.

In primary regulated supplies, the primary current at the opening is almost entirely diverted into the main secondary output because the auxiliary circuit only requires a negligible portion of it. However, in real multi–output FLYBACK converters, the split depends on the current imposed by the converter loads. If the sum of all these output currents leads to the well known triangular waveform, individual shapes can be quite strange in discontinuous mode when a load changes its nominal value. These phenomena are described in details by Lloyd DIXON in reference [7].

Putting the Model to Work

Thanks to ideal components, a Spice simulator can add the parasitic elements and shows how they interact with each other. **Figure 8** presents a complete offline AC/DC wall adapter implemented around the recent MC33363. The corresponding INTU-SOFT's IsSpice4 and MICROSIM's PSpice models are available for download at http://motserv.indirect.com/home2/models/bin/smps_circuits.html.
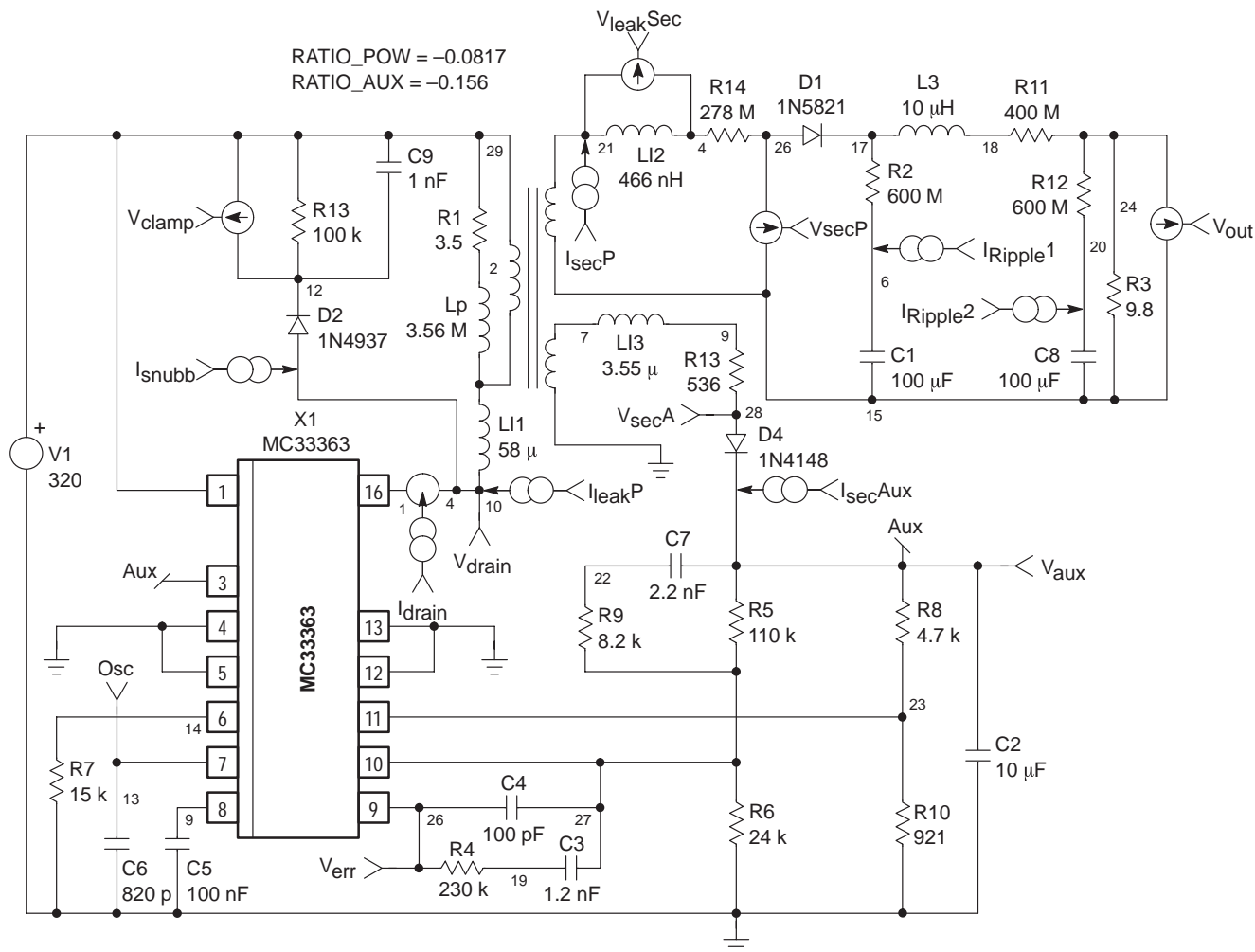
Figure 8. IsSpice4 Simulations Help to Understand How the Phenomena Take Place

In this schematic, the aforementioned parasitic inductive elements have been incorporated and simulated. **Figure 8b** and **8c** respectively detail the auxiliary/power voltages observed on the real PCB and the simulated ones.
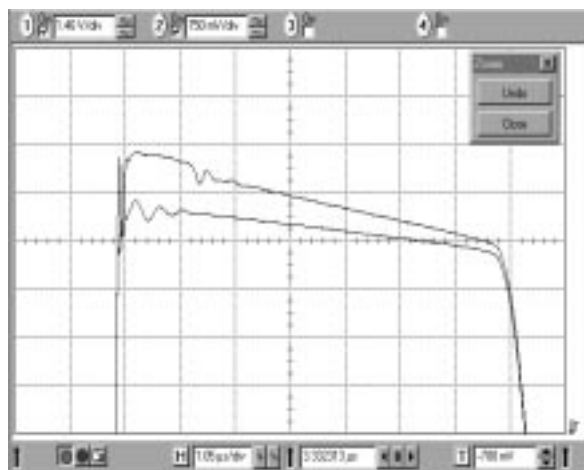

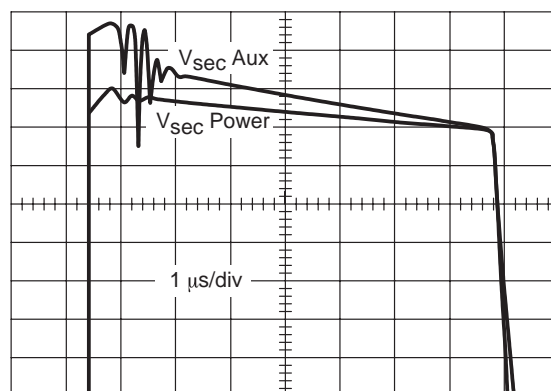**Figure 8b. Measured Auxiliary and Power Secondary Voltages**

Figure 8c. Simulated Secondary Waveforms

The good agreement between the figures, corroborated by the output voltage measurements, confirms the validity of the model and the measurement method. It also highlights the good secondary coupling coefficient of the transformer we used. However, a refined model should include the numerous capacitive leakage elements but also the core material model.

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Conclusion

When designing Switch–Mode Power Supplies, the understanding of how the various leakage elements interact within the system is the key to designing reliable and rugged systems. Thanks to easy measurement methods, the evaluation of these parasitic components becomes simple. With the help of a Spice simulator and adequate models, the effect can then immediately be visualized and properly compensated.

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