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Improving the Performance of Traditional Flyback-Topology With Two-Switch –Approach

Juha Pesonen

ABSTRACT

This application note explains the basic operation of the two-switch Flyback-power stage, and clarifies benefits and design restrictions not present in traditional Flyback-topology. Power stage operation is demonstrated with reference design PMP10037, implemented around a LM5015 High Voltage Monolithic Two-switch DC-DC Regulator. The reference design is available from the TI PowerLab Reference Design Library.

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1   Introduction

Due to simplicity and low parts count, Flyback-topology is often utilized in isolated switched-mode power supplies with an output power of 100 W or less. General operation principles of the Flyback-converter are used by power supply engineers for fast creation of new designs. Because of a low parts count, designs can be made with low cost. Due to its versatility and simplicity, Flyback-topology can be called the 'work horse' of isolated topologies. Even though Flyback-topology is easily used with different applications, there are some drawbacks. The voltage stress of the primary side transistor is high even in an ideal case, where the leakage effects of the transformer are not considered. When current flows on the secondary side, the drain voltage of the primary side transistor rises to the sum of the input voltage and the reflected voltage. When considering the parasitic ringing caused by transistor capacitances and the leakage inductance of the transformer, the voltage stress is even higher. Because the amplitude of the parasitic ringing is hard to predict, the designer must choose a transistor with a high voltage rating with high on-state resistance. High on-state resistance increases conduction losses and leads to a reduction of efficiency. Different kinds of snubbers and clamping-circuits can be used to reduce the transistor voltage stress. For voltage stress, snubbers and clamping-circuits are reasonable solutions, but the energy stored in the leakage inductance is absorbed, reducing efficiency. Figure 1 presents a simplified schematic of the traditional Flyback-power stage.
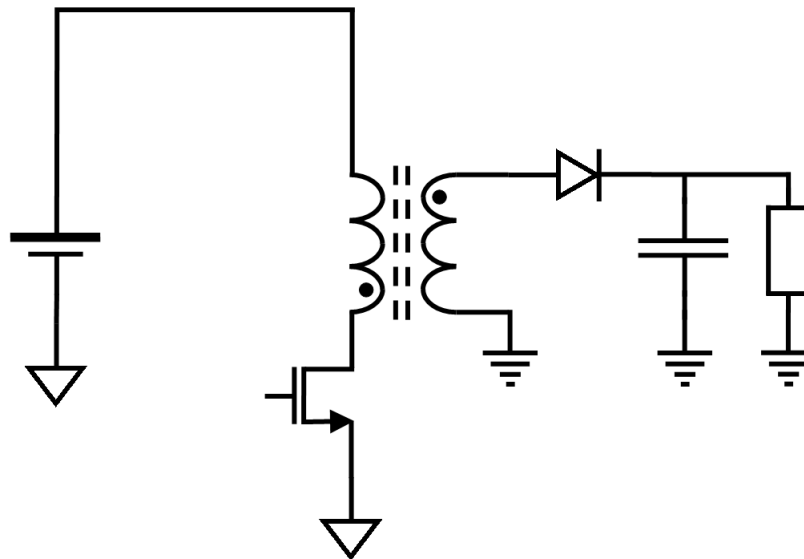

Figure 1. Traditional Flyback-Power Stage

The typical problems of Flyback-topology are overcome by using a two-switch Flyback-topology. When a second transistor is added between the input voltage and the transformer, the overall voltage stress is divided equally over both transistors. Instead of turning leakage energy into losses, it is now returned to the input supply via two diodes. Diodes also clamp drain-source voltages of both transistors to the input voltage, so the voltage rating of the transistors can be selected according to input voltage, without a bigger margin. Due to these improvements, the two-switch Flyback topology is an option over traditional Flyback topology. However, before using two-switch Flyback–topology, design restrictions and considerations must be clarified. Though similarities with traditional the Flyback-topology are notable, different design procedures must be followed. Figure 2 presents simplified schematic of a two-switch Flyback–powerstage.
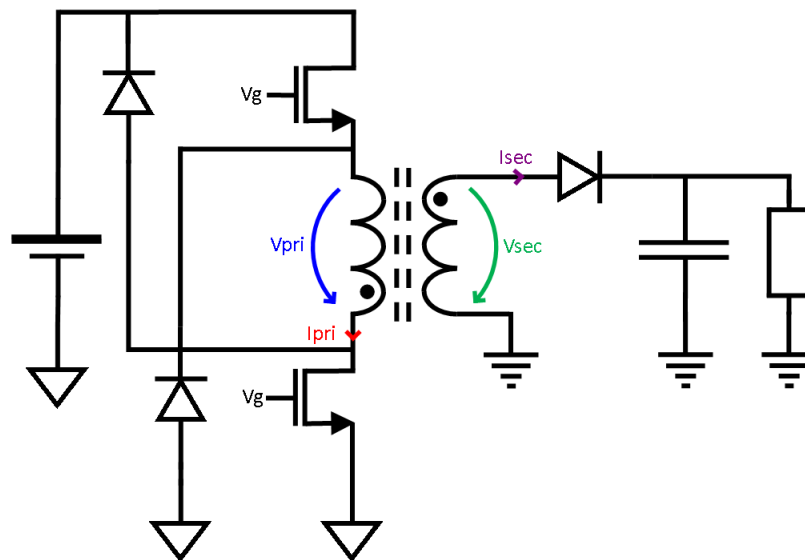
Figure 2. Two-Switch Flyback–Power Stage

2 Principle of Power Conversion

Basic operation of the two-switch Flyback–powerstage is similar to traditional Flyback-topology. In the beginning of the switching period, both transistors are closed and the primary transformer is connected between the input voltage and ground. Current starts to flow through the primary, and the diode on the secondary side is reverse biased due to the transformer polarity. Therefore, all the energy is stored to the transformer while the load current is supplied by the output capacitor. The time frame when the switches are closed is called the magnetizing period. When the switches are opened, stored energy is transferred to the output, supplying the load and the charging output capacitor. At the same time, the reflected voltage is applied over the primary. The time frame when the switches are open is called the demagnetizing period. Figure 3 presents currents and voltages in the power stage when working in continuous conduction mode.
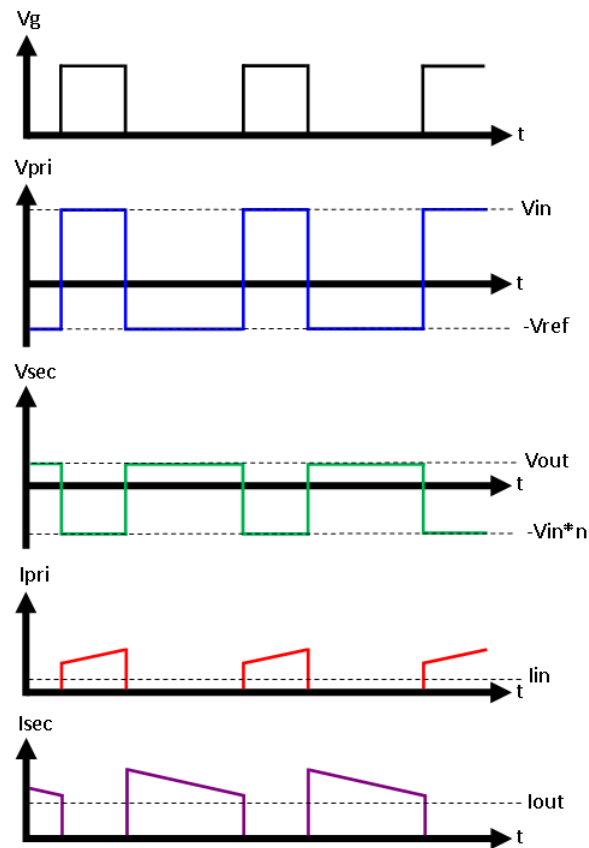
Figure 3. Power Stage Voltages and Currents

3   Transistor Voltage Stress

As mentioned earlier, the overall voltage stress is equally divided over both transistors. To define the voltage rating when selecting transistors, look at the schematic; the non-dot side of the primary cannot fall more than the amount of the diode forward voltage below ground. Similarly, the dot side of the primary cannot rise more than the amount of the diode forward voltage above the input voltage. The maximum voltage stress of both transistors is limited to the sum of the input voltage and the diode forward voltage. In a correctly-designed power stage, the diodes conduct only during switch off moments, when the drain-source –voltage of both transistors tends to rise because of parasitic ringing caused by leakage inductance.

After the parasitic ringing is dampened, the drain-source –voltages settle to a certain value for the rest of the demagnetizing period. To calculate the settled value for the drain-source –voltages, the primary side is described with Kirchhoff's second law according to Equation 1:

$$V_{in} - V_{DS} + V_{ref} - V_{DS} = 0$$

(1)

where V_{in} is the input voltage, V_{DS} is the drain-source –voltage and V_{ref} is the reflected voltage coming from the secondary side. Equation 1 can be modified to present the expression for the drain-source –voltage. This modification is done in Equation 2.

$$V_{DS} = \frac{V_{in} + V_{ref}}{2}$$

(2)

Based on these calculations, the behavior of the drain-source –voltages in continuous conduction mode is presented in Figure 4.
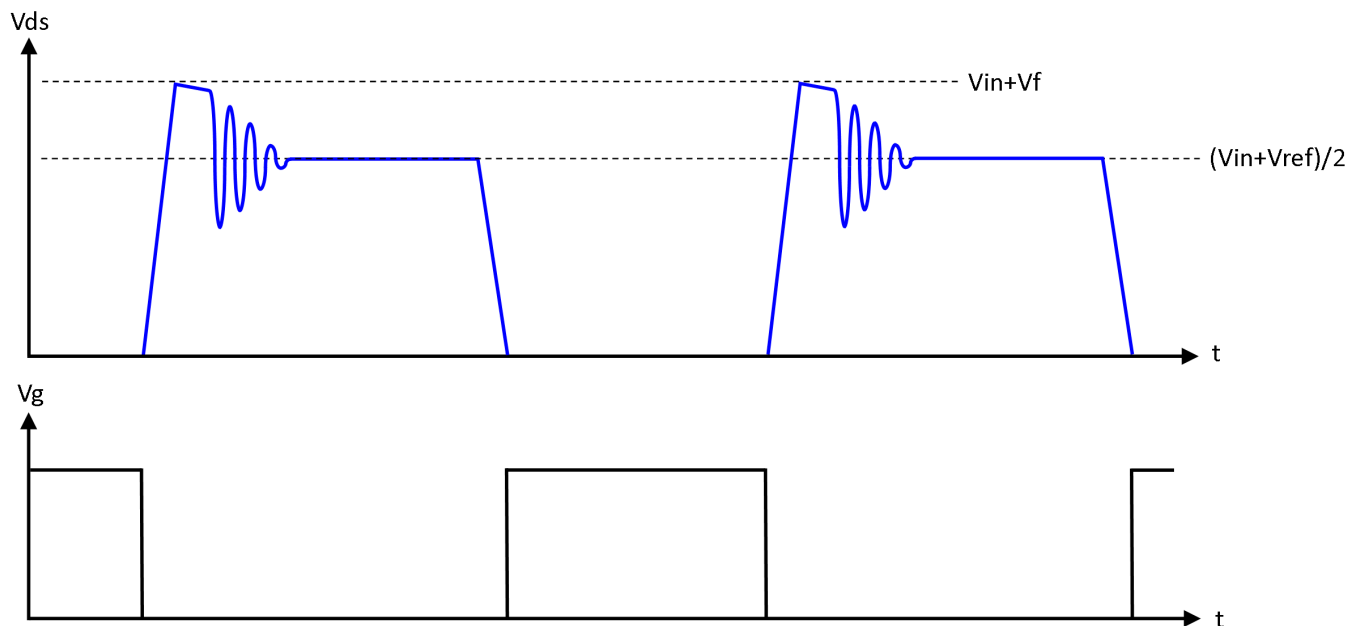
Figure 4. Behavior of Drain-Source –Voltages in Continuous Conduction Mode

Snubber-circuitry is not necessary because additional clamping voltage is not needed. In traditional Flyback-topology, the drain voltage must swing up to the sum of the input voltage and the reflected voltage: the drain voltage cannot be clamped straight to the input voltage. The problem is solved by creating an additional clamping voltage on top of the input voltage. Use a lossy RCD-clamp circuit to absorb the energy stored in the leakage inductance. In a two-switch Flyback-topology, the drain-source –voltages must swing up to only half when compared to a regular Flyback-topology, and therefore it is safe to clamp the drain-source –voltages to the input voltage. Instead of absorbing leakage energy and turning it into losses, the energy is now returned to the input supply.

When operating in discontinuous conduction mode, the drain-source –voltages ring when no current flows through the transformer. Ringing is caused by the transistor capacitances and transformer inductance. When the current stops flowing on the secondary side, there is no reflected voltage over the transformer primary. When no current flows through the transformer, the primary side can be described according to Equation 3.

$$V_{in} - V_{DS} - V_{DS} = 0$$

(3)

The value for the drain-source –voltages is calculated by Equation 4.

$$V_{DS} = \frac{V_{in}}{2}$$

(4)

The behavior of the drain-source –voltages in discontinuous conduction mode is presented in Figure 5.
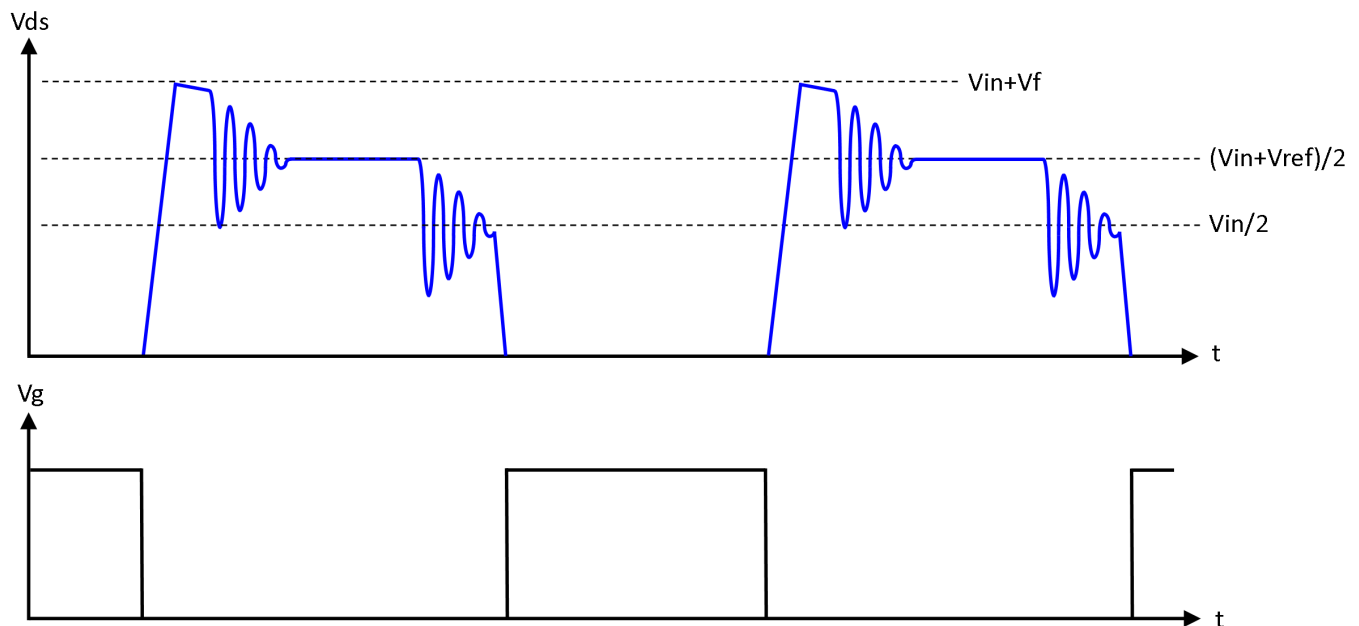
Figure 5. Behavior of Drain-Source –Voltages in Discontinuous Conduction Mode

4   Duty Cycle Limitation in Continuous Conduction Mode

In two-switch Flyback –topology, the diodes work as voltage clampers to reduce the voltage stress of the transistors. Despite this, the diodes cause a design restriction which is not present in traditional Flyback-topology. During the demagnetizing period, the primary side of the transformer is connected with diodes according to Figure 6.
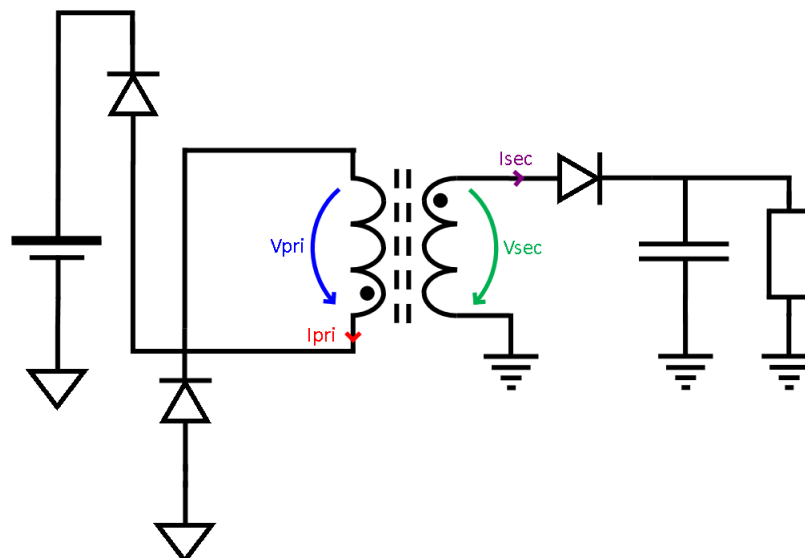

Figure 6. Power Stage During Demagnetizing Period

The current flows in the secondary side and the reflected voltage is applied over the transformer primary. The schematic in Figure 6 shows that the voltage range for the reflected voltage is limited. If the reflected voltage is bigger than the input voltage, it will not fit between the diodes. The diodes begin to clamp, interrupting transformer operation. The maximum value for the reflected voltage s reached by doing a few calculations. To ensure the reflected voltage is not forcing the diodes to clamp, the condition in Equation 5 must come true.

$$V_{DS} < V_{in} + V_f \tag{5}$$

where V_f is the diode forward voltage. Combine Equation 2 and Equation 5 to form Equation 6.

$$\frac{V_{in} + V_{ref}}{2} < V_{in} + V_f \tag{6}$$

The maximum value for the reflected voltage is obtained from Equation 6. This is done in Equation 7.

$$V_{ref} < V_{in} + 2V_f \tag{7}$$

which can be rounded to the input voltage, because it is the most dominating factor. To ensure that the condition in Equation 7 comes true, look at the continuous conduction mode large signal transfer function of Flyback-topology[1] in Equation 8.

$$V_{out} = V_{in} \cdot n \cdot \frac{D}{1-D} \tag{8}$$

where V_{out} is the output voltage, n is the transformer ratio and D is the duty cycle. The reflected voltage is presented with the output voltage and the transformer ratio according to Equation 9.

$$V_{ref} = \frac{V_{out}}{n} \tag{9}$$

Equation 8 can be modified to the form of Equation 10.

$$V_{ref} = V_{in} \cdot \frac{D}{1-D} \tag{10}$$

Equation 10 shows that the condition of Equation 7 comes true if the duty cycle is smaller than 0.5. If the duty cycle is larger than 0.5, the reflected voltage is too high and the condition of Equation 7 is violated. By combining Equation 7 with Equation 10 and neglecting the diode forward voltages, Equation 11 is formed.

$$V_{in} \cdot \frac{D}{1-D} < V_{in} \tag{11}$$

Equation 11 is modified to form Equation 12.

$$D < 0.5 \tag{12}$$

In two-switch Flyback –topology, the duty cycle is theoretically limited to values equal or less than 0.5 when operating in continuous conduction mode. Though it is a good design principle to limit the duty cycle range in traditional Flyback-topology to reduce component stress, there are no such theoretical limitations present.

If the power stage is operating in discontinuous conduction mode, the situation is different. In discontinuous conduction mode, the output voltage is a complicated function of switching frequency, primary inductance, load resistance and duty cycle. Thus the reflected voltage cannot be guaranteed to be lower than the input voltage just by limiting the duty cycle. However, the restriction for the reflected voltage is still valid, and this must be considered when selecting the turns ratio of the transformer.

5   Demonstration

Operation of two-switch Flyback –power stage is demonstrated with the PMP10037-reference design. The schematics, bill of materials, transformer specification, and comprehensive test results of the PMP10037 are freely available in the TI PowerLab Reference Design Library[2].

PMP10037 design specification:

- Input voltage range: 36 to 72 V, 48 V nominal
- Output voltage: 12 V
- Output current: 0.6 A
- Switching frequency: 300 kHz
- Operation mode: CCM

The PMP10037 is implemented around the LM5015 High Voltage Monolithic Two-switch DC-DC Regulator. The PMP10037 contains two integrated 75 V N-Channel MOSFETs and peak current-mode control. The LM5015 features all the functions necessary to implement an efficient high voltage two-switch Flyback –based switch-mode power supply, using a minimum of external components. The latest datasheet and application information can be downloaded from LM5015 product folder[3] on TI.com.

Figure 7 presents voltage waveforms measured from both sides of the transformer primary. Channel 1 is non-dot side (switched to input voltage) and channel 2 is dot side (switched to GND). To highlight the functionality of the diodes, the time scale is small, 100 ns/div. The input voltage is set to 36 V and the output is loaded with 0.6 A load.
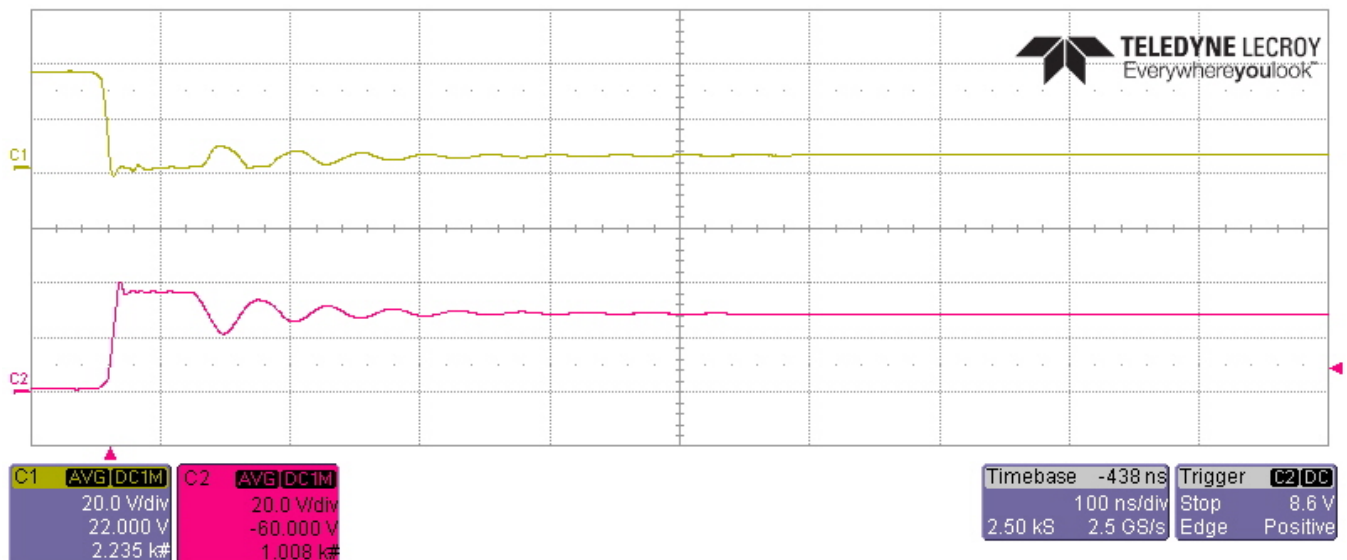
Figure 7. Zoomed Waveforms of Transformer Primary

When looking at Figure 7, the moment when the diodes are clamping is apparent. The drain-source –voltage of both transistors is clamped when it rises above the input voltage. After the ringing is dampened, both sides of the transformer primary settle until the drain-source –voltages are roughly 30 V, which complies with Equation 2. The maximum voltage stress for the transistors is the highest defined value for the input voltage, which is 72 V in this particular case. If the same requirements are covered with traditional Flyback-topology, the voltage stress would be 96 V when transformer leakage effects are not considered. After snubber-circuitry is added, some margin is still needed. The designer chooses a transistor with a voltage rating of roughly 150 V, twice as much when compared with the voltage rating of integrated MOSFETs in the LM5015.

Figure 8 presents the voltage waveforms of the transformer primary over a few switching cycles. Channel 1 is non-dot side and channel 2 is dot side. The input voltage is set to 48 V and the output is loaded with a 0.6 A load.
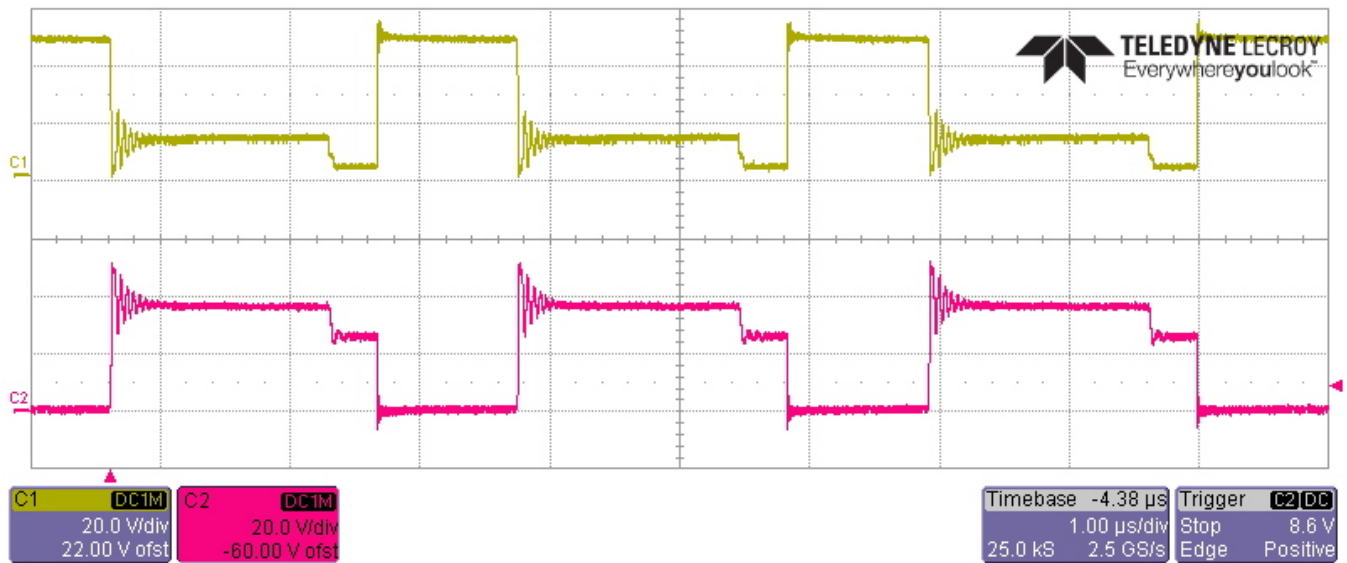
Figure 8. Waveforms of Transformer Primary

Both sides of the transformer primary go down rapidly at the end of the demagnetizing period. This is caused by the control IC, not power stage operation. To control the transistor between the input supply and the non-dot side of the transformer primary, an appropriate high-side driver is needed. In most cases the supply voltage for the high-side driver is generated with a bootstrap-circuit, and the LM5015 is not an exception. Usually the bootstrap-capacitor is self-charged when the low-side transistor conducts. This is possible only in topologies where transistors form a bridge-configuration and are controlled with a complementary PWM-signal. In two-switch Flyback-topology, this is not the case. Transistors are controlled by a simultaneous PWM-signal, and there is no bridge-configuration. To charge the bootstrap capacitor, an additional path from the non-dot side to ground is needed. Figure 9 presents the high-side driver of the LM5015. The bootstrap-capacitor is marked with "C".
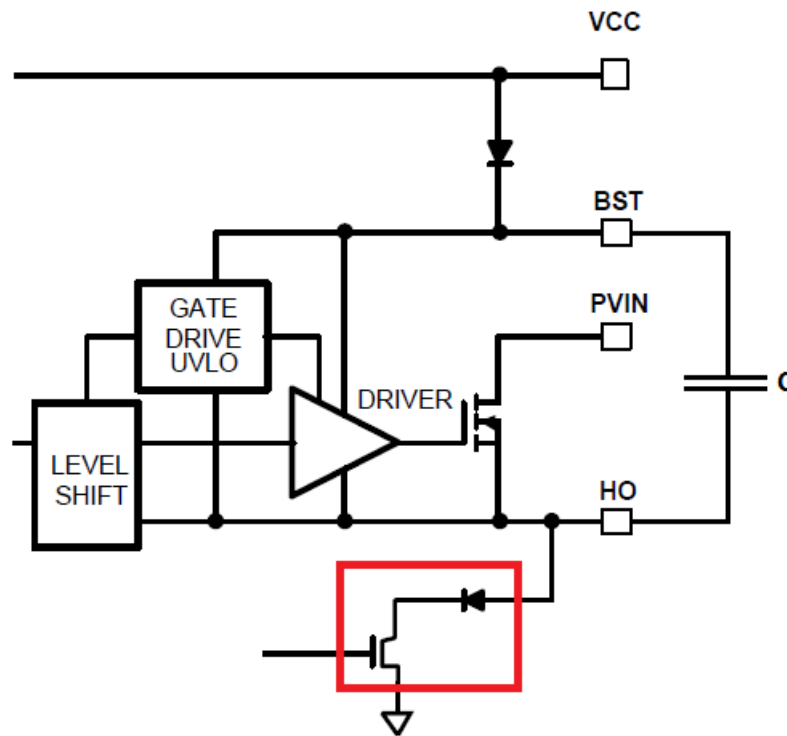
Figure 9. High-Side Driver of LM5015

The LM5015 includes internal circuitry for charging the bootstrap-capacitor. This circuitry consists of a transistor and a diode, and marked in Figure 9 with a red square. When the transistor conducts, the bootstrap-capacitor is charged from the supply voltage of the LM5015. The internal transistor is controlled, so that the bootstrap-capacitor is always charged at the end of the switching period. When the bootstrap-capacitor is charged, the non-dot side of the transformer primary is pulled low. To maintain a volt-second balance, the dot side of the transformer primary has to follow. As seen in Figure 8, when the non-dot side is pulled low at the end of the demagnetizing period, the voltage on the dot side also drops by the same amount. Despite this, the differential voltage over the transformer primary stays constant, and transformer operation is not violated in any way. The voltage waveform of the transformer primary is presented in Figure 10.
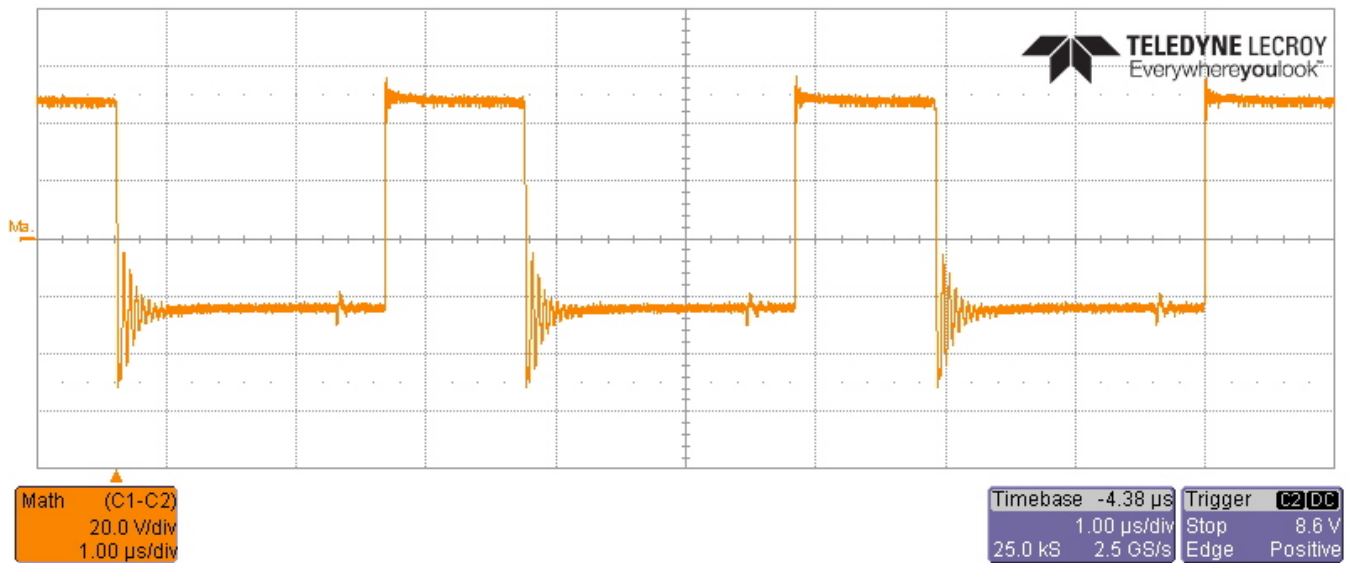
Figure 10. Differential Voltage Over Transformer Primary

If external transistors are used, driving the high-side transistor requires more effort. One solution for generating the supply voltage for the high-side driver is auxiliary winding, which generates the necessary supply voltage on top of the primary voltage. Another solution is to use a pulse transformer. However, these methods can cause unwanted issues, such as increasing complexity and size. The LM5015 integrates everything inside one chip keeping complexity, size, and costs low.

Figure 11 presents efficiency as a function of the load current, measured with different values of input voltage. Peak efficiency is over 90%, a good result when dealing with low output power. The typical peak efficiency of Flyback-converters with an output power lower than 10 W lies somewhere between 85% and 88%.
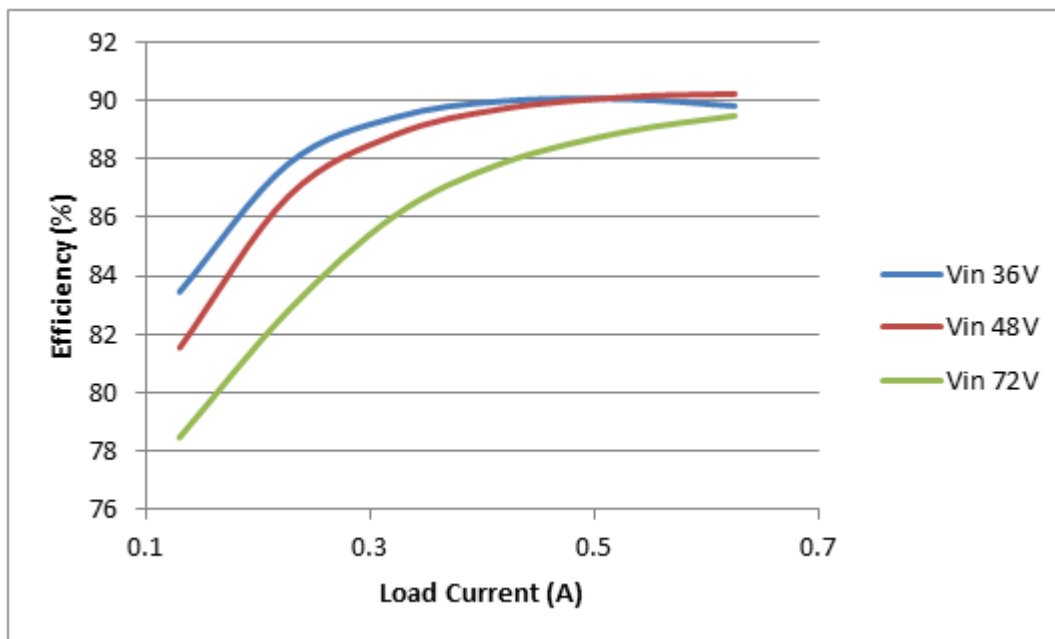
Figure 11. Efficiency as a Function of Load Current

6   Conclusion

This application report describes the drawbacks of traditional Flyback topology and clarifies how these drawbacks are overcome with a two-switch approach. The basic operation and design restrictions of the two-switch Flyback topology are discussed and the theory behind the design restrictions is also provided. The operation of a two-switch Flyback power stage is demonstrated with the PMP10037 reference design, implemented around the LM5015 High Voltage Monolithic Two-switch DC-DC Regulator.

While in the two-switch –approach, overall parts count is increased, a slightly larger parts count is not a major drawback. The energy stored in the leakage inductance is relatively small, so cheap diodes with a small package and a low current rating can be used. Generally transistors with a lower voltage rating are cheaper, and the on-state resistance of two transistors is smaller than one transistor with a double voltage rating, meaning performance improves while costs stay the same. There is also no need for lossy snubber-circuits in the two-switch approach.

7   References

[1] Robert W. Erickson & Dragan Maksimović: *Fundamentals of Power Electronics*, Second Edition. Kluwer Academic Publishers, 2004.

[2] PMP10037 in PowerLab Reference Design Library: http://www.ti.com/tool/pmp10037

[3] LM5015 product folder: http://www.ti.com/product/lm5015

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