

Note: If the reader is unfamiliar with the SMPS concept or with some of the terminology, he may refer to APPENDIX 1 page 15.

### INTRODUCTION

This application note, AN\_EVP001, was created to complement the IN-PLUG® product datasheets and help SMPS designers select the right AAI Controller IC for building low power, off-line flyback SMPS.

The goal is to build the most forgiving and cost-effective state of the art SMPS using AAI's patented approach that eliminates the necessity for several standard SMPS components.

The input mains line voltage can be either 110 volts AC (Domestic US) or 85 to 265 volts AC (International.) The line frequency can be 50, 60 or 400 Hertz nominal. Most designs will also operate from high-voltage DC with any polarity, like 140 volts or 120 to 400 volts.

The emphasis is on low-power, low cost SMPS with one isolated low-voltage output. However, please note that AAI ICs can also be used in higher power SMPS applications, especially when high output voltages and multiple outputs are required by applications as for example in CRT monitors.

When designing an SMPS, please check which governmental regulations your product will have to comply with, such as the following:

- Power Factor Correction (PFC),
- Electro-Magnetic Interference (EMI or EMC),
- Safety issues

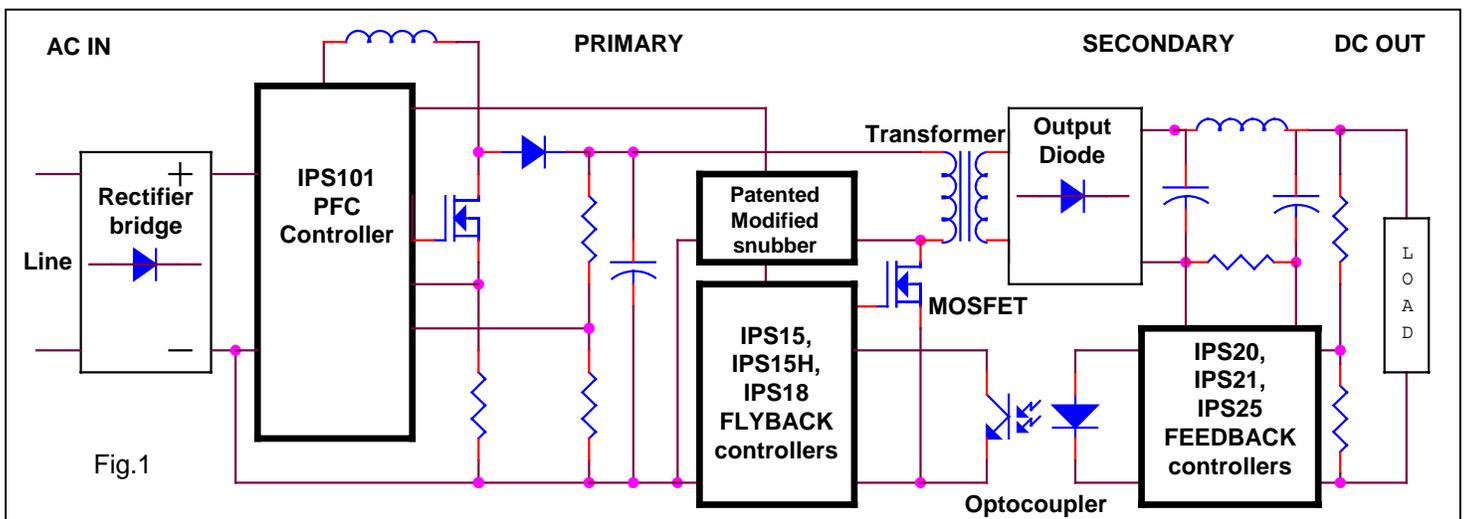
This application note deals with electro-magnetic interference and safety issues, but not with power

factor correction which is not yet required for SMPS below 75 watts of input power. Other IN-PLUG® products like the IPS101 are offered for higher power SMPS with power factor correction.

Excluding the PFC, an actual SMPS will include:

- The mains AC line input with the Diode Bridge, Storage Capacitor(s), EMI-reduction Choke as well as safety and protection components like Fuse or Fusible Resistor, MOV and X-Capacitor;
- The high-voltage Power MOSFET with its associated protection and EMI-reduction components;
- The line-side controller with its associated components;
- The secondary Power Diode, Tank Capacitors and Ripple-reduction components;
- The load-side controller with its associated components.
- The components between line-side and load-side, e.g., the Switching Flyback Transformer with its isolated secondary, the feedback Optocoupler and the Y-Capacitor.

**Please, see Figures 9 and 10 page 12, for detailed application schematics.**



## SELECTION OF FLYBACK AND FEEDBACK CONTROLLERS

### Line Side Flyback Controllers:

The IN-PLUG® IPS1x series (IPS15, IPS15H, IPS18) was designed for low-cost, high efficiency, low-power flyback off-line switching power supply solutions up to approximately 50 to 70W. It contains a shunt-regulator, a precision oscillator, a PWM with its associated comparator and loop compensation components as well as all the necessary biasing and protection circuitry (thermal shutdown, undervoltage and overcurrent).

It is optimized to operate with an optocoupler to provide the feedback from the secondary but could also be used with a bias winding which could sometimes be more economical. Typical applications include domestic and international power supply featuring and AC input ranging from 90 to 264V and DC from 100 to 350V.

IPS15H and IPS18 incorporate a special circuitry that involves a counter and some other digital blocks. This feature has been added to avoid delivering a high current to the load on an overload condition, which may result in damages to the SMPS, the load or both. IPS18 has been optimized for stand-by applications where the SMPS has to deliver a minimum amount of power while being required to comply with tight "green" regulations. When no load or a minimum load is connected to the SMPS, the controller enters a cycle skipping mode to keep the power consumption as low as possible.

Note: In low-power applications, the IPS1x series can eliminate MOV, Common Mode Choke, X and Y Capacitors and still enable SMPS to meet the strictest government regulations for EMI and input surge voltages.

### Load Side Feedback Controllers:

For applications requiring current limiting & voltage regulation, or voltage limiting & current regulation, select AAI's feedback controllers: IPS25, IPS20 or IPS21. Otherwise, select an appropriate zener diode or a TLC431 adjustable shunt regulator as shown in Figures 2 and 3 below.

- If the output current is below 700mA, → **select IPS25.**
- If the output current is 700mA and above, → **select IPS20 or IPS21.**

Note:

- Both the IPS20 and 21 incorporate the same trimmed temperature compensated 1.2V reference to set the output voltage. They only

differ by the characteristics of the current sensing references. The IPS20 features a trimmed 50mV reference with a positive temperature coefficient which closely matches that of a PCB copper trace. This copper trace sensing method can be used inexpensively with very low associated power losses. The IPS21 incorporates a constant 100mV reference for more conventional resistors

- The IPS25 can be used in high current application where efficiency is non-critical. (A 1.3 volts drop is required across the current-sensing resistor).
- The IPS20 and 21 can be used at any output current for maximum efficiency. However, an auxiliary supply is required to power the chip under a dead output short-circuit condition.

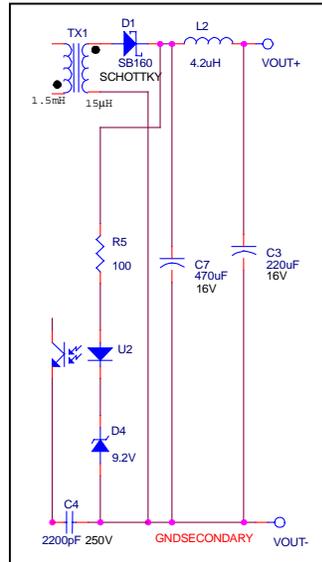


Fig. 2

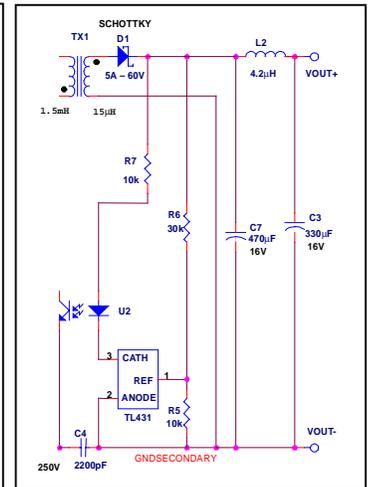


Fig. 3

## SELECTION OF THE MOSFET

### MOSFET Breakdown Voltage:

State of the art low-cost, high-voltage MOSFETs suitable for off-line flyback SMPS are available in breakdown voltages of 400, 600 & 800 volts.

- For Domestic applications, 400 volts is sufficient, though 600 volts are sometimes more readily available and equally suitable.
- For International application 600 volts can be used when the transformer turn-ratio is less than seven. 800 volt MOSFETs should be used with higher turns ratio transformers which generate larger voltage spikes due to higher leakage inductance and increased reflected voltages. With a careful design, an 800 volts MOSFET could somewhat increase the SMPS efficiency.

(See also transformer and snubber sections)

### ***MOSFET Current:***

For International applications, a good approximation on how to rate a MOSFET is as follows:

$$I_p (\text{MOSFET}) = \text{Input Power} / 20.$$

$I_p$  = peak current in the MOSFET

The input power can be deducted from the output power, assuming an efficiency of about 75% accounting for losses.

$$\text{Input Power} = 1.25 \times \text{Output Power}.$$

Please add some safety margin when choosing the MOSFET. Modern high-voltage MOSFETs have very good switching characteristics, but tend to de-saturate and behave like a current source when the application slightly exceeds their nominal current.

Example:

80W of output power requires  $100 \times 1.25 = 125\text{W}$  of input power and a  $125/20 = 5\text{A}$  MOSFET. Use a 5A or preferably a 6A MOSFET.

### ***MOSFET Gate Charge:***

The total gate charge is an important parameter which governs how much operating current will be needed to drive the gate. The IPS1x series is designed to drive MOSFETs with up to 50nC of total gate charge, but modern MOSFETs are available with typically 5nC for a 1A 600V device and 8-10nC for a 2A 600V device. The lower the gate charge, the faster the device and the lower the switching losses. Therefore, choosing a state of the art MOSFET with a low total gate charge has many advantages, usually at little or no extra cost.

### ***MOSFET Ron and Capacitance:***

Choosing a MOSFET with a low Ron resistance will reduce the conduction losses, but this is of little concern with a low-power flyback SMPS. Greater losses occur when the MOSFET's capacitance combines with the transformer's primary distributed capacitance. Therefore, it is better to compromise on Ron to choose a MOSFET with a low total gate charge and smaller Drain & Miller capacitance.

Note: Recent evolution in high-voltage MOSFET technology has achieved a substantial reduction of the Ron without increasing the drain capacitance.

### ***MOSFET Avalanche Rating:***

The avalanche rating of a high-voltage MOSFET defines how much energy the MOSFET can safely absorb when its maximum voltage is exceeded. The SMPS should be designed to never cause avalanche under normal operating conditions.

This is achieved by:

1. Choosing the right breakdown voltage for the MOSFET,

2. Designing the transformer for minimum leakage inductance,
3. Using a suitable Snubber Network
4. Clamping the spikes for higher power applications.

When all these requirements are fulfilled, the avalanche rating becomes an additional feature that increases the reliability of the SMPS and prevents accidental destruction in abnormal conditions like Line Surges, Lightning, etc.

**Refer to IXYS Polar N - Channel MOSFETs family to select the suitable MOSFET.**

## SELECTION OF THE OUTPUT DIODE

### ***Type of Diodes:***

- Schottky Diodes:  
Schottky Diodes are preferred due to their unique combination of low Forward Drop and extreme speed. Their utilization will increase the efficiency of the SMPS by reducing the losses in the diode itself and reducing the switching losses in the MOSFET. They can be used in Flyback SMPS operating in either Continuous or Discontinuous modes. Unfortunately, their breakdown voltages cannot economically exceed 60 volts, which prohibits their utilization in certain applications involving large input line voltages, transformers with small turn ratio, large output voltages.
- Fast Silicon Diodes:  
Fast Silicon Diodes, preferably with low series resistance (FRED Diodes), are very economical and can be found in a broad range of current and voltage ratings. They can be used in Flyback SMPS that do not operate in continuous mode.
- Ultra-Fast Silicon Diodes:  
Ultra-Fast Silicon Diodes are required for Flyback SMPS that are allowed to operate in continuous mode.

### ***Output Diode Current rating:***

To maximize the efficiency, the diode will have to handle a peak current equal to the peak current in the MOSFET, multiplied by the turn ratio of the transformer:

$$I_p(d) = I_p(\text{MOSFET}) \times N_p / N_s$$

Where  $I_p(d)$  is the peak current in the output diode,  $I_p(\text{MOSFET})$  is the Peak Current in the MOSFET,  $N_p$  is the number of turns of the primary transformer winding,  $N_s$  is the number of turns of the secondary transformer winding.

Using a diode for which the voltage drop is not guaranteed at the intended operating peak current will

result in excessive power dissipation. Due to the triangular shape of the current and the duty cycle, this will result in choosing a diode for which the nominal current could be as high as 2 to 3 times the output current of the SMPS.

Looking at the forward drop on the bench or on a curve tracer is also a good way to check the suitability of a given diode. The forward drop measured at the peak current should not exceed:

- 0.7 volts for a Schottky Diode,
- 1.2 volts for a Fast or Ultra-Fast Diode.

Refer to IXYS shottky diode DSS family to select the suitable diode.

### ***Output Diode Voltage Rating:***

The diode will have to block a voltage which is the sum of:

- The maximum output voltage ( $V_{out}$ ),
- The maximum rectified line voltage reflected to the secondary of the switching transformer when the MOSFET turns ON. ( $V_{AC} \times 1.414 \times N_s / N_p$ )
- Some spikes and ringing not completely eliminated by the snubber network.

## SELECTION OF THE TRANSFORMER

### ***Importance of a Good Transformer:***

The transformer is a key component of a good SMPS since it affects all aspects of the project, as well as the selection of other key components and the possible elimination of others. In a flyback SMPS, the Transformer operates more like an inductor with two (or more) windings than a typical transformer:

- the Primary winding is used to magnetize the core and store the energy on a cycle to cycle basis;
- the secondary winding demagnetizes the core and transfers the stored energy to the load side.

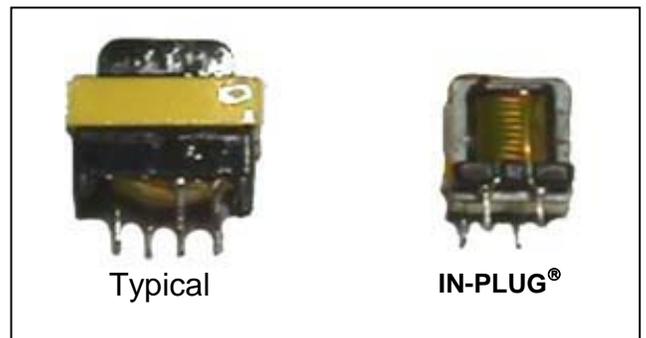
The energy is transferred sequentially from the line side to the load side while ensuring proper isolation from dangerous line voltages. However, despite the fact that it operates like an inductor, the bi-directional magnetic coupling between primary and secondary reflects voltages and spikes that must be accounted for in the breakdown voltages of associated switching components like MOSFET and diodes.

### ***Major Parameters Affected by the Transformer:***

1. The size of the SMPS,
2. The frequency of operation with effects on Size, Ripple and EMI,
3. The efficiency through the losses in the transformer itself and those induced in the MOSFET and the Snubber Network by the Leakage Inductance and the primary capacitances,
4. The voltage and current ratings of the MOSFET,
5. The type, voltage and current ratings of the output Diode,
6. The Y-Capacitor and the loop stability through the primary to secondary stray capacitance and common mode noise injection,
7. EMI compliance,
8. Compliance with Safety Agencies requirements.

### ***Important Transformer Parameters:***

1. Frequency of operation,
2. Maximum permissible power at Maximum temperature,
3. Maximum permissible duty cycle at Minimum input voltage and Maximum temperature,
4. Primary Inductance,
5. Primary Resistance,
6. Secondary Inductance (or  $N_p/N_s$  turn-ratio),
7. Secondary Resistance,
8. Leakage Inductance,
9. Primary distributed capacitance,
10. Primary to secondary capacitance,
11. Isolation voltage and construction technique suitable for Safety Agencies approval.



## HOW TO DESIGN A GOOD SWITCHING TRANSFORMER FOR LOW-POWER APPLICATIONS

Designing a good transformer for low-power, low-cost applications can be counter-intuitive. The following analyzes in the order of their importance the critical SMPS parameters that are directly impacted by the design of the transformer itself.

### *Safety Agencies Requirements:*

The approval of SMPS by safety agencies creates tremendous constraints on the selection of the transformer's core, the transformer's construction and some electrical parameters.

With larger size transformers, "margin tape" construction is considered the most cost-effective ways of complying with agencies requirements. This technique is usable down to a 19mm E-core. But for smaller E-cores, the margin tape construction substantially degrades the power capability and the leakage inductance.

For smaller and more compact SMPS, the only solution is to use "Triple Insulated" wires for the secondary and make sure that the "Creeping Distance" requirements are respected either at the Bobbin or at the PCB level. Using Triple Insulated wires, high performance transformers can be built for:

- 6 watts international applications with a 13mm E-core.
- 15 watts international application with a 16mm E-core.

### *Leakage Inductance:*

Leakage inductance should be as low as practical even if it means compromising with other parameters like copper losses.

The leakage inductance of optimized transformers tends to be higher with E-cores than with PQ cores which unfortunately do not exist in very small sizes. Therefore AAI has designed transformers using E-cores that take full advantage of the IPS15 ability to eliminate a number of external components from the PCB.

One technique to reduce the leakage inductance in E-core transformers is to use a turn-ratio  $N_p / N_s$  that does not exceed 7. This is easy to do for domestic applications and practically always allows the utilization of a Schottky diode for the load side.

It is also possible to use such low turn-ratios in many international applications. This will in-turn allow using of a 600V MOSFET; and in many cases, when the output voltage is 12 volts or less, a Schottky diode as well.

It is more difficult to design a low-leakage inductance transformer for applications requiring larger turns-ratio without sacrificing several other parameters. In fact, a turn-ratio in excess of 14 is not practical and should be avoided.

Note: AAI is developing a transformer intended to deliver 35 watts using a turn-ratio of 14 on a 20mm PQ core. This transformer shows excellent efficiency in flyback mode. A custom bobbin is presently being tooled in order to pass UL requirements.

### *Primary Distributed Capacitance:*

Reducing the primary distributed capacitance is important for the efficiency of the SMPS, as discussed in section "MOSFET Ron and Capacitance" p 3. The best results are obtained when the primary is wound with either 1 or 3 full layers of magnet wire. The one layer solution may concern people not used to switching transformers, since they involve current densities in the order of 10A/mm<sup>2</sup>. However, such values are perfectly acceptable for a single layer winding which evacuates the generated heat more easily than multi-layer ones.

### *Primary to Secondary Capacitance:*

In applications where the "Touch Current" must be reduced to under 100 $\mu$ A or in order to simply eliminate Y-capacitor, the primary to the secondary capacitance becomes critical.

For such applications, optimized transformers can be built that have acceptable leakage inductance and less than 16pF of primary to secondary capacitance. Unless the primary to secondary capacitance is 16pF or less, the Y-capacitor cannot be eliminated without causing loop stability problems that becomes difficult if not impossible to manage even with AAI's feedback controllers.

Note: Adding an internal shield is not an option because it will degrade the leakage inductance of a small-size transformer by a large factor.

### *Maximum Permissible Duty Cycle when Hot:*

Most switching transformers are made of ferrite materials which tend to rapidly lose their magnetic characteristics at elevated temperature. Since flyback transformers are designed with an air gap, the permeability of the ferrite material is of less importance than the magnetic induction at saturation and, to some extent, the losses at the operating frequency.

Since the flyback operation involves operating in a broad range of duty cycles, the core should not be allowed to saturate in the worst-case conditions as follows:

1. Minimum input voltage (100VDC or 180VDC);
2. 60% duty cycle meaning for example 6 $\mu$ s when operating at 100 KHz;
3. Core temperature at full load and maximum ambient temperature.

### ***Frequency of Operation:***

Higher frequency does not always mean better SMPS, and therefore in our opinion the choice of the operating frequency must be based on a number of considerations discussed in section "AAI patented snubber network p7.

### ***Maximum Permissible Power:***

As indicated in section "safety agency requirements" p5, AAI has designed modern and optimized transformers for small, low-cost SMPS in the following power ranges:

- 6 watts international applications with a 13mm E-core;
- 15 watts international application with a 16mm E-core;
- 35 watts international application with a 20mm PQ core.

Larger power ratings are possible, especially when multiple-outputs and relatively high output voltages are required. One typical example is a 75 watts SMPS for CRT monitors which involves 3 low-voltage outputs and one 75 volts output.

### ***Turn-Ratio:***

The turn-ratio is important for:

1. Reducing the leakage inductance as discussed in section "leakage inductance" p5,
2. Taking advantage of Schottky diodes in international applications delivering 12 volts or more,
3. Increasing the maximum power with a given core and MOSFET.

Two possible optimizations: one with turn-ratios not exceeding 7; another with turn-ratios between 10 and 14. Values greater than 14 are typically not practical.

### ***Primary Inductance***

Even though the primary inductance is key to the SMPS operation, it is not a high priority. Due to the gap in the core, the required inductance can be achieved in different ways, whereas there is less latitude for other requirements outlined above.

The optimum primary inductance can be calculated from the following formula:

$$L_p = 800 / (P \times F)$$

Where:

L<sub>p</sub> is the primary inductance in Henrys,  
P is the Input Power in Watts, not the output power. (In doubt use 75% efficiency),  
F is the frequency in Hertz.

Note:

1. Use the formula "as is" for international applications using a turn-ratio of 7.
2. Use  **$L_p = 1000 / (P \times F)$**  for international applications using a turn-ratio of 14.
3. Increase the inductance value by 30% for 220VAC only applications

### ***Primary Resistance:***

As indicated in section "important transformer parameters" p4, contrary to general belief, the primary resistance is not a concern in most applications. In fact, conduction losses which are small compared to the switching losses are due to:

1. the primary and secondary resistances,
2. the MOSFET R<sub>on</sub>,
3. the MOSFET current-sensing source resistance.

### ***Secondary Resistance:***

As indicated in section "Primary Resistance" above, the secondary resistance is not a critical parameter. However, when using magnet or triple-insulated wires for the secondary winding, it is important to adjust the diameter of the wire to entirely fill one full layer. Several wires in parallel in one layer are perfectly acceptable since they increase the copper section and somewhat decrease the leakage inductance at the same time.

## SNUBBER NETWORK

### Origin:

Originally, the purpose of the so-called "Snubber Network" was to reduce the switching losses of the bipolar transistors, which at the time were slow and would otherwise have died from excessive losses.

As bipolar transistors gave way to faster switches like MOSFETs, reducing the switching losses became less of an issue. However, designers found that a snubber network was then required to slow-down fast voltage transients that otherwise created unacceptable EMI.

Though it is not its primary function, up to certain power levels, the snubber network can reduce the over-voltage drain spikes created by the leakage inductance of the transformer. Power levels in excess of 15-20 watts, usually require the utilization of a clamp, especially when a 600 volts MOSFET is used in international applications.

### Typical Snubber Network:

Figure 4 and Figure 5 show two typical snubber networks connected between the MOSFET drain and the primary ground. It consists of a fast diode D1 and a capacitor C1 which voltage rating should equal that of the MOSFET and two resistors R1 and R2.

In the arrangement of Figure 4, R1 is in series with D1 and can be omitted in many applications. When the MOSFET is ON, the time constant  $R2 \times C1$  should be low enough for C1 to be almost discharged at the end of the ON time. When the MOSFET turns OFF, C1 charges through D1 and R1 and slows down the transition, reducing the switching losses and improving the EMI at the expense of some power dissipation in R1 and R2.

This is the snubber effect. R1, which value is usually low, increases the dynamic resistance of the diode D1 in order to prevent possible ringing created by too fast a transition in the diode. This further improves the EMI.

The arrangement of Figure 5 can be used when R1 in Figure 4 can be reduced to zero without harmful effects on EMI. Then, placing a high-value resistor R1 in parallel with D1 creates a damping effect on the transformer at the end of the demagnetization cycle. This can greatly attenuate the free oscillations that would otherwise normally occur and adversely affect the EMI signature.

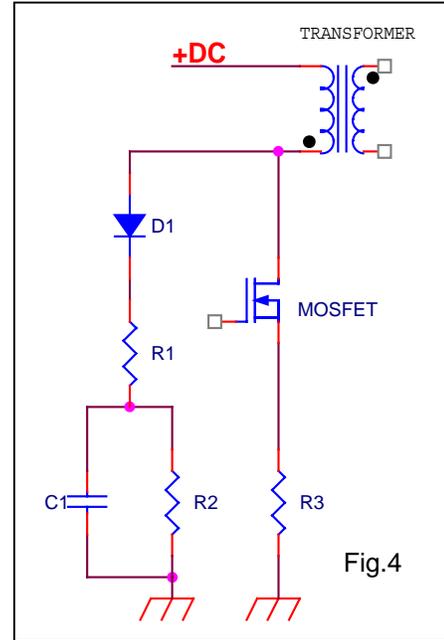


Fig.4

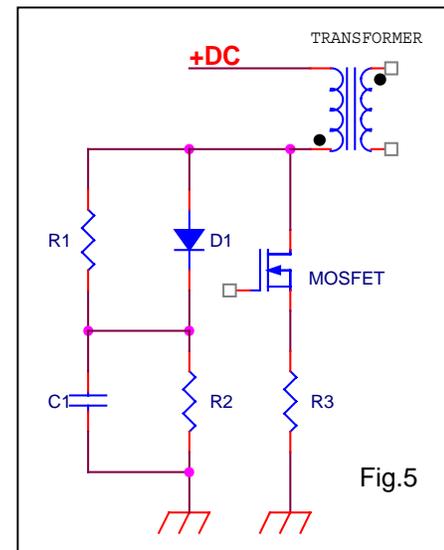


Fig.5

### AAI Patented Modified Snubber Network:

Since most SMPS require a snubber network, AAI has modified the classical schematics and patented the circuit in Figure 6 (Patent No. US 6,233,165 B1 has been granted by the US Patent Office on May 15 of 2001.) The circuit consists of two diodes D1 & D2 which are low-cost low-voltage signal diodes, one low-value capacitor and one resistor R1.

As before, when the MOSFET turns OFF, C1 charges through R1 to slow-down the transition; but C2, which is part of the Line-side controller IC circuitry, is also charged through D2. When the MOSFET turns ON, D1 becomes forward biased and C1 discharges to be ready for a new cycle.

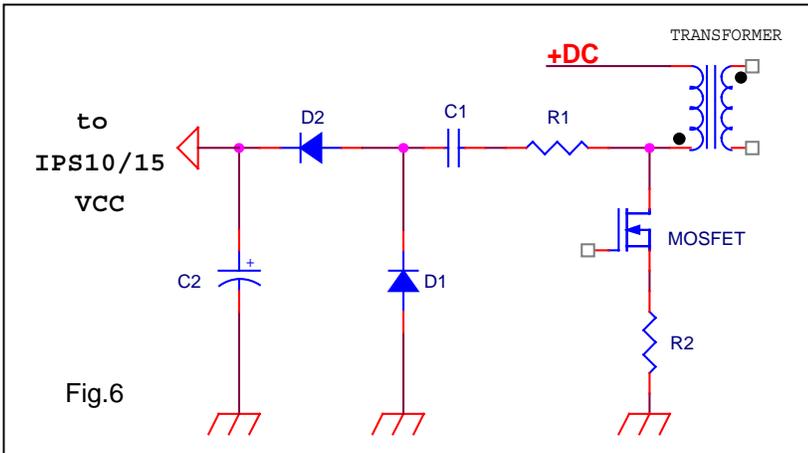
As switching occurs, there is enough power transferred to C2 to energize the entire line-side

circuitry without having to use an auxiliary power supply.

This arrangement totally eliminates the need for a third winding on the switching transformer (sometimes called "Bias Winding") and results in cost and space savings. It also makes the transformer more forgiving to design and manufacture.

**Important Note:**

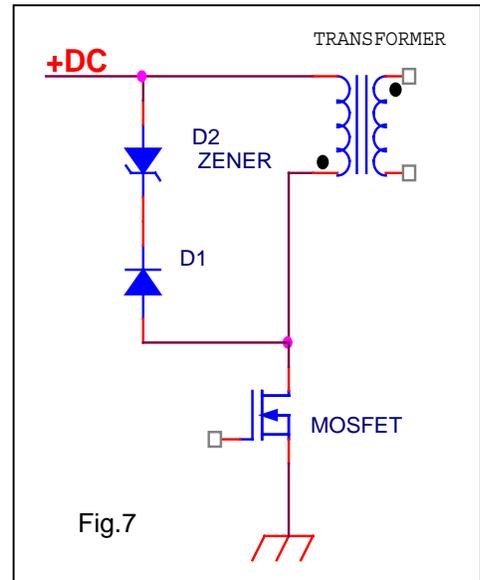
AAI's policy is to give a royalty-free, non-exclusive license to use the above-referenced patent to every user of IPS products who purchase their ICs either directly from AAI or through authorized distributors / representatives.



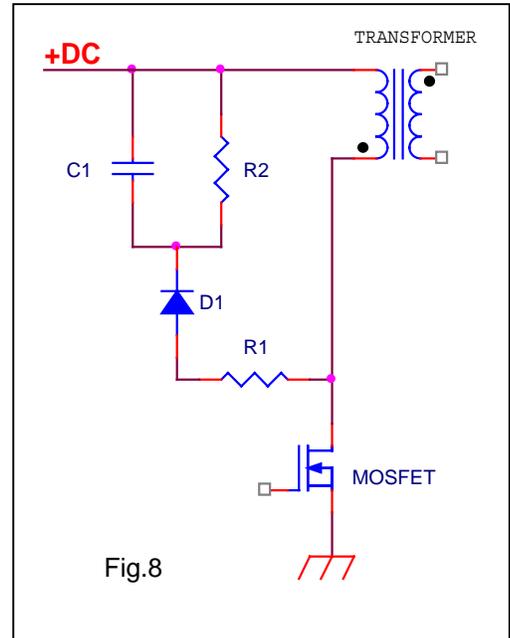
**DRAIN CLAMP**

By slowing down the transitions and creating some controlled losses, the snubber network tends to reduce, at the MOSFET drain, the voltage spike created by the transformer leakage inductance. For high power applications, however, and especially when using a 600 volt MOSFET in international applications, the utilization of a snubber network may not be sufficient to clamp the drain voltage to an absolutely safe limit, and an actual Clamp Circuit may be required to perform just that.

The two types of clamps commonly used for this purpose are shown on Figure 7 and Figure 8. Clamp Figure 7 consists of a high-speed high-voltage diode D1 in series with a Zener diode. The Zener voltage should be 20 to 30 volts above the output voltage reflected through the transformer;



the zener should be sized to dissipate the full energy stored in the leakage inductance of the transformer. Clamp Fig. 8 resembles a traditional snubber network with different component values, that returns to the +DC and not to GND. Here again R1 may be omitted in some applications if little or no ringing occurs when D1 starts to conduct.



Note: Even though the clamp in Figure 8 has a lower efficiency than the clamp in Figure 7, it is used more often because it is using readily available, low-cost components compared to high-voltage zener diodes, which have to be special ordered.

## CRITICAL DISCRETE COMPONENTS

### *Input Capacitor:*

The input tank capacitor serves a critical purpose: it is being charged to the peak line voltage at twice the line frequency and reduces the ripple of the rectified DC feeding the switching circuitry.

#### *Value:*

Use a capacitance value sufficient to keep the lower point of the rectified and filtered AC line above 100V. For SMPS up to 35 watts, 2  $\mu$ F per watt is usually enough. See also ESR and Temperature rating below.

#### *Voltage Rating:*

For Domestic Applications use 200 or 250 volts capacitors. For International Applications use 400 volts, or 450 should the SMPS be capable of surviving 300VAC Line Voltage for some period of time.

#### *ESR and Temperature Rating:*

It is critical to use a low ESR capacitor. Check with the manufacturer's rating to make sure that 2 $\mu$ F per watt is enough to handle the peak current. 85C rated capacitors usually require more than 2 $\mu$ F per watt and therefore capacitors rated 105C are often a better choice.

**EMI-reduction Tip:** Splitting the required input capacitance between two smaller size capacitors connected by a small inductor creates a low-cost filter that can pass the EMI requirements for low-power SMPS. This is especially true when combined with other EMI reduction techniques like:

- MOSFET gate-drive optimization featured by the IPS15;
- Optimum Snubber Network;
- Low leakage inductance transformer;
- "Low-noise" transformer design.

#### *Start-up Resistor:*

The IPS 10 and IPS 15 both feature a starting current of approximately 120  $\mu$ A. The starting current must be provided by an external start-up resistor. Once this resistor has pulled the VCC up to approximately 10 volts, the SMPS starts and the IC, which then requires more than 120 $\mu$ A, is powered through AAI's Patented Modified Snubber Network or other suitable means.

A 900K $\Omega$  resistor is necessary to ensure proper start-up at a line voltage of 85VAC (Minimum for the "International" range), whereas a 2 M $\Omega$  resistor is sufficient for 180VAC minimum line voltage.

### *Bias Resistor:*

This resistor allows the user to operate at the switching frequency which is best for the application in terms of SMPS size, efficiency and EMI compliance. Use the information provided in the IPS15, 15H and 18 datasheet documents to select the appropriate value.

**Please remember** that the switching frequency affects the following three parameters:

- The transmitted power in watts is:

$$1/2 \times L \times I^2 \times F,$$

where L is the primary inductance of the transformer, I is the peak current in the transformer, and F the operating frequency.

- Within limits, the higher the frequency the smaller the inductance and therefore the smaller the transformer.
- The switching losses which increase proportionally to the frequency and are described thereafter:

### *Switching Losses:*

Switching losses in the MOSFET heat the device and degrades its efficiency. In comparison, the conduction losses in the MOSFET are usually small. This means that the Ron of the MOSFET is not as critical a parameter as one may think for Flyback SMPS operating in discontinuous mode.

### *Additional Switching Losses:*

Additional switching losses caused by the capacitance seen by the drain of the MOSFET. Such capacitance consists of the distributed capacitance in the primary winding of the transformer in parallel with the MOSFET capacitance. Depending on the exact waveforms, this stray capacitance can generate up to an additional  $1/2 \times C \times V^2$  of losses. This further degrades the efficiency and generates more heat in the MOSFET. This effect underscores the importance of the following:

- Using a state of the art MOSFET with low stray capacitance;
- Giving preference to a low-capacitance design for the transformer.

### *EMI:*

EMI is the third important factor to take in account when selecting the operating frequency. Since the energy spectrum tends to show that harmonics of higher order are less energetic, operating at the lowest possible frequency for the application usually simplifies compliance with regulatory requirements. This is why, SMPS operating at high frequency often require some additional EMI filtering components to reduce the effect of harmonics of low order.

### ***Frequency Selection Tip:***

A good compromise for low-power SMPS below 30 watts is to operate around 70kHz. This usually results in a small enough transformer, and makes it easy to manage EMI. Operating above 120 kHz tends to significantly degrade the efficiency and should be avoided. Going below 25 kHz which could become dangerously close to frequencies that humans and pets can hear, should also be avoided.

### ***Gate Drive Resistors***

The internal driver of the IPS1x series has been designed to drive various MOSFETs up to 50 nC of total gate charge. For many small power application, this driver is greatly oversized, the MOSFET is driven too fast and results in an increase of the EMI generated by the switching.

The IPS1x series feature a split gate drive that allows two different resistors to independently control of the rise and fall time of the gate signal. When operating the SMPS in discontinuous mode, there is no current in the primary inductance when the MOSFET turns ON. By slowing down the turn ON time of the MOSFET, it becomes possible to significantly improve the EMI signature without noticeably degrading the efficiency. The turn OFF needs to remain fast but could be somewhat adjusted to obtain an acceptable tradeoff between efficiency and EMI.

### ***Line Overvoltage Resistors***

This feature is available with the IPS1x series where a resistor divider can be placed between the rectified AC line and the ground to drive the overvoltage sensing pin.

The divider should be calculated for this pin voltage to be 3.9 volts when the line voltage reaches the maximum value at which the SMPS is required to operate. Above this voltage, the MOSFET operation is inhibited allowing the SMPS to withstand higher line voltages without being permanently damaged.

Beware of possible non-linearity of SMD resistors when operating them at high voltage. Through-hole resistors might be a better choice. The total impedance should be very high (around 10 MEG) to limit the power dissipation.

### ***Output Filter:***

All SMPS require an output filter and, as a minimum for non-critical applications like replacement of wall-mounted transformers, one capacitor of 470 $\mu$ f or more. Most applications however are more demanding and require a LC or a RC filter to reduce the output ripple below the required level.

With the IPS25, which requires a voltage drop of 1.3 volts across the current-sensing resistor, it is usually possible to meet the ripple requirements by using this resistor as a filter element in the output and to avoid using an output inductor. The same applies to the IPS20/21 when self-powered, but an output inductor will usually be required when operated with an auxiliary supply for maximum efficiency at high output currents.

### ***Value:***

Since the SMPS operates at high frequency, the total output capacitance to be used depends more on the ESR and the manufacturer's rating in terms of maximum current.

### ***Voltage Rating:***

The capacitor should handle the required output voltage and the repetitive spikes generated by ESR at the peak diode current. Again give preference to a low ESR capacitor, especially for low output voltage SMPS.

### ***Temperature Rating:***

The ESR will generate additional power losses which will raise the capacitors temperature well above ambient. 105C rated capacitors are usually the best choice since they usually combine lower ESR with higher temperature rating. Undersizing the output capacitors will result in performance degradation and reliability issues.

**Please refer to the individual data-sheets for more details.**

## OTHER DISCRETE COMPONENTS

### *Fuse or Fusible Resistor, In-Rush Current:*

The fuse is essentially a protection device which should fail safe to prevent a fire hazard. Its voltage rating should correspond to the maximum line voltage the SMPS is rated to operate at. Its current rating should take into account the high peak current which results from the full-wave rectification as discussed in paragraph "Diode Bridge" below. It should be of a type approved by the relevant safety agencies.

In low power SMPS it is not unusual to replace the fuse by a Fusible Resistor that has a dual purpose:

- Limit the in-rush current to comply with some regulations and prolong the life of a line switch if any. (10Ω is the value most often used in 5 to 10W SMPS)
- Perform like a regular fuse should the SMPS fails.

With higher power SMPS, the in-rush current should be limited differently to prevent generating unacceptable losses once the in-rush current is over.

Suitable techniques include:

- Negative Temperature Coefficient Resistors;
- Series resistor shunted by a relay or an active device;

### *Surge Protector (MOV or Others):*

Surge protectors are typically used right after the fuse or the fusible resistor in order to protect the SMPS from damaging transients that could instantly destroy sensitive components like high-voltage MOSFETs.

Most frequently used protectors are Metal Oxide Varistors (MOVs). Other types of protectors appear less popular than MOVs for low-power, low-cost SMPS. They include Gas-Filled Protectors, Precision Spark Gaps and Semiconductor Devices.

The surge protector may be removed from certain low-power applications by taking advantage of the overvoltage sensing circuitry included in the IPS15 line-side controller.

### **X-Capacitor**

This capacitor is usually placed in parallel with the MOV and contributes to the low-pass filter which helps assure conducted EMI compliance.

This safety-agency listed capacitor which is bulky and expensive, may be removed from certain low-power applications by taking advantage of the EMI-reduction

features offered by the IPS15 line-side controller. Please refer to Section "Gate Drive Resistor" page 10.

### *Common-Mode Choke:*

A common-mode choke that is used in conjunction with one X-capacitor is the most common way to comply with conducted EMI regulations. This component may be replaced by a small and low-cost inductor for low-power SMPS using the IPS15. Please also refer to Section "Gate Drive Resistor" page 10.

### *Diode Bridge:*

The diode bridge performs the full-wave rectification of the AC line voltage. Despite the short conduction time of its diodes (on the order of 2mS) which results in high peak currents, a 1A current rating is usually adequate for most SMPS up to 35 watts. The voltage rating should be in accordance with that of the MOV being used.

An additional safety margin should be used when the MOV is omitted. Up to 800 volts for a diode bridge and 100 volts for discrete diodes would provide a cost effective margin of safety.

### *Y-Capacitor:*

The Y-CAP is usually connected between the SMPS primary and secondary to reduce the common mode injection of noise into the secondary which has an adverse effect on loop stability. The IPS1x series combined with a suitable design of the transformer allows the elimination of the Y-capacitor in low-power SMPS.

## Setting Output Voltage and Current Limiting

The IPS 20, IPS21 and IPS25 have been designed for ease of voltage and current adjustments. For these ICs, the output voltage can be adjusted with an external voltage divider that in turn allows remote sensing when desirable.

The maximum output current is set by a resistor across which a voltage of 50mV or 100mV should be developed respectively for the IPS20 and IPS21 and 1.3volts for the IPS25. These ICs feature:

- A maximum noise rejection through current mode drive of the optocoupler,
- Independent voltage and current feedback loop compensations.

Figure 9 - AC IN 85-260V, 0 - 5W OUT SMPS Using IPS15 and IPS25 Controllers:

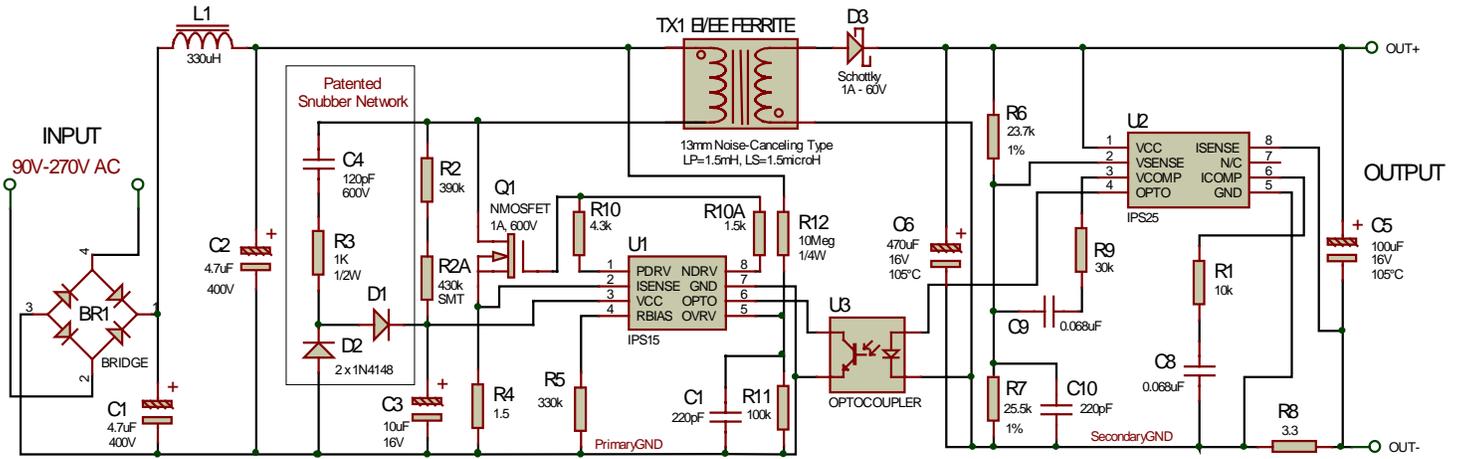
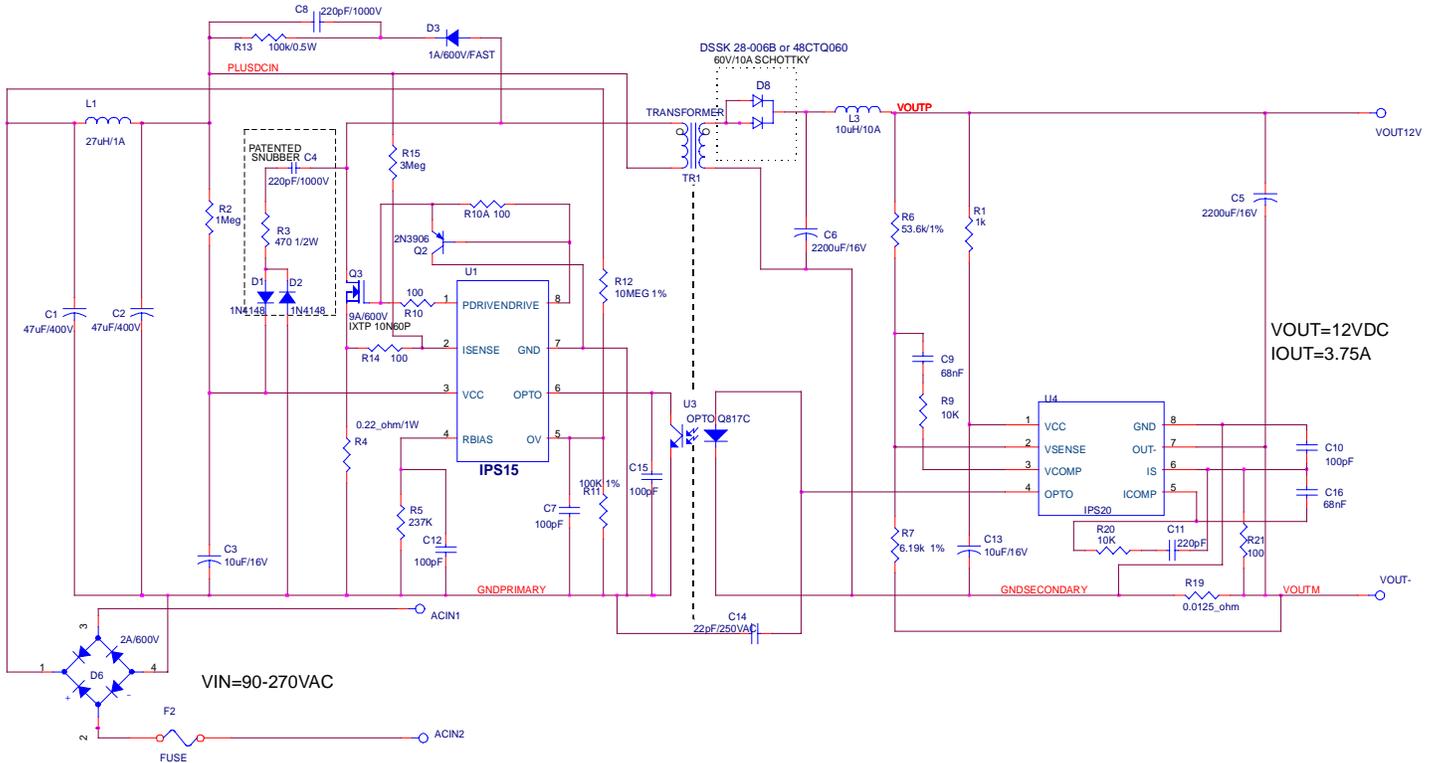


Figure 10: AC IN 85-260V, 45W OUT SMPS Using IPS15 and IPS20 or IP21 Controllers:



**APPENDIX 1 - AAI MODIFIED SNUBBER NETWORK PATENT INFORMATION**



US006233165B1

(12) **United States Patent**  
**Irissou et al.**

(10) **Patent No.:** **US 6,233,165 B1**  
(45) **Date of Patent:** **May 15, 2001**

(54) **POWER CONVERTER HAVING A LOW VOLTAGE REGULATOR POWERED FROM A HIGH VOLTAGE SOURCE**

(75) Inventors: **Pierre R. Irissou**, Sunnyvale; **Hans R. Camenzind**, Los Altos, both of CA (US)

(73) Assignee: **ASIC Advantage, Inc.**, Sunnyvale, CA (US)

(\*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

(21) Appl. No.: **09/570,658**

(22) Filed: **May 15, 2000**

(51) Int. Cl.<sup>7</sup> ..... **H02M 3/335**

(52) U.S. Cl. .... **363/97; 363/21.07**

(58) Field of Search ..... **363/20, 21, 97, 363/131**

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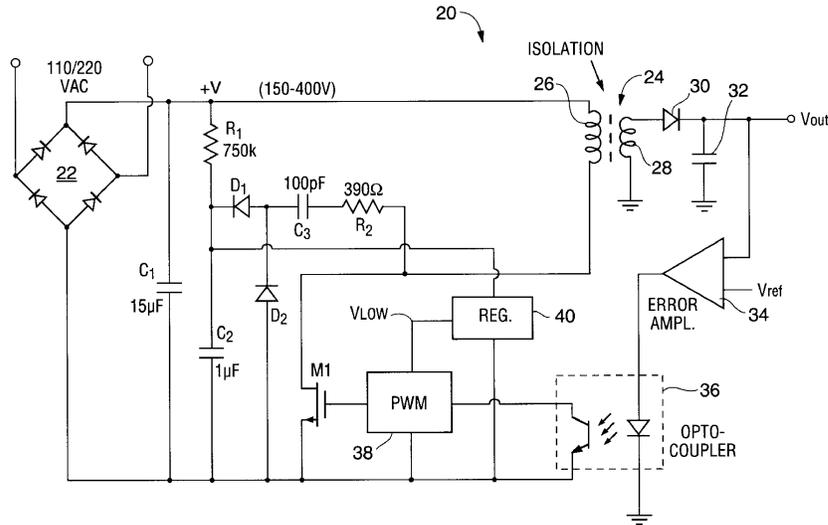
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Primary Examiner—Jeffrey Sterrett  
(74) Attorney, Agent, or Firm—Girard & Equitz LLP

(57) **ABSTRACT**

A power supply arrangement including a transformer and a switching element coupled to a primary winding of the transformer and a power supply. The power supply includes a capacitor having a first terminal coupled to a junction of the primary winding and the switching element and a first diode coupled between the capacitor and an output node. The power supply further includes a discharge element, such as a resistor or a diode, coupled between the junction of the first diode and the capacitor and a power supply common. A voltage regulator, such as a Zener diode, is connected to the output node to produce a regulated voltage powered by the voltage applied to the transformer primary winding, with the regulated voltage being used to power the control circuitry.

**27 Claims, 3 Drawing Sheets**



## APPENDIX 1 - SMPS OVERVIEW AND DEFINITIONS

### *Switch Mode Power Supply (SMPS):*

Switch Mode Power Supply (SMPS) are smaller and more efficient than their linear counterparts. The input is usually the AC mains line and the output is low-voltage DC. Most of the time the output is fully isolated from the mains line as required by applications and users' safety.

The relatively high frequency of operation allows a considerable size reduction of the magnetic components and tank capacitors while achieving high efficiency and low output ripple. These are prerequisites for reducing the size of SMPS.

### *Flyback:*

Flyback converter, sometimes also called "Boost Converter", is one of the oldest SMPS techniques. It is present in all TV sets where it generates several KV of high-voltage required by the CRT. The utilization of the flyback technique to generate isolated low voltages from an AC mains line is somewhat more recent and really took off when high-voltage Power MOSFETs with breakdown voltages of 600 volts and above became commonly available and very affordable.

There is a tendency to use the term "Flyback" when the output is isolated from the input. The isolation from the mains line voltage requires a switching magnetic component with a minimum of two windings which looks like a transformer but does not transfer the energy like one.

The term "Boost Converter" is generally used when the switching magnetic component is a simple inductor that can only generate an output voltage greater than the input voltage and is not electrically isolated from it.

This application note deals with isolated SMPS, and we will therefore exclusively use the term "Flyback".

The flyback approach has a number of unique characteristics which explain its wide acceptance. It has few drawbacks when dealing with low-power applications and can be used for higher power applications in certain conditions.

The first advantage of the flyback technique is the ability of a given design to operate within a wide range of input and output voltages. A factor of 10 is not uncommon, allowing a given SMPS to operate with a good efficiency from 85 to 265 volts AC at its input and still have an output voltage fully adjustable, for example, from 5 to 12 volts.

This type of performance is impossible with a linear power supply and a regular transformer, and difficult if not impossible with other types of SMPS. This

explains why flyback is the solution of choice for "Travel Power" applications.

The second advantage of the flyback technique is its simplicity. Only one high-voltage switching component is required, which cuts on size and cost and explains why Flyback is the solution of choice for minimum cost SMPS.

The first and main disadvantage of the flyback technique is the relatively large amount of electro magnetic interference generated by the intense switching which is taking place involving fast voltage and current transients. For low-power SMPS up to 10 watts, this application note will help the designer manage these undesirable effects by using a combination of inexpensive techniques. Power levels up to 35 watts can be managed with a few more EMI components. Power levels up to 75-100 watts are possible with more EMI filtering, especially when dealing with multiple outputs SMPS. Different techniques are definitely better suited than flyback when trying to achieve higher power levels with a fully EMI-compliant design.

The second disadvantage of the flyback technique has to do with the Power MOSFET and secondary Power Diode which must be sized to handle the peak current. The same components could deliver a lot more power with a different, but not as simple arrangement. This used to be a problem when power components were expensive, but has now become a very minor issue.

### *Flyback "Transformer":*

As indicated before, the switching magnetic component used in an isolated flyback SMPS has at least two windings which makes it look like a transformer. This is why it is often improperly called "Transformer". For convenience we will also call it a transformer even though it does not transfer the energy like a transformer does. With reference to Figure 1, the energy is transferred as follows:

1. When the MOSFET is ON, the rectified AC voltage is applied across the primary winding and the current starts to flow. The current increases linearly with time and the winding magnetically stores an energy equal to  $\frac{1}{2} \times L \times I^2$ . (L is the inductance of the primary winding of transformer TX1, I is the current in the winding and the MOSFET.)
2. When the MOSFET switches OFF, its drain voltage raises above the rectified DC voltage and the core is de-magnetized as current flows through the Secondary Power Diode and charges the output capacitor.

Figures 12 and 13 show the following waveforms: voltage on the MOSFET drain and current in the MOSFET and the primary of the transformer. We will discuss the differences between the two figures in the next paragraph but before we proceed let's emphasize a very important point which governs the selection of both the MOSFET and the Secondary Power Diode. The primary and secondary windings are coupled and there is a transformer effect taking place which cannot be ignored. As indicated before, the transformer effect is not used to transfer the energy from Primary to Secondary but it reflects voltages which have to be taken into account when selecting the breakdown voltage of both the MOSFET and the Secondary Power Diode.

When the MOSFET is ON, the full DC voltage divided by the turn ratio of the transformer appears across the secondary terminals and this voltage, plus the maximum output voltage of the SMPS plus some safety margin defines the minimum reverse breakdown voltage of the Secondary Power Diode. Similarly, when the MOSFET turns OFF, The secondary Power Diode becomes forward biased which applies the output voltage less a diode drop across the secondary winding. This voltage multiplied by the turn ratio between Primary and Secondary adds to the rectified DC and governs the breakdown voltage requirements for the MOSFET.

### *Discontinuous and Continuous Modes of Operation, Loop Stability:*

Looking at the MOSFET drain voltage on Figure 12, one can see three phases in the repeat pattern:

1. The MOSFET is ON and the energy is stored in the inductor;
2. The MOSFET is OFF and the energy is transferred to the load;
3. The MOSFET is still OFF and the inductor being totally demagnetized oscillates freely before a new cycle starts. This is a "dead time" typical of "discontinuous operation.

With the transformer core being fully demagnetized at the end of each cycle, it is possible to change the energy stored from Zero to Maximum within one cycle resulting in a very fast transient response.

In Figure 13, there is no longer any dead time. The very first cycle is the only one to start with a fully demagnetized transformer core resulting in a current increasing linearly from zero like before. In continuous mode however, a new cycle starts before the core is fully demagnetized resulting in not "dead time" in the cycle and a trapezoidal shape for the current waveform. Please note that the current waveform builds up progressively taking several cycles to reach a steady state. The energy transfer cannot therefore vary as quickly as before making the loop response of

the discontinuous mode much longer than that of the continuous one.

A complete SMPS resembles a servo-mechanism in terms of stability and must be designed to be perfectly stable over the entire range of input line voltages and loading conditions. At no load all flyback power supplies operate in discontinuous mode with a fast response but may enter continuous mode when operating with minimum line voltage and heavy loads. Such a situation should be avoided since it is almost impossible to compensate the loop to be stable for both discontinuous and continuous modes of operation. The current-limiting loop is usually somewhat less critical and could be slow enough to tolerate a certain amount of continuous mode as the SMPS operates in an overload or short-circuit condition.

Flyback SMPS can be designed to operate in discontinuous mode only where the energy transfer is higher with identical components. But this is beyond the scope of this application note.

Besides the loop stability issues, operating in discontinuous mode has three major advantages:

1. The transformer core being fully demagnetized at the end of each cycle, the MOSFET turns ON under very favorable conditions: there is no inductive current and only a small current spike caused by the stray capacitances. This results in very small switching losses at turn ON.
2. Since the MOSFET turn ON is not critical, it can be made relatively slow in order to reduce the EMI signature and save on EMI-reduction components. This is very easy to achieve with the IPS15 line-side PWM controller.
3. Since the transformer core is fully demagnetized when the MOSFET turns ON for a new cycle, there is no current flowing in the Secondary Power Diode which makes its reverse recovery time a non-critical parameter. In continuous mode however, an Ultra-Fast Diode or, voltage permitting, a Schottky Diode must be used to minimize the switching losses as the MOSFET turns ON when there is still current flowing in the diode.

APPENDIX 2 - VOLTAGE AND CURRENT MOSFET WAVEFORMS

Figure 12 - DISCONTINUOUS MODE:

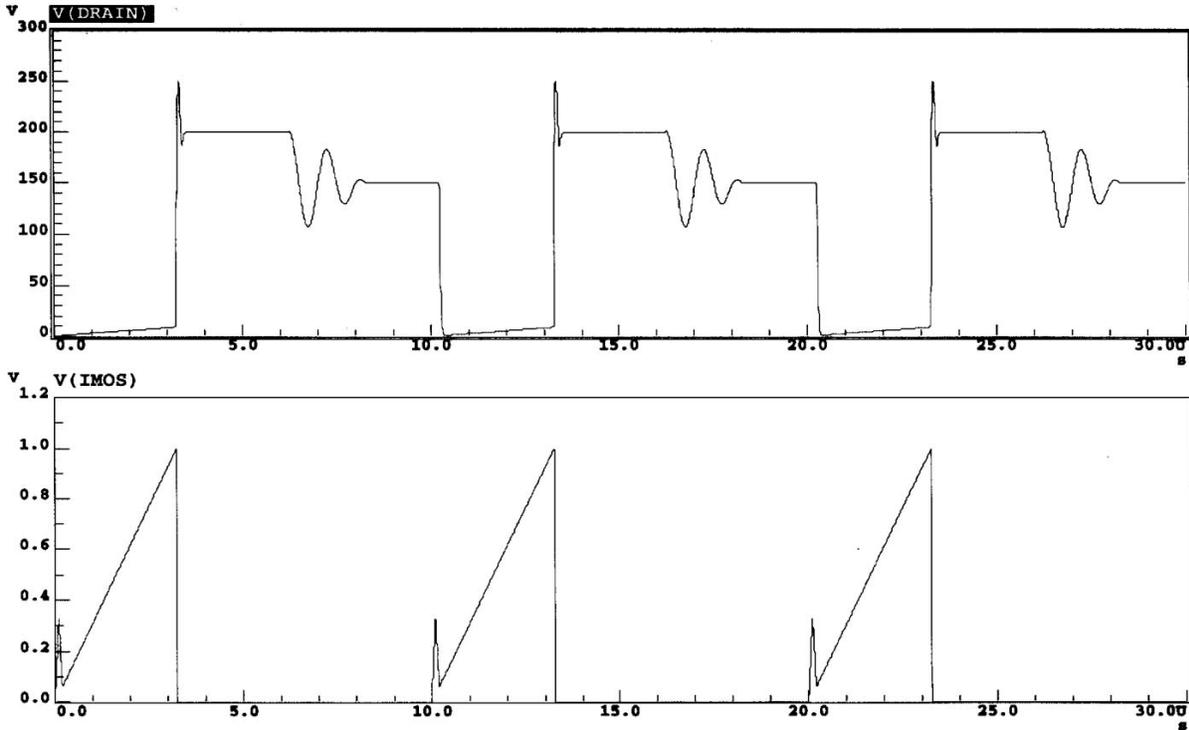
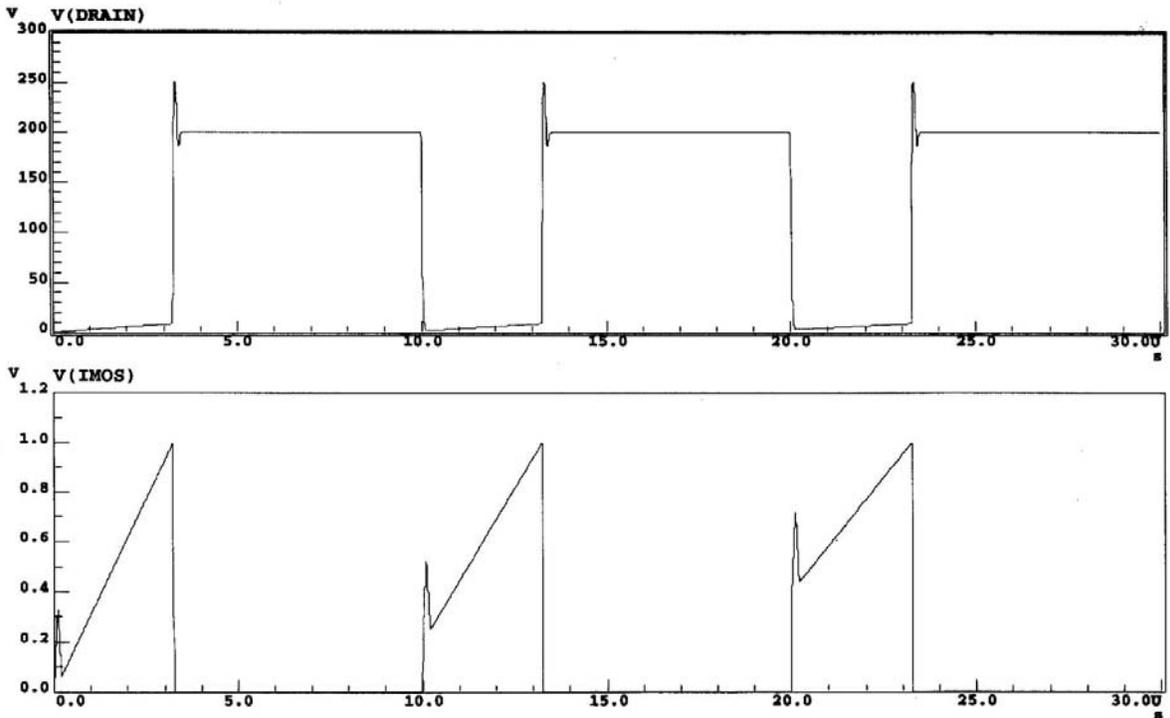


Figure 13 - CONTINUOUS MODE:



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