

## INTRODUCTION

This application note, AN-IPS-01, was created to complement the IN-PLUG<sup>®</sup> 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.

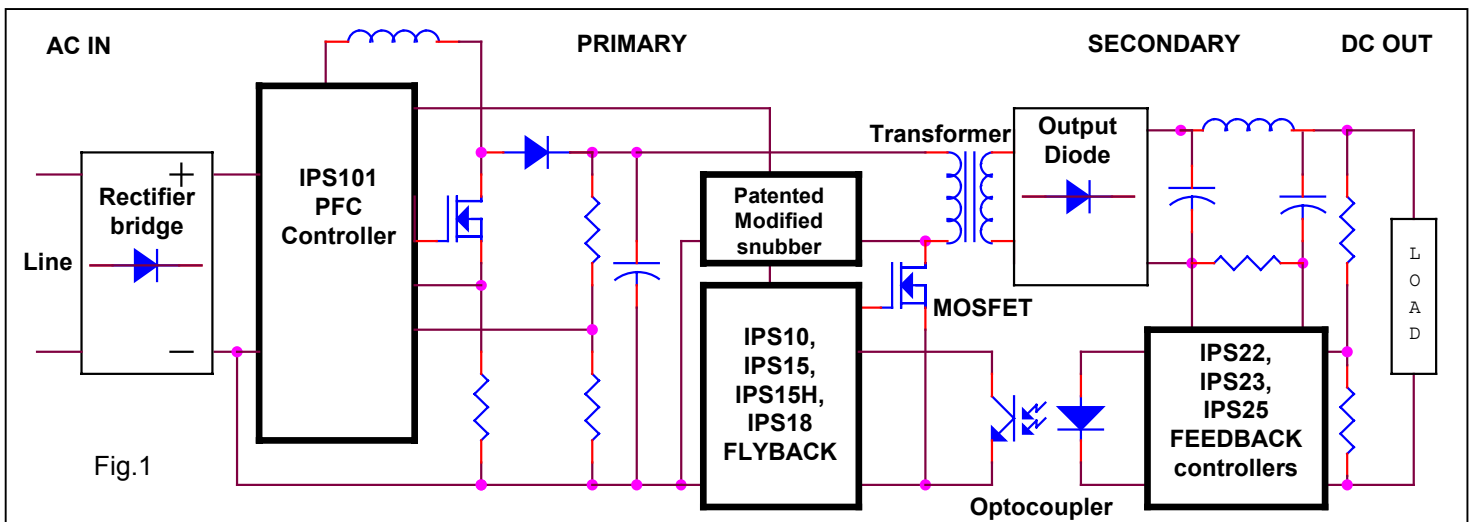
When designing an SMPS, please check which governmental regulations your product will have to comply with, such as the following:

- 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<sup>®</sup> 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.



## SELECTION OF FLYBACK AND FEEDBACK CONTROLLERS

Please, see Figures 9, 10 and 11 pages 12, 13 & 14 for detailed application schematics.

### Line Side Flyback Controllers:

- For applications where EMI and line over-voltage requirements are less stringent:  
→ select **IPS10**
- For applications with highest EMI and line over-voltage requirements:  
→ select **IPS15**

Note: In low-power applications, the IPS15 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, IPS22 or IPS23. 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 1A and above,  
→ select **IPS22 or IPS23**.
- If the output current is between 700mA and 1A, and if efficiency is critical,  
→ select **IPS22 or IPS23**.
- If the expected output current is between 700mA and 1A, and if efficiency is non-critical,  
→ select **IPS25**.

Note:

- Both the IPS22 and IPS23 integrate the same exact basic functions with one difference: the IPS23 has its VCC protected by a built-in 7.2V zener diode which can eliminate an external standard clamping zener when an auxiliary supply is used.
- 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 IPS22 or the IPS23 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.

(A drop of 250mV only is required across the current-sensing resistor.)

- All of AAI's standard feedback controller ICs have been designed for low cost, flexibility and performance requiring a minimum of external components. Their response can be tailored to match the most severe requirements in terms of regulation, transient response and loop stability.

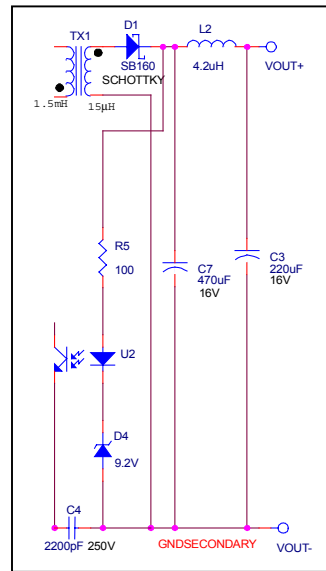


Fig. 2

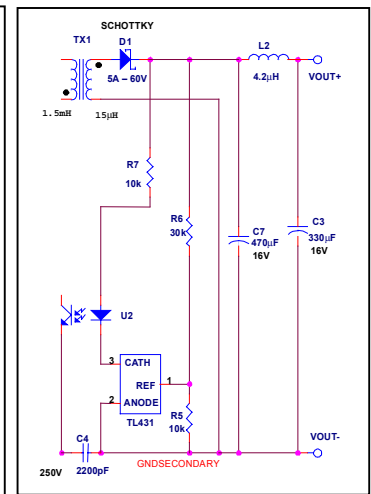


Fig. 3

## 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. IPS10 and IPS15 are 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.

## 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.

### **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.

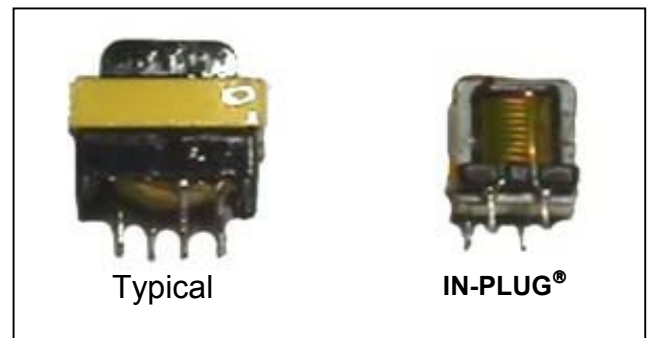
The energy is transferred sequentially from the line side to the load side while ensuring proper isolation from dangerous line voltages. However, despite the fact that it operates like an inductor, the bi-directional magnetic coupling between primary and secondary reflects voltages and spikes that must 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 tend 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  $6\mu\text{s}$  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:

$L_p$  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  $R_{on}$ ,
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.

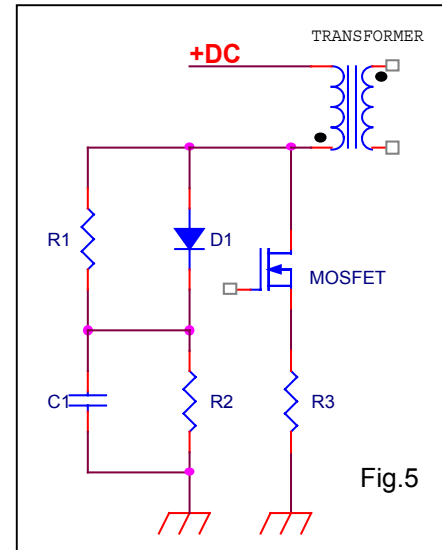
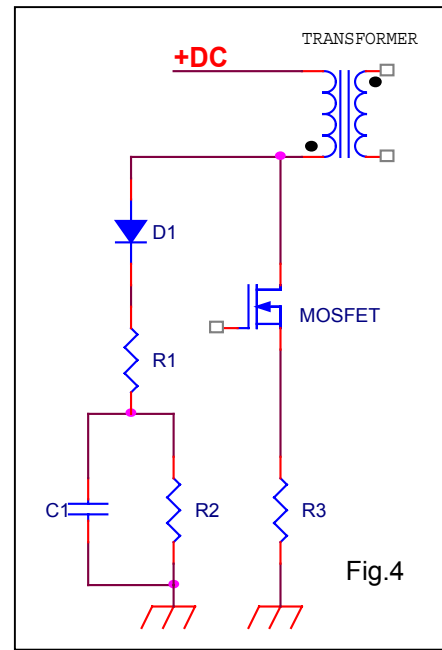
### Typical Snubber Network:

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.



### 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.

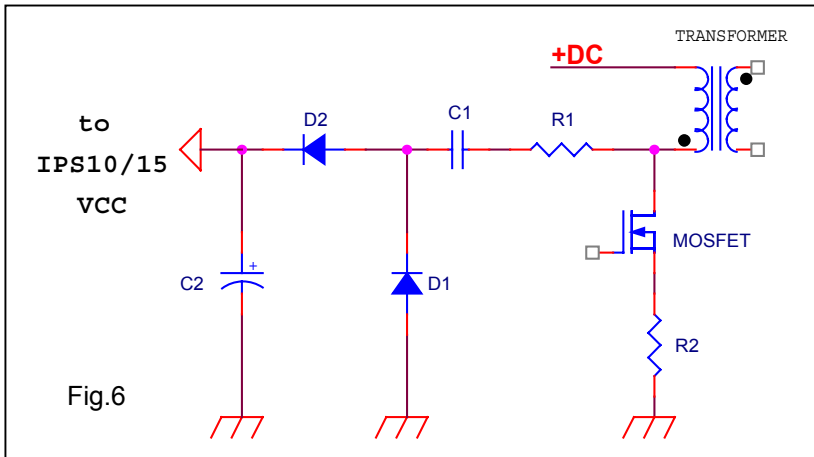
As before, when the MOSFET turns OFF, C1 charges through R1 to slow-down the transition; but C2, which is part of the Line-side controller IC circuitry, is also charged through D2. When the MOSFET turns ON, D1 becomes forward biased and C1 discharges to be ready for a new cycle.

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:**

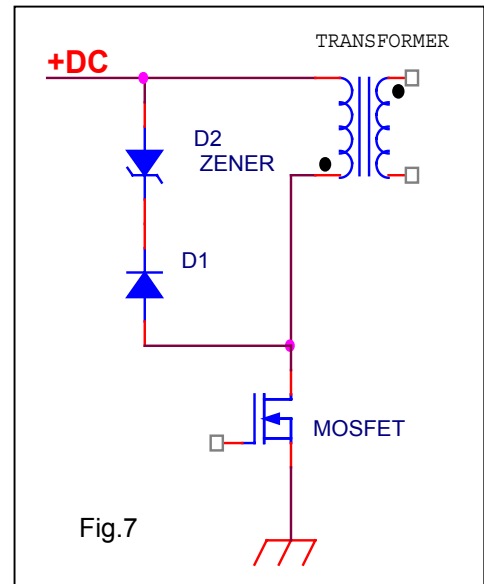
AAI's policy is to give a royalty-free, non-exclusive license to use the above-referenced patent to every user of IPS products who purchase their ICs either directly from AAI or through authorized distributors / representatives.



**DRAIN CLAMP**

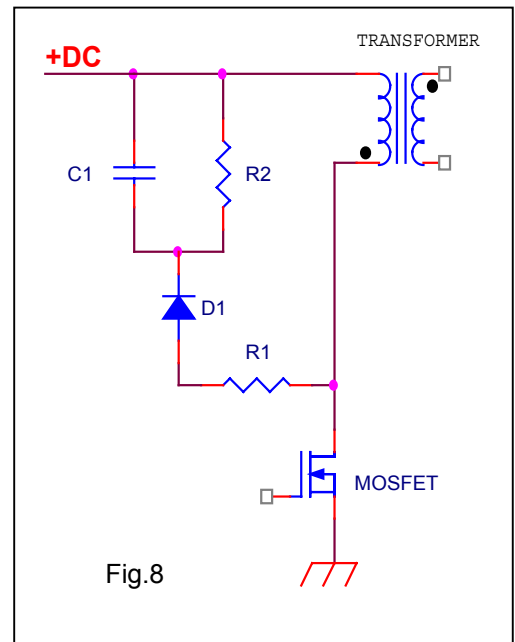
By slowing down the transitions and creating some controlled losses, the snubber network tends to reduce, at the MOSFET drain, the voltage spike created by the transformer leakage inductance. For high power applications, however, and especially when using a 600 volt MOSFET in international applications, 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 Figure7 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.





## **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 IPS10 and IPS15 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.

With the IPS 10, a resistor can be inserted between the gate drive pin and the actual MOSFET gate. This results in slowing down both the turn ON and the turn OFF of the MOSFET, which in turn improves the EMI but is not optimum for efficiency.

A better solution is to use the IPS15 which features 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 only available with the IPS 15 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 470µf 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 IPS22/23 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.

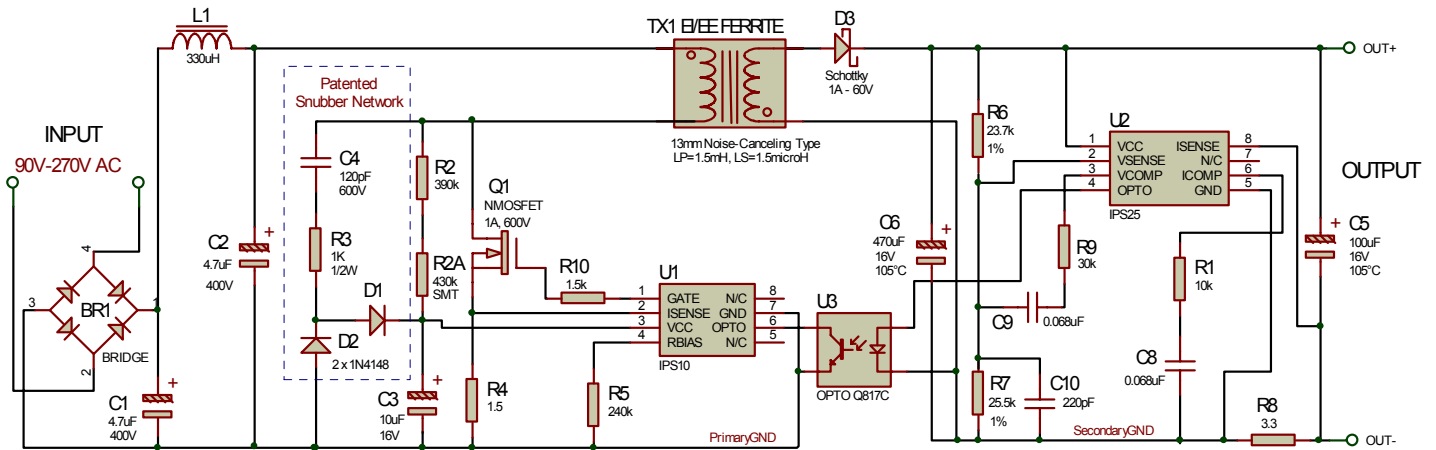
## **Temperature Rating:**

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.

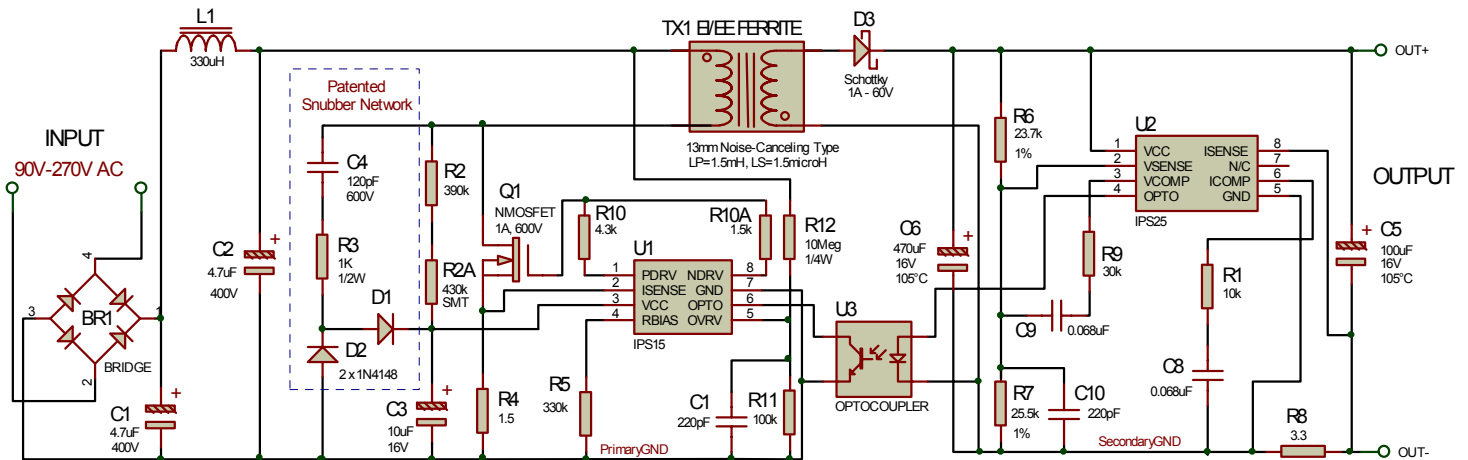
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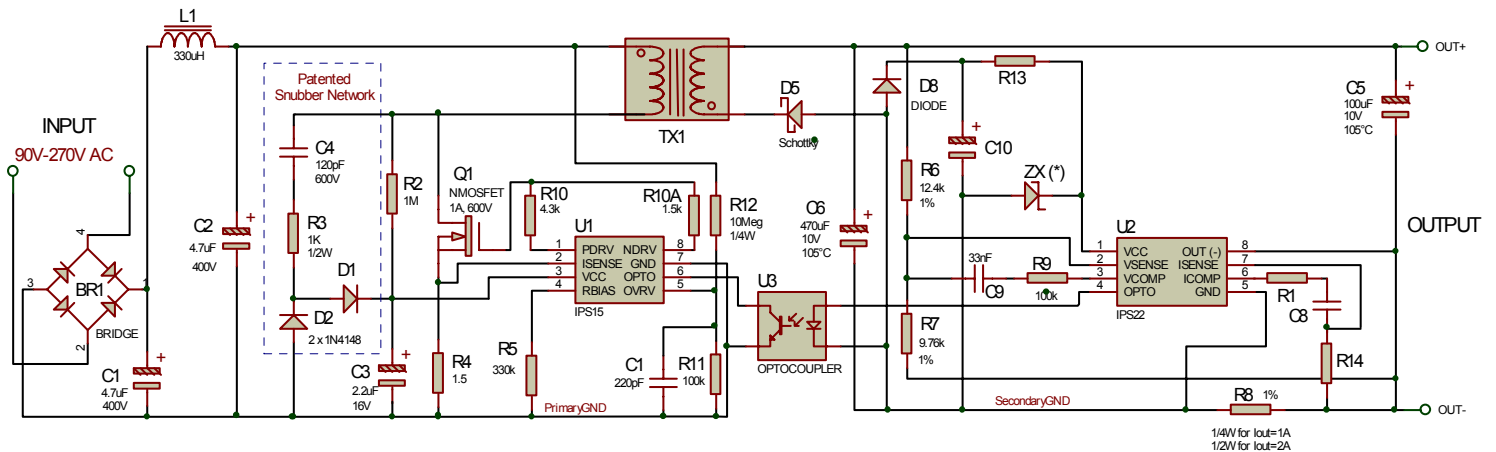
**Figure 9 - AC IN 85-260V, 0 - 5W OUT SMPS Using IPS10 and IPS25 Controllers:**



**Figure 10 - AC IN 85-260V, 0 - 5W OUT SMPS Using IPS15 and IPS25 Controllers:**



**Figure 11: AC IN 85-260V, 0 - 15W OUT SMPS Using IPS15 and IPS22 or IPS23 Controllers:**



1/4W for Iout=1A  
1/2W for Iout=2A

# 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**

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

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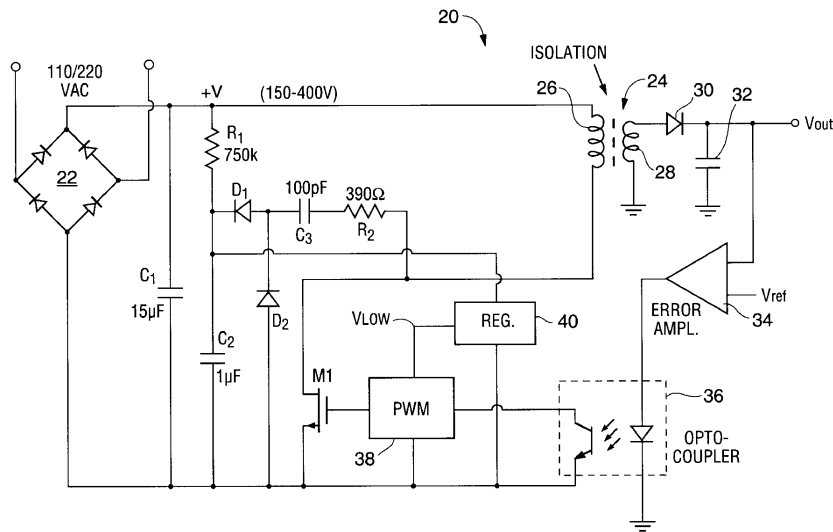
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(57) **ABSTRACT**

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 2 - 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  $1/2 \times L \times I^2$ . (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:

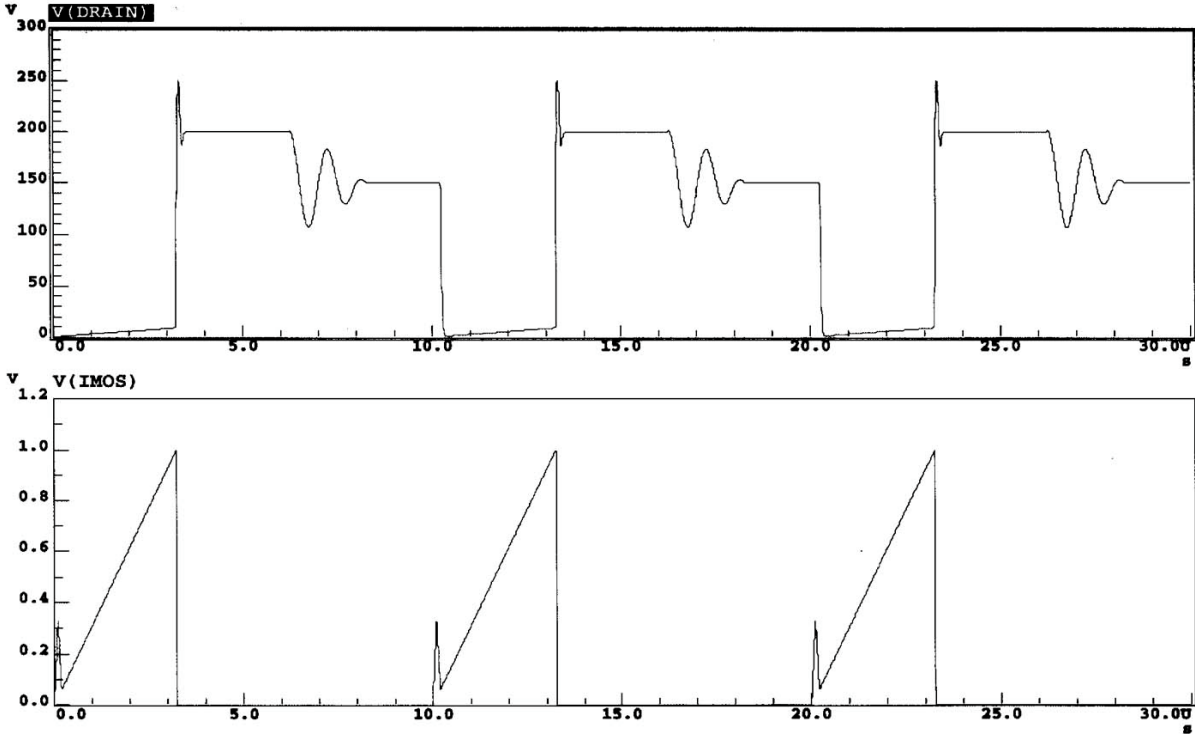
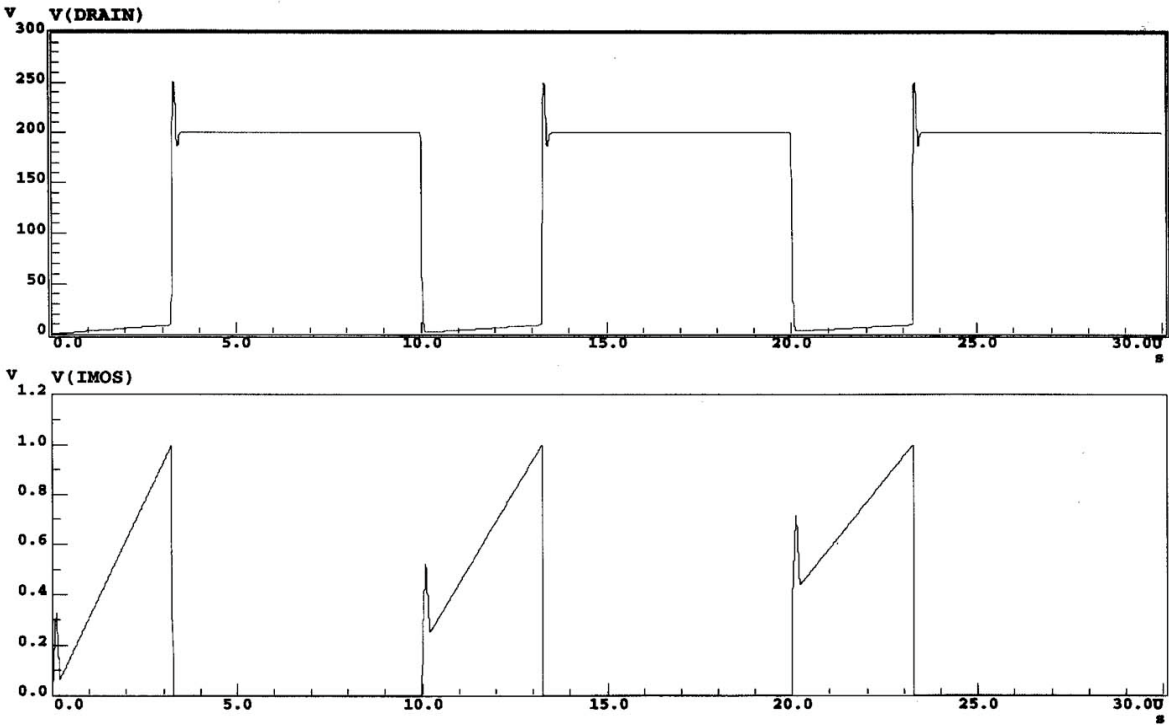


Figure 13 - CONTINUOUS MODE:





**Introduction:**

Complying with EMI regulations in terms of Conducted and Radiated Emissions, and Susceptibility to RF signals is now a requirement in most countries. Any new Switch Mode Power Supply (SMPS) design should be undertaken with full compliance in mind.

Because the reduction in physical sizes of both components and complete SMPS assemblies have greatly reduced the problems associated with Radiated emissions and Susceptibility to RF signals, the real challenge remains to meet the Conducted Emissions requirements. This can be achieved through "brute force" techniques or by using the "distributed" solutions as described below.

**Brute Force Solutions:**

Brute force solutions are the application of specific EMI reducing components to an SMPS that otherwise might not be required in the basic SMPS design. This is especially necessary for high-power SMPS. Brute Force components include:

- Common-mode line chokes and capacitors configured as low-pass filters whose purpose is to comply in the low-frequency part of the spectrum;
- Ferrite beads, RF inductors and capacitors to improve the compliance at high frequency;
- Clamps and dissipative networks whose purpose is to attenuate ringing and parasitic oscillations associated with the switching transformer, the high-speed diodes, etc;
- Ground planes, internal and external shields, feed-through capacitors and other shielding techniques.

**Distributed Solutions:**

Distributed Solutions are the optimization of standard SMPS building blocks (components and circuits) in reducing EMI without resorting to otherwise unnecessary "brute force" solutions. These distributed effects are exceptionally efficient in low-power SMPS applications as it is very possible to reduce EMI to adequate levels with a minimum or with no brute force component.

AAI has analyzed and presented below the contribution of every single critical component and operating condition of the SMPS that can affect conducted EMI signature. Also AAI proposes recipes to eliminate the application of large and expensive EMI-reducing components.

**Frequency of operation:      *Fixed frequency often leads to a lower SMPS optimization.***

The low end of the spectrum, which starts at 100kHz or 400kHz often creates the most stringent compliance problems. Since we are dealing with energy in fundamental and harmonics, the higher the harmonic rank, the lower the energy and the easier the compliance.

Depending on the frequency of operation, 100kHz could either be the fundamental, the second, the third or even the fourth harmonic and this is why the First Rule is to operate at the minimum switching frequency acceptable for the application. Unlike common assumptions, for a low-power SMPS delivering 3 to 5 Watts, operating at lower frequencies will not, by itself, result in larger transformers and filtering capacitors. The size of transformers and filtering capacitors are determined by other factors such as safety compliance for transformer and series resistance for capacitor.

Most integrated line-side controllers available on the market operate at an internally-defined frequency in order to save one pin and one external resistor.

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The problem is that the users of these integrated controllers who are now forced to optimize all other components on the board around a frequency that has been imposed by the controller itself but not necessarily optimized for the SMPS itself.

AAI's approach is to make the frequency of the SMPS operation externally adjustable. With all of AAI's IN-PLUG<sup>®</sup> Controllers, the frequency of operation can be selected anywhere from tens of kHz to well above 300kHz.

**Mode of operation:                      *Discontinuous mode operation brings much less noise.***

Flyback SMPS can operate in either "Discontinuous" or "Continuous" mode. The Continuous mode allows the SMPS to deliver more power than when operating in Discontinuous mode. But Continuous mode is noisier. When designing a Low-Power SMPS where the MOSFET is typically oversized for the application, AAI's recommendation is to operate in Discontinuous mode.

Flyback switching generates tremendous  $dV/dT$  and  $dI/dT$  and results in severe EMI problems. In Continuous mode, tremendous  $dI/dT$  occurs involving the transformer and the power diodes. This creates all sorts of ringing and forces the addition of several EMI-reducing components to comply with existing regulations. But in Discontinuous mode the MOSFET turns ON at zero current which substantially reduces the EMI signature without the addition of EMI-reducing components. Therefore, operating in Discontinuous mode improves the efficiency of low-power SMPS and eases-up the constraints related to the switching transformer.

**MOSFET Switch selection:                      *Over-rated MOSFETs mean more EMI.***

The smallest size MOSFET available from most suppliers is a 1A device, while a 250mA device, would be sufficient for 5W international applications. This means that MOSFETs are often over-rated for their application. Less efficient MOSFETs might be used with insignificant loss in performance while at the same time offering a dramatic reduction in the conducted EMI signature.

**MOSFET Gate Drive:                      *Independent gate turn-on and turn-off drive helps efficiently control a wider range of MOSFETs.***

Operating in Discontinuous mode removes the  $dI/dT$  when the MOSFET turns ON. In fact, there is no current in the inductor, which is the main source of noise in Continuous mode. The MOSFET, however, exhibits some current peaks associated with the stray capacitance in the transformer and the MOSFET itself.

This current peaking itself is a significant source of EMI that can be considerably improved by slowing down the turn ON of the MOSFET. Since there is no inductive current in Discontinuous mode when the MOSFET turns ON, it is possible to slow down its turn ON with little or no degradation of the efficiency.

But when the MOSFET turns OFF, the switching involves both  $dI/dT$  and  $dV/dT$ , and slowing down the turn OFF tends to create additional switching losses in the MOSFET. Because of these switching losses, slowing down the turn-OFF does not appear to be a good idea. The reality is that in low power SMPS, the minimum size MOSFET (1 watt or more) is usually over-rated and therefore can handle a little more dissipation without affecting the reliability of its operation. Slowing down the MOSFET complements the effect of the snubber network. While the effect on EMI is not quite the same, for equal losses, slowing down the MOSFET has more impact on the EMI signature than the snubber network.

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But because the slower switching MOSFET and the snubber network each affects different portions of the waveform, in combination, their EMI reducing effect is compounded.

The IPS15 line-side controller has been especially designed to offer the maximum flexibility for driving a wide range of MOSFETs with gate charges up to 50nC. This controller provides independent source and sink drive and allows the designer to independently optimize the turn ON and the turn OFF through the utilization of two external resistors. Only one resistor is required for turn ON adjustment. For applications requiring the fastest possible turn ON and turn OFF, no external resistors are necessary.

**Snubber Network:**      ***AAI's snubber network saves the clamp circuitry and eliminates the need for a transformer's third winding.***

The snubber network used to be mandatory for the reliable operation of SMPS when high-voltage Bipolar transistors were the only power switches available. The advent of MOSFETs changed the situation because they are so fast and their intrinsic switching losses are so small that they can usually switch large inductive loads without exceeding their SOA ratings. But not having a snubber network can result in excessive voltage spikes that may destroy the MOSFET by exceeding its avalanche rating. Therefore some form of drain clamp is required for the SMPS to reliably operate.

Adding a clamp will only protect the MOSFET, but will not improve the EMI signature because of the MOSFETs extremely fast switching. EMI can only be improved by incorporating a snubber network.

The following arrangements present various combinations suitable for different applications:

- No snubber and/or clamp are necessary for very low power SMPS of the order of 1 watt, where slowing down the MOSFET is sufficient to reduce EMI,
- A snubber alone in combination with a MOSFET gate-drive control is required for moderately low power SMPS of between 1 and 5 watts or more,
- Both the snubber and clamp are required with or without gate-drive control for higher power SMPS.

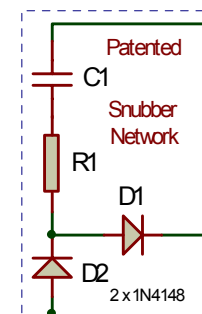
With the snubber playing such an important role in terms of both the SMPS's reliability of the operation and EMI signature, AAI has developed and patented a more efficient and more utilitarian snubber that:

- slows down the  $dV/dT$ :
- improves the EMI signature,
- reduces the voltage spikes and prevent MOSFET avalanche for low power applications,
- recycles some of the energy otherwise lost in the snubber network in order to power the line-side controller.

When combined with the precision cycle-to-cycle current limiting feature of the IN-PLUG<sup>®</sup> line-side controllers, AAI's snubber does not require the addition of a clamp circuit to ensure that the voltage and avalanche ratings of the MOSFET are not exceeded during both normal and abnormal conditions.

And by recycling the energy otherwise lost in the switching of the MOSFET to power the line side controller, AAI's snubber network eliminates the need for the transformer's third winding.

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AAI patented modified snubber network.

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Two-winding transformers can more readily comply with safety agencies' requirements than three-winding transformers. Two-winding transformers can also be designed to have less distributed capacitance, which in turn results in less EMI and less switching losses. And two-winding transformers can be designed to provide a "noise-canceling" effect that results in not only more EMI signature reduction, but reduced touch-current and the elimination of the Y-capacitors.

**Noise-Canceling Transformers: *Innovative noise-canceling solution for miniature two-winding transformers in the range 1 to 30 Watt.***

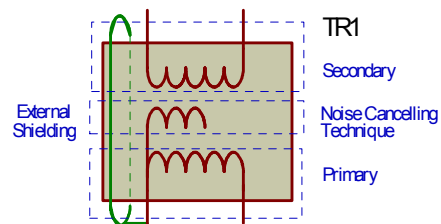
Conventional noise-canceling techniques can only be used to design large-size switching transformer and achieve:

- low distributed capacitance,
- enhanced common-mode noise rejection (one or more internal shield(s) involved),
- reduced emitted radiation and RF susceptibility (one or more external shield(s) necessary).

Yet there is a need for miniature low-noise transformers to help further reduce the EMI signature, these basic techniques are completely impractical because distributed capacitance would become enormous creating large power losses with EMI problems, and safety-agency requirements would not be possible to meet.

Given that AAI's modified snubber network eliminates the need for the transformers third-winding (also called "bias winding"), AAI has developed a noise-canceling solution for miniature two-winding transformers in the range 1 to 30 Watts that includes:

- minimization of the primary distributed capacitance (Efficiency and EMI),
- minimization of the primary to secondary capacitance (EMI, Touch-current, Y-capacitor, Loop stability),
- reduction of the primary to secondary common-mode noise injection through a combination of partial shielding and neutralization techniques (EMI and Loop stability),
- reduction of EMI signature through a partial external shielding, without increasing the distributed capacitance (Conducted and radiated EMI),
- minimization of the leakage inductance which is the main source of EMI and losses in flyback designs,
- compliance with safety agency requirements.



2-winding transformer built with noise-canceling techniques.

**Efficiency of Noise-Canceling Transformer Techniques:**

The 2 pictures below show the efficiency of AAI's Noise-Canceling Transformer techniques. Conducted EMI scans were done with the same basic core and size transformer on a 5W SMPS, one being regular (Figure 1A), one being modified with noise canceling techniques (Figure 1B). Designer can expect an EMI attenuation of more than 20dB. The center horizontal line represents the maximum EMI FCCB authorized level.

Working conditions are : 120V AC Input, 5V 700mA Output.

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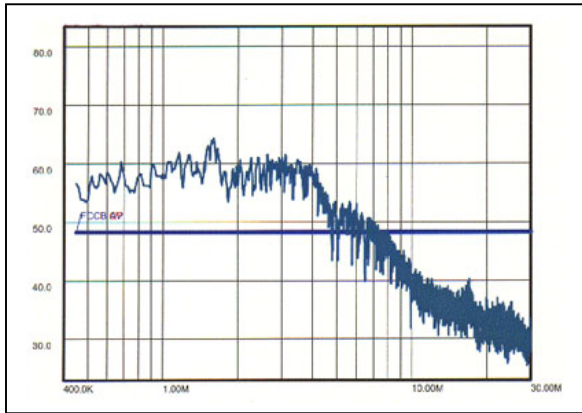


Figure 1A

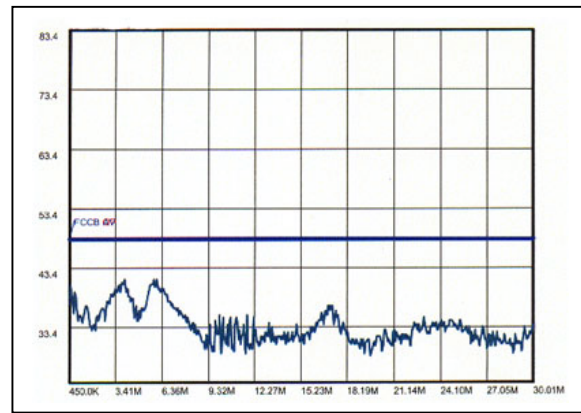


Figure 1B

**Start-up resistor:**                      ***A way to efficiently damp free oscillations.***

The start-up resistor required by many integrated controllers and all discrete implementations, can by itself reduce the EMI signature. Since all designs require a start-up current generated internally by a high-voltage current source or externally by a resistor, both of which burn current, one can take advantage of the burned energy to create additional damping of the drain waveform in an hard-to-control area.

This is achieved by returning the start-up resistor to the drain and not to the rectified DC voltage. This damping of the waveform is especially effective when the third winding has been eliminated through the utilization of the AAI's snubber network and the transformer has been of noise-canceling type (described below). In this case, the start-up resistor will considerably dampen the free oscillations that occur when all the energy stored in the inductance has been transferred to the secondary while at the same time helps improve the efficiency of the SMPS.

**Input filter:**                              ***Almost Free.***

It is very unusual to full comply with EMI requirements without incorporating some kind of input filter creating a low-pass effect. This is typically accomplished by the addition of one common mode choke and two capacitors that are sometimes replaced or complemented by two separate inductors and two capacitors.

When using all of the distributed EMI-management techniques described above, a much simpler and almost free input filter is usually sufficient to be fully compliant.

It consists in splitting the input electrolytic capacitor required to filter the rectified AC into two, and adding one low-cost inductor or in some cases, just a resistor.

Using two smaller capacitors instead of one has little or no cost impact and often results in a slimmer design. Depending on the ESR specifics, a lower temperature rating could also sometimes be used resulting in additional cost savings.

**PCB layout:**                              ***Tips to good Primary and Secondary PCB layout designs.***

It is important not to forget that there is no substitute for a good PCB layout. The primary circuitry that involves tremendous  $dV/dT$  and  $dI/dT$  which resembles lightning, should be handled carefully. Compacted layout, minimized loops and ground planes should be preferred everywhere possible.

The secondary circuitry that involves high currents and very large  $dI/dt$  should be also handled carefully. First, the loop between the transformer's secondary, the power diode and the first output capacitor should be as short as possible and as planar as possible. Since a small loop area is more important than the current density in the PCB copper, oversized traces on the PCB should be avoided as it may adversely affect both the EMI and the output ripple. Second, the transformer's exit terminals should be connected to the diode and the capacitor return with flying twisted wires. Ultimately a surface-mounted power diode should be preferred to a leaded one.

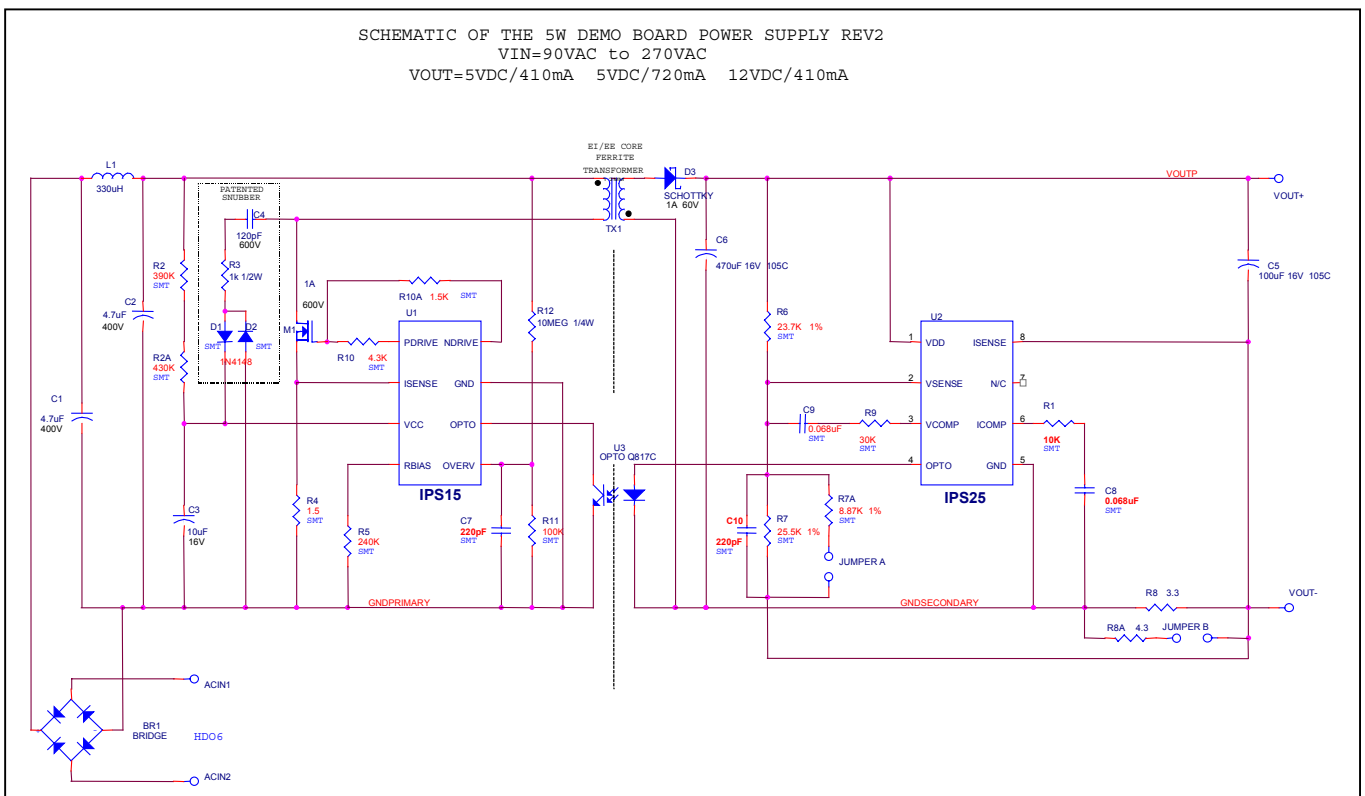
**Loop stability:** *No compromising on loop-stability or performance in combining AAI IN-PLUG® line-side and load-side controllers.*

The race for smaller and cheaper low-power SMPSs has generated a large number of designs and ICs which compromise on loop stability and regulation performance. AAI's IN-PLUG® line-side and load-side controllers can be combined to avoid any compromising. For simple SMPS applications, simple feedback techniques used in conjunction with our IPS10 line-side series help achieve very simple and inexpensive solutions that ensure excellent loop stability and prevent cycle-skipping which is always a source of problems.

For more sophisticated applications the IPS20 load-side series can provide superb voltage regulation, current limiting, transient response and low output ripple with a minimum of external components.

**Example of Distributed EMI-reduction design.**

Below is an SMPS schematic of a 5 Watts evaluation board which was designed by AAI in respect to the EMI-reduction distributed techniques expressed in this document. This board is available for evaluation with extra parts and documentation.



## How to verify the suitability of a specific flyback transformer for a given SMPS application.

### INTRODUCTION

This application note AN-IPS-03 was created to complement the IN-PLUG® application notes AN-IPS-01 & 02 and help SMPS designers ensure that a specific transformer planned to be used in a given application, will be suitable to meet the target features and will correctly operate in Discontinuous mode.

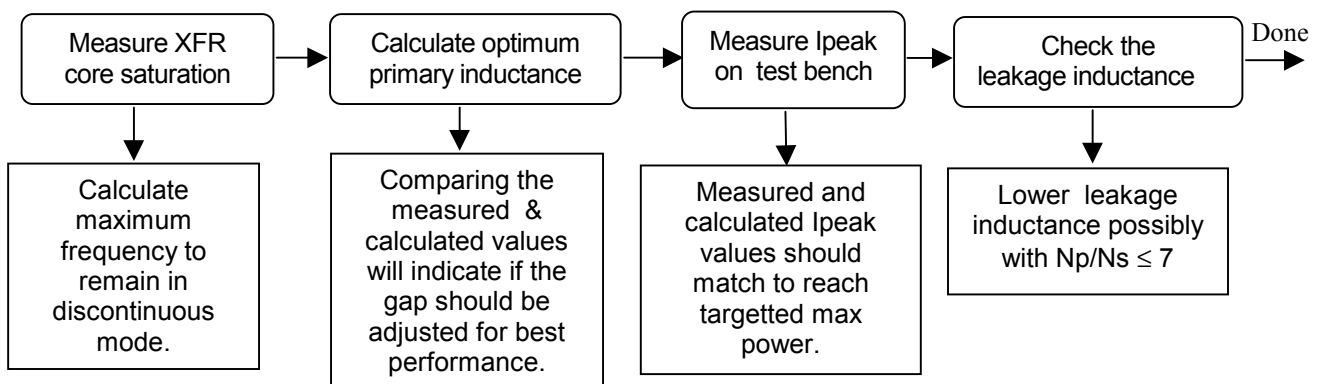
### FLYBACK SMPS MODE OF OPERATION

Flyback SMPS can operate in either "Discontinuous" or "Continuous" mode. The Continuous mode allows the SMPS to deliver more power than when operating in Discontinuous mode. But Continuous mode is noisier. When designing a Low-Power SMPS where the MOSFET is typically oversized for the application, AAI's recommendation is to operate in Discontinuous mode.

Flyback switching generates tremendous  $dV/dT$  and  $dI/dT$  and results in severe EMI problems. In Continuous mode, tremendous  $dI/dT$  occurs involving the transformer and the power diodes. This creates all sorts of ringing and forces the addition of several EMI-reducing components to comply with existing regulations. But in Discontinuous mode the MOSFET turns ON at zero current which substantially reduces the EMI signature without the addition of EMI-reducing components. Therefore, operating in Discontinuous mode improves the efficiency of low-power SMPS and eases-up the constraints related to the switching transformer. In addition, the loop stability is much easier to achieve in this mode.

To ensure that the designed transformer will correctly operate in Discontinuous mode, the user will have to verify the parameters below while following the indicated steps:

- Transformer turn-ratio:  $N = N_p / N_s$ ,
- Transformer primary inductance  $L_p$ ,
- Transformer gap,
- Transformer leakage inductance,
- Primary Power  $P_{in} = 1/2 \times L_p \times I_{peak} \times F$ ,
- Frequency of operation  $F$ .



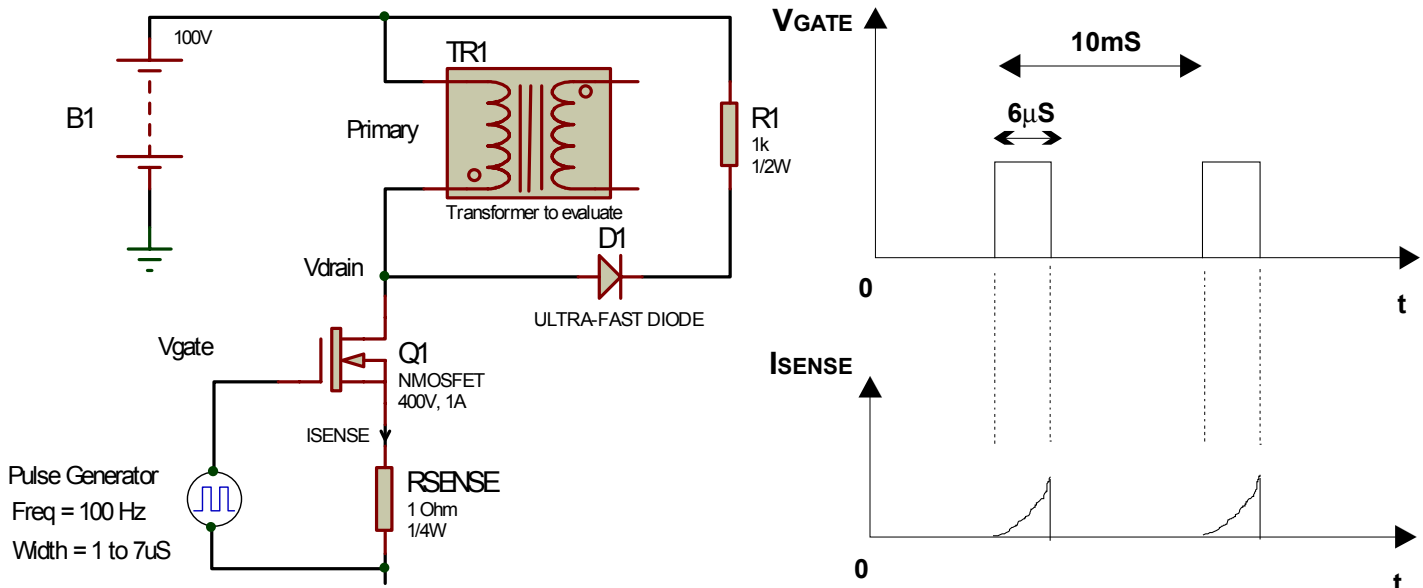
### HOW TO FIND THE LIMIT BETWEEN CONTINUOUS AND DISCONTINUOUS MODE

When the transformer is already calculated and built, the simplest way to verify its suitability for a given SMPS application, is to start with a lab bench measurement to determine the "Tsat" parameter as described below. "Tsat" is related to the core saturation region of the transformer.

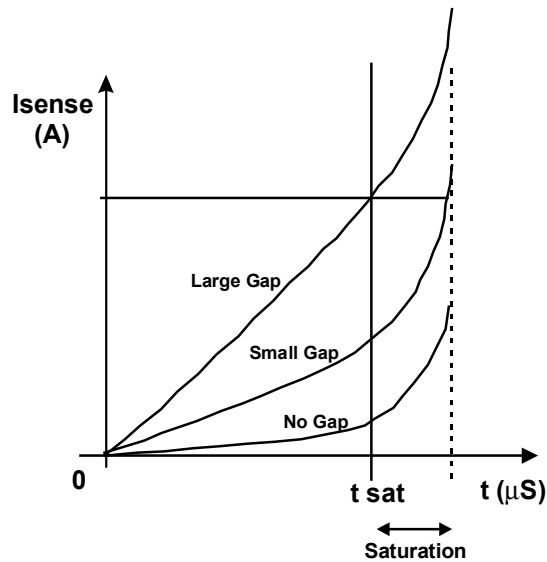
**Note:** The saturation of a ferrite core doesn't depend on the core gap.

## MAXIMUM TIME "Tsat" BEFORE CORE SATURATION

Referring to the bench test diagram below, at 100V supply, the saturation region can be found by measuring how long it takes to reach **Tsat**. No core gap could be used if practical, or simply the final gapped transformer if more suitable. The operation must be done at hot temperature to account for the ferrite performance degradation. This max time of T<sub>sat</sub> will help then determine the maximum frequency of operation.



Details of I<sub>sense</sub>



## MAXIMUM FREQUENCY OF OPERATION

In the SMPS application and assuming  $(V_{out} + 1) \times N_p/N_s = 100V$ , the maximum operating frequency to remain in discontinuous mode is  $1 / (2 \times T_{sat}) = F$ .

Where:  $N_p$  = primary turns,  
 $N_s$  = secondary turns,  
 $V_{out}$  = output voltage provided by the SMPS.



Note:

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.

### OPTIMUM PRIMARY INDUCTANCE

The optimum primary inductance value can be calculated by applying  $L_p = 800 / (P_{in} \times F)$ . For a desired output power and in absence of better information, an efficiency of 75% can be used ( $P_{in} = P_{out} / 0.75$ )

Where:  $L_p$  = primary inductance,  
 $P_{in}$  = primary power ( $P_{in} = 1/2 L_p I_{peak} F$ )  
 $F$  = frequency of operation.  
 $I_{peak}$  = peak current in the MOSFET

For more details, see section “Primary Inductance” page 6 of application note AN- IPS-01.

### GAP ADJUSTMENT

If the inductance of the transformer is different from the calculated value, it might be necessary to adjust the core gap to obtain the correct primary inductance.

Due to the non-linearity of the ferrite material, it is always advisable after adjusting the gap of the flyback transformer, to actually check it on the bench and not only to rely on the measured inductance.

### VERIFICATION ON THE TEST SET-UP

On the test set-up, the correct peak current can be verified by applying  $I_{peak} = P_{in} / 20$ . Then the input power must be verified by applying  $P_{in} = 1/2 \times L_p \times I_{peak} \times F$  to check if it well corresponds to the maximum input power.

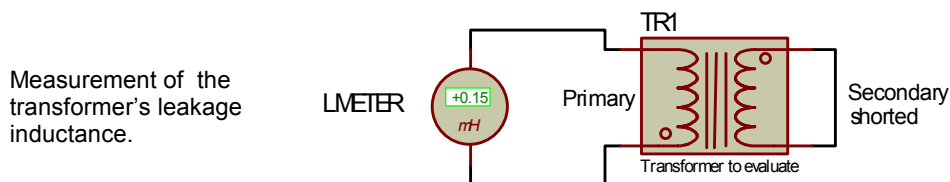
For more details on MOSFET current and input power, refer to section “MOSFET current” page 3 of application note AN- IPS-01.

**Note:** If the gap is too large, increase the number of primary turns or use a larger ferrite core.

### LEAKAGE INDUCTANCE

The leakage inductance is a key parameter affecting the efficiency, the voltage spikes, and EMI of a flyback SMPS. Low leakage inductance is relatively easy to achieve when  $N_p/N_s \leq 7$  for domestic applications and practically always allows the utilization of a Schottky diode for the load side. For  $N_p/N_s \geq 7$  special windings techniques might be required such as “split primary and/or secondary windings”, “secondary windings using foils instead of wires” etc...

For more details, refer to section “Leakage Inductance” page 5 of application note AN- IPS-01.



### PRIMARY CAPACITANCE AND PRIMARY TO SECONDARY CAPACITANCE

Refer to section “MOSFET current” page 5 of application note AN- IPS-01.