TOPSwitch® Flyback Power Supply Efficiency

Application Note AN-19

Introduction

Efficiency should be a major consideration when designing a flyback switching power supply. It can impact many aspects of a product design, from enclosure size and ventilation to safety qualification. Many design choices affect the efficiency of a given power supply design, including such seemingly disparate factors as the value of the bulk storage capacitor, transformer core geometry and construction, choice of output rectifier, and snubber and clipper networks. This application note will concentrate on efficiency comparisons between *TOPSwitch* and conventional power supply designs, efficiency measurement techniques, and general design considerations for constructing efficient flyback power supplies using *TOPSwitch*. For a more detailed treatment of flyback power supply and transformer design refer to application notes AN-16, AN-17 and AN-18.

TOPSwitch vs. Discrete Design

For a given output power, the efficiency attainable with a *TOPSwitch* flyback power supply is equal to or better than a conventional power supply design using a PWM IC controller

Design Type: (On Resistance/Voltage Rating)		3842 + MOSFET (1.2 Ω/600V)	*TOPSwitch* (3.6 Ω/700V)
DISCRETE vs. *TOPSwitch* PERFORMANCE (@ V_{IN} = 120 VAC)			
Output Power		34.27 W	34.21 W
Input Power		39.38 W	38.63 W
Efficiency		87%	88.6%
Power Loss		5.11 W	4.42 W
Operating Frequency		76 KHz	100 KHz
POWER LOSS BUDGET			
Conduction Loss	MOSFET $R_{DS(ON)}$	0.37 W	1.07 W
	Current Sense Resistor	0.16 W	None
Switching Loss	MOSFET CV^2f LOSS	0.43 W	0.32 W
	MOSFET Crossover Loss	1.08 W	Negligible
Other Losses	Start Up Circuit	0.03 W(1-2 W*)	Negligible
	PWM Controller	0.3 W	0.05 W
	Output Diode	1.00 W	0.98 W
	Clamp Circuit	1.20 W	1.07 W
	Miscellaneous Loss (Input Filter, Bridge, Transformer, Output Filter, Secondary Feedback)	0.54 W	0.93 W

**A more conventional single resistor start-up circuit will dissipate 1-2 Watts.*

Figure 1. Power Loss Comparison for Discrete and TOPSwitch Supplies.

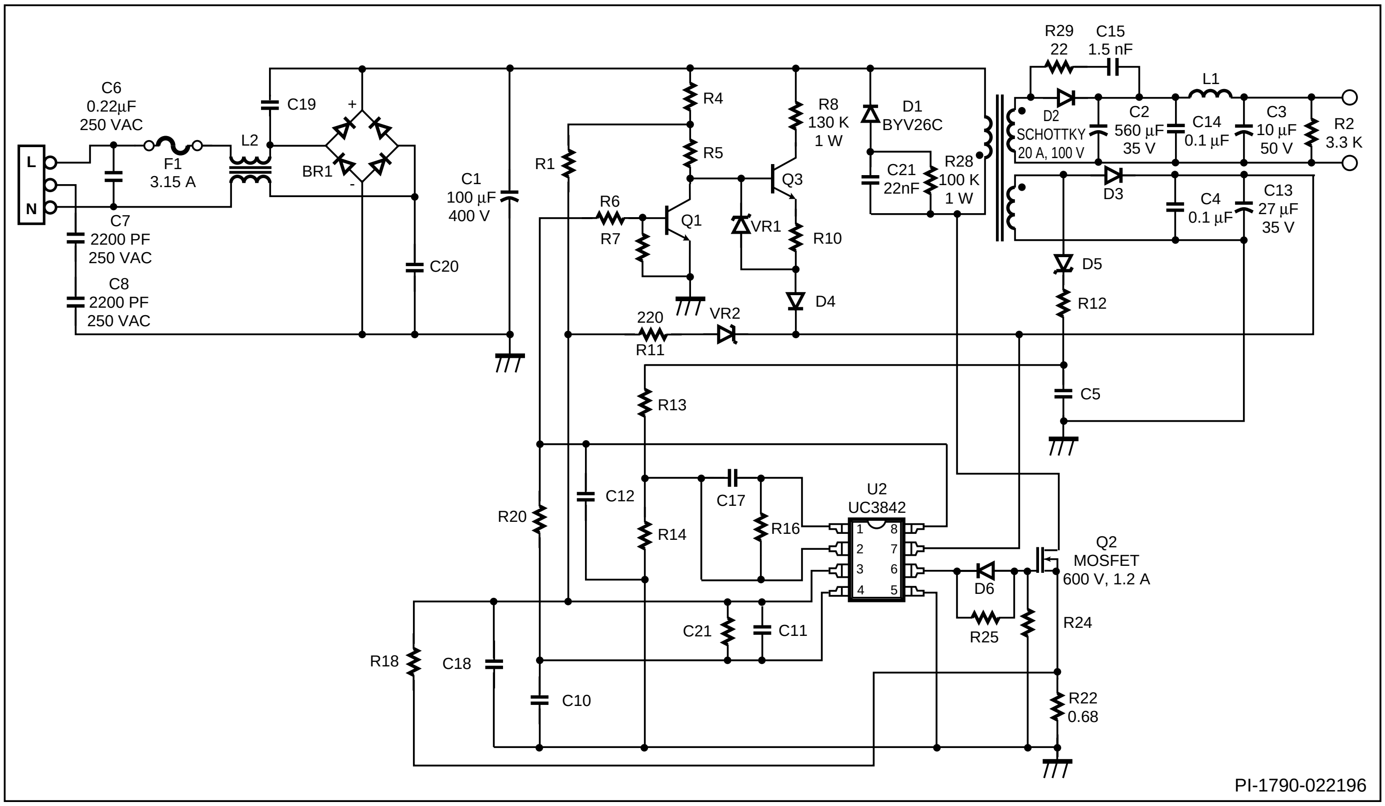

Figure 2. Discrete 3842 and MOSFET Flyback Power Supply.

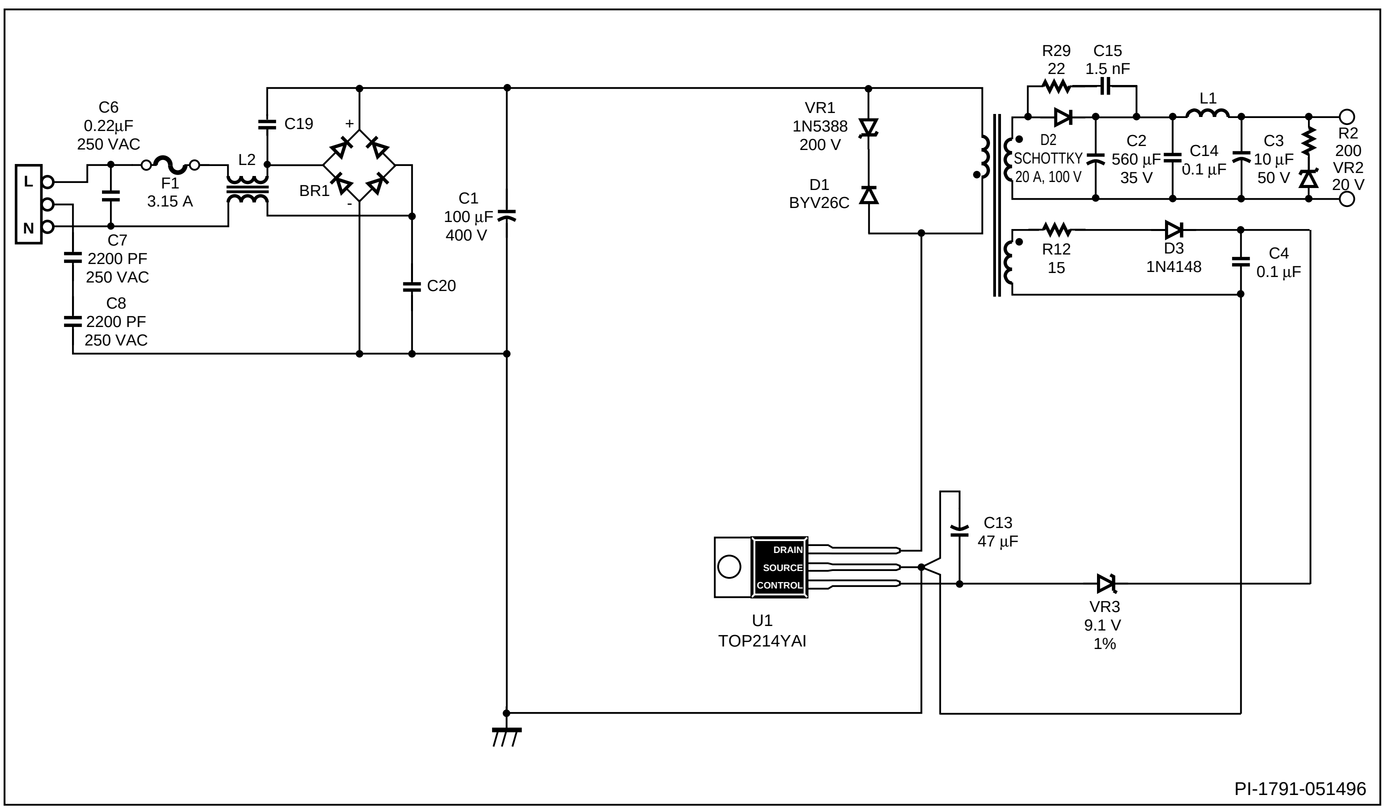

Figure 3. TOPSwitch Realization of Figure 2 Power Supply.

and discrete MOSFET, even though the discrete design may use a MOSFET with an $R_{DS(ON)}$ much lower than the equivalent *TOPSwitch*. The reasons for this become apparent when the power loss distribution for a discrete supply and *TOPSwitch* supply are examined and compared in detail. For this purpose, the power loss budget was measured for a commercially available 24V, 34W universal input switching power supply at an input voltage of 120 VAC. This discrete design, shown in Figure 2, uses a 3842 PWM IC controller and a 600 V, 1.2 ohm MOSFET, operating at a switching frequency of 76 KHz. The primary circuit of the supply was modified with a TOP214 IC, using the same transformer, rectifier, and output filter. The resulting design is shown in Figure 3. The power loss budget was measured for the converted supply at an input voltage of 120 VAC. The results for both designs are compared in Figure 1. **The efficiency of the *TOPSwitch* design is slightly better than the discrete design, even though the $R_{DS(ON)}$ of the TOP214 at 3.6 ohms is three times that of the MOSFET used in the discrete design.**

Part of the efficiency difference is due to the lower total losses in the *TOPSwitch* as compared to a discrete MOSFET. In Figure 1, the switch losses for the discrete MOSFET design and the *TOPSwitch* design are divided into two major components, conduction loss and switching loss. The conduction loss of the power MOSFET in the discrete design is 0.37 W, with an additional loss term of 0.16 W due to the current sense resistor. The *TOPSwitch* conduction loss is much higher at 1.07 W, due to the higher $R_{DS(ON)}$ of the TOP214. However, since the *TOPSwitch* uses the $R_{DS(ON)}$ of its internal MOSFET to sense current, there is no current sense resistor or its associated power loss. The switching losses are divided into two components, CV^2f losses and crossover losses. The CV^2f term represents the dissipation due to the stored energy in the parasitic capacitance of the transformer and the output capacitance of the MOSFET switch, which must be discharged by the MOSFET at the beginning of each cycle. The crossover loss term is due to the finite switching time of the MOSFET. During turn on and turn off, there is a short period when there is significant overlap of voltage and current across the MOSFET. A slow MOSFET will have a long overlap period at turn on and turn off, resulting in higher losses. The CV^2f losses of the *TOPSwitch* circuit are only 74% of the losses of the discrete circuit, even though the *TOPSwitch* is running at 100 KHz as compared to 76 KHz for the discrete design. Both designs use the same transformer, so the difference in CV^2f loss is due to the lower output capacitance of the TOP214, which is about one tenth of the output capacitance of a comparable discrete MOSFET. The switching crossover loss of the *TOPSwitch* is negligible, compared to the 1.08 W crossover loss of the discrete MOSFET. This difference is due

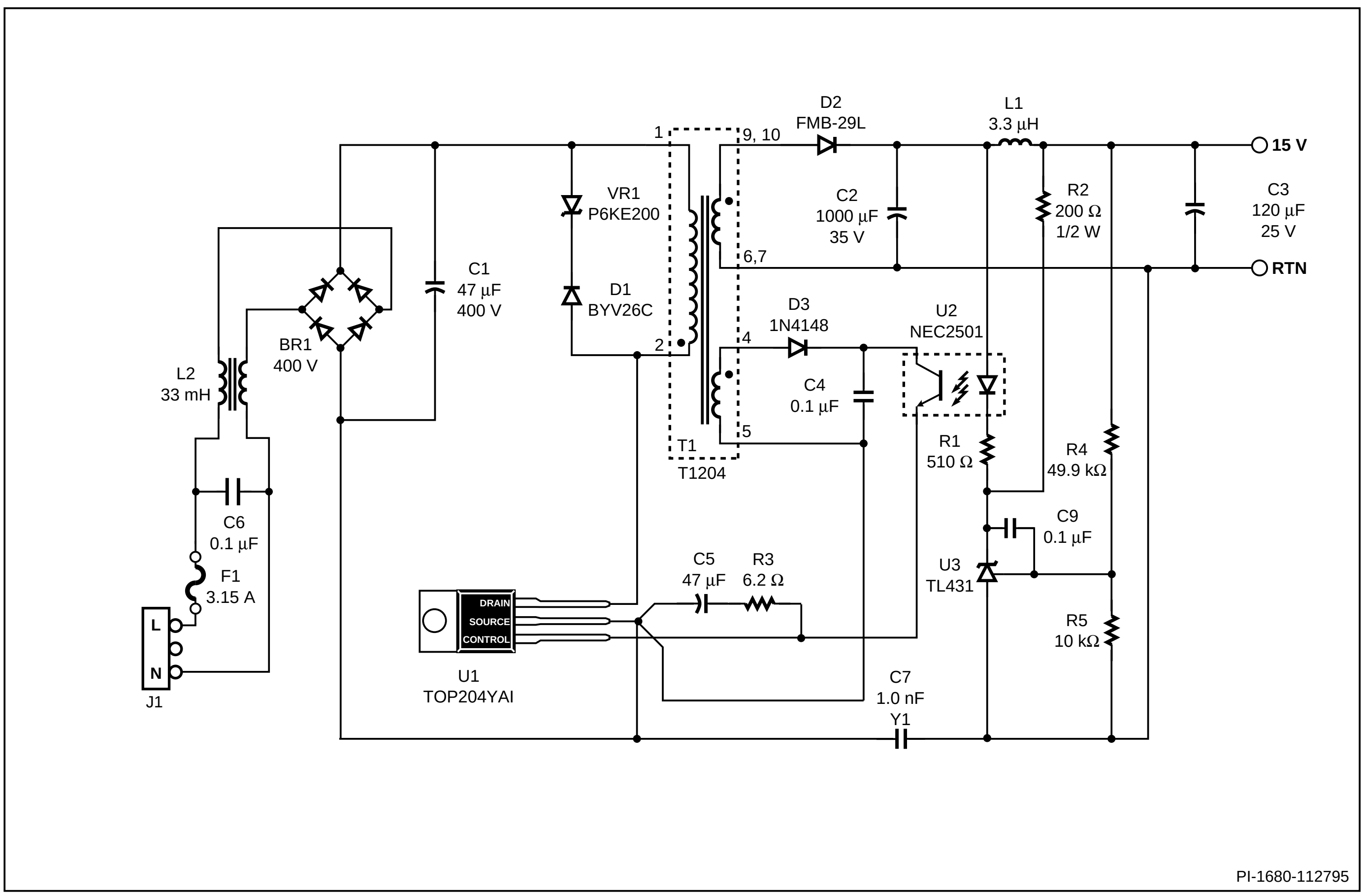

Figure 4. Schematic Diagram of the ST204A Power Supply.

to the low output and Miller capacitance of the *TOPSwitch* internal MOSFET, resulting in fast switching times. The sum of the conduction, CV^2f, and crossover losses is 1.39 W for the TOP214, as compared to 2.04 W for the discrete MOSFET. The higher conduction losses of the *TOPSwitch* are offset by lower switching losses.

The *TOPSwitch* and discrete designs also differ in the amount of energy that is consumed by the startup and control circuitry. These losses are shown in Figure 1. The losses in the *TOPSwitch* startup circuit are negligible, compared to 30 mW for the discrete design. The *TOPSwitch* has an internal startup supply that is automatically switched off when the *TOPSwitch* starts up, so that there are no losses due to the startup circuitry when *TOPSwitch* is in operation. Most 3842 and MOSFET power supply designs use a resistor connected to the high voltage bus to provide startup bias rather than the more sophisticated startup bias circuit shown in Figure 2. **Conventional 3842 and MOSFET supplies using resistive startup bias will dissipate a constant 1-2 watts in the bias resistor, making the advantages conferred by *TOPSwitch* even more apparent.** Controller power consumption for *TOPSwitch* is much less than a comparable 3842-based circuit (50 mW vs. 300 mW). The losses in the controller are due to the power consumed by the control circuit and the power required for the controller to drive the MOSFET switch. The *TOPSwitch* MOSFET is a low threshold device with low gate capacitance and almost negligible Miller capacitance, resulting in very low drive power requirements. The *TOPSwitch* controller is a low power CMOS design typically requiring only 5.7 V, 2.5 mA at maximum duty cycle, and 6.5 mA at minimum duty cycle. All of these factors contribute to the low power consumption of the *TOPSwitch* controller.

Output diode losses for the discrete and *TOPSwitch* supplies are roughly the same. Primary clamp circuit losses are lower for the *TOPSwitch* circuit, due to the higher operating frequency of the *TOPSwitch* design. For a fixed value of primary inductance, higher operating frequency results in lower peak operating current in the primary, reducing the amount of energy stored in the leakage inductance and therefore, the power lost in the clamp circuit. Miscellaneous losses in the *TOPSwitch* circuit are higher, due to the higher operating frequency of *TOPSwitch* (100 KHz vs. 76 KHz for the discrete design) and the higher output pre-load current.

TOPSwitch Power Supply Efficiency

For purposes of illustration, a more detailed efficiency analysis was performed using the ST204A reference design board. The ST204A is a 15 V, 30 W universal-input flyback power supply using the TOP204. A complete schematic diagram of this design is shown in Figure 4. As shown in Figure 5, this design has an efficiency of 80% or greater at full load over most of the 90-264 VAC operating range. At input voltages greater than 120 VAC, efficiency is between 85 and 87%. Various parts of this design will be examined with regard to their effect on the overall efficiency of the supply, and techniques will be presented for efficiency measurement and optimization.

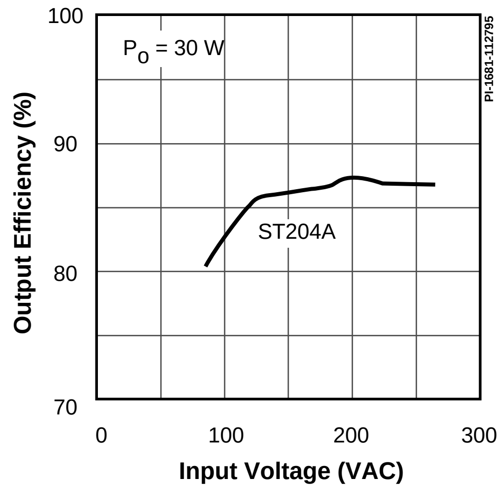

Figure 5. Efficiency vs. Input Voltage, 30 W Output.

Elements of Power Consumption

A power dissipation budget for the ST204A is shown in Figure 6. Power dissipation was measured for selected components at input voltages of 90, 120, and 240 VAC, for 30 W output power. Due to inevitable inaccuracies in measuring and estimating the dissipation of the various components, the sum of all the individual power loss components is different from the total power loss as measured at the supply input by 3-4%. A relatively small group of components is responsible for most of the power loss in the ST204A. These are the input common mode inductor (L2), input rectifier bridge (BR1), *TOPSwitch* (U1), drain voltage clamp Zener (VR1), and the output rectifier (D2). Other components dissipate a relatively small amount of power, but have a large effect on the overall efficiency of the supply. These are the input filter capacitor (C1) and transformer (T1).

Measurement Techniques

Measuring the true efficiency contribution of each component in a power supply is not always a straightforward process. In the case of the ST204A and the comparison study cited above, three measurement techniques were used to obtain the power budgets shown in Figures 1 and 6: direct measurement with a wattmeter, calculation from voltage and current measurements, and the DC thermal equivalent method.

Direct Measurement

Direct measurement of power dissipation is useful mostly for measuring the overall efficiency of a power supply. This measurement is best performed with a wattmeter which is designed to provide the average reading necessary to obtain

INPUT VOLTAGE:	90 VAC	120 VAC	240 VAC
ST204A PERFORMANCE vs. INPUT VOLTAGE			
Output Power	29.8 W	29.8 W	29.8 W
Input Power	37.8 W	36.1 W	35.3 W
Efficiency	78.8%	82.5%	84.4%
Power Loss	8.0 W	6.3 W	5.5 W
POWER LOSS BUDGET			
TOPSwitch (U1) (CV^2f+Conduction)	1.2 W	1 W	0.81 W
Clamp Diode (D1)	0.05 W	0.05 W	0.05 W
Clamp Zener (VR1)	1.3 W	1.3 W	1.3 W
Output Diode (D2)	1.5 W	1.5 W	1.5 W
Transformer (T1): Primary Secondary Core	(est.) 0.05 W 0.15 W 0.05 W	(est.) 0.04 W 0.16 W 0.05 W	(est.) 0.02 W 0.30 W 0.05 W
Input Choke (L2)	2.1 W	0.85 W	0.22 W
Diode Bridge (BR1)	1.02 W	0.85 W	0.41 W
Output Choke (L1)	0.08 W	0.08 W	0.08 W
Output Filter (C2)	0.11 W (est.)	0.09 W (est.)	0.05 W (est.)
Input Filter (C1)	0.32 W (est.)	0.22 W (est.)	0.13 W (est.)
Bias Winding and Feedback Circuits	0.28 W	0.30 W	0.36 W

Figure 6. Power Loss Budget for ST204A at 30 W Output Power.

input power measurements. RMS-reading meters are not suitable for use in efficiency measurements on AC circuits. Use of a RMS meter on the AC input of a power supply operating at less than unity power factor will provide a grossly understated value for efficiency. The wattmeter should have enough resolution in the power range of interest to perform accurate measurements. Also, the meter should be capable of handling waveforms with a crest factor (ratio of peak value to RMS value) of at least 3:1. The capacitor input filter used on most switching power supplies distorts the input current waveform, resulting in high crest factor and low power factor (typically 0.6 to 0.8, depending on AC line impedance and input voltage).

The meter should be connected as shown in Figure 7, with the voltage sensing leads as close as possible to the input of the supply, to avoid error due to voltage drop in the input cable. This voltage drop can cause an error of 1-2 efficiency points, even for a small power supply.

If a wattmeter is not available, overall efficiency measurements for a power supply can be made by applying high-voltage DC current to the input, and making voltage and current readings with conventional meters. Most switching power supplies run equally well with DC or AC input. However, do not use this method with a power supply having a fan or transformer connected directly across the AC input, as these components will present a short circuit to a DC input. Efficiency measurements will be 1-2 points higher with DC input than with AC input, as the power supply will be running at unity power factor, causing less stress on the input components. Also, there will be no line frequency ripple on the bulk filter capacitor, allowing the supply to run at a higher average input voltage and lower average current, dissipating less power. However, DC input measurements can be used to obtain approximate efficiency measurements. If a high-voltage DC supply is not available, a rectifier bridge and filter capacitor can be used to generate a high-voltage DC bus from the AC line. Ripple on this supply should be kept to less than 5% at the maximum power of the unit under test to reduce the error due to line frequency ripple.

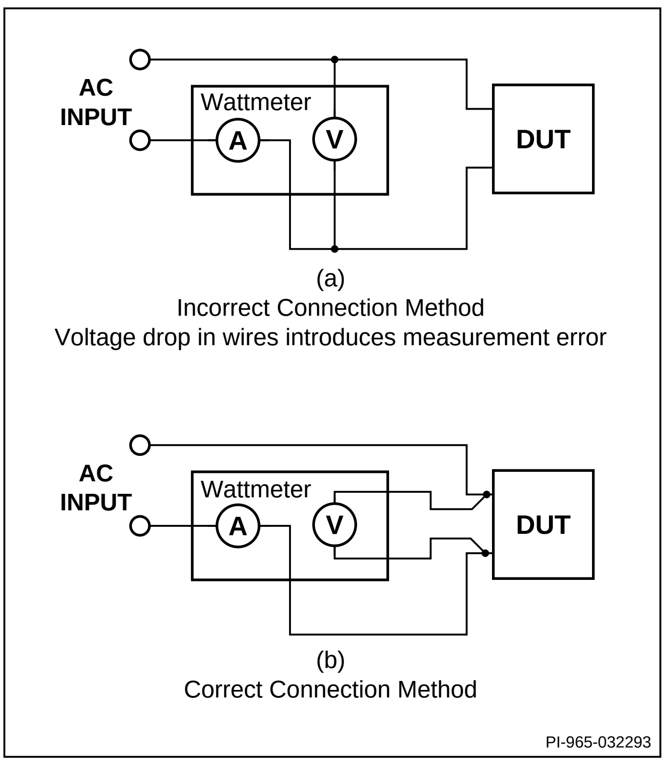

Figure 7. Wattmeter Connection Methods.

Calculation

In some cases, voltage and current measurements and calculation will be the most straightforward method of determining the power dissipation of a given component. In the ST204A, this method was used for C1 (input filter capacitor), and C2 (output filter capacitor), T1, L1 and for the primary bias and secondary feedback circuits. Some digital oscilloscopes have sufficiently advanced math functions to be able to directly calculate the average power from voltage and current waveforms. It should be noted that some active current probes have delays on the order of 50 nsec. This can cause a significant amount of error in measuring the instantaneous power dissipation of a switching device. A more reliable means of measuring power dissipation in such cases is to use the DC thermal equivalent method.

DC Thermal Equivalent Method

This is a very powerful method for obtaining estimates of power dissipation, especially for components such as switching transistors, MOSFETs, and power diodes, where there are reverse recovery and switching losses as well as forward conduction losses. To use this method, the temperature rise in a given component is measured. The same temperature rise is then induced in that component using a DC current. From DC voltage and current measurements, the average power dissipation in the component can then be determined. This method was used to determine the power loss in U1, D1, D2, VR1, L2, and BR1 of the ST204A circuit.

Designing for Higher Efficiency

Almost every step in a flyback power supply design can affect efficiency, from initial design considerations of maximum duty cycle and transformer primary inductance to the choice of components. The following paragraphs examine the effect of initial design parameters and component selection on efficiency. Where possible, general design guidelines are given for selecting components and initial design parameters.

Transformer Inductance

Higher efficiency can be achieved in a flyback design by reducing the RMS operating current of the primary *TOPSwitch*. Lower RMS primary current can be achieved by providing sufficient primary inductance in the transformer to allow the supply to run in the continuous mode of operation. This reduces both the peak and RMS currents in the primary, cutting dissipation in *TOPSwitch*, the transformer windings, the output rectifier, and the output filter capacitor. As an added benefit, the amount of energy stored in the leakage inductance of the transformer is reduced. This stored energy scales as the square of the peak current, and must be dissipated each switching cycle in the primary clamp circuit. Reducing the primary peak current by increasing the transformer primary inductance reduces primary clamp losses. Design techniques for choosing the optimum transformer inductance for a given application are presented in AN-16 and AN-17. Transformer construction techniques are shown in AN-18. The primary inductance of transformer T1 in the ST204A circuit is 627 µH. At 30 W output, this value of primary inductance results in continuous mode operation over much of the input voltage range. A comparison between discontinuous and continuous mode operation is shown in Appendix A.

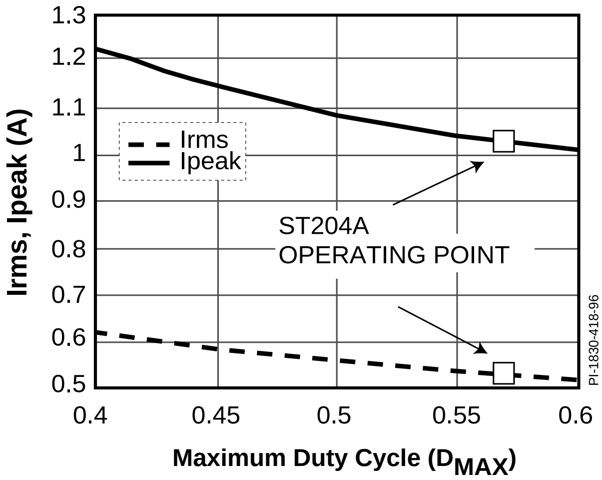

Figure 8. Primary Peak, RMS Current vs. Maximum Duty Cycle for ST204A.

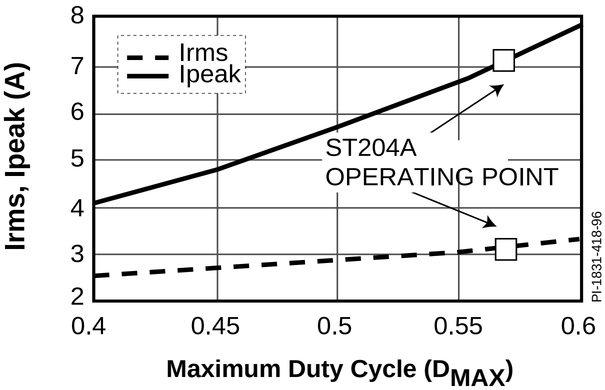

Figure 9. Secondary Peak, RMS Current vs. Maximum Duty Cycle for ST204A.

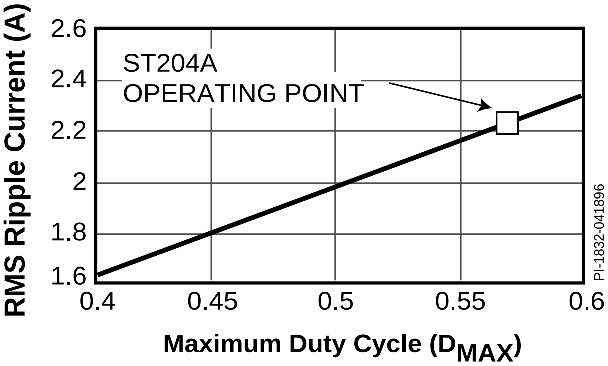

Figure 10. Output Filter Capacitor RMS Ripple current vs. Maximum Duty Cycle for ST204A.

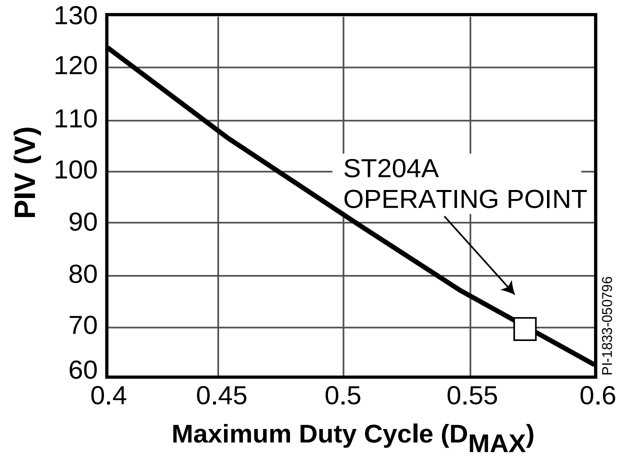

Figure 11. Output Diode Peak Inverse Voltage at Maximum Input Voltage vs. Maximum Duty Cycle for ST204A.

Maximum Duty Cycle

For a flyback power supply, the choice of maximum duty cycle at minimum input voltage (D_{MAX}) determines in part the apportioning of losses between the primary and secondary components of the power supply. The maximum duty cycle discussed here does not refer to the internal duty cycle limit of *TOPSwitch*, but to an operating point set externally to *TOPSwitch*. The power supply duty cycle for a given input and output voltage is set by the ratio of secondary to primary turns of the flyback transformer. A lower ratio of secondary to primary transformer turns (N_S/N_P) results in a higher D_{MAX}, governed by the formula:

$$D_{MAX} = \frac{V_O + V_D}{\left(V_{MIN} \times \frac{N_S}{N_P}\right) + V_O + V_D} \qquad (1)$$

D_{MAX} is the duty cycle for minimum input voltage, V_O is the output voltage, V_D is the forward voltage drop of the output rectifier, N_S is the number of transformer secondary turns, and N_P is the number of primary turns. The maximum *TOPSwitch* duty cycle limit provides a limit point for the duty cycle. The choice of secondary to primary turns ratio is not completely arbitrary. In addition to setting the operating duty cycle of the supply, the turns ratio also determines the magnitude of the reflected secondary voltage across the primary winding of the transformer during the *TOPSwitch* off time, according to the equation:

$$V_{OR} = \frac{N_P}{N_S} \times (V_O + V_D) \qquad (2)$$

V_{OR} is the reflected value of the secondary voltage across the primary of the transformer. During the *TOPSwitch* off time, the drain voltage is equal to the sum of the primary DC bus voltage and the reflected secondary voltage at the transformer primary, plus any leakage spikes. This voltage sets a practical limit to the transformer turns ratio, and hence the maximum duty cycle for a given *TOPSwitch* application. The limit can be expressed in terms of a maximum recommended value of reflected secondary voltage for a *TOPSwitch* application, and is used in the initial stages of transformer design to determine transformer turns ratio. For TOP100 series parts operating in the input voltage range of 85-132 VAC, the maximum recommended value of V_{OR} is 60 V. For TOP200 series parts operating in the input voltage range of 195-265 VAC or 85-265 VAC, the maximum recommended V_{OR} is 135 V. This sets a D_{MAX} for 115 V input supplies using TOP100 series parts of 40%. For 230 V input supplies using TOP200 series parts, the D_{MAX} limit set by reflected voltage considerations is also 40%. For universal input supplies (85-265 VAC) using TOP200 series parts, D_{MAX} is 60%. The effect of D_{MAX} and V_{OR} on power supply and transformer design is discussed in application notes AN-16 and AN-17. The peak *TOPSwitch* drain voltage can be readily controlled using a Zener clamp circuit. This circuit (VR1 and D1) is shown in Figure 4. For 115 V input applications, the recommended Zener voltage for VR1 is 90 V. For 230 V or universal input, the recommended Zener voltage is 200 V. Application Note AN-14 gives a table of Zener diode types suitable for use with *TOPSwitch*.

The ST204A power supply design shown in Figure 4 is a universal input design using a TOP204. The following paragraphs discuss some of the trade-offs involved in varying D_{MAX} for this design from a minimum of 40% to the maximum recommended value of 60%. Continuous mode operation was assumed, with a primary inductance of 627 mH, 100 KHz operating frequency, minimum primary DC bus voltage of 90 V, estimated efficiency of 80%, and 15 V, 30 W output power. The minimum input voltage condition represents the point of maximum current stress on *TOPSwitch* and other components in the ST204A. Figures 8-11 show primary and secondary peak and RMS currents and output capacitor RMS ripple current as a function of D_{MAX}. In Figure 11, the maximum output diode peak inverse voltage (for V_{IN} = 265 VAC) is shown as a function of D_{MAX}. The actual ST204A operating point (D_{MAX} = 0.57) is marked on each graph.

As shown in Figure 8, primary peak and RMS currents decrease from 1.22A and 0.62A for D_{MAX} = 40% to 1.1A and 0.53A for D_{MAX} = 60%. *TOPSwitch* conduction loss varies as the square of RMS drain current, so the higher D_{MAX} can cut conduction losses by as much as 27%, increasing efficiency or allowing use of a smaller *TOPSwitch.* Figures 9 and 10 show that although the primary current levels are decreased by a higher value of D_{MAX}, the secondary peak and RMS currents and the output capacitor ripple current increase with increasing D_{MAX}. This trade-off is characteristic of flyback power supplies in general. The increase in secondary RMS current does not affect the output rectifier to the same extent as the *TOPSwitch*, as the output diode voltage drop as a function of forward current is fairly constant. At higher D_{MAX}, the current stress on the output rectifier is increased, but the voltage stress is decreased, due to the smaller ratio of secondary to primary transformer turns. This can be shown by rearranging Equation (1) to obtain N_S/N_P as a function of D_{MAX}.

$$\frac{N_S}{N_P} = \frac{(1 - D_{MAX}) \times (V_O + V_D)}{D_{MAX} \times V_{MIN}} \tag{3}$$

For output voltages of 24 V or below, the lower secondary voltage stress resulting from a higher D_{MAX} may allow use of a Schottky rectifier instead of an ultrafast rectifier. The lower voltage drop of the Schottky rectifier reduces the output rectifier power dissipation despite the increased peak and RMS secondary currents resulting from the increase in D_{MAX}, and can result in an overall increase in supply efficiency. For the ST204A, the peak inverse voltage across the output diode as a function of D_{MAX} is shown in Figure 11. This curve was calculated for the maximum input voltage for ST204A (265 VAC) to show the maximum peak inverse voltage on the output rectifier. For a D_{MAX} of 0.5 or greater, the peak inverse voltage is low enough to allow use of a Schottky output diode with adequate voltage margin.

Though the RMS secondary current is not a very strong function of duty cycle, the peak secondary current rises dramatically with D_{MAX}. This high peak current can cause noise problems due to interaction with the parasitic inductances of the secondary circuit. Secondary circuit layout will be more critical for a power supply designed for a high D_{MAX}. The RMS ripple current in the output filter capacitor also increases with increasing D_{MAX}, as shown in Figure 10. This will necessitate using a capacitor with a higher ripple current rating for higher values of D_{MAX}. In general, using the highest recommended value of D_{MAX} for a given *TOPSwitch* and input voltage will maximize the power capability and/or efficiency of a *TOPSwitch* flyback supply by reducing the peak and RMS primary currents. When using the maximum recommended D_{MAX}, the output rectifier and filter capacitor must be sized to accommodate the increased secondary peak and RMS currents. The lower secondary voltage stress resulting from a larger value of D_{MAX} may allow the use of a Schottky output rectifier, resulting in higher overall supply efficiency. Setting the transformer turns ratio to the recommended V_{OR} value for a given *TOPSwitch* application will automatically set D_{MAX} to the maximum recommended value.

Operating Frequency

Operating frequency of a switching power supply should be carefully considered, due to the trade-off between the size advantage gained by high frequency operation and the decrease in efficiency due to transformer core and copper losses, diode and MOSFET switching losses, and snubber losses. 100 KHz appears to be the optimum operating frequency for supplies operating from 100/115/230 VAC mains, for reasons of cost, size, efficiency, and EMI. For this reason, *TOPSwitch* has a fixed operating frequency of 100 kHz.

Transformer Construction

In a flyback converter, the transformer is the main energy storage and processing element. Therefore, it has a large effect on the efficiency of the power supply as a whole. An efficient flyback transformer will have low DC losses, low AC losses, low leakage inductance, and low winding capacitance.

DC Losses

The only significant DC losses in a power transformer will be due to the copper losses in the transformer windings. For a high-efficiency design, the transformer wire gauge should be sufficiently large to reduce the copper losses to an acceptable level. A traditional design guideline is to size each winding for a current capacity of 200 to 500 circular mils per RMS ampere of current. The power windings in the transformer of the ST204A example circuit were designed for a current density of approximately 300 circular mils per RMS ampere.

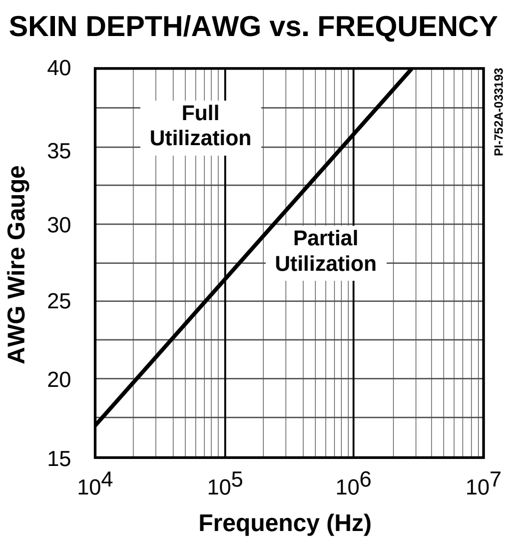

Figure 12. Skin Depth vs. Frequency.

AC Losses

AC losses in the transformer arise from skin effect losses in the transformer windings, and AC core losses. High frequency currents tend to flow close to the surface of a conductor rather than its interior. This phenomenon is called the skin effect. The penetration of AC current into a conductor varies as the square root of the frequency, so for a higher frequency, currents will flow closer to the surface of the conductor and leave the interior underutilized. The result is a higher effective resistance for AC current versus DC current. To minimize the AC copper losses in a transformer, no conductor should be used that has a thickness greater than 2 times the skin depth at the operating frequency of the supply. A graph of usable wire gauge as a function of frequency is shown in Figure 12. As an example, for 100 KHz, 26 AWG is the largest wire size that allows full utilization of the cross-section of the wire. High current windings at 100 KHz should be constructed using several strands of 26 AWG or smaller wire rather than one large diameter conductor in order to allow full utilization of the conductor. Foil conductors with a thickness less than or equal to twice the skin depth can also be used for high current windings to reduce skin effect losses.

Core losses can also add significantly to the power loss of a transformer. These losses are due to the AC component of the flux in the transformer. The AC flux density in a flyback transformer can be estimated using either of two formulas:

$$B_{AC} = \frac{0.4\pi \times N_P \times I_P \times K_{RP}}{2 \times l_g} = \frac{10^8 \times V_{MIN} \times D_{MAX}}{2 \times N_P \times A_e \times f_S} \quad (4)$$

B_{AC} is half of the peak to peak flux density in Gauss, N_P is the number of primary turns, I_p is the primary peak current, K_{RP} is the ripple current to peak current ratio (refer to Appendix A), l_g is the transformer core gap in centimeters, V_{MIN} is the minimum DC input voltage to the transformer, D_{MAX} is the duty cycle at minimum input voltage, A_e is the core cross-sectional area in square centimeters, and f_S is the operating frequency in Hertz. The first equation calculates the flux excursion using ampere-turns, and the second expression determines the flux excursion using the volt-second product. Both equations should yield the same result for the same operating conditions. The choice of using one expression versus the other is determined by the basic information about the transformer that is readily available. For a transformer designed to operate in the continuous mode, typical values of B_{AC} will be around 400-750 Gauss. For high-efficiency operation, a material should be selected that will keep core losses under 50 mW/cm^3 at 100 KHz. Refer to the core loss curves published by ferrite manufacturers in order to select a suitable core material. A few examples of appropriate ferrite materials for 100 KHz designs are Philips 3C85 and 3F3, Magnetics, Inc. P and R, TDK PC40, Siemens N67 , Tokin 2500B and 2500B2, and B2 material from Thomson.

Leakage Inductance

A very important consideration in designing a low-loss transformer is minimizing the amount of leakage inductance. A transformer with high leakage inductance will dissipate large amounts of energy in the primary clamp components. The energy dissipated in the clamp is wasted and detracts from the overall efficiency.

For a transformer meeting international insulation and safety requirements, a practical value for leakage inductance is about 1-3% of the open circuit primary inductance. Values much below 1% are difficult to attain and still meet safety regulations.

Designing transformers for low leakage inductance involves several considerations:

- Minimize number of turns
- Keep winding build (ratio of winding height to width) small
- Increase width of windings
- Minimize insulation between windings
- Increase coupling between windings

These considerations are also discussed in AN-17 and AN-18.

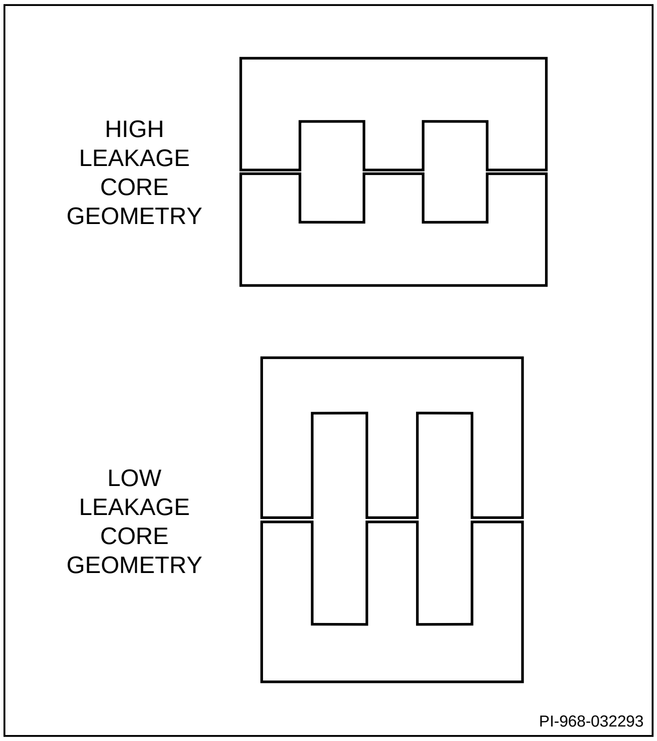

Figure 13. Leakage Characteristics of Different Core Shapes.

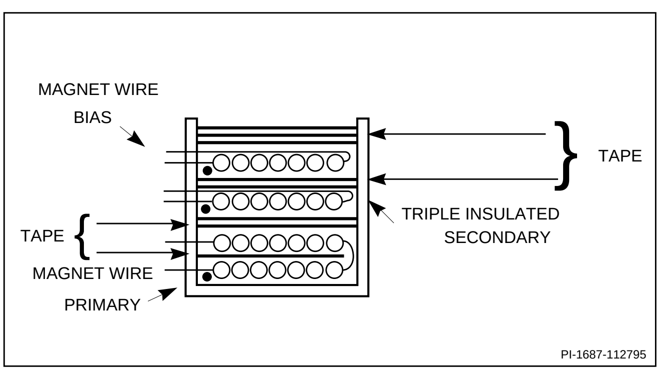

Figure 14. Cross Section of Transformer Using Triple Insulated Wire.

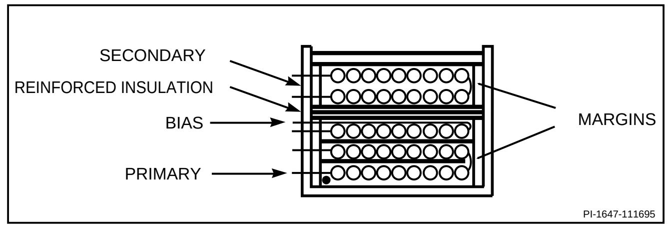

Figure 15. Cross Section of a Transformer Construction Using Magnet Wire and Margins.

Minimizing Winding Turns and Build

The first step in building a low-leakage transformer is choosing the proper core geometry. Since leakage inductance varies as the number of turns squared, it is important to choose a core that will not require too large of a number of turns to attain the desired primary inductance. The core should be sufficiently large so that the required primary turns can be wound using 2 layers or less. This minimizes leakage inductance and winding capacitance. Figure 13 shows a comparison between a high leakage and a low leakage core geometry. Cores with short, fat window areas should be avoided, as the winding build will be large, resulting in high leakage inductance. Such cores include: pot cores, RM cores, PQ cores (especially the short versions), and some E cores. More suitable cores are ETD cores, EER cores, EC cores, and many E cores.

In many cases, the optimum core size and shape for a transformer of a given rating will depend on whether a triple insulated wire or magnet wire is used in the transformer construction. Triple insulated wire has three separable layers of insulation, any two of which can withstand the full UL/IEC hipot test voltage. Triple insulated wire thus satisfies the requirements for a reinforced insulation per UL/IEC regulations, and can be used to construct a transformer without the creepage margins required in a design using conventional magnet wire. A cross -sectional comparison between a triple insulated wire transformer design and a conventional magnet wire design is shown in Figures 14 and 15. The triple insulated wire design uses a magnet wire primary and a triple insulated secondary. This is generally the most cost effective and space-efficient way to utilize the benefits of triple insulated wire, as it is larger in diameter and more costly than magnet wire. The secondary winding will require fewer turns of larger diameter wire than the primary, so the cost and space impact of the triple insulated wire is minimized. Note that in the triple insulated wire design , the full width of the transformer bobbin is usable, due to the reinforced insulation provided by the triple insulated wire. A transformer using a triple insulated wire design will generally be 1/2 to 2/3 of the size of a transformer of the same power capability using a magnet wire design due to the elimination of safety margins. Leakage inductance will also be less for the triple insulated design due to the improved utilization of space on the transformer bobbin.

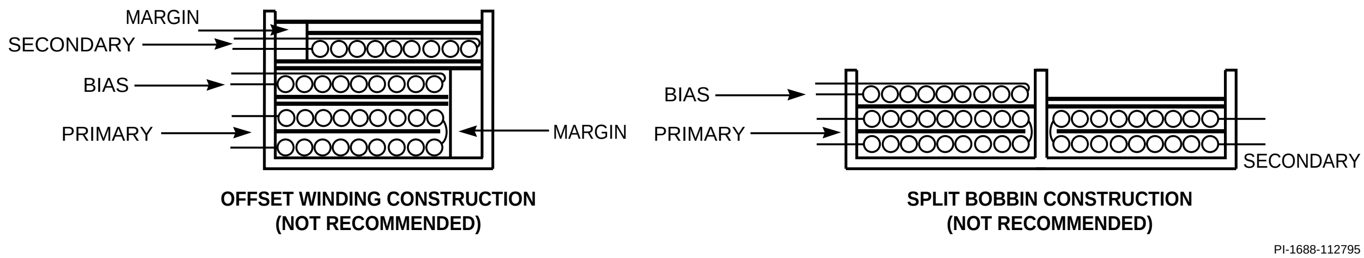

Figure 16. Offset and Split Bobbin Construction Techniques (Not Recommended).

The transformer in the ST204A uses triple insulated wire with an EI28 core for 30 W output. A magnet wire design would require a EI30 core to accommodate the creepage margins, resulting in a 33% size increase.

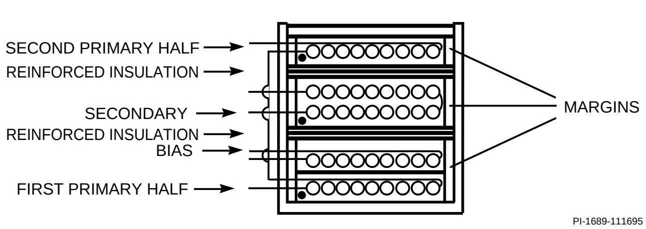

Figure 17. Split Primary Construction to Reduce Leakage Inductance.

Winding Arrangement for Minimum Leakage Inductance
The arrangement of windings in a transformer will have a large effect on the leakage inductance. Transformer windings should be arranged in concentric fashion for minimum leakage inductance, as shown in Figures 14 and 15. Offset or split bobbin construction (shown in Figure 16) should be avoided, as this technique will result in high leakage inductance and unacceptable losses in the primary clamp circuit.

In a multiple output transformer, the secondary with the highest output power should be placed closest to the primary for the best coupling and lowest leakage. For higher power applications (40 watts and above), it is recommended to use a split primary “sandwich” construction as shown in Figure 17 to minimize leakage inductance. Using a split primary will usually cut the leakage inductance to half that of a transformer with a single primary winding.

High power secondary windings consisting of only a few turns should be spaced across the width of the bobbin window instead of being bunched together, in order to maximize coupling with the primary. Using multiple parallel strands of wire is also a good technique of increasing the fill factor and coupling of a winding with few turns. In such cases, the wire size may be determined more by the requirement for a good fill factor rather than the RMS current rating of the wire. Where cost permits, using foil windings is also a good way of increasing coupling, although this method is practical only for low voltage, high current secondary windings. A more complete discussion of transformer leakage inductance can be found in several of the references listed at the conclusion of this document.

Transformer Capacitance
A high-efficiency transformer should have low interwinding capacitance. Energy stored in the parasitic capacitance of the transformer is absorbed by the *TOPSwitch* each cycle during the turn-on transition, reducing efficiency. Excess capacitance also will ring with stray inductance during switch transitions, causing noise problems. Capacitance effects are usually the most important in the primary winding, where the operating voltage (and consequent energy storage) is high. The primary winding should be the first winding on the transformer. This allows the primary winding to have a low mean length per turn, reducing the internal capacitance. The driven end of the primary winding (the end connected to the *TOPSwitch* Drain pin) should be the start of the winding rather than the finish. This takes advantage of the shielding effect of the second half of the primary winding and reduces capacitive coupling to adjacent windings. A layer of insulation between adjacent primary windings can cut the internal capacitance of the primary winding by as much as a factor of four, with consequent reduction of CV^2f losses for the *TOPSwitch*.

Primary Components

Inrush Limiters
There are several options that can be used to avoid excessive inrush current into the input filter capacitor of a power supply during initial turn on. In very low power supplies or supplies where there is no stringent limit on inrush current, it is possible to use a certain amount of fixed impedance, either in the form of a fixed resistor, or resistance built into the RFI filter inductors. The loss in efficiency can usually be tolerated, especially as these methods offer space and cost savings which can be vital. The ST204A utilizes the impedance of the input common mode choke (L2) in order to reduce inrush current. If tighter control of the inrush current is needed along with higher efficiency, a negative temperature coefficient thermistor can be used as an

inrush limiter. The thermistor presents a large impedance to the AC line upon initial startup, but self-heats and turns to a low resistance after a short period of time. The current rating of this thermistor should be carefully chosen. If a device with too high a current rating is chosen for a given application, it will not self-heat sufficiently, and will act as a considerable impedance between the supply and the AC line, wasting power. Care should be taken that the lead temperature of the thermistor where it enters the PC board does not exceed the safety rating of the board material at maximum ambient temperature. Many manufacturers make their inrush limiting thermistors with steel leads instead of copper to provide some thermal isolation from the PC board.

Input Diode Bridge

The input diode bridge on a switching power supply should have a rating at least equal to the RMS input current of the power supply at the lowest limit of the AC line voltage. This current can be estimated by:

$$I_{RMS} \approx \frac{P_{out}}{\eta \times V_{ACMIN} \times PF} \qquad (5)$$

I_{RMS} is the RMS input current, η is the estimated efficiency of the supply, V_{ACMIN} is the minimum RMS AC input voltage, and PF is an estimate of the input power factor. Practical values of PF range from 0.6 to 0.8, depending on line voltage and the effective impedance of the AC line feeding the supply. A small efficiency increase can be obtained by increasing the current rating of the input rectifier bridge, so that it operates at a lower current density. This results in a lower forward voltage drop, reducing power loss in the input rectifier

Input RFI Filter

As shown in the power loss tabulation of Figure 6, the losses in the common mode choke L2 can be significant at low input voltages, where the RMS input current is highest. In order to increase the efficiency at low line, the common mode choke can be changed to a physically larger size unit of the same inductance with lower internal resistance. The necessary current rating for the common mode choke can be estimated using Equation (5). The choice of common mode choke size will be a trade-off between supply size, cost, and efficiency.

Input Capacitor Selection

The choice of an input filter capacitor can have a bearing on the efficiency of a power supply, especially at low input AC voltage. If the input capacitor is too small, there will be a large ripple component on the capacitor at low line, resulting in lower average voltage available for the *TOPSwitch*, higher average operating current through all components in the power path, and lower efficiency. A rough guideline is to size the total input capacitance at 1 microfarad per output watt for a 230 VAC-only supply or 2 microfarads per watt for a 115/230 VAC dual range supply or a 115 VAC single range supply. For a universal input supply or a 100 VAC supply with a full wave bridge rectifier, use 3 microfarads per output watt.

Snubber and Clipper Networks

Snubbers and clippers are used in a power supply circuit in order to limit voltage swing and reduce EMI. A typical use of a clipper circuit can be seen in the ST204A circuit in Figure 4. D1 and VR1 act as a clipper to limit the peak value of the leakage spike generated when *TOPSwitch* turns off. In order to minimize the power loss in snubbers and clippers, design the transformer for low leakage inductance, and plan the circuit layout for low stray inductance. Techniques for circuit layout design can be found in AN-14. Low leakage construction techniques for transformers are shown in AN-18. In some cases a RC snubber circuit may be necessary on the output rectifier to reduce EMI. In such cases, use the minimum amount of capacitance in the snubber necessary for RFI reduction. Too large a value of snubber capacitor will reduce power supply efficiency, especially at high input line voltage.

***TOPSwitch* Dissipation**

Dissipation in the *TOPSwitch* is an important factor in determining the efficiency of the power supply. The techniques outlined in the preceding paragraphs for reducing primary RMS current will help to reduce dissipation due to conduction losses in the *TOPSwitch*. The conduction losses for *TOPSwitch* are greatest at minimum input voltage, so reducing conduction losses will have the most effect on efficiency at low input voltage. If cost allows, choosing a *TOPSwitch* with lower $R_{DS(ON)}$ will help to reduce conduction losses. Generous heat sinking of the *TOPSwitch* will also help to reduce conduction losses, as the $R_{DS(ON)}$ of *TOPSwitch* (or any other MOSFET) increases with increasing junction temperature. Reducing the parasitic capacitance of the transformer will help reduce *TOPSwitch* switching losses, which become more important at high input voltages.

Secondary Components

Output Rectifiers

Output rectifiers are a major source of inefficiency in a power supply, and should be carefully chosen. As shown by the ST204A power dissipation budget of Figure 6, losses generated by the output rectifier are about one fourth to one fifth of the total system loss. Rectifiers have two important loss mechanisms: forward conduction loss, and reverse recovery loss. Both loss mechanisms can have a significant effect on power supply efficiency.

Both the choice of output rectifier and operating parameters can affect the efficiency of the power supply. A supply designed to run in the continuous mode will have lower secondary RMS and peak current than a supply designed for discontinuous operation, resulting in lower forward conduction losses. However, the output rectifier will be forced to recover while there is a

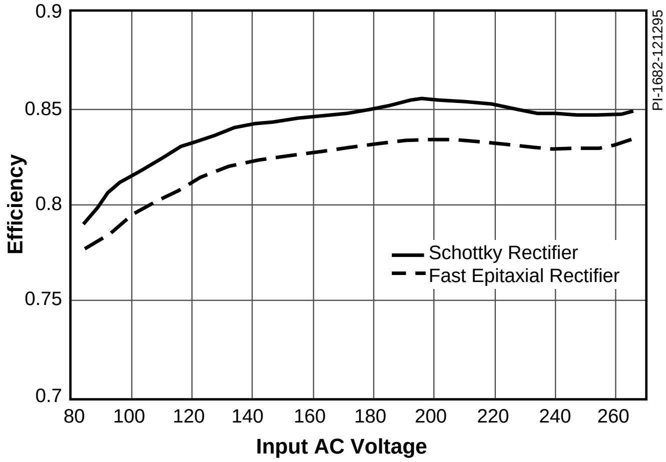

Figure 18. Efficiency vs. Input Voltage for Schottky and Fast Epitaxial Output Rectifier.

significant amount of current flowing through it, resulting in higher reverse losses in the continuous mode. Careful choice of the output rectifier can help to reduce reverse recovery losses. Where expense and voltage permit, Schottky rectifiers should be used if efficiency is of great importance. Schottky rectifiers have significantly lower forward losses than PN junction rectifiers. Since Schottky rectifiers are majority carrier devices, they do not have the reverse recovery losses exhibited by conventional fast recovery devices. However, Schottky rectifiers do have a comparatively large junction capacitance, which results in reverse current spikes similar to those generated by conventional fast recovery diodes, but without the internal losses of conventional rectifiers. The use of Schottky diodes is limited by their relatively low breakdown voltage ratings (≤100 V). If Schottky rectifiers cannot be used due to breakdown voltage considerations, the next best choice is a fast epitaxial rectifier with moderate forward losses and very fast reverse recovery times (< 50 nsec). These rectifiers are available with voltage ratings ranging from 50 V to 1000 V. Standard fast recovery rectifiers, with reverse recovery times in excess of 150 nsec, should not be used in *TOPSwitch* supplies unless there is no reverse recovery current drawn during diode turnoff. Under no circumstances should normal diffused-junction rectifiers be used except as AC input rectifiers.

The efficiency of the ST204A example circuit was measured using a Schottky output rectifier vs. a fast epitaxial rectifier to determine the difference in total power supply efficiency at an output power of 30 W. The two diodes compared were a Sanken FMB-29L Schottky rectifier and a Philips BYW29-200 fast epitaxial rectifier. The FMB-29L Schottky rectifier is rated at 90 V, 8A. The BYW29-200 is rated at 200 V, 8A, with a reverse recovery time of less than 25 nsec. Efficiency vs. AC input voltage for the two devices is shown in Figure 18. The supply efficiency at 115 VAC. using the FMB-29L Schottky rectifier is 84.05%. When the BYW29-200 is used, the efficiency is 82.53%. This efficiency difference is due mainly to the lower forward voltage drop of the Schottky rectifier.

Output Rectifier Current Rating

A general design rule for high efficiency is to choose the output rectifier current rating to be at least 3 times the rated DC output current of the supply. This reduces the power dissipation in the output rectifier by allowing it to run at a lower current density. The actual RMS and peak operating current of the output rectifier can be several times the DC output current of the power supply, as shown in Figure 9. This figure shows the secondary peak and RMS currents (identical to the output rectifier current) as a function of D_{MAX} for the ST204A, which is rated for an output currents of 2A. At D_{MAX} = 40%, RMS and peak secondary currents are 2.6A, and 4.2A, respectively, while at D_{MAX} = 60%, they rise to 3.3A and 7.8A. The voltage drop and power dissipation across the output diode is a function of the peak current through the diode. Rating the output diode current at 3 times the DC output current is consistent with the actual operating conditions. The output diode chosen for the ST204A (Sanken FMB-29L) is a 90 V Schottky diode rated at 8 amperes.

Output Capacitors

In flyback power supplies, the output filter capacitors carry a large amount of ripple current, as they must completely support the output current of the supply while the primary switch is on. For a supply operating in continuous mode, the ripple current in the output filter capacitor can be estimated as

$$I_{RIPPLE} \cong I_O \times \sqrt{\frac{D_{MAX}}{1 - D_{MAX}}} \qquad (6)$$

Operating in the continuous mode reduces the peak and RMS ripple current in the output capacitors, and consequently the peak and output ripple voltage at the switching frequency. Dissipation in the output capacitors can be calculated as:

$$P = I_{RIPPLE}^2 \times (ESR) \qquad (7)$$

where I_{RIPPLE} is the RMS capacitor ripple current in amperes, and ESR is the equivalent series resistance of the capacitor in ohms. For long life, under continuous duty conditions the capacitor current rating should be sized at 1.5 to 2 times the ripple current. The temperature and frequency derating factors published in the capacitor specifications should be taken into account when sizing the output capacitors, as well as the duty factor of the supply (continuous or intermittent operation). For intermittent operation, a less stringent derating factor can be used.

Summary

General Guidelines

- Design the power supply to run at the highest recommended D_{MAX}. For 100/115 VAC input supplies using TOP100 series parts and 230 V input supplies using TOP200 series parts, recommended D_{MAX} is 40%. For universal input supplies using TOP200 series parts, recommended D_{MAX} is 60%. Choose D_{MAX} by adjusting the secondary to primary turns ratio for a set value of reflected secondary voltage (V_{OR}) during the *TOPSwitch* off time. For 100/115 VAC input supplies using the TOP100 series, V_{OR} = 60 V. For 230 VAC and universal input supplies, V_{OR} = 135 V.

Specific Design Considerations

Primary Circuit Losses

- Choose the transformer primary inductance to run in the continuous mode versus the discontinuous mode to reduce peak and RMS primary currents.

- Choose the inrush limiting thermistor to run as hot as possible (given sensible component deratings) to reduce power loss in the thermistor.

- Choose an adequately rated input diode bridge to reduce power loss.

- Size the input filter capacitor at 3 µF/output watt for 100 VAC and universal input, 1 µF/watt for 230 VAC input, and 2 µF/watt for 115 VAC only input with or without a voltage doubler.

- Optimize the *TOPSwitch* for cost and efficiency.

- Use snubbing and damping networks only as required to keep leakage spikes and EMI within limits.

Transformer Losses

- Size the windings in the transformer for low DC loss.

- Use multiple strands of small wire or foil windings where necessary to combat skin effect (AC) losses.

- Choose core geometry and winding technique carefully to minimize leakage inductance.

- Keep the AC flux swing in the transformer to reasonable limits to avoid excessive core loss, use a low loss core material, and if there is room, select a larger core size to reduce core loss.

Secondary Losses

- Choose an output rectifier with a current rating at least 3 times the continuous output current rating. Use secondary rectifiers with the lowest forward voltage drop and shortest recovery time that cost allows. Use Schottky rectifiers when possible.

- Eliminate snubbers on the secondary rectifiers, or design for the minimum acceptable snubber dissipation to keep leakage spikes to within safe limits.

- Derate the output filter capacitors for 1.5-2 times the output capacitor ripple current for low losses and long life under continuous operation.

APPENDIX A

Peak Primary Currents for Discontinuous-mode Versus Continuous-mode Flyback

There are two basic modes of operation for a flyback converter. The first is the discontinuous mode, in which the energy stored in the flyback transformer is completely discharged each switching cycle. The second is the continuous mode, where the transformer begins each switching cycle with a non-zero value of stored energy. Representative primary current waveforms are shown for each case in Figure 19. Each mode has its particular advantages, but the continuous flyback runs at a lower value of primary and secondary current for the same operating parameters than a discontinuous flyback. In application notes AN-16 and AN-17, K_{RP} is defined as the ratio of the primary ripple current to the primary peak current. K_{RP} can be defined in terms of the initial primary current (I_I), and the peak current (I_P) as follows:

$$K_{RP} = \frac{I_P - I_I}{I_P} = \frac{I_R}{I_P}$$

I_R denotes the ripple current of the primary current waveform. K_{RP} can be used to describe quantitatively the degree of continuous operation of a flyback supply. K_{RP} can assume values from zero to unity. K_{RP}=1 indicates discontinuous operation, while K_{RP}<1 indicates continuous operation. K_{RP}=0 implies the ultimate extreme for continuous operation, with an infinite primary inductance and a rectangular primary current waveform. For a given input voltage range, the value of K_{RP} can be selected within the suggested range to optimize either the *TOPSwitch* or the transformer size. A low value of K_{RP} indicates more continuous operation, with a relatively high primary inductance. This results in low peak and RMS primary currents, and optimizes the design for the smallest possible *TOPSwitch*, at the cost of a larger transformer. A larger K_{RP} will result in less continuous operation, lower primary inductance, and higher primary peak and RMS currents. This optimizes the transformer size at the expense of a larger *TOPSwitch*. A K_{RP} range of 0.4-1.0 is recommended for 100/115 VAC. and universal input *TOPSwitch* supplies, while a K_{RP} of 0.6-1.0 is optimum for 230 VAC supplies.

Two design examples are given below showing the differences in peak and RMS current for a universal input power supply running in the discontinuous mode at K_{RP} = 1, and in the continuous mode at K_{RP} = 0.4. For a more detailed description of flyback operation, please refer to AN-16 and AN-17.

Discontinuous-mode Operation

Operating Parameters:

- K_{RP} = 1.0
- V_{MIN}= Minimum DC bus voltage = 90 VDC
- D_{MAX} = Duty cycle at V_{MIN} = 0.6
- P_O = output power = 30 W
- η= efficiency = 80%

I_P can be expressed as a function of I_R and K_{RP}, and also as a function of basic supply parameters and I_R, as shown below:

$$I_P = \frac{I_R}{K_{RP}} = \frac{P_O}{V_{MIN} \times D_{MAX} \times \eta} + \frac{I_R}{2}$$

Using the above equation to solve for I_P, the result is:

$$I_P = \frac{P_O \times \frac{2}{2 - K_{RP}}}{V_{MIN} \times D_{MAX} \times \eta}$$

For V_{MIN} = 90VDC, D_{MAX} = 0.6, P_O = 30W, K_{RP} = 1.0, and η = 0.8, I_P is:

$$I_P = \frac{30 \times \frac{2}{2-1}}{90 \times 0.6 \times 0.8} = 1.39A$$

Solving for I_{RMS}:

$$I_{RMS} = I_P \times \sqrt{D_{MAX} \times \left(\frac{K_{RP}^2}{3} - K_{RP} + 1 \right)}$$

$$= 1.39 \times \sqrt{0.6 \times \left(\frac{1}{3} - 1 + 1 \right)} = 0.62A$$

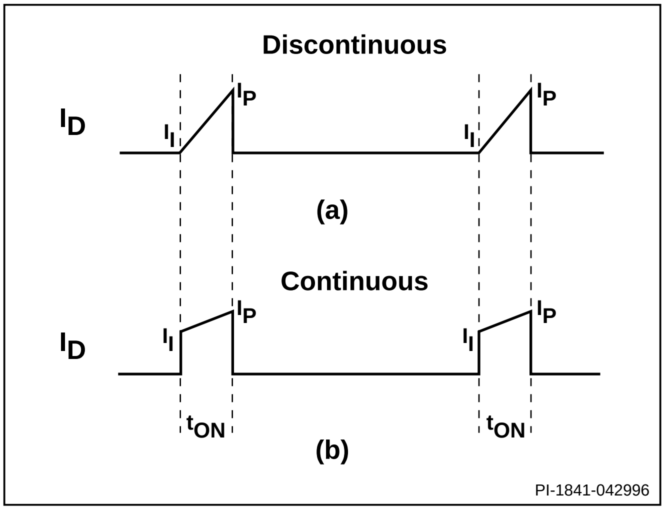

Figure 19. Comparison of Primary Current Waveforms for (a) Discontinuous, and (b) Continuous Mode Operation.

Continuous-mode Operation

Operating Parameters:

- K_{RP} = 0.4
- V_{MIN}= Minimum DC bus voltage = 90 VDC
- D_{MAX} = Duty cycle at V_{MIN} = 0.6
- P_O = output power = 30 W
- η = efficiency = 80%

The initial conditions are the same as before, only the K_{RP} is now 0.4, denoting highly continuous operation. With K_{RP} = 0.4, the new value for I_P is:

$$I_P = \frac{30 \times \dfrac{2}{2-0.4}}{90 \times 0.6 \times 0.8} = 0.87A$$

The new value of I_{RMS} is:

$$I_{RMS} = 0.87 \times \sqrt{0.6 \times \left(\frac{0.4^2}{3} - 0.4 + 1 \right)} = 0.54A$$

The peak current for the continuous mode of operation is 63% of the peak current for the discontinuous mode. The RMS current for continuous operation is 87% of the RMS current for the discontinuous case, resulting in 24% less switch dissipation for a given *TOPSwitch*. Continuous mode operation could allow the use of a smaller *TOPSwitch* for the same output power, or allow the *TOPSwitch* to run with lower losses. For a continuous design, the AC component of the primary current is lower than for a discontinuous design, reducing skin effect and core losses in the transformer.

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