Flyback Transformer Design For *TOPSwitch*® Power Supplies

Application Note AN-17

POWER INTEGRATIONS, INC.

When developing *TOPSwitch* flyback power supplies, transformer design is usually the biggest stumbling block. Flyback transformers are not designed or used like normal transformers. Energy is stored in the core. The core must be gapped. Current effectively flows in either the primary or secondary winding but never in both windings at the same time.

Why use the flyback topology? Flyback power supplies use the least number of components. At power levels below 75 watts, total flyback component cost is lower when compared to other techniques. Between 75 and 100 Watts, increasing voltage and current stresses cause flyback component cost to increase significantly. At higher power levels, topologies with lower voltage and current stress levels (such as the forward converter) may be more cost effective even with higher component counts.

Flyback transformer design, which requires iteration through a set of design equations, is not difficult. Simple spreadsheet iteration reduces design time to under 10 minutes for a transformer

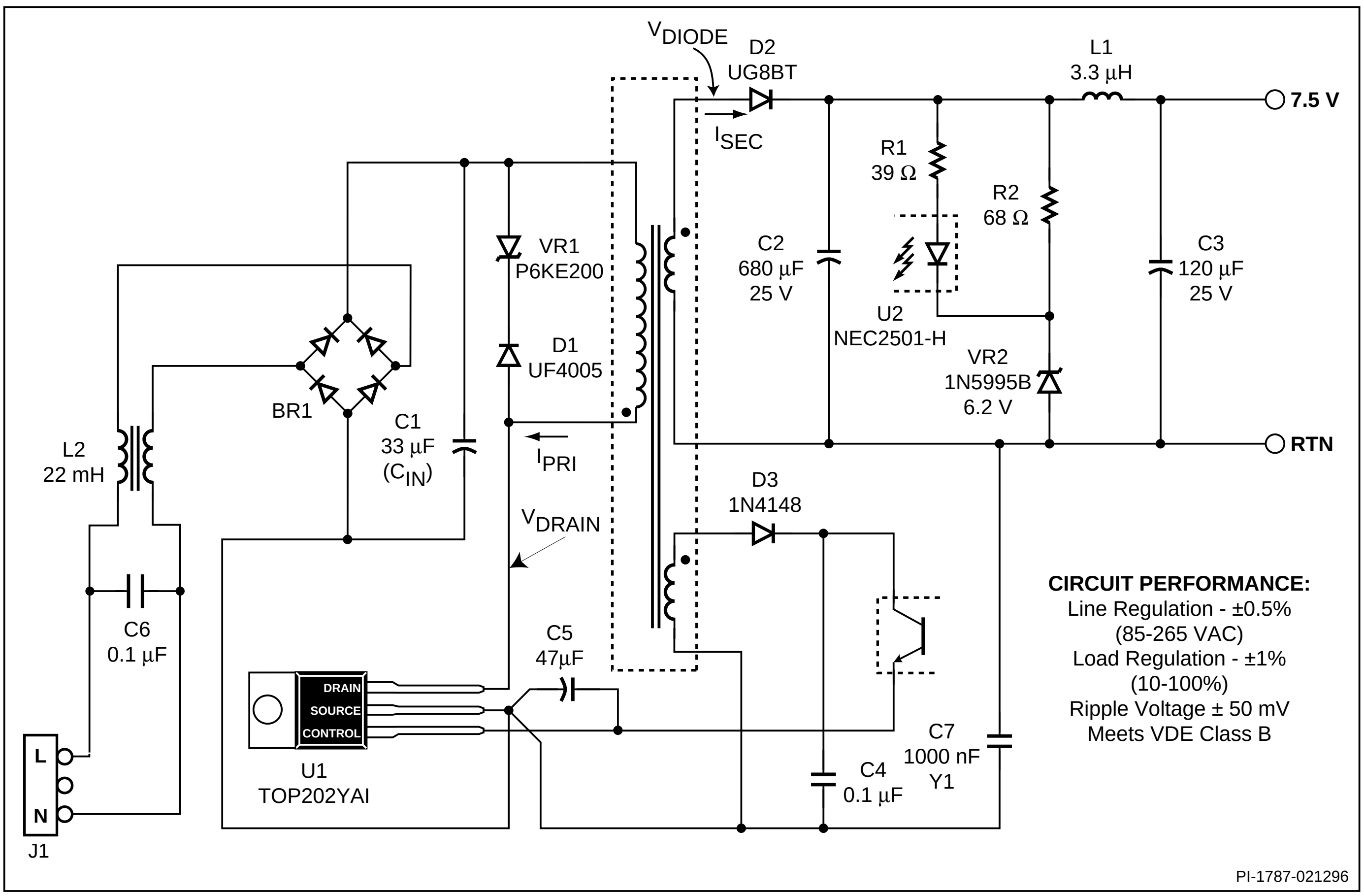

Figure 1. ST202A Power Supply Operates from Universal Input Voltage and Delivers 15 Watts.

that usually works the first time. This method, used for continuous mode as well as discontinuous mode designs, has three distinct steps:

1) Identify and estimate a set of independent variables (input) depending on application details, transformer core, and selected *TOPSwitch*.

2) Identify and calculate a set of dependent parameters (output).

3) Iterate specified independent variables until selected dependent parameters fall within defined limits for a practical flyback transformer.

A simple PC spreadsheet (available from Power Integrations for Excel or compatible spreadsheet programs) automates the transformer design method presented in this application note. (Note: this improved version has been completely revised and may give slightly different answers compared to earlier versions. Refer to the last page of this application note for a complete description of the changes.)

A new parameter, the ratio of primary ripple current to peak current (K_{RP}), is introduced to describe the *TOPSwitch* drain current waveform shape and simplify subsequent calculations such as RMS current and AC flux density.

Application specific independent variables include minimum and maximum AC input voltage, line frequency, *TOPSwitch* switching frequency, output and bias voltages, output power, bridge rectifier conduction time, size of input energy storage capacitor, power supply efficiency and power loss allocation between primary and secondary circuitry. Variables depending on the transformer core and construction include effective core cross sectional area and magnetic path length, ungapped effective inductance, bobbin physical winding width, margin width (for creepage distance and safety isolation), number of primary layers, and number of secondary turns. Variables depending on *TOPSwitch* include switching frequency, reflected output voltage, ripple to peak current ratio, and *TOPSwitch* voltage drop.

For a given application and transformer core, 20 of these 23 independent variables will be calculated or estimated once and then remain fixed during iteration. Only three variables, number of secondary turns N_S, ripple to peak current ratio K_{RP}, and number of primary winding layers L will be changed during the iteration process.

Dependent parameters are divided into four groups: DC input voltage, primary current waveform shape, transformer design, and voltage stress. DC input voltage parameters are simply the minimum and maximum DC input voltage after the AC mains have been rectified and filtered. Primary current waveform shape parameters include maximum duty cycle, average current, peak current, ripple current, and RMS current to completely define transformer primary current and determine operation in either continuous or discontinuous mode. Transformer design parameters include primary inductance, number of primary turns, number of bias winding turns, gapped effective inductance, maximum flux density, AC flux density, ungapped core relative permeability, estimated gap length, effective bobbin width, insulated primary wire diameter, insulation thickness, bare conductor cross section, primary current capacity, and secondary design parameters. Voltage stress parameters determine the maximum *TOPSwitch* off-state drain voltage and output rectifier peak inverse voltage.

Of all these dependent parameters, only three require examination and comparison within limits during iteration. Maximum flux density B_M, gap length L_G, and primary current capacity CMA are checked with each iteration until all three parameters are within specified limits. The remaining dependent parameters are either intermediate calculations or parameters used by the manufacturer for construction or the designer for specifying components.

Understanding primary and secondary current waveform shape in both continuous and discontinuous mode operation is necessary before beginning transformer design.

Figure 1 shows a typical flyback power supply using the TOP202 *TOPSwitch* from Power Integrations, Inc. *TOPSwitch* combines an integrated high voltage MOSFET switch with a complete switching power supply controller and protection circuitry in a single 3 pin TO220 package. The *TOPSwitch* power supply operates from 85 to 265 VAC and delivers 15 Watts at 7.5 Volt output. AC power is rectified and filtered by BR1 and C1 (C_{IN}) to create the high voltage DC bus applied to the primary winding of T1. The other side of the transformer primary is driven by *TOPSwitch*. D1 and VR1 clamp voltage spikes caused by transformer leakage inductance. D2, C2, L1, and C3 rectify and filter the power secondary. *TOPSwitch* bias voltage is provided by D3 and C4 which rectify and filter the bias winding. EMI filter components L2, C6, and C7 reduce conducted emission currents. Bypass capacitor C5 filters internal *TOPSwitch* gate charge current spikes and also compensates the control loop. Regulation is achieved when the output voltage rises sufficiently above Zener diode voltage (VR2) to cause optocoupler photodiode current to flow. Optocoupler phototransistor current flows into the *TOPSwitch* control pin to directly control the duty cycle and output voltage. R1 together with series impedances of VR2 and *TOPSwitch* determine the control loop DC gain. R2 and VR2 provide a slight preload to improve regulation at light loads.

Figures 2 and 3 show typical voltage and current waveforms taken from the same power supply delivering 15 Watts from 110 VAC input voltage but with different flyback transformer

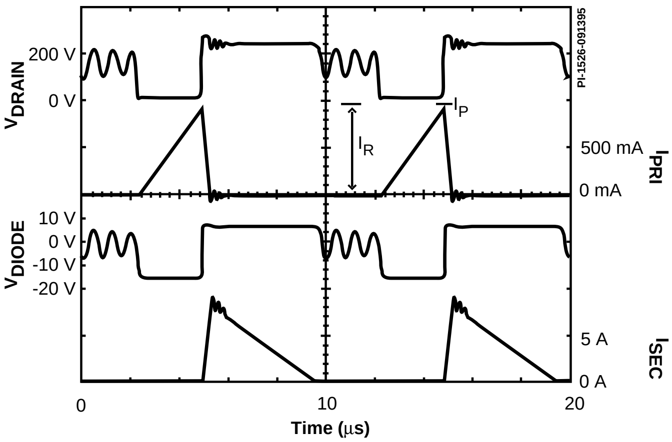

Figure 2. Voltage and Current Waveforms for Transformer Primary and Secondary in Discontinuous Mode.

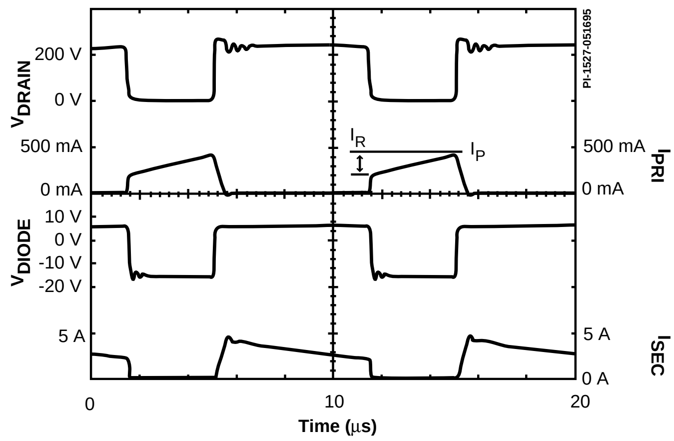

Figure 3. Voltage and Current Waveforms for Transformer Primary and Secondary in Continuous Mode.

primary inductance. *TOPSwitch* turns on to effectively apply the DC input voltage across the transformer winding with the "dot" side at lower potential than the "no-dot side". Primary current I_{PRI} increases linearly with a rate of change (di/dt) that varies directly with DC input voltage and inversely with primary inductance. Ripple current I_R is defined as the incremental linear current rise (di) over the entire *TOPSwitch* on time (t_{ON}). Peak primary current I_P is the final value occurring as *TOPSwitch* turns off. Energy, proportional to the square of peak current I_P, is stored by magnetic field in the transformer core as if the primary winding were a simple inductor. The secondary winding carries a reflected voltage proportional to primary voltage by turns ratio with the same "dot" polarity. While *TOPSwitch* is on, output diode D2 and bias diode D3 are reverse biased which prevents secondary current flow. When *TOPSwitch* turns off, the decreasing magnetic field induces an abrupt voltage reversal on all transformer windings such that the "dot" side is now higher potential than the "no-dot" side. Diode D2 and D3 become forward biased and secondary current rises quickly to a peak value (proportional by the inverse turns ratio to primary peak current I_P). Primary current immediately drops to zero. *TOPSwitch* drain voltage quickly rises to a voltage equal to the sum of the DC input voltage and reflected output voltage. Secondary winding current now linearly decreases at a rate that varies directly with output voltage and inversely with secondary inductance. Duty cycle is defined as the ratio of *TOPSwitch* on time t_{ON} to switching period T. D can also be calculated from t_{ON} and switching frequency f_S as shown.

$$D = \frac{t_{ON}}{T} = t_{ON} \times f_S$$

Figure 2 shows *TOPSwitch* and output diode triangular current waveforms which define "discontinuous" mode of operation resulting from low primary inductance. The secondary current linearly decreases to zero before *TOPSwitch* turns on again. The stored energy is completely delivered to the load. *TOPSwitch* drain voltage V_{DRAIN} relaxes and rings back towards the DC bus voltage when no current is flowing in either primary or secondary.

Figure 3 shows trapezoidal current waveforms which define "continuous" mode of operation resulting from high primary inductance. Secondary current is still flowing when *TOPSwitch* turns on at the beginning of the next cycle. The stored energy is not completely delivered to the load. Energy (due to non-zero magnetic field) remains in the core when *TOPSwitch* turns on again which causes the initial step in *TOPSwitch* current. Note that *TOPSwitch* drain voltage V_{DRAIN} stays at a high value equal to the sum of the DC input voltage and reflected output voltage until *TOPSwitch* turns on again.

Current never flows in the primary and secondary winding at the same time. Neither primary or secondary current is actually continuous. In flyback power supplies, continuous/discontinuous mode refers to magnetic field continuity in the transformer core over one complete switching cycle. (The flyback power supply is an isolated version of the simple buck-boost converter where continuous and discontinuous modes are easily defined by inductor current continuity.)

Each primary current waveform has a peak value (I_P), a ripple current value (I_R), an average or DC value (I_{AVG}), and an RMS value (I_{RMS}). I_P determines the number of primary turns and the core size necessary to limit peak flux density and must also be below *TOPSwitch* peak current limit. I_{AVG} is the average or DC primary current (as well as the power stage DC input current) which is proportional to output power. I_{RMS} causes power losses due to winding resistance and *TOPSwitch* R_{DSON}. The ratio (K_{RP}) of ripple current I_R to peak current I_P defines the continuous or discontinuous waveform. K_{RP} also simplifies subsequent calculations. Transformers designed for discontinuous operation have a higher peak current and a ripple current to peak current ratio K_{RP} of one. Practical continuous designs have lower peak currents and a ripple to peak current ratio K_{RP} of less than one but typically greater than 0.4. K_{RP} is inversely proportional to primary inductance so a continuous design with lower K_{RP} will have a higher inductance. Continuous transformer designs have a practical primary inductance upper limit approximately four times that of a discontinuous design at the same input voltage and output power due to the difference in peak currents and value of K_{RP}.

The primary current waveforms shown in Figures 2 and 3 deliver the same output power and therefore (assuming equal efficiency) must have equal I_{AVG}. The discontinuous current waveform has a higher peak value and therefore must have a higher RMS current value. Discontinuous mode requires less inductance and reduces transformer size but operates with higher losses and lower efficiency due to higher RMS currents. Continuous mode requires higher inductance and larger transformer size but offers improved efficiency and lower power losses. The trade-off between transformer size and power supply efficiency depends on the packaging and thermal environment in each application.

Some control loop comments regarding continuous mode are in order here. Most designers tend to avoid the continuous mode whenever possible because the feedback control loop is more difficult to analyze. Discontinuous mode power supplies are modeled with a single pole response and are simple to stabilize. Continuous mode offers improved efficiency, reduced losses, lower component temperatures, or higher output power but analysis is more difficult because a right half plane zero and complex pole pair all shift with duty cycle. However, stabilizing a continuous mode *TOPSwitch* power supply is quite straightforward. Adequate phase margins are achievable over all line and load combinations because the 70% maximum *TOPSwitch* duty cycle DC_{MAX} (from the data sheet) limits right half plane zero and complex pole pair migration. Phase margin is generally higher than expected once the damping effect of effective series power path resistance and output capacitor ESR is taken into account. Crossover bandwidths of 1 KHz (or wider) are easily achievable with phase margins of at least 45 degrees. Refer to AN-14 for circuit techniques to use in continuous mode designs.

Transformer core, winding, and safety issues must also be discussed before beginning design.

Transformer core and construction parameters depend on the selected core and winding techniques used in assembly. Physical height and cost are usually most important when selecting cores. This is especially true in AC mains adapter power supplies normally packaged in sealed plastic boxes. Applications allowing at least 0.75 inches of component height can use low cost EE or EI cores from Magnetics, Inc., Japanese vendors TDK and Tokin, or European vendors Philips, Siemens, and Thomson. Applications requiring lower profile can benefit from EFD cores available from the European vendors. EER cores offer a large window area, require few turns, and have bobbins available with high pin counts for those applications requiring multiple outputs. ETD cores are useful in the higher power designs when space is not a problem. PQ cores are more expensive but take up slightly less PC board space and require less turns than E cores. Safety isolation requirements make pot cores, RM cores, and toroids generally not suitable for flyback power supplies operating from the AC mains.

Flyback transformers must provide isolation between primary and secondary in accordance with the regulatory agencies of the intended market. For example, information technology equipment must meet the requirements of IEC950 in Europe and UL1950 in the U.S. These documents specify creepage and clearance distances as well as insulation systems used in transformer construction. 5 to 6 mm creepage distance is usually sufficient between primary and secondary (check with the appropriate agency and specification). Isolation is usually specified by electric strength and is tested with a voltage of typically 3000 VAC applied for 60 seconds. Two layers of insulation (Basic and Supplementary) can be used between primary and secondary if each layer exceeds the electric strength requirement. Three layers of insulation (reinforced) can also be used if all combinations of two layers (out of total three layers) meets the electric strength requirement.

Figure 4a shows the margin winding technique used in most flyback transformers. The margin is usually constructed with layers of tape slit to the width of the desired margin and wrapped in sufficient layers to match the winding height. The margin is generally half the required primary to secondary creepage distance (2.5 mm in this example). Cores and bobbins should be selected large enough that the actual winding width is at least twice the total creepage distance to maintain transformer coupling and reduce leakage inductance. The primary is wound between the margins. To reduce the risk of interlayer voltage breakdown due to insulation abrasion, improve layer to layer insulation, and decrease capacitance, the primary layers should be separated by at least one layer of UL listed polyester film tape (3M 1298) cut to fit between the margins. Impregnation with varnish or epoxy can also improve the layer to layer insulation and electric strength but does not reduce capacitance. The bias winding may then be wound over the primary. Supplementary or reinforced insulation consisting of two or three layers of UL listed polyester film tape cut to the full width of the bobbin may then be wrapped over the primary and bias windings. Margins are again wound. The secondary winding is wound between the margins. Another two or three layers of tape is added to secure the windings. Insulation sleeving may be needed over the leads of one or all windings to meet creepage distance requirements at lead exits. Nylon or

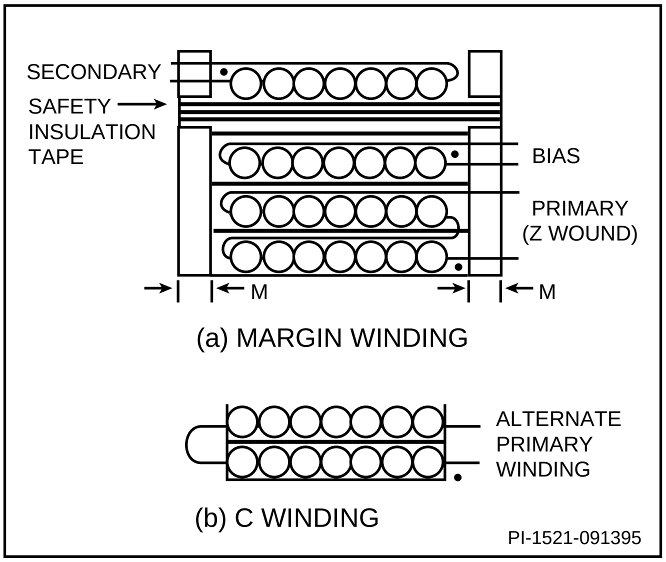

Figure 4. Margin Wound Transformer.

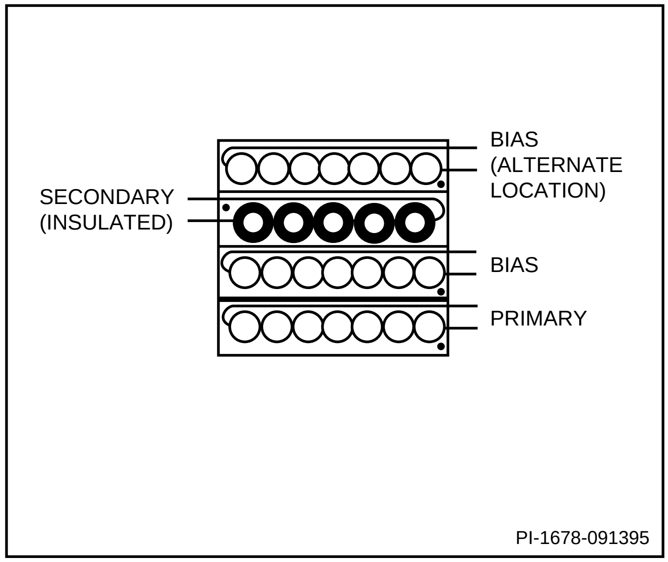

Figure 5. Triple Insulated Wire Wound Transformer.

Teflon sleeving with a minimum wall thickness of 0.41 mm should be used to meet the safety agency requirements. Consider the core as isolated dead metal (which means the core is conductive but not part of any circuit and safely insulated from the consumer). The sum of distance from primary winding (or lead exits) to the core added to the distance from the core to the secondary (or lead exits) must be equal to or greater than the required creepage distance.

Both Z winding (Figure 4a) and C winding (Figure 4b) techniques for multiple primary layers are shown. Note that the "dot" side which connects to *TOPSwitch* is buried under the second layer for self shielding to reduce EMI (common mode conducted emission currents). Z winding decreases transformer capacitance, decreases AC *TOPSwitch* losses, and improves efficiency but is more difficult and costly to wind. The C winding is easier and lower cost to wind but at the expense of higher loss and lower efficiency.

Figure 5 shows a new technique using double or triple insulated wire on the secondary to eliminate the need for margins (insulated wire sources can be found at the end of this application note). In double insulated wire, each layer is usually capable of meeting the electric strength requirement of the safety agency. In triple insulated wire, all three combinations of two layers taken together must usually meet the electric strength requirement. Special care is necessary to prevent insulation damage during winding and soldering. This technique reduces transformer size and eliminates the labor cost of adding margins but has higher material cost and may increase winding costs. The primary winding is wound over the full width of the bobbin flange. The bias winding can be wound if desired over the primary. One layer of tape is usually necessary between primary or bias and secondary to prevent abrasion of the insulated wire. The double or triple insulated wire is then wound. Another layer of tape is added to secure insulated winding.

Figure 5 also shows an alternate position for the bias winding wound directly over the secondary to improve coupling to the secondary winding and reduce leakage inductance (to improve load regulation in bias winding feedback circuits). Note that because the bias winding is a primary circuit, margin wound transformers must have another layer of supplementary or reinforced insulation between the secondary and alternate bias winding.

Refer to AN-18 for more information regarding transformer construction guidelines.

Flyback transformer design now begins by specifying the three groups of independent variables shown in the spreadsheet (Figure 6).

Application Variables:

Output power P_O, output voltage V_O, AC mains frequency f_L, *TOPSwitch* switching frequency f_S (100KHz), minimum (V_{ACMIN}), and maximum (V_{ACMAX}) AC mains voltage come directly from the application.

For efficiency (η), start with an estimate based on measurements in similar power supplies (or use a value of 0.8 if data is unavailable).

Efficiency can be used to calculate total power loss P_L in the power supply as shown below. Some power losses occurring in series primary components such as the bridge rectifier, common

	A	B	C	D	E	F
1		**INPUT**		**OUTPUT**		
2	**ENTER APPLICATION VARIABLES**					
3	VACMIN	85			Volts	Minimum AC Input Voltage
4	VACMAX	265			Volts	Maximum AC Input Voltage
5	fL	60			Hertz	AC Mains Frequency
6	fS	100000			Hertz	TOPSwitch Switching Frequency
7	VO	7.5			Volts	Output Voltage
8	PO	15			Watts	Output Power
9	n	0.8				Efficiency Estimate
10	Z	0.5				Loss Allocation Factor
11	VB	10.4			Volts	Bias Voltage
12	tC	3.2			mSeconds	Bridge Rectifier Conduction Time Estimate
13	CIN	33			uFarads	Input Filter Capacitor
14						
15	**ENTER TOPSWITCH VARIABLES**					
16	VOR	85			Volts	Reflected Output Voltage
17	VDS	10			Volts	TOPSwitch on-state Drain to Source Voltage
18	VD	0.4			Volts	Output Winding Diode Forward Voltage Drop
19	VDB	0.7			Volts	Bias Winding Diode Forward Voltage Drop
20	**KRP**	**0.92**				Ripple to Peak Current Ratio (0.4 < KRP < 1.0)
21						
22	**ENTER TRANSFORMER CORE/CONSTRUCTION VARIABLES**					
23		**EE22-Z**				Core Type
24	AE	0.41			cm^2	Core Effective Cross Sectional Area
25	LE	3.96			cm	Core Effective Path Length
26	AL	2400			nH/T^2	Ungapped Core Effective Inductance
27	BW	8.43			mm	Bobbin Physical Winding Width
28	M	0			mm	Safety Margin Width (Half the Primary to Secondary Creepage Distance)
29	**L**	**2**				Number of Primary Layers
30	**NS**	**5**				Number of Secondary Turns
31						
32	**DC INPUT VOLTAGE PARAMETERS**					
33	VMIN			93	Volts	Minimum DC Input Voltage
34	VMAX			375	Volts	Maximum DC Input Voltage
35						
36	**CURRENT WAVEFORM SHAPE PARAMETERS**					
37	DMAX			0.51		Duty Cycle at Minimum DC Input Voltage (VMIN)
38	IAVG			0.20	Amps	Average Primary Current
39	IP			0.74	Amps	Peak Primary Current
40	IR			0.68	Amps	Primary Ripple Current
41	IRMS			0.32	Amps	Primary RMS Current
42						
43	**TRANSFORMER PRIMARY DESIGN PARAMETERS**					
44	LP			623	uHenries	Primary Inductance
45	NP			54		Primary Winding Number of Turns
46	NB			7		Bias Winding Number of Turns
47	ALG		215		nH/T^2	Gapped Core Effective Inductance
48	**BM**			**2085**	Gauss	Maximum Flux Density (2000 < BM < 3000)
49	BAC		959		Gauss	AC Flux Density for Core Loss Curves (0.5 X Peak to Peak)
50	ur		1845			Relative Permeability of Ungapped Core
51	**LG**			**0.22**	mm	Gap Length (Lg >> 0.051 mm)
52	BWE		16.86		mm	Effective Bobbin Width
53	OD			0.31	mm	Maximum Primary Wire Diameter including insulation
54	INS		0.05		mm	Estimated Total Insulation Thickness (= 2 * film thickness)
55	DIA			0.26	mm	Bare conductor diameter
56	AWG			30	AWG	Primary Wire Gauge (Rounded to next smaller standard AWG value)
57	CM		102		Cmils	Bare conductor effective area in circular mils
58	**CMA**			**321**	Cmils/Amp	Primary Winding Current Capacity (200 < CMA < 500)
59						
60	**TRANSFORMER SECONDARY DESIGN PARAMETERS**					
61	ISP			7.95	Amps	Peak Secondary Current
62	ISRMS			3.36	Amps	Secondary RMS Current
63	IO			2.00	Amps	Power Supply Output Current
64	IRIPPLE			2.70	Amps	Output Capacitor RMS Ripple Current
65						
66	CMS		1079		Cmils	Secondary Bare Conductor minimum circular mils
67	AWGS			19	AWG	Secondary Wire Gauge (Rounded up to next larger standard AWG value)
68	DIAS			0.91	mm	Secondary Minimum Bare Conductor Diameter
69	ODS			1.69	mm	Secondary Maximum Insulated Wire Outside Diameter
70	INSS		0.39		mm	Maximum Secondary Insulation Wall Thickness
71						
72	**VOLTAGE STRESS PARAMETERS**					
73	VDRAIN			573	Volts	Maximum Drain Voltage Estimate (Includes Effect of Leakage Inductance)
74	PIVS			42	Volts	Output Rectifier Maximum Peak Inverse Voltage
75	PIVB			59	Volts	Bias Rectifier Maximum Peak Inverse Voltage
76						
77	**ADDITIONAL OUTPUTS**					
78	VX	12			Volts	Auxiliary Output Voltage
79	VDX	0.7			Volts	Auxiliary Diode Forward Voltage Drop
80	NX			8.04		Auxiliary Number of Turns
81	PIVX			68	Volts	Auxiliary Rectifier Maximum Peak Inverse Voltage
82						

Figure 6. Spreadsheet for ST202A Flyback Transformer Design.

mode choke, and *TOPSwitch* are not associated directly with energy stored in the flyback transformer core. The remaining power losses, occurring in the output rectifier and clamp Zener diode when energy is released from the flyback transformer, are now defined as secondary loss P_{LS}. Loss Allocation Factor Z, defined below as the ratio of secondary loss P_{LS} to total loss P_L, is a scaling factor which distributes the losses between primary and secondary. Loss allocation factor Z is typically between 0.4 and 0.6 which means that secondary loss P_{LS} is usually 40% to 60% of total power supply loss P_L.

$$P_L = P_O \times \left(\frac{1-\eta}{\eta}\right)$$

$$Z = \frac{P_{LS}}{P_L}$$

Bias voltage V_B is determined by the feedback control circuit and is usually between 10 volts and 30 volts (see AN-16).

For bridge rectifier conduction time t_C, 3 milliSeconds is typical (measure on a similar power supply or set equal to zero for a conservative first design).

For input filter capacitor C_{IN}, start with a standard value in microFarads between two and three times the output power in Watts (appropriate for universal or 115 VAC input). For example: 30µF to 45µF is a suitable capacitance range for a 15 Watt supply. 33µF is the lowest standard value within the range.

TOPSwitch Variables:

Reflected output voltage V_{OR} appears across the transformer primary when *TOPSwitch* is off and current is flowing through the secondary and output rectifier diode. Transformers optimized for *TOPSwitch* applications should be designed with a maximum reflected voltage V_{OR} of 60V or less for the TOP1XX series and 135V or less for the TOP2XX series. For more information, refer to AN-16.

V_{DS} is the on-state *TOPSwitch* voltage from the data sheet (typically 10 volts) at the specified value for peak *TOPSwitch* drain current I_P.

Output rectifier forward voltage drop V_D depends on output voltage. For lower output voltages (typically 8 Volts and below) a Schottky diode is commonly used and V_D is typically 0.4 Volts. In some cases, a Schottky diode can be used for output voltages as high as 12V depending on input voltage range and transformer turns ratio. For higher output voltage, an ultrafast recovery PN junction diode is normally used and V_D is typically 0.7 Volts.

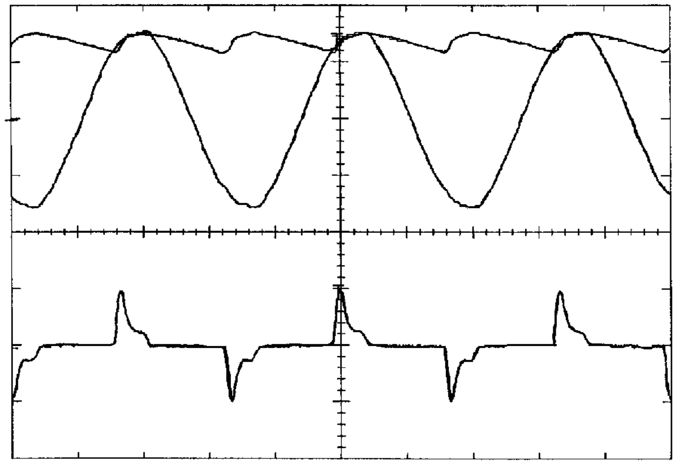

Figure 7. Bridge AC Current, AC Voltage, and DC Voltage Waveforms.

Bias winding diode forward voltage drop (V_{DB}) is also typically 0.7 Volts

Ripple current to peak current ratio K_{RP} determines how far into the continuous mode a flyback transformer will operate. Continuous mode transformers optimized for *TOPSwitch* applications operating from 100/115 VAC or universal input voltage should have a minimum K_{RP} of 0.4. Applications operating from 230 VAC input voltage should have a minimum K_{RP} of 0.6. Discontinuous mode transformers optimized for *TOPSwitch* applications always have a K_{RP} equal to 1.0.

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

Transformer Core/Construction Variables:

The following effective parameters are specified by the core and bobbin manufacturer in data sheets: cross sectional area A_e (cm^2), path length L_e (cm), ungapped inductance A_L (specified in either mH/(1000 turns)^2 or nH/T^2), and physical bobbin winding width B_W (mm).

Margin width M, determined by insulation methods and regulatory requirements discussed above, is usually between 2.5 to 3.0 mm for margin wound or set to zero for insulated wire wound transformers.

For number of layers L, one or two layers of primary winding are normally used. Higher number of layers increase cost, increase capacitance, reduce coupling, and increase leakage inductance.

Number of secondary turns N_S is a key iteration variable. One turn per Volt of output voltage is a good value to begin with for N_S (for example: start with 5 turns for a +5V output).

The four groups of dependent parameters can now be calculated.

DC Input Voltage Parameters:

Minimum DC input voltage V_{MIN} depends on the AC input voltage, bridge rectifier, and energy storage capacitor. Figure 7 shows how C_{IN} charges to the peak of the AC input voltage during a short conduction time t_C. Because of full wave rectification, C_{IN} has a ripple voltage at twice line frequency. C_{IN} must supply the entire average primary current during the discharge time between the peaks of the AC input voltage. Minimum DC voltage V_{MIN} can be found from the following equation where P_O is the power supply output power, η is an estimate of efficiency, f_L is line voltage frequency, V_{ACMIN} is the minimum AC mains voltage, C_{IN} is the value of the filter capacitor, and t_C is an estimate for conduction time. As an example, for 60 Hz, 85 VAC input voltage, efficiency of 0.8, 15 Watt output power, 33 uF input filter capacitance, and estimated conduction time of 3.2 mS, V_{MIN} is 93 Volts DC.

$$V_{MIN} = \sqrt{(2 \times V_{ACMIN}^2) - (\frac{2 \times P_O \times (\frac{1}{2 \times f_L} - t_C)}{\eta \times C_{IN}})}$$

$$= \sqrt{(2 \times 85^2) - (\frac{2 \times 15 \times (\frac{1}{2 \times 60} - 3.2mS)}{0.8 \times 33\mu F})} = 93V$$

Maximum DC input voltage V_{MAX} is simply the peak value of the highest AC input voltage (V_{ACMAX}) expected in the application. Operation from 265 VAC input results in a maximum DC bus voltage V_{MAX} of 375 Volts DC.

$$V_{MAX} = V_{ACMAX} \times \sqrt{2} = 265 \times \sqrt{2} = 375V$$

Current Waveform Shape Parameters:

D_{MAX} is the actual duty cycle occurring when the *TOPSwitch* power supply delivers maximum output power from minimum input voltage. D_{MAX} has an upper limit equal to the minimum value of the *TOPSwitch* Data Sheet parameter DC_{MAX} (64%). D_{MAX} is calculated from reflected voltage V_{OR}, minimum DC input voltage V_{MIN}, and *TOPSwitch* on-state Drain to Source voltage V_{DS}:

$$D_{MAX} = \frac{V_{OR}}{V_{OR} + (V_{MIN} - V_{DS})}$$

Average current I_{AVG} is calculated from minimum DC input voltage V_{MIN}, output power P_O, and efficiency η:

$$I_{AVG} = \frac{P_O}{\eta \times V_{MIN}}$$

Peak primary current Ip is calculated from average current I_{AVG}, ripple to peak current ratio K_{RP}, and maximum duty cycle D_{MAX}:

$$I_P = I_{AVG} \times \frac{2}{(2 - K_{RP}) \times D_{MAX}}$$

Ripple current I_R is calculated from average current I_{AVG}, peak primary current I_P, and maximum duty cycle D_{MAX}:

$$I_R = 2 \times (I_P - \frac{I_{AVG}}{D_{MAX}})$$

RMS current I_{RMS} is calculated from maximum duty cycle D_{MAX}, peak primary current I_P, and ripple to peak ratio K_{RP}. I_{RMS} can also be calculated directly from D_{MAX}, I_P, and ripple current I_R.

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

$$= \sqrt{D_{MAX} \times (I_P^2 - (I_P \times I_R) + \frac{I_R^2}{3})}$$

Transformer Design Parameters:

Primary inductance L_P (in μH) is determined by the flyback transformer energy equation defined below. The flyback transformer stores energy proportional to the square of primary current. When *TOPSwitch* is on, primary current linearly ramps up over a current range, defined earlier as ripple current I_R, and increases the energy stored in the flyback transformer core. When *TOPSwitch* turns off, the stored energy increment associated with ripple current I_R is delivered to the load and secondary losses (rectifier and clamp). Inductance L_P can now by calculated from output power P_O, efficiency η, loss allocation factor Z, peak current I_P, switching frequency f_S, and ripple current to peak current ratio K_{RP} (which determines I_R).

$$L_P = 10^6 \times \frac{P_O \times (\frac{(Z \times (1-\eta)) + \eta}{\eta})}{f_S \times I_P^2 \times K_{RP} \times (1 - \frac{K_{RP}}{2})}$$

Primary inductance L_P (in µH) can also be determined from a simple function of ripple current I_R, effective primary voltage (V_{MIN}-V_{DS}), maximum duty cycle D_{MAX}, and switching frequency f_S as shown below but the resulting value for primary inductance may be slightly different due to the selected value for loss allocation factor Z and *TOPSwitch* on-state Drain to Source voltage V_{DS}. The energy equation given above is preferred for selecting the value of inductance L_P while the ripple current equation given below is best for verifying the L_P value using in-circuit measurements.

$$L_{P(MEASURED)} = 10^6 \times \frac{(V_{MIN} - V_{DS}) \times D_{MAX}}{I_R \times f_S}$$

Number of primary turns N_P depends on number of secondary turns N_S, output voltage V_O, diode forward voltage drop V_D, effective primary voltage (V_{MIN}-V_{DS}), and maximum duty cycle D_{MAX}:

$$N_P = N_S \times \frac{V_{MIN} - V_{DS}}{V_O + V_D} \times \frac{D_{MAX}}{1 - D_{MAX}}$$

The number of bias winding turns N_B is calculated from the output voltage V_O, output diode voltage V_D, secondary number of turns N_S, target bias voltage V_B, and bias diode voltage V_{BD}:

$$N_B = \frac{V_B + V_{BD}}{V_O + V_D} \times N_S$$

A_{LG} is the effective inductance for the gapped core in nH/T^2. Some core vendors offer standard gapped core sets with specified A_{LG}. The transformer manufacturer either procures the gapped core for the given A_{LG} value or grinds the gap to meet the inductance specification in the finished transformer. A_{LG} is also used to simplify subsequent calculations. A_{LG} is calculated from primary inductance L_P (in µH) and number of primary turns N_P. Note that A_{LG} is specified in nH/(turn)^2.

$$A_{LG} = 1000 \times \frac{L_P}{N_P^2}$$

Maximum flux density B_M is a dependent iteration variable to be manipulated between the limits of 2000 and 3000 Gauss by varying number of secondary turns N_S which directly varies number of primary turns N_P as previously shown. B_M is calculated from peak current I_P, number of primary turns N_P, effective gapped inductance A_{LG}, and effective core cross sectional area A_e. B_M can also be calculated from effective primary voltage (V_{MIN}-V_{DS}), output voltage V_O, output diode voltage V_D, and maximum duty cycle D_{MAX}:

$$B_M = \frac{N_P \times I_P \times A_{LG}}{10 \times A_e}$$

$$= N_S \times \frac{I_P \times A_{LG}}{10 \times A_e} \times \frac{V_{MIN} - V_{DS}}{V_O + V_D} \times \frac{D_{MAX}}{1 - D_{MAX}}$$

B_{AC} is the AC flux density component. The equation gives peak AC flux density (rather than peak to peak) to use with core loss curves provided by the core vendor. B_{AC} can be calculated from maximum flux density B_M and ripple to peak current ratio K_{RP}. B_{AC} can also be calculated from effective primary voltage (V_{MIN}-V_{DS}), duty cycle, frequency, effective core cross sectional area, and number of primary turns N_P:

$$B_{AC} = \frac{B_M \times K_{RP}}{2} = \frac{(V_{MIN} - V_{DS}) \times D_{MAX} \times 10^8}{2 \times f_S \times A_e \times N_P}$$

Relative permeability μ_r of the ungapped core must be calculated to estimate the gap length L_g. μ_r is found from core parameters A_e (cm^2), L_e (cm), and ungapped effective inductance A_L:

$$\mu_r = \frac{A_L \times L_E}{0.4 \times \pi \times A_e \times 10}$$

Gap length L_g is the air gap ground into the center leg of the transformer core. Grinding tolerances and A_{LG} accuracy place a minimum limit of 0.051 mm on L_g. L_g (in mm) is calculated from number of primary turns N_P, core effective cross sectional area A_e, primary inductance L_P (in µH), core effective path length L_e, and relative permeability μ_r:

$$L_g = \left(\frac{0.4 \times \pi \times N_P^2 \times A_e}{L_P \times 100} - \frac{L_e}{\mu_r} \right) \times 10$$

Effective bobbin width BW_E takes into account physical bobbin width BW, margins M, and number of layers L:

$$BW_E = L \times (BW - (2 \times M))$$

Primary insulated wire diameter OD in mm is found from effective bobbin width BW_E and number of primary turns N_P:

$$OD = \frac{BW_E}{N_P}$$

The bias winding is usually wound with the same wire diameter as the primary to reduce the number of different wire gauges necessary for production.

Actual magnet wire outside diameter OD is slightly larger than the diameter DIA of the bare copper conductor. Insulation thickness varies inversely with bare copper conductor American Wire Gauge (AWG) size which means that smaller diameter conductors have thinner insulation thickness. Data from several different manufacturers were tabulated to generate an empirical expression for total insulation thickness INS (in mm) as a function of heavy insulated magnet wire outside diameter (in mm).

$$INS = (0.0594 \times LOG(OD)) + 0.0834$$

$$DIA = OD - INS$$

Another empirical equation determines the AWG for magnet wire with a given bare conductor diameter DIA. Integer AWG values are the standard sizes of available wire so the calculated AWG value should always be rounded up to the next integer or standard value (the next smaller standard conductor diameter) before proceeding with the current capacity or CMA calculation.

$$AWG = 9.97 \times (1.8277 - (2 \times LOG(DIA)))$$

Magnet wire for transformer winding usually has the cross sectional area specified in circular mils. A circular mil is the cross sectional area of a wire with a diameter of 1 mil (or 0.0254 mm). The effective cross sectional area in circular mils (CM) of a standard AWG size bare conductor wire is found from the following simple expression.

$$CM = 2^{\left(\frac{50-AWG}{3}\right)}$$

"Circular mils per Amp" or CMA is a convenient way to specify winding current capacity. CMA, which is the inverse of current density, is simply the ratio of cross sectional area in circular mils to the RMS value of primary current. CMA should be between 200 and 500 and is calculated from cross sectional wire area in CM and RMS primary current I_{RMS}.

$$CMA = \frac{CM}{I_{RMS}}$$

This completes all calculations necessary for the primary winding. Secondary peak current, RMS current, average output current, output capacitor ripple current, and secondary minimum and maximum conductor diameter must also be calculated.

Peak secondary current I_{SP} is a simple function of peak primary current I_P, primary turns N_P, and secondary turns N_S.

$$I_{SP} = I_P \times \frac{N_P}{N_S}$$

Secondary RMS current I_{SRMS} is found from maximum duty cycle D_{MAX}, secondary peak current I_{SP}, and ripple to peak current ratio K_{RP} (K_{RP} is identical for primary and secondary).

$$I_{SRMS} = I_{SP} \times \sqrt{(1 - D_{MAX}) \times (\frac{K_{RP}^2}{3} - K_{RP} + 1)}$$

Output current I_O is simply the ratio of output power P_O to output Voltage V_O:

$$I_O = \frac{P_O}{V_O}$$

Output capacitor ripple current I_{RIPPLE} is not a true transformer parameter but is needed for capacitor selection and easy to calculate from other transformer parameters. I_{RIPPLE} is found from secondary RMS current I_{SRMS} and output current I_O.

$$I_{RIPPLE} = \sqrt{I_{SRMS}^2 - I_O^2}$$

Minimum secondary bare conductor diameter DIA_S (in mm) based on previously calculated current capacity CMA and secondary RMS current must be determined.

From the primary CMA and secondary RMS current I_{SRMS}, the minimum secondary bare conductor CM_S is calculated.

$$CM_S = CMA \times I_{SRMS}$$

Minimum secondary AWG_S is then calculated from another empirical equation. Secondary calculated wire gauge AWG_S is always rounded down to the next integer value which selects the next larger standard wire size.

$$AWG_S = 9.97 \times (5.017 - LOG(CM_S))$$

(Secondary conductors larger than 26 AWG should not be used due to skin effects. Refer to AN-18 for suggestions on parallel conductor techniques.)

Bare conductor diameter (in mm) is now determined.

$$DIA_S = \sqrt{\frac{4 \times 2^{\left(\frac{50-AWG_S}{3}\right)}}{1.27 \times \pi}} \times \frac{25.4}{1000}$$

The maximum wire outside diameter OD_S (in mm) for a single layer based on number of secondary turns and bobbin width must also be calculated:

$$OD_S = \frac{BW - (2 \times M)}{N_S}$$

Secondary wire insulation thickness can now be calculated from the bare conductor outside diameter (determined by CMA) and the insulated wire outside diameter (determined by number of turns and effective bobbin width). Note that secondary insulation thickness INS_S (in mm) is the insulation wall thickness rather than the total insulation thickness used in the primary winding calculation.

$$INS_S = \frac{OD_S - DIA_S}{2}$$

Obviously, if insulation thickness INS_S is not a positive number, another transformer design iteration is necessary with either more secondary layers, a smaller number of secondary turns, or a transformer core with a wider bobbin.

For insulated wire secondaries, INS_S must be equal to or greater than insulation thickness of the selected wire.

Parallel combinations of wire with half the diameter may be easier to wind and terminate but the effective secondary CMA will be half the value of the single winding.

Voltage Stress Parameters:

Maximum drain voltage is the sum of maximum DC input voltage V_{MAX}, an estimated drain clamp voltage term based on V_{OR}, and an estimated voltage term related to typical blocking diode forward recovery. Refer to AN-16 for more detail.

$$V_{DRAIN} = V_{MAX} + (1.4 \times 1.5 \times V_{OR}) + 20V$$

Maximum peak inverse voltage PIV_S for the output rectifier is determined by transformer primary and secondary number of turns N_P and N_S, maximum DC input voltage V_{MAX}, and output voltage V_O.

$$PIV_S = V_O + (V_{MAX} \times \frac{N_S}{N_P})$$

Maximum peak inverse voltage PIV_B for the bias rectifier is determined from a similar equation using number of bias turns N_B.

$$PIV_B = V_B + (V_{MAX} \times \frac{N_B}{N_P})$$

Additional or auxiliary output winding number of turns N_X and rectifier diode peak inverse voltage PIV_X can be determined from the desired value for auxiliary output voltage V_X, auxiliary rectifier diode forward voltage drop V_{DX}, output voltage V_O, output rectifier diode forward voltage drop V_D, and number of secondary turns N_S.

$$N_X = \frac{V_X + V_{DX}}{V_O + V_D} \times N_S$$

$$PIV_X = V_X + (V_{MAX} \times \frac{N_X}{N_P})$$

Iteration can now be used to reach a final and acceptable solution for the flyback transformer design.

Iterate number of secondary turns N_S or primary ripple to peak current ratio K_{RP} until maximum flux density B_M is between indicated limits and check that gap length L_g is higher than indicated minimum value. B_M will decrease and L_g will increase as N_S or K_{RP} is increased.

Examine primary current capacity in Circular Mils per Amp (CMA). If CMA is below the specified lower limit of 200, consider increasing number of primary layers from one to two or use the next larger core size and perform new iteration. If CMA is greater than 500, consider using the next smaller core size. (CMA greater than 500 simply means that the wire diameter is oversized for the expected RMS current).

The transformer design is now complete. The transformer manufacturer needs the following information:

- Core part number and gapped effective inductance A_{LG}
- Bobbin part number
- Wire gauge and insulation style on all windings
- Safety or Electric strength and Creepage distance specifications
- Primary Inductance L_P
- Number of turns (N_P, N_S, N_B, etc.) for each winding
- Bobbin pin connections
- Winding layer placement and winding instructions
- Temperature class of operation (class A is 105 °C, class B is 130 °C, etc.)

Spreadsheet Improvements

The order of the spreadsheet has been changed to simplify the iteration process. Reflected voltage V_{OR} and ripple to peak current ratio K_{RP} are now independent variables which make peak current Ip and duty cycle D_{MAX} dependent variables. Loss allocation factor Z is introduced to distinguish between power losses occurring before energy is stored in the transformer (primary losses) and power losses occurring after energy is released from the transformer (secondary losses). Primary inductance L_P is now calculated from output power P_O, K_{RP}, efficiency η, and loss allocation factor Z. The spreadsheet now takes into account primary magnet wire insulation thickness as well as the discrete steps of standard AWG wire sizes. Metric dimensions are used throughout (with the exception of Circular mils for wire cross sectional area). Drain Voltage V_{DRAIN} now includes an estimate for the effect of leakage inductance induced voltage spikes on typical primary clamp circuits.

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McLyman, C., Transformer and Inductor Design Handbook, Marcel Dekker, Inc. 1978

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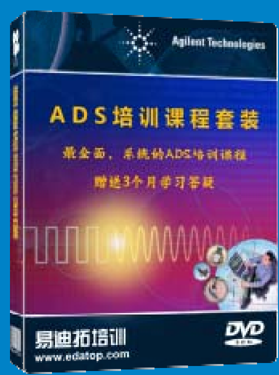

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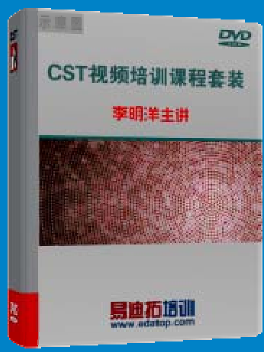

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