

# Designing Multiple Output Flyback Power Supplies with *TOPSwitch*<sup>®</sup>

## Application Note AN-22



### Introduction

Many *TOPSwitch* flyback power supply applications require two or more outputs to supply a variety of secondary circuits. Typical consumer applications of these multiple output converters include television and related products such as set-top decoders and video cassette recorders (VCRs). Industrial applications generally require a number of outputs to supply

analog and digital low voltage circuitry. Motor control applications often require several separately isolated outputs to supply half-bridge drivers and control circuitry.

When compared to single output flyback supplies, multiple output applications demand further design considerations to

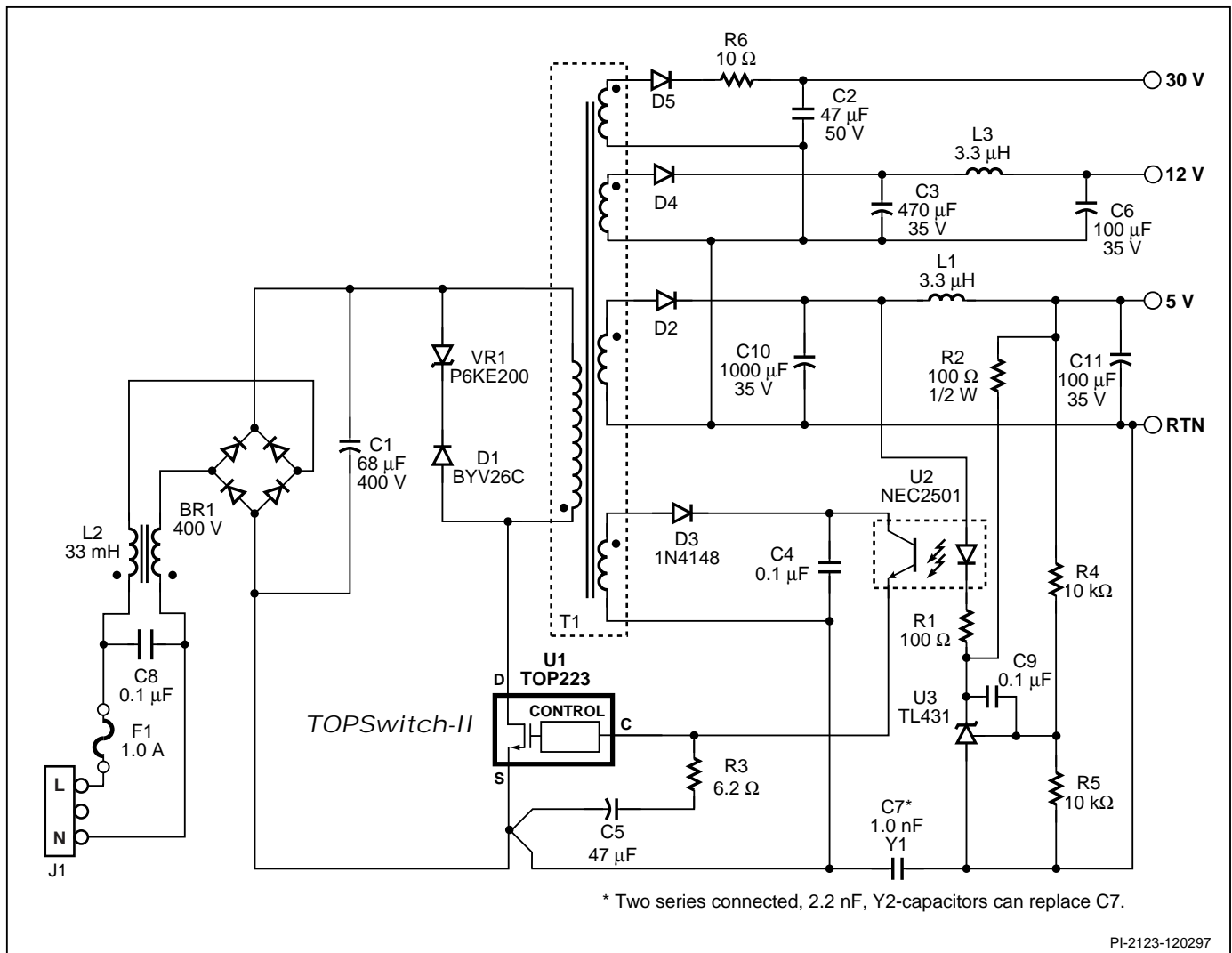


Figure 1. Schematic Diagram of 85-265 VAC, 25 W Power Supply Using TOP223.

POWER SUPPLY SPECIFICATIONS		
Input Voltage:		85-265 VAC
Output 1	Voltage	5 VDC $\pm$ 5%
	Current	0.40 A to 2.00 A
Output 2	Voltage	12 VDC $\pm$ 10%
	Current	0.12 A to 1.20 A
Output 3	Voltage	30 VDC $\pm$ 10%
	Current	0.01 A to 0.02 A
Total Output Power:		25 W

Table 1. Outline Power Supply Specification.

optimize the performance. The design of multiple output power supplies always requires some breadboarding to verify transformer designs, feedback techniques and system behavior.

This Application Note provides guidelines to streamline the decision making process and to reduce development effort for an optimized design. An example multiple output power supply design illustrates the procedure. All essential aspects are considered.

The design begins with system specifications that define regulation requirements, followed by selection of an appropriate feedback scheme. It then moves to calculation of transformer parameters and application of construction techniques specific to multiple output supplies, aided by reference to Application Notes AN-17 and AN-18 for detailed descriptions.

A discussion of output cross regulation includes measurements and test results. Additional EMI considerations are presented with reference to AN-15 and AN-16. There is also a listing of general tips which may be appropriate to specific designs.

Appendix A provides some additional reminders for use of the transformer design spreadsheet, while Appendix B contains special techniques for use with output voltages of 3.3 V and 5 V. Appendix C gives complete construction details of the transformer used in the hardware examples.

## Design Procedure

The design procedure for multiple output power supplies is a simple extension of the single output case. The circuitry on the primary side of the transformer is the same for either application. Additional steps in the design for multiple outputs are needed only to calculate turns ratios and wire sizes for the extra windings. Transformer construction has more degrees of freedom than in the single output case. The designer can apply several circuit techniques to adjust output regulation characteristics as needed.

POWER SUPPLY FEEDBACK TECHNIQUES			
Feedback Technique	Output Regulation		
	Main Output	Other Outputs	Notes
Primary (Basic or Enhanced)	$\pm$ 10%	Wider than $\pm$ 10%	Any lightly loaded output may be post-regulated to get $\pm$ 5% or better regulation
Opto/Zener	$\pm$ 5%	Wider than $\pm$ 10%	With 2% Zener
Opto/TL431	$\pm$ 5%	Tighter than $\pm$ 10%	Proportional feedback from two or more outputs optional

Table 2. Choice of Feedback Technique Depends on Requirements for Output Regulation.

## Regulation Requirements

Specification of the regulation requirements on all outputs is essential to successful design of the circuit configuration and transformer. Requirements differ significantly depending on the application.

One output usually requires tighter regulation than the others. Usually the 5 V supply for logic circuitry requires regulation of  $\pm 5\%$  or less, while other outputs have a wider tolerance of typically  $\pm 10\%$ . Many applications now require both 3.3 V and 5 V outputs, with  $\pm 5\%$  regulation specifications. There are several techniques which can be used to achieve this performance, and they are discussed in more detail in Appendix B of this application note.

While a 5 V output may have the most stringent regulation specification, a different winding often has a higher output load specification. Consideration must therefore be given to the required cross regulation between these outputs, because it will influence the transformer winding technique for an optimum design.

Table 1 gives an outline specification for a 25 W power supply with three outputs. Note that the 5 V output has the highest current and the tightest regulation, but the 12 V output delivers the highest power. The techniques presented here can be extended to any number of outputs. Some specific considerations for more outputs are discussed later.

The next step of the design is to determine the most appropriate feedback technique. As a quick reference for deciding the optimum feedback technique, Table 2 provides broad design rules which can be used, based on the required output tolerances of a specific application. If no tighter than  $\pm 10\%$  tolerance is required on all outputs, a primary side feedback scheme may be employed. This technique eliminates the need for an optocoupler by using the primary bias winding of the transformer to derive information about the regulated output on the secondary. This type of feedback scheme is detailed in AN-16. It is difficult, however, to achieve the output voltage tolerance of  $\pm 5\%$  with this scheme alone.

If outputs requiring  $\pm 5\%$  are only lightly loaded, primary side feedback may be used with a linear post regulator on these outputs at the expense of some drop in efficiency. From the specification in Table 1, however, the 2 A peak load on the 5 V output would lead to excessive dissipation in a linear regulator; therefore, the remainder of this application note will concentrate on feedback that uses an optocoupler.

There are two common techniques to generate a secondary reference with optocoupler feedback. The first uses a simple Zener diode as a secondary reference. This technique is described in the supporting literature for Power Integrations'

RD5 reference design board. The output voltage is determined by the Zener voltage, the forward voltage of the optocoupler's LED and the series resistor that sets the loop gain. A 2% tolerance Zener diode allows  $\pm 5\%$  tolerance on the regulated output voltage. However, it is often necessary to improve cross regulation by providing feedback from more than one output. The second technique uses a TL431 precision shunt regulator to offer more flexibility in such cases.

The TL431 precision shunt regulator integrates an accurate 2.5 V bandgap reference with an amplifier and driver into a single device. It is popular as a secondary referenced error amplifier. The TL431 also introduces the possibility of combining feedback from two or more outputs simultaneously to its reference pin. This can be a useful technique when it is required to employ one output as the primary source of feedback but also introduce a proportion of the feedback from another output. This advanced technique is described in more detail later.

This application note, therefore, focuses on the use of the TL431 shunt regulator. Figure 1 shows a schematic in a typical application with an optocoupler to provide tight regulation on the 5 V output of a multiple output power supply.

## Transformer Design

The choice of *TOPSwitch* and calculation of the primary transformer characteristics is independent of the number of outputs. As such, the Power Integrations standard transformer design spreadsheets (available from your local Power Integrations representative or on the Power Integrations Web Page at [www.powerint.com](http://www.powerint.com)) can be used to define the basic transformer specification in terms of the transformer core, primary inductance, primary turns and the output volts per turn. This basic design can then be extended to define the turns and wire selection on other outputs.

Two spreadsheets are available: one for discontinuous conduction mode (DCM) designs and one for continuous conduction mode (CCM) designs. Refer to AN-16 and AN-17 in the Power Integrations 1996-97 Data Book and Design Guide for further explanation of converter operation and use of the spreadsheets.

Operation in DCM results in smaller transformer core sizes for a given output power, but the smallest size is often not the most desirable choice in multiple output power supplies. Transformer hardware is usually selected to allow optimum circuit board layout. This motivation drives the selection of a transformer bobbin with the best arrangement of the number of pins and the pin spacing.

Designing for CCM provides the optimum utilization of the *TOPSwitch* silicon for a given output power. Therefore, this



	A	B	C	D	E	F
1	Rev 2.1	INPUT		OUTPUT		CONTR2P1.XLS: TOPSwitch Continuous Flyback Transformer Design Spreadsheet
2	ENTER APPLICATION VARIABLES					AN-22
3	VACMIN	8.5			Volts	Minimum AC Input Voltage
4	VACMAX	26.5			Volts	Maximum AC Input Voltage
5	fL	50			Hertz	AC Mains Frequency
6	fS	100000			Hertz	TOPSwitch Switching Frequency
7	VO	5			Volts	Output Voltage
8	PO	2.5			Watts	Output Power
9	n	0.8				Efficiency Estimate
10	Z	0.5				Loss Allocation Factor
11	VB	12			Volts	Bias Voltage
12	tC	3			mSeconds	Bridge Rectifier Conduction Time Estimate
13	CIN	68			uFarads	Input Filter Capacitor
14						
15	ENTER TOPSWITCH VARIABLES					
16	VOR	11.0			Volts	Reflected Output Voltage
17	ILIMITMAX	1.65	TOP224		Amps	From TOPSwitch Data Sheet
18	VDS	1.0			Volts	TOPSwitch on-state Drain to Source Voltage
19	VD	0.7			Volts	Output Winding Diode Forward Voltage Drop
20	VDB	0.7			Volts	Bias Winding Diode Forward Voltage Drop
21	KRP	0.45				Ripple to Peak Current Ratio (0.4 < KRP < 1.0)
22						
23	ENTER TRANSFORMER CORE/CONSTRUCTION VARIABLES					
24		ETD29				Core Type
25	AE	0.76			cm^2	Core Effective Cross Sectional Area
26	LE	7.2			cm	Core Effective Path Length
27	AL	2100			nH/T^2	Ungapped Core Effective Inductance
28	BW	19			mm	Bobbin Physical Winding Width
29	M	3			mm	Safety Margin Width (Half the Primary to Secondary Creepage Distance)
30	L	2				Number of Primary Layers
31	NS	4				Number of Secondary Turns
32						
33	DC INPUT VOLTAGE PARAMETERS					
34	VMIN			90	Volts	Minimum DC Input Voltage
35	VMAX			37.5	Volts	Maximum DC Input Voltage
36						
37	CURRENT WAVEFORM SHAPE PARAMETERS					
38	DMAX			0.58		Duty Cycle at Minimum DC Input Voltage (VMIN)
39	Iavg			0.35	Amps	Average Primary Current
40	IP			0.78	Amps	Peak Primary Current
41	IR			0.35	Amps	Primary Ripple Current
42	IRMS			0.46	Amps	Primary RMS Current
43						
44	TRANSFORMER PRIMARY DESIGN PARAMETERS					
45	LP			1339	uHenries	Primary Inductance
46	NP			77		Primary Winding Number of Turns
47	NB			9		Bias Winding Number of Turns
48	ALG		225		nH/T^2	Gapped Core Effective Inductance
49	BM			1771	Gauss	Flux Density at PO, VMIN
50	BP			3767	Gauss	Peak Flux Density (BP < 4200)
51	BAC		399		Gauss	AC Flux Density for Core Loss Curves (0.5 X Peak to Peak)
52	ur		1583			Relative Permeability of Ungapped Core
53	LG			0.38	mm	Gap Length (Lg >> 0.051 mm)
54	BWE		26		mm	Effective Bobbin Width
55	OD			0.34	mm	Maximum Primary Wire Diameter including insulation
56	INS		0.06		mm	Estimated Total Insulation Thickness (= 2 * film thickness)
57	DIA			0.28	mm	Bare conductor diameter
58	AWG			30	AWG	Primary Wire Gauge (Rounded to next smaller standard AWG value)
59	CM		102		Cmils	Bare conductor effective area in circular mils
60	CMA			219	Cmils/Amp	Primary Winding Current Capacity (200 < CMA < 500)
61						
62	TRANSFORMER SECONDARY DESIGN PARAMETERS					
63	ISP			14.98	Amps	Peak Secondary Current
64	IS RMS			7.62	Amps	Secondary RMS Current
65	IO			5.00	Amps	Power Supply Output Current
66	IRIPPLE			5.75	Amps	Output Capacitor RMS Ripple Current
67						
68	CMS		1667		Cmils	Secondary Bare Conductor minimum circular mils
69	AWGS			17	AWG	Secondary Wire Gauge (Rounded up to next larger standard AWG value)
70	DIAS			1.15	mm	Secondary Minimum Bare Conductor Diameter
71	ODS			3.25	mm	Secondary Maximum Insulated Wire Outside Diameter
72	INSS		1.05		mm	Maximum Secondary Insulation Wall Thickness
73						
74	VOLTAGE STRESS PARAMETERS					
75	VDRain			626	Volts	Maximum Drain Voltage Estimate (Includes Effect of Leakage Inductance)
76	PIVS			24	Volts	Output Rectifier Maximum Peak Inverse Voltage
77	PIVB			55	Volts	Bias Rectifier Maximum Peak Inverse Voltage
78						
79	ADDITIONAL OUTPUTS					
80	VX	12			Volts	Auxiliary Output Voltage
81	VDX	0.7			Volts	Auxiliary Diode Forward Voltage Drop
82	NX			8.91		Auxiliary Number of Turns
83	PIVX			55	Volts	Auxiliary Rectifier Maximum Peak Inverse Voltage

Figure 2. Spreadsheet to Design Transformers for Single Output and Multiple Output Flyback Converters.



example uses the spreadsheet for continuous conduction mode. The techniques described in the following sections to extend the standard single output transformer design to multiple outputs are the same for either spreadsheet.

## Spreadsheet Transformer Design

Figure 2 shows the spreadsheet for a transformer that meets the output power and input voltage specification of Table 1. A full explanation of the use of the spreadsheet is provided in AN-17, but a brief overview will suffice for this explanation.

The first section of the spreadsheet is used to input the application variables. Note that only the 5 V output is needed to determine the number of turns of the primary, while the total output power for all outputs is specified in this section to select the transformer core, primary inductance and wire gauge.

Initial design requirements may not be firm enough to determine which *TOPSwitch* will be used in the final product. The designer usually has to choose between two likely candidates (see AN-21). In all designs, whether single or multiple output, the transformer design should accommodate the largest *TOPSwitch* that might be used with it. A designer may find it necessary to use the larger *TOPSwitch* (with a lower on-resistance) to permit the use of a smaller heatsink, for example.

Thus, although the circuit of Figure 1 specifies the TOP223Y, the spreadsheet uses the upper current limit value for the TOP224Y/TOP224P. The higher value is used here to ensure flexibility to allow the use of the TOP224 should the application require it. The change may be necessary if mechanical restrictions in the available space of the power supply's enclosure force the use of a smaller heatsink.

The upper current limit is subsequently used in the spreadsheet to determine the peak flux density  $B_p$ , which should be limited to prevent excessive core saturation under overload and start up conditions.

The ferrite core used here is the industry standard ETD29. This is used as an example only. Other standard cores such as the EE or EER families can be substituted as desired.

The design is based on a margin wound construction, where 3 mm margins are provided at each side of the bobbin to give a total of 6 mm primary to secondary creepage distance. This is the standard creepage distance allowed for mains input power supplies meeting IEC950 (or equivalent) isolation. Local safety agency requirements for creepage and clearance should be obtained before committing a design to manufacture.

Other transformer construction techniques, such as slotted bobbin, concentric bobbin or the use of triple insulated wire,

are equally applicable. The bobbin style does not influence the calculation of the primary inductance, but specific bobbin width must be input to determine the physical space available for the primary winding. Although triple insulated wire techniques are not normally favored in applications requiring a high number of secondary turns, transformer suppliers should be consulted for advice on the optimum construction technique in a particular application.

The spreadsheet defines two layers for the primary winding to minimize construction costs. If other cores with reduced bobbin widths are used, additional layers may be necessary to satisfy recommendations for current capacity (CMA). It should be noted that an even number of layers will ease construction because the start and finish of the primary winding will be at the same side of the bobbin.

The remaining sections of the spreadsheet provide the transformer design that results from the input variables described above. The key parameters that must be checked before a design can be deemed acceptable are detailed in AN-17 and summarized in Appendix A.

Since the spreadsheet is written for single output supplies, the 'Transformer Secondary Design Parameters' show values assuming the total output power is provided by the 5 V output. It is therefore necessary to extend these calculations to account for the partitioning of output power defined in the power supply specification of Table 1. The following section provides the equations necessary to assign appropriate numbers of turns and wire gauges to each output.

## Calculation of Secondary Turns

From the spreadsheet, the 5 V output winding is defined as having 4 turns. The voltage on the cathode of D2 in Figure 1 is 5 V. Therefore, 4 turns produce the output voltage plus the forward drop of the output diode D2.

The volts per turn  $V_{PT}$  is defined as:

$$V_{PT} = \frac{(V_O + V_D)}{N_S} \quad (1)$$

where:

$V_{PT}$  = volts per turn

$V_O$  = output voltage (5 V)

$V_D$  = output diode forward voltage drop (typically 0.7 V for ultra fast PN power diodes and 0.4 V for Schottky diodes)



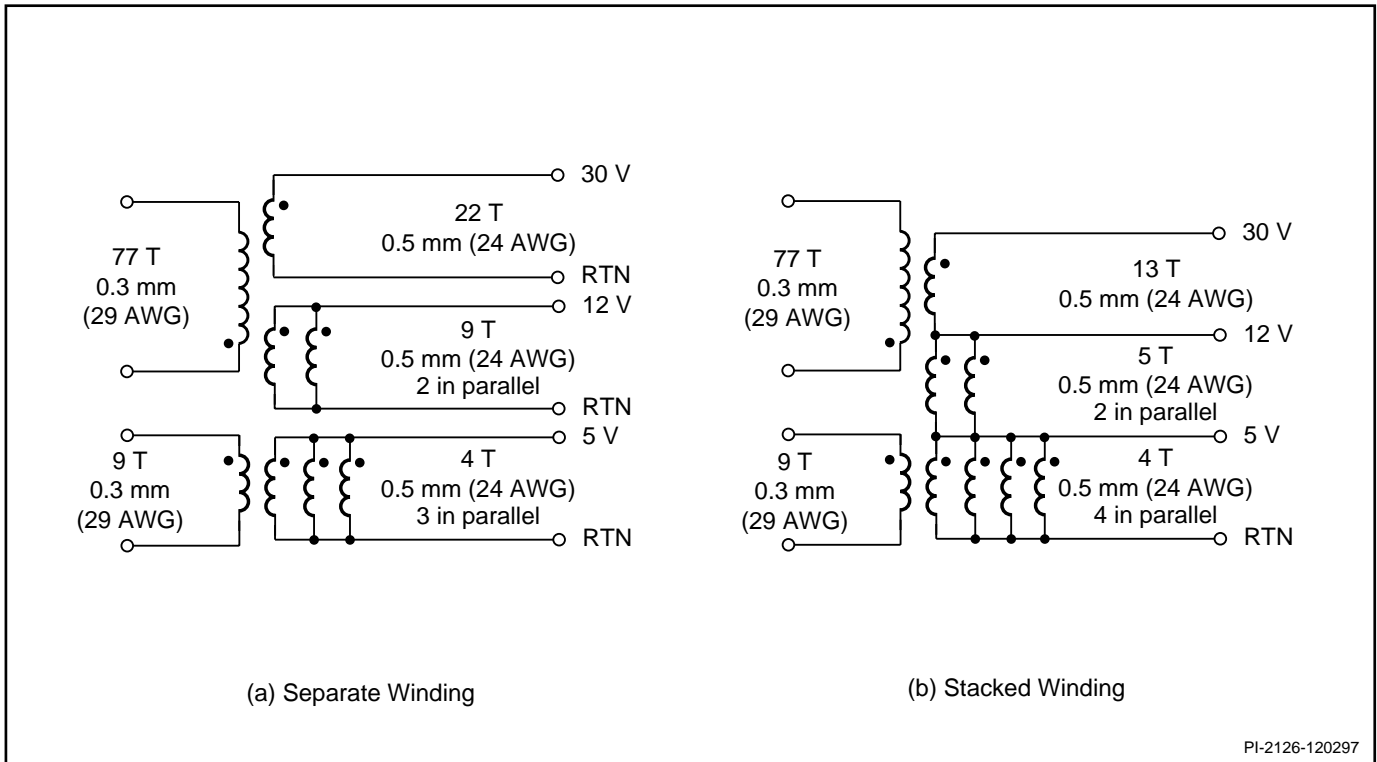


Figure 3. Transformer Winding Diagrams Showing Two Techniques for the Secondary Winding.

$N_s$  = number of secondary turns (4 turns for the 5 V output)

Substitution of these values into (1) gives:

$$V_{PT} = 1.43 \text{ V per turn}$$

This value is used to calculate the turns required by the other outputs.

Simple rearrangement of (1) gives:

$$N_s = \frac{(V_o + V_D)}{V_{PT}} \quad (2)$$

For the 12 V output,

$$V_o = 12 \text{ V}$$

$$V_D = 0.7 \text{ V}$$

Substituting in (2):

$$N_{s12} = \frac{(12 + 0.7)}{1.43} = 8.9 \text{ turns}$$

A practical transformer requires integer numbers of turns; therefore, the 12 V output uses 9 turns.

For the 30 V output,

$$V_o = 30 \text{ V}$$

$$V_D = 0.7 \text{ V}$$

Substituting in (2) gives:

$$N_{s30} = \frac{(30 + 0.7)}{1.43} = 21.5 \text{ turns}$$

Select 22 turns for the 30 V winding.

This last result highlights a frequently encountered problem in multiple output transformers. An integer number of turns, such as 21 or 22, will make the output voltage lower or higher respectively than desired. Since this is a high voltage output with a large number of turns, the difference between the desired value and the integer value amounts to only about 2%. The resulting change in output voltage is not significant, and will be masked by other factors such as cross regulation and diode characteristics. However, it is worth mentioning the options available should this problem be encountered with

lower voltage outputs where the requirement for integer numbers of turns can introduce a significant deviation from the desired value.

1. If the output in question requires a high degree of accuracy, then a higher output voltage can be defined in equation (2) and a linear post regulator employed to achieve the output voltage.
2. If the tolerance is less critical, a series resistor and a Zener diode of appropriate value can be used as a shunt regulator for low power outputs.
3. The fundamental transformer design could be modified such that the main 5 V output uses a number of turns which yields an integer number of turns on the other windings when calculated using equations (1) and (2).
4. The choice of rectifier on the main regulated output can be used to influence the volts per turn. If a Schottky diode with a forward voltage of typically 0.4 V were employed on the 5 V output, the  $V_{PT}$  from (1) would be 1.35. A standard PN diode on the 30 V output would from equation (2) yield 22.7 turns, which is closer to the integer number 23.

Use of the Schottky diode with 4 turns on the 5 V output, however, would decrease the accuracy of the 12 V output. The required number of turns would move farther away from an integer value, from 8.9 to 9.4 turns.

The designer can investigate alternative integer turns ratios with both Schottky and PN diodes by repeating the spreadsheet design for other values of secondary turns. If a need for higher efficiency calls for a Schottky diode on the 5 V output, then 3 turns on the 5 V output with 7 and 17 turns for the 12 V and 30 V outputs respectively may give acceptable results.

Designers often use the "golden ratios" of 3:7:9 with a Schottky diode for the 5 V output and a PN diode for the 12 V output, or 4:9:11 with all PN diodes to achieve outputs of 5, 12 and 15 V. Another useful ratio is 2:3 for outputs of 3.3 and 5 V with Schottky diodes on each. The turns could be in the ratio of 3:4 if the 3.3 V output uses a PN diode and the 5 V uses a Schottky diode. All designs need to be tested thoroughly to verify acceptability.

In practice, if tight tolerance is required on windings other than the main feedback output, some form of post regulation or combined feedback circuitry is often necessary. These issues of cross regulation are discussed later in the section on circuit performance.

In this case, as mentioned above, the choice of 22 turns for the 30 V output will not introduce a significant inaccuracy. The

final choice of turns on each output is therefore shown in Figure 3(a), and summarized as follows:

5 V	—	4 turns
12 V	—	9 turns
30 V	—	22 turns

Figure 3 illustrates two winding diagrams: one with separate windings for each output and one with stacked output windings. These two configurations are discussed in detail later in the section on transformer construction.

### Choice of Output Wire Gauge

Appropriate wire gauge for the outputs is determined on the basis of the maximum continuous RMS current rating for each winding. The analysis of the distribution of current in the various outputs can be very complex, but a few reasonable assumptions make the task easy.

The waveshapes of the currents in the individual output windings are determined by the impedances in each circuit. Leakage inductance, rectifier characteristics and capacitor values are some of the parameters that affect the magnitude and duration of the currents. The average currents are always equal to the DC load current, while the RMS values are functions of peak magnitudes and conduction times. The RMS values determine the power dissipation in the windings. For ordinary multiple output designs it is valid to make the reasonable simplifying assumption that all output currents have the same shape as for the single output case. This is the case of greatest dissipation.

Ultimately the final design of the transformer has to be decided on the basis of tests and consultation with transformer suppliers. However, the first order analysis that assumes the same waveshape for all output currents provides a start point for the choice of wire gauge.

The single output design of the spreadsheet calculates the RMS current in the secondary as if the 5 V winding supplied all the power. However, from the specification of Table 1, the 5 V output supplies a maximum of 10 W. The actual currents in the multiple output application are computed from quantities on the single output spreadsheet.

Since we assume the currents in the output windings have the same shape, each will have the same ratio of RMS to average as the single output case. If  $K_{RA}$  is the ratio of RMS to average current, then

$$K_{RA} = \frac{I_{SRMS}}{I_O} = \frac{7.62 \text{ A}}{5 \text{ A}} = 1.524 \quad (3)$$



where  $I_{SRMS}$  and  $I_o$  are from the spreadsheet.

To find the RMS current in a winding, we simply multiply its average current by  $K_{RA}$ .

$$I_{RMSX} = I_X \times K_{RA} \quad (4)$$

Hence, the RMS current in the 5 V winding is

$$I_{RMS5} = 2.0 \text{ A} \times 1.524 = 3.05 \text{ A}$$

and the RMS current on the 12 V winding is

$$I_{RMS12} = 1.2 \text{ A} \times 1.524 = 1.83 \text{ A}$$

Similar calculations for the 30 V output yield

$$I_{RMS30} = 30.5 \text{ mA}$$

The wire diameter can be chosen on the basis of the total dissipation in the output winding. One can find the resistance of the winding from the resistance per unit length of a particular wire gauge and the length of the wire associated with each output winding. However, a calculation based on the current density can be used to make a first estimate of the required wire gauge on each output.

A current density between 4 and 10 A/mm<sup>2</sup> (corresponding to

493 and 197 CMA respectively) allows acceptable power dissipation in the majority of applications, depending on the conditions of maximum ambient temperature and efficiency requirements. In the United States, it is common to use the reciprocal of current density expressed as circular mils per ampere (CMA). One mil is 0.001 inch, and the area in circular mils is the square of the wire diameter in mils. One circular mil is  $7.854 \times 10^{-7} \text{ in}^2$  or  $5.067 \times 10^{-4} \text{ mm}^2$ .

Based on 9 A/mm<sup>2</sup> (219 CMA), using the RMS current calculated above, the minimum bare copper diameter for each output is:

5 V output — 0.66 mm (22 AWG)

12 V output — 0.51 mm (24 AWG)

30 V output — 0.07 mm (41 AWG)

The above calculations define the minimum wire diameter specifications. However, practical considerations of transformer manufacture determine the actual wire gauges used. For example, two or three parallel windings on the higher current outputs can reduce the required wire diameter while optimizing coverage of the bobbin. These issues are discussed in detail next.

## Transformer Construction

Primary winding techniques are well documented in AN-18

WINDING TECHNIQUE	ADVANTAGES	DISADVANTAGES
Separate Output Windings	<ol style="list-style-type: none"> <li>1. Flexibility in winding placement; Output with highest current can be positioned closest to primary to minimize energy lost from leakage inductance.</li> </ol>	<ol style="list-style-type: none"> <li>1. Poor regulation of lightly loaded outputs due to peak charging.</li> <li>2. Generally higher manufacturing costs.</li> <li>3. More pins on bobbin.</li> </ol>
Stacked Output Windings	<ol style="list-style-type: none"> <li>1. Improved cross regulation.</li> <li>2. Generally lowest cost manufacturing technique.</li> </ol>	<ol style="list-style-type: none"> <li>1. Winding with lowest or highest voltage output must be placed closest to the primary winding – no flexibility to reduce leakage inductance of outputs with higher currents.</li> </ol>

Table 3. Comparison of Secondary Winding Techniques in Margin Wound Transformers.





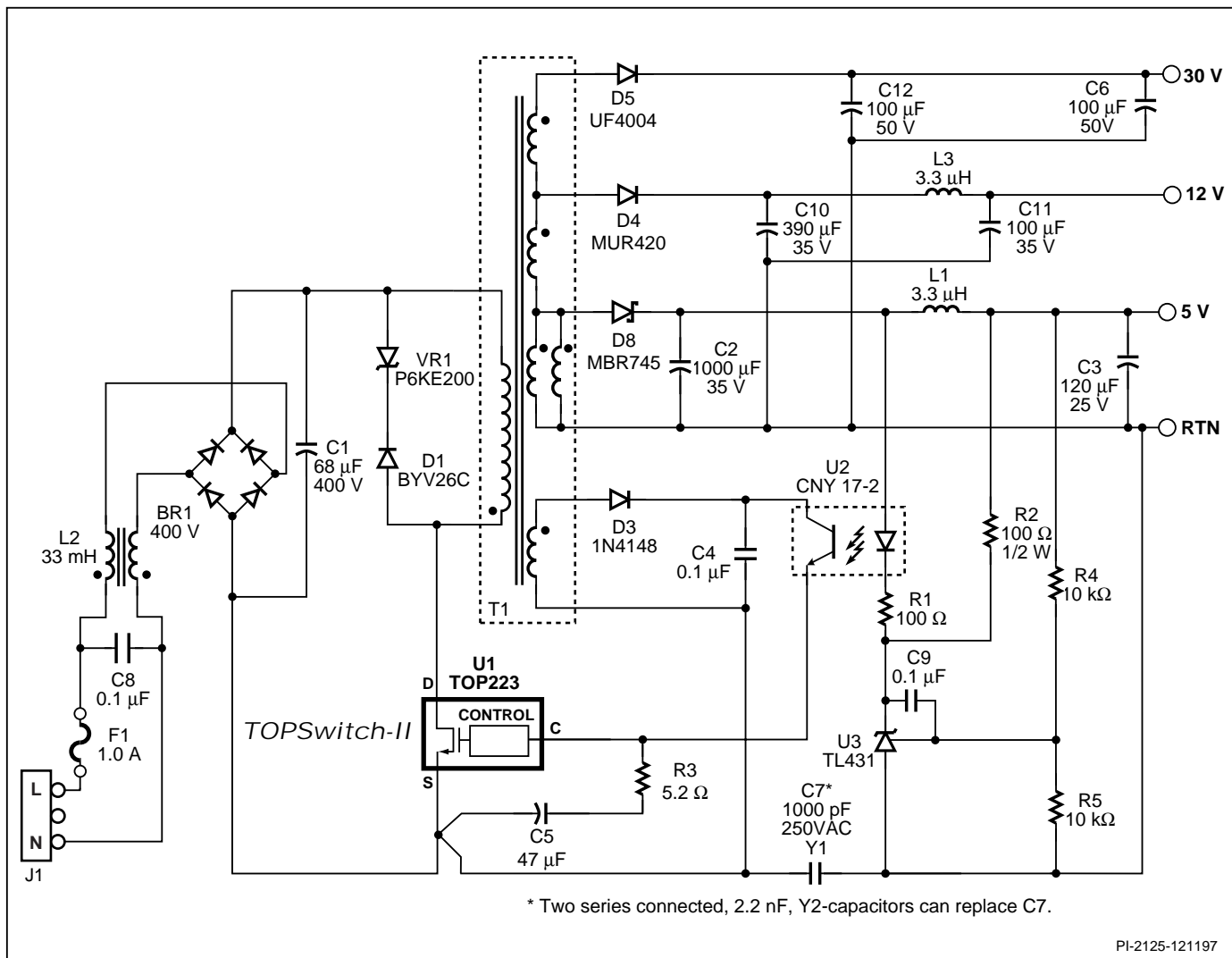


Figure 4. Schematic of Multiple Output 25 W Power Supply with Stacked Secondary Windings.

and are not influenced by the number of output windings. There are, however, two secondary winding techniques commonly used in margin wound transformers. These are described below and summarized in Table 3. Other transformer constructions such as slotted bobbin and concentric bobbin designs may demand other considerations. The designer should consult with the specific transformer supplier to insure the optimum technique in each case.

### Separate Output Windings

The winding diagram of Figure 3(a) shows each output wound as a separate coil. In this way each winding conducts only current associated with the specific load on that output. Since each output is wound as a separate operation, this construction technique provides flexibility in the placement of output windings relative to the primary winding. This freedom can be an important consideration in multiple output transformers to minimize the leakage inductance.

The leakage inductance of a transformer is the inductance associated with flux which does not link all windings. As such, this flux does not contribute to the transfer of energy. In single output transformer structures, all the leakage is usually measured on the primary by shorting the output winding and measuring the resulting inductance of the primary. This provides a good estimate of the energy which the primary clamp circuitry will dissipate. In Figure 1, components D1 and VR1 are specified for clamping the leakage energy.

However, in a multiple output design, there are leakage inductances associated with each output winding according to its coupling to the primary and to other secondary windings. Placement of the output windings should be made to minimize the leakage inductance associated with outputs that provide the most current. For example, in the circuit design of this application note, the 5 V and 12 V outputs handle most of the power with 2 A and 1.2 A respectively, while the 30 V output has a load of only 20 mA. The windings therefore should be

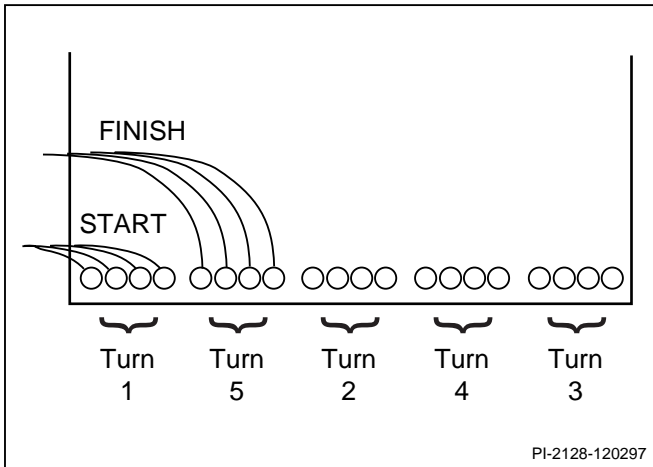


Figure 5. Cross Section of Bobbin Showing Five Interleaved Turns of Four Parallel Conductors on a Single Layer.

arranged such that the 5 V and 12 V outputs have the best coupling to the primary winding.

An arrangement that has the 30 V winding closest to the primary may show the same primary leakage inductance as the preferred structure when measured with the standard technique of shorting all outputs together. In the application, however, efficiency will be reduced since the leakage inductance associated with the 5 V and 12 V outputs will be higher.

The use of separate output windings provides complete flexibility in the winding arrangement. In this case the optimum configuration for separate layers might be to wind the 5 V output first followed by the 12 V winding and finally the 30 V output. That is, the winding with the greatest output current would go next to the primary. An even better arrangement would have the two highest current windings share a single layer using the nesting technique illustrated in Appendix C.

Separate windings, however, tend to increase the cost of the transformer since every output winding is a separate operation. The alternative stacking technique described below improves the regulation, particularly on lightly loaded outputs.

### Stacked Output Windings

Figure 3(b) shows a stacked output winding configuration, which is generally favored by transformer manufacturers. The windings of the 5 V output provide the return and part of the windings for the 12 V output. Similarly, the 30 V output uses the turns of the 5 and 12 V outputs and additional turns to make up the full winding. The wire for each output must be sized to accommodate its output current plus the sum of the currents for the other outputs stacked on top of it.

The stacked configuration improves cross regulation while

reducing construction costs. Consider this example where the 5 V output is fully loaded but the 12 V and 30 V outputs have minimum load applied. With separate output windings, the capacitors on the 12 V and 30 V outputs would tend to peak charge under the influence of leakage inductance. However, with a stacked winding, the fact that the 5 V output forms part of the 12 V and 30 V windings reduces the impedance of these windings and reduces the effect of peak charging.

The only disadvantage of this winding technique is that there is little flexibility in the placement of the windings relative to the primary. Either the 30 V or 5 V winding must form the start of the output windings closest to the primary. In this case, since the 5 V has the highest loading, it is defined as the start of the secondary winding.

Since the stacking technique generally offers the best cross regulation, the winding construction of Figure 3(b) was chosen for the example circuit in this application note, as illustrated in Figure 4. The only difference between T1 in Figures 1 and 4 is the use of the stacked winding technique on the transformer in Figure 4.

### Construction to Improve Cross Regulation

The cross regulation is a measure of how well the output voltages regulate under the influence of varying load conditions on other outputs. The quality of cross regulation depends on the coupling between the various output windings. The better coupled these windings are, the better the cross regulation.

As such, it is recommended that each individual winding is wound to cover the complete bobbin width. Therefore, the easiest way to wind the transformer is to use several parallel wires of the same gauge to insure the bobbin is well covered.

In this case, the total copper area used by the 5 V winding must handle the total RMS current of all outputs.

The total output RMS current is:

$$I_{RMSTOT} = I_{RMS5} + I_{RMS12} + I_{RMS30} = 5.03 \text{ A}$$

This summation is possible only when the currents have the same shape, which is a valid simplifying assumption for the design.

Based on a current density of 9 A/mm<sup>2</sup> (219 CMA), the copper diameter of a single wire would need to be 1.03 mm (20 AWG). However, if the wire is split into several parallel sections, each carrying an equal share of the current, we may use a smaller diameter wire which is much easier to handle during manufacture.

Also, the multiple parallel strands of thinner wire can be placed flat for good coverage of the bobbin as shown in Figure 5. This will insure that the winding is well coupled to the primary and to the other secondaries that are wound afterwards.

In this example we chose to split the 5 V winding into six conductors to fit the pin arrangement of the bobbin. One pin can accommodate three wires. Since each wire carries one sixth the current, or 0.84 A RMS, we may use a wire diameter of 0.4 mm (27 AWG), which is much easier to handle during manufacture.

The 12 V winding must handle a total of 1.86 A RMS ( $I_{RMS12} + I_{RMS30}$ ). To maintain a maximum current density below 9 A/mm<sup>2</sup> (219 CMA) we can use the same 0.4 mm (27 AWG) wire with the number of parallel strands reduced to 2. Again this should be wound evenly across the bobbin with turns distributed to provide the optimum coupling with the 5 V and primary winding. Appendix C shows how to put both windings on the same layer for best coupling.

Finally, the 30 V winding is added across the entire bobbin width. This winding carries the current for only the 30 V load; therefore, we can use a single strand of the 0.4 mm diameter (27 AWG) wire. If desired, a thinner wire gauge may be specified to reduce the volume occupied by the winding. The same wire may be used in all windings to reduce cost. Appendix C illustrates these methods with complete construction details of the transformer used in this Application Note.

The techniques detailed above should be used in the transformer construction to optimize cross regulation. However, additional external circuit techniques to further enhance cross regulation are discussed in the section on circuit performance.

## Output Rectifier Specification

As in single output converters, the proper choice of output rectifiers in multiple output converters is essential to achieve desired performance and reliability. It is important to use only Schottky and ultra fast PN junction rectifiers. The effects of the reverse recovery characteristics on the primary circuit are amplified in multiple output applications because the output rectifiers are effectively in parallel. Refer to AN-19 for a discussion of how the selection of output rectifiers influences efficiency.

The specification on each output rectifier diode is determined on the basis of the required voltage and current rating. The peak inverse voltage (PIV) on each diode is given by:

$$PIV_x = V_x + \left( V_{MAX} \times \frac{N_x}{N_p} \right) \quad (5)$$

where  $V_x$  is the voltage of the particular output,  $N_x$  is the number of output turns on the particular output and  $N_p$  is the transformer primary turns.  $V_{MAX}$  is the maximum primary DC rail voltage, which for 230 VAC input applications is typically 375 VDC (peak value of 265 VAC).

For the transformer in this example,

$$N_p = 77 \text{ turns}$$

$$V_{MAX} = 375 \text{ V}$$

Hence, for the 5 V output,

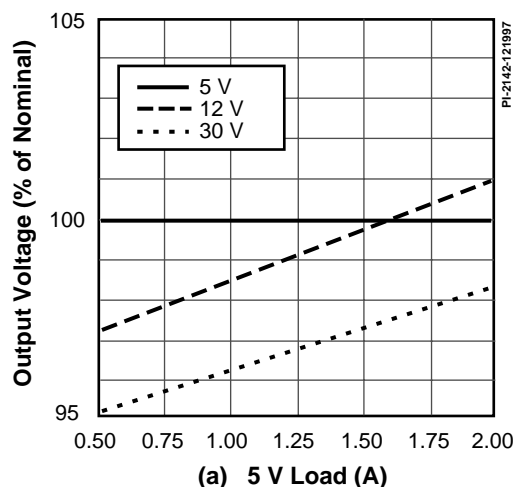


Figure 6 (a). Cross Regulation with Feedback from 5 V Only. Response to a 5 V Load.

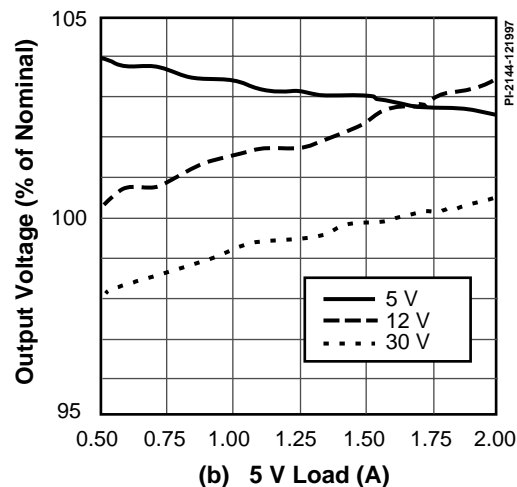


Figure 6 (b). Cross Regulation with Feedback from 5 V and 12 V. Response to a 5 V Load.



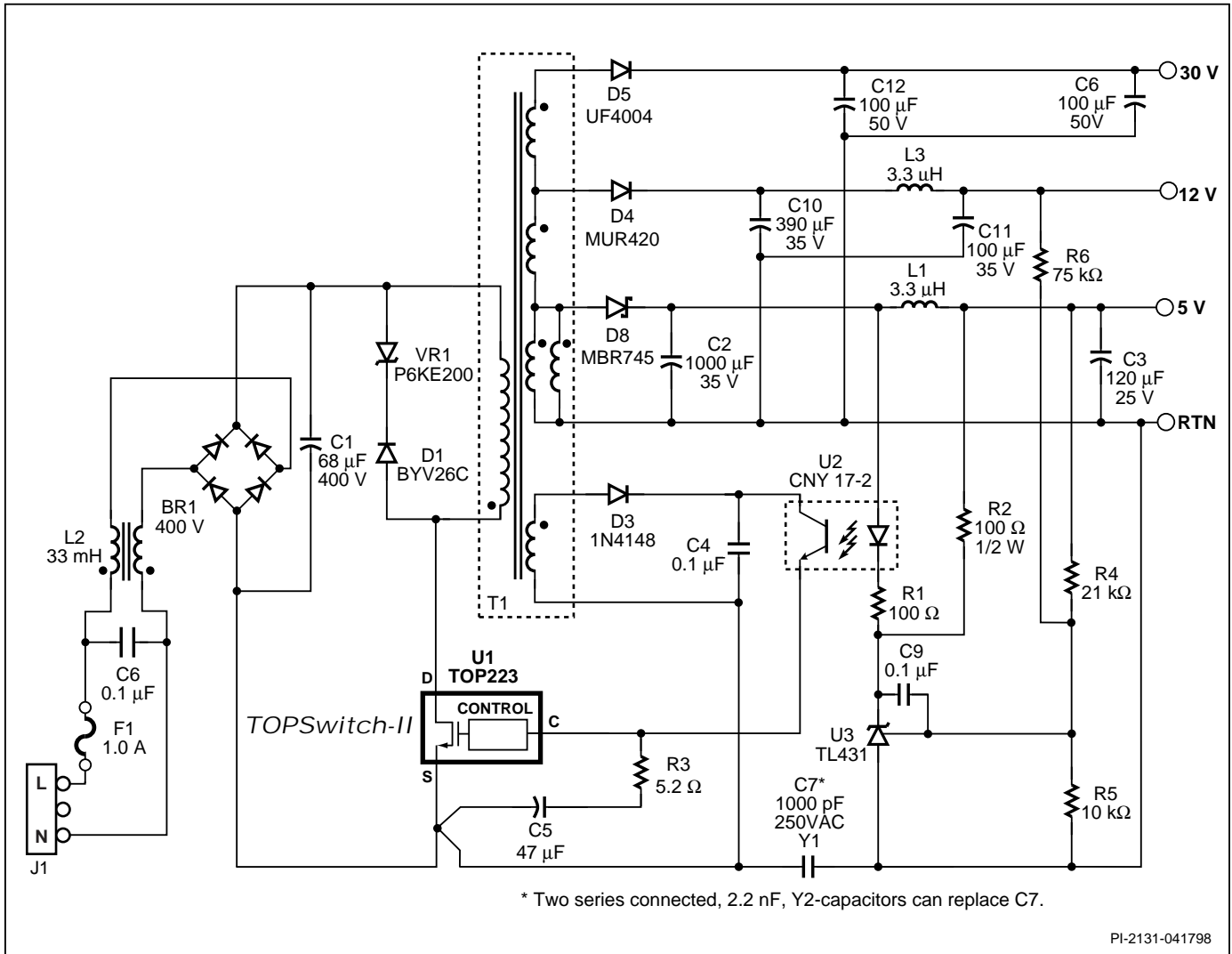


Figure 7. Modified Schematic with Feedback from Both 5 V & 12 V Outputs.

$$PIV_5 = 5 + \left( 375 \times \frac{4}{77} \right) = 25 \text{ V}$$

For the 12 V output,

$$PIV_{12} = 12 + \left( 375 \times \frac{9}{77} \right) = 56 \text{ V}$$

For the 30 V output,

$$PIV_{30} = 30 + \left( 375 \times \frac{22}{77} \right) = 137 \text{ V}$$

The diodes chosen for each output should have a reverse voltage rating  $1.25 \times PIV_x$ . This insures that the peak reverse

voltage never exceeds 80% of the rating of a particular diode. Hence, in this case, the diode on the 5 V output should be rated for more than 30 V, the 12 V output more than 70 V, and the 30 V output more than 171 V. Peak reverse voltages should be measured on all diodes under maximum load and startup conditions to ensure that ratings are not exceeded.

The rule of thumb for the diode current rating is to choose a device with a DC current rating at least three times the average DC output current of the particular output. From the current specifications of Table 1 and the voltage requirements above, the following minimum ratings should be defined in this case:

- 5 V output diode — 6.0 A, 30 V
- 12 V output diode — 3.6 A, 70 V
- 30 V output diode — 60 mA, 171 V



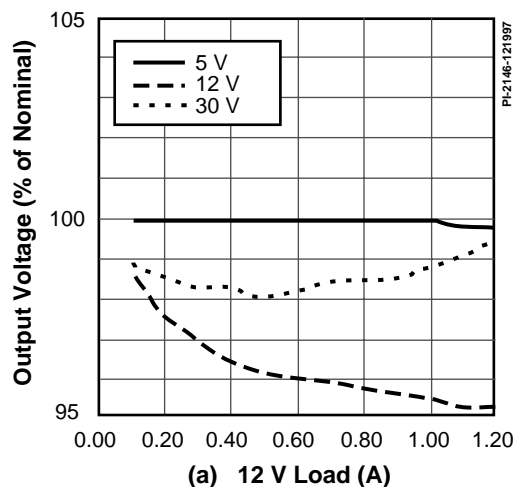


Figure 8(a). Cross Regulation with Feedback from 5 V Only.  
Response to Variation of 12 V Load

For reverse voltage ratings less than 100 V, Schottky diodes can be used to minimize power losses. As discussed earlier, Schottkys can also be used to improve the relative accuracy of output voltages when calculating the number of turns. Schottky diodes are more expensive than PN junction diodes. The circuit of Figure 1 uses ultra fast recovery PN diodes for the lowest cost, while the circuit in Figure 4 uses a Schottky diode on the 5 V output with the same transformer design. Circuit performance may be improved with a transformer designed specifically for a Schottky diode on the 5 V output.

In this example many possible diodes are available to achieve the required characteristics. The devices in the example of Figure 4 are:

5 V output:	MBR745 7.8 A, 45 V Motorola
12 V output:	MUR420 4.0 A, 200 V Motorola
30 V output:	UF4004 1.0 A, 400 V General Semiconductor

Other suitable diodes are available from different manufacturers. Tests with a number of diodes are recommended to verify the optimum devices in each application.

## Circuit Performance

The volts per turn defined in Equation (1) is an approximation based on the forward voltage of the output diode. This value

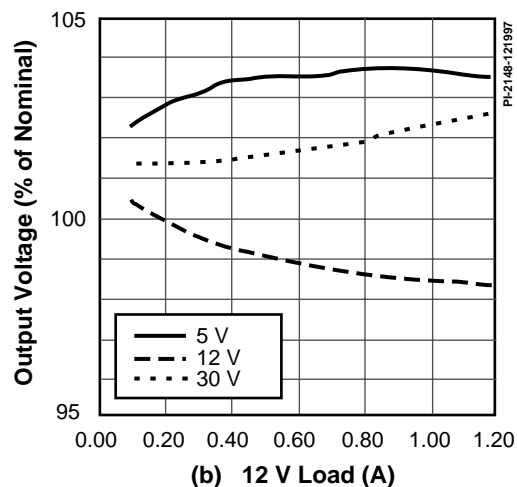


Figure 8(b). Cross Regulation with Feedback from 5 V and 12 V.  
Response to Variation of 12 V Load

changes with load current and temperature. As the outputs have varying loads, the output diodes will exhibit different forward voltages depending on the load conditions on the particular output. Changing load conditions on the 5 V output, for example, will inherently influence the voltages on the other outputs.

In addition, secondary effects such as voltage spikes from leakage inductance and quality of coupling between output windings, lead to reduced voltage accuracy on outputs which do not provide feedback through the optocoupler.

The basic circuit of Figure 4 derives feedback only from the 5 V output. As a consequence, the other output voltages vary as the 5 V output current changes. The influence on the 12 V output is shown in Figure 6(a). Use of a Schottky diode in a circuit designed for a PN diode emphasizes the effect of a change in voltage drop, as illustrated in this example.

The 5 V output voltage is well controlled since it exclusively provides the feedback signal. The 12 V output, however, is seen to vary by  $\pm 2\%$  as the 5 V load is varied between 25% and 100% (0.5 amps to 2.0 amps). For this test the 12 V output load was held constant at 0.6 amps. The 12 V and 30 V outputs are also below their nominal values because of the lower drop of the Schottky diode.

Transformer construction techniques to optimize output cross regulation were discussed earlier. However, it is often necessary to further enhance cross regulation using external circuit techniques. For example, if improved regulation is required on the 12 V output, a simple technique is to derive the feedback from both 5 V and 12 V outputs. In this example, as in most applications, higher accuracy is required on one of the outputs. Here it is assumed that the main output is still the 5 V, but some feedback may be drawn from the 12 V output to improve its

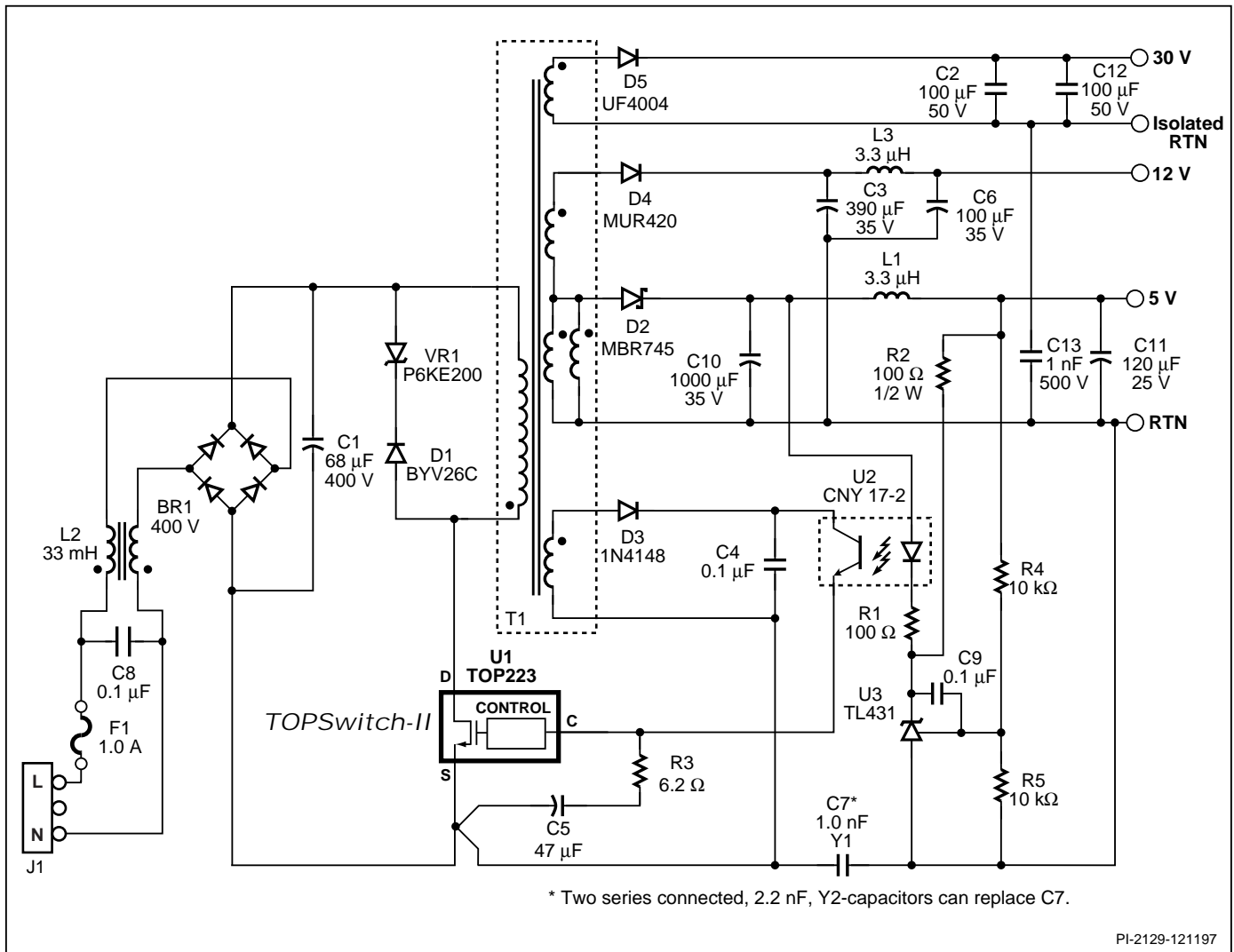


Figure 9. Modified Schematic of Figure 4 with Isolated 30 V Output and C13 for Common Mode Current Return.

load regulation. The schematic of Figure 7 illustrates a simple modification to the original circuit of Figure 4, where resistor R6 is introduced from the 12 V output to the reference pin of the TL431 shunt regulator.

Figure 6(b) illustrates the improvement obtained by employing this new feedback scheme where the load regulation on the 12 V output is improved to  $\pm 1.5\%$ . The effect would be more dramatic if the transformer had greater leakage inductance on the output windings.

The value of R6 is generally determined through iteration and depends on the degree of feedback desired from the second output. Introducing feedback from a second winding has a detrimental effect on the regulation of the main output. In this example the change in the 5 V output increases from effectively 0% in Figure 6(a) to  $\pm 0.75\%$  in Fig 6(b).

A good rule of thumb as a start point for tests is to choose R6

such that it yields about 10% of the current in R4 (with the TL431 reference pin at 2.5 V).

In this example the current in R4 before modification is:

$$I_{R4} = \frac{(5 - 2.5) V}{10 \text{ k}\Omega} = 250 \mu\text{A}$$

To emphasize the effect, we let the 12 V output provide 50% of this amount through R6. Assuming that the TL431 reference pin is still at 2.5 V

$$R6 = \frac{(12 - 2.5) V}{125 \mu\text{A}} = 76 \text{ k}\Omega$$

A standard resistor value of 75.0 kohm was chosen for R6 in Figure 7.



Note that since the sum of the currents through R4 and R6 is a constant equal to  $2.5\text{V}$  divided by R5, the additional feedback that R6 introduces from the  $12\text{V}$  output will tend to reduce the regulated value of the  $5\text{V}$  output. This requires that R4 is in turn adjusted to retain the  $5\text{V}$  output at the desired level. To retain the voltage at the reference pin of the TL431 at  $2.5\text{V}$ , the value of R4 must therefore be increased to reduce its current by  $50\%$ .

$$R4 = \frac{(5 - 2.5)\text{V}}{(250 - 125)\ \mu\text{A}} = 20\ \text{k}\Omega$$

A slightly larger precision  $21.0\ \text{k}\Omega$  resistor was specified in the circuit of Figure 7 to compensate for the small penalty in regulation on the  $5\text{V}$  output.

Figure 8 shows load regulation measurements before and after these circuit modifications with the  $5\text{V}$  output load held constant at  $1\ \text{A}$  and the  $12\text{V}$  output load varied from  $10\%$  to

$100\%$ . In Figure 8(a), the  $5\text{V}$  output is very stable since this is exclusively providing output feedback, while the  $12\text{V}$  output drops by  $4\%$  over the load range. In Figure 8(b), the introduction of R6 maintains tighter regulation on the  $12\text{V}$  output ( $\pm 1.5\%$  variation with load), whereas this additional feedback introduces a  $\pm 0.75\%$  variation in the  $5\text{V}$  output voltage over the same load range.

The degree of feedback required from each output can thus be determined depending on the application requirements for output voltage tolerance. Breadboard evaluation is necessary to adjust component values for the desired performance.

## EMI Considerations

In general, the EMI considerations in a multiple output TOPSwitch power supply do not differ from those of a single output supply, and are covered in detail in AN-15. There are, however, specific multiple output power supplies where

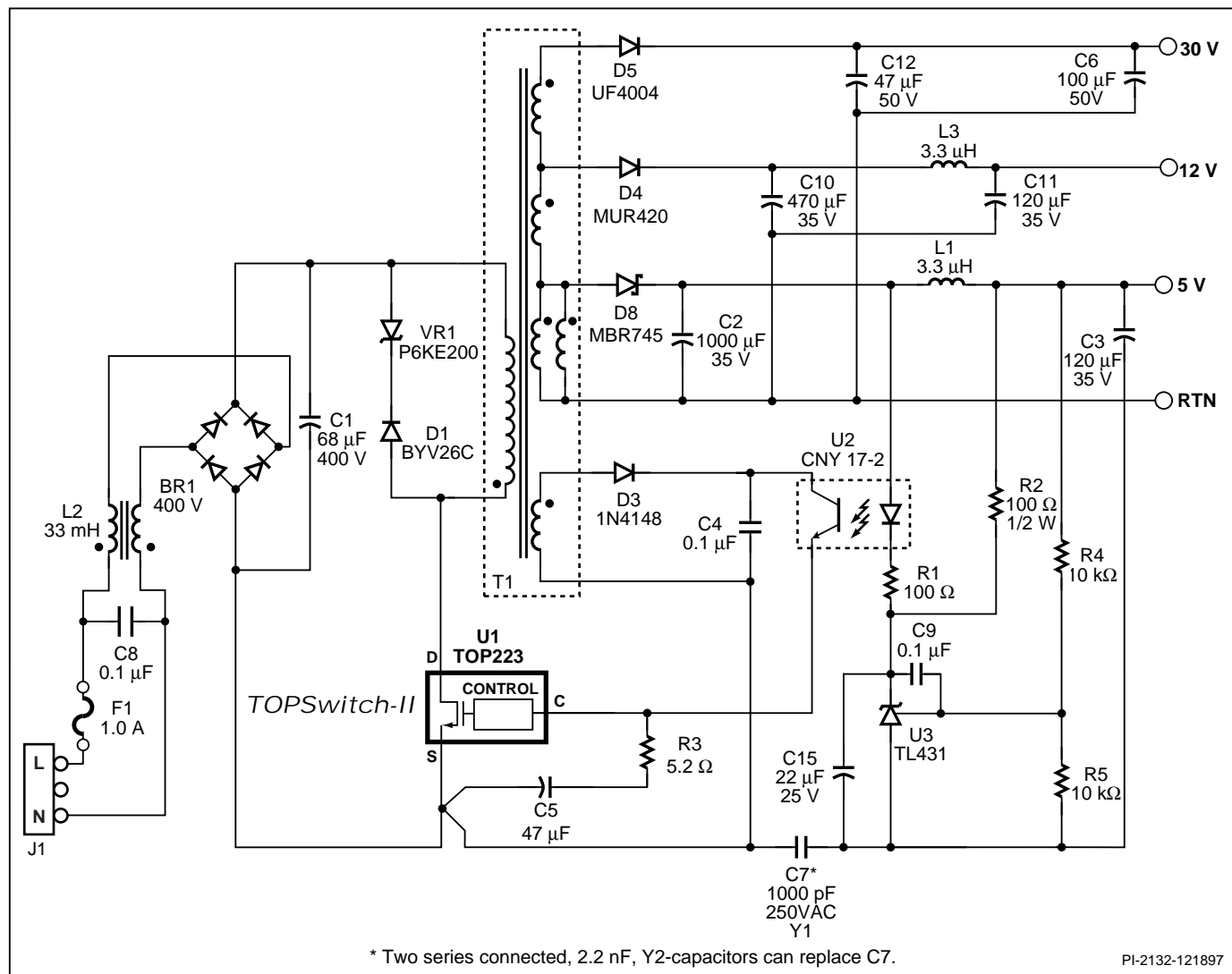


Figure 10. Modified Schematic of Figure 4 with Soft-Start Capacitor C15 Added.

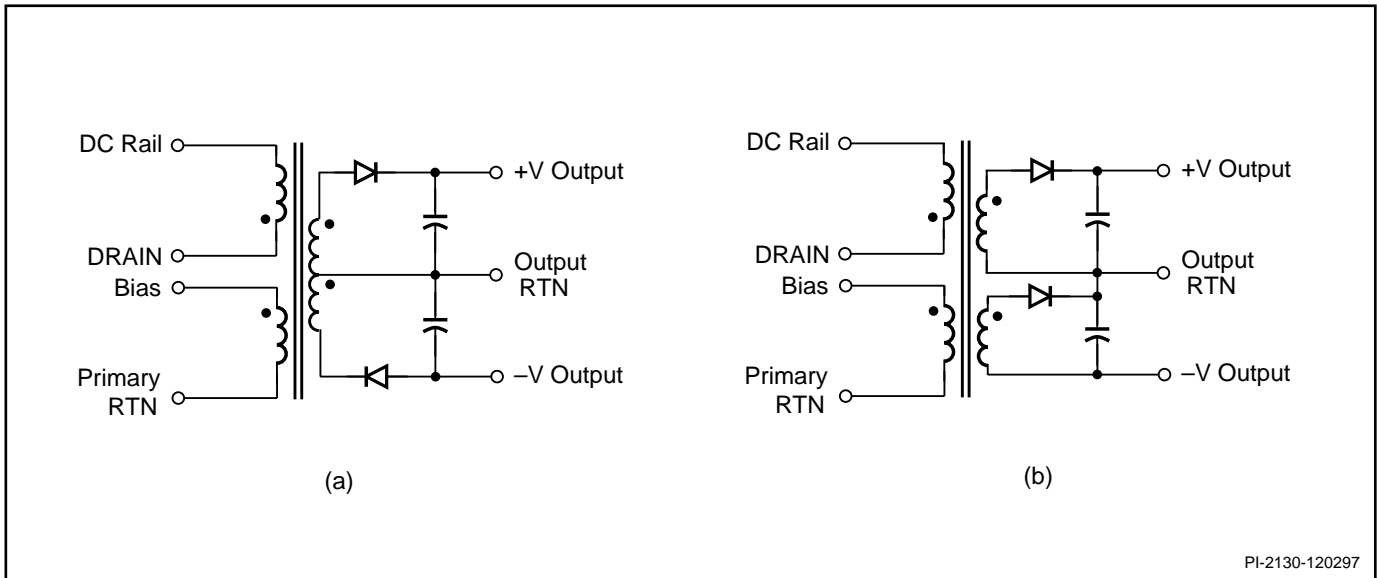


Figure 11 (a), (b). Two Configurations to Get Negative Outputs.

additional measures are necessary to optimize the EMI performance. This is particularly true when the outputs are galvanically isolated from each other. In motor control circuits, for example, several isolated outputs may be required to supply high side drivers in an inverter output stage.

In these cases it is important that displacement currents driven by the *TOPSwitch* DRAIN node through the transformer's interwinding capacitance have a low impedance return path from a specific output to the primary side of the power supply. This consideration demands that each isolated output provide a low impedance path for common mode displacement currents to return from its own return to the primary return (*TOPSwitch* SOURCE potential). This low impedance path can usually be provided from the output's return through a capacitor (suitably rated for the isolation voltage required on a particular output) to the main secondary return, from where a safety Y capacitor is connected to the primary return rail. This configuration is shown in Figure 9, where the isolated 30 V output has a 500 V capacitor, C13, connected between its return rail and that of the main power supply output.

If these low impedance capacitive paths are not provided on each isolated output, then the common mode displacement currents transferred through the transformer interwinding capacitance will return to their source on the primary of the transformer through any alternative route that is available. The common mode currents may split many times on their route to the DRAIN node. If a capacitive return path is not present, there is the risk that enough of the displacement current will flow through the AC input conductors to fail regulatory emission specifications.

The need for additional capacitors in this type of circuit depends on the transformer's interwinding capacitance. Additional capacitors from an isolated output may not be necessary if its capacitance to the primary is low enough. However, tests are essential to verify the necessity of additional components.

One other EMI consideration related to output diode snubbers is worthy of note. Output diodes are always a source of additional noise that depends on their forward and reverse recovery characteristics, particularly the  $di/dt$  and  $dv/dt$  during recovery. Many diodes are now available with so called 'soft recovery' characteristics which are designed to limit switching noise. It is often desirable, however, to further snub the diode characteristics with external components.

These external snubbers are usually a single capacitor, or series resistor and capacitor in parallel with the output diodes.

In many cases the snubbing circuitry can be limited to a single output diode to achieve the desired reduction in switching noise. In such cases, the highest voltage winding with significant loading should be chosen for the snubber circuitry. In this example, the 12 V output diode would be chosen since the capacitors on that output have lower ESR than the capacitors on the 30 V output. It also has the best overall coupling with the primary winding because it is physically closest. During the primary switching events, these snubber components are an AC current path in series with the output electrolytic capacitors. They therefore provide a low impedance AC path across the transformer output winding and the output diode to confine the noise currents created by primary switching events.



## Additional Tips

Following are some tips which can be considered and tested where necessary to improve circuit performance.

### Optocoupler Connection

In multiple output power supplies, the current for the optocoupler LED is often supplied via the loop gain setting resistor from an output other than the main feedback voltage. In Figure 4, this connection of R1 is made to the 5 V winding. This technique introduces some AC feedback from the 5 V winding, which helps reduce variation on that output during transient load conditions.

R1 and R2 may be connected to the 12 V output instead of the 5 V output (with their values changed appropriately). Ripple current from this output has a path to the TL431 reference pin via R1 and C9. This type of connection, however, will often introduce loop instability with very light loads on the 12 V output. The reason is that the 12 V output is subject to peak charging from energy in leakage inductance as its load approaches zero. Peak charging effectively uncouples the output so that it is no longer related to the 5 V output by the turns ratio. If instability is observed during light or no load conditions on the 12 V output, two options are available:

1. The optocoupler LED should be supplied from the 5 V winding (with the value of R1 selected to maintain acceptable AC gain) or
2. A dummy or minimum load resistor can be added to the 12 V output to eliminate the effects of peak charging. Dummy loads are usually added to improve regulation at light loads. R2 is used for this purpose on the 5 V output in Figure 4. R2 might be moved to the 12 V output if one dummy load is sufficient to meet specifications. The value of this resistor should be adjusted as necessary to allow for the load range of a particular application.

### Soft Start Circuitry

Soft start circuitry is often useful to avoid output voltage overshoot during power supply turn on. This is achieved simply by introducing a capacitor from the TL431 cathode to anode as shown by C15 in Figure 10.

Note a discharge path is required for this capacitor to insure the soft start function is reset when the output voltages decay at turn off. This function is provided in Figure 10 by the minimum load resistor R2.

When introducing soft start, it is useful to supply the optocoupler LED from a higher voltage output, such as the 12 V rail in this case, since this will insure that C15 begins to charge and provide the soft start function as soon as possible after the power supply starts to operate. The issues of minimum load on the higher voltage output, discussed above, must be considered when doing this to insure loop stability under all conditions.

### Improving Regulation in Lightly Loaded Outputs

Some outputs, such as the 30 V output in this example, can have very light loads even under maximum load conditions. These are prone to peak charging, which can produce output voltages much higher than expected by the turns ratio of the transformer output. The degree of this peak charging is strongly influenced by the loads on the other outputs.

The output in question can simply be clamped with a Zener diode between the output and secondary return. However, a lower cost and more efficient solution is to provide some low pass filtering that will reject the short voltage spikes from leakage inductance to prevent charging of the output capacitors. The introduction of a resistor in series with D5 in Figure 4 will provide this function. Values from 10 to 100 ohms should be tested to determine the optimum. See R6 in Figure 1.

### Negative Outputs

Negative outputs are often required in a system for operational amplifiers or other analog circuitry. Two simple configurations generally used to provide these outputs are shown in Figure 11.

Figure 11(a) shows the most usual configuration, where the direction of the output diode is reversed such that that diode's cathode is connected to the transformer's output pin. The other end of the negative winding is connected to the common secondary return using the same dot convention as the other output windings. An alternative technique connects the anode of the output diode to the return end of the winding with the cathode connected to the common secondary return as shown in Figure 11(b). The alternate, however, is not available with stacked windings.

The calculation of the number of output turns is identical to that for positive outputs, and the same transformer construction techniques are used to optimize cross regulation. Since negative outputs are often lightly loaded, the techniques to improve regulation in lightly loaded outputs detailed above are often useful. Alternatively, the output can simply be post regulated with a linear regulator.

## Appendix A

### Key Spreadsheet Variables.

The following key variables in the transformer design spreadsheet of Figure 2 should be checked before a transformer design can be deemed acceptable:

$D_{MAX}$  — Must be less than the *TOPSwitch* data sheet minimum value of 64% (0.64).

$I_p$  — To allow for thermal effects, this should be no greater than 90% of the data sheet minimum current limit specification for the chosen *TOPSwitch* at 25°C. In this example, the minimum current limit for the TOP223Y is specified as 0.9 A, so the spreadsheet value of 0.78 meets the above criterion.

$B_p$  — This must be below the recommended value of 4200 gauss to avoid excessive core saturation at the peak *TOPSwitch* current limit. Here the value of 3767 gauss is well within this requirement.

$L_G$  — Although the guidance of transformer vendors should be sought, airgaps of <0.051mm are not recommended because such small gaps make it difficult to hold a reasonable tolerance on the specified primary inductance.

CMA — Values between 200 and 500 allow reasonable temperature rise in the windings. Smaller values indicate higher temperatures from greater losses in the copper. The value of 219 circular mils per amp in Figure 2 meets the recommended lower limit of 200.

Again, AN-17 should be consulted for full details on the use of the spreadsheet of Figure 2.



## Appendix B

### 3.3 V and 5 V Outputs

An increasing number of applications require that both 3.3 V and 5 V outputs in multiple output power supplies, both requiring  $\pm 5\%$  regulation to supply digital control circuitry. Several commonly used techniques to achieve this performance are described below.

#### Linear Regulator

The simplest, though least efficient technique, is to design only a 5 V output winding with wire capable of supplying the RMS current for both the 5 V and 3.3 V outputs. A linear regulator is then placed on this 5 V output, regulating down to 3.3 V as shown in Figure 1. Integrated 3.3 V regulators are now available from a number of suppliers with varying current capabilities. A simple emitter follower regulator could also be employed using discrete components.

The disadvantage of this technique is reduced power supply efficiency, although it simplifies the transformer construction and reduces the number of output pins.

#### Transformer Turns Ratio

Two techniques are commonly used to design separate transformer windings for each output. Each has the required turns ratio relationship to provide the regulation required.

##### 1. Copper wire

If 3 turns are defined for the 3.3 V output and an ultra fast PN junction diode is specified for this output, the calculation of

the volts per turn provides a solution where 4 turns are used with a Schottky diode for the 5 V output.

$$V_{PT} = \frac{(V_o + V_D)}{N_s} \quad (1)$$

If

$$V_o = 3.3 \text{ V}$$

$$V_D = 0.7 \text{ V}$$

$$N_s = 3 \text{ turns}$$

Then from (1) find

$$V_{PT} = 1.33 \text{ V per turn}$$

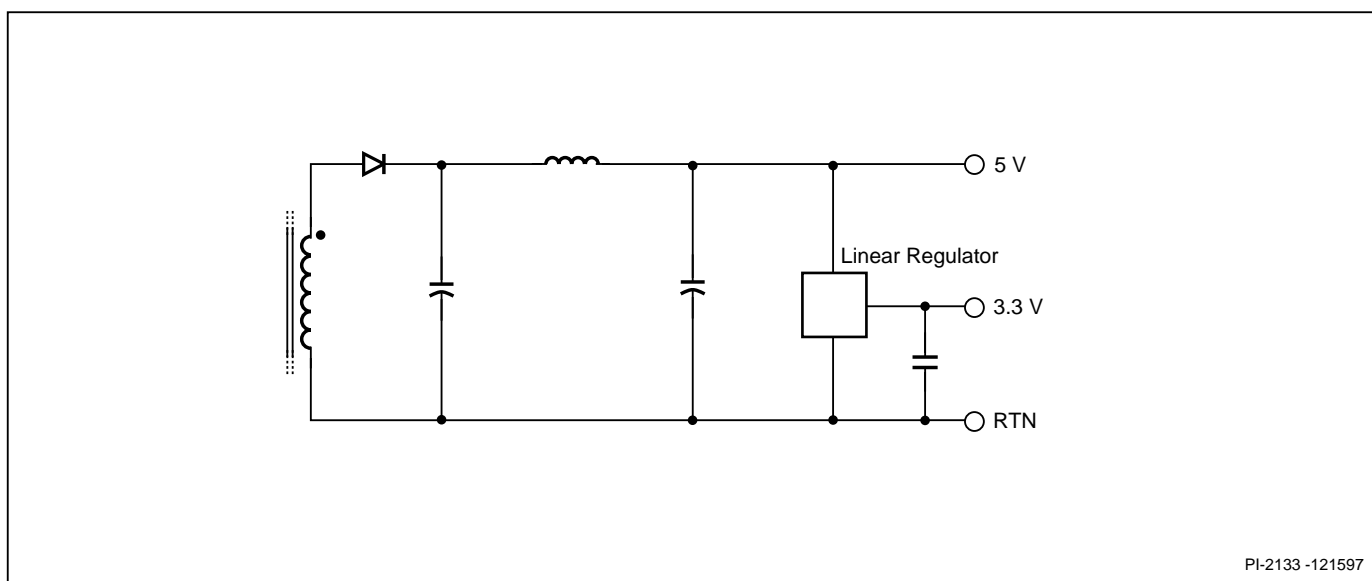
rearranging (1) to calculate the turns required for the 5 V output yields:

$$N_s = \frac{V_o + V_D}{V_{PT}} \quad (2)$$

If

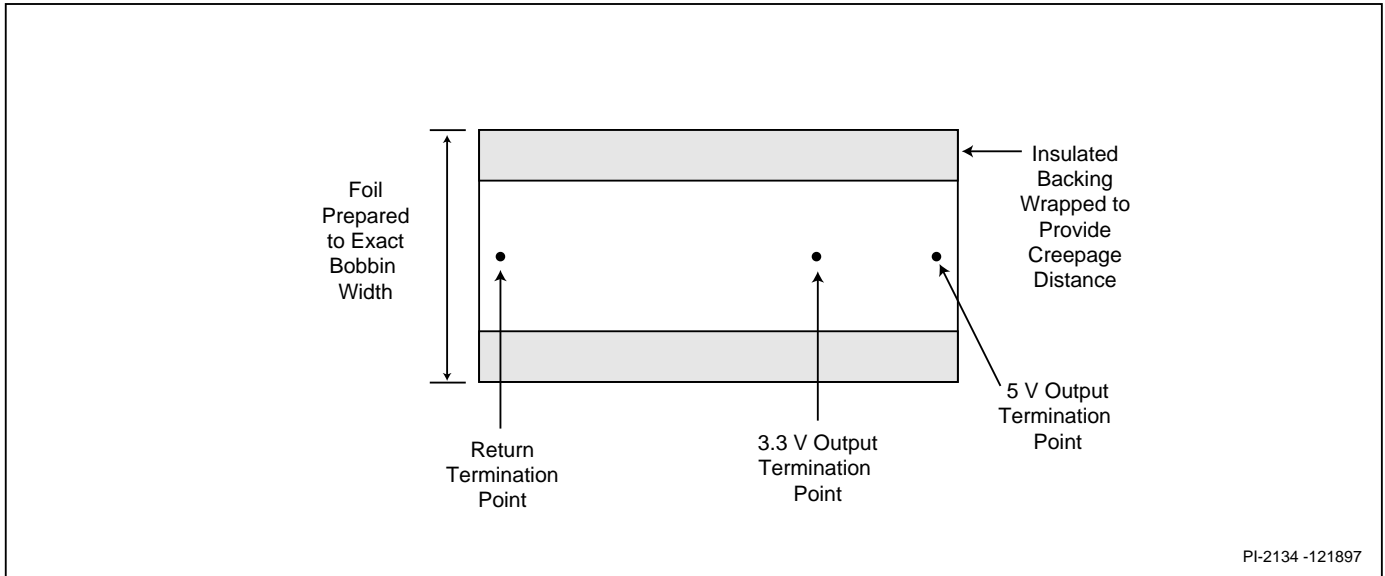
$$V_s = 5 \text{ V}$$

$$V_D = 0.4 \text{ V}$$



PI-2133 -121597

Figure 1. Derivation of 3.3 V Output from 5 V with Linear Regulator.



PI-2134 -121897

Figure 2. Preparation of Foil Windings for 5 V and 3.3 V Outputs.

$$V_{PT} = 1.33 \text{ V per turn}$$

From (2), the turns required on the 5 V output are:

$$N_s = 4.06 \text{ turns}$$

This result demonstrates that this choice of turns and output diodes yields an almost perfect integer turns ratio between the 3.3 V and 5 V outputs. It is a very popular solution for this reason.

The coupling between the output windings is still a crucial factor to insure that the turns ratio calculated above does indeed result in the required output cross regulation. Since so few turns are involved in these outputs, it is usual for multiple parallel wire strands to be used on each output winding, and for the 3.3 V and 5 V outputs to be constructed as separate windings. Stacked windings are not appropriate in this case. As discussed in the body of the application note, windings should be constructed in 2 layers and interleaved across the bobbin width to optimize coupling with the primary winding.

## 2. Foil windings

An alternative technique is to use foil instead of multiple strands of copper wire. Using this technique, the turns ratio of 3 turns on the 3.3 V output and 4 turns on the 5 V output is retained. The foil is cut to the required length with appropriate termination points included prior to winding. The foil is then

wound as a single operation. Termination to the transformer pins is performed afterwards. Figure 2 illustrates this technique.

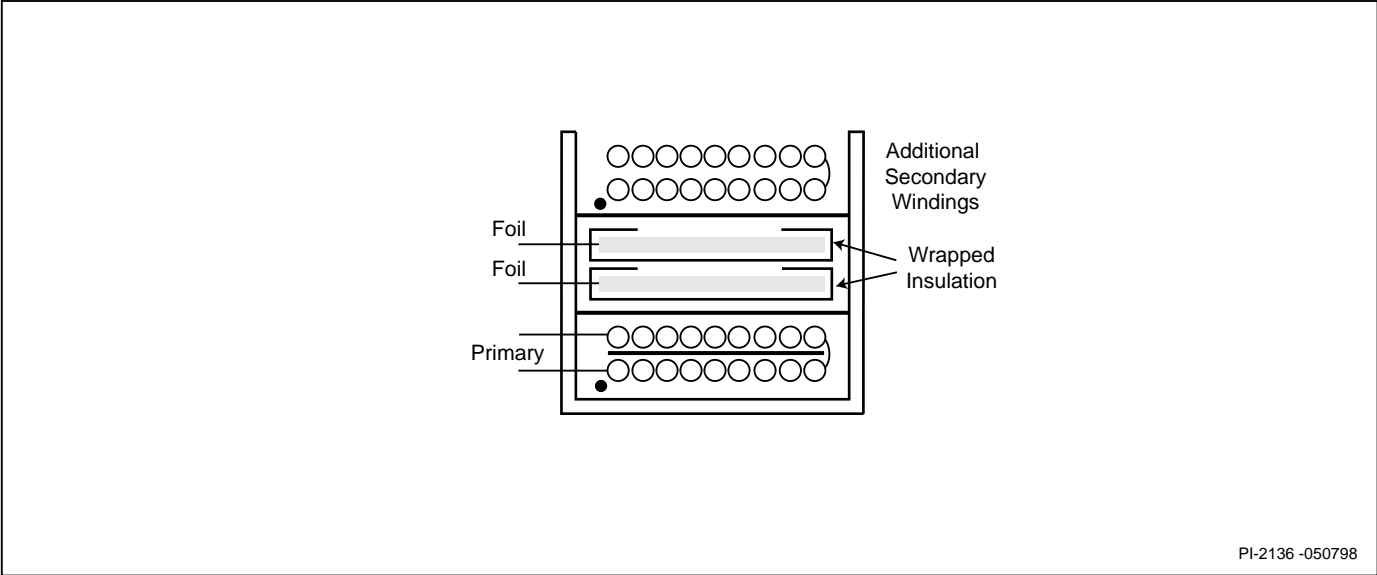
The foil is prepared to fit the bobbin width of the chosen transformer exactly, and is backed with insulation material which is wrapped around the foil to provide creepage distances appropriate to the required isolation requirements of the application.

Although this technique may add some cost to the transformer construction, the fact that the foil is prepared to the exact bobbin width provides excellent coupling with the primary winding. In addition, the 3.3 V and 5 V windings have very good mutual coupling that improves cross regulation. This mutual coupling makes the stacked winding construction the preferred technique when using foil windings.

As shown in Figure 3, subsequent output windings can be stacked on the foil windings, though the total RMS current requirements must be accounted for in the choice of the foil.

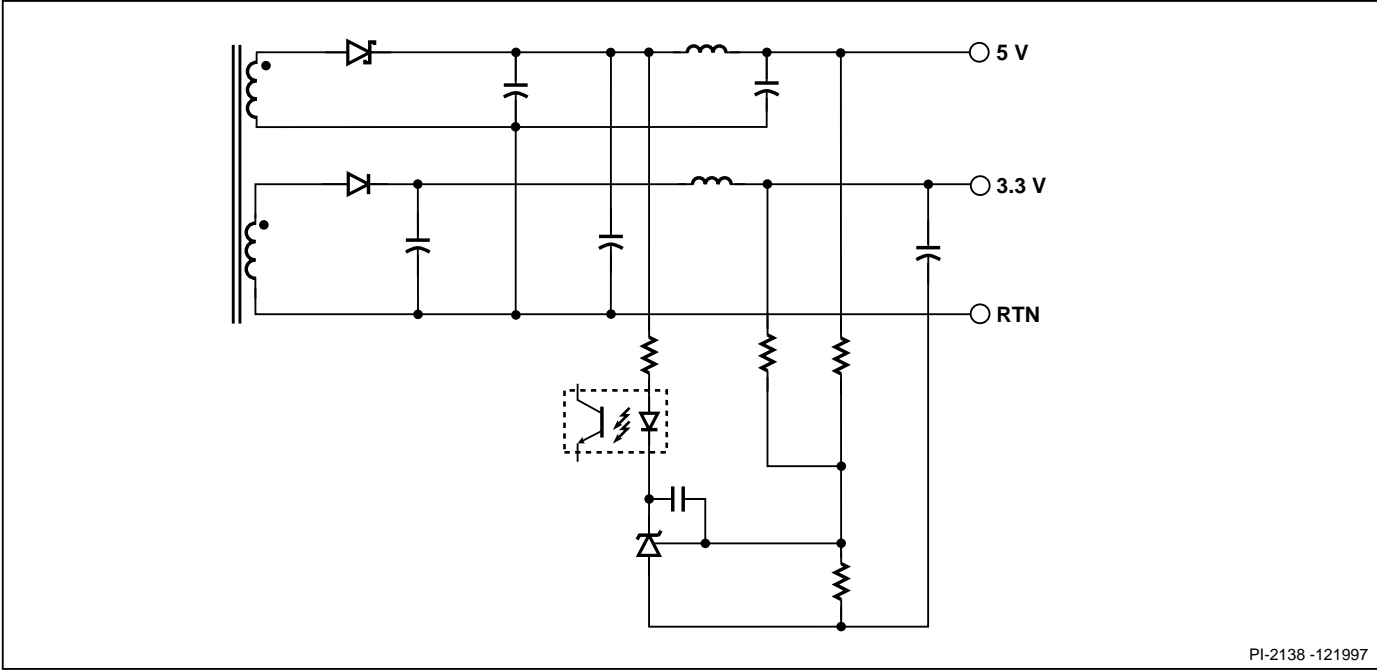
Independent of whether copper wire or foil winding techniques are used, the output feedback configuration must be determined according to the load and regulation requirements.

Figure 4 shows the use of a TL431 where the feedback is derived from both the 3.3 V and 5 V outputs. The proportion of feedback from each output can be adjusted as required, and is discussed in detail in the body of the Application Note.



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Figure 3. Winding Arrangement with Foil and Wire for Multiple Outputs.

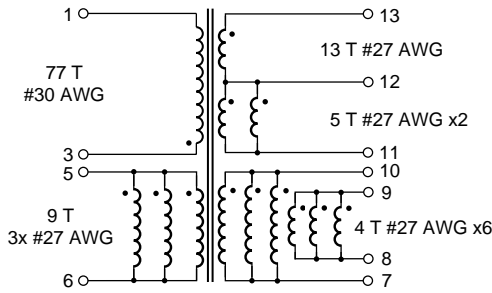


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Figure 4. Use of Feedback from Both Outputs with TL431 to Improve Regulation on 3.3 V Output.

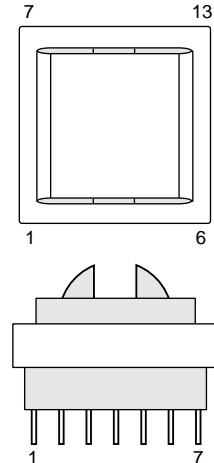
## Appendix C Transformer Construction Details

### MARGIN WOUND TRANSFORMER



CORE# - ETD29 (Philips)  
GAP FOR  $A_L$  OF 225 nH/T<sup>2</sup>  
BOBBIN# 4322 021 3438 (Philips)

PIN	FUNCTION
1	HIGH-VOLTAGE DC BUS
3	TOPSwitch DRAIN
5	VBIAS
6	PRIMARY-SIDE COMMON
7, 8	RETURN
9, 10	+5V OUTPUT
11	+5V OUTPUT CONNECTION
12	+12V OUTPUT
13	+30V OUTPUT



### ELECTRICAL SPECIFICATIONS

Electrical Strength	60 Hz, 1 minute, from pins 1-6 to pins 7-13	3000 VAC
Creepage	Between pins 1-6 and pins 7-13	5.0 mm (min)
Primary Inductance	All windings open	1340 $\mu$ H, $\pm$ 10%
Resonant Frequency	All windings open	1 MHz (min)
Primary Leakage Inductance	Pins 7 through 13 shorted	34 $\mu$ H (max)

NOTE: All inductance measurements should be made at 100 kHz

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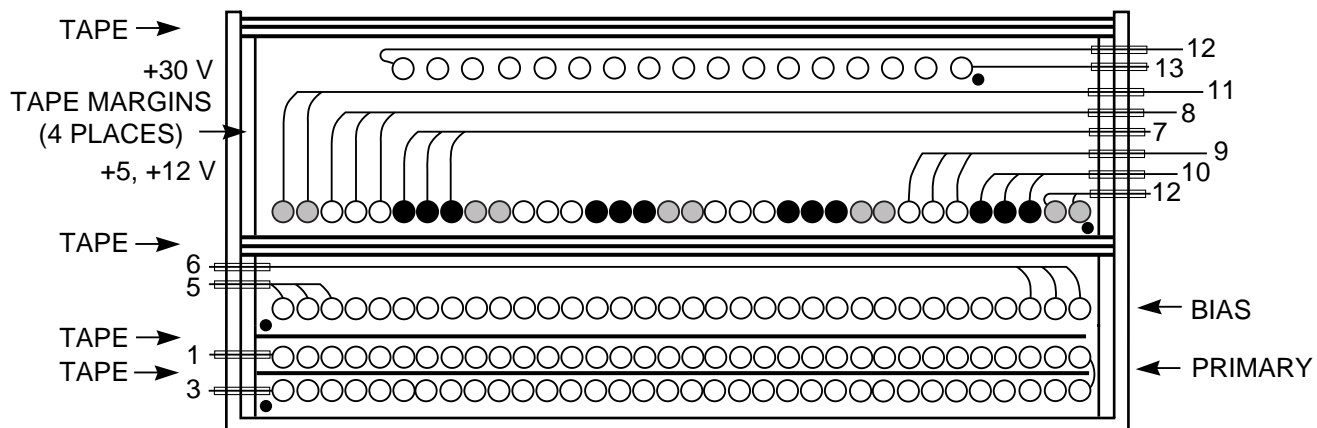
### PARTS LIST FOR TRANSFORMER DESIGN EXAMPLE

Item	Amt.	Description	Part #	Manufacturer
1	1pr.	Core, ETD29	4312 020 3750*	Philips
2	1ea.	Bobbin, ETD29-1S-13P, 13 pin	4322 021 3438	Philips
3	A/R	Wire, 30 AWG Heavy Nyleze		
4	A/R	Wire, 27 AWG Heavy Nyleze		
5	A/R	Tape, Epoxy 2.5 mm wide	#10	3M
6	A/R	Tape, Polyester 14 mm wide	#1298	3M
7	A/R	Tape, Polyester 19 mm wide	#1298	3M
8	A/R	Varnish		

\*Gap for  $A_L$  of 225 nH/T<sup>2</sup>  $\pm$  5%

PI-2154-050798

## MARGIN WOUND TRANSFORMER CONSTRUCTION



### WINDING INSTRUCTIONS

Primary and Bias Margins	Tape Margins with item [5]. Match height with Primary and Bias windings.
Double Primary Layer	Start at Pin 3. Wind 39 turns of item [3] from left to right. Wind in a single layer. Apply 1 layer of tape, item [6], for basic insulation. Wind remaining 38 turns in the next layer from right to left. Finish on Pin 1.
Basic Insulation	1 Layer of tape [6] for insulation.
Bias Winding	Start at Pin 5. Wind 9 Parallel Trifilar turns of item [4] from left to right. Wind uniformly, in a single layer, across entire width of bobbin. Finish on Pin 6.
Reinforced Insulation	3 Layers of tape [7] for insulation.
Output Margins	Tape Margins with item [5]. Match height with all output windings
+5 V and +12 V Winding	Start with two sets each containing three wires item [4], and one pair of wires item [4]. Terminate first set of three wires to pin 9 and the second set of three wires to pin 10. Terminate the pair of wires to pin 12. Wind the combination of eight wires in parallel right to left evenly across the bobbin, with the pair of wires closest to the right side of the bobbin. After four turns of the combination of eight wires, terminate the first set of wires to pin 8 and the second set of wires to pin 7. Continue to wind the pair of wires one more turn for five turns total. Finish at pin 11.
Basic Insulation	1 Layer of tape [6] for basic insulation.
+30 V Winding	Start at Pin 13. Wind 13 turns of item [4] from right to left. Wind uniformly, in a single layer, across entire width of bobbin. Finish on Pin 12.
Outer Assembly	3 Layers of tape [7] for insulation.
Final Assembly	Assemble and secure core halves. Impregnate uniformly with varnish.

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