Constant Current/Constant Power Regulation Circuits for *TOPSwitch*®

Design Note DN-14

Current control circuits are used to accurately control output current of a power supply. Battery chargers are the largest single application for current controlled power supplies requiring either constant-current or constant-power output characteristics. Motor drive applications can also require power supplies with controlled output current.

This design note concentrates on three basic secondary current control circuits for *TOPSwitch* flyback converters. A simple, low cost circuit using a transistor current sense and regulation circuit is presented first. The second circuit features highly accurate current control using an op amp and secondary bias winding, which can also be modified to approximate a constant power output characteristic. The third circuit also features highly accurate current control using an op amp with secondary bias derived from the existing output winding. All three circuits share similar power stage components. The AC line voltage is rectified and filtered by BR1 and C1. The resulting DC voltage is applied to the drain of *TOPSwitch* through the primary winding of T1. *TOPSwitch* combines a high voltage power MOSFET and control circuit in a single 3-terminal package. VR1 and D1 clamp *TOPSwitch* drain voltage to a safe level. C5 provides bypassing and frequency compensation while also determining the auto-restart period during fault conditions. D3 and C4 rectify and filter the *TOPSwitch* bias supply (note that D3 has a rated voltage of 200 V). C6, L2, and C7 control EMI. For details of *TOPSwitch* flyback converter design, refer to application notes AN-14, AN-16, AN-17, and AN-18.

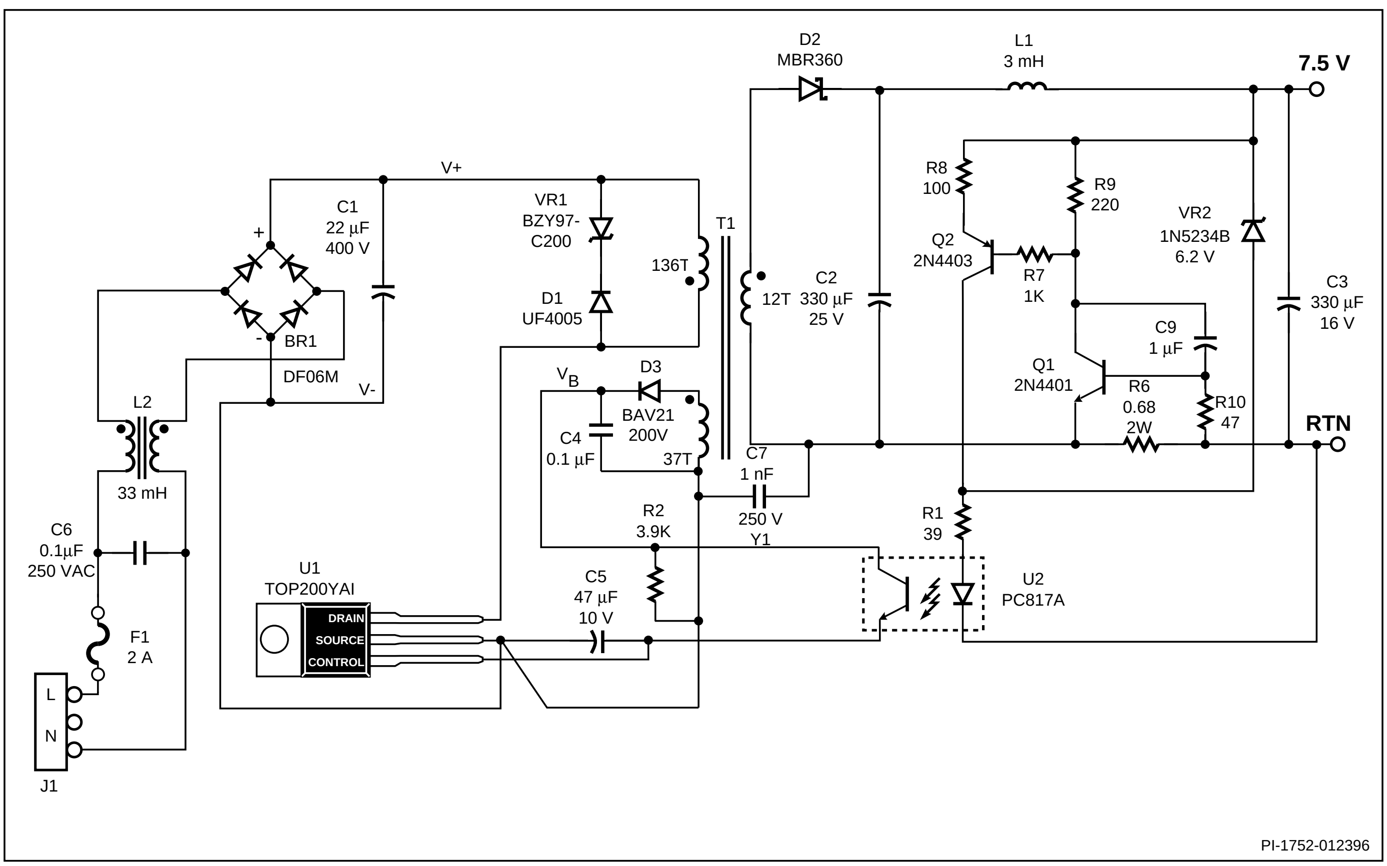

Figure 1. Simple, Low Cost Constant Voltage/Constant Current 7.5 V, 1A Supply Using the TOP200.

Transistor Current Limit Circuit

Figure 1 shows a 7.5 V constant voltage/constant current power supply circuit using the TOP200. The supply operates at a nominal output voltage of 7.5 volts or controlled output current of 1 A. This circuit can be used when low cost and simplicity are key factors, and an output current accuracy of +/-8% over a temperature range of 0-50°C can be tolerated. Due to the power dissipated in current sense resistor R6, this current control circuit should be used for output currents of 1A or less. For output currents greater than 1A, or if output current accuracy and temperature stability are a major concern, the op amp current control circuits shown in Figure 5 or Figure 11 should be used.

Circuit Description

At low output currents, the supply runs in constant voltage mode. Output voltage is determined by the voltage drop across VR2 and the light emitting diode (LED) inside optocoupler U2. Output current is sensed by R6. When the voltage drop across R6 becomes sufficient to turn on Q1, R9 drops enough voltage to turn on Q2, which sources current into optocoupler U2, reducing output voltage. VR2 then no longer conducts current, and the power supply enters constant current mode. R9 sets the collector current in Q1 required to turn on Q2, and establishes the operating point of Q1. R8 establishes the operating point of Q2 and determines the gain of the current control circuit. C9 compensates the current control loop. R2 provides a small preload on the primary bias winding to limit the output voltage when the supply is unloaded.

During turn on and auto-restart, transient peak currents can saturate Q1. R10 limits the current through the base-emitter junction of Q1, while R7 prevents current from being diverted away from R1 through the base-emitter junction of Q2 when Q1 is saturated.

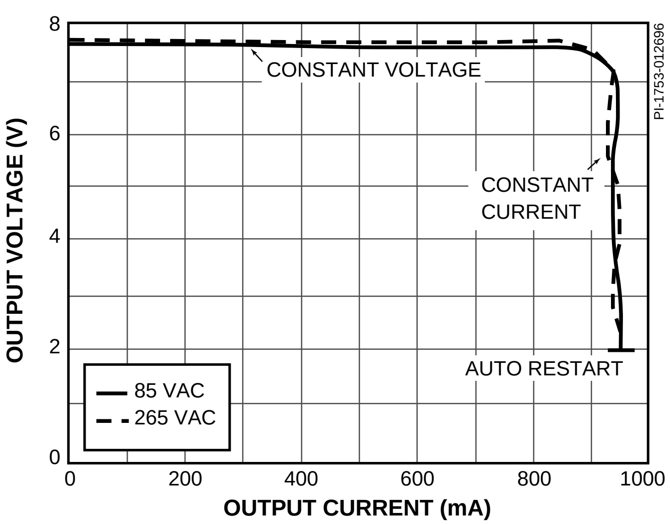

Figure 2. Output Voltage vs. Output Current Characteristic for Simple, Low Cost Transistor Circuit shown in Figure 1.

Performance

Figure 2 shows the measured performance of the simple, low cost transistor circuit. Voltage and current data taken at 85 VAC and 265 VAC are virtually identical. Q1 and Q2 control the output current until output voltage drops to approximately 2 V. At this point, Q1 and Q2 can no longer source sufficient optocoupler current, causing output current to become unregulated (but still limited by *TOPSwitch* primary current limit). With unregulated output current, voltage across R6 increases, further decreasing effective bias voltage for the current control circuit. This interaction between the current control circuit and current sense resistor R6 is regenerative, resulting in cutoff of current to optocoupler U2, forcing *TOPSwitch* into auto-restart. As a result, the simple, low cost transistor circuit transitions cleanly from constant current control to auto-restart at typically 2 to 2.4 volts output, depending on the input voltage. In auto-restart, average output current is much lower than the linearly controlled output current.

Analysis of Simple, Low Cost Circuit

Voltage and current control circuits and their bias supplies will be examined to show derivation of component values and circuit behavior for various operating conditions**.**

Voltage Control Circuit

V_{VR2}, V_{LED}, and the voltage drop V_{R1} (across R1) determine the output voltage:

$$V_{R1} = I_{R1} \times R1 \qquad (1)$$

$$V_O = V_{VR2} + V_{LED} + V_{R1} \qquad (2)$$

V_{VR2} and V_{LED} are the respective voltage drops across Zener diode VR2 and the U2 LED. I_{R1} is the current through R1 required to maintain output voltage regulation for a given value of input voltage and output load current. VR2 is a 1N5234B Zener diode rated at 6.2 V. The U2 LED typically drops 1.2 V at room temperature. The approximate contribution of V_{R1} can be determined by estimating the amount of control current required by *TOPSwitch* for normal duty cycle operation. I_{R1} will vary with input voltage, output load current, and the current transfer ratio (CTR) of optocoupler U2. From the *TOPSwitch* data sheet, nominal values of control pin current I_C required to operate *TOPSwitch* range from 2.5 mA for maximum duty cycle operation to 6.5 mA for minimum duty cycle operation. If *TOPSwitch* is biased to the middle of this range, control pin current I_C is 4.5 mA as shown in Equation (3).

$$I_C = \frac{2.5mA + 6.5mA}{2} = 4.5mA \qquad (3)$$

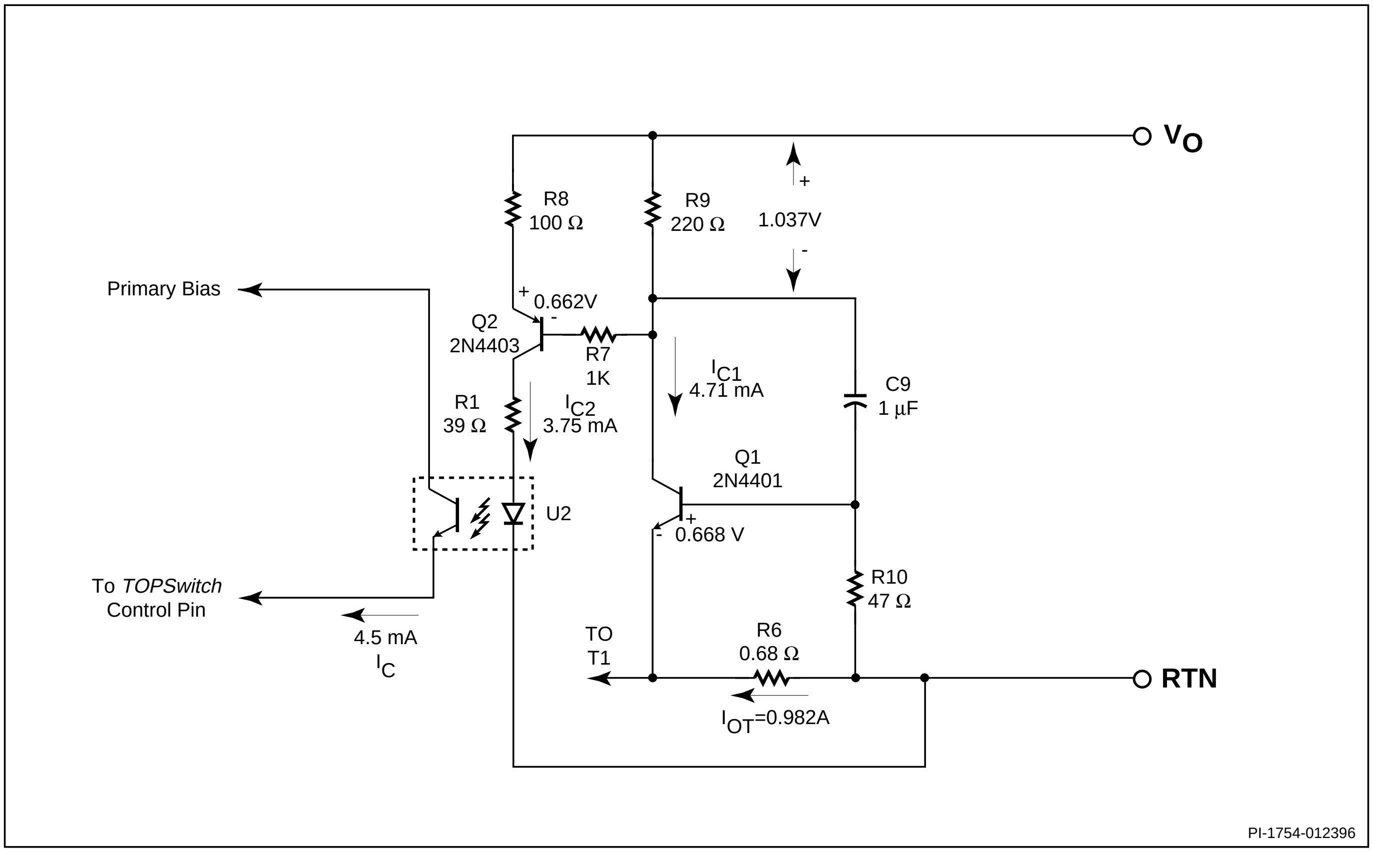

Figure 3. Voltage and Currents for Simple, Low Cost Current Control Circuit.

This current is supplied to the Control pin of the *TOPSwitch* from the emitter of optocoupler U2. For a given control pin current I_C, the I_{R1} value is calculated:

$$I_{R1} = \frac{I_C}{CTR} \tag{4}$$

I_{R1} is the required optocoupler input current for a given control pin current I_C and optocoupler CTR. The optocoupler shown has a CTR range of 80% to 160%. Assume that the actual value of CTR is in the middle of this range, or 120%. This means that a current of 3.75 mA must be supplied to the input of the optocoupler to bias *TOPSwitch* into the middle of its operating range. At this value of optocoupler LED current, voltage drop across R1 is 0.15 V, resulting in a nominal output voltage of 7.55 V according to Equation (2).

$$\begin{aligned} V_O &= V_{VR2} + V_{LED} + V_{R1} \\ &= 6.2V + 1.2V + 0.15V = 7.55V \end{aligned}$$

Current Control Circuit

Q1, Q2, C9, R1, and R6 - R10 perform the current control function. The value of R6 and the V_{BEQ1} voltage drop of Q1 set the target output current I_{OT}:

$$I_{OT} = \frac{V_{BEQ1}}{R6} \tag{5}$$

The V_{BE} voltage drop of Q1 and Q2 can be estimated from Equation (6):

$$V_{BE} = \frac{kT}{q} \times \log_e\left[\frac{I_C}{I_S}\right] \tag{6}$$

$$\text{where } \frac{kT}{q} = 0.0262V \;\; at \;\; 25°C$$

I_S is 4×10^{-14} amperes for a small signal silicon transistor and $\log_e$ indicates natural logarithm. I_C is actually collector current which is approximately equal to emitter current, especially in the case of the high gain transistors used here. For an approximate value for Q2 collector current I_{C2}, assume that *TOPSwitch* is biased by the current control circuit to the middle of the nominal operating range (I_C = 4.5 mA).

The current control circuit is shown in detail in Figure 3. For a nominal CTR value of 120% for U2, an LED current of 3.75 mA is required. This sets the collector current of transistor Q2 to approximately 3.75 mA. From this value of current, one can estimate the V_{BEQ2} voltage drop of Q2 using Equation (6):

$$V_{BEQ2} = \frac{kT}{q} \times \log_e\left[\frac{I_{C2}}{I_S}\right] = 0.0262V \times \log_e\left[\frac{3.75mA}{4 \times 10^{-14}A}\right]$$
$$= 0.662V \qquad (7)$$

A voltage drop V_{R8} of 375 millivolts is required across R8 to establish the required collector current of 3.75 mA in Q2. The voltage drop V_{R9} across R9 must be equal to the voltage drop across R8 plus the V_{BEQ2} drop of Q2 in order to establish the required value of 3.75 mA in Q2:

$$V_{R9} = V_{R8} + V_{BEQ2}$$
$$= 0.375V + 0.662V = 1.037V$$

The current required to develop 1.037 V across R9 is 4.71 mA which is also Q1 collector current I_{C1}:

$$I_{C1} = I_{R9} = \frac{V_{R9}}{R9} = \frac{1.037V}{220\Omega} = 4.71mA$$

From this value, one can estimate the V_{BEQ1} voltage drop of Q1 in order to establish a value for current sense resistor R6:

$$V_{BEQ1} = 0.0262V \times \log_e\left[\frac{4.71mA}{4 \times 10^{-14}A}\right] = 0.668V$$

With I_{OT} set at 1A and the V_{BEQ1} obtained using Equation (6), the value for R6 is found by rearranging Equation (5):

$$R6 = \frac{V_{BEQ1}}{I_{OT}} = \frac{0.668V}{1.0A} = 0.668\Omega \qquad (8)$$

The closest standard 5% value resistor value is 0.68Ω which results in a typical room temperature target current I_{OT} of:

$$I_{OT} = \frac{0.668V}{0.68\Omega} = 0.982A \qquad (9)$$

Figure 2 shows measured performance which is very close to the target value of output current.

Junction Temperature Effects

Variations in the junction temperature of Q1 will cause variations in the value of controlled output current. The base-emitter voltage drop of Q1 changes at a rate of approximately 2 mV/°C, decreasing with increasing temperature.

If the ambient temperature rises by 25°C, the value of controlled output current I_O will change by approximately 8%:

$$I_O = \frac{V_{BEQ1} - (\Delta T \times 2mV/°C)}{R6} \qquad (10)$$
$$= \frac{0.668V - (25°C \times 2mV/°C)}{0.68\Omega} = 0.909A$$

For applications requiring current control that is more stable with temperature, the op amp current control circuits shown in Figure 5 or Figure 11 should be used.

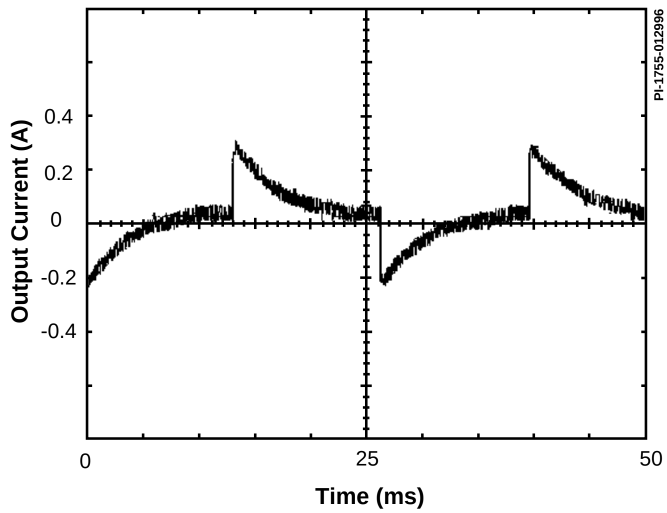

Figure 4. Output Current Transient Response for Simple, Low Cost Current Control Circuit.

Current Limit Frequency Compensation

C9 provides frequency compensation for the current control circuit. The value of C9 can be optimized in the circuit by applying a stepped *resistive* load to the output of the supply when operating in current control mode and examining the resulting output current using a DC current probe. The transient response should be well-damped, with no ringing or oscillation. Figure 4 shows the stepped resistive load response.

Primary Bias Supply

D3 and C4 provide a bias voltage for the primary circuit. The bias winding is connected in flyback polarity. This arrangement provides a predictable and well regulated source of voltage for the primary control circuit. The output voltage of the bias winding will be almost independent of input voltage and will track the secondary output voltage according to:

$$V_B = ((V_O + V_{D2} + (I_O \times R6)) \times \frac{N_B}{N_S}) - V_{D3} \qquad (11)$$

V_B is the bias voltage, V_{D2} is the forward voltage drop of D2, N_B is the number of transformer bias turns, N_S is the number of secondary turns, and V_{D3} is the forward voltage drop of D3. In a current controlled power supply, the output voltage is adjusted by the current control circuit to maintain a constant current at the output. For an increased output load (lower load resistance), the output voltage will be adjusted to a lower value to maintain constant current. The primary bias voltage will track the output voltage, and decrease as the output load is increased. The bias winding should be sized to provide enough voltage to properly bias the optocoupler and *TOPSwitch* Control pin at the lowest output voltage occurring just before the supply transitions from constant current to auto-restart. From Figure 2, this output voltage is typically 2 V. *TOPSwitch* maximum Control pin voltage is 6.1 V. The output voltage for the primary bias winding should be set to accommodate the *TOPSwitch* control pin voltage, plus approximately 3 V to keep the output transistor of optocoupler U2 properly biased. The absolute minimum bias voltage available should be at least 9 V. The required voltage and turns ratio for the bias winding can be determined by rearranging Equation (11):

$$N_B = \frac{V_B + V_{D3}}{V_O + V_{D2} + (I_{OT} \times R6)} \times N_S \qquad (12)$$

In the Figure 1 example N_S = 12 turns, V_B = 9 V, V_O = 2 V, V_{D2} = 0.6 V, V_{D3} = 1 V, I_{OT} = 0.982A, and R6 = 0.68 Ω.

$$N_B = \frac{9V + 1V}{2V + 0.6V + (0.982A \times 0.68\Omega)} \times 12 = 37 \text{ turns}$$

Under maximum output voltage and current conditions (V_O = 7.5V, I_O = 0.982A), this turns ratio will result in a V_B of:

$$V_B = \left((7.5V + 0.6V + (0.982A \times 0.68\Omega)) \times \frac{37}{12}\right) - 1V$$
$$= 26V \qquad (13)$$

The above voltage is the calculated nominal value for the primary bias voltage. In an actual application, the value of the bias voltage may be higher, due to peak charging of the primary bias supply by the primary voltage leakage spike. The influence of this leakage spike will be greatest at the highest output power of the supply. This occurs when output voltage V_O and output current I_O are both at maximum values which is also the transition region from constant voltage to constant current as shown in Figure 2. To determine the highest value of the primary bias supply, it should be measured in the application at this operating point.

Optocoupler Voltage Rating

It is necessary to consider the voltage stress across the output transistor of optocoupler U2 to make certain that the absolute maximum voltage rating of U2 is not exceeded. The maximum voltage V_{U2} across the output transistor of optocoupler U2 will be equal to the maximum bias voltage minus the *TOPSwitch* minimum Control pin voltage (5.5 V).

$$V_{U2} = V_{BIAS} - V_C = 26V - 5.5V = 20.5V \qquad (14)$$

A table of suitable optocouplers for use with *TOPSwitch* designs can be found in application note AN-14. Voltage ratings of these optocouplers range from a low value of 30 V to a maximum of 90 V. The voltage stress calculated above will allow any of the optocouplers in the table to be used with adequate voltage margin. When designing battery chargers with higher output voltages, it may be necessary to select an optocoupler with a higher output voltage rating from the table.

Op amp Current Control Circuit with Secondary Bias Winding

Figure 5 shows a 15 V, 2A constant voltage/constant current supply using the LM358 dual op amp for both voltage and current control. This circuit provides higher accuracy compared with the simple transistor circuit. Power loss is lower and efficiency is better because smaller resistance values can be used for sense resistor R6.

Circuit Description

Output voltage is rectified and filtered by D2, C2, C3, and L1. Output voltage is sensed by R4 and R5 and then compared by op amp U4B to reference U3. The output of op amp U4B drives current through D6 and R1 into the LED of optocoupler U2 to control *TOPSwitch* duty cycle. C9, R9, and R12 compensate the voltage control loop of the supply.

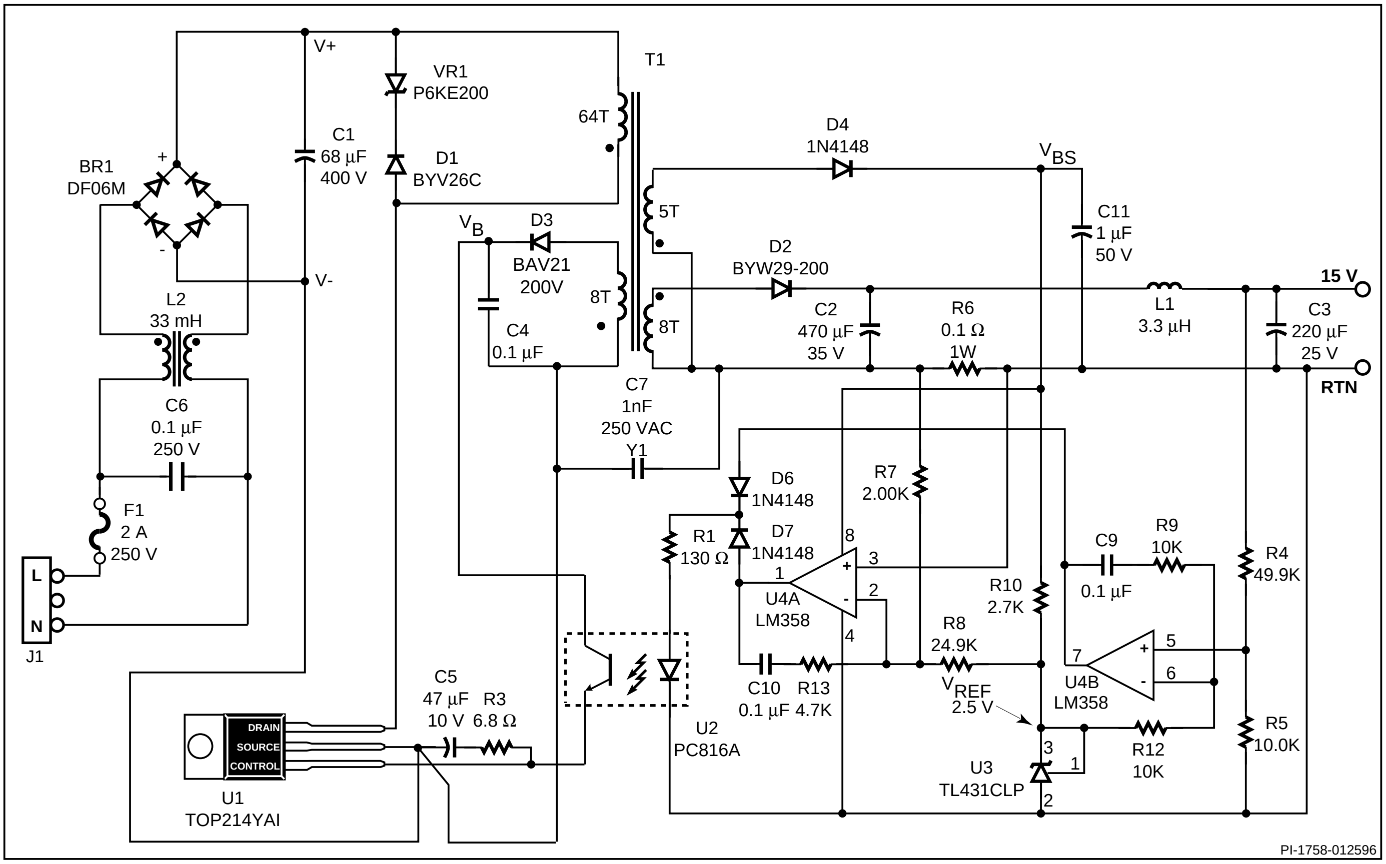

Figure 5. High Accuracy Constant Voltage/Constant Current 15V, 2A Supply Using the TOP214.

Output current is sensed by R6 . The voltage across R6 is compared by op amp U4A to the divider voltage determined by R7, R8, and reference U3. The output of op amp U4A drives current through D7 and R1 into the LED of optocoupler U2 to control duty cycle. The current control loop is compensated by C10 and R13. Diodes D6 and D7 "or" the op amp outputs together, such that the op amp with the highest output voltage drives optocoupler current and controls *TOPSwitch* duty cycle.

D4 and C11 generate bias supply V_{BS} for the secondary control circuit.

Performance

The output voltage and output current characteristic of the op amp current control circuit is shown in Figure 6 for input voltages of 85 VAC and 265 VAC. The curves for the two voltage extremes are almost identical. The slight difference in output current between the two extremes of input voltage is caused by the increased dissipation in the TL431 reference at high line voltage. This is caused by the increased amount of current through U3 due to the increase in secondary bias voltage V_{BS} at high line AC input voltage.

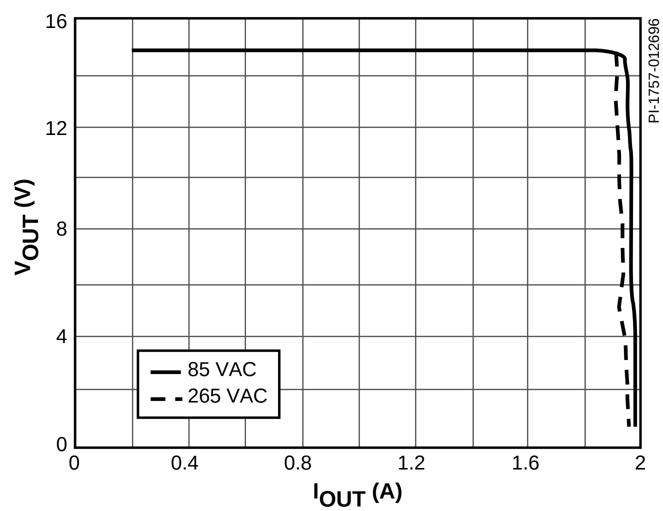

Figure 6. Output Voltage vs. Output Current Characteristic for High Accuracy Circuit shown in Figure 5.

Voltage Control Circuit

The output voltage of the supply is determined by R4, R5, and the reference voltage V_{REF} of U3 according to Equation (15):

$$V_O = V_{REF} \times \left(\frac{R4 + R5}{R5} \right) \tag{15}$$

In normal use, a battery may be connected to a battery charger disconnected or unplugged from AC input voltage. Total series resistance of R4 and R5 must be high enough to minimize effective battery discharge current. However, total series resistance of R4 and R5 should not be excessively high in order to avoid noise pickup. A value of 10kΩ is chosen for R5. This is a compromise between the above concerns. V_{REF} is the reference voltage of the U3 TL431 shunt regulator, which has a nominal value of 2.495 V. With these two values, R4 can be calculated by rearranging Equation (15):

$$\begin{aligned} R4 &= \frac{V_O - V_{REF}}{V_{REF}} \times R5 \\ &= \frac{15V - 2.495V}{2.495} \times 10k\Omega = 50.1k\Omega \end{aligned} \tag{16}$$

The closest standard 1% resistor value for R4 is 49.9K.

Current Control Circuit

The target output current I_{OT} is determined by V_{REF}, R6, R7, and R8 according to Equation (17):

$$I_{OT} = \frac{V_{REF} \times R7}{R6 \times R8} \tag{17}$$

The value for R7 is chosen as a compromise between loading effects on reference U3, noise pickup, and error introduced by the input bias current of op amp U4 . A value of 2 kΩ is a good compromise. R6 should be chosen to allow a relatively large current sense signal, but not so large that the dissipation at maximum output current is a concern. With R6 of 0.1 Ω, for a 2A output current, the available current sense signal is 0.2 V. Power dissipation is 0.4W, which requires a resistor power rating of 1 W. After values for R7 and R6 are chosen, R8 can be calculated by rearranging Equation (17):

$$\begin{aligned} R8 &= \frac{V_{REF} \times R7}{I_{OT} \times R6} \\ &= \frac{2.495V \times 2.00k\Omega}{2A \times 0.1\Omega} = 24.95k\Omega \end{aligned} \tag{18}$$

The closest standard 1% resistor value is 24.9 kΩ.

Primary and Secondary Bias Supplies

D4 and C11 rectify and filter the output of a secondary bias winding on T1 to bias the secondary control circuitry. The bias winding is connected in the forward polarity, so that D4 conducts when *TOPSwitch* is on. This means that the output voltage V_{BS} of the secondary bias winding will track the value of the input DC bus voltage V_{IN}, rather than the output voltage V_O, according to Equation (19):

$$V_{BS} = \left(V_{IN} \times \frac{N_{BS}}{N_P} \right) - V_{D4} \tag{19}$$

V_{BS} is the secondary bias voltage, V_{IN} is the voltage of the primary DC bus, N_{BS} is the number of turns in the T1 secondary bias winding, N_P is the number of transformer primary turns, and V_{D4} is the forward voltage drop of secondary bias rectifier D4. In op amp current control circuits, bias supplies must be independent of the output voltage to maintain control over output current when the power supply output is shorted by a low but non-zero impedance (soft short) and the output voltage is very low. If bias current were supplied from the output voltage or from a flyback winding (proportional to output voltage), the bias voltage under shorted output conditions would be too low for the LM358 op amp to source sufficient current through the optocoupler to control *TOPSwitch* duty cycle. Output current would become unregulated and exceed the target output current.

The value of current sense resistor R6 used for the op amp constant current circuit is selected to minimize power dissipation, and is generally 4-5 times smaller than the value used for a transistor current control circuit. If the op amp current control circuit becomes unregulated (due to insufficient bias voltage), voltage drop across R6 is too low to cut off the optocoupler and force the power supply into auto-restart. Instead, output current will stabilize at some value above the target output current, depending on actual impedance of the soft short. When the power supply output is shorted with zero impedance (hard short), output voltage may fall to a sufficiently low value to cut off the optocoupler and force the power supply into auto-restart. This results in an output voltage/output current characteristic as shown in Figure 7. If the current control circuit is powered from a bias source that is independent of output voltage, there is always sufficient voltage to bias the current control circuit, even if the output of the supply is soft or hard shorted. This prevents the out-of-control behavior depicted in Figure 7.

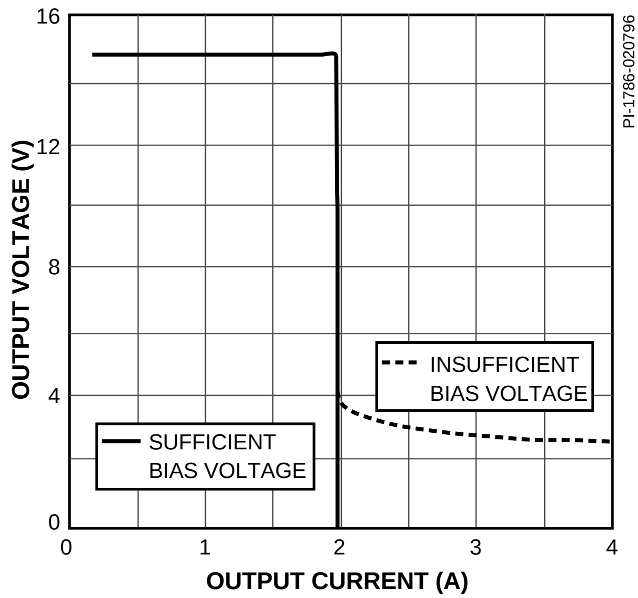

Figure 7. Constant Current Characteristic with Loss of Current Control at Low Output Voltage caused by Insufficient Secondary BiasVoltage.

At the minimum AC input voltage, secondary bias voltage V_{BS} must be high enough to bias the secondary control circuitry and keep the output voltage and output current under control. The LM358 op amp used for U4 has a minimum supply voltage rating of 3 volts but is characterized only for a supply voltage of 5 V or above. Therefore, 5 V will be used as the minimum bias voltage. This voltage must be sustained at the lowest AC input voltage, which is 85 VAC. The minimum value of DC bus voltage can be calculated using the formula:

$$V_{MIN} = \sqrt{(2 \times V_{ACMIN}^2) - \frac{2 \times P_O \times \left(\left(\frac{1}{2 \times f_L}\right) - t_C\right)}{\eta \times C1}} \quad (20)$$

V_{MIN} is the value of the DC bus voltage, V_{ACMIN} is the minimum AC input line voltage, P_O is the output power, η is the estimated supply efficiency, f_L is the AC line frequency, t_C is the conduction time of the input rectifier, and C1 is the input filter capacitance. Assuming η is 80%, f_L is 50 Hz, t_c is 3mS (a typical value), C1 is 68 uF, V_{ACMIN} is 85 VAC, and P_O is 30 W, the resulting value of minimum voltage is:

$$V_{MIN} = \sqrt{(2 \times 85^2) - \frac{2 \times 30 \times \left(\left(\frac{1}{2 \times 50}\right) - 3 \times 10^{-3}\right)}{0.8 \times 68 \times 10^{-6}}}$$
$$= 82V$$

From this value of minimum bus voltage, the number of secondary bias winding turns N_{BS} can be determined by rearranging Equation (19):

$$N_{BS} = \frac{N_P \times (V_{BS} + V_{D4})}{V_{IN}} \quad (21)$$

In this example, N_p is 64 turns, V_{BS} is 5 V, V_{D4} is 1 V, and V_{MIN} is 82 V. The resulting bias turns are:

$$N_{BS} = \frac{64 \times (5V + 1V)}{82V} = 4.7 \cong 5 \text{ turns}$$

This value can then be used to check the secondary bias voltage at high line (265 VAC) using Equation (19). The peak value of the DC bus voltage V_{MAX} at high line is given by Equation (22) where V_{ACMAX} is the maximum AC input voltage. All other values have already been determined. The maximum bias voltage is :

$$V_{MAX} = V_{ACMAX} \times \sqrt{2} = 265 \times \sqrt{2} = 375V \quad (22)$$

$$V_{BSMAX} = \left(V_{MAX} \times \frac{N_{BS}}{N_P}\right) - V_{D4}$$
$$= \left(375V \times \frac{5}{64}\right) - 1 = 28.3V \quad (23)$$

This voltage is within the rating of the LM358 op amp.

D3 and C4 provide a similar primary bias voltage V_B for the optocoupler and *TOPSwitch*. The primary bias voltage V_B is also supplied using a forward winding in order to maintain output voltage and current control at low output voltage. Determining the necessary number of turns for the primary bias is a procedure similar to that for the secondary bias winding as described above. The minimum voltage required at the primary winding is the *TOPSwitch* maximum Control pin voltage of 6.1 V, plus approximately 3 V to keep the optocoupler output transistor out of saturation. Thus the minimum primary bias voltage should be at least 9 V. Primary bias rectifier D3 has a voltage drop of 1 V. The minimum input DC bus voltage V_{MIN} has already been calculated as 82 V. The number of turns for the primary bias winding can be calculated from Equation (24):

$$N_B = \frac{N_P \times (V_B + V_{D3})}{V_{MIN}} = \frac{64 \times (9V + 1V)}{82V} \quad (24)$$
$$= 7.8 \cong 8$$

The maximum primary bias voltage can be determined using Equation (19):

$$V_{BMAX} = \left(V_{MAX} \times \frac{N_B}{N_P} \right) - V_{D3}$$
$$= \left(375V \times \frac{8}{64} \right) - 1V = 45.9V$$

The actual voltage across the optocoupler output transistor is equal to the bias voltage minus the *TOPSwitch* minimum Control pin voltage:

$$V_{U2} = V_{BMAX} - V_C = 45.9V - 5.5V = 40.4V \quad (25)$$

Optocoupler U2 should have a voltage rating of at least 50 volts. Transformer leakage inductance may cause actual bias voltage to be slightly higher while averaging effects may cause actual bias voltage to be slightly lower. Measure bias voltage at extremes of input voltage (85 VAC, 265 VAC) and full range of output current (no load to short circuit). If V_B is less than 9 V at 85 VAC input and short circuit load current, increase number of bias turns. If V_B is too high for optocoupler voltage rating at 265 VAC input, and V_B is higher than 9 V at 85 VAC input with short circuit load current, then the number of bias turns can be reduced. Otherwise, select an optocoupler with higher voltage rating. Refer to AN-14 for a table of suitable optocouplers.

Optocoupler Series Resistor

R1, in series with the optocoupler LED, defines the current available to control *TOPSwitch* and also helps to set the voltage and current control loop gains. The value for R1 is dependent on LM358 op amp positive saturation voltage V_{SAT}, forward voltage V_{D7} of diode D7, and forward LED voltage V_{LED} of optocoupler U2. R1 should be sized to allow full control of *TOPSwitch* at the minimum CTR value for U2. The value of R1 can be estimated from Equation (26):

$$R1 = \frac{(V_{SAT} - V_{D7} - V_{LED}) \times CTR_{MIN}}{I_{C(MAX)}} \quad (26)$$

From the *TOPSwitch* data sheet, $I_{C(MAX)}$ is the sum of external bias current and dynamic control current over temperature and is approximately 10mA. The optocoupler U2 has a minimum current transfer ratio (CTR) of 80%. V_{BS} has a minimum value of 5 V, and the V_{SAT} of the output stage of the LM358 is typically 3.5 V for a source current of 10mA, according to the data sheet. V_{LED} is typically 1.2 V, and V_{D7} is typically 0.65 V. This results in a R1 value of:

$$R1 = \frac{(3.5V - 0.65V - 1.2V) \times 0.8}{10mA} = 132\Omega$$

This is the maximum value for R1. The actual value of R1 used is 130Ω. The minimum value of R1 depends on control loop gain considerations. Too small a value for R1 leads to voltage and current control loops that oscillate and are difficult to stabilize.

Voltage Loop and Current Loop Frequency Compensation

C9, R9, and R12 compensate the voltage control loop. The values shown can be used as starting values for new circuits. Stability of the control loop can be tested by applying a 25% *resistive* load step to the output and observing the resulting voltage transient. The response should be well damped and free of ringing and oscillation.

C10 and R13 compensate the current control loop and the values shown can be used as starting values for new circuits. The value of C10 can be optimized in the circuit by applying a stepped *resistive* load to the output of the supply when operating in current control mode and examining the resulting output current using a DC current probe. The transient response should be well-damped, with no ringing or oscillation. Figure 4 shows a typical characteristic response for a well damped current control loop.

Constant Power Control

Constant power control is attractive for charger supplies in battery operated appliances such as portable computers, camcorders, etc., when it is desirable to simultaneously operate the appliance and charge the internal battery. In a constant power control circuit, the power supply output current increases as the output voltage of the supply decreases in a manner such that the product of output voltage and output current is constant. This allows a high rate of charge for a depleted battery with low terminal voltage, tapering off to a lower rate as the battery reaches its fully-charged state. Lead-acid batteries can efficiently accept a high rate of charge when in a depleted state, but the optimum rate of charge decreases as their terminal voltage rises to the fully charged condition. Current in excess of the optimum charge rate causes excessive battery power dissipation and heating. Constant power control offers a method of quickly and efficiently charging a battery, and minimizes power dissipation in the battery from the discharged state to full charge. A constant power supply also has advantages when working with the DC-DC converters used in the front-end of almost all notebook computer power supplies. These converters are used to efficiently generate the working voltages used inside the computer from the internal battery or external power supply. Most DC-DC converters have a negative resistance input characteristic, since for a given output power, current draw increases as the input voltage decreases. This characteristic can be incompatible with conventional constant current charging circuits, since the combination of a negative resistance input characteristic and a constant current source can cause abnormally low output voltage or poor energy delivery to the battery. This is not a problem with constant power control, as the current delivery capability increases at lower output voltage.

A common method of accomplishing constant power regulation is to use a discontinuous flyback supply. A discontinuous flyback converter operating with constant peak current is inherently a constant power device, and can be described by Equation (27):

$$P_O = \frac{L_P \times I_{LIMIT}^2 \times f_S}{2} \tag{27}$$

where P_O is the output power, L_P is the primary inductance, I_{LIMIT} is the peak primary current limit, and f_S is the switching frequency. This is a simple transfer function, but variations in primary inductance, operating frequency, and peak current limit from supply to supply cause unacceptable variations in output power. Another disadvantage is that the supply must operate in discontinuous mode. For a given output power, a discontinuous flyback circuit will operate at a higher peak switch current than a continuous flyback circuit, with a higher power loss and lower efficiency due to increased switch dissipation.

If the constant power control function is moved to the secondary side of the power supply, the control is dependent only on output voltage, and output current becomes independent of conditions on the primary side. Thus, a secondary referenced constant power circuit can be used with a continuous flyback supply with lower power loss and increased efficiency.

Op Amp Power Control Circuit

A secondary referenced constant power control circuit is shown in Figure 8. The circuit is a slight variant of the 15 V, 2A constant current power supply of Figure 5. Resistor R11, R12, and Zener diode VR2 have been added to the output current control circuit to allow the output current to change as a function of output voltage. With the proper choice of values, this circuit closely approximates a constant power function over a 2:1 change in output voltage which is sufficient for most battery charger circuits.

Circuit Description

The circuit approximates a constant power contour with an accuracy of +/-10% over a 2:1 output voltage swing using inexpensive, standard components. R6 monitors the output current of the supply. V_{REF}, R7, and R8 set the highest value of output current. VR2, R11, and R12 sense the output voltage to change the value of the output current in inverse proportion to the output voltage. When the output voltage drops below the Zener voltage of VR2, the circuit reverts to constant current mode, with an output current equal to twice the current value at highest output voltage.

In order to choose the proper component values, circuit behavior is defined at the inception of current control and also when output voltage is at half-value or $V_O/2$. Since the circuit is operating in constant power mode, the output current at $V_O/2$ should be twice the target output current I_{OT} at full output voltage V_O. Using this requirement, the necessary circuit behavior can be described using the two equations below:

At nominal output voltage V_O:

$$\begin{aligned}&(V_O - V_{VR2}) \times K_2 = \\ &((V_{REF} + (I_{OT} \times R_6)) \times K_1) - (I_{OT} \times R_6)\end{aligned} \tag{28}$$

At $V_O/2$:

$$\begin{aligned}&\left(\frac{V_O}{2} - V_{VR2}\right) \times K_2 = \\ &((V_{REF} + (2 \times I_{OT} \times R_6)) \times K_1) - (2 \times I_{OT} \times R_6)\end{aligned} \tag{29}$$

V_O is the nominal output voltage, I_{OT} is the target output current at nominal output voltage, V_{VR2} is the Zener voltage of VR2, V_{REF} is the reference voltage of U3. K_1 and K_2 are defined as follows:

$$K_1 = \frac{R_7}{R_7 + R_8} \quad K_2 = \frac{R_{11}}{R_{11} + R_{12}} \tag{30}$$

Equation (28) describes the behavior of the circuit at the inception of current control with V_O at nominal value and Equation (29) describes the behavior at $V_O/2$. There are two equations and seven variables, so some simplifying assumptions need to be made in order to derive circuit values. V_O and I_{OT} are predetermined by the power supply specification. R_6 is chosen in advance as a compromise between having an adequate current sense signal and limiting the power dissipation at the maximum output current. A 0.1 Ω, 3 watt resistor is shown, which limits power dissipation at 4A maximum output current to 1.6 watts. V_{REF} is the nominal TL431 reference value of 2.495 V. If V_{VR2} is chosen to be equal to $V_O/2$, K_1 can be easily calculated from values already known using Equation (29). This V_{VR2} value provides the closest fit to a constant power function over the 2:1 output voltage range.

If V_{VR2} is equal to $V_O/2$, then the left-hand side of the Equation (29) is equal to zero. The right-hand side of the Equation (29) can be rearranged to yield a solution for K_1:

$$K_1 = \frac{2 \times I_{OT} \times R6}{(V_{REF} + (2 \times I_{OT} \times R_6))} \qquad (31)$$

Using the values already chosen for the variables, K_1 is :

$$K_1 = \frac{2 \times 2A \times 0.1\Omega}{(2.495V + (2 \times 2A \times 0.1\Omega))} = \frac{0.4}{2.895} = 0.138$$

Once K_1 has been determined, K_2 is determined by rearranging Equation (28) as follows:

$$K_2 = \frac{\left((V_{REF} + (I_{OT} \times R_6)) \times K_1\right) - (I_{OT} \times R_6)}{(V_O - V_{VR2})} \qquad (32)$$

Using the above value for K_1 and a value of 7.5 V for V_{VR2}, K_2 can be calculated as:

$$K_2 = \frac{((2.495V + (2A \times 0.1\Omega)) \times 0.138) - (2A \times 0.1\Omega)}{(15V - 7.5V)}$$

$$= \frac{0.372 - 0.2}{7.5} = 0.0229$$

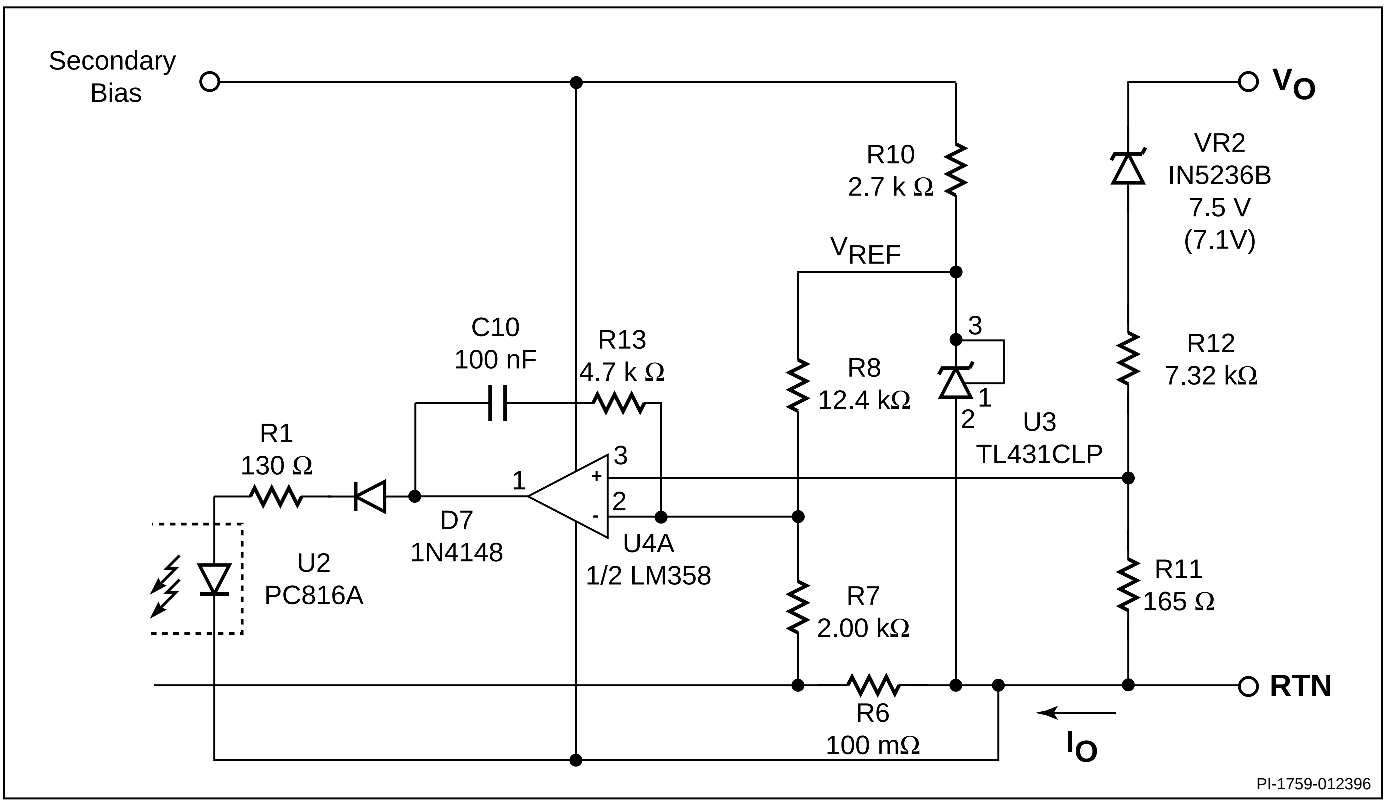

Figure 8. Op Amp Implementation of the Constant Power Circuit.

Choosing R7 and R8

The value for R7 is chosen as a compromise between loading effects on reference U3, noise pickup, and error introduced by the input bias current of op amp U4 . A value of 2kΩ is a good compromise. If R7 is chosen to be 2kΩ, R8 can be calculated by rearranging Equation (30):

$$\begin{aligned} R8 &= \frac{R7 \times (1 - K_1)}{K_1} \\ &= \frac{2k\Omega \times (1 - 0.138)}{0.138} = 12.49k\Omega \end{aligned} \qquad (33)$$

The closest standard 1% resistor value is 12.4kΩ. The resulting K_1 value is :

$$K_1 = \frac{R7}{R7 + R8} = \frac{2k\Omega}{2k\Omega + 12.4k\Omega} = 0.139$$

This value of K_1 should be used to recalculate K_2:

$$\begin{aligned} K_2 &= \frac{((2.495V + (2A \times 0.1\Omega)) \times 0.139) - (2A \times 0.1\Omega)}{(15V - 7.5V)} \\ &= \frac{0.375 - 0.2}{7.5} = 0.0233 \end{aligned}$$

Choosing R11 and R12

A battery may be connected to a battery charger disconnected or unplugged from AC input voltage. Total series resistance of R11 and R12 must be high enough to minimize effective battery discharge current. At the same time, the current through R11 and R12 must be sufficient to bias Zener VR2. A compromise current of 1 mA is chosen for this circuit. This will affect the voltage drop of VR2. Assuming that the bias current of the op amp U4 is negligible, the current through R11 and R12 is as follows:

$$\frac{V_O - V_{VR2}}{R11 + R12} = 1mA \qquad (34)$$

For V_O of 15 V and V_{VR2} of 7.5 V, the sum of R11 and R12 is:

$$\begin{aligned} R11 + R12 &= \frac{V_O - V_{VR2}}{1mA} \\ &= \frac{15V - 7.5V}{1mA} = 7.5k\Omega \end{aligned} \qquad (35)$$

R11 will be equal to:

$$\begin{aligned} R11 &= K2 \times (R11 + R12) \\ &= 0.0233 \times 7.5k\Omega = 175\Omega \end{aligned} \qquad (36)$$

174Ω is the closest standard 1% resistor value. R12 is calculated using this value:

$$\begin{aligned} R12 &= 7.5k\Omega - R11 \\ &= 7.5k\Omega - 174\Omega = 7.326k\Omega \end{aligned} \qquad (37)$$

7.32kΩ is the closest 1% value.

Correcting for Zener Voltage of VR2

The voltage drop for Zener diodes is specified in the data sheets at a test current I_{ZT}. Values of operating current less than this test current result in a Zener voltage that is less than the specified value.

The 1N5236B Zener used for VR2 is specified at an I_{ZT} of 20mA. At a current of 1 mA, the Zener voltage is approximately 7.1 V rather than the specified value of 7.5 V. If this Zener is used without adjusting the values in the power control circuit, the initial output current will be lower than expected. The actual voltage across VR2 can be measured in circuit and used to provide corrected values for K_2 and R11. For a VR2 voltage of 7.1 V, the corrected value of K_2 is:

$$\begin{aligned} K_2 &= \frac{((2.495V + (2A \times 0.1\Omega)) \times 0.139) - (2A \times 0.1\Omega)}{(15V - 7.1V)} \\ &= \frac{0.375 - 0.2}{7.9} = 0.0222 \end{aligned}$$

If R11 and R12 total series resistance is still assumed to be 7.5kΩ, the new value for R11 is:

$$R11 = K2 \times (R11 + R12) = 0.0222 \times 7.5k\Omega = 167\Omega$$

165Ω is a standard 1% resistor value. R12 is then calculated:

$$R12 = 7.5k\Omega - R11 = 7.5k\Omega - 165\Omega = 7.34k\Omega$$

7.32kΩ is the closest standard 1% resistor value.

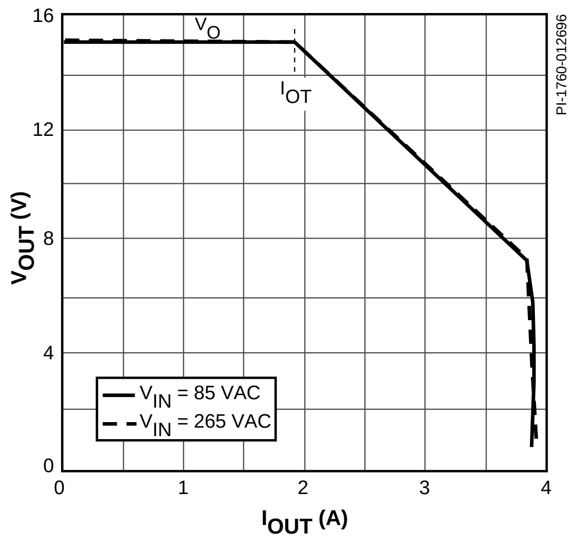

Figure 9. Output Voltage vs. Output Current for Constant Power Circuit Shown in Figure 8.

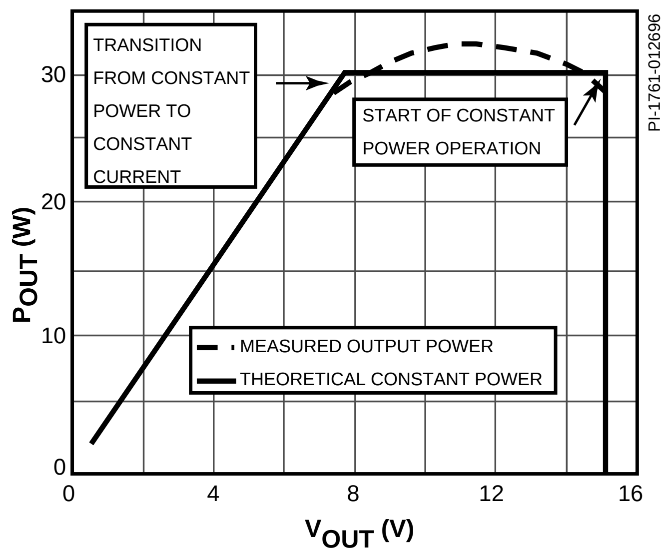

Figure 10. Comparison of Constant Power Circuit with True Constant Power Characteristic.

Circuit Performance

Actual performance of the constant power circuit is shown in Figures 9 and 10. Figure 9 shows the output voltage and output current characteristic for the constant power regulator at minimum and maximum line voltage. Output voltage and output current data at the two input extremes of 85 VAC and 265 VAC are almost identical. Figure 10 shows the output power and output voltage data of the constant power regulator compared to an ideal constant power circuit that reverts to constant current below 7.5 V output. The constant power circuit of Figure 8 approximates a true constant power characteristic with an error of +/- 10%.

Op amp Current Limit Circuit with No Secondary Bias Winding

Figure 11 shows another constant voltage/constant current supply using an op amp for current control. A Zener diode is used for voltage control. This circuit also provides more accurate control over output current compared to the simple transistor circuit. Power loss is lower and efficiency is better because smaller resistance values can be used for sense resistor R6.

Circuit Description

Output voltage is rectified and filtered by D2, C2, C3, and L1. Note that diode D2 has been inverted and moved from the dot side to the non-dot side of the secondary winding. This connection has exactly the same flyback rectification and filtering effect as the “flyback” circuit shown in Figure 1 while offering the advantage of generating a secondary bias voltage without using a secondary bias transformer winding.

D4, R15, VR4, Q1, and C11 combine a rectifier and filter with a simple, linear regulator to generate a clamped secondary bias voltage V_{BS}. D4 conducts current only when *TOPSwitch* is on (which means that rectification occurs in the “forward” mode).

Output voltage is determined by voltage drops across VR2, D8, and the LED inside optocoupler U2 (similar to circuit shown in Figure 1). Diode D8 blocks current in the forward direction through VR2 which may occur when the power supply output is short circuited.

Output current is sensed by R6. Voltage across R6 is compared by op amp U4 to the divider voltage determined by R7, R8, and 5.1 V Zener diode VR3. The op amp output drives current through D7 and R1 into the LED of optocoupler U2 to control duty cycle. The current control loop is compensated by C10 and R13. (Current control is similar to the circuit shown in Figure 5.)

Performance

The output voltage and output current characteristic of the op amp control circuit is shown in Figure 12 for input voltages of 85 VAC and 265 VAC. The curves for the two voltage extremes are almost identical. Differences between 85 VAC and 265 VAC data are minimal because the clamped secondary bias voltage V_{BS} results in constant dissipation in the Zener diode VR3 over the input voltage range.

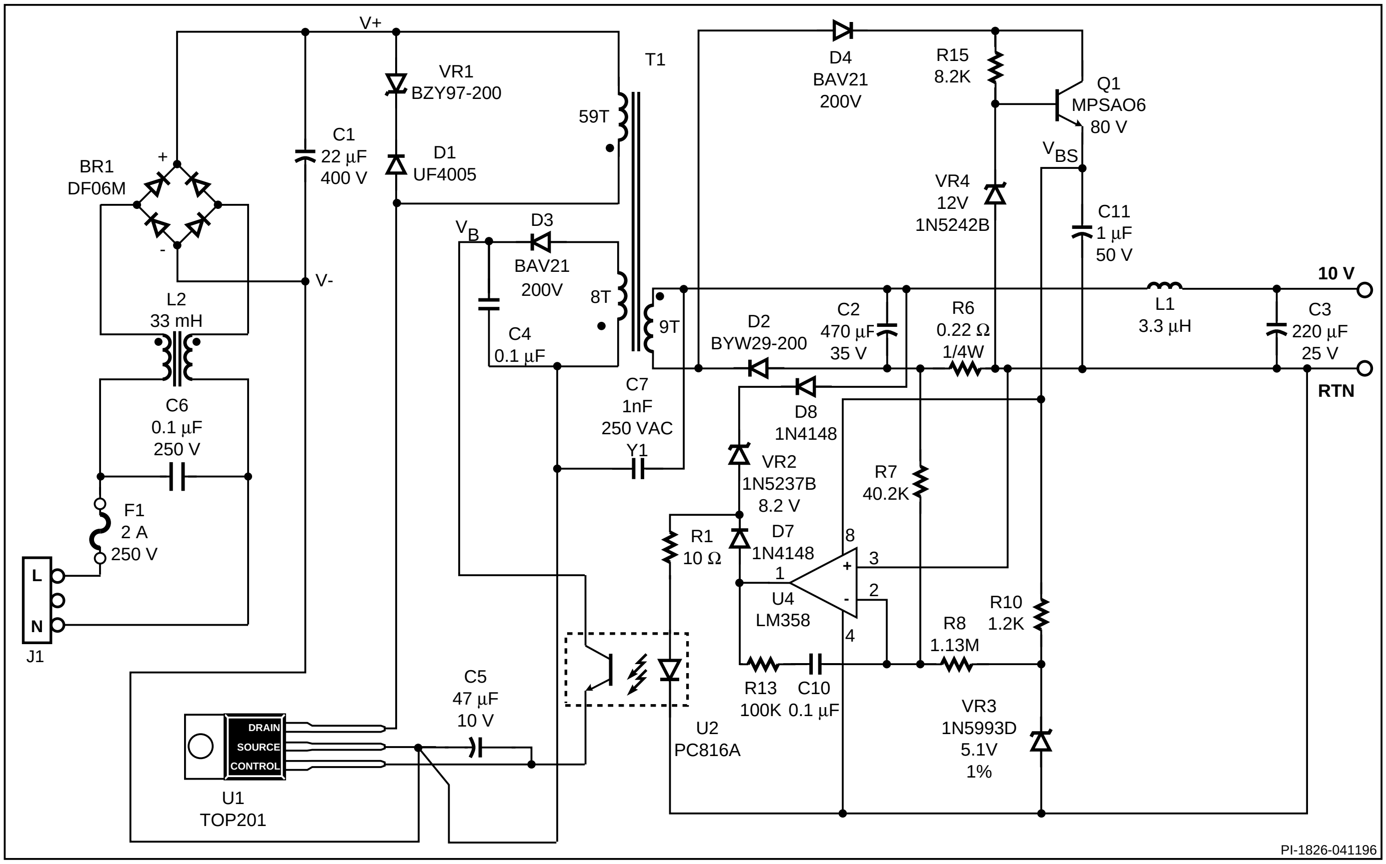

Figure 11. High Accuracy Constant Voltage/Constant Current 10V, 800 mA Supply Using the TOP201 with Doubler Diode Connection for Secondary Bias.

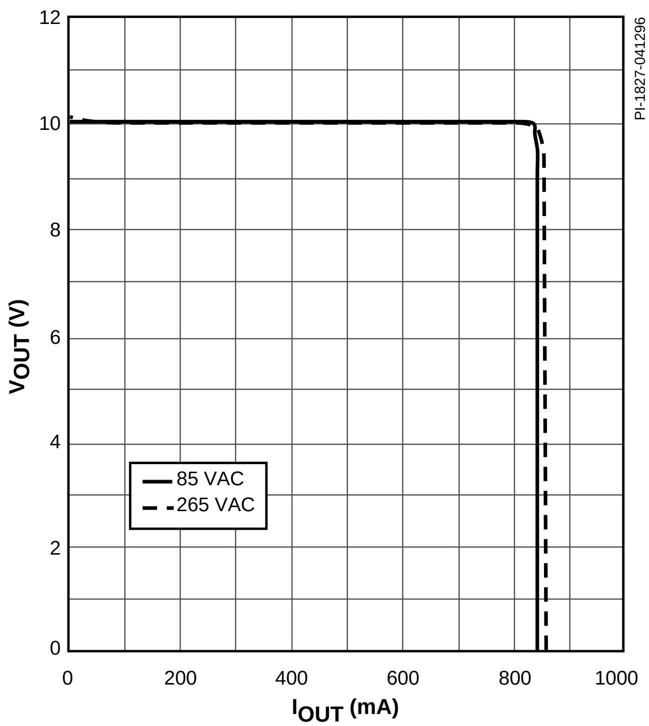

Figure 12. Output Voltage vs. Output Current Characteristic for High Accuracy Circuit with Doubler Bias Shown in Figure 11.

Voltage Control Circuit

V_{VR2}, V_{D8}, V_{LED}, and the voltage drop V_{R1} (across R1) determine the output voltage. V_{VR2} is the typical 8.2 V voltage drop across Zener diode VR2, V_{D8} is the 0.65 V voltage drop across blocking diode D8, and V_{LED} is the 1.2 V voltage drop across the optocoupler LED. Refer back to the simple, low cost circuit analysis (Equations 1-4) to determine the voltage V_{R1} across R1 as a function of optocoupler CTR and *TOPSwitch* Control pin current I_C.

$$\begin{aligned} V_O &= V_{VR2} + V_{D8} + V_{LED} + V_{R1} \\ &= 8.2V + 0.65V + 1.2V + (3.75mA \times 10\Omega) \\ &= 10.1V \end{aligned} \quad (38)$$

Current Control Circuit

The target output current I_{OT} is determined by VR3, R6, R7, and R8 according to Equation (39):

$$I_{OT} = \frac{V_{VR3} \times R7}{R6 \times R8} \quad (39)$$

The value for R7 is chosen relatively high (40.2kΩ) to tailor op amp frequency response. R6 should be chosen to allow a relatively large current sense signal, but not so large that the dissipation at maximum output current is a concern. R6 is 0.22Ω, resulting in an available current sense signal of 0.176 V for an output current of 800 mA. Power dissipation is 0.14 W requiring a resistor power rating of 0.25W. R8 can be calculated from the above values by rearranging Equation (39):

$$R8 = \frac{V_{VR3} \times R7}{I_{OT} \times R6} = \frac{5.1V \times 40.2k\Omega}{800mA \times 0.22\Omega} = 1.165M\Omega \quad (40)$$

The closest standard 1% resistor value is 1.13MΩ.

Primary Bias Supply

D3 and C4 rectify the primary bias winding in the forward direction to provide primary bias voltage V_B for the optocoupler and *TOPSwitch*. To calculate the number of turns for the primary bias winding, refer back to the previous op amp control circuit analysis (Equations 24-25).

$$N_B = \frac{N_P \times (V_B + V_{D3})}{V_{MIN}} = \frac{59 \times (9V + 1V)}{82V} = 7.2 \cong 8T$$

$$V_{BMAX} = (V_{MAX} \times \frac{N_B}{N_P}) - V_{D3} = (375V \times \frac{8}{59}) - 1V = 49.9V$$

$$V_{U2} = V_{BMAX} - V_C = 49.9V - 5.5V = 44.4V$$

The optocoupler selected for U2 must have a voltage rating of at least 60 volts. Measure bias voltage under conditions described for previous op amp circuit, adjust bias winding turns if necessary, and determine optocoupler voltage rating. Refer to AN-14 for a table of suitable optocouplers.

Secondary Bias Supply

Diode D4 has a peak cathode voltage V_{CD4} depending on DC input voltage V_{IN}, transformer primary turns N_P, secondary turns N_S, D4 forward voltage V_{D4}, and output voltage V_O:

$$V_{CD4} = V_O + (\frac{N_S}{N_P} \times V_{IN}) - V_{D4} \quad (41)$$

At low input voltage, C11 charges up through Q1 to approximately D4 cathode voltage V_{D4}, which is less than Zener voltage of VR4. At high input voltage, VR4 clamps the base of Q1 to approximately 12 V to limit the maximum voltage across C11. Transistor Q1 has maximum collector to emitter voltage V_{CE} at high line input voltage according to Equation (42):

$$V_{CE} = V_{CD4} - (V_{VR4} - V_{BEQ1}) \quad (42)$$

With high line input voltage of 265 VAC, maximum DC bus voltage V_{MAX} was found previously (using Equation (22)) to be 375 V. The transformer has 59 primary turns (N_P) and 9 secondary turns (N_S). For an output voltage V_O of 10 V, Zener diode VR4 voltage of 12 V, D4 forward voltage V_{D4} of 1 V, and base emitter voltage V_{BE1} of 0.65 V, maximum collector to emitter voltage V_{CE} of transistor Q1 is found from Equation (42) to be:

$$V_{CE} = 10V + (\frac{9}{59} \times 375V) - 1V - 12V + 0.65V = 54.8V$$

Q1 must have a collector to emitter voltage rating of at least 60 V but preferably 80 V.

To properly bias the op amp, minimum secondary bias voltage V_{BSMIN} must be at least 5 V at low line input voltage with the output under short circuit conditions. With low line input voltage of 85 VAC, minimum DC bus voltage V_{MIN} was found previously (using Equation (20)) to be 82 V. Zener diode VR4 does not clamp secondary bias voltage V_{BS} at low line input voltage and output voltage V_O is essentially zero under short circuit conditions. V_{BSMIN} is found from V_{MIN}, primary turns N_P, secondary turns N_S, D4 forward voltage V_{D4}, and base to emitter voltage V_{BEQ1} of Q1:

$$V_{BSMIN} = (V_{MIN} \times \frac{N_S}{N_P}) - V_{D4} - V_{BEQ1} = (82V \times \frac{9}{59}) - 1V - 0.65V = 10.9V \quad (43)$$

Averaging effects or Q1 turn on time may cause actual secondary bias voltage to be slightly lower. Measure bias voltage V_{BS} at extremes of input voltage (85 VAC, 265 VAC) and full range of output current (no load to short circuit). If V_{BS} is less than 5 V at 85 VAC with short circuit load current, increase number of secondary bias turns. Measure D4 cathode voltage V_{CD4} to verify Q1 collector-emitter voltage is within rating. If necessary, select a higher voltage transistor for Q1.

Current Loop Frequency Compensation
Resistor R1 must have a low value for the Zener diode voltage control loop (using VR2) to regulate accurately. R1 also determines the current loop DC gain which is higher than the previous op amp circuit shown in Figure 5. R7, R8, and R13 have been scaled higher in absolute resistance value (relative to value of capacitor C10) to reduce bandwidth and stabilize the current control loop.

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2. Power Integrations, AN-16, "*TOPSwitch* Flyback Design Methodology"

3. Power Integrations, AN-17, "Flyback Transformer Design for *TOPSwitch* Power Supplies"

4. Power Integrations, AN-18, "*TOPSwitch* Flyback Transformer Construction Guide

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