Simple Bias Supplies Using the TOP200

Design Note DN-8

POWER
INTEGRATIONS, INC.

Power systems often require significant amounts of bias power from a small, separate, regulated power supply. Applications which use bias supplies include computer power supplies, uninterruptible power systems, motor controllers, electric power meters, and appliance controllers. A housekeeping power supply is often required in equipment designed for long periods of low power quiescent operation. The supply provides power for separate circuitry waiting for input. One important application is the “Green PC” which enters a low power sleep mode until awakened by keyboard entry or I/O activity. Fax machines, televisions, VCRs, video monitors, laser printers, and copiers are other applications that often require small housekeeping supplies.

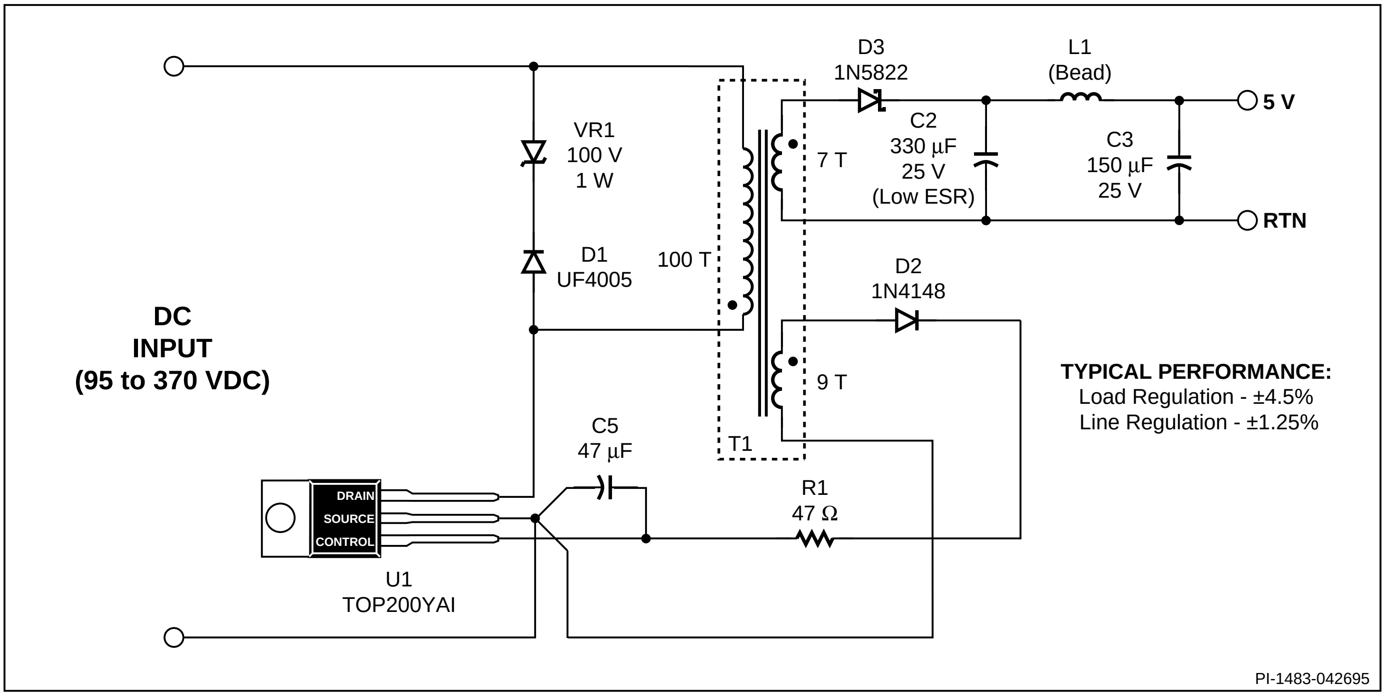
Two common methods for supplying bias power are linear supplies utilizing a line frequency transformer or switching power supplies using a blocking oscillator circuit. These approaches can encounter difficulties when wide range input, high efficiency, and/or small size is required. Using *TOPSwitch*®, a primary regulated switching power supply can be constructed that approaches the simplicity of a linear design, but with wide range input, good efficiency, and very small size and weight.

Application Circuits

For the sake of clarity, the following application circuits will assume an AC input voltage has been rectified and filtered elsewhere in the system such that high voltage DC is available. If the circuit is located far away from the bulk high-voltage supply, a decoupling capacitor of 0.1 to 1 µF is recommended across the high-voltage input. This capacitor should be placed so that it is close to U1 and T1, and the loop between the decoupling capacitor, U1, and T1 is as small as possible.

The circuit descriptions below assume some familiarity with the functioning of *TOPSwitch*. For greater detail on the *TOPSwitch* integrated circuit, please refer to the data sheets for TOP100-4 and TOP200-4, as well as AN-14.

The circuits shown in Figures 1 and 2 are simple 5 V, 5 watt bias supplies using the TOP200. Both circuits are full-range input flyback power supplies that employ primary-side regulation from a transformer bias winding. This approach is best for low-cost applications requiring isolation and operation within a narrow range of load power. Line and load regulation of ±5%



TYPICAL PERFORMANCE:
Load Regulation - ±4.5%
Line Regulation - ±1.25%

Figure 1. Minimum Part Count Bias Supply Using the TOP200.

December 1995

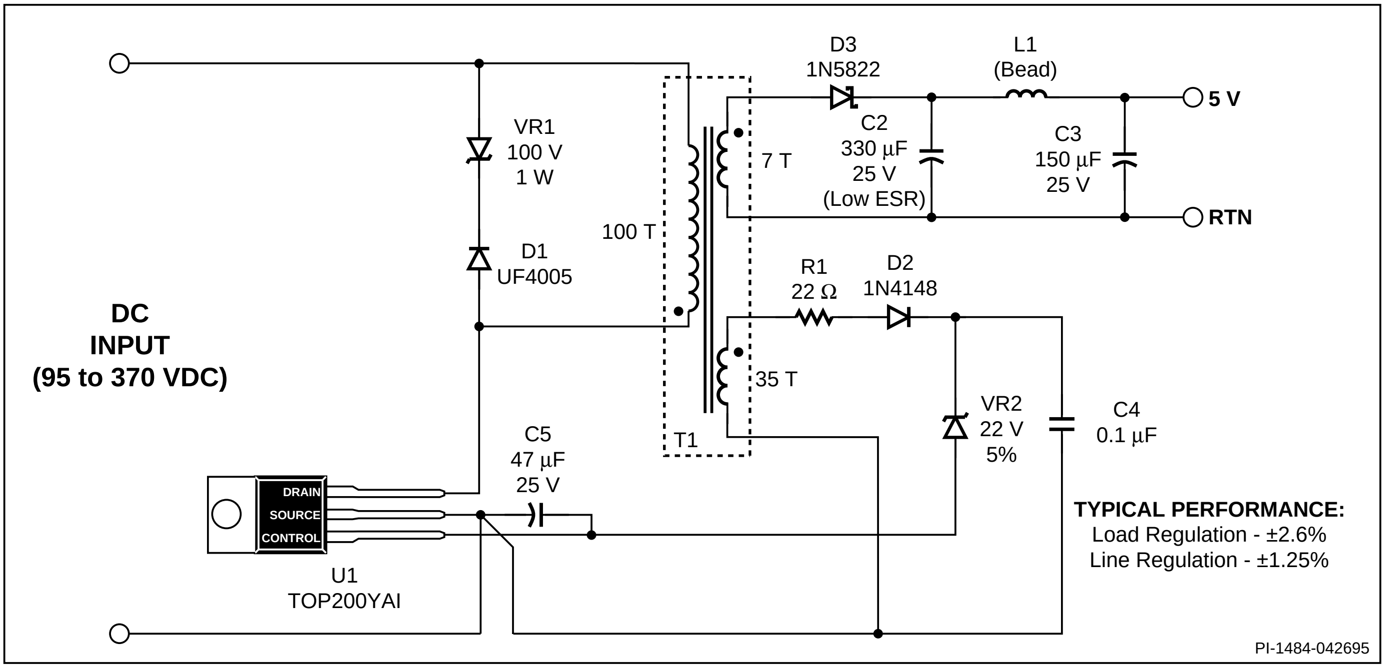


Figure 2. Improved Load Regulation Bias Supply Using the TOP200.

Application Circuits (cont.)

or better can be achieved from 10% to 100% of rated load. Careful attention is required in transformer construction and layout to reduce the effects of stray inductance.

Voltage feedback in both circuits is obtained from the transformer bias winding, which eliminates the need for optocouplers and secondary-referenced error amplifiers. High-voltage DC is applied to the primary winding of T1. The other side of the transformer primary is driven by the integrated high-voltage MOSFET transistor within the TOP200 (U1). Both circuits operate at a switching frequency of 100 kHz, set by the internal oscillator of the TOP200. The clamp circuit implemented by VR1 and D1 limits the leading-edge voltage spike caused by transformer leakage inductance to a safe value. The 5 V power secondary winding is rectified and filtered by D3, C2, C3, and L1 to create the 5 V output voltage.

The circuits in Figures 1 and 2 differ only in the manner in which voltage feedback from the bias winding is used to control the output voltage. In the circuit of Figure 1, the output of the T1 bias winding is rectified and filtered by R1, D2, and C5. The voltage across C5 is regulated by U1, and is determined by the voltage drop across the 5.7 V internal shunt regulator at the CONTROL pin of U1. A voltage on C5 greater than the drop across the CONTROL pin of U1 will cause current to flow into the CONTROL pin of U1. The operating duty cycle is inversely proportional to the control current, with the result that U1 will adjust its duty cycle to minimize the control current to a value that maintains the 5.7 V control voltage. The output voltage of the bias winding is reflected by the turns ratio between the bias winding and the output winding to determine the output voltage. C5 is used to bypass the CONTROL pin of U1. It also provides loop compensation for the power supply by shunting AC currents around the CONTROL pin, and also determines the auto-restart frequency of U1 during start-up and short circuit conditions. See the TOP100-4 or TOP200-4 data sheets for more detail on these functions.

The circuit of Figure 2 is slightly more complex, and should be used when tighter load regulation is required. C4 and VR2 have been added, and the supply is now regulated by the combination of the internal 5.7 V shunt regulator at the CONTROL pin of U1 and the voltage drop across VR2.

Performance Comparison

Line and load regulation curves are shown below in Figures 3 and 4. The minimum parts count supply has looser load regulation than the circuit in Figure 2, with ±4.5% load regulation from 10% to 100% of full load. Line regulation from 95 to 370 VDC is ±1.25%. Load regulation for the circuit in Figure 2 is ±2.6%, from 10% to 100% of full load. Its output line regulation is the same as that of Figure 1. If minimum parts count and lowest cost are the most important features, the minimum parts count circuit of Figure 1 can be used. When output load regulation is paramount, the circuit of Figure 2 should be chosen.

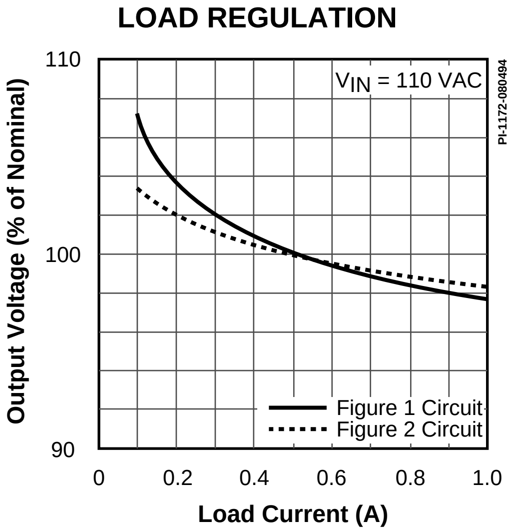


Figure 3. Load Regulation Performance Comparison.

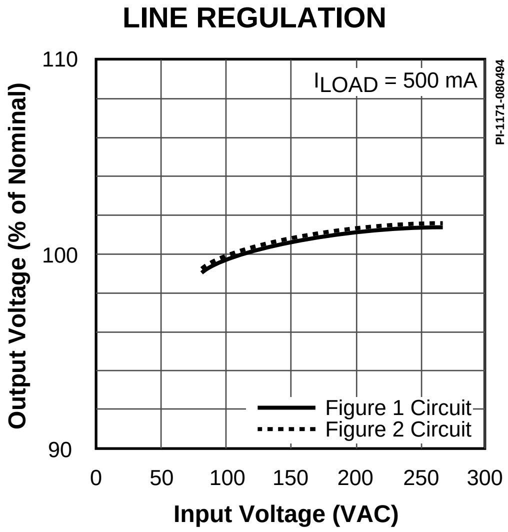


Figure 4. Line Regulation Performance Comparison.

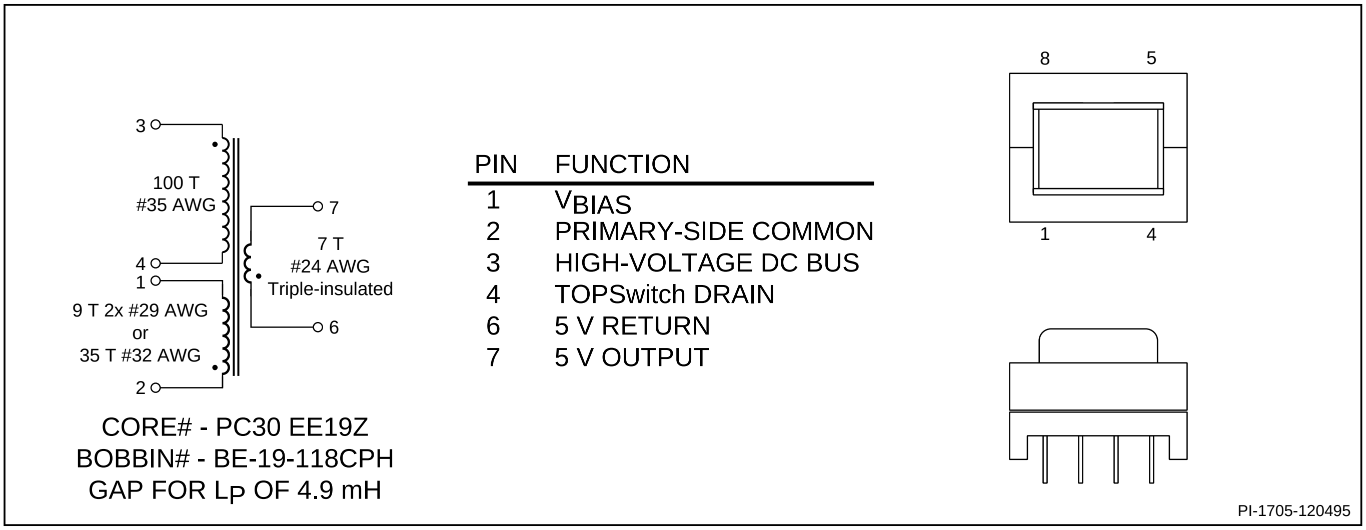
Design Parameters

Transformer Design

A practical transformer design for bias and housekeeping power supplies must have the following features:

- Low-cost winding techniques
- High inductance
- Low capacitance
- Adequate electric strength
- Low radiated emissions (EMI)
- Bias/feedback winding closely coupled to output winding

Application Note AN-7 details many techniques for practical high-frequency transformer design. The transformer used in the design examples of Figures 1 and 2 is shown in Figure 5. The overall design for each transformer is the same, with the exception of the number of turns in the bias winding. The minimum parts count circuit of Figure 1 requires a bias winding of 9 turns, while the circuit of Figure 2 requires a bias winding of 35 turns.



PIN	FUNCTION
1	V_{BIAS}
2	PRIMARY-SIDE COMMON
3	HIGH-VOLTAGE DC BUS
4	TOPSwitch DRAIN
6	5 V RETURN
7	5 V OUTPUT

CORE# - PC30 EE19Z
BOBBIN# - BE-19-118CPH
GAP FOR L_P OF 4.9 mH

Figure 5. T1 Transformer Pinout and Physical Dimensions.

Design Parameters (cont.)

This transformer has several special design features worthy of note. The first of these is the use of triple-insulated wire for the transformer secondary. This allows the elimination of tape margins for safety spacing, enabling the full width of the bobbin to be used for the primary and bias windings. This allows a smaller core to be used for a given power and primary inductance level. Because of the better utilization of the bobbin width and the consequent reduction of winding build, acceptable values of leakage inductance can be obtained with a conventional two layer primary winding rather than a split primary.

The transformer primary is applied first. The switching MOSFET is connected to the start of the primary, so that the half of the primary winding that has the largest voltage excursion is shielded from the other windings by the remainder of the primary winding. This reduces the capacitive coupling from the primary to the secondary, and reduces radiated coupling to other parts of the power supply circuit. The secondary is wound on top of the primary using triple-insulated wire, and the bias winding is applied last. This particular winding order is important to the proper functioning of the supply. Since the bias winding is applied last, it is closely coupled to the secondary, but only loosely coupled to the primary. This allows the voltage on the bias winding to track the voltage on the output of the supply. Because the bias winding is only loosely coupled to the primary, it is less subject to peak charging due to the primary turn-off leakage spike. The magnitude of the primary leakage spike is a function of the peak primary current. At high output currents, peak charging from this leakage spike in the bias supply tends to cause *TOPSwitch* to reduce its operating duty cycle, resulting in output droop at high output power. Placing the bias winding last allows the leakage spike to be filtered effectively with a single resistor in series with the bias winding.

The design and wire size of the bias winding is selected to form a single layer completely traversing the width of the transformer bobbin. This is done by selecting a wire size much larger than needed to handle the current delivered by the bias supply. In the case of the transformer for the simple circuit of Figure 1 with only 9 turns on the bias winding, a bifilar construction is used to increase the fill factor. Having the bias winding extend the full width of the bobbin maximizes the coupling to the secondary winding directly beneath it, improving regulation. Since the bias winding fills the width of the bobbin, there is no opportunity for variation in the position of the bias winding, and the unit-to-unit consistency is improved.

Selecting the Bias Winding Filter Resistor

The bias winding filter resistor (R1) plays an important part in determining the load regulation of the supply. Under ideal conditions, the load regulation of the supply would be determined by the difference in voltage drop between the bias rectifier and output rectifier, as well as the difference in resistance between the bias and output windings. Under actual conditions, the transformer leakage inductance causes a voltage spike to occur when the primary switching MOSFET turns off. With no filter on the feedback winding, this spike peak charges the bias supply and causes the secondary output to droop excessively as the output load is increased. With R1 in place, the turn-off spike is filtered out, and the bias voltage more accurately reflects the final value of the primary reflex voltage, rather than the initial turn off spike. This improves the load regulation of the output. However, this gain in load regulation does have a drawback. The same filtering process that makes the bias winding less sensitive to the turn-off leakage spike also makes the bias supply less sensitive to peak charging of the output at low or zero output load. This means that the output voltage rise at low or zero load will be higher with the filter resistor than without it. These two effects must be traded off when selecting a value for the filter resistor. There is generally an optimum value of resistance, obtainable by experiment, that provides acceptable load regulation without resulting in excessive output voltage rise at zero load. In the case of the circuits of Figure 1 and Figure 2, the optimum value of filter resistance was found by experiment to be 22 Ω. This resistance will vary depending on the transformer leakage inductance, the degree of coupling between the primary and bias windings, and the amount of capacitance after R1. The optimum value will generally be somewhere between 10 to 100 Ω. A reasonable starting value for experimentation in a new design is an intermediate value such as 47 Ω.

Selecting Components for Determining Output Voltage

In a primary-regulated bias supply, the output voltage is regulated by controlling the voltage of the primary bias winding. Since the output winding and bias winding are coupled together, the output voltage will tend to track the bias voltage. They are related by the turns ratio between the bias winding and the output winding. The output voltage of the supply of Figure 2 can be expressed as follows:

$$V_{OUT} = \left(\left(5.7V + V_{VR2} + V_{D2} + I_C \left(15 + Z_{VR2} \right) \right) \times \frac{N_S}{N_B} \right) - V_{D3} \quad (1)$$

V_{OUT} is the output voltage of the supply and V_{VR2} is the VR2 Zener voltage. 5.7 V is the voltage of the U1 internal shunt regulator voltage. I_C is the current through VR2 flowing into the CONTROL pin of U1, which provides the operating current for U1, and also establishes its operating duty cycle. 15 Ω is the dynamic impedance of the *TOPSwitch* internal shunt regulator, and Z_{VR2} is the dynamic impedance of Zener diode VR2. N_S is the number of turns in the T1 secondary winding, and N_B is the number of turns in the bias winding. V_{D3} is the voltage drop of the output rectifier D3. In the case of the circuit of Figure 1, VR2 is not used, and the terms V_{VR2} and Z_{VR2} fall out of the expression, leaving:

$$V_{OUT} = \left(\left(5.7V + V_{D2} + \left(I_C \times 15\right)\right) \times \frac{N_S}{N_B}\right) - V_{D3} \qquad (2)$$

The *TOPSwitch* control current (I_C) has a range of 2.5 mA minimum at maximum duty cycle to 6.5 mA for minimum duty cycle. The control current and impedance terms introduce a variation of around 1% in the final value of the output voltage. This can be ignored or taken into account with an approximate midrange value for control current (I_C) of 5 mA when choosing component values to set the output voltage of the supply.

In the circuit example of Figure 2, a relatively high value of Zener voltage (22 V) was selected for VR2. There are two reasons behind this choice. The first is that use of accurate Zener with voltage of several times the *TOPSwitch* control voltage will tend to reduce the effect of variations in the value of the 5.7 V CONTROL pin voltage of *TOPSwitch*, since the *TOPSwitch* control voltage will be less of a factor in determining the output voltage as compared to VR2. A 1% tolerance Zener can be selected to further tighten the absolute tolerance of the output voltage for this circuit. Higher voltage Zeners also tend to have a sharper "knee" in their V-I characteristic than low voltage Zeners. This provides higher gain in the control loop for tighter control of load regulation, since a sharper breakdown characteristic will result in a larger change of current into the control pin of U1 for a given change in bias voltage. This is borne out by the load regulation of the circuit in Figure 2. The high value of bias voltage also tends to provide a pre-load to the supply when the output load is small. When the output load on the supply is reduced to a small value, the feedback control circuit will source current into the CONTROL pin in order to maintain regulation. Since the bias voltage is relatively high, the current sourced into the CONTROL pin represents a substantial amount of power (almost 200 mW if minimum output duty cycle is programmed). This loads the supply and tends to curb output voltage rise for low or zero load conditions.

Factors Affecting Line and Load Regulation

For optimum load regulation, a value of transformer primary inductance should be chosen such that the power supply remains in the continuous mode over as much of its specified input line and output load range as possible. For a flyback supply operating in the continuous mode, the duty cycle is independent of output load to a first order, and varies only in response to input voltage. This is important because of the effect of variations in *TOPSwitch* control current on the output voltage of the supply. These effects are represented by the Zener impedance terms and the *TOPSwitch* control impedance terms in Equations (1) and (2) above. The product of I_C and the various circuit impedances forms an error term that causes regulation to be less than perfect. Since the CONTROL pin input current I_C must vary to control the duty cycle, any change in duty cycle will cause a change in the voltage drop across the Zener and/or *TOPSwitch* input dynamic impedance, causing a shift in output voltage and a consequent deterioration in load regulation. If the power supply operates in the discontinuous mode, its operating duty cycle will be a strong function of output loading. This means that the *TOPSwitch* control current will also vary with load, and regulation will suffer. The circuit of Figure 2 can be used as an example. It uses a 22 V Zener to set the output voltage in conjunction with the *TOPSwitch* control voltage. The impedance of a typical 22 V Zener (1N5251) is 30 Ω. The *TOPSwitch* CONTROL pin dynamic impedance is 15 Ω. From the *TOPSwitch* data sheet, it can be found that the *TOPSwitch* control function dynamic range is 4 mA. This is the control current required to drive the *TOPSwitch* duty cycle from maximum to minimum. The variation in the reference voltage due to this current is:

$$\Delta V_{REF} = \Delta I_C \times \left(15 + Z_{VR2}\right) \qquad (3)$$

$$\Delta V_{REF} = 4\text{mA} \times 45\Omega = 180\text{mV}$$

This will cause an output voltage change of:

$$\Delta V_{OUT} = \Delta V_{REF} \times \frac{N_S}{N_B} \qquad (4)$$

$$\Delta V_{OUT} = 180\text{mV} \times \frac{7}{35} = 36\text{mV}$$

This translates to a variation of 0.7% over the range of duty cycle. This variation in voltage drop with duty cycle, as well as changes to the average value of the bias voltage due to duty cycle changes, gives rise to the variation in output voltage with input line voltage observed in Figure 4.

Output Filter Components

In each of the circuits shown in this Design Note, the output voltage is filtered with a π-network consisting of C2, C3, and L1. C2 is chosen for low ESR and high ripple current capability. The RMS output ripple current carried by C2 is approximately equal to the DC output current in amperes, and the capacitor should be chosen accordingly. A second stage filter is necessary to reduce the 100 kHz and high frequency ripple and noise to 50 mV. Without the second stage of the filter, the fundamental switching component exceeds 100 mV at maximum load, with high frequency spike noise of several hundred millivolts. The inductor used for L1 is not critical. Any inductor of 0.5 to 2 µH with sufficient current capability will work. In the original prototype, both a small ferrite bead on a wire lead with an inductance of 1 µH, and 0.5 µH air-core inductor were tried with virtually identical results. For higher output powers, the ferrite bead would be unsuitable due to saturation.

An electrolytic capacitor is the best choice for C3. This choice is dictated both by cost and damping considerations. The original circuit utilized a 0.1 µF ceramic capacitor in the second-stage filter. This choice turned out to be unsuitable due to the high Q of the ceramic capacitor, which resulted in high frequency, lightly damped ringing at the output of the supply. Electrolytic capacitors have sufficient ESR to damp the resonance with the filter inductor.

The value of filter capacitors will have an effect on the stability of the supply, as they roll off the gain of the bias voltage control loop. Insufficient output capacitance will cause the bias regulation loop to oscillate. Stability of the supply can be checked by step loading the bias supply with a pulsed load of around 5 mA and checking the response at the bias filter capacitor (C5 in Figure 1 and C4 in Figure 2).

NOTES

WORLD HEADQUARTERS
Power Integrations, Inc.
477 N. Mathilda Avenue
Sunnyvale, CA 94086
USA
Main: 408•523•9200
Customer Service:
Phone: 408•523•9265
Fax: 408•523•9365

AMERICAS
For Your Nearest Sales/Rep Office
Please Contact Customer Service
Phone: 408•523•9265
Fax: 408•523•9365

EUROPE & AFRICA
Power Integrations (Europe) Ltd.
Mountbatten House
Fairacres
Windsor SL4 4LE
United Kingdom
Phone: 44•(0)•1753•622•208
Fax: 44•(0)•1753•622•209

JAPAN
Power Integrations, K.K.
Keihin-Tatemono 1st Bldg.
12-20 Shin-Yokohama 2-Chome, Kohoku-ku
Yokohama-shi, Kanagawa 222 Japan
Phone: 81•(0)•45•471•1021
Fax: 81•(0)•45•471•3717

ASIA & OCEANIA
For Your Nearest Sales/Rep Office
Please Contact Customer Service
Phone: 408•523•9265
Fax: 408•523•9365

APPLICATIONS HOTLINE
World Wide 408•523•9260

APPLICATIONS FAX
Americas 408•523•9361
Europe/Africa
44•(0)•1753•622•209
Japan 81•(0)•45•471•3717
Asia/Oceania 408•523•9364