

Introduction

This application note for the MIC38C43 provides a design tool for the project engineer. This note also provides a template for future projects: the user need only scale the appropriate sections. Designing an off-line discontinuous flyback power supply for the first time can become overwhelming. The task involves many disciplines of electrical engineering—magnetics, filter, feedback, layout, and thermal issues to name a few. The design procedure can be broken down into two parts. The reader starts with (1) an in-depth, block-by-block theory of operation and follows with (2) a step-by-step practical example. This example includes specifying the major components, including the magnetics. The schematic, PCB layout, electrical waveforms, and a bill of materials are also included.

The **MIC38C4x** family of BiCMOS current-mode PWM controllers represents another technological advance in MICREL's switch-mode power supply line of ICs. Among power supply designers, the bipolar 384x family is probably the best known of the current-mode PWM controllers currently in use today. However, instead of using the standard bipolar technology used by our competitors, the MIC38C4x ICs use a BiCMOS process. Comparing the MIC38C4x to its bipolar competitors, the start-up current has been reduced from 500 μ A to 50 μ A, and operating current is 4mA instead of 11mA. This allows

the power supply designer to increase efficiency, reduce board space, and reduce the start-up resistors from 1W to 1/4W in most wide input applications. The MIC38C4x family has many other value-added features such as 500kHz switching, -40°C to $+85^{\circ}\text{C}$ operation, 40ns rise and 30ns fall times, and pin-for-pin compatibility with the bipolar UC3842/43/44/(A) and also the UC2842/3/4. Factory trimmed oscillator discharge current and bandgap reference (1%) simplifies design and builds confidence for the designer where high volume manufacturing demands repeatability and ultralow standard deviations in these tolerance-critical areas. The output section of the MIC38HC4x version has been beefed-up with a 1A peak current capability, eliminating the need for a gate-drive transformer in many applications.

These are applications such as VCRs, power tools, laptop and notebook computers, digital cameras, appliances, battery charges, video monitors, and dc-dc converters.

This MIC38C43 reference design is an example of an ultra-low-cost, off-line, isolated, discontinuous-mode, flyback power supply using the Micrel MIC38C43BN current-mode PWM controller IC. This design example features a universal input (85Vac to 265Vac) in a scalable, manufacturable design. Figure 2 is the circuit diagram.

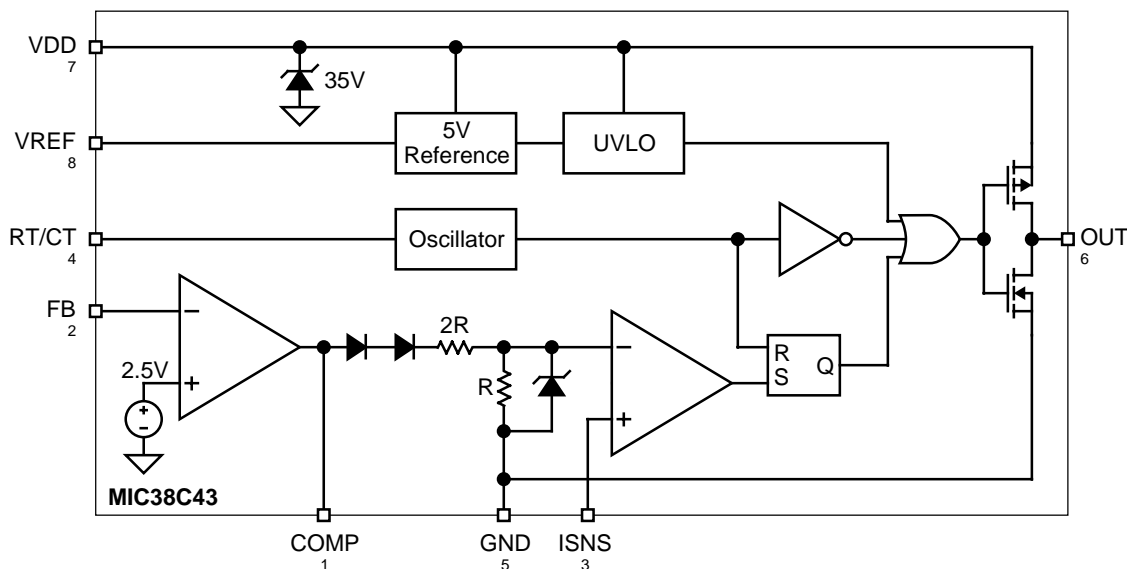


Figure 1. MIC38C43 Functional Diagram

Discontinuous Flyback Supply's Building Blocks

The off-line flyback power supply can be broken down into building blocks to make it easier to understand and design.

The following is a block-by-block description of the Micrel MIC38C43 reference design.

MIC38C43 Controller Block (Figure 1)

The MIC38C43 uses current-mode control to adjust the PWM waveform duty cycle that determines the output voltage. The P-channel and N-channel MOSFETs (output inverter) are alternately switched at a constant frequency; only the duty cycle (on time) is adjusted.

A switching cycle begins when the oscillator generates a reset pulse. This pulse resets the RS latch and turns on Q1 (the power supply main switch) via the logic formed by the OR gate and inverter output. During this time, the inductor current (T1 primary) is increasing and stores energy in the inductor. The PWM comparator measures the voltage at I_{SNS} (current sense pin) and compares it to the output of the error amplifier. When the current-sense waveform intersects the error amplifier output waveform, the PWM comparator will set the RS latch, which turns off Q1 via the logic and inverter. Energy is then discharged from the inductor until the next switching cycle begins. By varying the P-channel MOSFET on-time (duty cycle), the inductor current is adjusted to maintain the output voltage.

Input Block (Figure 2)

The input stage of the supply consists of a common-mode choke, an X-type (line-to-line) safety film capacitor, two Y-type safety ceramic capacitors, one (line-to-ground) and one (neutral-to-ground). The input stage also contains a slow-blow fuse, bulk capacitor, and a diode bridge.

A slow-blow type fuse is used to avoid potential nuisance blowing of the fuse. This is due to the instantaneous current drawn by the bulk capacitor at initial turn-on. The slow-blow fuse also allows for a lower, safer current rating.

Noise/EMI (common-mode currents) normally reflected back onto the ac line are attenuated by L1 and Y-type capacitors C2 and C3. The Y-type capacitors are sized as large as possible to reduce the common-mode noise and still comply with UL/CSA/VDE leakage current requirements. This is why off-line supplies that must pass safety agency testing usually use either 4.7nF or 2.2nF for 120Vac or 230Vac operation, respectively. X-type capacitor C1 and the leakage inductance of L1 attenuate the differential-mode EMI.

Note: X-type and Y-type agency approved safety capacitors must be used in these locations if agency approval is a requirement.

The fixed impedance of L1 (common-mode choke) provides sufficient inrush protection for this cost-conscious design. Other devices, such as NTCs and thermistors, are available if the instantaneous input current becomes an issue.

Output Block (Figure 2)

On the primary side, when the load increases, the control loop increases the duty cycle. On the secondary side Schottky

rectifier D4 produces a dc voltage. Energy is then stored in output capacitor C11. C11 is sized to withstand the ripple current and peak voltage. A low-pass filter formed by L2 and C12 attenuates the high-frequency harmonics to an acceptable level.

Feedback Blocks (FB1 and FB2)

Feedback for this design is current mode. This means there are two feedback loops: one that monitors the output voltage, and a faster loop that monitors the current through the inductor. Whether it is a forward, buck, boost or a flyback topology, the current-mode controller will require monitoring of both loops for standard operation.

Historically, in contrast to current-mode control, the older voltage-mode control monitors only the output voltage and current. This secondary-control scheme causes propagation delays from the supply output to the controller feedback pin. These delays cause second-order effects that are not acceptable, especially in today's high frequency switching supplies that operate at ever higher switching frequencies. The advantages of current-mode control over voltage-mode control are improved line and load regulation, a simpler compensation network, and better overall transient and load response.

Inner Feedback Loop (Figure 2, Feedback Loop 1)

The inner loop (FB1) consists of U1, T1 primary, Q1, R1, R8, and C7. When Q1 turns on, the current in the primary ramps up to some peak current, based on the line and load conditions at that time. Next, current flows through the T1 primary, Q1, R8 and finally to ground. The current flowing through sense resistor R8 develops a voltage drop proportional to the on-time of Q1. This signal is then fed back to the I_{SNS} pin of U1. The R1 and C7 form a low pass filter used to attenuate turn-on spikes and any noise that might corrupt the I_{SNS} signal. Since this is a current-mode controller, current is sensed on a cycle-by-cycle basis, allowing for tighter line and load regulation. When the supply goes into current limit due to a fault condition, output power is limited by folding back the current to less than its maximum operating point. Adjusting R8 sets the current limit. If the current limit is too high, increasing this resistor value will decrease the current limit and vice versa.

Outer Feedback Loop (Figure 2 Feedback Loop 2)

The output voltage is regulated using a TL431 (U3) shunt regulator that senses any perturbation in output voltage via the R10-R11 voltage divider. The voltage divider is configured for a divided down voltage of 2.5V for an output voltage of 5.0V. The divider output voltage is compared to the internal 2.5V reference of the TL431 that sinks the LED cathode of an optoisolator. R12 sets overall dc gain and keeps the LED current within its safe operating area. The TL431 biases the cathode of the optoisolator's diode more negative by pulling it closer to ground, and lets up when the output voltage comes into regulation. The optoisolator transistor is biased within its linear region causing a voltage drop across R6, which is seen by the controller's error amp input. This is compared to the internally generated 2.5V reference. Components C13, C14, and R15, via the CTR (current transfer ratio) of the optoisolator set the overall control-loop frequency response.

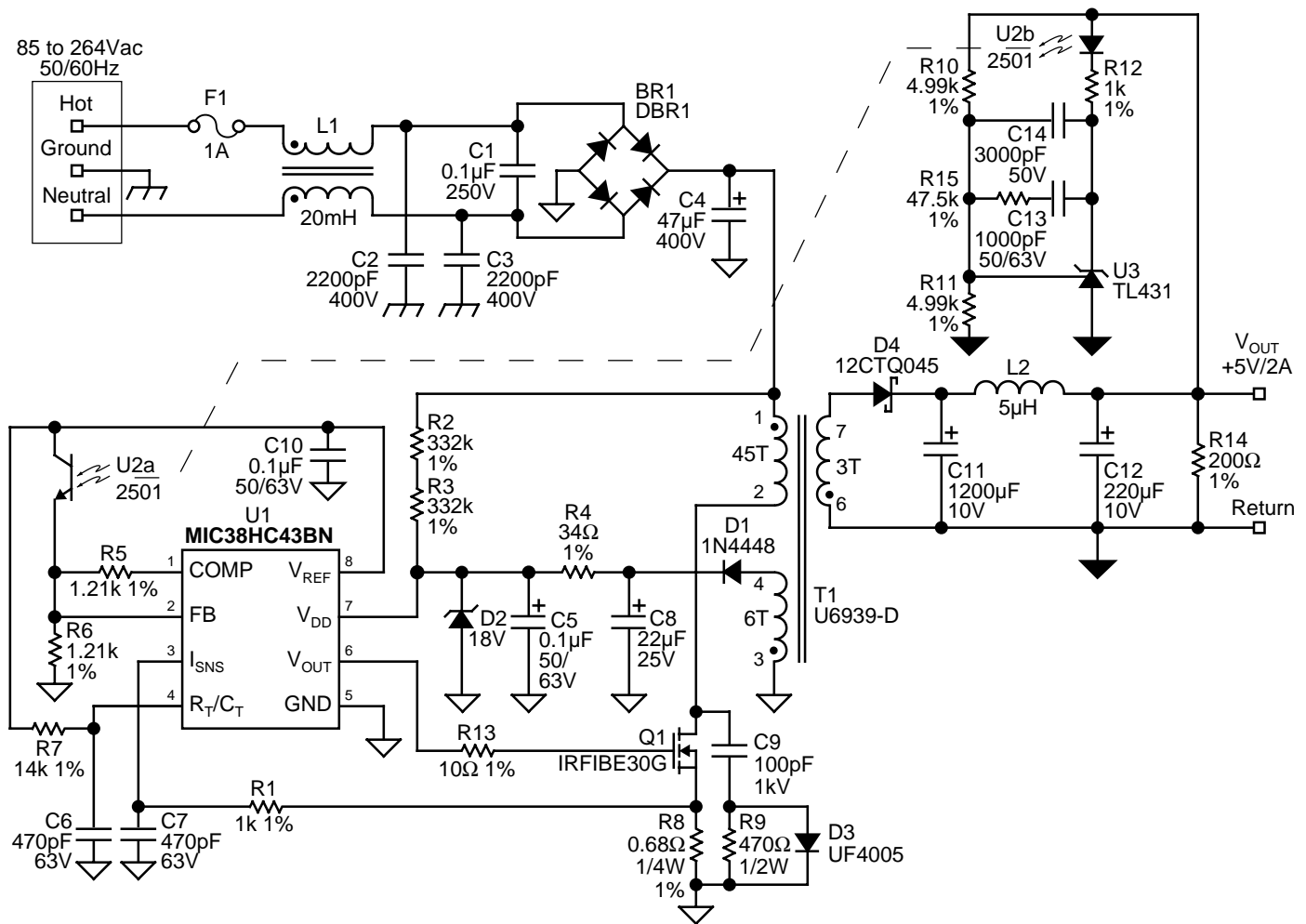


Figure 2. Universal-Input Discontinuous Flyback Supply

Universal-Input Discontinuous Flyback

Design Example

Input Specification

- (1) $V_{IN(min)} = 85V_{ac}$, $V_{IN(max)} = 265V_{ac}$
- (2) $f_{LINE(min)} = 50Hz$, $f_{LINE(max)} = 60Hz$
 $f_{SW} = 200kHz$

Output Specification

- (3) $V_{OUT} = 5.0 \pm 1\%$
- (4) $I_{OUT(min)} = 0.2A$, $I_{OUT(max)} = 2A$
- (5) Load regulation = $\pm 0.2\%$
- (6) Line regulation = $\pm 0.2\%$
- (7) $V_{OUT(ripple)} = 10mV$

Voltage Reference

Given the need for tight load and line regulation, select the TL431 precision voltage reference for the feedback loop. The 38C4x reference provides constant voltage for the R_T/C_T oscillator.

Calculate the peak voltage at $V_{IN(min)}$ and $V_{IN(max)}$.

Determine the peak voltage at $V_{IN(min)}$

$$(8) \quad V_{IN(peak)} = V_{IN} \times \sqrt{2}$$

$$V_{IN(peak)} = 85V \times \sqrt{2}$$

$$(8a) \quad V_{IN(peak)} = 120.21V$$

For the average voltage, subtract half of the ripple voltage and the forward drop of the full bridge rectifier:

$$(9) \quad V_{DC(low)} = V_{IN(min)} - \frac{V_{RIPPLE}}{2} - V_{F(diode)}$$

where:

$$V_{RIPPLE} = 15\% V_{IN(peak)}$$

$$V_{F(diode)} = 1V$$

then:

$$V_{DC(low)} = 120V - \frac{120V \times 0.15}{2} - 1V$$

round to 100V

$$(9a) \quad V_{DC(low)} = 100V$$

Calculate the DC Rail at High Line AC

Determine the peak voltage at $V_{IN(max)}$. Using formula (8):

$$(10) \quad V_{IN(peak)} = V_{IN(max)} \sqrt{2}$$

$$V_{IN(peak)} = 265V \sqrt{2}$$

$$(10a) V_{IN(peak)} = 374.71V$$

For the average voltage, subtract half of the ripple voltage and the forward drop of the full bridge rectifier, using Equation (11):

$$(11) V_{DC(high)} = 375V - \frac{375 \times 0.015}{2} - 1$$

Round to 345V.

$$(11a) V_{DC(high)} = 345V$$

Input/Bulk Capacitor

The input capacitor is used to hold up the rectified line voltage in an off-line application. The size of this capacitor is based on ripple current, the peak voltage across the capacitor and the energy storage requirements for a particular application.

It is a good design rule to keep the ripple voltage on the input capacitor between 10% and 15% when operating under worst case conditions (low-line, full-load). This means that at the minimum line input voltage with a full load on the supply's output, the peak-to-peak voltage ripple will make up only 15% of the peak voltage seen across the capacitor. The V_{P-P} ripple is a function of the capacitors ESR and the ripple current. If the capacitor is too small, the ripple component will become too large. This causes the main switch to see a lower average voltage and results in lower overall system reliability and efficiency.

Calculate the necessary capacitance in microfarads. Assume 80% efficiency.

$$E_{IN} = \frac{P_{IN}}{2 f_{MIN}}$$

$$f_{MIN} = 50Hz$$

$$P_O = 10W$$

$$P_{IN} = 10W \times 1.2 = 12W$$

Then:

$$E_{IN} = \frac{12}{2 \times 50} = 0.12 \text{ joules}$$

$$C = \frac{2 E_{IN}}{V_{PK}^2 - V_{MIN(av)}^2}$$

$$C = \frac{2 \times 0.12J}{120^2 - 100^2} = 54.54\mu F$$

Select the closest standard value, a 47 μF , 400V, 105°C aluminum electrolytic for the C4 bulk capacitor.

Designing the Flyback Transformer

A TDK PC40 EE22-Z core and a TDK BE-22-118CP vertical bobbin were selected for this design.

The maximum recommended on-time is 48% of the full period. This keeps the regulator operating in discontinuous mode under worst case conditions. The on-time is used in Equation 2 to calculate the maximum primary inductance.

Note: By using a maximum on-time of 48% at 200kHz we have allowed for less rise and fall time, however, this is outweighed by the fact that the core volume goes down by half for every factor of four we increase operating frequency.

$$t_{ON} = D_{max} \frac{1}{f_{OP}}$$

then:

$$t_{ON} = 2.4 \times 10^{-6}s$$

where:

$$D_{max} = 0.48$$

$$f_{OP} = 200kHz$$

$$t = \frac{1}{f_{OP}} = 5 \times 10^{-6}$$

The maximum primary inductance for discontinuous operation under worst case conditions (low line voltage, full load) is:

$$L_{PRI} = \frac{(V_{IN(min)} t_{ON})^2}{2.5 t P_{OUT}} 0.9$$

then:

$$L_{PRI} = 414.72\mu H$$

where:

$$V_{IN(min)} = 100V$$

$$t_{ON} = 2.4 \times 10^{-6}$$

$$P_{OUT} = 10W$$

Peak primary current is used to determine the center gap for the core and the number of primary turns.

$$I_{PEAK} = \frac{(V_{IN(min)} t_{ON})}{L_{PRI}}$$

Use 100V for some margin at the ac low line.

where:

$$V_{IN} = 100V$$

$$t_{ON} = 2.4 \times 10^{-6}$$

$$L_{PRI} = 414\mu H$$

then:

$$I_{PEAK} = 0.579A$$

Center Gap Formula (Figure 4a)

$$l_{GAP} = \frac{0.4 \pi L_{PRI} I_{PK}^2 \times 10^8}{A_e (B_{MAX})^2}$$

At this time it is important to note that a flyback transformer operates in the first quadrant and is considered unipolar, unlike a push-pull topology that operates in two quadrants. This subject is greater than the scope of this paper and should be researched before designing your magnetics. For further information please consult your magnetics supplier's data book.

where:

$$\Delta B_{\text{MAX}} = 1300 \text{ (gauss)}$$

Conversion note: 1300 gauss = 130 millitesla

The core area is found in Figure 4a under A_e .

$$A_e = 0.41 \text{ cm}^2$$

then:

$$l_{\text{GAP}} = 0.0252 \text{ cm} = 9.917 \times 10^{-3} \text{ inches}$$

Number of Primary Turns

$$N_{\text{PRI}} = \frac{B_{\text{max}} l_{\text{GAP}}}{0.4 \pi I_{\text{PEAK}}}$$

then:

$$N_{\text{PRI}} = 45 \text{ turns}$$

Number of Secondary Turns

$$N_{\text{SEC}} = \frac{N_{\text{PRI}} (V_{\text{OUT}} + V_{\text{DIODE}}) (1 - D_{\text{max}})}{(V_{\text{IN (min)}}) D_{\text{max}}}$$

where:

$$V_{\text{DIODE}} = 0.6 \text{ V}$$

$$V_{\text{OUT}} = 5 \text{ V}$$

$$V_{\text{CC}} = 11 \text{ V}$$

then:

$$N_{\text{SEC}} = 3 \text{ turns}$$

Number of Tertiary Turns

Once the controller has gone through its start-up, the tertiary winding powers the MIC38HC43 controller.

$$N_{\text{BIAS}} = \frac{N_{\text{PRI}} (V_{\text{CC}} + V_{\text{DIODE}}) (1 - D_{\text{max}})}{(V_{\text{IN (min)}}) D_{\text{max}}}$$

then:

$$N_{\text{BIAS}} = 6 \text{ turns}$$

Use the primary wire size to minimize the number of different wire sizes and reduce cost.

Primary Wire Size

All wire needs to be at least 300 circular mils per ampere (CMA). The area of 30AWG is 100 circular mils.

$$\text{CMA}_{\text{PRI}} = \frac{\text{CM}_{\text{PRI}}}{\text{PRI}_{\text{RMS}}}$$

where:

$$\text{CM}_{\text{PRI}} = 100 \text{ CM}$$

$$\text{PRI}_{\text{RMS}} = 0.22$$

The diameter of 30AWG single-build nylese is 0.0112 inches.

then:

$$\text{CMA}_{\text{PRI}} = \frac{100 \text{ CM}}{0.22 \text{ A}} = 454 \text{ CM/A [OK]}$$

Secondary Wire Size

The secondary also needs to be at least 300CMA.

$$\text{CMA}_{\text{SEC}} = \frac{\text{CM}_{\text{SEC}}}{\text{SEC}_{\text{RMS}}}$$

where:

$$\text{secondary diameter} = 0.0228 \text{ inch}$$

$$I_{\text{PEAK(sec)}} = 7.76 \text{ A}$$

$$\text{SEC}_{\text{RMS}} = 3.14 \text{ A}_{\text{RMS}}$$

$$\text{CM}_{\text{SEC}} = 1020 \text{ CM}$$

The secondary is bifilar 23AWG which is 0.0226 inch diameter and is $510 \text{ CM} \times 2 = 1020$ circular mils.

Therefore;

$$\text{CMA}_{\text{SEC}} = 1020 \text{ CM} / 3.14 \text{ A} = 325 \text{ CM/A [OK]}$$

Primary Turns Per Layer

The number of primary turns/layer is based on wire size and the bobbin width (BW). This is the "C" dimension in the TDK catalog. See Figure 4d.

$$\text{PRI}_{\text{Tperlayer (max)}} = \frac{\text{BW}}{\text{PRIAWG}_{\text{DIA}}}$$

where:

$$\text{BW} = 0.332 \text{ inch}$$

$$\text{PRIAWG}_{\text{DIA}} = 0.0112 \text{ inch}$$

Single-build 30AWG is 0.0112 inches diameter.

then:

$$\text{PRI}_{\text{T/Layer(max)}} = 29.643 \text{ T/layer} < 45 \text{ turns [not OK]}$$

So split the primary in to two layers, (1) 22 turns and (2) 23 turns.

Secondary Turns Per Layer

The secondary uses triple insulation to meet safety agency approvals. To keep the magnetics compact and efficient we have gone to a bifilar secondary approach.

$$\text{SEC}_{\text{MAX(dia)}} = 0.035 \text{ inch with triple insulation}$$

$$2 \times \text{SEC}_{\text{MAX(dia)}} = 0.070 \text{ inch}$$

$$\text{SEC}_{\text{T/LAYER (max)}} = \frac{\text{BW}}{\text{SEC}_{\text{DIA (max)}}}$$

then:

$$\text{SEC}_{\text{T/LAYER(max)}} = 4.74 \text{ T/Layer [OK < 3T } N_{\text{SEC}}]$$

Winding Instructions

At this point we should have two primary winding layers wound and separated by mylar tape. Tape over the second primary winding and add the 6-turn tertiary winding. Then, tape over the tertiary winding and add the 3-turn secondary.

Core Loss

From page 9 of the '96 TDK catalog, the core loss is 0.07 mW/cm^3 at 200kHz.

Gauss level = 1300 gauss, then: $1300/10 = 130 \text{ milliteslas}$

TDK uses sine wave data for its core loss charts. The flyback topology operates in one quadrant and is considered a unipolar device.

therefore:

$$\frac{130\text{mT}}{2} = 65\text{mT}$$

$$C_{\text{LOSS}} = 0.07 \text{ (watt/cm}^3\text{)} \\ \text{(see Figure 5)}$$

$$CV = 1.61\text{cm}^3 \\ \text{(see Figure 4a)}$$

$$P_{\text{LOSS(core)}} = C_{\text{LOSS}} CV$$

then:

$$P_{\text{LOSS(core)}} = 0.113\text{W}$$

Bobbin Stack-Up

This is the bobbin stack up of windings and tape versus bobbin window height for a TDK EE22-PC40 core. Designing magnetics is an iterative process. If the wire does not fit, then you will have to go to a larger core and recalculate. (See Figure 4b and 4c for bobbin dimensions.)

$$\text{BobH}_{\text{MIN}} = \\ \left[(\text{PRIAWG}_{\text{DIA}} \text{ PL}) + (\text{SEC}_{\text{DIA (max)}} \text{ SL}) \right] + \\ (\text{BIAS}_{\text{DIA}} \text{ BL}) + (\text{TL } T_{\text{TH}})$$

where:

$$A = 0.492 \text{ inch}$$

$$B = 0.311 \text{ inch}$$

$$\text{BobH}_{\text{ACTUAL}} = A - B/2$$

$$\text{PRIAWG}_{\text{DIA}} = 0.0112 \text{ inch}$$

$$\text{BobH}_{\text{ACTUAL}} = 0.0905 \text{ inch}$$

$$\text{PL} = \text{primary layer}$$

$$\text{PL} = 2$$

$$\text{SL} = \text{secondary layer}$$

$$\text{SL} = 1$$

$$\text{SEC}_{\text{MAX(dia)}} = 0.035 \text{ inch}$$

$$\text{TL} = \text{tape layers}$$

$$\text{TL} = 4$$

$$T_{\text{TH}} = \text{tape thickness} = 0.001 \text{ inch}$$

$$\text{BIAS}_{\text{DIA}} = 0.0112 \text{ inch}$$

$$\text{BL} = 1$$

then:

$\text{BobH}_{\text{ACTUAL}}$ is the bobbin's maximum window winding area. The BobH_{MIN} is the actual stack-up of windings and insulation, therefore the $\text{BobH}_{\text{ACTUAL}}$ must be greater than BobH_{MIN} .

$$\text{BobH}_{\text{ACTUAL}} = 0.0905 \text{ inch}$$

$$\text{BobH}_{\text{MIN}} = 0.0724 \text{ inch [OK]}$$

Input Diode Bridge

When selecting a diode bridge for an off-line supply, the peak current and peak reverse voltage seen by the bridge must be calculated. The peak current must be calculated at the minimum input voltage of the off-line supply, with full load applied to the output. Assume a conservative power factor of 0.6 and efficiency of 80%. We know from calculating the magnetics that primary I_{PK} is:

$$I_{\text{PK}} = 0.579A_{\text{PK}}$$

Use a 2× margin.

Therefore:

$$I_{\text{DIODE(min)}} = 2 (0.579A_{\text{PK}})$$

$$I_{\text{DIODE(min)}} = 1.12A_{\text{PK}}$$

The reverse voltage rating of the bridge must be calculated at the high line-input supply voltage. The load is not important for this calculation.

$$V_{\text{PK(rev)}} = 1.414 \times V_{\text{IN(max)}}$$

$$V_{\text{PK(rev)}} = 1.414 \times 265$$

$$V_{\text{PK(rev)}} = 375\text{V}$$

Select a 400V, 1.5A molded diode bridge.

Output Rectifier

The output diode in a flyback supply must be chosen carefully to optimize the power system's total efficiency. There are two ways in which a rectifier consumes power: forward conduction loss and reverse recovery loss.

Voltage Rating

To select the correct rectifier for a specific application, the four main considerations are forward voltage drop (V_F), reverse recovery time, maximum reverse peak working voltage (V_{RWM}), and maximum forward current (I_F). The maximum reverse voltage is across the rectifier when the supply is at high line voltage, therefore, the V_{RWM} calculation is:

$$V_{\text{RWM}} = V_O + \left(V_{\text{IN(peak)}} \frac{N_{\text{SEC}}}{N_{\text{PRI}}} \right)$$

$$V_{\text{RWM}} = 5\text{V} + \left(375 \times \frac{3}{45} \right) = 30\text{V}$$

The next standard value with some margin is 40V.

Current Rating

The calculation for the maximum forward peak current is done at $V_{\text{IN(min)}}$ and, with the supply at full load.

$$I_{\text{PK}} = V_{\text{DC(low)}} \left(\frac{D_{\text{max}}}{L_{\text{PRI}} \times f_{\text{OP}}} \right)$$

$$\text{Low line dc rail} = 100\text{Vdc}$$

$$I_{\text{PK}} = 100\text{V} \left(\frac{48}{414 \times 10^{-6} \times 200 \times 10^3} \right) = 0.579\text{A}$$

The peak current the rectifier will see is approximately

$$I_{PK(pri)} \times n = 0.579 \times 15 = 8.69A_{PK}$$

Therefore, for this application, we will use a 10A Schottky center tap diode with a minimum V_R of 40V to allow for some margin. The diode must withstand the rms current and the peak current. The use of a TO-220 package rectifier with a center tap allows the diodes to be paralleled and heat sunked for improved reliability. Another design rule to approximate the current rating for the output diode in a discontinuous design is:

$$I_D = 4 \times I_{OUT}$$

For a continuous design, the rule is:

$$I_D = 3 \times I_{OUT}$$

As in any design, there are trade-offs to be made, such as choosing discontinuous or continuous mode. This supply was designed to run in the discontinuous mode to simplify the feedback control system and to reduce reverse recovery losses in the rectifier. If this were a continuous design, the rectifier would be forced to recover while there was still substantial current flowing through it.

Output Hold-Up Capacitor

When reviewing Marcon Capacitor's '96 catalog, notice that at the end of each section there is derating information based on temperature and frequency (see Figure 3b). At 45°C and 100kHz the catalog shows that we can multiply the 105°C ripple current rating for this series of Marcon capacitor (CEE) by 2.4 for a ambient of 45°C, worst case. Note even though the catalog's derating graph only goes to 100kHz, clearly the trend shows that we can multiply by at least 2.4 for a ambient of 45°C.

Calculate ripple current rating:

$$I_{SEC(rms)} = I_{PRI(rms)} \times n$$

where:

$$I_{PRI(rms)} = 0.19A_{RMS}$$

$$n = \frac{N_{PRI}}{N_{PRI}} = \frac{45}{3} = 15$$

then:

$$I_{SEC(rms)} = 0.19 \times 15 = 2.85A_{rms}$$

So we need a capacitor that is rated for 2.85A_{RMS} current at 45°C ambient.

Then:

$$I_{RIPPLE \text{ rating}} = 2.85/2.4 = 1.1875A_{RMS} \text{ at } 105^\circ\text{C} \text{ (see Figure 3a)}$$

The corresponding ESR is 0.29, therefore the V_{RIPPLE} will be:

$$ESR \times I_{SEC(rms)} = V_{RIPPLE}$$

$$0.29 \times 2.85 = 826mV$$

Note: We will use this later for calculating the output filter.

Calculate Voltage Spike on Output Capacitor

$$I_{PEAK(pri)} = 0.579A$$

$$V_{SPK} = N \times I_{PEAK(pri)} \times ESR$$

$$15 \times 0.579 \times 0.29 = 2.52V_{PK}$$

So the output capacitor will see $2.5 + 5.0 = 7.5V$ maximum

Voltage rating = 10V

A CEEFM1A122M6 is a 1200μF, 10V aluminum electrolytic capacitor.

Output Filter

The output filter is a low pass filter used to attenuate the residual noise left on the dc bus after the main hold-up capacitor. The output inductor should be rated for 3A continuous.

$V_{RIPPLE} = 800mV$. We need to reduce to 10mV.

We know our fundamental is 200kHz, so we set the filter to have a break frequency of 5kHz. We also know a single-stage low-pass filter will roll off at 20dB/decade.

We will be attenuating out past 500kHz, therefore, from 5kHz to 50kHz is 1 decade and from 50kHz to 500kHz is two decades.

We have the break frequency set so it cuts in before 5kHz. By the switching frequency's second harmonic of 400kHz, the noise is attenuated by almost -40dB. Assume a capacitor value of 220μF.

$$L = \left(\frac{1}{2 \pi f \sqrt{C}} \right)^2 = 4.6\mu H$$

Let's round this to the next standard value, 5μH.

$$f_{BREAK} = \frac{1}{2 \pi \sqrt{L C}}$$

$$f_{BREAK} = \frac{1}{2 \pi \sqrt{(5 \times 10^{-6}) \times 220 \times 10^{-6}}} = 4.8kHz$$

$$-40dB = 20 \log \frac{800mV}{X} \rightarrow 10^{-2} = \frac{800mV}{X}$$

$$\frac{800mV}{100} = X = 8mV \text{ of peak-to-peak ripple}$$

A minimum output load of 10% is used to prevent, (1) continuous operation, and (2) switching spikes from ratcheting up the voltage on the output capacitor. Remember, if all else fails, use Klinkoff's "variable constant!"

Part No.	Rated Voltage (Vdc)	Capacitance (µF) @120Hz, 20°C	Max. Leakage Current (µA) @20°C	Tan δ	Max. ESR (Ω) @120Hz, 20°C	Max. Impedance @-10°C @20°C	Max. Ripple Current @100kHz @120Hz	Nominal Case Size D × L (mm)
CEEFM1A122M6	10	1,200	360	0.21	0.290	0.126 0.063	1,190 1,083	10 × 25

Figure 3a. Capacitor C11 Specifications

+45°C	+65°C	+85°C	+105°C
2.4	2.19	1.73	1

Figure 3b. Capacitor Maximum Ripple Current at 100kHz

TDK Ferrite Core and Bobbin Information

Part No.	Dimensions mm (inches)						Effective Parameters			
	A	B	C	D	E min.	F	H	C _i (mm ⁻¹)	A _e (mm ²)	l _e (mm)
PC40EE22-Z	22.0 ±0.3 (0.866 ±0.012)	9.35 ±0.15 (0.368 ±0.006)	5.75 ±0.25 (0.226 ±0.010)	5.75 ±0.25 (0.226 ±0.010)	13.0 (0.512)	5.35 ±0.15 (0.211 ±0.006)	4.3 (0.169)	0.970	41.0	39.6
										1610
										2400 ±25% 2180 ±25%

Figure 4a. Transformer T1 Core Dimensions

Part No.	Dimensions mm (inches)		
	A	B	C
BE-22-118CP	12.5 (0.492)	7.9 (0.311)	8.45 (0.332)

Figure 4b. Transformer T1 Bobbin Dimensions

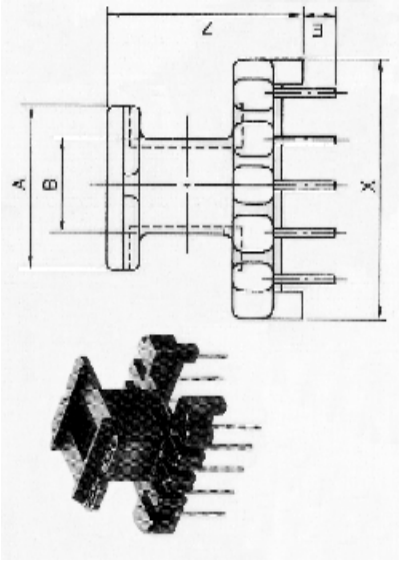


Figure 4c. Transformer T1 Bobbin Dimensions

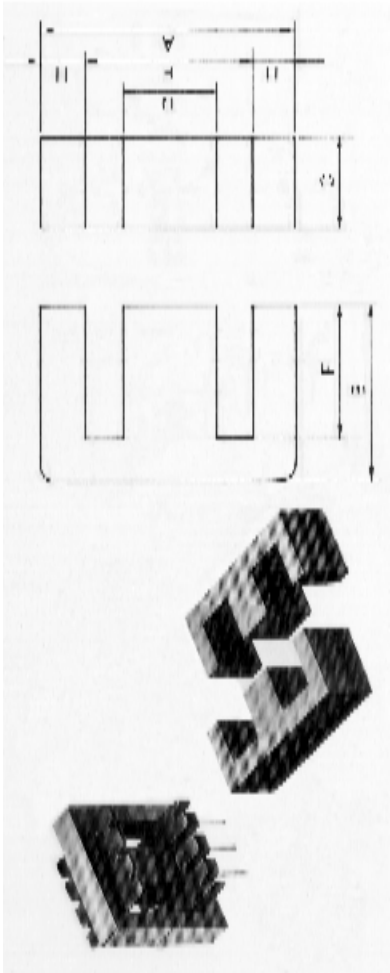


Figure 4d. Transformer T1 Core Dimensions

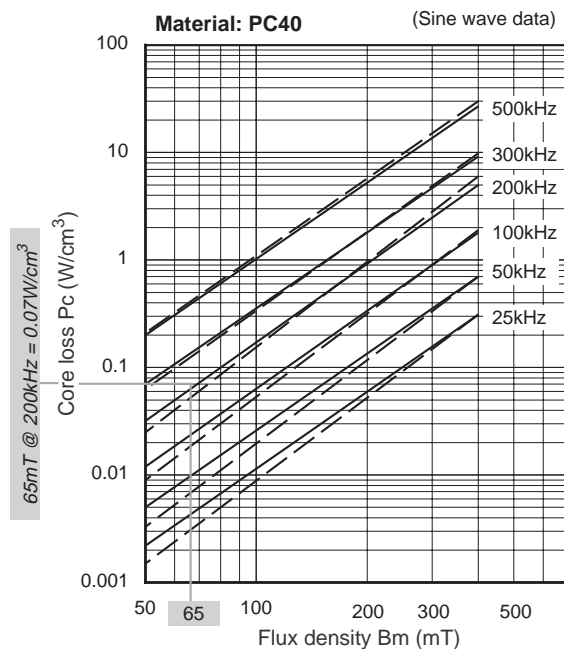


Figure 5. Transformer T1 Core Loss

Transient Load Response

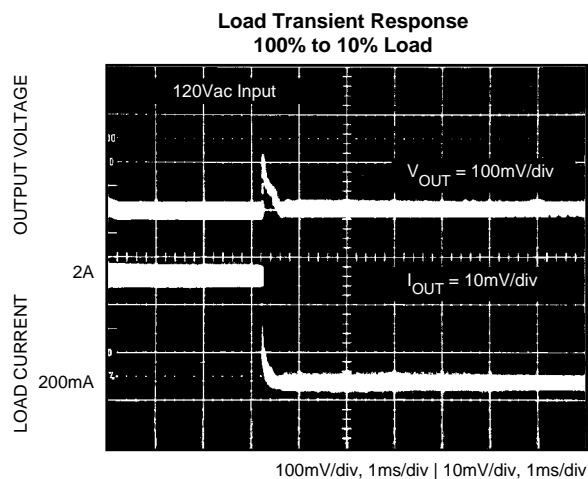


Figure 6a.

Drain Voltage Waveforms

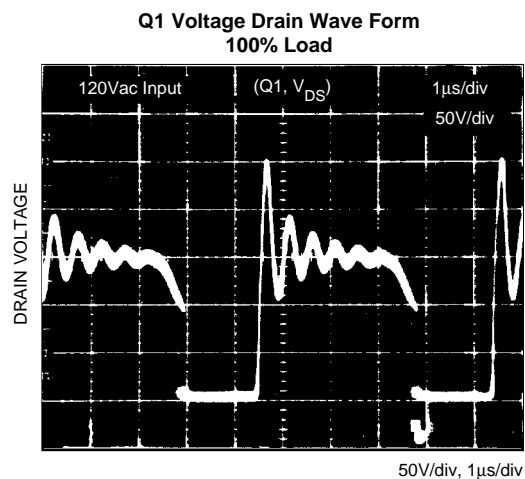


Figure 7a.

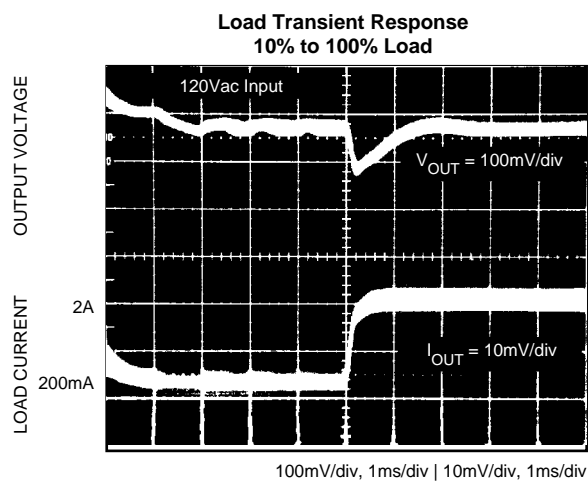


Figure 6b.

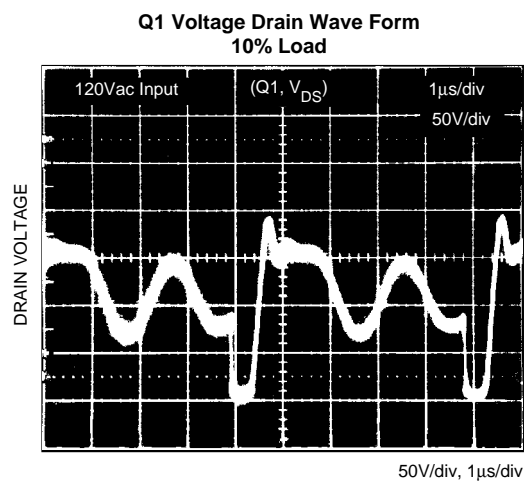


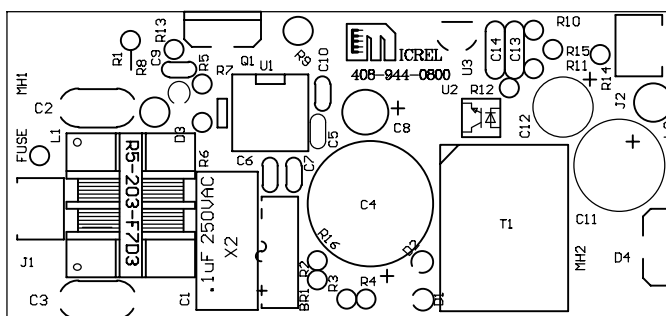
Figure 7b.

Bill of Materials

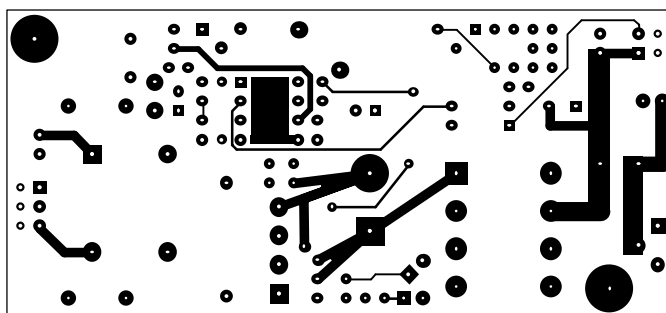
Item	Part Number	Manufacturer	Description	Value	Quantity
C1	CFKC22E104M	Nitsuko	film capacitor	0.1μF, 250Vac, X-type	1
C2, C3	DE7100F222MVA1-KC	Murata	ceramic capacitor	2200pF, 400V, Y-type	2
C4	KMG400VB47018X20L	Nippon Chemicon	aluminum electrolytic	47μF, 400V	1
C5, C10	ECU-S1J104KBB	Panasonic	ceramic capacitor	0.1μF, 63V	2
C6, C7	ECU-S1J471JCA	Panasonic	ceramic NPO/COG capacitor	470pF ±5%, 63V	2
C8	CEBSM1E220M	Marcon	aluminum electrolytic, 5×11	22μF, 25V	1
C9	ECK-D3A101KBN		ceramic capacitor	100pF, 1kV	1
C11	CEEFM1A122M6	Marcon	aluminum electrolytic, 10×25	1200μF, 10V	1
C12	CEEFM1A2211M4	Marcon	aluminum electrolytic, 6.3×11	220μF, 10V	1
C13	ECU-S1H104KBB	Panasonic	NPO/COG ceramic capacitor	1000pF, 63V	1
C14	ECU-S1H332JCB		COG ceramic capacitor	3300pF ±5%, 50V	1
D1	1N4448	Motorola	diode	100V	1
D2	1N4746	Motorola	Zener diode	18V, 5W	1
D3	UF4005	General Instruments	UFR diode	600V, 1A	1
D4	12CTQ045	International Rectifier	diode	45V, 12A	1
DBR1	KBPO4M	General Instrument	molded bridge	400V, 1.5A	1
F1	263 001	Littlefuse	Picofuse	250V, 1A	1
J1	1725669	Phoenix	connector		1
J2	1725656	Phoenix	connector		1
L1	R5-203F7D3	FDK	inductor	20mH, 800mA	1
L2	CTX-5-3-FR	Coiltronics	radial inductor	5μH, 3A	1
Q1	IRFIBE30G	International Rectifier	N-channel MOSFET	800V, 3Ω	1
R1			metal film resistor	1.0k ±1%, 1/4W	1
R2, R3			metal film resistor	332k ±1%, 1/4W	1
R4			metal film resistor	34.0Ω ±1%, 1/4W	1
R5, R6			metal film resistor	1.21k ±1%, 1/4W	1
R7			metal film resistor	14k ±1%, 1/4W	1
R8			metal film resistor	0.68Ω ±1%, 1/4W	1
R9			metal film resistor	470Ω ±5%, 1/2W	1
R10, R11			metal film resistor	4.99k ±1%, 1/4W	1
R12			metal film resistor	1.0k ±1%, 1/4W	1
R13			metal film resistor	10.0k ±1%, 1/4W	1
R14			metal film resistor	200Ωk ±1%, 1/4W	1
R15			metal film resistor	47.5k ±1%, 1/4W	1
T1	U6939-C	Coilcraft	transformer*		1
U1	MIC38C43BN	Micrel	current-mode PWM controller		1
U2	2501	NEC	optoisolator		1
U3	TL431	Motorola	voltage reference		1
—	user dependent		heat sink		1

* tel: (847) 639-6400, contact Frank Concialadi

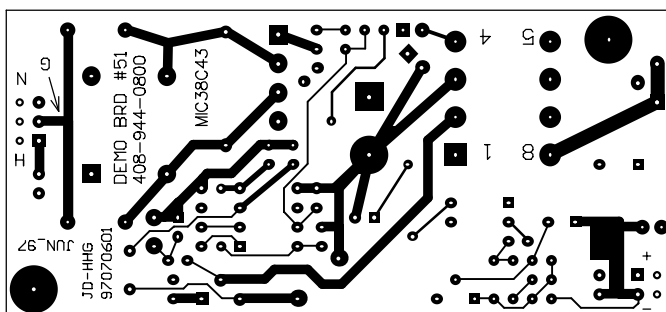
Printed Circuit Layout



Silk Screen



Component Side



Solder Side

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