



UC3844, 45 UC2844, 45

High Performance Current Mode Controllers

The UC3844, UC3845 series are high performance fixed frequency current mode controllers. They are specifically designed for Off-Line and dc-to-dc converter applications offering the designer a cost effective solution with minimal external components. These integrated circuits feature an oscillator, a temperature compensated reference, high gain error amplifier, current sensing comparator, and a high current totem pole output ideally suited for driving a power MOSFET.

Also included are protective features consisting of input and reference undervoltage lockouts each with hysteresis, cycle-by-cycle current limiting, a latch for single pulse metering, and a flip-flop which blanks the output off every other oscillator cycle, allowing output deadtimes to be programmed for 50% to 70%.

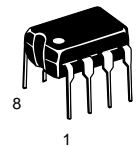
These devices are available in an 8-pin dual-in-line plastic package as well as the 14-pin plastic surface mount (SO-14). The SO-14 package has separate power and ground pins for the totem pole output stage.

The UCX844 has UVLO thresholds of 16 V (on) and 10 V (off), ideally suited for off-line converters. The UCX845 is tailored for lower voltage applications having UVLO thresholds of 8.5 V (on) and 7.6 V (off).

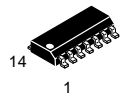
- Current Mode Operation to 500 kHz Output Switching Frequency
- Output Deadtime Adjustable from 50% to 70%
- Automatic Feed Forward Compensation
- Latching PWM for Cycle-By-Cycle Current Limiting
- Internally Trimmed Reference with Undervoltage Lockout
- High Current Totem Pole Output
- Input Undervoltage Lockout with Hysteresis
- Low Startup and Operating Current
- Direct Interface with Motorola SENSEFET Products

HIGH PERFORMANCE CURRENT MODE CONTROLLERS

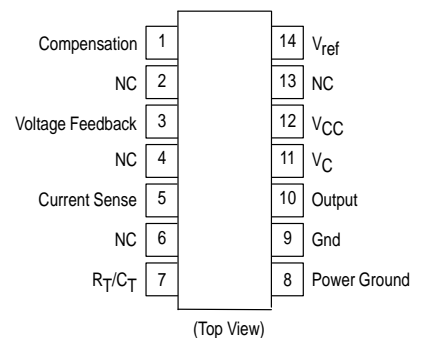
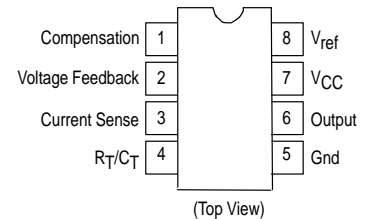
N SUFFIX
PLASTIC PACKAGE
CASE 626



D SUFFIX
PLASTIC PACKAGE
CASE 751A
(SO-14)

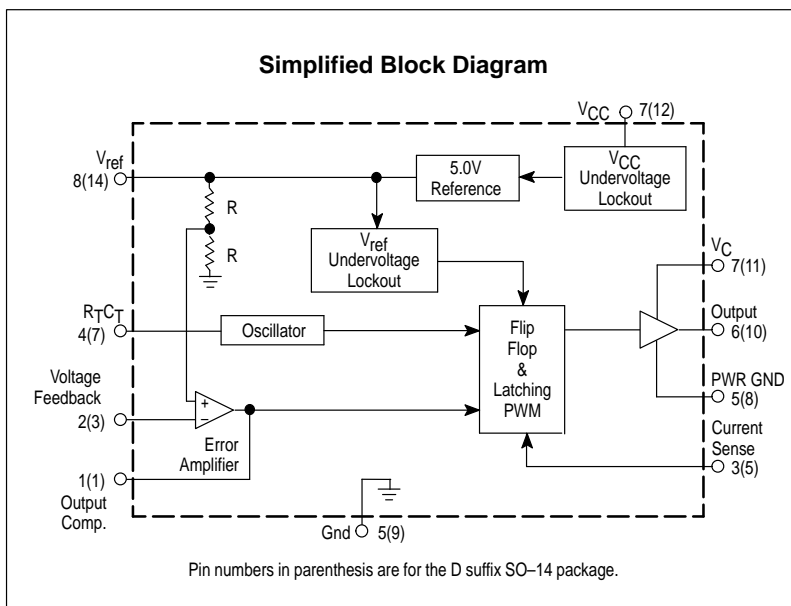


PIN CONNECTIONS



ORDERING INFORMATION

Device	Operating Temperature Range	Package
UC3844D	T _A = 0° to +70°C	SO-14
UC3845D		SO-14
UC3844N		Plastic
UC3845N	T _A = -25° to +85°C	Plastic
UC2844D		SO-14
UC2845D		SO-14
UC2844N		Plastic
UC2845N		Plastic



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Figure 1. Timing Resistor versus Oscillator Frequency

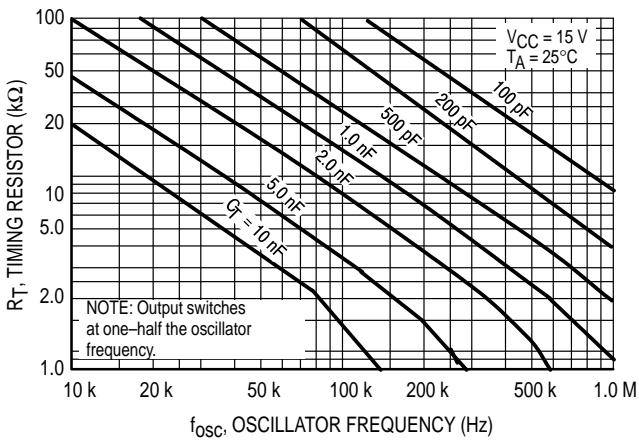


Figure 2. Output Deadtime versus Oscillator Frequency

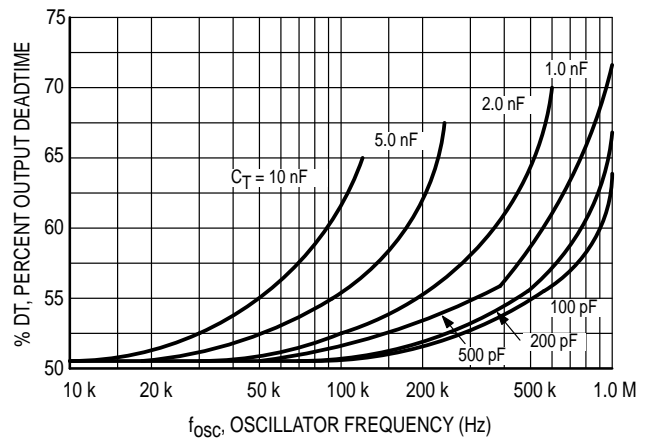


Figure 3. Error Amp Small Signal Transient Response

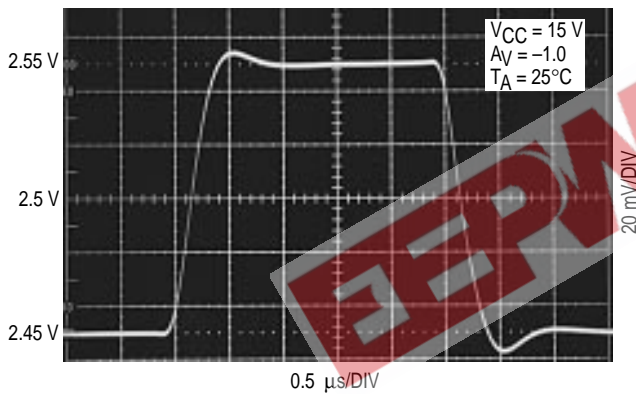


Figure 4. Error Amp Large Signal Transient Response

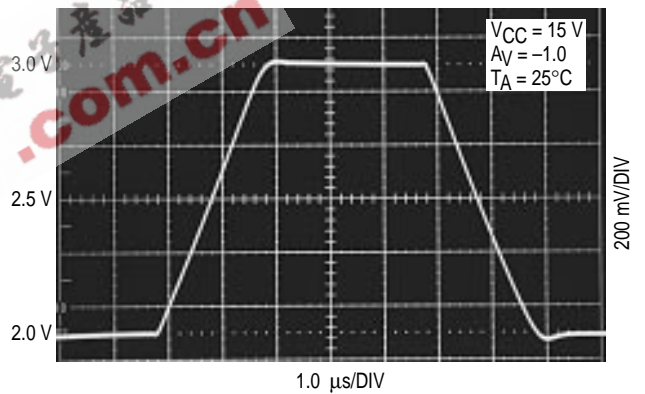


Figure 5. Error Amp Open Loop Gain and Phase versus Frequency

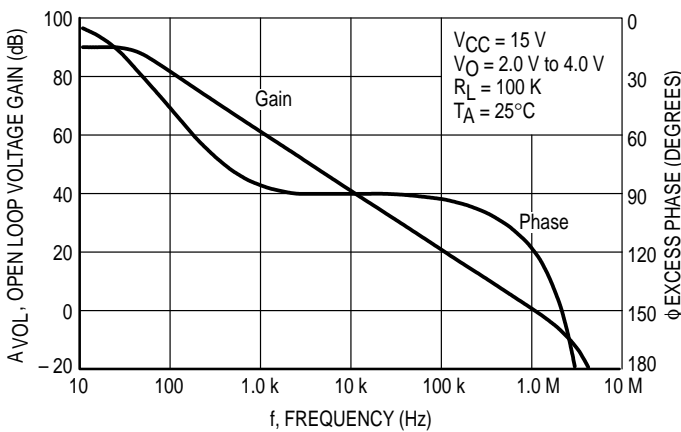
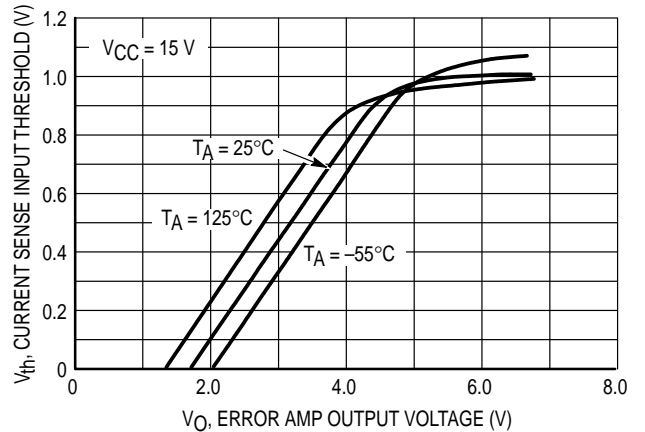


Figure 6. Current Sense Input Threshold versus Error Amp Output Voltage



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Figure 7. Reference Voltage Change versus Source Current

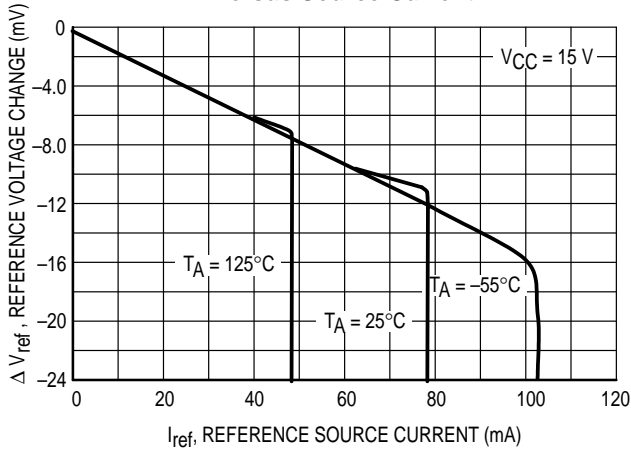


Figure 8. Reference Short Circuit Current versus Temperature

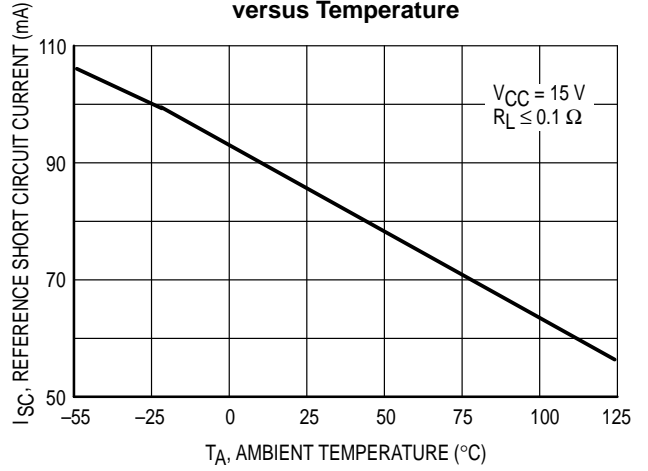


Figure 9. Reference Load Regulation

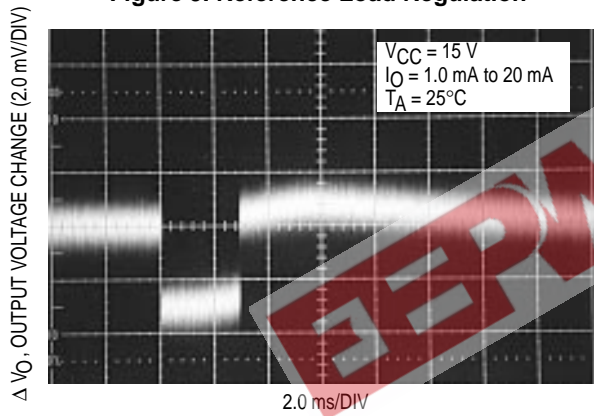


Figure 10. Reference Line Regulation

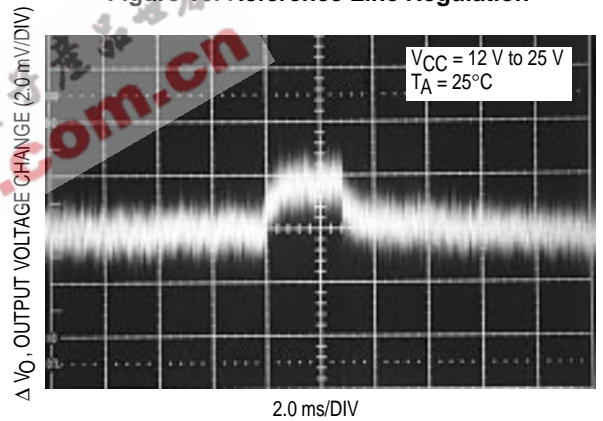


Figure 11. Output Saturation Voltage versus Load Current

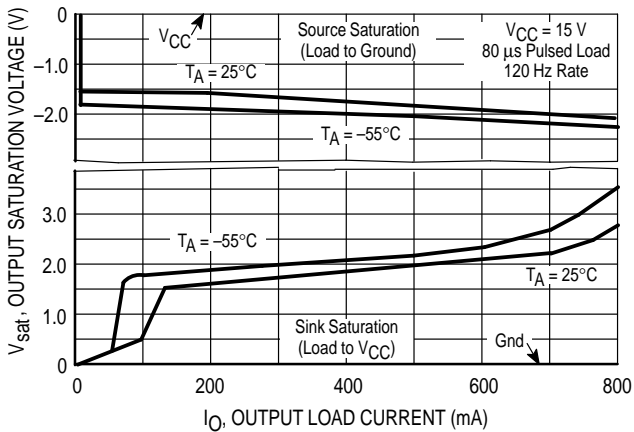
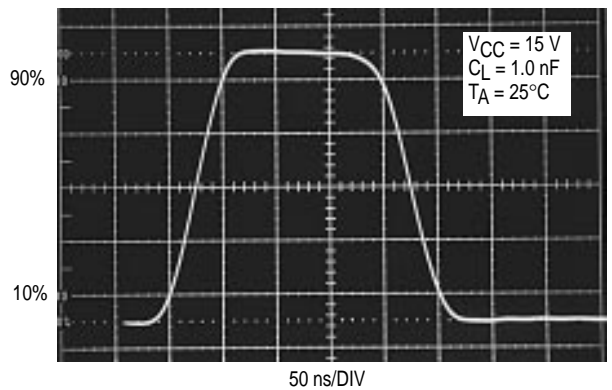
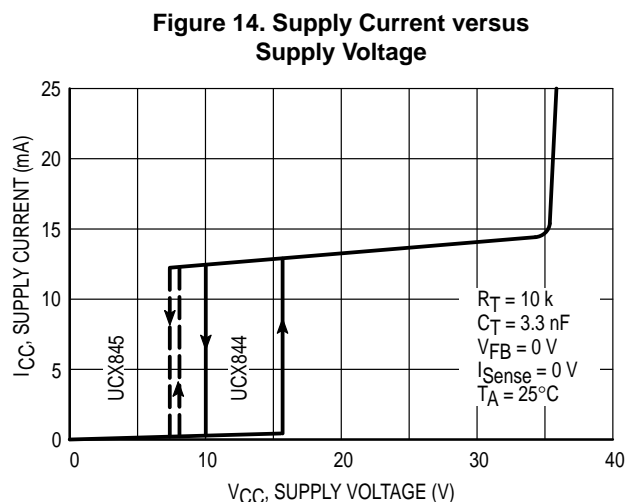
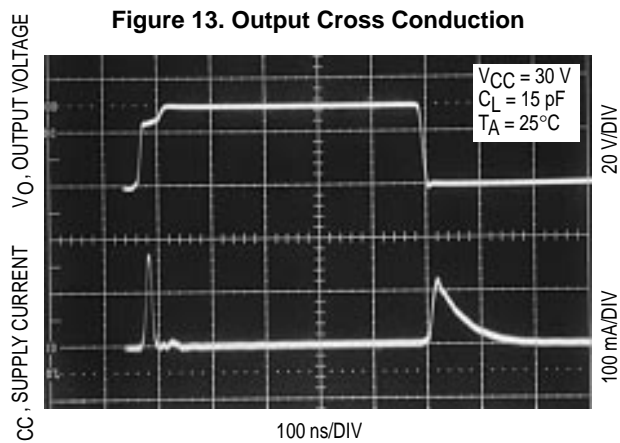


Figure 12. Output Waveform



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PIN FUNCTION DESCRIPTION

Pin		Function	Description
8-Pin	14-Pin		
1	1	Compensation	This pin is Error Amplifier output and is made available for loop compensation.
2	3	Voltage Feedback	This is the inverting input of the Error Amplifier. It is normally connected to the switching power supply output through a resistor divider.
3	5	Current Sense	A voltage proportional to inductor current is connected to this input. The PWM uses this information to terminate the output switch conduction.
4	7	R_T/C_T	The Oscillator frequency and maximum Output duty cycle are programmed by connecting resistor R_T to V_{ref} and capacitor C_T to ground. Operation to 1.0 MHz is possible.
5	–	Gnd	This pin is combined control circuitry and power ground (8-pin package only).
6	10	Output	This output directly drives the gate of a power MOSFET. Peak currents up to 1.0 A are sourced and sunk by this pin. The output switches at one-half the oscillator frequency.
7	12	V_{CC}	This pin is the positive supply of the control IC.
8	14	V_{ref}	This is the reference output. It provides charging current for capacitor C_T through resistor R_T .
–	8	Power Ground	This pin is a separate power ground return (14-pin package only) that is connected back to the power source. It is used to reduce the effects of switching transient noise on the control circuitry.
–	11	V_C	The Output high state (V_{OH}) is set by the voltage applied to this pin (14-pin package only). With a separate power source connection, it can reduce the effects of switching transient noise on the control circuitry.
–	9	Gnd	This pin is the control circuitry ground return (14-pin package only) and is connected to back to the power source ground.
–	2,4,6,13	NC	No connection (14-pin package only). These pins are not internally connected.

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OPERATING DESCRIPTION

The UC3844, UC3845 series are high performance, fixed frequency, current mode controllers. They are specifically designed for Off-Line and dc-to-dc converter applications offering the designer a cost effective solution with minimal external components. A representative block diagram is shown in Figure 15.

Oscillator

The oscillator frequency is programmed by the values selected for the timing components R_T and C_T . Capacitor C_T is charged from the 5.0 V reference through resistor R_T to approximately 2.8 V and discharged to 1.2 V by an internal current sink. During the discharge of C_T , the oscillator generates an internal blanking pulse that holds the center input of the NOR gate high. This causes the Output to be in a low state, thus producing a controlled amount of output deadtime. An internal flip-flop has been incorporated in the UCX844/5 which blanks the output off every other clock cycle by holding one of the inputs of the NOR gate high. This in combination with the C_T discharge period yields output deadtimes programmable from 50% to 70%. Figure 1 shows R_T versus Oscillator Frequency and figure 2, Output Deadtime versus Frequency, both for given values of C_T . Note that many values of R_T and C_T will give the same oscillator frequency but only one combination will yield a specific output deadtime at a given frequency.

In many noise sensitive applications it may be desirable to frequency-lock the converter to an external system clock. This can be accomplished by applying a clock signal to the circuit shown in Figure 17. For reliable locking, the free-running oscillator frequency should be set about 10% less than the clock frequency. A method for multi unit synchronization is shown in Figure 18. By tailoring the clock waveform, accurate Output duty cycle clamping can be achieved to realize output deadtimes of greater than 70%

Error Amplifier

A fully compensated Error Amplifier with access to the inverting input and output is provided. It features a typical dc voltage gain of 90 dB, and a unity gain bandwidth of 1.0 MHz with 57 degrees of phase margin (Figure 5). The noninverting input is internally biased at 2.5 V and is not pinned out. The converter output voltage is typically divided down and monitored by the inverting input. The maximum input bias current is $-2.0 \mu\text{A}$ which can cause an output voltage error that is equal to the product of the input bias current and the equivalent input divider source resistance.

The Error Amp Output (Pin 1) is provide for external loop compensation (Figure 28). The output voltage is offset by two diode drops ($\approx 1.4 \text{ V}$) and divided by three before it connects to the inverting input of the Current Sense Comparator. This guarantees that no drive pulses appear at the Output (Pin 6) when Pin 1 is at its lowest state (V_{OL}). This occurs when the power supply is operating and the load is removed, or at the beginning of a soft-start interval (Figures 20, 21). The Error

Amp minimum feedback resistance is limited by the amplifier's source current (0.5 mA) and the required output voltage (V_{OH}) to reach the comparator's 1.0 V clamp level:

$$R_{f(\text{min})} \approx \frac{3.0 (1.0 \text{ V}) + 1.4 \text{ V}}{0.5 \text{ mA}} = 8800 \Omega$$

Current Sense Comparator and PWM Latch

The UC3844, UC3845 operate as a current mode controller, whereby output switch conduction is initiated by the oscillator and terminated when the peak inductor current reaches the threshold level established by the Error Amplifier Output/Compensation (Pin1). Thus the error signal controls the inductor current on a cycle-by-cycle basis. The current Sense Comparator PWM Latch configuration used ensures that only a single pulse appears at the Output during any given oscillator cycle. The inductor current is converted to a voltage by inserting the ground referenced sense resistor R_S in series with the source of output switch Q1. This voltage is monitored by the Current Sense Input (Pin 3) and compared a level derived from the Error Amp Output. The peak inductor current under normal operating conditions is controlled by the voltage at pin 1 where:

$$I_{pk} = \frac{V_{(\text{Pin } 1)} - 1.4 \text{ V}}{3 R_S}$$

Abnormal operating conditions occur when the power supply output is overloaded or if output voltage sensing is lost. Under these conditions, the Current Sense Comparator threshold will be internally clamped to 1.0 V. Therefore the maximum peak switch current is:

$$I_{pk(\text{max})} = \frac{1.0 \text{ V}}{R_S}$$

When designing a high power switching regulator it becomes desirable to reduce the internal clamp voltage in order to keep the power dissipation of R_S to a reasonable level. A simple method to adjust this voltage is shown in Figure 19. The two external diodes are used to compensate the internal diodes yielding a constant clamp voltage over temperature. Erratic operation due to noise pickup can result if there is an excessive reduction of the $I_{pk(\text{max})}$ clamp voltage.

A narrow spike on the leading edge of the current waveform can usually be observed and may cause the power supply to exhibit an instability when the output is lightly loaded. This spike is due to the power transformer interwinding capacitance and output rectifier recovery time. The addition of an RC filter on the Current Sense Input with a time constant that approximates the spike duration will usually eliminate the instability; refer to Figure 23.

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Figure 15. Representative Block Diagram

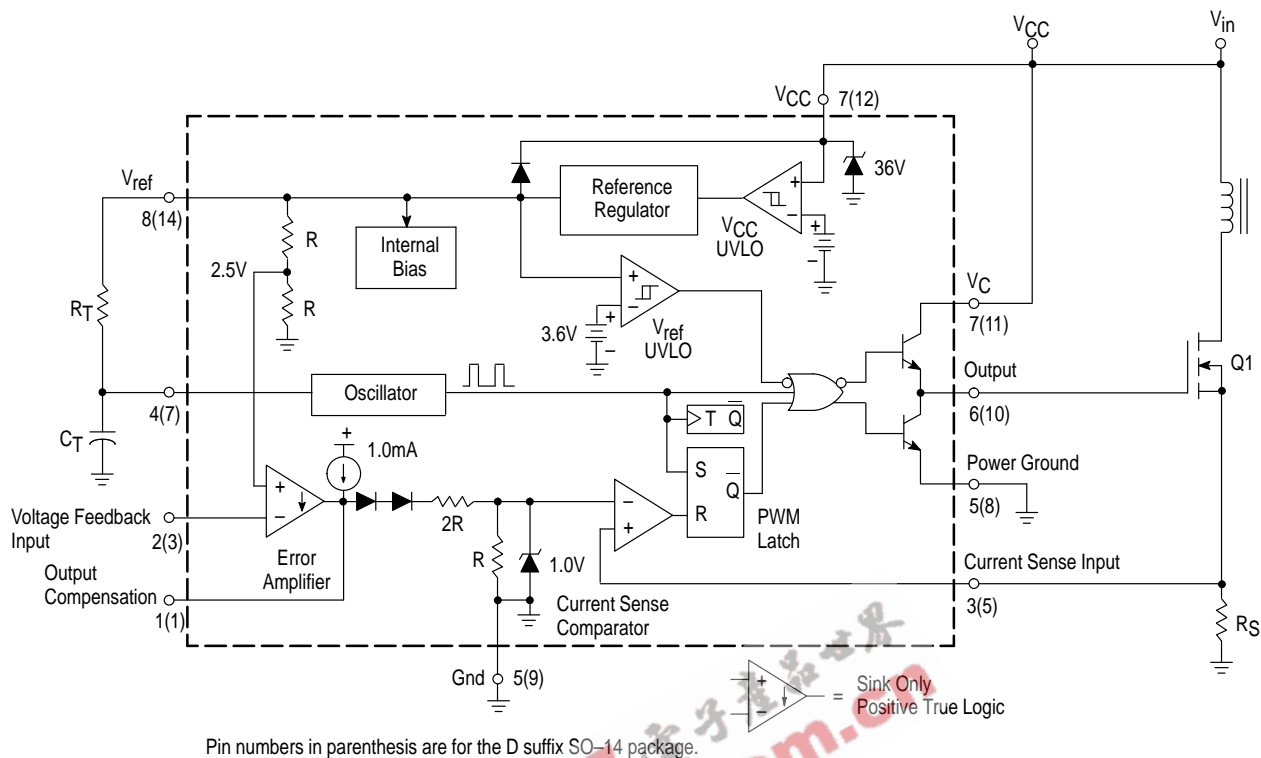
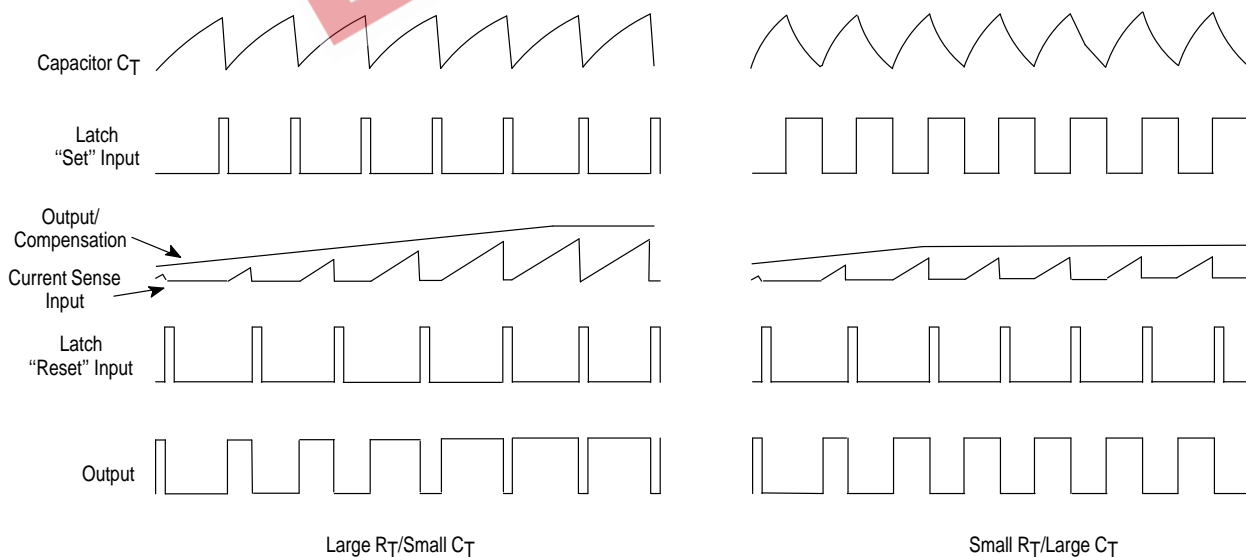


Figure 16. Timing Diagram



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Undervoltage Lockout

Two undervoltage lockout comparators have been incorporated to guarantee that the IC is fully functional before the output stage is enabled. The positive power supply terminal (V_{CC}) and the reference output (V_{ref}) are each monitored by separate comparators. Each has built-in hysteresis to prevent erratic output behavior as their respective thresholds are crossed. The V_{CC} comparator upper and lower thresholds are 16 V/10 V for the UCX844, and 8.4 V/7.6 V for the UCX845. The V_{ref} comparator upper and lower thresholds are 3.6 V/3.4 V. The large hysteresis and low startup current of the UCX844 makes it ideally suited in off-line converter applications where efficient bootstrap startup techniques later required (Figure 29). The UCX845 is intended for lower voltage dc-to-dc converter applications. A 36 V zener is connected as a shunt regulator from V_{CC} to ground. Its purpose is to protect the IC from excessive voltage that can occur during system startup. The minimum operating voltage for the UCX844 is 11 V and 8.2 V for the UCX845.

Output

These devices contain a single totem pole output stage that was specifically designed for direct drive of power MOSFETs. It is capable of up to ± 1.0 A peak drive current and has a typical rise and fall time of 50 ns with a 1.0 nF load. Additional internal circuitry has been added to keep the Output in a sinking mode whenever undervoltage lockout is active. This characteristic eliminates the need for an external pull-down resistor.

The SO-14 surface mount package provides separate pins for V_C (output supply) and Power Ground. Proper implementation will significantly reduce the level of switching transient noise imposed on the control circuitry. This becomes particularly useful when reducing the $I_{pk(max)}$ clamp level. The separate V_C supply input allows the designer

added flexibility in tailoring the drive voltage independent of V_{CC} . A zener clamp is typically connected to this input when driving power MOSFETs in systems where V_{CC} is greater than 20 V. Figure 22 shows proper power and control ground connections in a current sensing power MOSFET application.

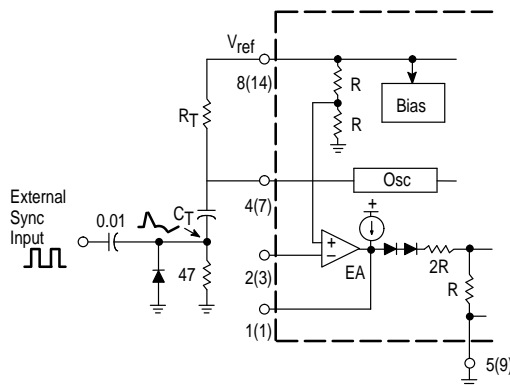
Reference

The 5.0 V bandgap reference is trimmed to $\pm 1.0\%$ tolerance at $T_J = 25^\circ\text{C}$ on the UC284X, and $\pm 2.0\%$ on the UC384X. Its primary purpose is to supply charging current to the oscillator timing capacitor. The reference has short circuit protection and is capable of providing in excess of 20 mA for powering additional control system circuitry.

Design Considerations

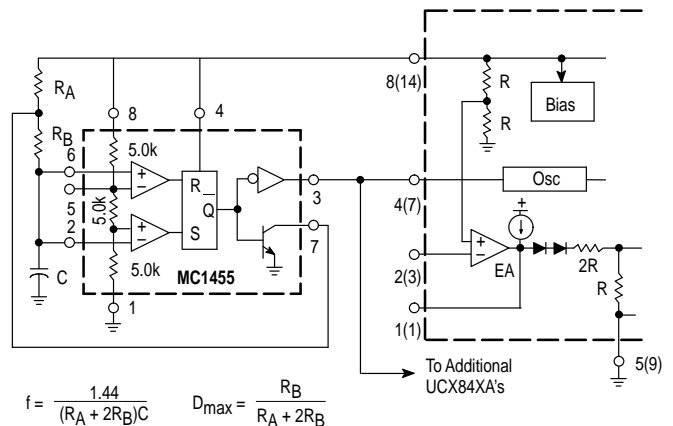
Do not attempt to construct the converter on wire-wrap or plug-in prototype boards. High frequency circuit layout techniques are imperative to prevent pulsewidth jitter. This is usually caused by excessive noise pick-up imposed on the Current Sense or Voltage Feedback inputs. Noise immunity can be improved by lowering circuit impedances at these points. The printed circuit layout should contain a ground plane with low-current signal and high-current switch and output grounds returning on separate paths back to the input filter capacitor. Ceramic bypass capacitors (0.1 μF) connected directly to V_{CC} , V_C , and V_{ref} may be required depending upon circuit layout. This provides a low impedance path for filtering the high frequency noise. All high current loops should be kept as short as possible using heavy copper runs to minimize radiated EMI. The Error Amp compensation circuitry and the converter output voltage divider should be located close to the IC and as far as possible from the power switch and other noise generating components.

Figure 17. External Clock Synchronization



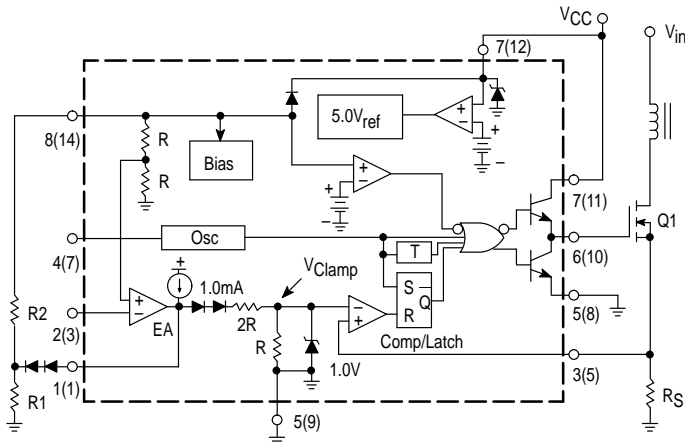
The diode clamp is required if the Sync amplitude is large enough to cause the bottom side of CT to go more than 300 mV below ground.

Figure 18. External Duty Cycle Clamp and Multi-Unit Synchronization



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Figure 19. Adjustable Reduction of Clamp Level



$$V_{Clamp} = \frac{1.67}{\left(\frac{R_2}{R_1} + 1\right)} + 0.33 \times 10^{-3} \left(\frac{R_1 R_2}{R_1 + R_2}\right)$$

$$I_{pk(max)} \approx \frac{V_{Clamp}}{R_S}$$

Where: $0 \leq V_{Clamp} \leq 1.0 \text{ V}$

Figure 20. Soft-Start Circuit

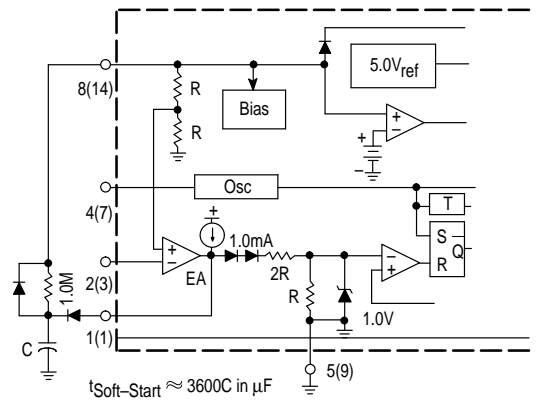
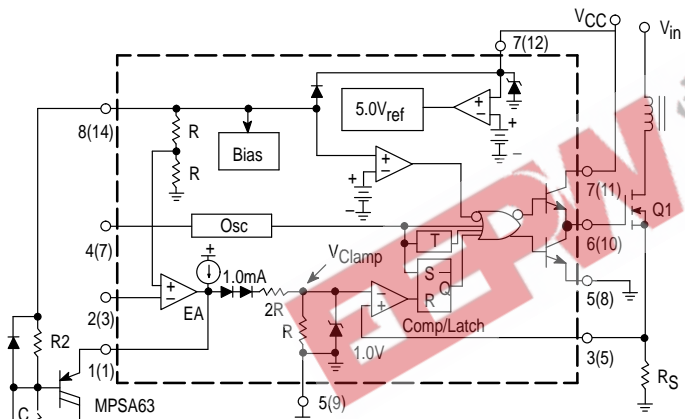


Figure 21. Adjustable Buffered Reduction of Clamp Level with Soft-Start

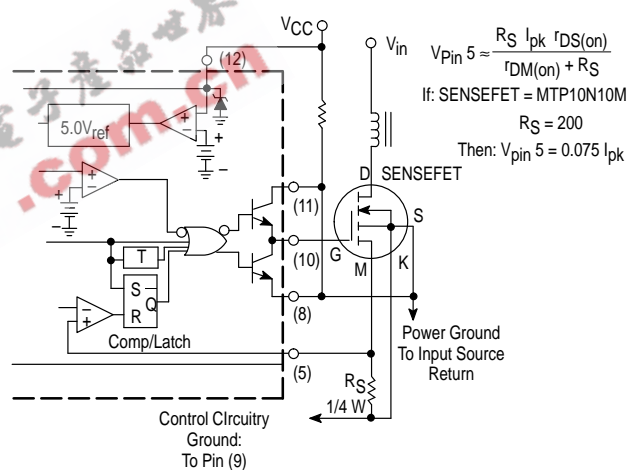


$$V_{Clamp} = \frac{1.67}{\left(\frac{R_2}{R_1} + 1\right)} + 0.33 \times 10^{-3} \frac{R_1 R_2}{R_1 + R_2}$$

$$I_{pk(max)} \approx \frac{V_{Clamp}}{R_S} \quad \text{Where: } 0 \leq V_{Clamp} \leq 1.0 \text{ V}$$

$$t_{Softstart} = -\ln \left[1 - \frac{V_C}{3V_{Clamp}} \right] C \frac{R_1 R_2}{R_1 + R_2}$$

Figure 22. Current Sensing Power MOSFET

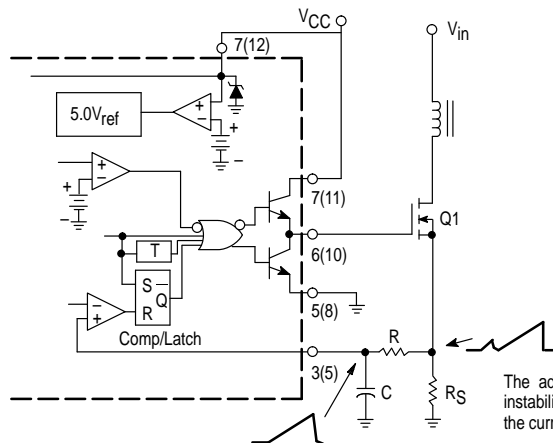


$$V_{Pin 5} = \frac{R_S I_{pk} r_{DS(on)}}{r_{DM(on)} + R_S}$$

If: SENSEFET = MTP10N10M
 $R_S = 200$
 Then: $V_{pin 5} = 0.075 I_{pk}$

Virtually lossless current sensing can be achieved with the implement of a SENSEFET power switch. For proper operation during over current conditions, a reduction of the $I_{pk(max)}$ clamp level must be implemented. Refer to Figures 19 and 21.

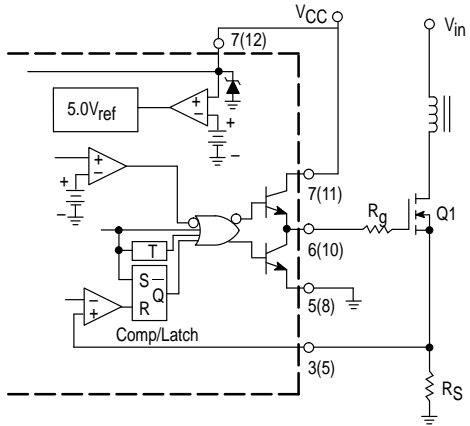
Figure 23. Current Waveform Spike Suppression



The addition of the RC filter will eliminate instability caused by the leading edge spike on the current waveform.

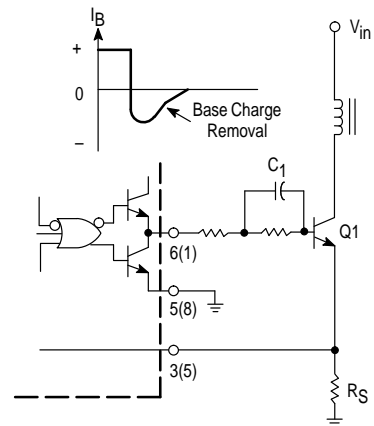
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Figure 24. MOSFET Parasitic Oscillations



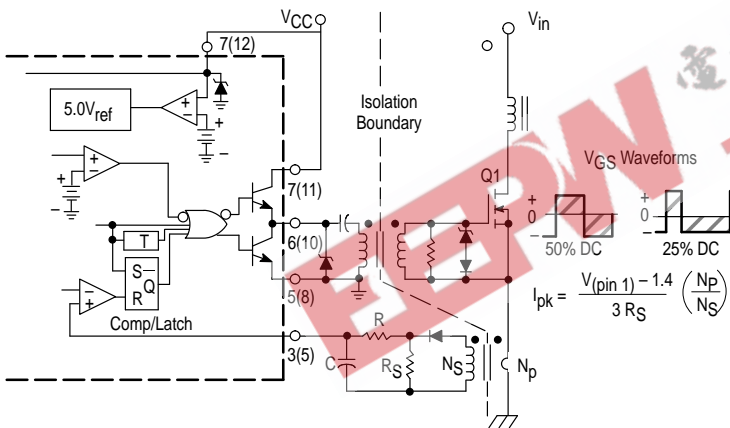
Series gate resistor R_g will damp any high frequency parasitic oscillations caused by the MOSFET input capacitance and any series wiring inductance in the gate-source circuit.

Figure 25. Bipolar Transistor Drive



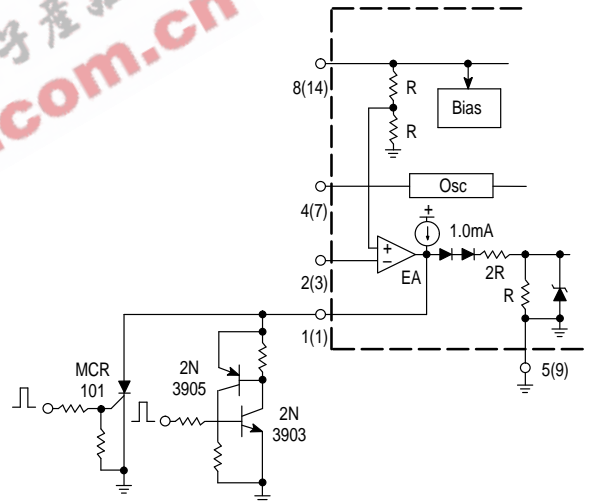
The totem-pole output can furnish negative base current for enhanced transistor turn-off, with the addition of capacitor C_1 .

Figure 26. Isolated MOSFET Drive



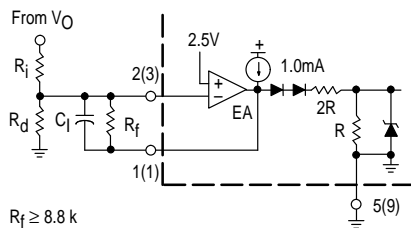
$$I_{pk} = \frac{V(\text{pin } 1) - 1.4}{3 R_S} \left(\frac{N_p}{N_s} \right)$$

Figure 27. Latched Shutdown

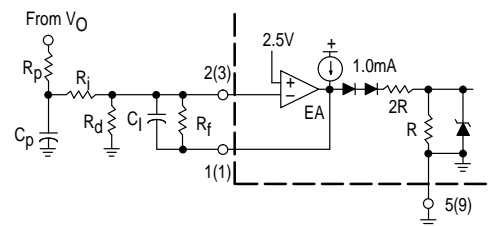


The MCR101 SCR must be selected for a holding of less than 0.5 mA at $T_A(\text{min})$. The simple two transistor circuit can be used in place of the SCR as shown. All resistors are 10 k.

Figure 28. Error Amplifier Compensation



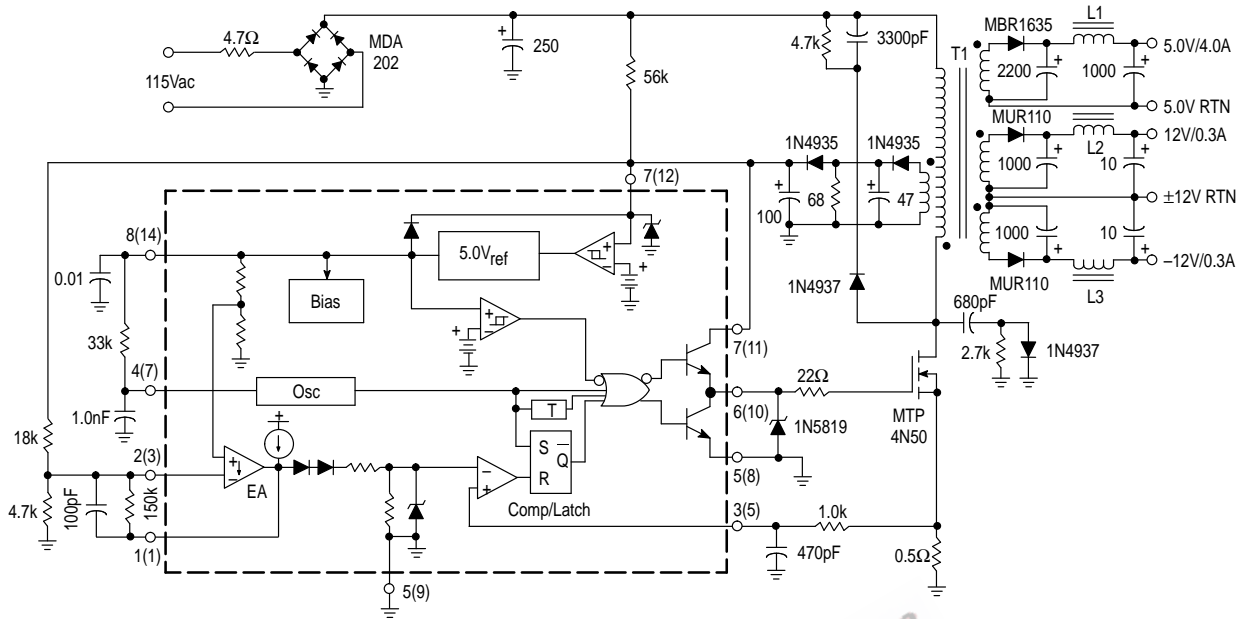
Error Amp compensation circuit for stabilizing any current-mode topology except for boost and flyback converters operating with continuous inductor current.



Error Amp compensation circuit for stabilizing current-mode boost and flyback topologies operating with continuous inductor current.

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Figure 29. 27 Watt Off-Line Flyback Regulator



T1 – Primary: 45 Turns # 26 AWG
 Secondary ± 12 V: 9 Turns # 30 AWG
 (2 strands) Bifilar Wound
 Secondary 5.0 V: 4 Turns (six strands)
 #26 Hexfilar Wound
 Secondary Feedback: 10 Turns #30 AWG
 (2 strands) Bifilar Wound
 Core: Ferroxcube EC35-3C8
 Bobbin: Ferroxcube EC35PCB1
 Gap ≈ 0.01" for a primary inductance of 1.0 mH

L1 – 15 μH at 5.0 A, Coilcraft Z7156.
 L2, L3 – 25 μH at 1.0 A, Coilcraft Z7157.

Test	Conditions	Results
Line Regulation: 5.0 V ± 12 V	$V_{in} = 95 \text{ Vac to } 130 \text{ Vac}$	$\Delta = 50 \text{ mV or } \pm 0.5\%$ $\Delta = 24 \text{ mV or } \pm 0.1\%$
Load Regulation: 5.0 V ± 12 V	$V_{in} = 115 \text{ Vac, } I_{out} = 1.0 \text{ A to } 4.0 \text{ A}$ $V_{in} = 115 \text{ Vac, } I_{out} = 100 \text{ mA to } 300 \text{ mA}$	$\Delta = 300 \text{ mV or } \pm 3.0\%$ $\Delta = 60 \text{ mV or } \pm 0.25\%$
Output Ripple: 5.0 V ± 12 V	$V_{in} = 115 \text{ Vac}$	40 mV _{pp} 80 mV _{pp}
Efficiency	$V_{in} = 115 \text{ Vac}$	70%

All outputs are at nominal load currents, unless otherwise noted.

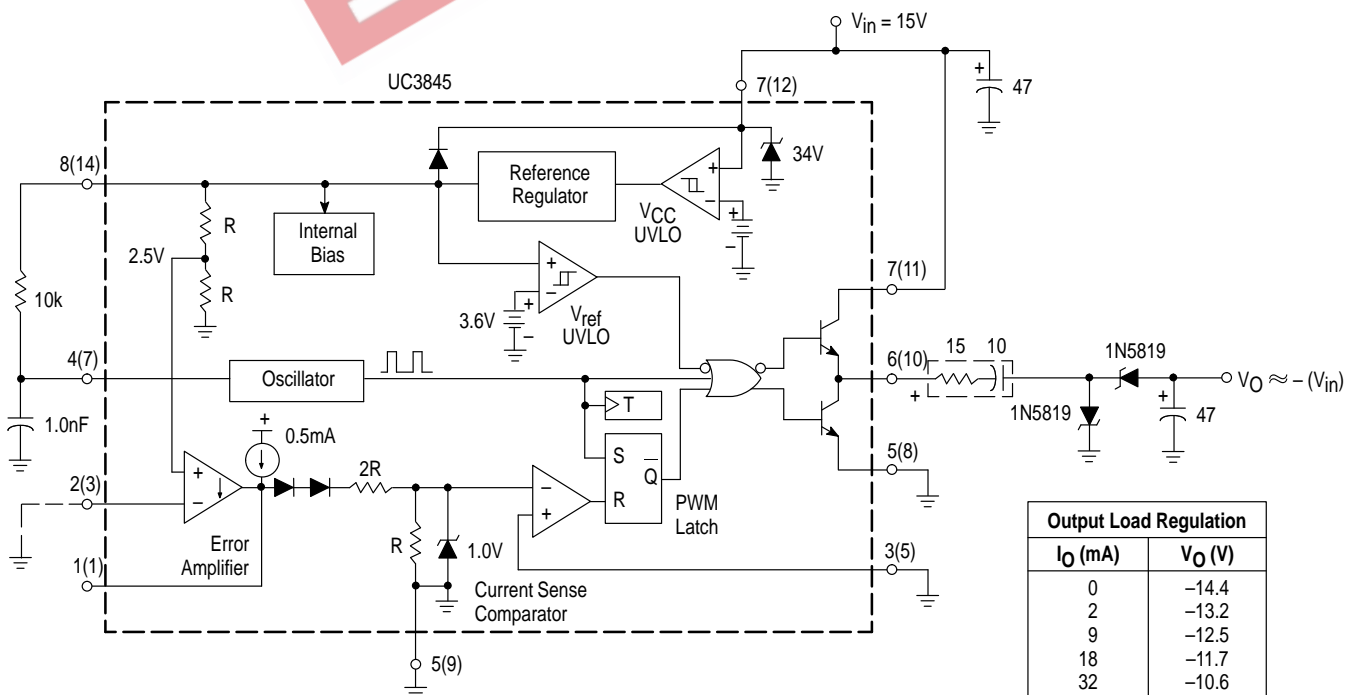
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Figure 30. Step-Up Charge Pump Converter



The capacitor's equivalent series resistance must limit the Drive Output current to 1.0 A. An additional series resistor may be required when using tantalum or other low ESR capacitors. The converter's output can provide excellent line and load regulation by connecting the R2/R1 resistor divider as shown.

Figure 31. Voltage-Inverting Charge Pump Converter



The capacitor's equivalent series resistance must limit the Drive Output current to 1.0 A. An additional series resistor may be required when using tantalum or other low ESR capacitors.

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OUTLINE DIMENSIONS

N SUFFIX
PLASTIC PACKAGE
CASE 626-05
ISSUE K

NOTE 2: SEATING PLANE

NOTES:
1. DIMENSION L TO CENTER OF LEAD WHEN FORMED PARALLEL.
2. PACKAGE CONTOUR OPTIONAL (ROUND OR SQUARE CORNERS).
3. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.

DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	9.40	10.16	0.370	0.400
B	6.10	6.60	0.240	0.260
C	3.94	4.45	0.155	0.175
D	0.38	0.51	0.015	0.020
F	1.02	1.78	0.040	0.070
G	2.54 BSC		0.100 BSC	
H	0.76	1.27	0.030	0.050
J	0.20	0.30	0.008	0.012
K	2.92	3.43	0.115	0.135
L	7.62 BSC		0.300 BSC	
M	— 10°		— 10°	
N	0.76	1.01	0.030	0.040

\oplus \varnothing 0.13 (0.005) M T A M B M

D SUFFIX
PLASTIC PACKAGE
CASE 751A-03
(SO-14)
ISSUE F

NOTE 2: SEATING PLANE

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: MILLIMETER.
3. DIMENSIONS A AND B DO NOT INCLUDE MOLD PROTRUSION.
4. MAXIMUM MOLD PROTRUSION 0.15 (0.006) PER SIDE.
5. DIMENSION D DOES NOT INCLUDE DAMBAR PROTRUSION. ALLOWABLE DAMBAR PROTRUSION SHALL BE 0.127 (0.005) TOTAL IN EXCESS OF THE D DIMENSION AT MAXIMUM MATERIAL CONDITION.

DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	8.55	8.75	0.337	0.344
B	3.80	4.00	0.150	0.157
C	1.35	1.75	0.054	0.068
D	0.35	0.49	0.014	0.019
F	0.40	1.25	0.016	0.049
G	1.27 BSC		0.050 BSC	
J	0.19	0.25	0.008	0.009
K	0.10	0.25	0.004	0.009
M	0° 7°		0° 7°	
P	5.80	6.20	0.228	0.244
R	0.25	0.50	0.010	0.019

\oplus 0.25 (0.010) M T B S A S

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