MOTOROLA

UC3842A, 43A
UC2842A, 43A

High Performance Current Mode Controllers

The UC3842A, UC3843A series of high performance fixed frequency current mode controllers 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 a trimmed oscillator for precise duty cycle control, 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.

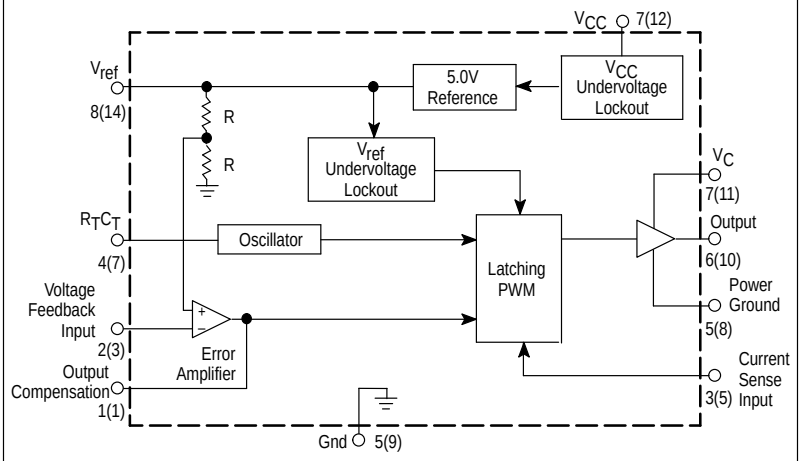
Also included are protective features consisting of input and reference undervoltage lockouts each with hysteresis, cycle–by–cycle current limiting, programmable output deadtime, and a latch for single pulse metering.

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 UCX842A has UYLO thresholds of 16 V (on) and 10 V (off), ideally suited for off–line converters. The UCX843A is tailored for lower voltage applications having UVLO thresholds of 8.5 V (on) and 7.6 V (off).

- Trimmed Oscillator Discharge Current for Precise Duty Cycle Control
- Current Mode Operation to 500 kHz
- Automatic Feed Forward Compensation
- Latching PWM for Cycle–By–Cycle Current Limiting
- Internally Trimmed Reference with Undervoltage Lockout
- High Current Totem Pole Output
- Undervoltage Lockout with Hysteresis
- Low Startup and Operating Current
- Direct Interface with Motorola SENSEFET Products

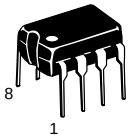
Simplified Block Diagram



Pin numbers in parenthesis are for the D suffix SO–14 package.

HIGH PERFORMANCE CURRENT MODE CONTROLLERS

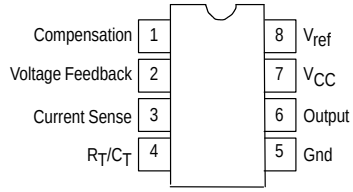
N SUFFIX
PLASTIC PACKAGE
CASE 626



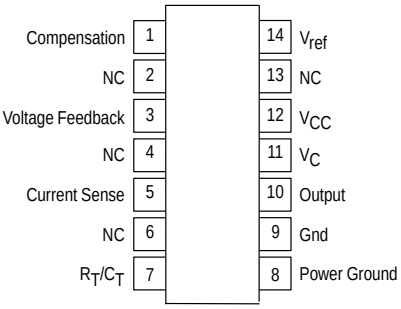
D SUFFIX
PLASTIC PACKAGE
CASE 751A
(SO–14)



PIN CONNECTIONS



(Top View)



(Top View)

ORDERING INFORMATION

Device	Operating Temperature Range	Package
UC3842AD	T_A = 0° to +70°C	SO–14
UC3843AD		SO–14
UC3842AN		Plastic
UC3843AN		Plastic
UC2842AD	T_A = – 25° to +85°C	SO–14
UC2843AD		SO–14
UC2842AN		Plastic
UC2843AN		Plastic

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MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Total Power Supply and Zener Current	$(I_{CC} + I_Z)$	30	mA
Output Current, Source or Sink (Note 1)	I_O	1.0	A
Output Energy (Capacitive Load per Cycle)	W	5.0	μJ
Current Sense and Voltage Feedback Inputs	V_{in}	– 0.3 to + 5.5	V
Error Amp Output Sink Current	I_O	10	mA
Power Dissipation and Thermal Characteristics D Suffix, Plastic Package Maximum Power Dissipation @ T_A = 25°C Thermal Resistance, Junction–to–Air N Suffix, Plastic Package Maximum Power Dissipation @ T_A = 25°C Thermal Resistance, Junction–to–Air	 P_D $R_{\theta JA}$ P_D $R_{\theta JA}$	 862 145 1.25 100	 mW °C/W W °C/W
Operating Junction Temperature	T_J	+ 150	°C
Operating Ambient Temperature UC3842A, UC3843A UC2842A, UC2843A	T_A	 0 to + 70 – 25 to + 85	°C
Storage Temperature Range	T_{stg}	– 65 to + 150	°C

ELECTRICAL CHARACTERISTICS (V_{CC} = 15 V, [Note 2], R_T = 10 k, C_T = 3.3 nF, T_A = T_{low} to T_{high} [Note 3], unless otherwise noted.)

Characteristics	Symbol	UC284XA			UC384XA			Unit
		Min	Typ	Max	Min	Typ	Max	
REFERENCE SECTION								
Reference Output Voltage (I_O = 1.0 mA, T_J = 25°C)	V_{ref}	4.95	5.0	5.05	4.9	5.0	5.1	V
Line Regulation (V_{CC} = 12 V to 25 V)	Reg_{line}	–	2.0	20	–	2.0	20	mV
Load Regulation (I_O = 1.0 mA to 20 mA)	Reg_{load}	–	3.0	25	–	3.0	25	mV
Temperature Stability	T_S	–	0.2	–	–	0.2	–	mV/°C
Total Output Variation over Line, Load, Temperature	V_{ref}	4.9	–	5.1	4.82	–	5.18	V
Output Noise Voltage (f = 10 Hz to 10 kHz, T_J = 25°C)	V_n	–	50	–	–	50	–	μV
Long Term Stability (T_A = 125°C for 1000 Hours)	S	–	5.0	–	–	5.0	–	mV
Output Short Circuit Current	I_{SC}	– 30	– 85	– 180	– 30	– 85	– 180	mA
OSCILLATOR SECTION								
Frequency T_J = 25°C T_A = T_{low} to T_{high}	f_{osc}	 47 46	 52 –	 57 60	 47 46	 52 –	 57 60	kHz
Frequency Change with Voltage (V_{CC} = 12 V to 25 V)	$\Delta f_{osc}/\Delta V$	–	0.2	1.0	–	0.2	1.0	%
Frequency Change with Temperature T_A = T_{low} to T_{high}	$\Delta f_{osc}/\Delta T$	–	5.0	–	–	5.0	–	%
Oscillator Voltage Swing (Peak–to–Peak)	V_{osc}	–	1.6	–	–	1.6	–	V
Discharge Current (V_{osc} = 2.0 V) T_J = 25°C T_A = T_{low} to T_{high}	I_{dischg}	 7.5 7.2	 8.4 –	 9.3 9.5	 7.5 7.2	 8.4 –	 9.3 9.5	mA

NOTES: 1. Maximum Package power dissipation limits must be observed.
2. Adjust V_{CC} above the Startup threshold before setting to 15 V.
3. Low duty cycle pulse techniques are used during test to maintain junction temperature as close to ambient as possible
T_{low} = 0°C for UC3842A, UC3843A; –25°C for UC2842A, UC2843A
T_{high} = +70°C for UC3842A, UC3843A; +85°C for UC2842A, UC2843A

ELECTRICAL CHARACTERISTICS (V_{CC} = 15 V, [Note 2], R_T = 10 k, C_T = 3.3 nF, T_A = T_{low} to T_{high} [Note 3], unless otherwise noted.)

Characteristics	Symbol	UC284XA			UC384XA			Unit
		Min	Typ	Max	Min	Typ	Max	
ERROR AMPLIFIER SECTION								
Voltage Feedback Input (V_O = 2.5 V)	V_{FB}	2.45	2.5	2.55	2.42	2.5	2.58	V
Input Bias Current (V_{FB} = 2.7 V)	I_{IB}	–	–0.1	–1.0	–	–0.1	–2.0	μA
Open Loop Voltage Gain (V_O = 2.0 V to 4.0 V)	A_{VOL}	65	90	–	65	90	–	dB
Unity Gain Bandwidth (T_J = 25°C)	BW	0.7	1.0	–	0.7	1.0	–	MHz
Power Supply Rejection Ratio (V_{CC} = 12 V to 25 V)	PSRR	60	70	–	60	70	–	dB
Output Current								mA
Sink (V_O = 1.1 V, V_{FB} = 2.7 V)	I_{Sink}	2.0	12	–	2.0	12	–	
Source (V_O = 5.0 V, V_{FB} = 2.3 V)	I_{Source}	–0.5	–1.0	–	–0.5	–1.0	–	
Output Voltage Swing								V
High State (R_L = 15 k to ground, V_{FB} = 2.3 V)	V_{OH}	5.0	6.2	–	5.0	6.2	–	
Low State (R_L = 15 k to V_{ref}, V_{FB} = 2.7 V)	V_{OL}	–	0.8	1.1	–	0.8	1.1	
CURRENT SENSE SECTION								
Current Sense Input Voltage Gain (Notes 4 & 5)	A_V	2.85	3.0	3.15	2.85	3.0	3.15	V/V
Maximum Current Sense Input Threshold (Note 4)	V_{th}	0.9	1.0	1.1	0.9	1.0	1.1	V
Power Supply Rejection Ratio	PSRR							dB
V_{CC} = 12 to 25 V (Note 4)		–	70	–	–	70	–	
Input Bias Current	I_{IB}	–	–2.0	–10	–	–2.0	–10	μA
Propagation Delay (Current Sense Input to Output)	$t_{PLH(in/out)}$	–	150	300	–	150	300	ns
OUTPUT SECTION								
Output Voltage								V
Low State (I_{Sink} = 20 mA)	V_{OL}	–	0.1	0.4	–	0.1	0.4	
(I_{Sink} = 200 mA)		–	1.6	2.2	–	1.6	2.2	
High State (I_{Sink} = 20 mA)	V_{OH}	13	13.5	–	13	13.5	–	
(I_{Sink} = 200 mA)		12	13.4	–	12	13.4	–	
Output Voltage with UVLO Activated	$V_{OL(UVLO)}$							V
V_{CC} = 6.0 V, I_{Sink} = 1.0 mA		–	0.1	1.1	–	0.1	1.1	
Output Voltage Rise Time (C_L = 1.0 nF, T_J = 25°C)	t_r	–	50	150	–	50	150	ns
Output Voltage Fall Time (C_L = 1.0 nF, T_J = 25°C)	t_f	–	50	150	–	50	150	ns
UNDERVOLTAGE LOCKOUT SECTION								
Startup Threshold	V_{th}							V
UCX842A		15	16	17	14.5	16	17.5	
UCX843A		7.8	8.4	9.0	7.8	8.4	9.0	
Minimum Operating Voltage After Turn–On	$V_{CC(min)}$							V
UCX842A		9.0	10	11	8.5	10	11.5	
UCX843A		7.0	7.6	8.2	7.0	7.6	8.2	
PWM SECTION								
Duty Cycle								%
Maximum	DC_{max}	94	96	–	94	96	–	
Minimum	DC_{min}	–	–	0	–	–	0	
TOTAL DEVICE								
Power Supply Current (Note 2)	I_{CC}							mA
Startup:								
(V_{CC} = 6.5 V for UCX843A,								
14 V for UCX842A) Operating		–	0.5	1.0	–	0.5	1.0	
		–	12	17	–	12	17	
Power Supply Zener Voltage (I_{CC} = 25 mA)	V_Z	30	36	–	30	36	–	V

NOTES: 2. Adjust V_{CC} above the Startup threshold before setting to 15 V.

3. Low duty cycle pulse techniques are used during test to maintain junction temperature as close to ambient as possible
   T_{low} = 0°C for UC3842A, UC3843A; –25°C for UC2842A, UC2843A
   T_{high} = +70°C for UC3842A, UC3843A; +85°C for UC2842A, UC2843A

4. This parameter is measured at the latch trip point with V_{FB} = 0 V.

5. Comparator gain is defined as: $A_V \frac{\Delta V \text{ Output Compensation}}{\Delta V \text{ Current Sense Input}}$

Figure 1. Timing Resistor versus Oscillator Frequency

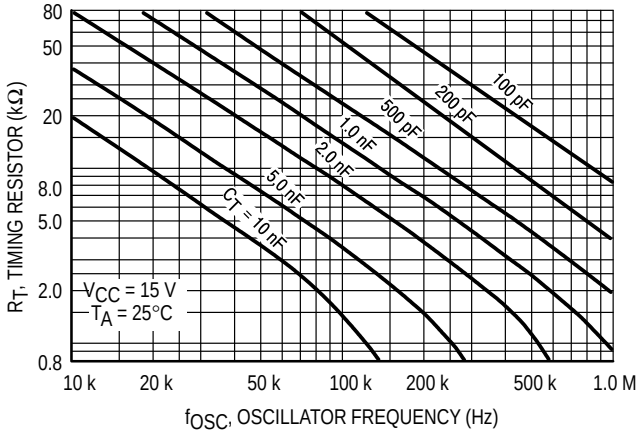


Figure 2. Output Deadtime versus Oscillator Frequency

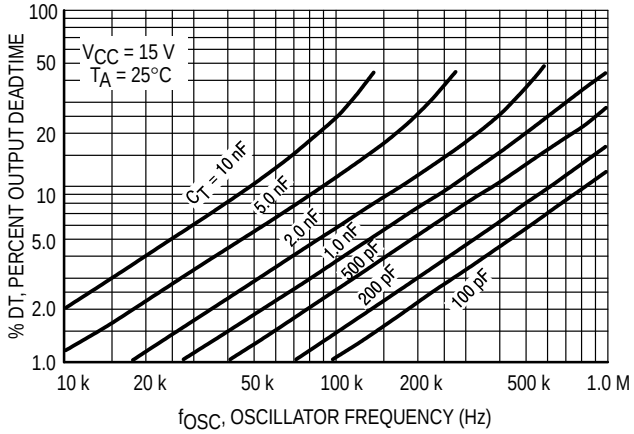


Figure 3. Oscillator Discharge Current versus Temperature

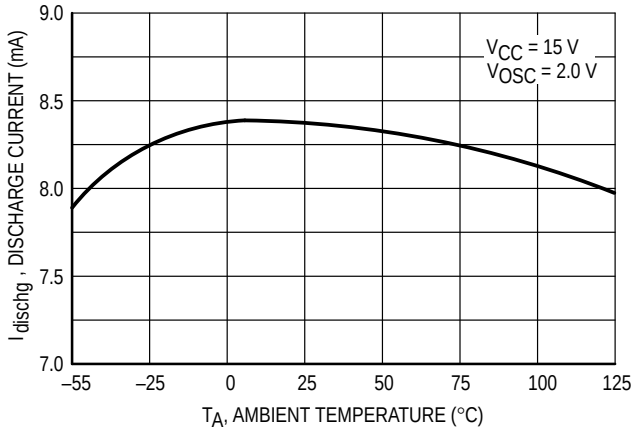


Figure 4. Maximum Output Duty Cycle versus Timing Resistor

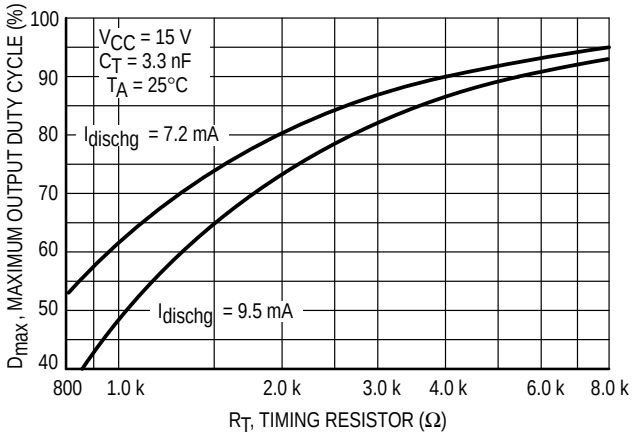


Figure 5. Error Amp Small Signal Transient Response

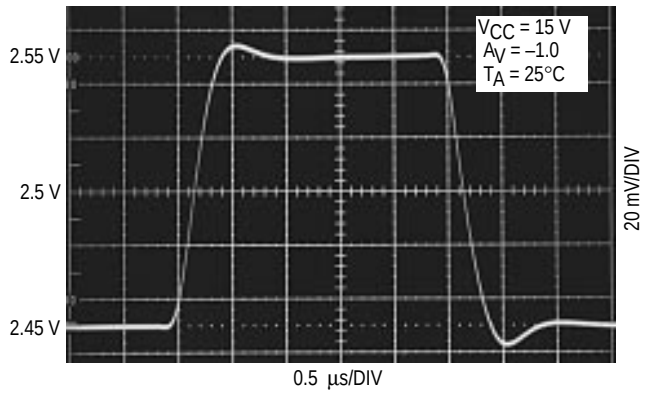


Figure 6. Error Amp Large Signal Transient Response

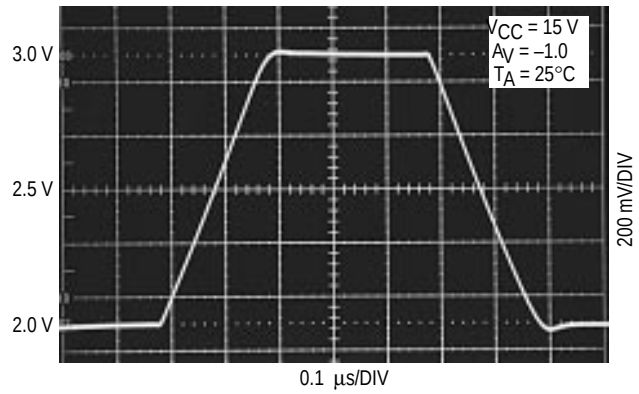


Figure 7. Error Amp Open Loop Gain and Phase versus Frequency

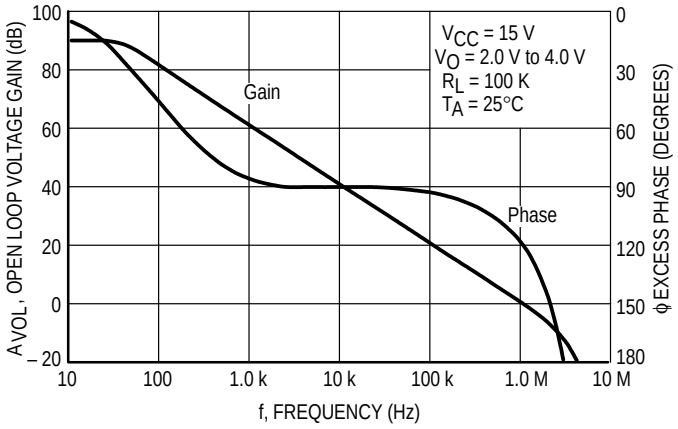


Figure 8. Current Sense Input Threshold versus Error Amp Output Voltage

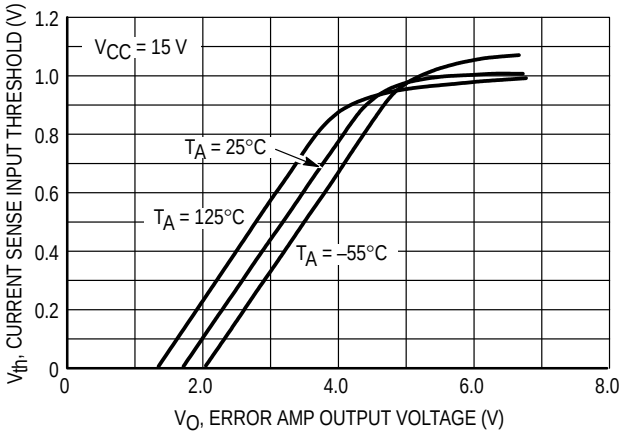


Figure 9. Reference Voltage Change versus Source Current

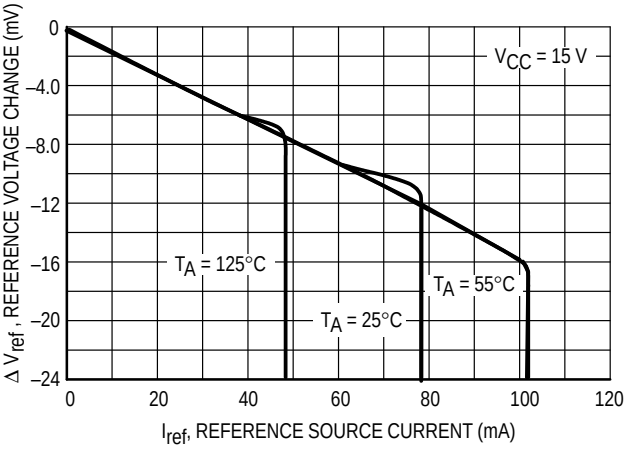


Figure 10. Reference Short Circuit Current versus Temperature

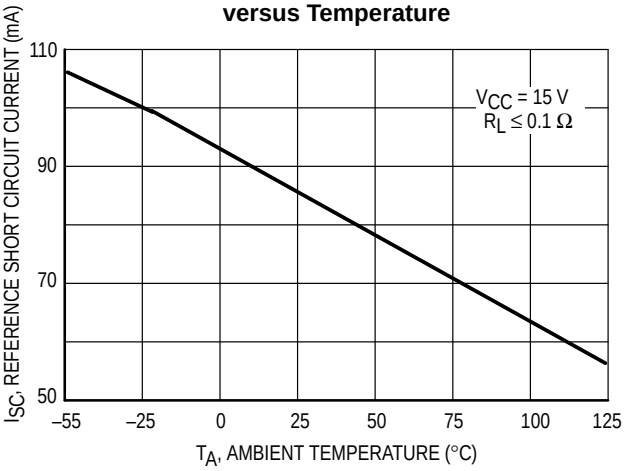


Figure 11. Reference Load Regulation

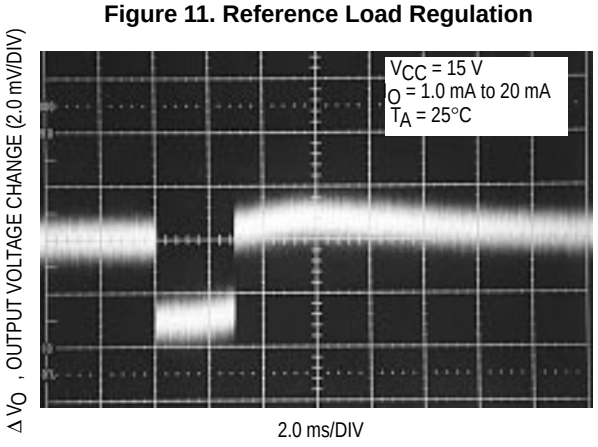


Figure 12. Reference Line Regulation

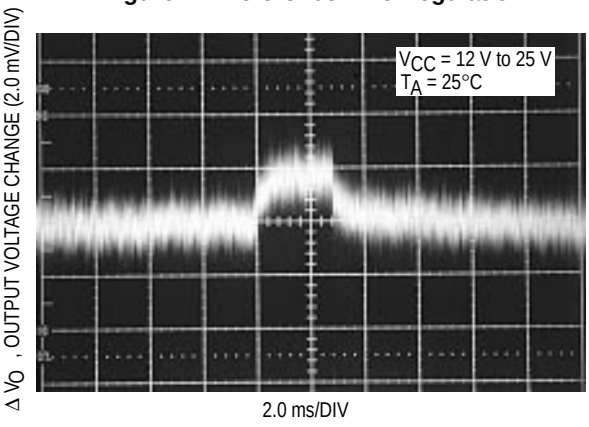


Figure 13. Output Saturation Voltage versus Load Current

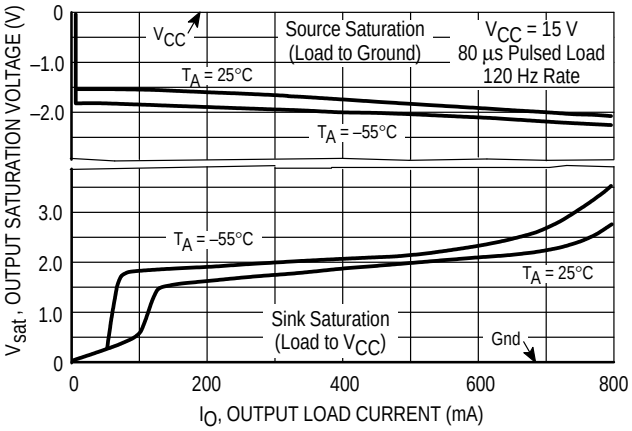


Figure 14. Output Waveform

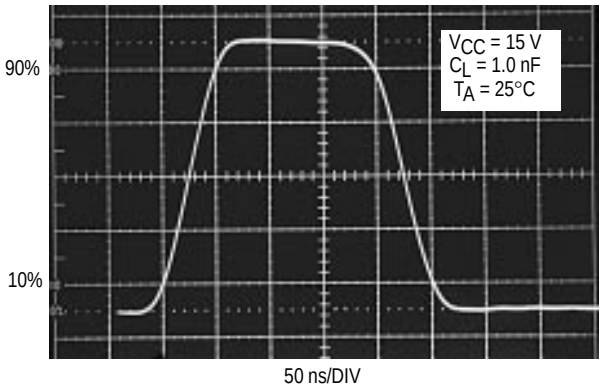


Figure 15. Output Cross Conduction

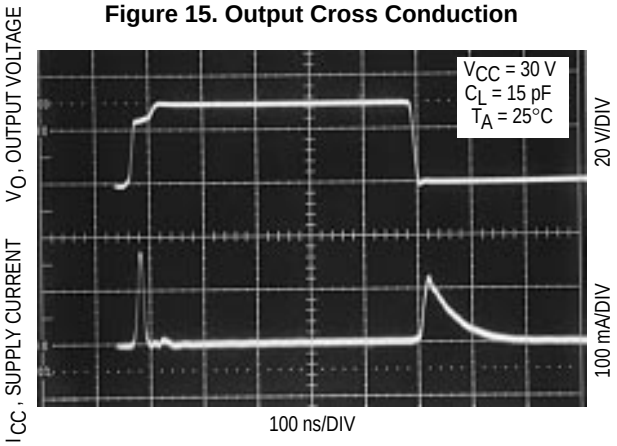


Figure 16. Supply Current versus Supply Voltatage

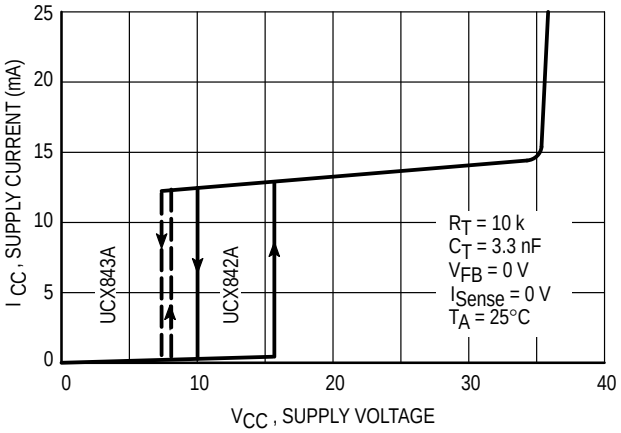
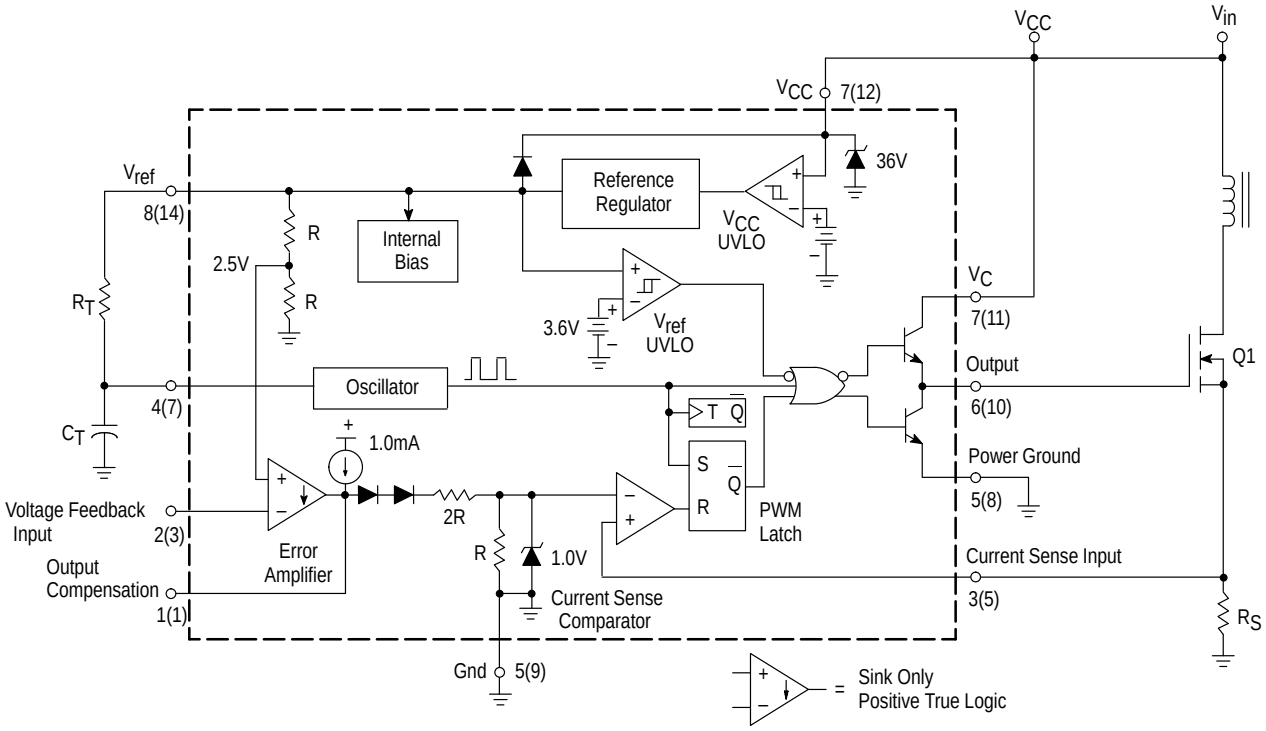
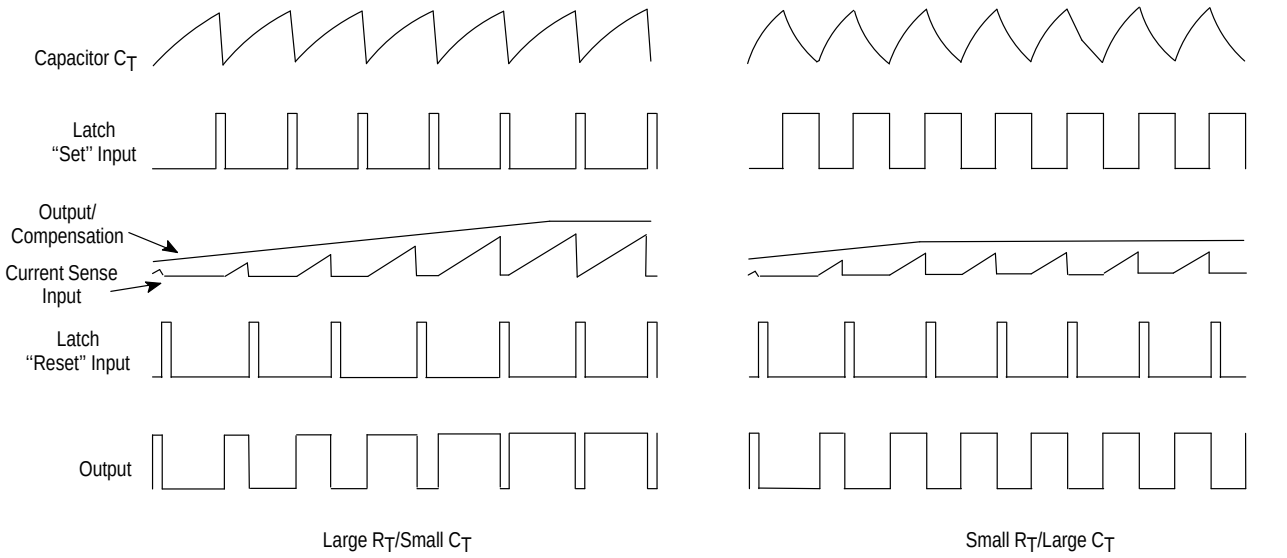


Figure 17. Representative Block Diagram



Pin numbers in parenthesis are for the D suffix SO–14 package.

Figure 18. Timing Diagram



OPERATING DESCRIPTION

The UC3842A, UC3843A 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 17.

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 and 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. 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. The oscillator thresholds are temperature compensated, and the discharge current is trimmed and guaranteed to within ± 10% at T_J = 25°C. These internal circuit refinements minimize variations of oscillator frequency and maximum output duty cycle. The results are shown in Figures 3 and 4.

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 20. 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 21. By tailoring the clock waveform, accurate Output duty cycle clamping can be achieved.

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 7). 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 μA 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 30). The output voltage is offset by two diode drops (≈ 1.4 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 23, 24). 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:

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

Current Sense Comparator and PWM Latch

The UC3842A, UC3843A 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 peak 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:

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

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:

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

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 22. 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 $I_{pk(max)}$ 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 26.

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 500 kHz is possible.
5	–	Gnd	This pin is the 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.
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 back to the power source ground.
–	2,4,6,13	NC	No connection (14–pin package only). These pins are not internally connected.

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 UCX842A, and 8.4 V/7.6 V for the UCX843A. The V_{ref} comparator upper and lower thresholds are 3.6V/3.4 V. The large hysteresis and low startup current of the UCX842A makes it ideally suited in off–line converter applications where efficient bootstrap startup techniques are required (Figure 33). The UCX843A is intended for lower voltage dc to dc converter applications. A 36 V zener is connected as a shunt regulator form 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 UCX842A is 11 V and 8.2 V for the UCX843A.

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 an 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 that 20 V. Figure 25 shows proper power and control ground connections in a current sensing power MOSFET application.

Reference

The 5.0 V bandgap reference is trimmed to ±1.0% tolerance at T_J = 25°C on the UC284XA, and ± 2.0% on the UC384XA. 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.

Current mode converters can exhibit subharmonic oscillations when operating at a duty cycle greater than 50% with continuous inductor current. This instability is independent of the regulators closed–loop characteristics and is caused by the simultaneous operating conditions of fixed frequency and peak current detecting. Figure 19A shows the phenomenon graphically. At t_0, switch conduction begins, causing the inductor current to rise at a slope of m_1. This slope is a function of the input voltage divided by the inductance. At t_1, the Current Sense Input reaches the threshold established by the control voltage. This causes the switch to turn off and the current to decay at a slope of m_2 until the next oscillator cycle. The unstable condition can be shown if a pertubation is added to the control voltage, resulting in a small ΔI (dashed line). With a fixed oscillator period, the current decay time is reduced, and the minimum current at switch turn–on (t_2) is increased by ΔI + ΔI m2/m1. The minimum current at the next cycle (t_3) decreases to (ΔI + ΔI m_2/m_1) (m_2/m_1). This pertubation is multiplied by $m_2.m_1$ on each succeeding cycle, alternately increasing and decreasing the inductor current at switch turn–on. Several oscillator cycles may be required before the inductor current reaches zero causing the process to commence again. If m_2/m_1 is greater than 1, the converter will be unstable. Figure 19B shows that by adding an artificial ramp that is synchronized with the PWM clock to the control voltage, the ΔI pertubation will decrease to zero on succeeding cycles. This compensation ramp (m_3) must have a slope equal to or slightly greater than $m_2/2$ for stability. With $m_2/2$ slope compensation, the average inductor current follows the control voltage yielding true current mode operation. The compensating ramp can be derived from the oscillator and added to either the Voltage Feedback or Current Sense inputs (Figure 32).

Figure 19. Continuous Current Waveforms

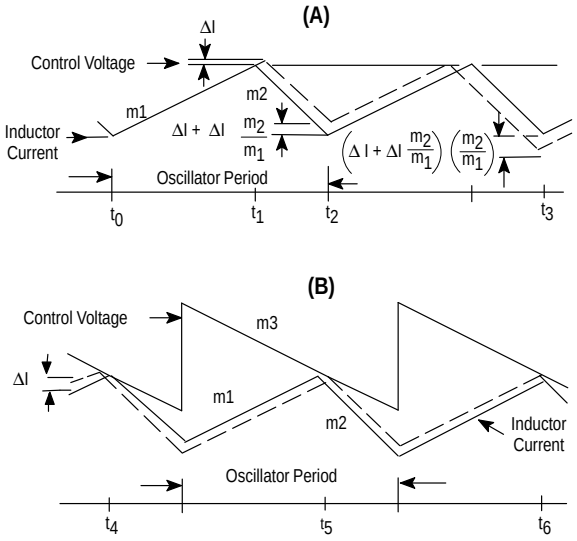
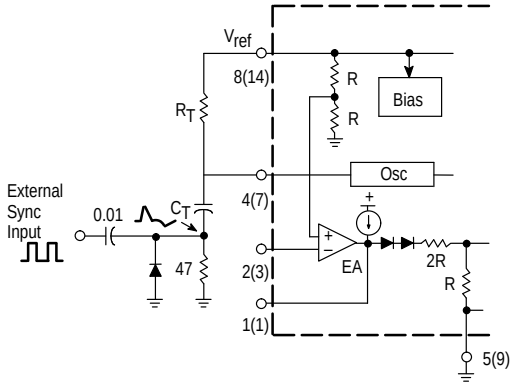
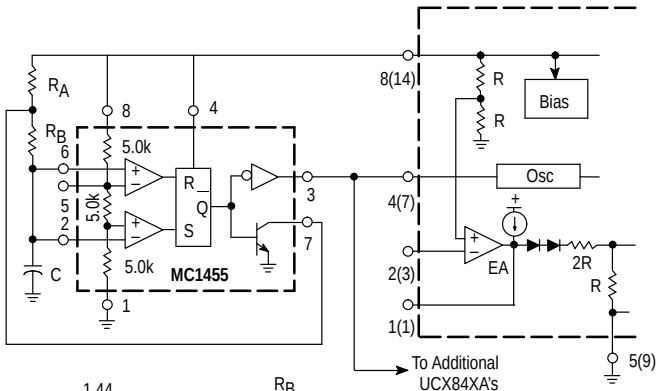


Figure 20. External Clock Synchronization



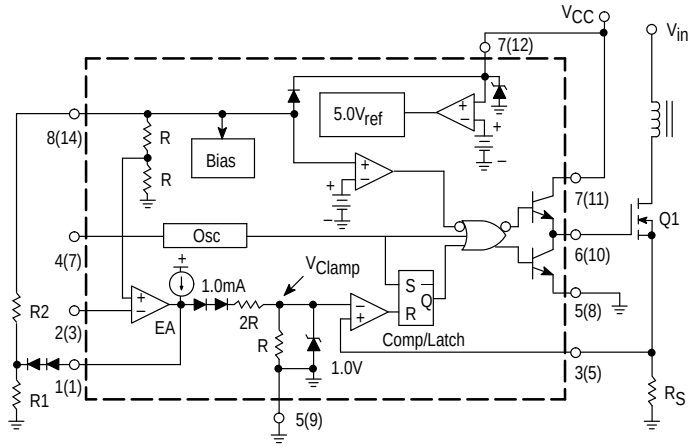
The diode clamp is required if the Sync amplitude is large enough to cause the bottom side of CT to go more than 300 mV below ground.

Figure 21. External Duty Cycle Clamp and Multi Unit Synchronization



$$f = \frac{1.44}{(R_A + 2R_B)C} \qquad D_{max} = \frac{R_B}{R_A + 2R_B}$$

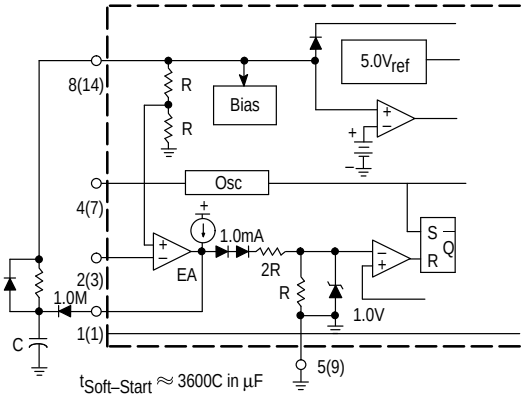
Figure 22. Adjustable Reduction of Clamp Level



$$V_{Clamp} = \frac{1.67}{\left(\frac{R_2}{R_1} + 1\right)} + 0.33 \times 10 - 3\left(\frac{R_1\ R_2}{R_1 + R_2}\right) \qquad I_{pk(max)} = \frac{VClamp}{RS}$$

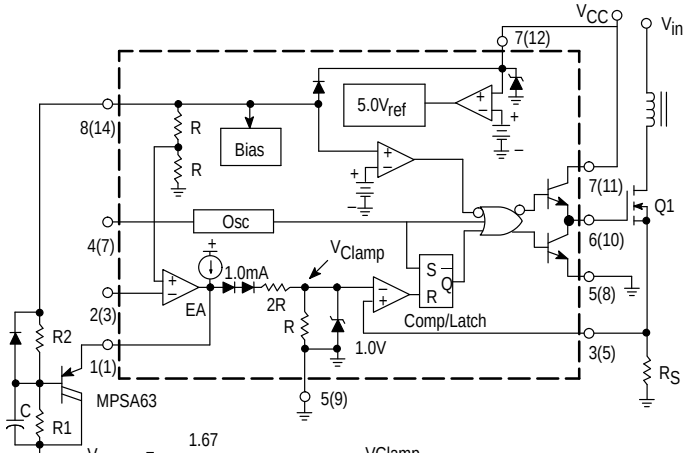
Where: $0 \le V_{Clamp} \le 1.0$ V

Figure 23. Soft–Start Circuit



$t_{Soft–Start} \approx 3600C$ in μF

Figure 24. Adjustable Buffered Reduction of Clamp Level with Soft–Start

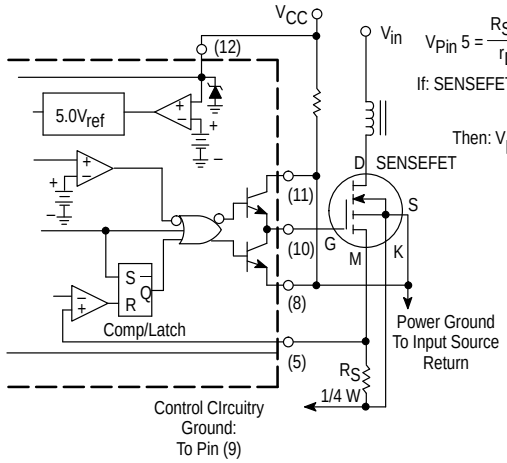


$$V_{Clamp} = \frac{1.67}{\left(\frac{R_2}{R_1} + 1\right)} \qquad I_{pk(max)} = \frac{VClamp}{RS}$$

Where: $0 \le V_{Clamp} \le 1.0$ V

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

Figure 25. Current Sensing Power MOSFET



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

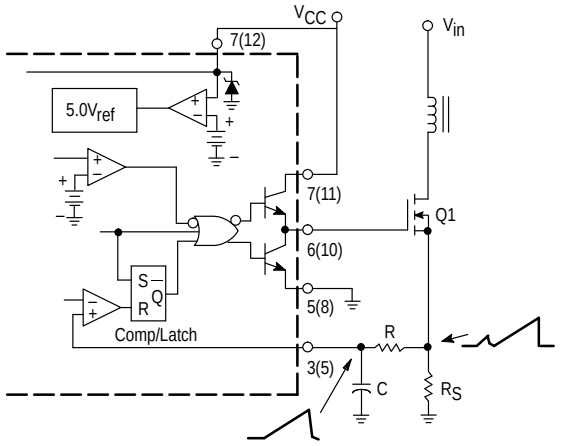
If: SENSEFET = MTP10N10M

$R_S = 200$

Then: $V_{pin\ 5} = 0.075\ I_{pk}$

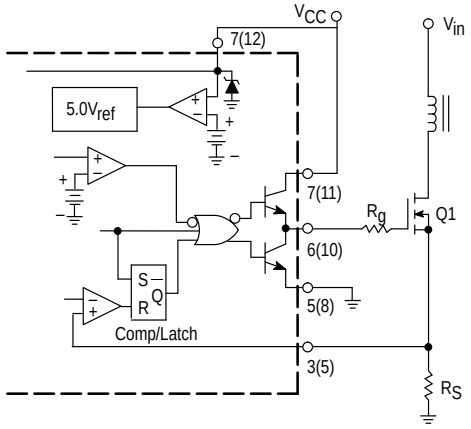
Virtually lossless current sensing can be achieved with the implementation of a SENSEFET power switch. For proper operation during over current conditions, a reduction of the $I_{pk(max)}$ clamp level must be implemented. Refer to Figures 22 and 24.

Figure 26. Current Waveform Spike Suppression



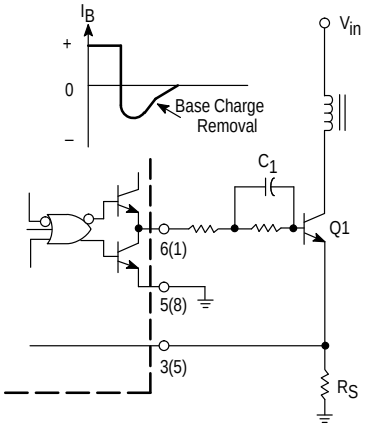
The addition of the RC filter will eliminate instability caused by the leading edge spike on the current waveform.

Figure 27. 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 28. 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 29. Isolated MOSFET Drive

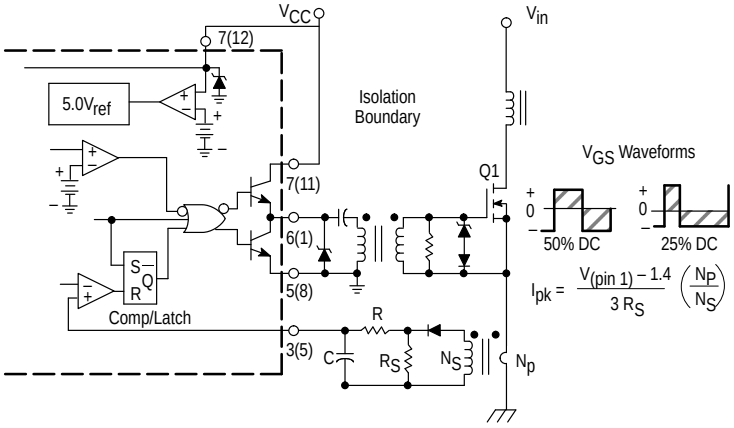
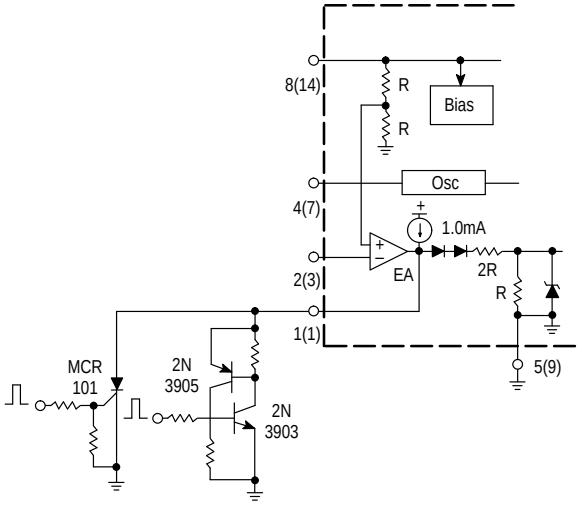
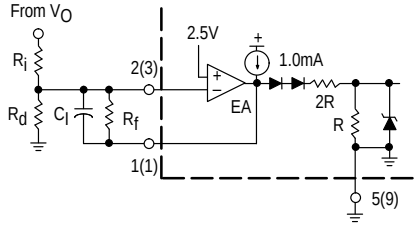


Figure 30. Latched Shutdown

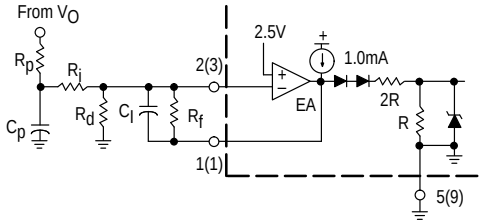


The MCR101 SCR must be selected for a holding of less than 0.5 mA at $T_{A(min)}$. The simple two transistor circuit can be used in place of the SCR as shown. All resistors are 10 k.

Figure 31. Error Amplifier Compensation

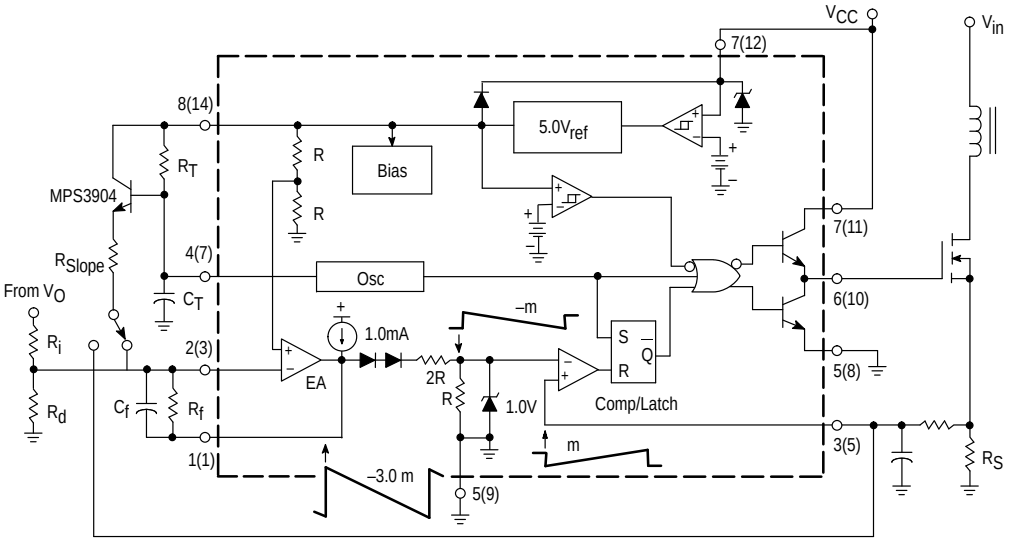


Error Amp compensation circuit for stabilizing any current–mode topology except for boost and flyback converters operating with continuous inductor current.



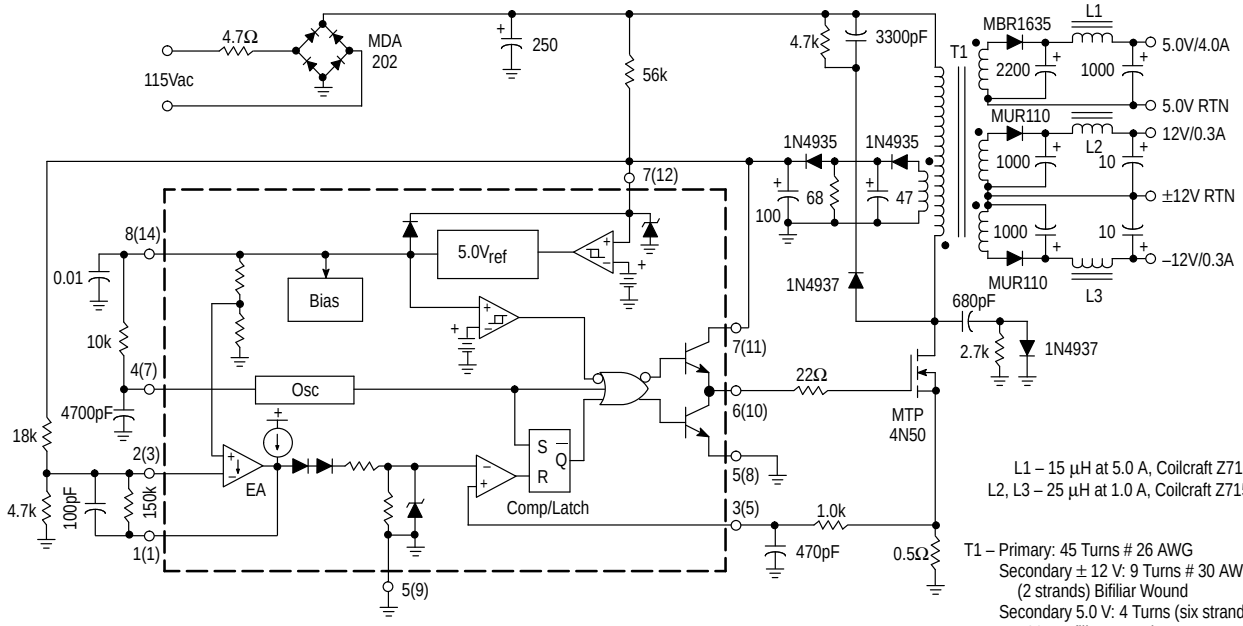
Error Amp compensation circuit for stabilizing current–mode boost and flyback topologies operating with continuous inductor current.

Figure 32. Slope Compensation



The buffered oscillator ramp can be resistively summed with either the voltage feedback or current sense inputs to provide slope compensation.

Figure 33. 27 Watt Off–Line Flyback Regulator



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

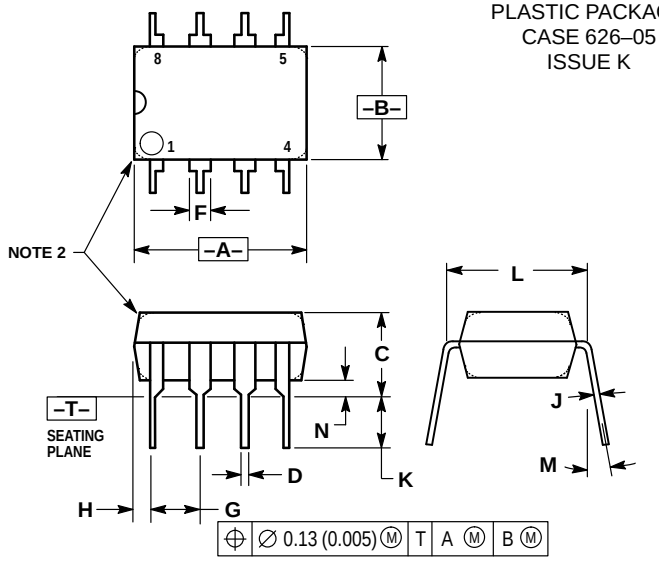
T1 – Primary: 45 Turns # 26 AWG
Secondary ± 12 V: 9 Turns # 30 AWG
(2 strands) Bifiliar Wound
Secondary 5.0 V: 4 Turns (six strands)
#26 Hexfiliar Wound
Secondary Feedback: 10 Turns #30 AWG
(2 strands) Bifiliar Wound
Core: Ferroxcube EC35–3C8
Bobbin: Ferroxcube EC35PCB1
Gap ≈ 0.01" for a primary inductance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

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

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

OUTLINE DIMENSIONS

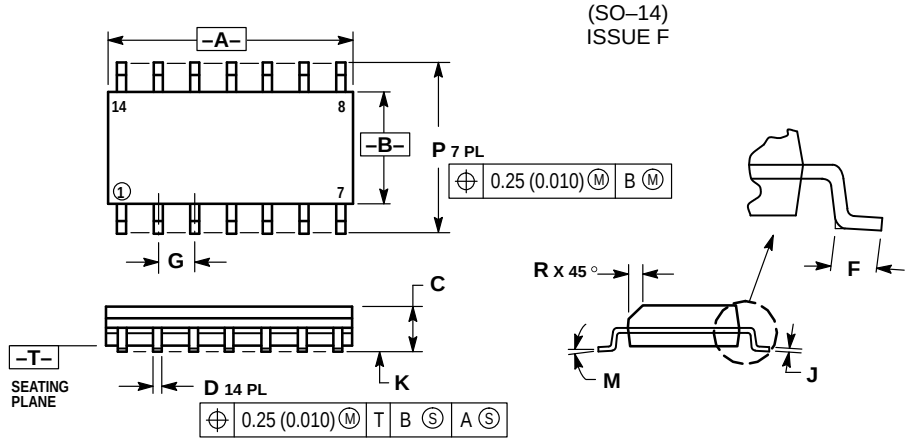
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PLASTIC PACKAGE
CASE 626–05
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2. PACKAGE CONTOUR OPTIONAL (ROUND OR SQUARE CORNERS).
3. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.

	MILLIMETERS		INCHES	
DIM	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

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



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.

	MILLIMETERS		INCHES	
DIM	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

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