TS3842B/3843B

High Performance Current Mode Controller



DIP-8



SOP-8

Designed for Off-Line and DC-to-DC converter applications.

General Description

The TS3842B and TS3843B series are high performance fixed frequency current mode controllers. This is specifically designed for Off-Line and DC-to-DC converter applications offering the designer a cost effective solution with minimal external components. This 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.

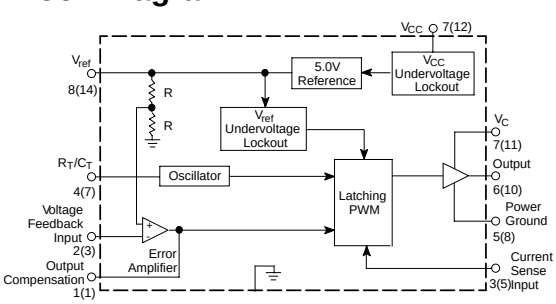
This device is available in 8-pin dual-in-line plastic packages as well as the 8-pin plastic surface mount (SOP-8). The SOP-8 package has separate power and ground pins for the totem pole output stage.

The TS3842B has UVLO thresholds of 16V (on) and 10V (off), ideally suited for off-line converters.

Features

- ✧ Trimmed Oscillator Discharge Current for Precise Duty Cycle Control
- ✧ Current Mode Operation to 500KHz
- ✧ 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 Hystersis
- ✧ Low Start-Up and Operating Current

Block Diagram



The document contains information on a new product.Specifications and information herein are subject to change without notice.

Ordering Information

DEVICE	OPERATING TEMPERATURE (Ambient)	PACKAGE
TS3842/3843BCD	-20 to +85℃	DIP-8
TS3842/3843BCS		SOP-8

Absolute 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)	Io	1.0	A
Output Energy (Capacitive Load per Cycle)	W	5.0	µJ
Current Sense and Voltage Feedback Inputs	Vin	-0.3 to +5.5	V
Error Amp Output Sink Current	Io	10	mA
Power Dissipation and Thermal Characteristics			
Plastic DIP			
Maximum Power Dissipation @ T_A=25℃	P_D	862	mW
Thermal Resistance Junction to Air	$R_{\theta JA}$	145	℃/W
Plastic SOP			
Maximum Power Dissipation @ T_A=25℃	P_D	1.25	W
Thermal Resistance Junction to Air	$R_{\theta JA}$	100	℃/W
Operating Junction Temperature	T_J	0 to +150	℃
Operating Ambient Temperature	T_A	-20 to +85	℃
Storage Temperature Range	Tstg	-25 to +150	℃

Electrical Characteristics

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

CHARACTERISTIC	SYMBOL	MIN	TYP	MAX	UNIT
REFFRENCE SECTION					
Reference Output Voltage (Io=1.0mA,T_J = 25℃)	Vref	4.9	5.0	5.1	V
Line Regulation (V_{CC} =12V to 25V)	Regline	-	2.0	20	mV
Load Regulation (Io =1.0mA to 20mA)	Regload	-	3.0	25	mV
Temperature Stability	Ts	-	0.2	-	mV/℃
Total Output Variation over Line,Load ,and Temperature	Vref	4.82	-	5.18	V
Output Noise Voltage (f = 10Hz to 10kHz, T_J=25℃)	Vn	-	50	-	μV
Long Term Stability (T_A=125℃ for 1000 Hours)	S	-	5.0	-	mV
Output Short Circuit Current	Isc	-30	-85	180	mA
OSCILLATOR SECTION					
Frequency					
T_J=25℃	Fosc	47	52	57	KHz
T_A=T_{low} to T_{high}		46	-	60	
Frequency Change with Voltage (V_{CC} =12V to 25V)	△fosc/△V	-	0.2	1.0	%
Frequency Change with Temperature T_A=T_{low} to T_{high}	△fosc/△T	-	5.0	-	%
Oscillator Voltage Swing (Peak-to-Peak)	Vosc	-	1.6	-	V
Discharge Current (Vosc=2.0V)					
T_J=25℃	Idischg	7.5	8.4	9.3	mA
T_A=T_{low} to T_{high}		7.2	-	9.5	

Electrical Characteristics

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

CHARACTERISTIC	SYMBOL	MIN	TYP	MAX	UNIT
ERROR AMPLIFIER SECTION					
Voltage Feedback Input (Vo=2.5V)	V_{FB}	2.42	2.5	2.58	V
Input Bias Current (V_{FB}=5.0V)	I_{IB}	-	-0.1	-2.0	μA
Open-Loop Voltage Gain (Vo=2.0V to 4.0V)	A_{VOL}	65	90	-	dB
Unity Gain Bandwidth (T_J=25℃)	BW	0.7	1.0	-	MHz
Power Supply Rejection Radio (V_{CC}=12V to 25V)	PSRR	60	70	-	dB
Output Current					
Sink (Vo=1.1V, V_{FB}=2.7V)	I_{sink}	2.0	12	-	mA
Source (Vo=5.0V, V_{FB}=2.3V)	I_{Source}	-0.5	-1.0	-	
Output Voltage Swing					
High State (R_L=15K to ground, V_{FB}=2.3V)	V_{OH}	5.0	6.2	-	V
Low State (R_L=15K to Vref, V_{FB}=2.7V)	V_{OL}	-	0.8	1.1	
CURRENT SENSE SECTION					
Current Sense Input Voltage Gain (Note 4&5)	Av	2.85	3.0	3.15	V/V
Maximum Current Sense Input Threshold(Note 4)	V_{th}	0.9	1.0	1.1	V
Power Supply Rejection Radio V_{CC}=12V to 25V,Note 4	PSRR	-	70	-	dB
Input Bias Current	I_{IB}	-	-2.0	-10	μA
Propagation Delay(Current Sense Input to Output)	$t_{PLH(IN/OUT)}$	-	150	300	ns

Electrical Characteristics

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

CHARACTERISTIC	SYMBOL	MIN	TYP	MAX	UNIT
OUTPUT SECTION					
Output Voltage					
Low State (Isink=20mA)	V_{OL}	-	0.1	0.4	
(Isink=200mA)		-	1.6	2.2	V
High State (Isource=20mA)	V_{OH}	13	13.5	-	
(Isource=200mA)		12	13.4	-	
Output Voltage with UVLO Activated V_{CC}=6.0V,Isink=1.0mA	V_{OL}(UVLO)	-	0.1	1.1	V
Output Voltage Rise Time (C_L=1.0nF,T_J=25℃)	tr	-	50	150	ns
Output Voltage Fall Time (C_L=1.0nF,T_J=25℃)	tf	-	50	150	ns
UNDERVOLTAGE LOCKOUT SECTION					
Start-Up Threshold					
TS3842B	Vth	14.5	16	17.5	V
TS3843B		7.8	8.4	9.0	
Minimum Operating Voltage After Turn-On					
TS3842B	$V_{CC(min)}$	8.5	10	11.5	V
TS3843B		7.0	7.6	8.2	
PWM SECTION					
Duty Cycle					
Maximum	DCmax	94	96	-	%
Minimum	DCmin	-	-	0	
TOTAL DEVICE					
Power Supply Current					
Start-Up, V_{CC}= 14V	I_{CC}	-	0.25	0.5	mA
Operating (Note 2)		-	12	17	
Power Supply Zener Voltage (I_{CC}=25mA)	Vz	30	36	-	V

Note: 1. Maximum package power dissipation limits must be observed.
2. Adjust V_{cc} above the Start-Up threshold before setting to 15V.
3. Low duty cycle pulse technique are used during test to maintain junction temperature as close to ambient as possible.
T_{low} = -20℃ ,T_{high} = +85℃
4. This parameter is measured at the latch trip point with V_{FB} = 0V.
5. Comparator gain is defined as : $Av = \frac{\Delta V \text{ Output Compensation}}{\Delta V \text{ Current Sense Input}}$

Figure 1 - Timing Resistor vs. Oscillator Frequency

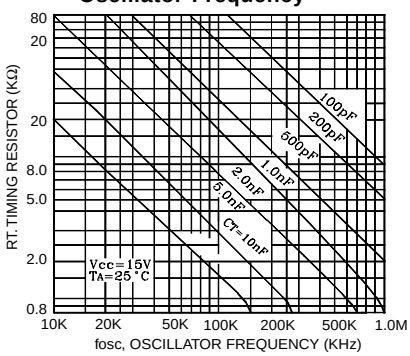


Figure 2 - Output Dead Time vs. Oscillator Frequency

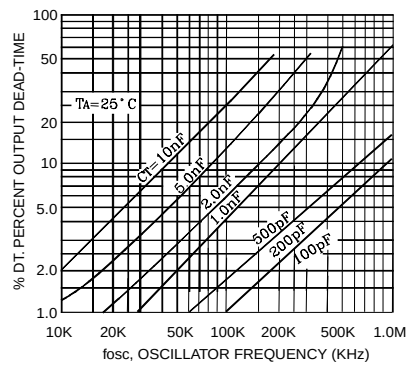


Figure 3 - Oscillator Discharge Current vs. Temperature

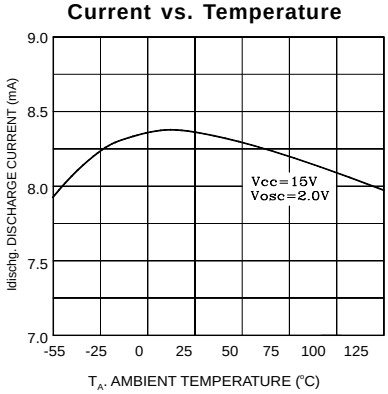


Figure 4 - Maximum Output Duty Cycle vs. Timing Resistor

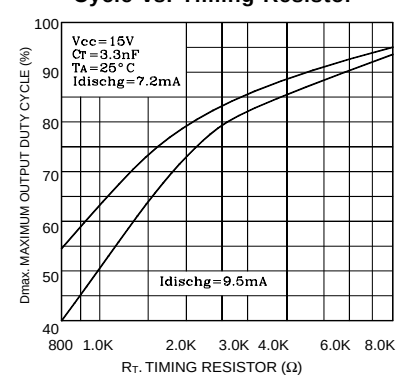


Figure 5 - Error Amp Small Signal Transient Response

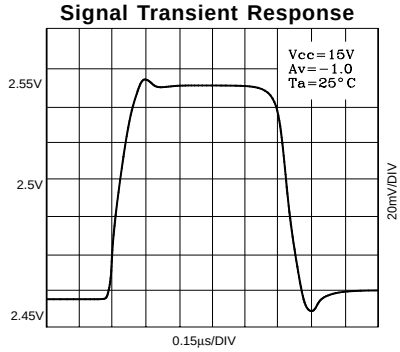


Figure 6 - Error Amp Large Signal Transient Response

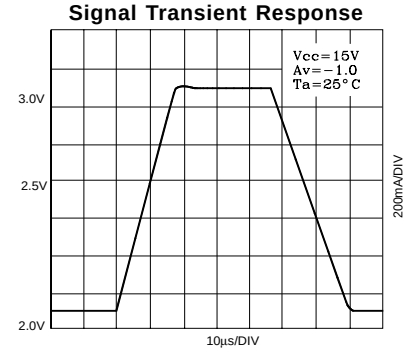


Figure 7 - Error Amp Open-Loop Gain and Phase vs. Frequency

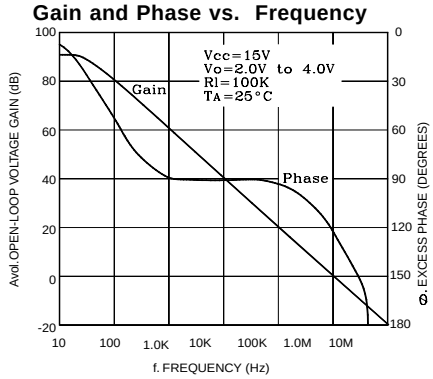


Figure 8 - Current Sense Input Threshold vs. Error Amp Output voltage

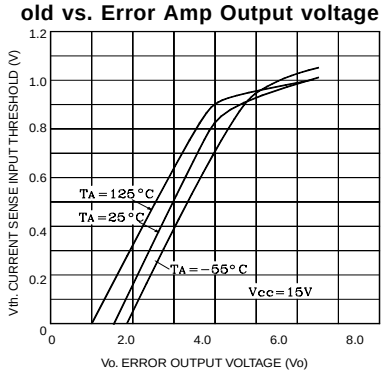


Figure 9 - Reference Voltage Change vs. Source Current

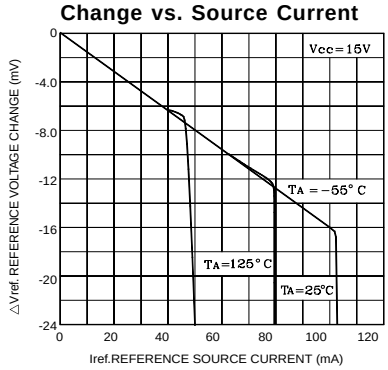


Figure 10 - Reference Short Circuit Current vs. Temperature

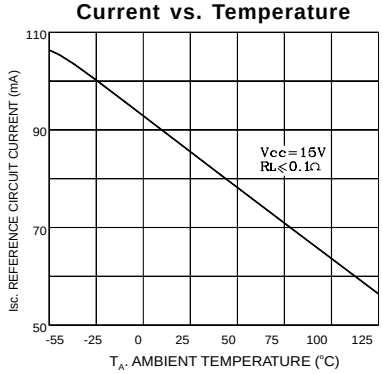


Figure 11 - Reference Load Regulation

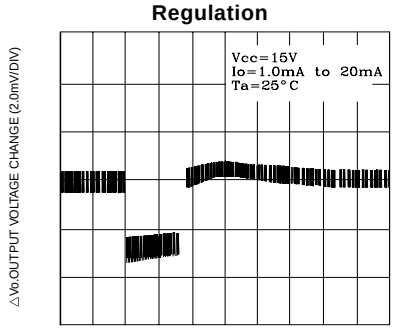


Figure 12 - Reference Line Regulation

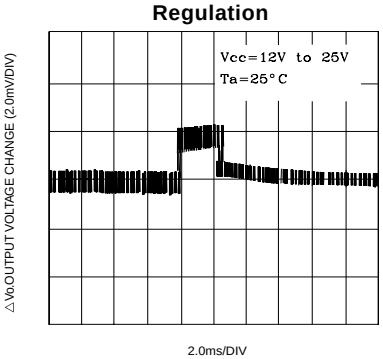


Figure 13 - Output Saturation Voltage vs. Load Current

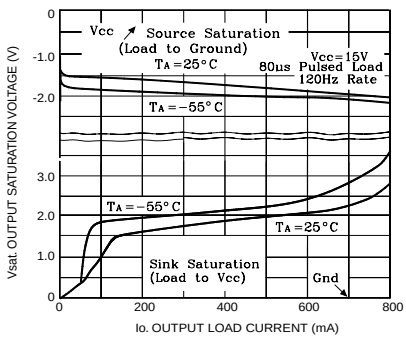


Figure 14 - Output Waveform

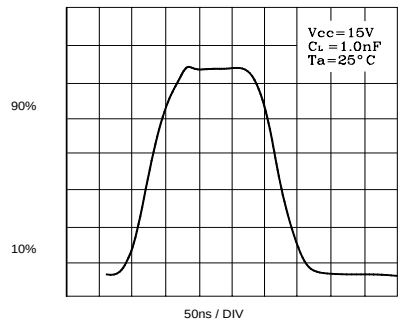


Figure 15 - Output Cross Conduction

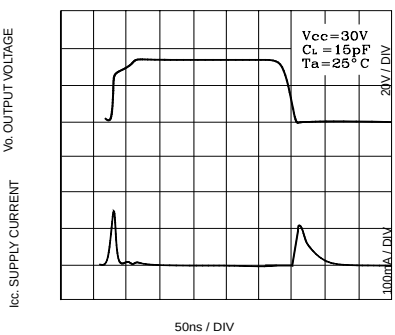


Figure 16 - Supply Current vs. Supply Voltage

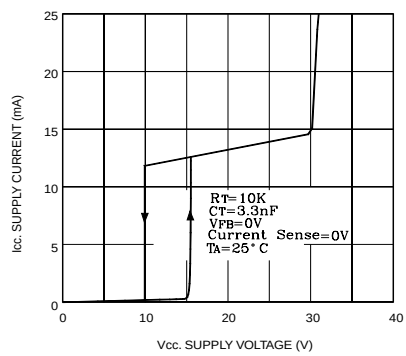


Figure 17- Representative Block Diagram

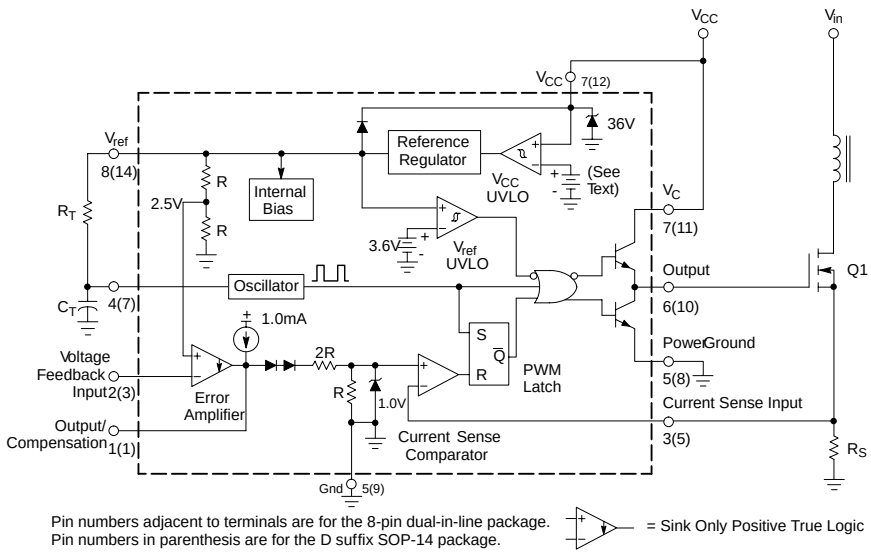
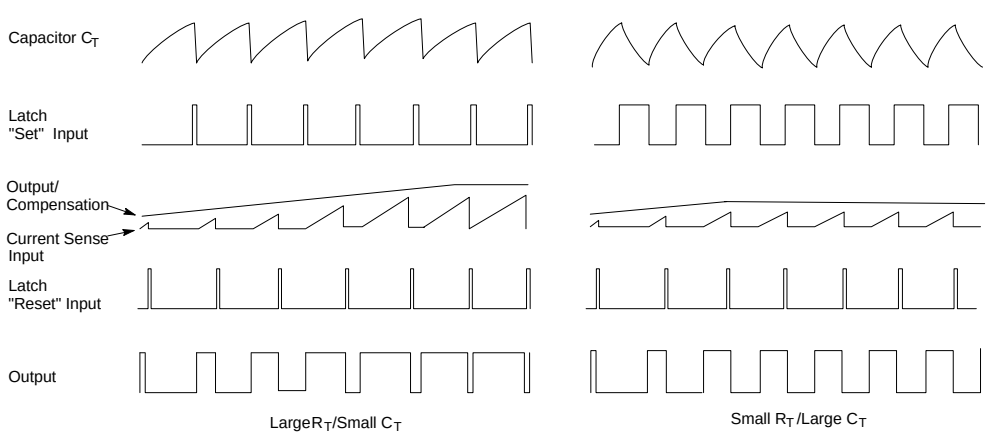


Figure 18 - Timing Diagram



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 (Vcc) and the reference output (Vref) are each monitored by separate comparators. Each has built-in hysteresis to prevent erratic output behavior as their respective thresholds are crossed. The large hysteresis and low start-up current of the TS3842B makes it ideally suited in off-line converter applications where efficient bootstrap start-up technique (Figure 33). 36V zener is connected as a shunt regulator from Vcc to ground. Its purpose is to protect the IC from excessive voltage that can occur during system start-up. The minimum operating voltage for the TS3842B is 11V.

Output

These devices contain a single totem pole output stage that was specifically designed for direct drive of power MOSFET's. It is capable of up to ±1.0A peak drive current and has a typical rise and fall time of 50 ns with a 1.0nF 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 SOP-8 surface mount package provides separate pins for Vc(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 Ipk(max) clamp level. The separate Vc supply input allows the designer added flexibility in tailling the drive voltage independent of Vcc. A zener clamp is typically connected to this input when driving power MOSFETs in systems where Vcc is greater than 20V. Figure 25 shows proper power and control ground connections in a current sensing power MOSFET application.

Reference

The 5.0V bandgap reference is trimmed to ±2.0% on the TS3842B. Its primary purpose to supply charging current to the oscillator timing capacitor. The reference has short circuit protection and is capable of providing in excess of 20mA for powering additional control system circuitry.

Design Considerations

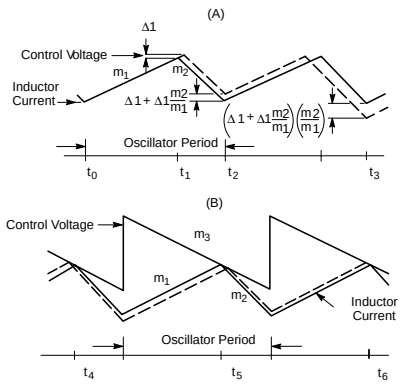
Do not attempt to construct the converter on wire wrap or plug-in prototype boards.

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 separate paths back to the input filter capacitor. Ceramic bypass capacitors (0.1μF) connected directly to Vcc,Vc, and Vref may be required depending upon circuit layout.

Undervoltage Lockout(contd.)

This provides a low impedance path for filtering the high frequency noised. 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.

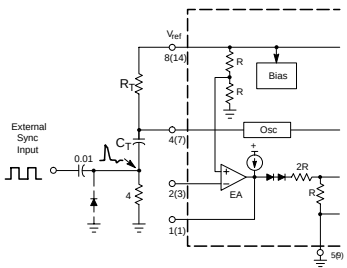
Figure 19 - Continuous Current Waveforms



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 19(A) 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 t1, 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. This unstable condition can be shown if a perturbation is added to the control voltage, resulting in a small $\Delta 1$(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 $\Delta 1+\Delta 1 m_2/m_1$. The minimum current at the next cycle (t_3) decreases to $(\Delta 1+ \Delta 1 m_2/m_1)(m_2/m_1)$. This perturbation 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 19(B) shows that by adding an artificial ramp that is synchronized with the PWM clock to the control voltage, the $\Delta 1$ perturbation will decrease to zero on succeeding cycles. This compensating 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).

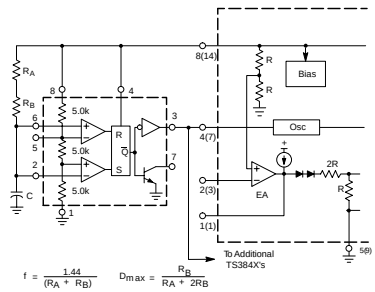
Undervoltage Lockout(contd.)

Figure 20 - External Clock Synchronization



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

Figure 21 - External Duty Cycle Clamp and Multi Unit Synchronization



Firuge 22-Adjustable Reduction of Clamp Level

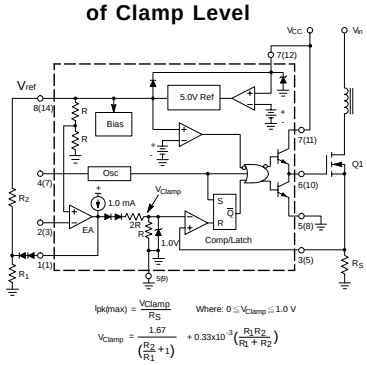
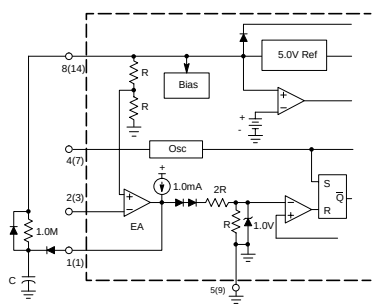


Figure 23- Soft-Start Circuit



$I_{Soft\text{-}Start}$=3600c in μF

Figure 24- Adjustable Buffered Reduction of Clamp Level with Soft-Start

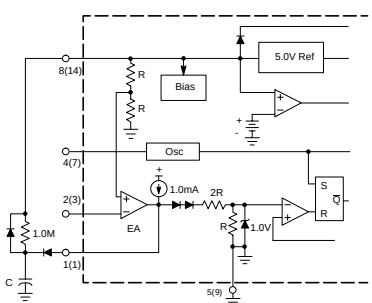
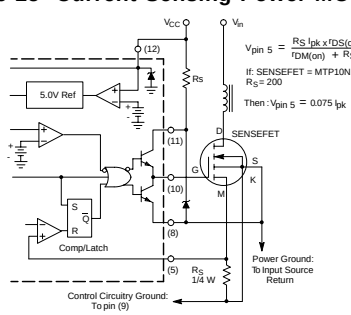


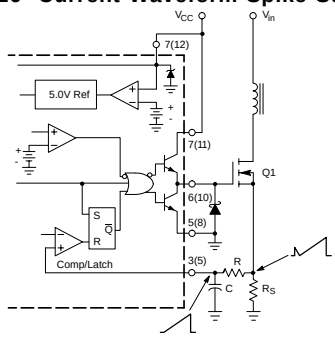
Figure 25- Current Sensing Power MOSFET



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 Ipk(max) clamp level must be implemented. Refer to Figure 22 and 24.

Undervoltage Lockout(contd.)

Figure 26- Current Waveform Spike Suppression



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

Figure 27- MOSFET Parasitic Oscillations

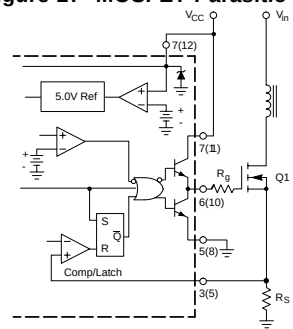
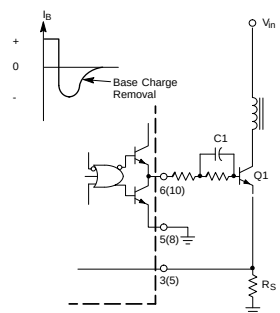


Figure 28- Bipolar Transistor Drive



The totem-pole output can finish negative base current for enhanced transistor turn-off with the additions of capacitor C1.

Figure 29- Isolated MOSFET Drive

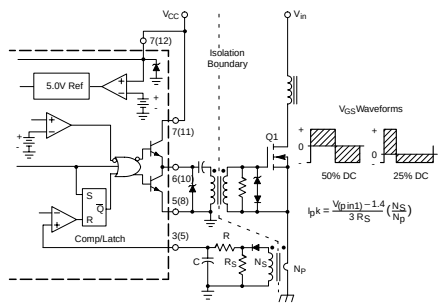
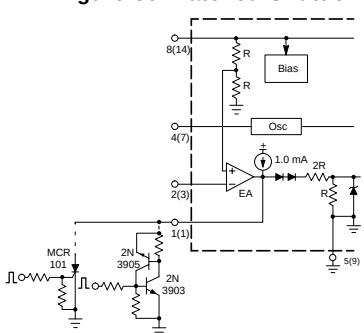


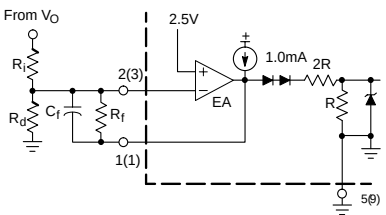
Figure 30- Latched Shutdown



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

Undervoltage Lockout(contd.)

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.

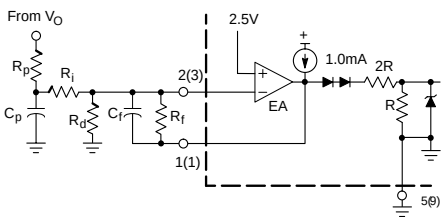
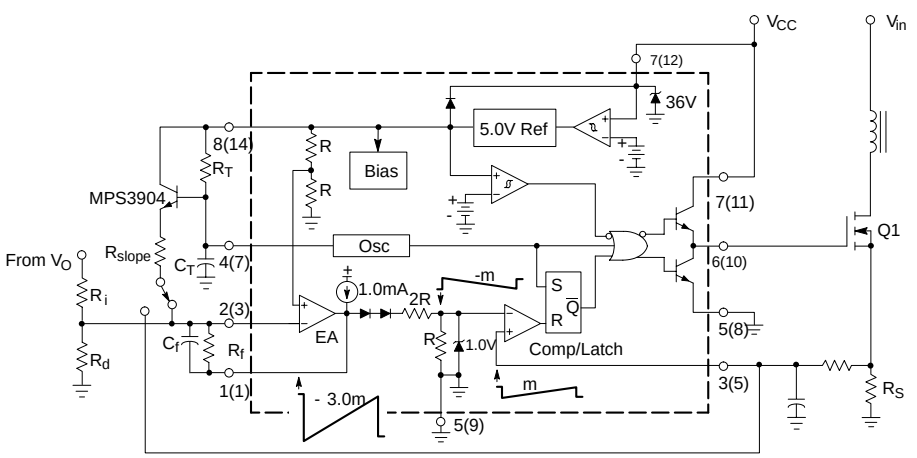


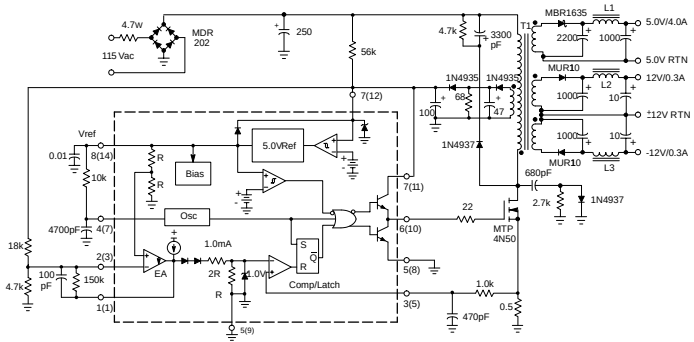
Figure 32-Slope Compenstaion



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

Undervoltage Lockout(contd.)

Figure 33-27 Watt Off-Line Regulation



TEST	CONDITIONS	RESULTS
Line Regulation: 5.0V	Vin=95 to 130 Vac	=50mV or ±0.5%
±12V		=24mV or ±0.1%
Load Regulation: 5.0V	Vin=115Vac, Iout =1.0A to 4.0A	=300mV or ±3.0%
±12V	Vin=115Vac,Iout=100mA to 300mA	=60mV or ±0.25%
Output Ripple: 5.0V	Vin=115Vac	40mVp-p
±12V		80 Vp-p
Efficiency	Vin=115Vac	70%

All outputs are at nominal load currents unless otherwise noted.

T1

Primary : 45 Turns #26 AWG

Secondary ±12V :
9 Turns #30 AWG (2 strands) Bifiliar Wound.

Secondary 5.0V :
4 Turns (six strands) #26 Hexfiliar Wound.

Secondary Feedback :
10 Turns #30 AWG (2 strands) Bifiliar Wound.

Core :
Ferroxcube EC35-3C8

Bobbin :
Ferroxcube EC35PCB1

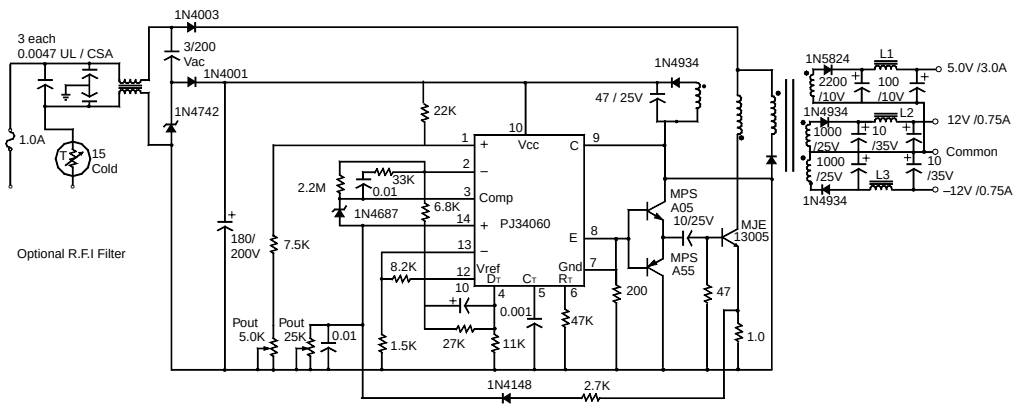
Gap :
≅@0.10” for a primary inductance of 1.0mH.

L1:
15μH at 5.0A, Coilcraft 27156.

L2,L3:
25μH at 1.0A, Coilcraft 27157.

Undervoltage Lockout(contd.)

Figure 34-33 Watt Off-Line Flyback Converter with Soft-Start and Primary Power Limiting



TEST	CONDITIONS	RESULTS
Line Regulation 5.0V	Vin=95 to 135 Vac, Io=3.0A	20mV 0.40%
Line Regulation± 12V	Vin=95 to 135 Vac, Io=±0.75A	52mV 0.26%
Line Regulation 5.0V	Vin=115 Vac, Io=1.0 to 4.0A	476mV 9.5%
Line Regulation± 12V	Vin=115 Vac, Io=±0.4 to ±0.9A	300mV 2.5%
Line Regulation 5.0V	Vin=115 Vac, Io=3.0A	45 mVp-p P.A.R.D.
Line Regulation± 12V	Vin=115 Vac, Io=±0.75A	75 mV p-p P.A.R.D.
Efficiency	Vin=115 Vac, Io(5.0V)=3.0A Io(±12V)=±0.75A	74%

T1

Coilcraft 11-464-16, 0.025" gap in each leg

Bobbin :

Coilcraft 37-573

Windings:

Primary, 2 each:

75 turns #26 Awg Bifilar wound

Feedback:

15 turns #26 Awg

Secondary , 5.0V:

6 turns #22 Awg Bifilar wound

Secondary , 5.0V:

14 turns #24 Awg Bifilar wound

L1

Coilcraft Z7156. 15μF @ 5.0A

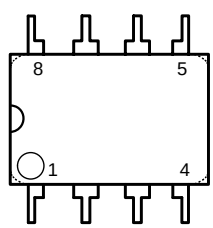
L2,L3

Coilcraft Z7157. 25μF @ 1.0A

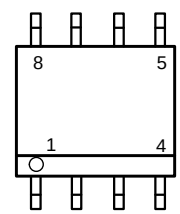
Pin Function Description

PIN NO.	FUNCTION	DESCRIPTION
1	Compensation	This pin is the Error Amplifier output and is made available for loop compensation
2	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	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	R_T/C_T	The Oscillator Frequency and maximum Output duty are programmed by connecting resistor R_T to Vref and capacitor C_T to ground operation to 500kHz is possible
5	Gnd	This pin is the combined control circuitry and power ground (8-pin package only).
6	Output	This output directly drives the gate of a power MOSFET. Peak current up to 1.0A are sourced and sunk by this pin.
7	Vcc	This pin is the positive supply of the control IC.
8	Vref	This pin is the reference output. It provides charging current for capacitor C_T through resistor R_T.

DIP-8



SOP-8



PIN : 1. Compensation
2. Voltage feedback
3. Current Sense
4. R_T / C_T
5. Gnd
6. Output
7. V_{CC}
8. V_{ref}

Operating Description

The TS3842B 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 5.0V reference through resistor R_T to approximately 2.8V and discharge to 1.2V 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. Figure 1 shows R_T versus Oscillator Frequency and Figure 2, Output Deadtime versus Oscillator 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℃. These internal circuit refines minimum variations of oscillator frequency and maximum output duty cycle. The results are shown in Figure 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 tailling 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 90dB, and a unity gain bandwidth of 1.0MHz with 57 degrees of phase margin (Figure 7). The non-inverting input is internally biased at 2.5V 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 -2.0μA 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 provided for external loop compensation (Figure 31). The output voltage is offset by two diode drops (≈1.4V) 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 (Figure 23,24). The Error Amp minimum feedback resistance is limited by the amplifier's source current (0.5mA) and the required output voltage (V_{OH}) to reach the comparator's 1.0V clamp level:

$$R_{f(MIN)} = [3x(1.0V)+1.4V] / 0.5mA = 8800\Omega$$

Current Sense Comparator and PWM Latch

The TS3842B 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 (Pin 1). 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 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 to 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:

$$I_{PK} = [V(Pin\ 1) - 1.4V] / 3R_S$$

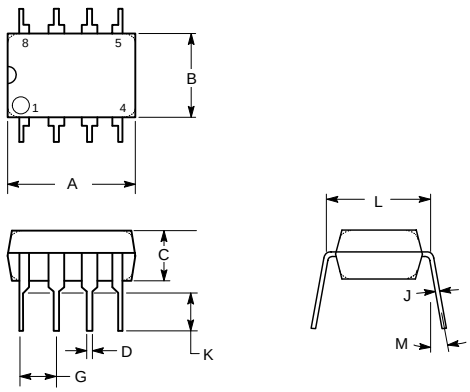
Abnormal operating conditions occur when the power supply output is overloaded or if output voltage sensing is lost. Under these conditions, the Current Sense Comparator threshold will be internally clamped to 1.0V. Therefore the maximum peak switch current is:

$$I_{PK\ (MAX)} = 1.0V / R_S$$

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.

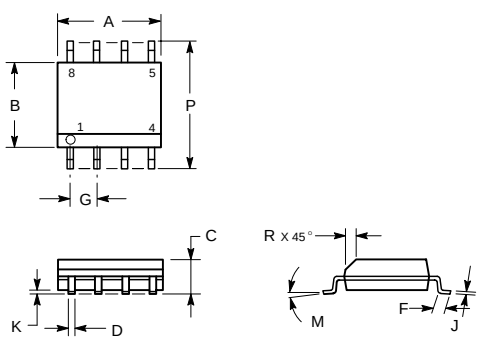
A narrow spike on the leading edge of the current waveform can usually be observed and may cause the power supply to exhibit an instability when the output is lightly loaded. This spike is due to the power transformer interwinding capacitance and output rectifier recovery time. The addition of an RC filter on the Current Sense Input with a time constant that approximates the spike duration will usually eliminate the instability: refer to Figure 26.

DIP-8



SYMBOLS	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	9.07	9.32	0.357	0.367
B	6.22	6.48	0.245	0.255
C	3.18	4.43	0.125	0.135
D	0.35	0.55	0.019	0.020
G	2.54BSC		0.10BSC	
J	0.29	0.31	0.011	0.012
K	3.25	3.35	0.128	0.132
L	7.75	8.00	0.305	0.315
M	-	10°	-	10°

SOP-8



SYMBOLS	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	4.80	5.00	0.189	0.196
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	1.27BSC		0.05BSC	
K	0.10	0.25	0.004	0.009
M	0°	7°	0°	7°
P	5.80	6.20	0.229	0.244
R	0.25	0.50	0.010	0.019