

SEPIC Constant-Current/ Constant-Voltage Battery Charger

FEATURES

- **Charger Input Voltage May Be Higher, Equal to or Lower Than Battery Voltage**
- Charges Any Number of Cells Up to 20V
- 1% Voltage Accuracy for Rechargeable Lithium Batteries
- 100mV Current Sense Voltage for High Efficiency
- **Battery Can Be Directly Grounded**
- **500kHz Switching Frequency Minimizes Inductor Size**
- Charging Current Easily Programmable or Shut Down

APPLICATIONS

- Charging of NiCd, NiMH, Lead-Acid or Lithium Rechargeable Cells
- Precision Current Limited Power Supply
- Constant-Voltage/Constant-Current Supply
- Transducer Excitation
- Universal Input CCFL Driver

DESCRIPTION

The LT[®]1513 is a 500kHz current mode switching regulator specially configured to create a constant-current/constant-voltage battery charger. In addition to the usual voltage feedback node, it has a current sense feedback circuit for accurately controlling output current of a fly-back or SEPIC (Single-Ended Primary Inductance Converter) topology charger. These topologies allow the current sense circuit to be ground referred and completely separated from the battery itself, simplifying battery switching and system grounding problems. In addition, these topologies allow charging even when the input voltage is lower than the battery voltage. The LT1513 can also drive a CCFL Royer converter with high efficiency in floating or grounded mode.

Maximum switch current on the LT1513 is 3A. This allows battery charging currents up to 2A for a single lithium-ion cell. Accuracy of 1% in constant-voltage mode is perfect for lithium battery applications. Charging current can be easily programmed for all battery types.

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TYPICAL APPLICATION

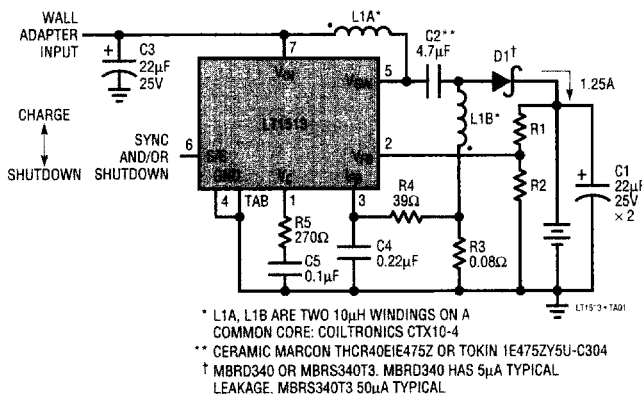
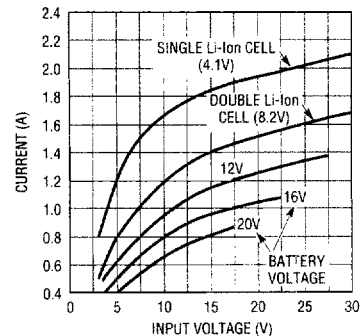


Figure 1. SEPIC Charger with 1.25A Output Current

Maximum Charging Current



INDUCTOR = 10µH
 ACTUAL PROGRAMMED CHARGING CURRENT WILL BE INDEPENDENT OF INPUT VOLTAGE IF IT DOES NOT EXCEED VALUES SHOWN

LT1513-1 TA02

ABSOLUTE MAXIMUM RATINGS

Supply Voltage	30V
Switch Voltage	40V
S/S Pin Voltage	30V
FB Pin Voltage (Transient, 10ms)	±10V
V _{FB} Pin Current	10mA
I _{FB} Pin Voltage (Transient, 10ms)	±10V
Storage Temperature Range	-65°C to 150°C
Ambient Temperature Range	
LT1513C	0°C to 70°C
LT1513I	-40°C to 85°C
Operating Junction Temperature Range	
LT1513C	0°C to 125°C
LT1513I	-40°C to 125°C
Short Circuit	0°C to 150°C
Lead Temperature (Soldering, 10 sec)	300°C

PACKAGE/ORDER INFORMATION

	<p>ORDER PART NUMBER</p> <p>LT1513CR LT1513IR</p>
	<p>Consult factory for Military grade parts.</p>

ELECTRICAL CHARACTERISTICS

V_{IN} = 5V, V_C = 0.6V, V_{FB} = V_{REF}, I_{FB} = 0V, V_{SW} and S/S pins open, unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS		
V _{REF}	FB Reference Voltage	Measured at FB Pin V _C = 0.8V	●	1.233	1.245	1.257	V	
			●	1.228	1.245	1.262	V	
	FB Input Current	V _{FB} = V _{REF}	●		300	550	nA	
					600	nA		
	FB Reference Voltage Line Regulation	2.7V ≤ V _{IN} ≤ 25V, V _C = 0.8V	●	0.01	0.03	%/V		
V _{IREF}	I _{FB} Reference Voltage	Measured at I _{FB} Pin V _{FB} = 0V, V _C = 0.8V	●	-107	-100	-93	mV	
			●	-110	-100	-90	mV	
	I _{FB} Input Current	V _{IREF} = V _{IREF} (Note 2)	●	10	25	35	μA	
	I _{FB} Reference Voltage Line Regulation	2.7V ≤ V _{IN} ≤ 25V, V _C = 0.8V	●	0.01	0.05	%/V		
g _m	Error Amplifier Transconductance	ΔI _C = ±25μA	●	1100	1500	1900	μmho	
			●	700		2300	μmho	
	Error Amplifier Source Current	V _{FB} = V _{REF} - 150mV, V _C = 1.5V	●	120	200	350	μA	
	Error Amplifier Sink Current	V _{FB} = V _{REF} + 150mV, V _C = 1.5V	●		1400	2400	μA	
	Error Amplifier Clamp Voltage	High Clamp, V _{FB} = 1V Low Clamp, V _{FB} = 1.5V		1.70	1.95	2.30	V	
				0.25	0.40	0.52	V	
A _v	Error Amplifier Voltage Gain			500		V/V		
	V _C Pin Threshold	Duty Cycle = 0%		0.8	1	1.25	V	
f	Switching Frequency	2.7V ≤ V _{IN} ≤ 25V 0°C ≤ T _J ≤ 125°C T _J < 0°C		450	500	550	kHz	
					430	500	580	kHz
					400		580	kHz
	Maximum Switch Duty Cycle		●	85	95	%		
	Switch Current Limit Blanking Time			130	260	ns		
BV	Output Switch Breakdown Voltage	0°C ≤ T _J ≤ 125°C T _J < 0°C		40	47		V	
					35		V	
V _{SAT}	Output Switch ON Resistance	I _{SW} = 2A	●	0.25	0.45	Ω		
I _{LIM}	Switch Current Limit	Duty Cycle = 50%	●	3.0	3.8	5.4	A	
		Duty Cycle = 80% (Note 1)	●	2.6	3.4	5.0	A	
ΔI _N /ΔI _{SW}	Supply Current Increase During Switch ON Time			15	25	mA/A		

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ELECTRICAL CHARACTERISTICS

$V_{IN} = 5V$, $V_C = 0.6V$, $V_{FB} = V_{REF}$, $I_{FB} = 0V$, V_{SW} and S/S pins open, unless otherwise noted.

SYMBOL	PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS	
	Control Voltage to Switch Current Transconductance			4		A/V	
	Minimum Input Voltage		●	2.4	2.7	V	
I_Q	Supply Current	$2.7V \leq V_{IN} \leq 25V$	●	4	5.5	mA	
	Shutdown Supply Current	$2.7V \leq V_{IN} \leq 25V$, $V_{S/S} \leq 0.6V$, $T_J \geq 0^\circ C$ $T_J < 0^\circ C$	●	12	30 50	μA μA	
	Shutdown Threshold	$2.7V \leq V_{IN} \leq 25V$	●	0.6	1.3	2	V
	Shutdown Delay		●	5	12	25	μs
	S/S Pin Input Current	$0V \leq V_{S/S} \leq 5V$	●	-10	15	μA	
	Synchronization Frequency Range		●	600	800	kHz	

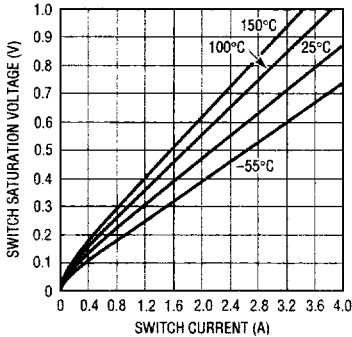
The ● denotes specifications which apply over the full operating temperature range.

Note 1: For duty cycles (DC) between 50% and 85%, minimum guaranteed switch current is given by $I_{LIM} = 1.33(2.75 - DC)$.

Note 2: The I_{FB} pin is servoed to its regulating state with $V_C = 0.8V$.

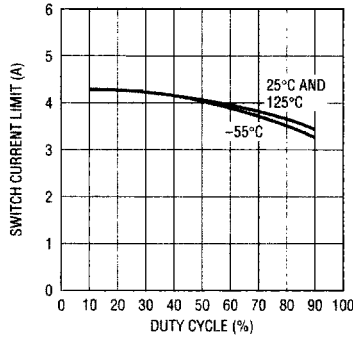
TYPICAL PERFORMANCE CHARACTERISTICS

Switch Saturation Voltage vs Switch Current



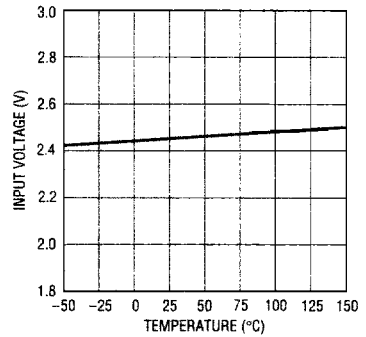
LT1513-601

Switch Current Limit vs Duty Cycle



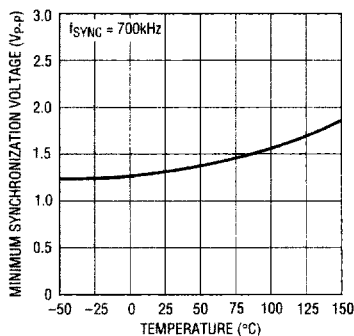
LT1513-602

Minimum Input Voltage vs Temperature



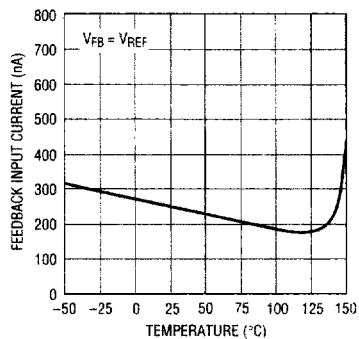
LT1513-603

Minimum Peak-to-Peak Synchronization Voltage vs Temperature



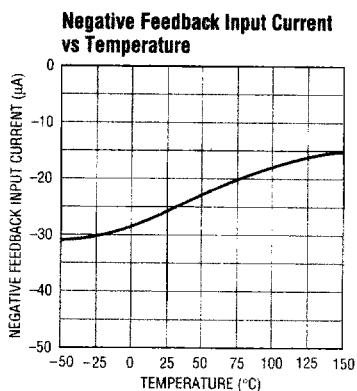
LT1513-604

Feedback Input Current vs Temperature

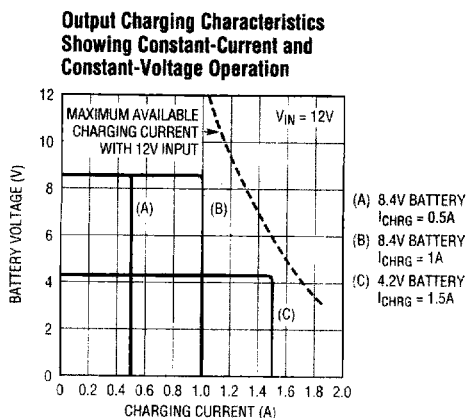


LT1513-605

TYPICAL PERFORMANCE CHARACTERISTICS



LT1513-026



15-3 907

PIN FUNCTIONS

V_C (Pin 1): The compensation pin is primarily used for frequency compensation, but it can also be used for soft starting and current limiting. It is the output of the error amplifier and the input of the current comparator. Peak switch current increases from 0A to 3.6A as the V_C voltage varies from 1V to 1.9V. Current out of the V_C pin is about 200µA when the pin is externally clamped below the internal 1.9V clamp level. Loop frequency compensation is performed with a capacitor or series RC network from the V_C pin *directly to the ground pin* (avoid ground loops).

FB (Pin 2): The feedback pin is used for positive output voltage sensing. The R1/R2 voltage divider connected to FB defines Li-Ion float voltage at full charge, or acts as a voltage limiter for NiCd or NiMH applications. FB is the inverting input to the voltage error amplifier. Input bias current is typically 300nA, so divider current is normally set to 100µA to swamp out any output voltage errors due to bias current. The noninverting input of this amplifier is tied internally to a 1.245V reference. The grounded end of the output voltage divider should be connected directly to the LT1513 ground pin (avoid ground loops).

I_{FB} (Pin 3): The current feedback pin is used to sense charging current. It is the input to a current sense amplifier that controls charging current when the battery voltage is below a programmed limit. During constant-current operation, the I_{FB} pin regulates at -100mV. Input resis-

tance of this pin is 5kΩ, so filter resistance (R4, Figure 1) should be less than 50Ω. The 39Ω, 0.22µF filter shown in Figure 1 is used to convert the pulsating current in the sense resistor to a smooth DC current feedback signal.

GND (Pin 4): The ground pin is internally connected to the TAB and both must be connected directly to a ground plane. The TAB of the surface mount R package should be soldered directly to the plane. It is also important that the compensation network, the output voltage divider, the output capacitor and the input capacitor be connected directly to this ground plane.

V_{SW} (Pin 5): The switch pin is the collector of the power switch, carrying up to 3A of current with fast rise and fall times. Keep the traces on this pin as short as possible to minimize radiation and voltage spikes. In particular, the path in Figure 1 which includes SW to C2, D1, C1 and around to the LT1513 ground pin should be as short as possible to minimize voltage spikes at switch turn-off.

S/S (Pin 6): This pin can be used for shutdown and/or synchronization. It is logic level compatible, but can be tied to V_{IN} if desired. It defaults to a high ON state when floated. A logic low state will shut down the charger to a micropower state. Driving the S/S pin with a continuous logic signal of 600kHz to 800kHz will synchronize switching frequency to the external signal. Shutdown is avoided in this mode with an internal timer.

APPLICATIONS INFORMATION

The LT1513 is an IC battery charger chip specifically optimized to use the SEPIC converter topology. A complete charger schematic is shown in Figure 1. The SEPIC topology has unique advantages for battery charging. It will operate with input voltages above, equal to or below the battery voltage, has no path for battery discharge when turned off, and eliminates the snubber losses of flyback designs. It also has a current sense point that is ground referred and need not be connected directly to the battery. The two inductors shown are actually just two identical windings on one inductor core, although two separate inductors can be used.

A current sense voltage is generated with respect to ground across R3 in Figure 1. The average current through R3 is always identical to the current delivered to the battery. The LT1513 current limit loop will servo the voltage across R3 to -100mV when the battery voltage is below the voltage limit set by the output divider R1/R2. Constant-current charging is therefore set at $100\text{mV}/R3$. R4 and C4 filter the current signal to deliver a smooth feedback voltage to the I_{FB} pin. R1 and R2 form a divider for battery voltage sensing and set the battery float voltage. The suggested value for R2 is 12.4k. R1 is calculated from:

$$R1 = \frac{R2(V_{\text{BAT}} - 1.245)}{1.245 + R2(0.3\mu\text{A})}$$

V_{BAT} = battery float voltage
 $0.3\mu\text{A}$ = typical FB pin bias current

A value of 12.4k for R2 sets divider current at $100\mu\text{A}$. This is a constant drain on the battery when power to the charger is off. If this drain is too high, R2 can be increased to 41.2k, reducing divider current to $30\mu\text{A}$. This introduces an additional uncorrectable error to the constant voltage float mode of about $\pm 0.5\%$ as calculated by:

$$V_{\text{BAT}} \text{ Error} = \frac{\pm 0.15\mu\text{A}(R1)(R2)}{1.245(R1 + R2)}$$

$\pm 0.15\mu\text{A}$ = expected variation in FB bias current around the nominal $0.3\mu\text{A}$ typical value.

With $R2 = 41.2\text{k}$ and $R1 = 228\text{k}$, ($V_{\text{BAT}} = 8.2\text{V}$), the error due to variations in bias current would be $\pm 0.42\%$.

A second option is to disconnect the divider when charger power is off. This can be done with a small Nfet as shown in Figure 3. D2, C6 and R6 form a peak detector to drive the gate of the FET to about the same as the battery voltage. If power

is turned off, the gate will drop to 0V and the only drain on the battery will be the reverse leakage of the catch diode D1. See Diode Selection for a discussion of diode leakage.

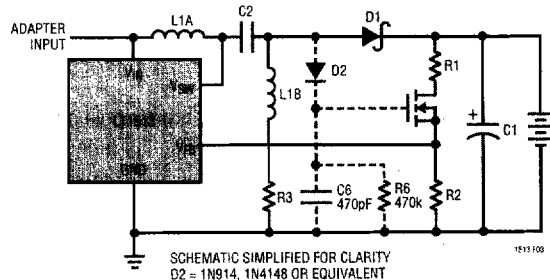


Figure 3. Eliminating Divider Current

Maximum Input Voltage

Maximum input voltage for the LT1513 is partly determined by battery voltage. A SEPIC converter has a maximum switch voltage equal to input voltage plus output voltage. The LT1513 has a maximum input voltage of 30V and a maximum switch voltage of 40V, so this limits maximum input voltage to 30V, or $40\text{V} - V_{\text{BAT}}$, whichever is less.

Shutdown and Synchronization

The dual function S/S pin provides easy shutdown and synchronization. It is logic level compatible and can be pulled high or left floating for normal operation. A logic low on the S/S pin activates shutdown, reducing input supply current to $12\mu\text{A}$. To synchronize switching, drive the S/S pin between 600kHz and 800kHz.

Inductor Selection

L1A and L1B are normally just two identical windings on one core, although two separate inductors can be used. A typical value is $10\mu\text{H}$, which gives about 0.5A peak-to-peak inductor current. Lower values will give higher ripple current, which reduces maximum charging current. $5\mu\text{H}$ can be used if charging currents are at least 20% lower than the values shown in the maximum charging current graph. Higher inductance values give slightly higher maximum charging current, but are larger and more expensive. A low loss toroid core such as Kool M μ , Molypermalloy or Metglas is

Kool M μ is a registered trademark of Magnetics, Inc.
 Metglas is a registered trademark of AlliedSignal Inc.

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recommended. Series resistance should be less than 0.04Ω for each winding. "Open core" inductors, such as rods or barrels are not recommended because they generate large magnetic fields which may interfere with other electronics close to the charger.

Input Capacitor

The SEPIC topology has relatively low input ripple current compared to other topologies and higher harmonics are especially low. RMS ripple current in the input capacitor is less than 0.25A with $L = 10\mu\text{H}$ and less than 0.5A with $L = 5\mu\text{H}$. A low ESR 22 μF , 25V solid tantalum capacitor (AVX type TPS or Sprague type 593D) is adequate for most applications with the following caveat. Solid tantalum capacitors can be destroyed with a very high turn-on surge current such as would be generated if a low impedance input source were "hot switched" to the charger input. If this condition can occur, the input capacitor should have the highest possible voltage rating, at least twice the surge input voltage if possible. Consult with the capacitor manufacturer before a final choice is made. A 4.7 μF ceramic capacitor such as the one used for the coupling capacitor can also be used. These capacitors do not have a turn-on surge limitation. The input capacitor must be connected directly to the V_{IN} pin and the ground plane close to the LT1513.

Output Capacitor

It is assumed as a worst case that all the switching output ripple current from the battery charger could flow in the output capacitor. This is a desirable situation if it is necessary to have very low switching ripple current in the battery itself. Ferrite beads or line chokes are often inserted in series with the battery leads to eliminate high frequency currents that could create EMI problems. This forces all the ripple current into the output capacitor. Total RMS current into the capacitor has a maximum value of about 1A, and this is handled with the two paralleled 22 μF , 25V capacitors shown in Figure 1. These are AVX type TPS or Sprague type 593D surface mount solid tantalum units intended for switching applications. Do not substitute other types without ensuring that they have adequate ripple current ratings. See Input Capacitor section for details of surge limitation on solid tantalum capacitors if the battery may be "hot switched" to the output of the charger.

Coupling Capacitor

C2 in Figure 1 is the coupling capacitor that allows a SEPIC converter topology to work with input voltages either higher or lower than the battery voltage. DC bias on the capacitor is equal to input voltage. RMS ripple current in the coupling capacitor has a maximum value of about 1A at full charging current. A conservative formula to calculate this is:

$$I_{\text{COUP(RMS)}} = \frac{I_{\text{CHRG}}(V_{\text{IN}} + V_{\text{BAT}})(1.1)}{2(V_{\text{IN}})}$$

(1.1 is a fudge factor to account for inductor ripple current and other losses)

With $I_{\text{CHRG}} = 1.2\text{A}$, $V_{\text{IN}} = 15\text{V}$ and $V_{\text{BAT}} = 8.2\text{V}$, $I_{\text{COUP}} = 1.02\text{A}$.

The recommended capacitor is a 4.7 μF ceramic type from Marcon or Tokin. These capacitors have extremely low ESR and high ripple current ratings in a small package. Solid tantalum units can be substituted if their ripple current rating is adequate, but typical values will increase to 22 μF or more to meet the ripple current requirements.

Diode Selection

The switching diode should be a Schottky type to minimize both forward and reverse recovery losses. Average diode current is the same as output charging current, so this will be under 2A. A 3A diode is recommended for most applications, although smaller devices could be used at reduced charging current. *Maximum diode reverse voltage will be equal to input voltage plus battery voltage.*

Diode reverse leakage current will be of some concern during charger shutdown. This leakage current is a direct drain on the battery when the charger is not powered. High current Schottky diodes have relatively high leakage currents (5 μA to 500 μA) even at room temperature. The latest very-low-forward devices have especially high leakage currents. It has been noted that surface mount versions of some Schottky diodes have as much as ten times the leakage of their through-hole counterparts. This may be because a low forward voltage process is used to reduce power dissipation in the surface mount package. In any case, check leakage specifications carefully before making a final choice for the switching diode. Be aware that diode manufacturers want to specify a maximum leakage current that is ten times higher than the typical leakage. It is very difficult to get them

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to specify a low leakage current in high volume production. This is an on going problem for all battery charger circuits and most customers have to settle for a diode whose typical leakage is adequate, but theoretically has a worst-case condition of higher than desired battery drain.

Thermal Considerations

Care should be taken to ensure that worst-case conditions do not cause excessive die temperatures. Typical thermal resistance is 30°C/W for the R package but this number will vary depending on the mounting technique (copper area, airflow, etc.).

Average supply current (including driver current) is:

$$I_{IN} = 4mA + \frac{(V_{BAT})(I_{CHRG})(0.024)}{V_{IN}}$$

Switch power dissipation is given by:

$$P_{SW} = \frac{(I_{CHRG})^2(R_{SW})(V_{BAT} + V_{IN})(V_{BAT})}{(V_{IN})^2}$$

R_{SW} = Output switch ON resistance

Total power dissipation of the die is equal to supply current times supply voltage, plus switch power:

$$P_{D(TOTAL)} = (I_{IN})(V_{IN}) + P_{SW}$$

For $V_{IN} = 10V$, $V_{BAT} = 8.2V$, $I_{CHRG} = 1.2A$, $R_{SW} = 0.3\Omega$,

$$I_{IN} = 4mA + 24mA = 28mA$$

$$P_{SW} = 0.64W$$

$$P_D = (10)(0.028) + 0.64 = 0.92W$$

Programmed Charging Current*

LT1513 charging current can be programmed with a PWM signal from a processor as shown in Figure 5. C6 and D2 form a peak detector that converts a positive logic signal to a negative signal. The average negative signal at the input to R5 is equal to the processor V_{CC} level multiplied by the inverse PWM ratio. This assumes that the PWM signal is a CMOS output that swings rail-to-rail with a source resistance less than a few hundred ohms. The negative voltage is converted to a current by R5 and R6 and filtered by C7. This current multiplied by R4 generates a voltage that subtracts from the 100mV sense voltage of the LT1513. This is not a high precision technique because of the errors in V_{CC} and the diode voltage, but it can typically be used to adjust charging current over a 20% to 100% range with good

* A new version of the LT1513 is available with accurate charging current directly proportional to PWM duty cycle. Contact factory.

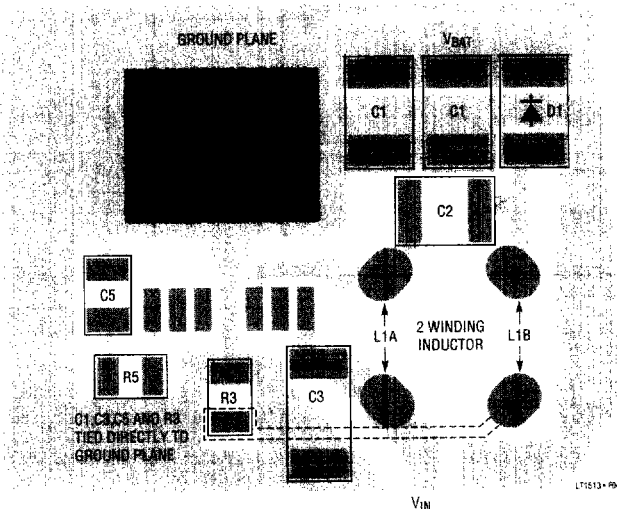


Figure 4. LT1513 Suggested Partial Layout for Critical Thermal and Electrical Paths

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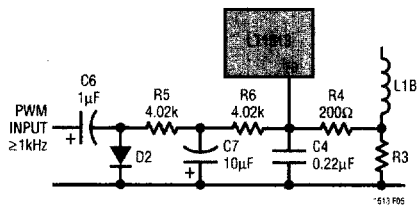


Figure 5

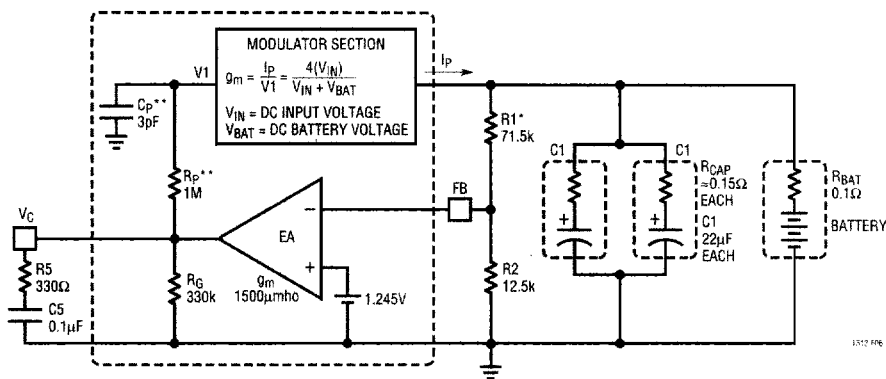
repeatability (full charging current accuracy is not affected). To reduce the load on the logic signal, R4 has been increased from 24Ω to 200Ω. This causes a known increase in full-scale charging current (PWM = 0) of 3% due to the 5k input resistance of the I_{FB} pin. Note that 100% duty cycle gives full charging current and that very low duty cycles (especially zero!) will not operate correctly. Very low duty cycle (<10%) is a problem because the peak detector requires a finite uptime to reset C6.

Voltage Mode Loop Stability

The LT1513 operates in constant-voltage mode during the final phase of charging lithium-ion and lead-acid batteries.

This feedback loop is stabilized with a series resistor and capacitor on the V_C pin of the chip. Figure 6 shows the simplified model for the voltage loop. The error amplifier is modeled as a transconductance stage with g_m = 1500µmho (from the Electrical Characteristics). Amplifier output resistance is modeled with a 330k resistor. The power stage (modulator section) of the LT1513 is modeled as a transconductance whose value is 4(V_{IN})/(V_{IN} + V_{BAT}). This is a very simplified model of the actual power stage, but it is sufficient when the unity-gain frequency of the loop is low compared to the switching frequency. The output filter capacitor model includes its ESR (R_{CAP}). A series resistance (R_{BAT}) is also assigned to the battery model.

Analysis of this loop normally shows an extremely stable system for all conditions, even with 0Ω for R5. The one condition which can cause reduced phase margin is with a very large battery resistance (>5Ω), or with the battery replaced with a resistive load. The addition of R5 gives good phase margin even under these unusual conditions. R5 should not be increased above 330Ω without checking for two possible problems. The first is instability in the constant current region (see Constant-Current Mode Loop Stability),



* FOR 8.4V BATTERY. ADJUST VALUE OF R1 FOR ACTUAL BATTERY VOLTAGE
 ** R_P AND C_P MODEL PHASE DELAY IN THE MODULATOR

THIS IS A SIMPLIFIED AC MODEL FOR THE LT1513 IN CONSTANT-VOLTAGE MODE. RESISTOR AND CAPACITOR NUMBERS CORRESPOND TO THOSE USED IN FIGURE 1. R_P AND C_P MODEL THE PHASE DELAY IN THE MODULATOR. C3 IS 3pF FOR A 10µH INDUCTOR. IT SHOULD BE SCALED PROPORTIONALLY FOR OTHER INDUCTOR VALUES (6pF FOR 20µH). THE MODULATOR IS A TRANSCONDUCTANCE WHOSE GAIN IS A FUNCTION OF INPUT AND BATTERY VOLTAGE AS SHOWN.

AS SHOWN, THIS LOOP HAS A UNITY-GAIN FREQUENCY OF ABOUT 250Hz. UNITY-GAIN WILL MOVE OUT TO SEVERAL KILOHERTZ IF BATTERY RESISTANCE INCREASES TO SEVERAL OHMS. R5 IS NOT USED IN ALL APPLICATIONS, BUT IT GIVES BETTER PHASE MARGIN IN CONSTANT-VOLTAGE MODE WITH HIGH BATTERY RESISTANCE.

Figure 6. Constant-Voltage Small-Signal Model

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and the second is subharmonic switching where switch duty cycle varies from cycle to cycle. This duty cycle instability is caused by excess switching frequency ripple voltage on the V_C pin. Normally this ripple is very low because of the filtering effect of C_5 , but large values of R_5 can allow high ripple on the V_C pin. Normal loop analysis does not show this problem, and indeed small signal loop stability can be excellent even in the presence of subharmonic switching. The primary issue with subharmonics is the presence of EMI at frequencies below 500kHz.

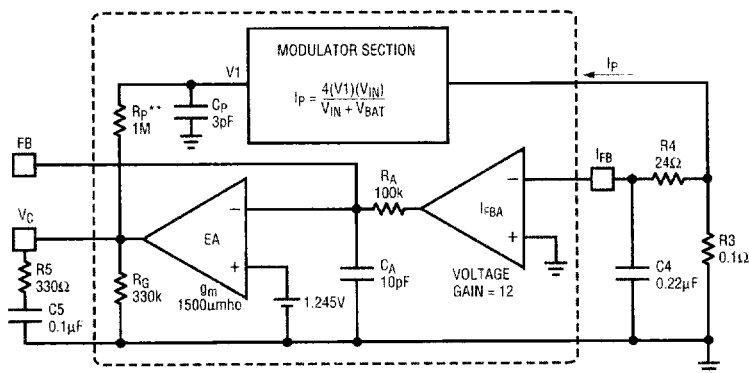
Constant-Current Mode Loop Stability

The LT1513 is normally very stable when operating in constant-current mode (see Figure 7), but there are certain conditions which may create instabilities. The combination of higher value current sense resistors (low programmed charging current), higher input voltages, and the addition of a loop compensation resistor (R_5) on the V_C pin may create an unstable current mode loop. (A resistor is sometimes added in series with C_5 to improve loop phase margin when the loop is operating in voltage mode.) Instability results

because loop gain is too high in the 50kHz to 150kHz region where excess phase occurs in the current sensing amplifier and the modulator. The I_{FBA} amplifier (gain of -12.5) has a pole at approximately 150kHz. The modulator section consisting of the current comparator, the power switch and the magnetics, has a pole at approximately 50kHz when the coupled inductor value is $10\mu\text{H}$. Higher inductance will reduce the pole frequency proportionally. The design procedure presented here is to roll off the loop to unity-gain at a frequency of 25kHz or lower to avoid these excess phase regions.

The suggested way to control unity loop frequency is to increase the filter time constant on the I_{FB} pin (R_4/C_4 in Figures 1 and 7). The filter resistor cannot be arbitrarily increased because high values will affect charging current accuracy. Charging current will increase by 1% for each 40Ω increase in R_4 . There is no inherent limitation on the value of C_4 , but if this capacitor is ceramic, it should be an X7R type to maintain its value over temperature. X7R dielectric requires a larger footprint.

4



THIS IS A SIMPLIFIED AC MODEL FOR THE LT1513 IN CONSTANT-CURRENT MODE. RESISTOR AND CAPACITOR NUMBERS CORRESPOND TO THOSE USED IN FIGURE 1. R_P AND C_P MODEL THE PHASE DELAY IN THE PowerPath. C_3 IS 3pF FOR A $10\mu\text{H}$ INDUCTOR. IT SHOULD BE SCALED PROPORTIONALLY FOR OTHER INDUCTOR VALUES (6pF FOR $20\mu\text{H}$). THE PowerPath IS A TRANSCONDUCTANCE WHOSE GAIN IS A FUNCTION OF INPUT AND BATTERY VOLTAGE AS SHOWN.

THE CURRENT AMPLIFIER HAS A FIXED VOLTAGE GAIN OF 12. ITS PHASE DELAY IS MODELED WITH R_A AND C_A .

THE ERROR AMPLIFIER HAS A TRANSCONDUCTANCE OF $1500\mu\text{mho}$ AND AN INTERNAL OUTPUT SHUNT RESISTANCE OF 330Ω .

AS SHOWN, THIS LOOP HAS A UNITY-GAIN FREQUENCY OF ABOUT 27kHz. R_5 IS NOT USED IN ALL APPLICATIONS, BUT IT GIVES BETTER PHASE MARGIN IN CONSTANT VOLTAGE MODE.

1517FB

Figure 7. Constant-Current Small-Signal Model

APPLICATIONS INFORMATION

The formula for calculating the minimum value for the filter capacitor C4 is:

$$C4 = \frac{(R3)(4)(V_{IN})(12)(1500\mu)(R5)}{2\pi(f)(R4)(V_{IN} + V_{BAT})}$$

V_{IN} = Highest input voltage

1500μ = Transconductance of error amplifier (EA)

f = Desired unity-gain frequency

V_{BAT} = Battery voltage

For example, assume $V_{IN(MAX)} = 15V$, $R3 = 0.4\Omega$ (charging current set to 0.25A), $R4 = 24\Omega$, $R5 = 330\Omega$ and $V_{BAT} = 8V$,

$$C4 = \frac{0.4(4)(15)(12)(0.0015)(330)}{6.3(25000)(39)(15 + 8)} = 1\mu F$$

The value for C4 could be reduced to a more manageable size by increasing R4 to 75Ω and reducing R5 to 300Ω , yielding $0.47\mu F$ for C4. The 2% increase in charging current can be ignored or factored into the value for R3.

More Help

Linear Technology Field Application Engineers have a CAD spreadsheet program for detailed calculations of circuit operating conditions. In addition, our Applications Department is always ready to lend a helping hand. The LT1371 data sheet may also be helpful. The LT1513 is identical except for the current amplifier circuitry.

TYPICAL APPLICATION

This Cold Cathode Fluorescent Lamp driver uses a Royer class self-oscillating sine wave converter to driver a high voltage lamp with an AC waveform. CCFL Royer converters have significantly degraded efficiency if they must operate at low input voltages, and this circuit was designed to handle input voltages as low as 2.7V. Therefore, the LT1513 is connected to generate a negative current through L2 that allows the Royer to operate as if it were connected to a constant higher voltage input.

The Royer output winding and the bulb are allowed to float in this circuit. This can yield significantly higher efficiency in situations where the stray bulb capacitance to surrounding enclosure is high. To regulate bulb current in Figure 8, Royer *input* current is sensed with R2 and filtered with R3 and C6. This negative feedback signal is applied to the I_{FB} pin of the LT1513. For more information on this circuit contact the LTC Applications Department and see Design Note 133. Considerable written application literature on Royer CCFL circuits is also available from other LTC Application and Design Notes.

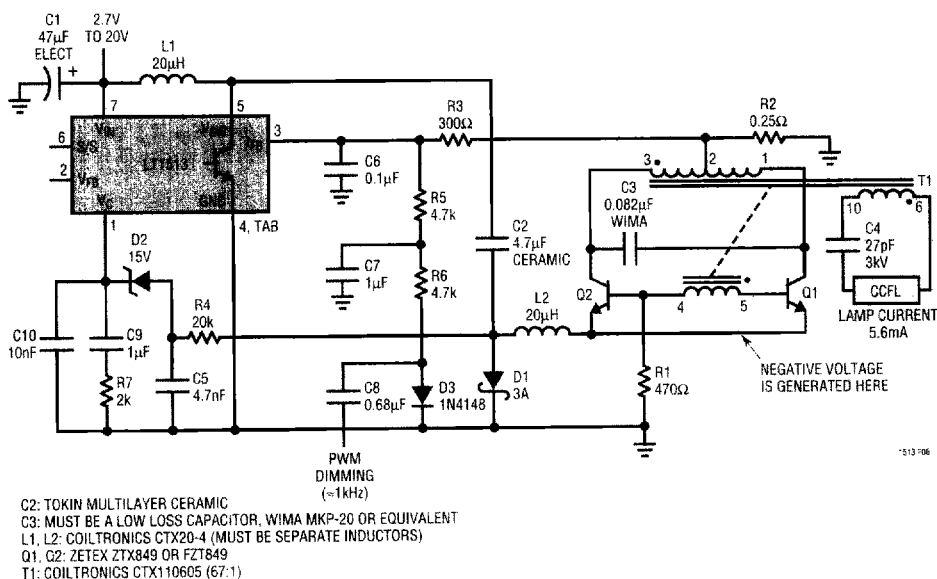


Figure 8. CCFL Power Supply for Floating Lamp Configuration Operates on 2.7V

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RELATED PARTS

PART NUMBER	DESCRIPTION	COMMENTS
LT1239	Backup Battery Management System	Charges Backup Battery and Regulates Backup Battery Output when Main Battery Removed
LTC [®] 1325	Microprocessor Controlled Battery Management System	Can Charge, Discharge and Gas Gauge NiCd, NiMH and Pb-Acid Batteries with Software Charging Profiles
LT1510	1.5A Constant-Current/Constant-Voltage Battery Charger	Step-Down Charger for Li-Ion, NiCd and NiMH
LT1511	3.0A Constant-Current/Constant-Voltage Battery Charger with Input Current Limiting	Step-Down Charger that Allows Charging During Computer Operation and Prevents Wall-Adapter Overload
LT1512	SEPIC Constant-Current/Constant-Voltage Battery Charger	Step-Up/Step-Down Charger for Up to 1A Current