NCP1582, NCP1582A, NCP1583

Low Voltage Synchronous Buck Controllers

The NCP158x is a low cost PWM controller designed to operate from a 5 V or 12 V supply. This device is capable of producing an output voltage as low as 0.8 V. This 8−pin device provides an optimal level of integration to reduce size and cost of the power supply. Features include a 0.7 A gate driver and an internally set 350 kHz (NCP1582, NCP1582A) and a 300 kHz (NCP1583) oscillator. The NCP158x also incorporates an externally compensated transconductance error amplifier and a programmable soft−start function. Protection features include short circuit protection (SCP) and under voltage lockout (UVLO). The NCP158x comes in an 8−pin SOIC package.

Features

- Input Voltage Range from 4.5 V to 13.2 V
- 350 kHz (NCP1582, NCP1582A), 300 kHz (NCP1583) Internal Oscillator
- Boost Pin Operates to 30 V
- Voltage Mode PWM Control
- 0.8 V ±1.5% Internal Reference Voltage
- Adjustable Output Voltage
- Programmable Soft−Start
- Internal 0.7 A Gate Drivers
- 80% Max Duty Cycle
- Input UVLO
- $R_{DS(on)}$ Current Sensing for Short Circuit Protection
- These are Pb−Free Devices

Applications

- Graphics Cards
- Desktop Computers
- Servers/Networking
- DSP and FPGA Power Supply
- DC−DC Regulator Modules

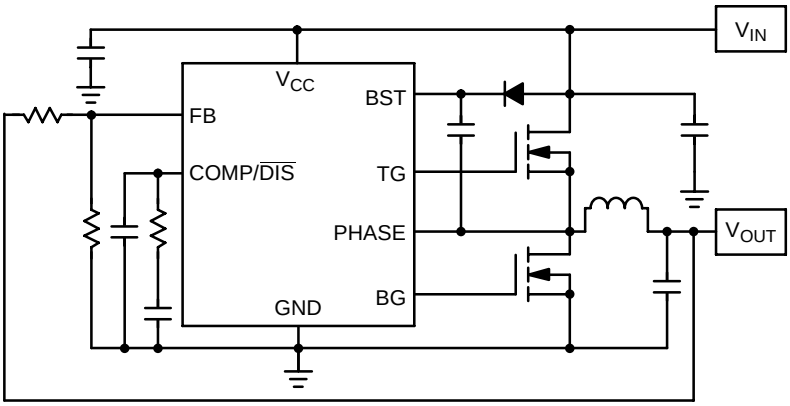


Figure 1. Typical Application Diagram

ON Semiconductor®

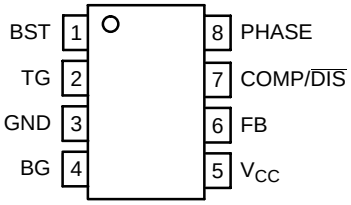
http://onsemi.com

MARKING DIAGRAM

SOIC−8
D SUFFIX
CASE 751

x = 2, 2A or 3
A = Assembly Location
L = Wafer Lot
Y = Year
W = Work Week
▪ = Pb−Free Device

PIN CONNECTIONS



(Top View)

NCP158x Series	Oscillator Frequency	SCP Trip Voltage
NCP1582	350 kHz	−350 mV
NCP1582A	350 kHz	−450 mV
NCP1583	300 kHz	−350 mV

ORDERING INFORMATION

Device	Package	Shipping†
NCP1582DR2G	SOIC−8 (Pb−Free)	2500/Tape & Reel
NCP1582ADR2G	SOIC−8 (Pb−Free)	2500/Tape & Reel
NCP1583DR2G	SOIC−8 (Pb−Free)	2500/Tape & Reel

†For information on tape and reel specifications, including part orientation and tape sizes, please refer to our Tape and Reel Packaging Specifications Brochure, BRD8011/D.

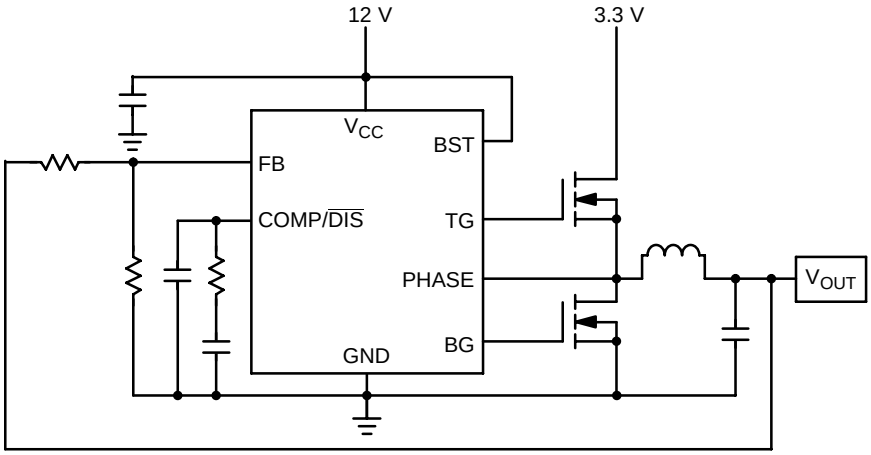


Figure 2. Typical VGA Card Application Diagram

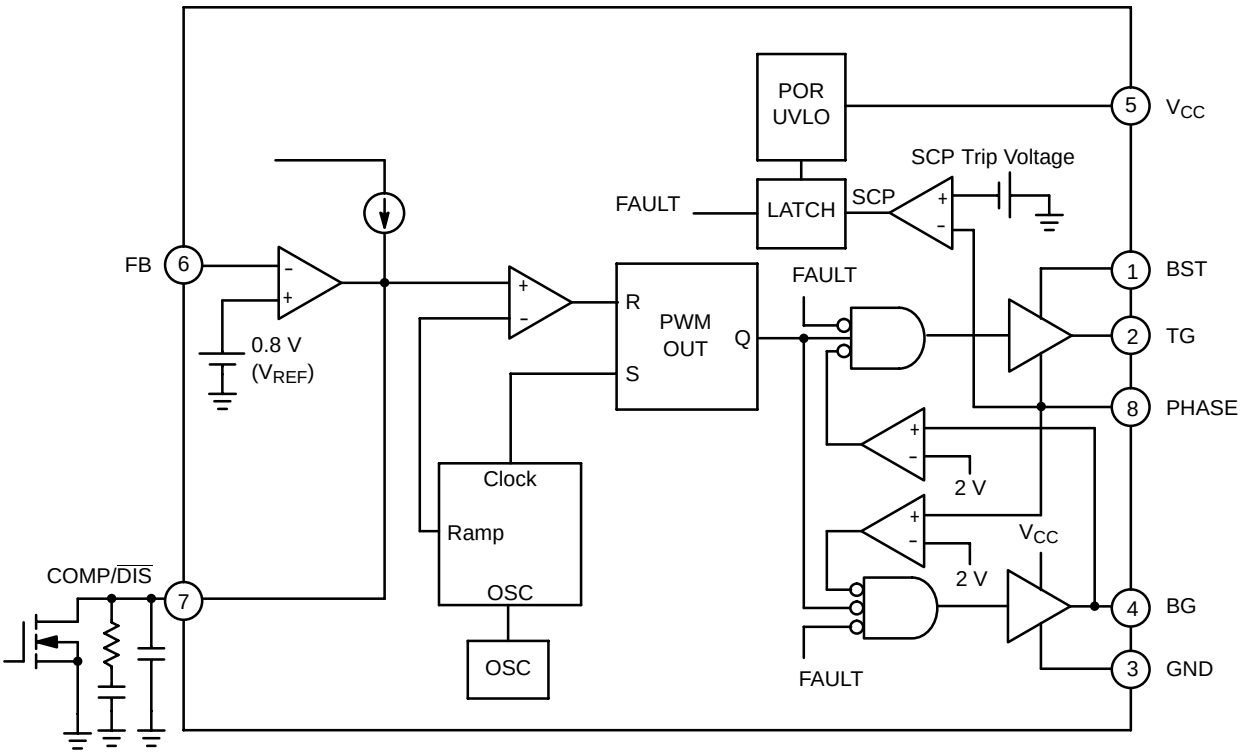


Figure 3. Detailed Block Diagram

PIN FUNCTION DESCRIPTION

Pin No.	Symbol	Description
1	BST	Supply rail for the floating top gate driver. To form a boost circuit, use an external diode to bring the desired input voltage to this pin (cathode connected to BST pin). Connect a capacitor (C_{BST}) between this pin and the PHASE pin. Typical values for C_{BST} range from 0.1 μF to 1 μF. Ensure that C_{BST} is placed near the IC.
2	TG	Top gate MOSFET driver pin. Connect this pin to the gate of the top N−Channel MOSFET.
3	GND	IC ground reference. All control circuits are referenced to this pin.
4	BG	Bottom gate MOSFET driver pin. Connect this pin to the gate of the bottom N−Channel MOSFET.
5	V_{CC}	Supply rail for the internal circuitry. Operating supply range is 4.5 V to 15 V. Decouple with a 1 μF capacitor to GND. Ensure that this decoupling capacitor is placed near the IC.
6	FB	This pin is the inverting input to the error amplifier. Use this pin in conjunction with the COMP pin to compensate the voltage−control feedback loop. Connect this pin to the output resistor divider (if used) or directly to Vout.
7	COMP/$\overline{DIS}$	Compensation Pin. This is the output of the error amplifier (EA) and the non−inverting input of the PWM comparator. Use this pin in conjunction with the FB pin to compensate the voltage−control feedback loop. The compensation capacitor also acts as a soft−start capacitor. Pull this pin low with an open drain transistor for disable.
8	PHASE	Switch node pin. This is the reference for the floating top gate driver. Connect this pin to the source of the top MOSFET.

ABSOLUTE MAXIMUM RATINGS

Pin Name	Symbol	V_{MAX}	V_{MIN}
Main Supply Voltage Input	V_{CC}	15 V	−0.3 V
Bootstrap Supply Voltage Input	BST	30 V wrt/GND 15 V wrt/PHASE	−0.3 V
Switching Node (Bootstrap Supply Return)	PHASE	24 V	−0.7 V −5 V for < 50 ns
High−Side Driver Output (Top Gate)	TG	30 V wrt/GND 15 V wrt/PHASE	−0.3 V wrt/PHASE
Low−Side Driver Output (Bottom Gate)	BG	15 V	−0.3 V −2 V for < 200 ns
Feedback	FB	5.5 V	−0.3 V
COMP/$\overline{DISABLE}$	COMP/$\overline{DIS}$	5.5 V	−0.3 V

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Thermal Resistance, Junction−to−Ambient	$R_{\theta JA}$	165	°C/W
Thermal Resistance, Junction−to−Case	$R_{\theta JC}$	45	°C/W
Operating Junction Temperature Range	T_J	−40 to 150	°C
Operating Ambient Temperature Range	T_A	−40 to 85	°C
Storage Temperature Range	T_{stg}	−55 to +150	°C
Lead Temperature Soldering (10 sec): Reflow (SMD styles only) Pb−Free (Note 1)		260 peak	°C
Moisture Sensitivity Level	MSL	1	−

Stresses exceeding Maximum Ratings may damage the device. Maximum Ratings are stress ratings only. Functional operation above the Recommended Operating Conditions is not implied. Extended exposure to stresses above the Recommended Operating Conditions may affect device reliability.

1. 60−180 seconds minimum above 237°C.

ELECTRICAL CHARACTERISTICS ($0°C < T_A < 70°C$, $-40°C < T_J < 125°C$ (Note 2), $4.5\ V < V_{CC} < 13.2\ V$, $4.5\ V < BST < 26.5\ V$, $C_{TG} = C_{BG} = 1.0$ nF(REF:NTD30N02), for min/max values unless otherwise noted.)

Characteristic	Conditions	Min	Typ	Max	Unit
Input Voltage Range	—	4.5		13.2	V
Boost Voltage Range	—	4.5		26.5	V
Supply Current					
Quiescent Supply Current	V_{FB} = 1.0 V, No Switching V_{CC} = 13.2 V	−	1.0	1.75	mA
Boost Quiescent Current	V_{FB} = 1.0 V, No Switching	−	140	−	μA
Under Voltage Lockout					
UVLO Threshold	V_{CC} Rising Edge	3.85	4.2		V
UVLO Hysteresis	—	−	0.5		V
Switching Regulator					
VFB Feedback Voltage, Control Loop in Regulation	T_A = 0 to 70°C −40 to 125°C	0.788	0.8 0.8	0.812	V
Oscillator Frequency (NCP1582, NCP1582A)	T_A = 0 to 70°C −40 to 125°C	300	350 350	400	kHz
Oscillator Frequency (NCP1583)	T_A = 0 to 70°C −40 to 125°C	275	300 300	325	kHz
Ramp−Amplitude Voltage		−	1.1	−	V
Minimum Duty Cycle		−	0	−	%
Maximum Duty Cycle		70	75	80	%
Minimum Pulse Width	Static Operating	100		150	ns
Blanking Time			50		ns
BG Minimum On Time			~500		ns
Error Amplifier (GM)					
Transconductance				5.0	mmho
Open Loop DC Gain		55	70	−	DB
Output Source Current Output Sink Current	V_{FB} = 0.8 V V_{FB} > 0.8 V	80 80	120 120		μA μA
Input Offset Voltage		−2.0	0	2.0	mV
Input Bias Current			0.1	1.0	μA
Unity Gain Bandwidth			4.0		Mhz
Soft−Start					
SS Source Current	V_{FB} < 0.8 V	5.0	10	15	μA
Switch Over Threshold			100		% of Vref
Current Limit					
Trip Voltage (NCP1582, NCP1583)	Vphase to ground		−350		mV
Trip Voltage (NCP1582A)	Vphase to ground		−450		mV
Gate Drivers					
Upper Gate Source	Vgs = 6.0 V	−	0.7		A
Upper Gate Sink	Vugate wrt Phase = 1.0 V		2.4		Ω
Lower Gate Source	Vgs = 6.0 V	−	0.7		A
Lower Gate Sink	Vlgate wrt GND = 1.0 V		2.2		Ω
PHASE Falling to BG Rising Delay	V_{CC} = 12 V, PHASE < 2.0 V, BG > 2.0 V	−	30	90	ns
BG Falling to TG Rising Delay	V_{CC} = 12 V, BG < 2.0 V, TG > 2.0 V	−	30	60	ns
Enable Threshold			0.4		V

2. Specifications to −40°C are guaranteed via correlation using standard quality control (SQC), not tested in production.

TYPICAL OPERATING CHARACTERISTICS

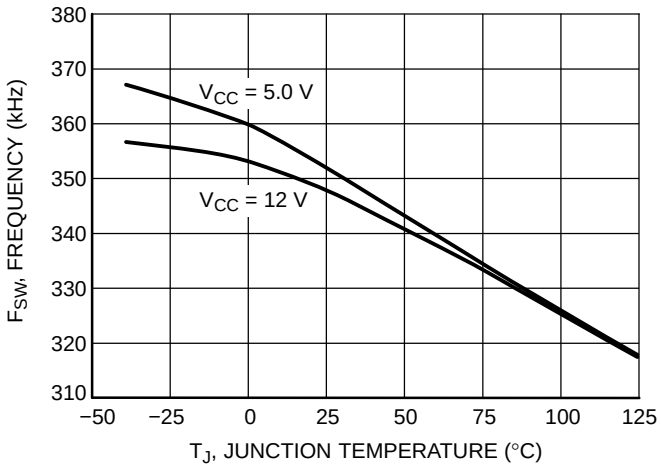


Figure 4. Oscillator Frequency (F_{SW}) vs. Temperature

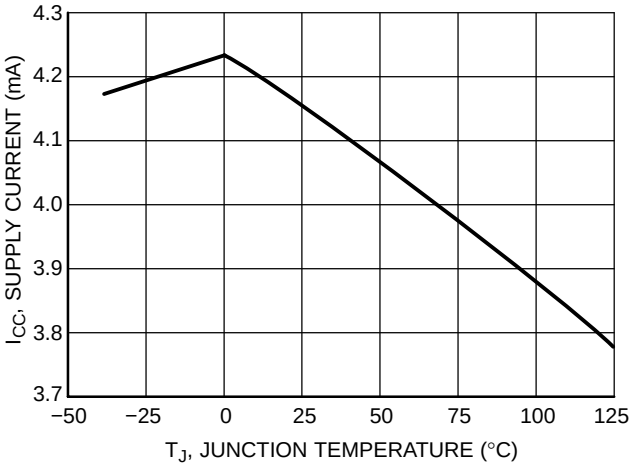


Figure 5. I_{CC} vs. Temperature

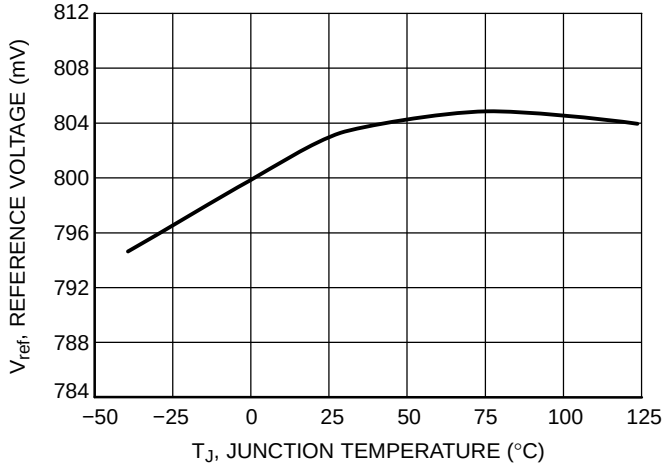


Figure 6. Reference Voltage (V_{ref}) vs. Temperature

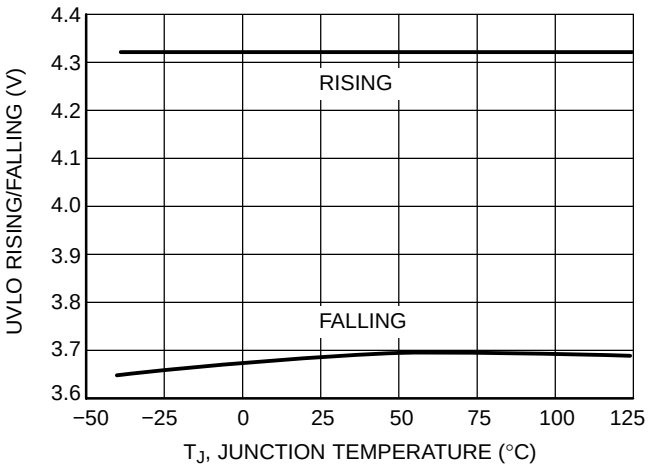


Figure 7. UVLO vs. Temperature

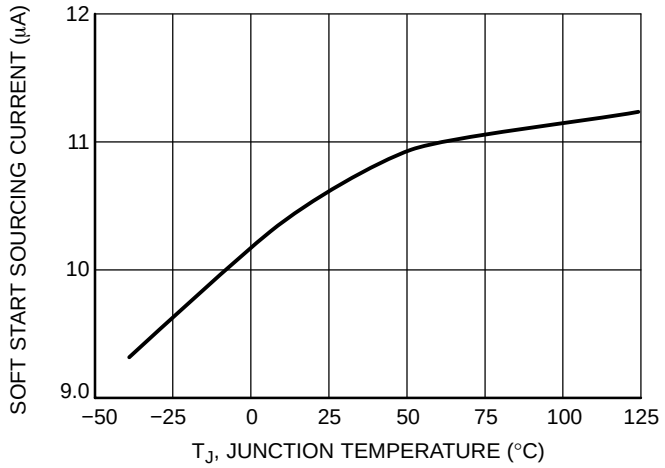


Figure 8. Soft Start Sourcing Current vs. Temperature

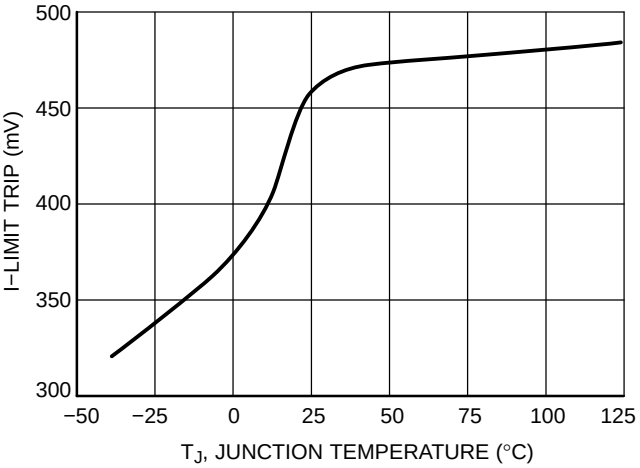


Figure 9. I−Limit vs. Temperature

DETAILED OPERATING DESCRIPTION

General

The NCP158x is an 8−pin PWM controller intended for DC−DC conversion from 5.0 V & 12 V buses. The NCP158x has a 0.7 A internal gate driver circuit designed to drive N−channel MOSFETs in a synchronous−rectifier buck topology. The output voltage of the converter can be precisely regulated down to 800 mV 1.5% when the V_{FB} pin is tied to V_{OUT}. The switching frequency is internally set. A high gain operational transconductance error amplifier (OTA) is used.

Duty Cycle and Maximum Pulse Width Limits

In steady state DC operation, the duty cycle will stabilize at an operating point defined by the ratio of the input to the output voltage. The NCP158x can achieve an 80% duty cycle. There is a built in off−time which ensures that the bootstrap supply is charged every cycle. The NCP158x, which is capable of a 100 nsec pulse width (min.), can allow a 12 V to 0.8 V conversion at 350 kHz.

Input Voltage Range (V_{CC} and BST)

The input voltage range for both V_{CC} and BST is 4.5 V to 13.2 V with respect to GND and PHASE, respectively. Although BST is rated at 13.2 V with respect to PHASE, it can also tolerate 26.5 V with respect to GND.

External Enable/Disable

When the Comp pin voltage falls or is pulled externally below the 400 mv threshold, it disables the PWM Logic and the gate drive outputs. In this disabled mode, the operational transconductance error amplifier's (EOTA) output source current is reduced and limited to the Soft Start current of 10 μA.

Normal Shutdown Behavior

Normal shutdown occurs when the IC stops switching because the input supply reaches UVLO threshold. In this case, switching stops, the internal SS is discharged, and all GATE pins go low. The switch node enters a high impedance state and the output capacitors discharge through the load with no ringing on the output voltage.

External Soft Start

The NCP158x features an external soft start function, which reduces inrush current and overshoot of the output voltage. Soft start is achieved by using the internal current source of 10 μA. (typ), which charges the external integrator capacitor of the transconductance amplifier. Figure 10 is a typical soft start sequence. This sequence begins once V_{CC} surpasses its UVLO threshold. During Soft Start, as the Comp Pin rises through 400 mV, the PWM Logic and gate drives are enabled. When the feedback voltage crosses 800 mV, the EOTA will be given control to switch to its higher regulation mode output current of 120 μA. In the event of an overcurrent during soft start, the overcurrent logic will override the soft start sequence and will shut down the PWM logic and both the high side and low side gates.

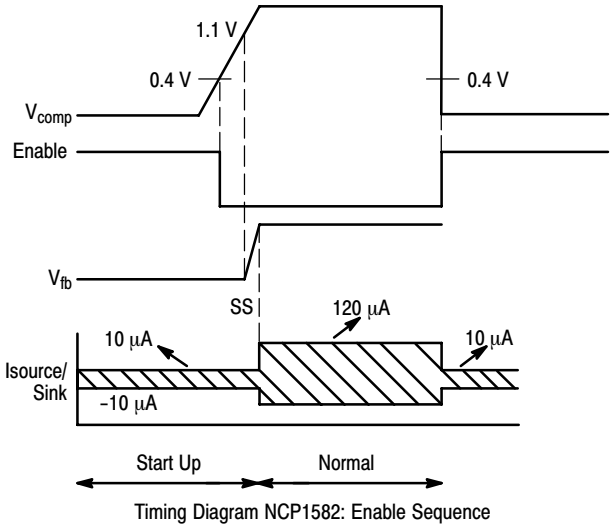


Figure 10. Soft Start Implementation

UVLO

Under Voltage Lockout (UVLO) is provided to ensure that unexpected behavior does not occur when V_{CC} is too low to support the internal rails and power the converter. For the NCP158x, the UVLO is set to ensure that the IC will start up when V_{CC} reaches 4.2 V and shutdown when V_{CC} drops below 3.7 V. This permits operation when converting from a 5.0 input voltage.

Current Limit Protection

In case of a short circuit or overload, the low−side (LS) FET will conduct large currents. The controller will shut down the regulator in this situation for protection against overcurrent. The low−side R_{DSon} sense is implemented by comparing the voltage at the Phase node when BG starts going low to an internally generated fixed voltage. If the phase voltage is lower than SCP trip voltage, an overcurrent condition occurs and a counter is initiated. When the counter completes, the PWM logic and both HS−FET and LS−FET are turned off. The controller will retry to see if the short circuit or overload condition is removed through the soft start cycle. The minimum turn−on time of the LS−FET is set to be 500 ns. The trip thresholds have a −95 mV, +45 mV process and temperature variation.

Drivers

The NCP158x includes 0.7 A gate drivers to switch external N−channel MOSFETs. This allows the NCP158x to address high−power as well as low−power conversion requirements. The gate drivers also include adaptive non−overlap circuitry. The non−overlap circuitry increase efficiency, which minimizes power dissipation, by minimizing the body diode conduction time.

A detailed block diagram of the non−overlap and gate drive circuitry used in the chip is shown in Figure 11.

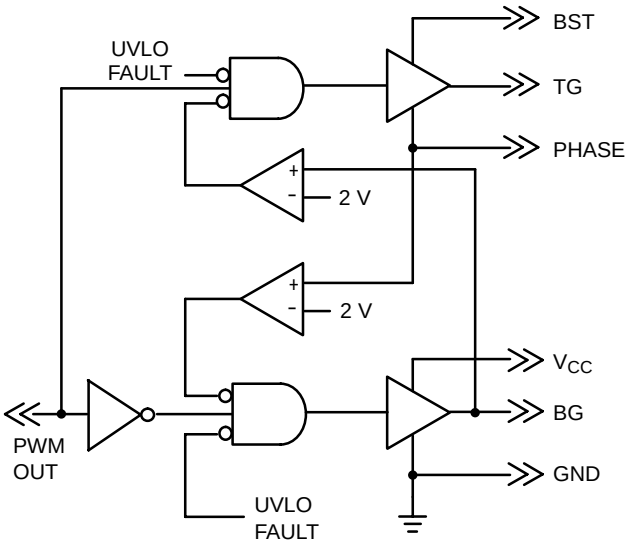


Figure 11. Block Diagram of Gate Driver and Non−Overlap Circuitry

Careful selection and layout of external components is required, to realize the full benefit of the onboard drivers. The capacitors between V_{CC} and GND and between BST and SWN must be placed as close as possible to the IC. The current paths for the TG and BG connections must be optimized. A ground plane should be placed on the closest layer for return currents to GND in order to reduce loop area and inductance in the gate drive circuit.

APPLICATION SECTION

Input Capacitor Selection

The input capacitor has to sustain the ripple current produced during the on time of the upper MOSFET, so it must have a low ESR to minimize the losses. The RMS value of this ripple is:

$$Iin_{RMS} = I_{OUT}\sqrt{D \times (1 - D)},$$

where D is the duty cycle, Iin_{RMS} is the input RMS current, & I_{OUT} is the load current. The equation reaches its maximum value with D = 0.5. Losses in the input capacitors can be calculated with the following equation:

$$P_{CIN} = ESR_{CIN} \times Iin_{RMS}^2,$$

where P_{CIN} is the power loose in the input capacitors & ESR_{CIN} is the effective series resistance of the input capacitance. Due to large d_I/d_t through the input capacitors, electrolytic or ceramics should be used. If a tantalum must be used, it must be surge protected. Otherwise, capacitor failure could occur.

Calculating Input Start−up Current

To calculate the input start up current, the following equation can be used.

$$I_{inrush} = \frac{C_{OUT} \times V_{OUT}}{t_{SS}},$$

where I_{inrush} is the input current during start−up, C_{OUT} is the total output capacitance, V_{OUT} is the desired output voltage, and t_{SS} is the soft start interval.

If the inrush current is higher than the steady state input current during max load, then the input fuse should be rated accordingly, if one is used.

Calculating Soft Start Time

To calculate the soft start time, the following equation can be used.

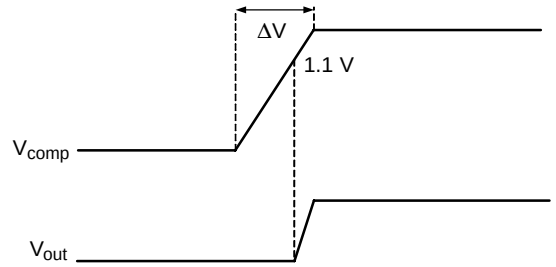
$$t_{SS} = \frac{(C_P + C_C) * \Delta V}{I_{SS}}$$

Where C_C is the compensation as well as the soft start capacitor,

C_P is the additional capacitor that forms the second pole.

I_{SS} is the soft start current

ΔV is the comp voltage from zero to until it reaches regulation.



The above calculation includes the delay from comp rising to when output voltage becomes valid.

To calculate the time of output voltage rising to when it reaches regulation; ΔV is the difference between the comp voltage reaching regulation and 1.1 V.

Output Capacitor Selection

The output capacitor is a basic component for the fast response of the power supply. In fact, during load transient, for the first few microseconds it supplies the current to the load. The controller immediately recognizes the load transient and sets the duty cycle to maximum, but the current slope is limited by the inductor value.

During a load step transient the output voltage initial drops due to the current variation inside the capacitor and the ESR. (neglecting the effect of the effective series inductance (ESL)):

$$\Delta V_{OUT-ESR} = \Delta I_{OUT} \times ESR_{COUT},$$

where $V_{OUT-ESR}$ is the voltage deviation of V_{OUT} due to the effects of ESR and the ESR_{COUT} is the total effective series resistance of the output capacitors.

A minimum capacitor value is required to sustain the current during the load transient without discharging it. The voltage drop due to output capacitor discharge is given by the following equation:

$$\Delta V_{OUT-DISCHARGE} = \frac{\Delta I_{OUT}^2 \times L_{OUT}}{2 \times C_{OUT} \times (V_{IN} \times D - V_{OUT})},$$

where $V_{OUT-DISCHARGE}$ is the voltage deviation of V_{OUT} due to the effects of discharge, L_{OUT} is the output inductor value & V_{IN} is the input voltage.

It should be noted that $\Delta V_{OUT-DISCHARGE}$ and $V_{OUT-ESR}$ are out of phase with each other, and the larger of these two voltages will determine the maximum deviation of the output voltage (neglecting the effect of the ESL).

Inductor Selection

Both mechanical and electrical considerations influence the selection of an output inductor. From a mechanical perspective, smaller inductor values generally correspond to smaller physical size. Since the inductor is often one of the largest components in the regulation system, a minimum inductor value is particularly important in space−constrained applications. From an electrical perspective, the maximum current slew rate through the output inductor for a buck regulator is given by:

$$SlewRate_{LOUT} = \frac{V_{IN} - V_{OUT}}{L_{OUT}}.$$

This equation implies that larger inductor values limit the regulator's ability to slew current through the output inductor in response to output load transients. Consequently, output capacitors must supply the load current until the inductor current reaches the output load current level. This

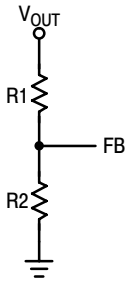
results in larger values of output capacitance to maintain tight output voltage regulation. In contrast, smaller values of inductance increase the regulator's maximum achievable slew rate and decrease the necessary capacitance, at the expense of higher ripple current. The peak−to−peak ripple current is given by the following equation:

$$Ipk - pk_{LOUT} = \frac{V_{OUT}(1 - D)}{L_{OUT} \times 350\ kHz},$$

where $Ipk-pk_{LOUT}$ is the peak to peak current of the output. From this equation it is clear that the ripple current increases as L_{OUT} decreases, emphasizing the trade−off between dynamic response and ripple current.

Feedback and Compensation

The NCP158x allows the output of the DC−DC converter to be adjusted from 0.8 V to 5.0 V via an external resistor divider network. The controller will try to maintain 0.8 V at the feedback pin. Thus, if a resistor divider circuit was placed across the feedback pin to V_{OUT}, the controller will regulate the output voltage proportional to the resistor divider network in order to maintain 0.8 V at the FB pin.



The relationship between the resistor divider network above and the output voltage is shown in the following equation:

$$R_2 = R_1 \times \left(\frac{V_{REF}}{V_{OUT} - V_{REF}}\right).$$

Resistor R1 is selected based on a design tradeoff between efficiency and output voltage accuracy. For high values of R1 there is less current consumption in the feedback network, However the trade off is output voltage accuracy due to the bias current in the error amplifier. The output voltage error of this bias current can be estimated using the following equation (neglecting resistor tolerance):

$$Error\% = \frac{0.1\ \mu A \times R_1}{V_{REF}} \times 100\%.$$

Once R1 has been determined, R2 can be calculated.

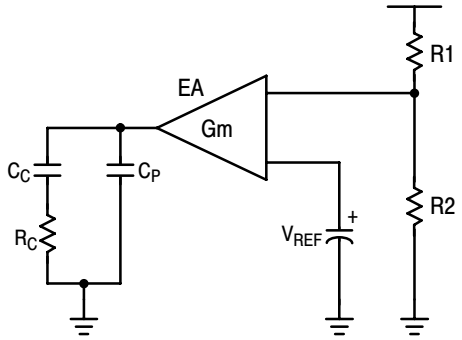


Figure 12. Type II Transconductance Error Amplifier

Figure 12 shows a typical Type II transconductance error amplifier (EOTA). The compensation network consists of the internal error amplifier and the impedance networks ZIN (R_1, R_2) and external Z_{FB} (R_c, C_c and C_p). The compensation network has to provide a closed loop transfer function with the highest 0 dB crossing frequency to have fast response (but always lower than $F_{SW}/8$) and the highest gain in DC conditions to minimize the load regulation. A stable control loop has a gain crossing with −20 dB/decade slope and a phase margin greater than 45°. Include worst−case component variations when determining phase margin. Loop stability is defined by the compensation network around the EOTA, the output capacitor, output inductor and the output divider. Figure 13. shows the open loop and closed loop gain plots.

Compensation Network Frequency:

The inductor and capacitor form a double pole at the frequency

$$F_{LC} = \frac{1}{2\pi \cdot \sqrt{L_O \cdot C_O}}$$

The ESR of the output capacitor creates a "zero" at the frequency,

$$F_{ESR} = \frac{1}{2\pi \cdot ESR \cdot C_O}$$

The zero of the compensation network is formed as,

$$F_Z = \frac{1}{2\pi \cdot R_C C_C}$$

The pole of the compensation network is calculated as,

$$F_P = \frac{1}{2\pi \cdot R_C \cdot C_P}$$

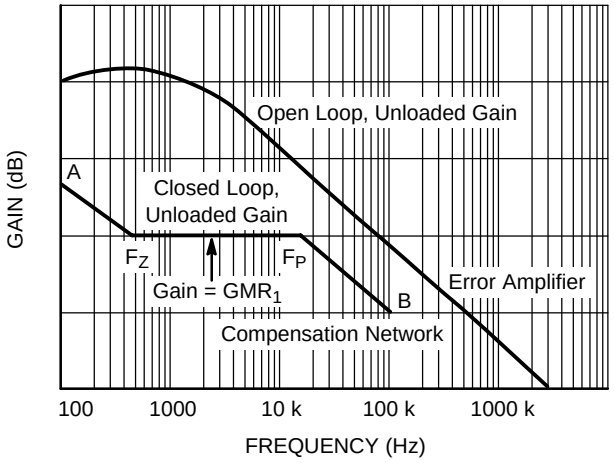


Figure 13. Gain Plot of the Error Amplifier

Thermal Considerations

The power dissipation of the NCP158x varies with the MOSFETs used, V_{CC}, and the boost voltage (V_{BST}). The average MOSFET gate current typically dominates the control IC power dissipation. The IC power dissipation is determined by the formula:

$$P_{IC} = (I_{CC} \cdot V_{CC}) + P_{TG} + P_{BG}.$$

Where:

P_{IC} = control IC power dissipation,
I_{CC} = IC measured supply current,
V_{CC} = IC supply voltage,
P_{TG} = top gate driver losses,
P_{BG} = bottom gate driver losses.

The upper (switching) MOSFET gate driver losses are:

$$P_{TG} = Q_{TG} \cdot f_{SW} \cdot V_{BST}.$$

Where:

Q_{TG} = total upper MOSFET gate charge at V_{BST},
f_{SW} = the switching frequency,
V_{BST} = the BST pin voltage.

The lower (synchronous) MOSFET gate driver losses are:

$$P_{BG} = Q_{BG} \cdot f_{SW} \cdot V_{CC}.$$

Where:

Q_{BG} = total lower MOSFET gate charge at V_{CC}.

The junction temperature of the control IC can then be calculated as:

$$T_J = T_A + P_{IC} \cdot \theta_{JA}.$$

Where:

T_J = the junction temperature of the IC,
T_A = the ambient temperature,
θ_{JA} = the junction−to−ambient thermal resistance of the IC package.

The package thermal resistance can be obtained from the specifications section of this data sheet and a calculation can be made to determine the IC junction temperature. However, it should be noted that the physical layout of the board, the proximity of other heat sources such as MOSFETs and inductors, and the amount of metal connected to the IC, impact the temperature of the device. Use these calculations as a guide, but measurements should be taken in the actual application.

Layout Considerations

As in any high frequency switching converter, layout is very important. Switching current from one power device to another can generate voltage transients across the impedances of the interconnecting bond wires and circuit traces. These interconnecting impedances should be minimized by using wide, short printed circuit traces. The critical components should be located as close together as possible using ground plane construction or single point grounding. The figure below shows the critical power components of the converter. To minimize the voltage overshoot the interconnecting wires indicated by heavy lines should be part of ground or power plane in a printed circuit board. The components shown in the figure below should be located as close together as possible. Please note that the capacitors C_{IN} and C_{OUT} each represent numerous physical capacitors. It is desirable to locate the NCP158x within 1 inch of the MOSFETs, Q1 and Q2. The circuit traces for the MOSFETs' gate and source connections from the NCP158x must be sized to handle up to 2 A peak current.

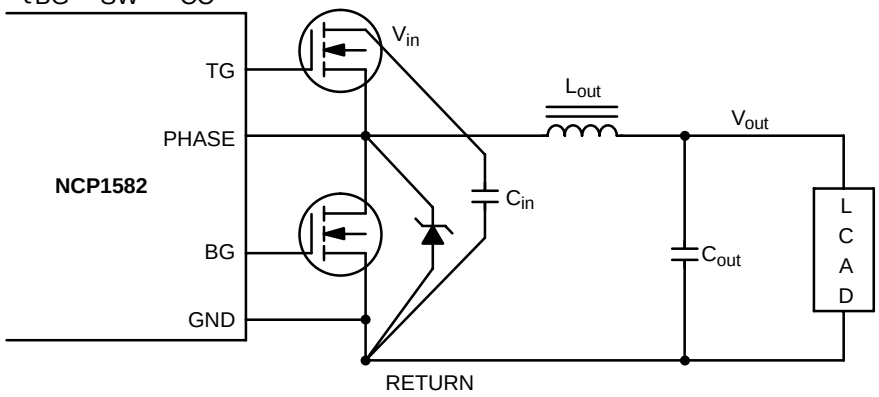


Figure 14. Components to be Considered for Layout Specifications

Design Example

Switching Frequency F_{SW} = 350 KHZ
Output Capacitance C_{ESR} = 45 mΩ/Each
Output Capacitance C_{out} = 6630 μF
Output Inductance L_{out} = 0.75 μH
Input Voltage V_{in} = 12 V
Output Voltage V_{out} = 3.3 V

Choose the loop gain crossover frequency;

$$F_{CO} = \frac{1}{10} * F_{SW} = 35\ kHz$$

The corner frequency of the output filter is calculated below;

$$F_{LC} = \frac{1}{2 * \pi * \sqrt{0.75\ \mu H * 6630\ \mu F}} = 2.3\ kHz$$

Let R_C = 1500

Check that the ESR zero frequency is not too high;

$$F_{ESR} = \frac{1}{2 * \pi \cdot CESR \cdot C_O} < \frac{F_{SW}}{5}$$

This condition is mandatory for loop stability.

Zero of the compensation network is calculated as follows;

$$F_Z = F_{LC}$$

$$C_C = \frac{1}{2 * \pi * F_Z * R_C}$$

$$= \frac{1}{2 * \pi * 2.3\ kHz * 1500} = 46\ nF$$

The compensation capacitor also acts as the soft start capacitor. By adjusting the value of this compensation capacitor, the soft start time can be adjusted.

Pole of the compensation network is calculated as follows;

$$F_P = 5 * F_{CO} = 175\ kHz$$

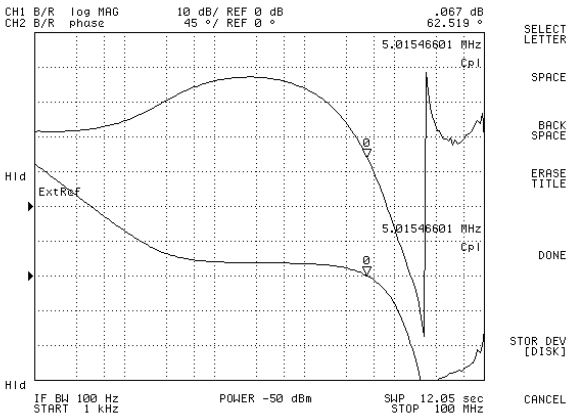
$$C_P = \frac{1}{2 * \pi * F_P * R_C}$$

$$= \frac{1}{2 * \pi * 175\ kHz * 1500} = 700\ pF$$

The recommended compensation values are;

R_C = 1500, C_C = 46 nF, C_P = 700 pF

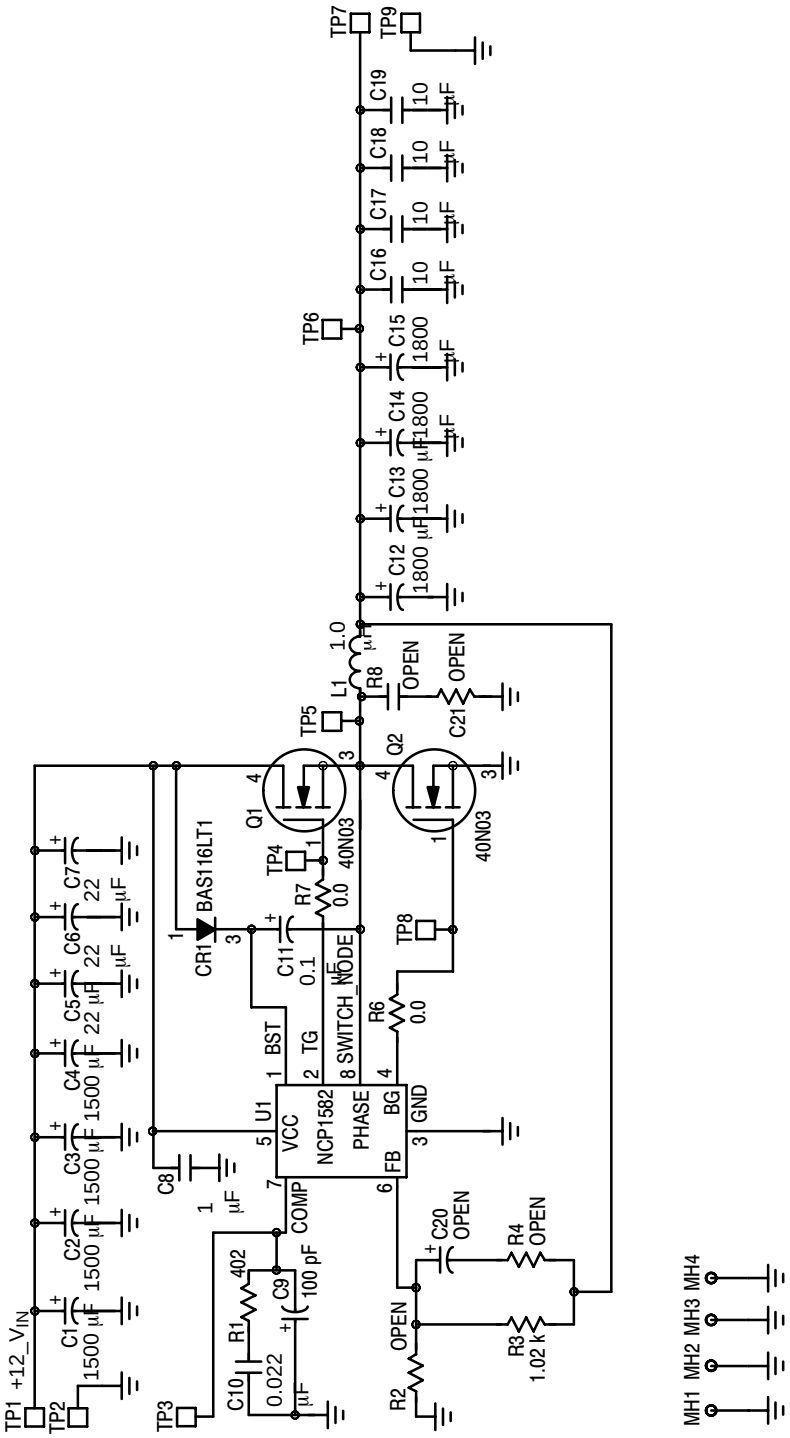
The NCP158x bode plot as measured from the network analyzer is shown below.



Top plot: Phase−Frequency (Phase Margin = 62.519°)
Bottom plot: Gain−Frequency (UGBW= 5 MHz)

Figure 15. Typical Bode plot of the Open−loop Frequency Response of the NCP158x

Demo Board PCB Layout



Bill of Materials

Item Number	Part Reference	Value	Quantity	MFG
1	C1 C2 C3 ~~C4~~	1500 μF	4	PANASONIC
2	C5 C6 C7	22 μF	3	TDK
3	C8	1.0 μF	1	TAIYO YUDEN
4	C9	100 pF	1	AVX
5	C10	0.022 μF	1	KEMET
6	C11	0.1 μF	1	AVX
7	C12 C13 C14 ~~C15~~	1800 μF	4	PANASONIC
8	C16 C17 C18 C19	10 μF	4	KEMET
9	C20	OPEN	1	–
10	C21	OPEN	1	–
11	CR1	BAS116LT1	1	ON SEMICONDUCTOR
12	L1	0.75 μH	1	TOKO
13	Q1 Q2	40N03	2	ON SEMICONDUCTOR
14	R1	402	1	DALE
15	R2	OPEN	1	–
16	R3	1.02 K	1	DALE
17	R4	OPEN	1	–
18	R6 R7	0	2	DALE
19	R8	OPEN	1	–
20	U1	NCP158x	1	ON SEMICONDUCTOR

TYPICAL PERFORMANCE CHARACTERISTICS

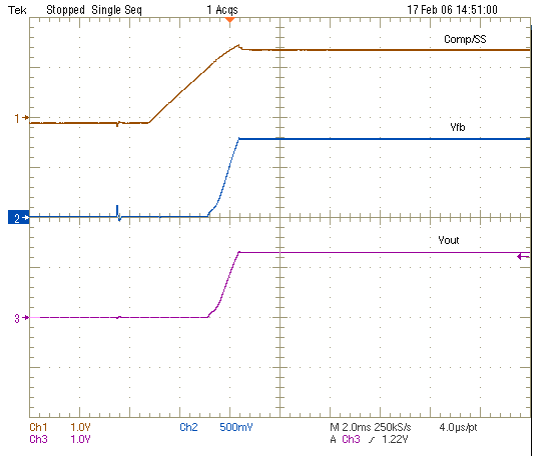


Figure 16. Start Up

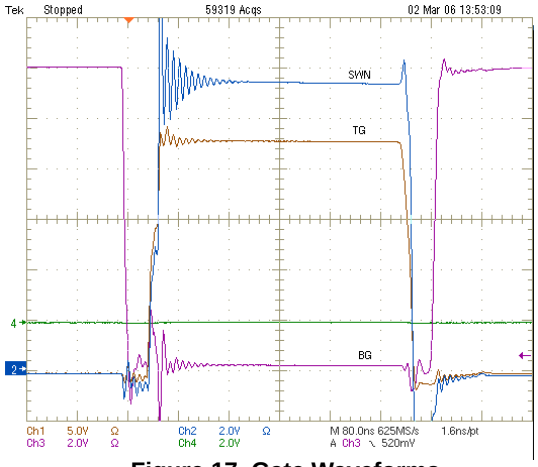


Figure 17. Gate Waveforms 15 A Load Sustaining

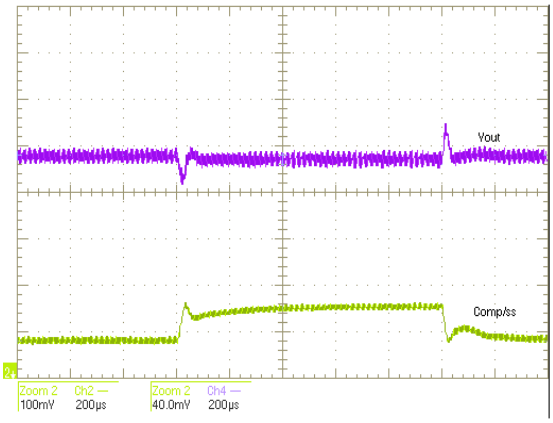


Figure 18. Transient Response (0−10 A Step Load)

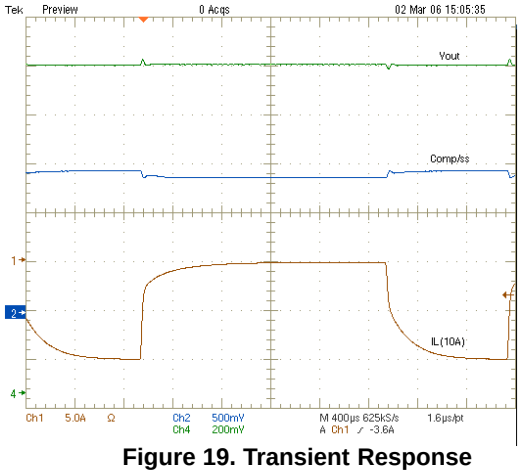


Figure 19. Transient Response

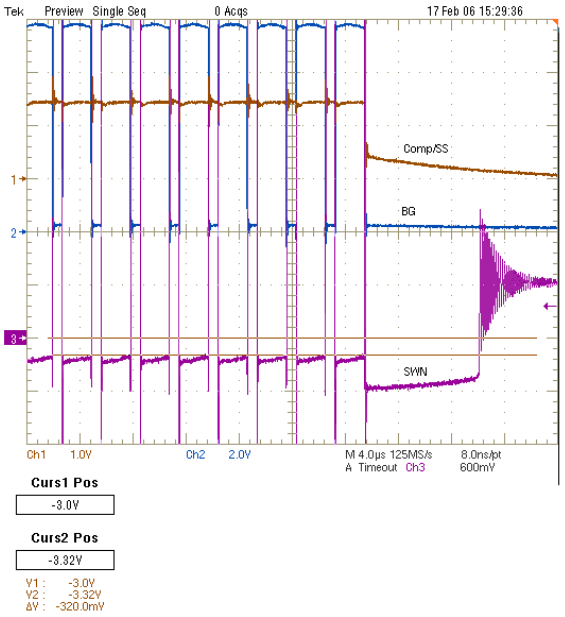


Figure 20. Over Current Protection (22 A DC Trip)

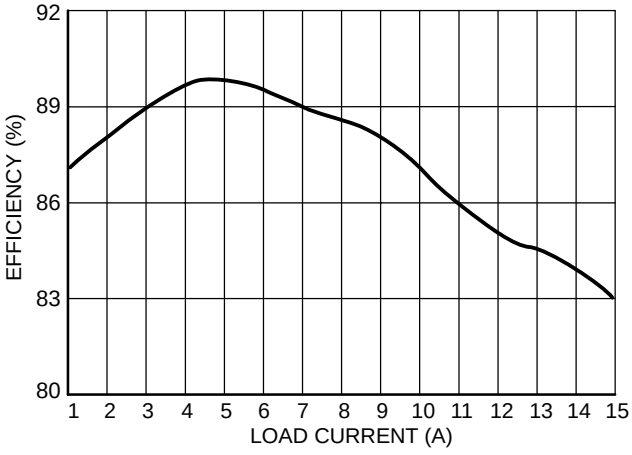
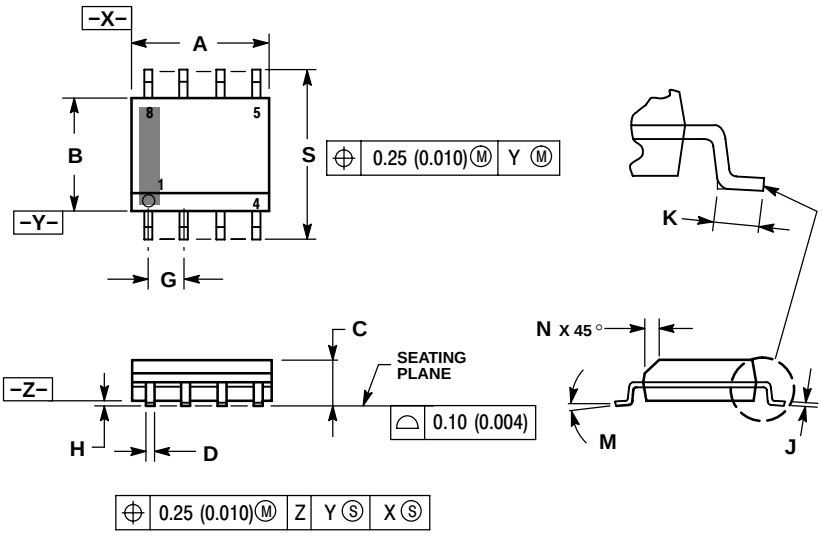


Figure 21. Efficiency vs. Load Current

PACKAGE DIMENSIONS

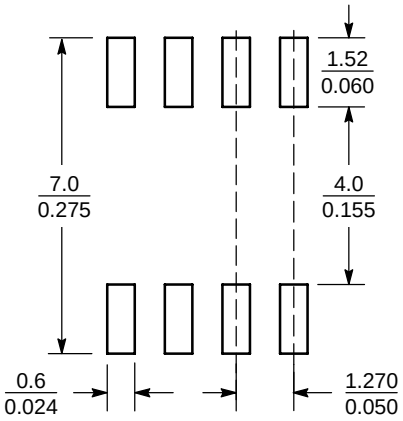
SOIC−8
D SUFFIX
CASE 751−07
ISSUE AH



NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: MILLIMETER.
3. DIMENSION 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.
6. 751−01 THRU 751−06 ARE OBSOLETE. NEW STANDARD IS 751−07.

	MILLIMETERS		INCHES	
DIM	MIN	MAX	MIN	MAX
A	4.80	5.00	0.189	0.197
B	3.80	4.00	0.150	0.157
C	1.35	1.75	0.053	0.069
D	0.33	0.51	0.013	0.020
G	1.27 BSC		0.050 BSC	
H	0.10	0.25	0.004	0.010
J	0.19	0.25	0.007	0.010
K	0.40	1.27	0.016	0.050
M	0 °	8 °	0 °	8 °
N	0.25	0.50	0.010	0.020
S	5.80	6.20	0.228	0.244

SOLDERING FOOTPRINT*



SCALE 6:1 (mm / inches)

*For additional information on our Pb−Free strategy and soldering details, please download the ON Semiconductor Soldering and Mounting Techniques Reference Manual, SOLDERRM/D.

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