Application Note 53

180 Watt Boost Converter

By Mark Ziegenfuss

General Description

The MIC2196 controller is used to implement a non-isolated boost converter (Fig. 1). A boost converter has a higher output voltage than its input voltage. As the input voltage varies the converter's output voltage is held constant by the feedback loop over its output current range. The output current of a boost converter is less than its input current. Neglecting the losses, the ratio of input current to output current is the same ratio of output voltage to input voltage. As the input voltage decreases the input current increases (effectively a negative input impedance) resulting in higher RMS currents in the converter. The under voltage lockout is used to prevent operation below about 10.5V. The output current is discontinuous with high AC RMS currents which require large output capacitors to smooth out. The input current is continuous with a triangular wave shape. When the controller is off and the output voltage goes below the input voltage there is a current path through the inductor and through the fly back diode to the output. No current limit exists for this current path so care must be taken not to short circuit the output. The input of the converter is 12V and the output is set at 26V (set by the R3 and R10 divider). The output current is 7 amps max. The maximum input voltage to the MIC2196 is 14v. Table 1 is a summary of the specifications of the 180 Watt boost converter. The parts list of the 180 boost converter is shown in Table 2.

Parameter	Min	Typ	Max
VIN	$10.5V_{DC}$	$12V_{DC}$	$14V_{DC}$
Output voltage	$25V_{DC}$	$26V_{DC}$	$27V_{DC}$
Output current	0		7A
Power out	0		180W
efficiency		92%	
Output ripple			$1V_{PP}$
Switching Freq		400kHz	

Table 1 Design Specifications

Schematic

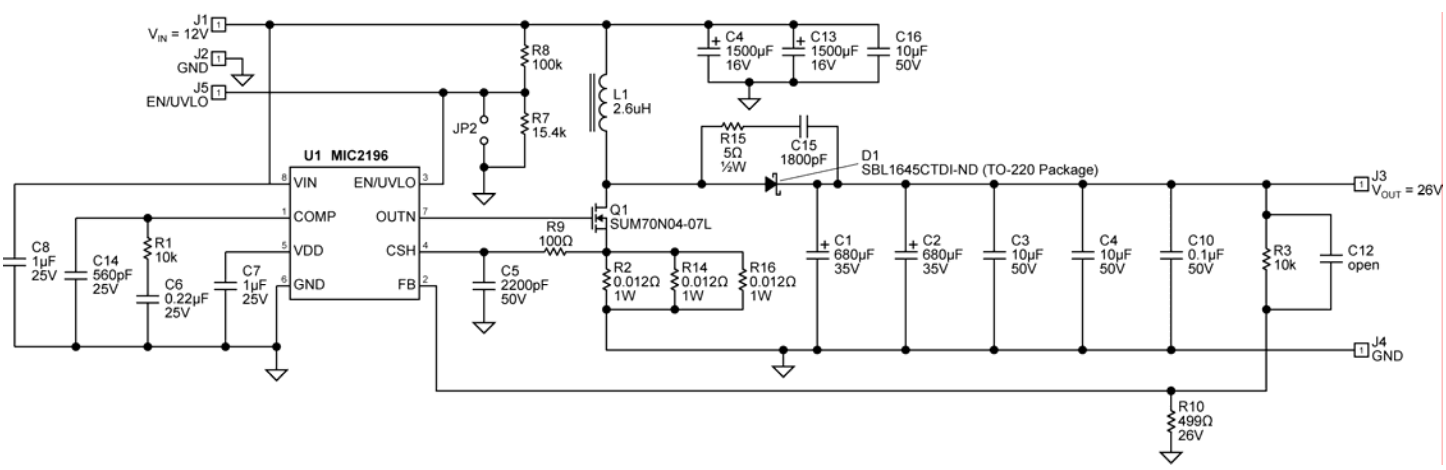

Figure 1 180 Watt Boost Schematic Diagram

Operation

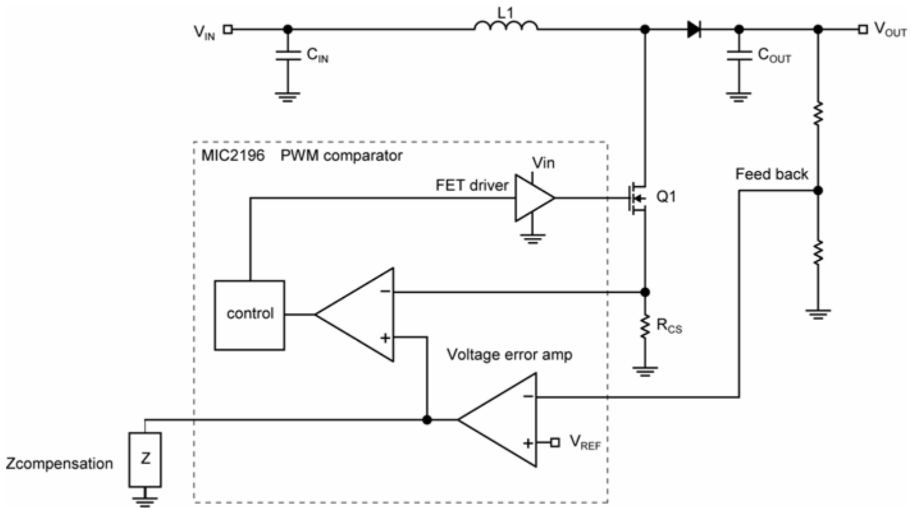

Figure 2 MIC2196 Boost Block Diagram

Basic Control Loop

The control loop is shown in Figure 2. The major components are the Inductor L1, the FET Q1, the fly back diode D1, the MIC2196 controller, the input and output capacitors. The MIC2196 is a peak current mode PWM controller. The current through the current sense resistor Rcs ramps up from zero to Ipk when Q1 is on (Ton). The MIC2196 PWM regulates the output voltage by comparing the ramp to the output of the voltage error amp so as the output voltage decreases the error amp output increases allowing the ramp to travels up farther thereby increasing the Ton time. On a pulse to pulse scale the pulse width increases as the output voltage decreases in order to regulate the output voltage. The error amplifier is a transconductance type.

Cycle-By-Cycle Current limit

The MIC2196 features cycle-by-cycle current limit. An over current comparator monitors the voltage at the cs pin. If this reaches 0.11V the comparator will terminate the gate drive to Q1 mid pulse.

Front Edge Blanking

Front edge blanking is employed to prevent premature current limit. R9 and C5 form a low pass filter to help filter the leading edge spikes.

Under Voltage Lockout

The MIC2196 uses an under voltage lockout circuit (UVLO) that monitors the Vin rail. This is programmable by the voltage divider at the UVLO pin. The UVLO circuit disables the output gate drive when the UVLO pin is Below 1.5V. When this pin is below 0.9V the controller is forced into a complete micro power shut down.

This converter will not turn on until Vin reaches about 10.5 Volts. This prevents excessive inrush currents at low input voltages where the duty cycle would approach 100% if the controller where allowed to operate below the far below the threshold voltage. There is 100mv of hysterisis to prevent any instability during the application and removal of Vin.

Slope Compensation

Slope compensation is required for duty cycles greater than 50% to prevent instabilities present in peak current mode control. The MIC2196 employs internal slope compensation so the user does not have to provide this slope compensation externally.

Internal Error Amplifier

An error amplifier is internal to the MIC2196. This is a high gain, high bandwidth transconductance amplifier. Because this is a current mode controller a relatively simply compensation scheme is employed, two poles a one zero as shown in Figure 3.

Design Equations

$V_{IN} = 12V$

$V_{OUT} = 26V$

$I_O = 7A$

$R_{LOAD} = V_{OUT} / I_O$

$F_{SW} = 400K$

$T = 1/F_{SW}$

$\eta = 0.92$

$V_{REF}=1.245V$

gm=0.2mA/V

$R_{DSON}=15m\Omega$

$DCR=4.97m\Omega$

The duty cycle D is found by;

$$D = \frac{(V_{OUT} - V_{IN})}{V_{OUT}} = 0.54$$

$$D' = 1 - D$$

This converter is designed to regulate in the continuous conduction mode CCM or discontinuous conduction mode DCM. A continuous minimum load current and a minimum inductance L are defined.

$$P_{IN} = \frac{L \times I_{PK}^{2}}{2\times T} + \frac{V_{IN}\times I_{PK}\times T_{OFF}}{2\times T} = \frac{P_{OUT}}{\eta} = \frac{V_{OUT}\times I_{OUT}}{\eta} = 198\text{watt}$$

Where

$$I_{PK} = \frac{V_{IN} \times T_{ON}}{L} \text{ And } T = T_{ON} + T_{OFF}$$

Substituting and collecting like terms yields

$$L \geq \frac{V_{IN}^{2} \times (V_{OUT} - V_{IN}) \times \eta}{2 \times F_{SW} \times I_{OUT} \times V_{OUT}^{2}} = 0.5\mu H$$

The minimum value of L to stay in the continuous conduction mode CCM for a given load current is given by the above formula. The selected value of L will have to be greater than this value. The other criteria selecting L is the maximum ripple current. There is a trade off between size and ripple current. L = 2.6µH is a good compromise.

The inductor peak and average currents are found by;

$$I_{Lp\text{-}p} = \frac{V_L \times (V_{OUT} - V_{IN})}{L \times F \times V_{OUT}} = 6.5A_{p-p}$$

Where;

$$V_L = V_{IN} - I_{IN(ave)} \times (DCR + R_{DSON})$$

DCR is the winding resistance of the inductor and Rdson is the FET on resistance

The input current is calculated by

$$I_{IN(ave)} = \frac{I_{OUT} \times V_{OUT}}{\eta \times V_{IN}} = 16.5A$$

The inductor current is continuous with a ramp on top as shown in Figure 3.

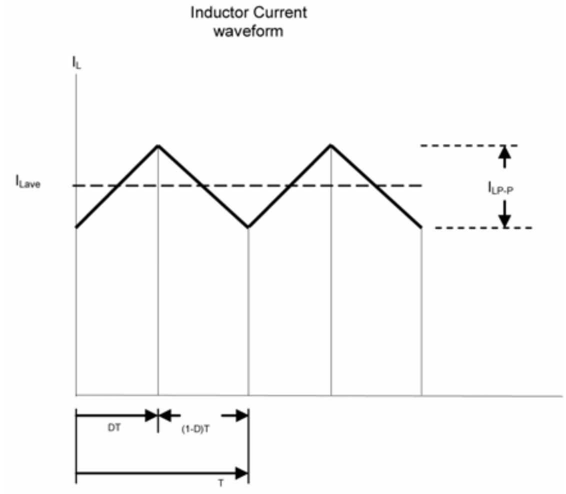

Figure 3 Inductor Current Wave form

The peak input current equals the peak inductor current and is calculated by

$$I_{Lpeak} = I_{L(ave)} + \frac{I_{Lp-p}}{2} = 20\,A$$

With the current levels defined selection of the power FET is possible.

The voltage stress on the FET is Vout=26V. Use a 40V FET with a current rating greater than 20A. Siliconix SUM70N04-07L is used; V_{DS} = 40V R_{DSON} @ 175˚C = 15mΩ, Q_g = 75nC, C_{OSS} = 320pF @ V_{OSS} = 25V

Power MOSFET Losses

The total power losses in the MOSFET is the sum of the conduction loss and the switching loss

PFET=switching loss + conduction loss

The conduction loss is

$$PFETon = I_{L(ave)}^{2} \times R_{DSON} \times D = 2.3W$$

The switching loss contains two parts the switching loss due to current and the switching loss due to C_{OSS} the drain source capacitance of the Power FET.

Internal to the MIC2196 is a 6 amp FET driver with a 2 ohm output resistance.

The switching loss due to current is

$$P_{CUR} = I \times V \times f_{SW} \times T_{SW}$$

$$\text{Where} \begin{cases} P_{CUR} = \text{Power loss due to current} \\ I = \text{Current} \\ V = \text{Voltage} \\ f_{SW} = \text{Switching frequency} \\ T_{SW} = \text{Switching time} = Q_g/I_{gatedrive} \\ I_{GATEDRIVE} = 6A \end{cases}$$

The switching power loss due to C_{OSS}

$$P_{COSS} = \frac{2\,C_{OSS}\sqrt{V_{OSS}}}{3} f_{SW}\ V^{3/2}$$

$$\text{Where} \begin{cases} C_{OSS} = 320\text{pf} \\ V_{OSS} = 25\text{V} \end{cases}$$

C_{OSS} and Voss can be found from the FET's Data sheet

$PFET_{SW} = P_{CUR} + P_{COSS}$ = 2.2W calculated

$PFET = PFET_{SW} + PFET_{ON}$ = 4.5W calculated

The switching losses vary from the calculated value as the switching time T_{SW} vary from FET to FET. The calculated value for T_{SW} is 12.5ns and the observed T_{SW} is about 25ns. This is partly because of the output impedance of the gate driver is not zero. Although very low (2ohms) the output impedance of the gate driver in effect increases T_{SW}.

Control Loop Stability and Compensation

Unlike voltage mode control current mode control does not have the two complex poles created by the LC output filter. The inductor is effectively taken out of the small signal transfer function (up until the about ½ the switching frequency). This simplifies the compensation needed for stability because in current mode there is not the180 degree phase shift associated with the LC output filter.

The small signal closed loop gain is,

$$G_{CL}(s) = G_{MOD} \times G_{MIC2196} \times G_{FEEDBACK} \times G_{ErrorAmp}$$

$$G_{MOD} = \frac{\hat{v}}{\hat{i}} = \frac{D'R_{LOAD}}{2} \frac{[1 - \frac{sL}{D'^2 R_{LOAD}}]}{[1 + \frac{sR_{LOAD}C_{OUT}}{2}]}$$

This is the current mode control-to-output small signal gain of the boost modulator (plant). It has a right half plane zero. $RHP_{ZERO} = \frac{sL}{D'^2 R_{LOAD}} = 48.4\text{Khz}$

It is essential that the closed loop transfer function's cross over frequency be much lower than the right half plane zero to ensure stability.

In addition to the gain of error amplifier the MIC2196 has a small signal gain of 60dB or $G_{MIC2196} = 1000$.

The feedback gain is the feedback resistor divider network or simply, $G_{FEEDBACK} = V_{REF}/V_{OUT}$ = 0.0479 or -26dB

The error amplifier small signal gain is

$G_{ErrorAmp}(s) = g_m \times Z_{COMP}$, where

$$Z_{COMP} = (R_1 + \frac{1}{sC_6}) \| (\frac{1}{sC_{14}})$$

As stated earlier the error amp has two poles and one zero (referred to as a type II error amp).

The computer generated transfer function of the error is shown in Figure 4.

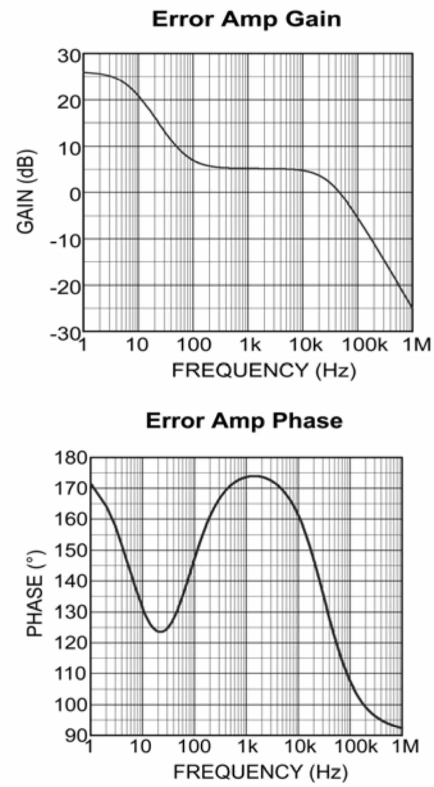

Figure 4 Error Amp Gain and Phase

The closed loop transfer function of the converter $G_{CL}(s) = G_{MOD} * G_{MIC2196} * G_{FEEDBACK} * G_{ErrorAmp}$ is shown in Figure 5. It has a DC gain of 55db, crosses over at 4.5khz with 75 degrees of phase margin, and 25db of gain margin.

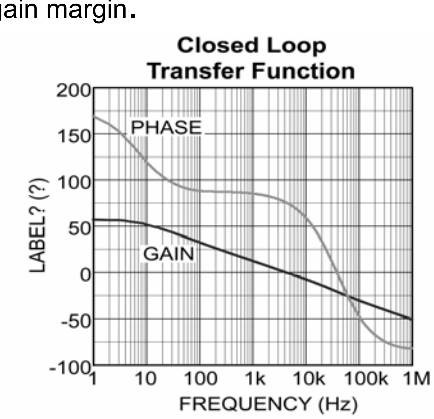

Figure 5 Closed Loop Gain and Phase

Heat Dissipation

High power converters have special concerns with regard to power dissipation. All though the efficiency is high (92%) at 180 watt output this converter has 16 watts of power dissipation. The power dissipation is mostly in the power FET and the diode. The I^2R losses in the inductor, PCB traces and current sense resistor all contribute to the power dissipation. Other losses include the ESR in the filter capacitors. The gate charge in the Power FET is dissipated in the gate driver of the MIC2196. Heat radiators (heat sinks) on the FET and Diode are needed to keep their junction temperatures at safe levels.

Bill of Materials

Ref	Part Description	Manufacturer	Part Number	Qty
U1	Boost controller	Micrel, Inc.	MIC2196BM	1
Q1	MOSFET	Vishay Siliconix.	TO-263	1
D1	16A, 45V schottky diode	Diodes. Inc	SBL1645CTDI-ND (TO-220 Package)	1
L1	2.6uH, 24A inductor	Sumida	CDEP-147-2R6	1
C1, C2	680 µF, 35V, aluminum	Chemi-Con	EKY-350ELL681MK20S	2
C5	22000pF, 50V ceramic cap	Vishay BC components	VJ0805Y222KXAMT	2
C6	0.22uF, 25V ceramic cap	muRata	GR M 21 B R7 1E 224 K A01B	2
C7, C8	1uF/25V, ceramic cap	muRata	GR M 21 B R7 1E 105 K A01B	3
C7, C8		Vishay Vitramon.	VJ0805S105KXJAT	OR
C4,C13	1500uF, 16V Aluminum cap	Chemi-Con	EKY-160E##152MK15S	2
C14	560pF,25V,cer cap	Murata	GR M 21 B R7 1E 562 K A01B	1
C3, C9,C16	10uf, 50V	Murata	GR M 32 D F5 1H 106 Z A01	2
C10	0.1uf, 50V	muRata	GR M 32 N 1X 1H 104 J Z01	1
C15	1800pf, 50V	muRata	VJ0805Y182KXAMT	1
R1,R3	10K (0805 size), 1%	Vishay Dale	CRCW08051002FRT1	2
R2, R4,R16	0.012 ohms (2512 size), 1%	Vishay Dale	WSL-2512-R012-F	3
R8	100K (0805 size), 1%	Vishay Dale	CRCW08051003FRT1	1
R7	15.4K (0805 size), 1%	Vishay Dale	CRCW08051542FRT1	1
R10	499 (0805 size), 1%	Vishay Dale	CRCW08054990FRT1	1
R9	100 (0805 size), 1%	Vishay Dale	CRCW08051000FRT1	1
R15	49.9, .5watt (2010 size) 1%	Vishay Dale	CRCW201049R9FKTA	1
HS1	Heat radiator (heat sink)	Thermalloy	533422B02552	1
HS2	Heat radiator (heat sink)	Thermalloy	533722B02552G	1
C11,C12	NOT USED			2
R5,R6, R12, R13	NOT USED			4

Table 2 180 Watt Boost Parts Lists

Notes:

1. **Micrel Semiconductor tel: 408-944-0800**
2. Vishay Corp. tel: 206-452-5664
3. Diodes, Inc. tel: 805-446-4800
4. muRata tel: 800-831-9172
5. Sumida tel: 408-321-9660
6. AVX tel: 843-448-9411
7. TDK tel: 847-803-6100

MICREL, INC. 2180 FORTUNE DRIVE SAN JOSE, CA 95131 USA
TEL +1 (408) 944-0800 FAX +1 (408) 474-1000 WEB http:/www.micrel.com

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