



Isolated Precision Regulation of a Single 1.8 Volt Output from a Universal Line Input

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APPLICATION NOTE

INTRODUCTION

The following Application Note describes an off-line switching power supply utilizing a precision programmable reference to regulate a 1.8 volt output. The center of the app note is the MC33363B, a monolithic SMPS controller with a 700 volt power switch, and the NCP100, a sub-one volt precision programmable reference. The system design and analysis will be described in detail.

The design requirements are for a universal off-line converter with a 1.8 volt, 1.0 ampere single output with less than 50 millivolts ripple and operates at 100 kHz. Most of the components selected are surface mount. This design is separated into the generic circuit of the off-line converter and the feedback network.

Generic Off-Line Conversion Circuit

To reduce the total number of components and thus minimize circuit board area, an SMPS controller with an integrated power switch is selected. With the low output power requirement, the MC33363B is selected as the control and power switch IC. The MC33363B contains an externally programmable frequency and current limit, internal startup circuit and can handle up to 8 watts of output power.

The basic off-line conversion circuit is illustrated in Figure 1. The line input is filtered through the EMC circuit then rectified and filtered. The rectified voltage is converted to a lower voltage via the transformer and the MC33363B. The secondary of the transformer is rectified to a DC voltage and filtered. The output voltage controls the duty cycle of the switcher via the isolated feedback network. The calculated values are a starting point and do not replace bench testing.

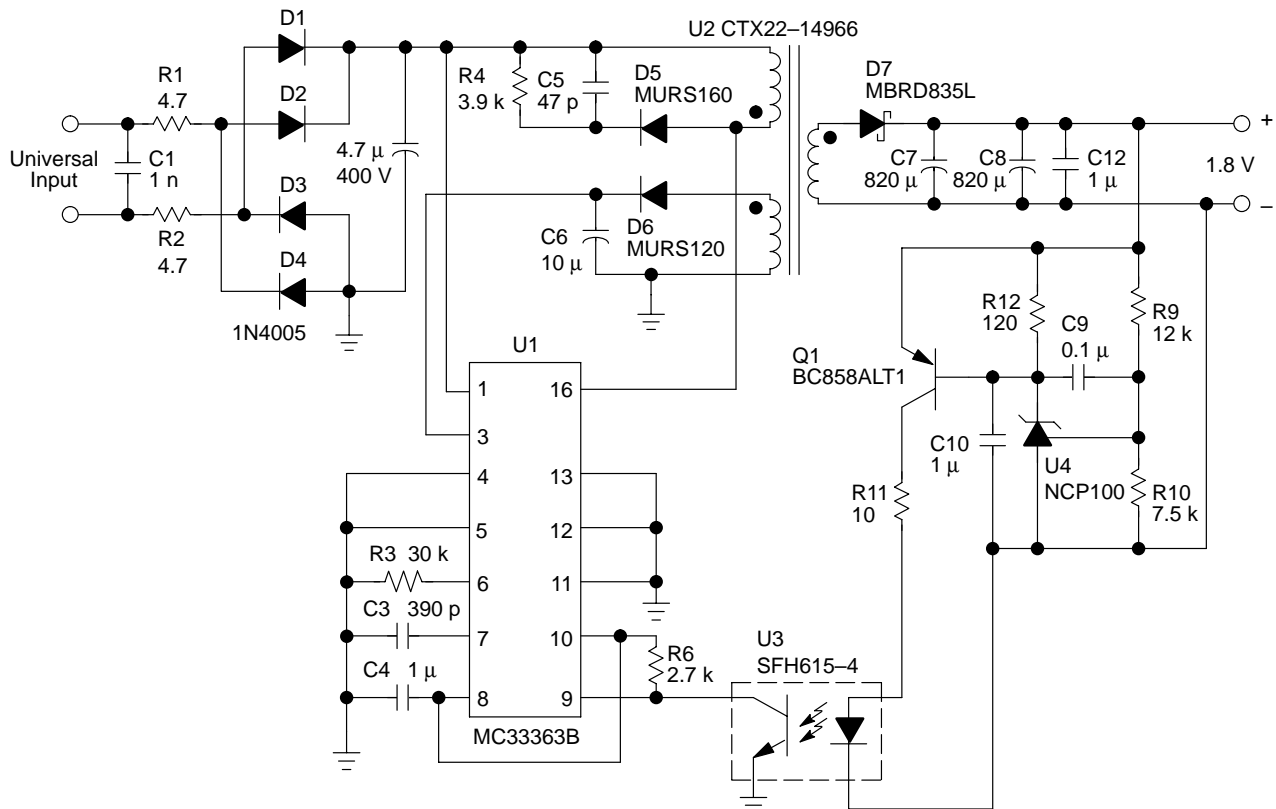


Figure 1. MC33363B Basic Flyback Circuit Schematic

Since the power level is low, an RC EMC filter is utilized. The RC filter is not as efficient as an LC, but it uses less board area and will cost less. 1N4005 600 volt fast diodes provide the full wave rectification of the AC input. The bulk capacitor is the filter. The selection of the capacitor is determined from three factors: input voltage, ripple current and maximum output ripple. The equations for each are as follows.

$$V_{dc} = V_{ac} \cdot \sqrt{2} \quad (\text{eq. 1})$$

$$I_{rms} = \sqrt{I_{pk}^2 \cdot \frac{D}{3}} \quad (\text{Discontinuous mode only}) \quad (\text{eq. 2})$$

$$C_{in} = \frac{k \cdot P_{in} (AV)}{f \cdot (V_{ripple} (p-p))^2} \quad (\text{eq. 3})$$

$$P_{in} = \frac{P_{out}}{\eta} \quad (\text{eq. 4})$$

Where k is 1 for AC inputs the peak to peak ripple is 6.0 volts and η is estimated to be 70% with the low power level of the board and the RC input filter. Solving for the equations above and using 50% for the maximum duty cycle, V_{dc} maximum is 375 volts, V_{dc} minimum is 120 volts, I_{peak} and I_{rms} are solved for in the transformer calculations, P_{in} is 3.2 watts, and C_{bulk} is 0.9 μF .⁽¹⁾

The transformer converts the rectified line voltage to the output voltage and is controlled by the MC33363B. To design the transformer, the following data is necessary. V_{dc} minimum is 100 volts (allowing for bulk voltage ripple and power switch voltage drop), the frequency is 100 kHz, the maximum duty cycle of the power switch and the reset time for the secondary are both 45% to maintain discontinuous conduction, and the secondary diode forward voltage drop is 0.45 volts (estimated).

$$P_{sec} = (V_{out} + V_{diode}) \cdot I_{out} \quad (\text{eq. 5})$$

$$P_{in} = \frac{\frac{1}{2} (V_{dc} \cdot T_{on})^2}{L_{pri} \cdot T}, \text{ or} \quad (\text{eq. 6})$$

$$L_{pri} = \frac{\frac{1}{2} (V_{dc} \cdot T_{on})^2}{P_{in} \cdot T} \quad (\text{eq. 7})$$

$$I_{peak} = \sqrt{2 \cdot P \cdot \frac{T}{L}} \quad (\text{eq. 8})$$

$$\frac{L_{pri}}{L_{sec}} = \left(\frac{N_p}{N_s} \right)^2 \quad (\text{eq. 9})$$

For the primary side of the transformer, L_{pri} is 3.16 mH, I_{peak} is 142 milliamperes, t_{on} is 4.5 μsec and I_{rms} is 55 milliamperes-rms. For the secondary, P_{sec} is 2.25 watts, L is 2.28 μH , I_{peak} is 4.44 amperes and I_{rms} is 1.72 amperes-rms. The turns ratio is 37.2. The auxiliary winding is set to 12 volts and 10 milliamperes. Using the secondary turns ratio, the auxiliary turns ratio is calculated as 6.6 in reference to the primary.

With the peak primary current and the frequency determined, the components surrounding the MC33363B can be selected. Referring to Figures 2 and 3, R_T will be set to 30 k Ohms and C_T will be set to 390 pF. The Over Voltage Protection, pin 11, will not be used. Its main function is for loss of optocoupler protection. Pin 8, the reference voltage, requires a 1.0 μF ceramic capacitor for stability. The voltage compensation and voltage feedback, pins 9 and 10, will be covered in the feedback section of this document. The V_{cc} pin requires a 10 μF capacitor for stability.

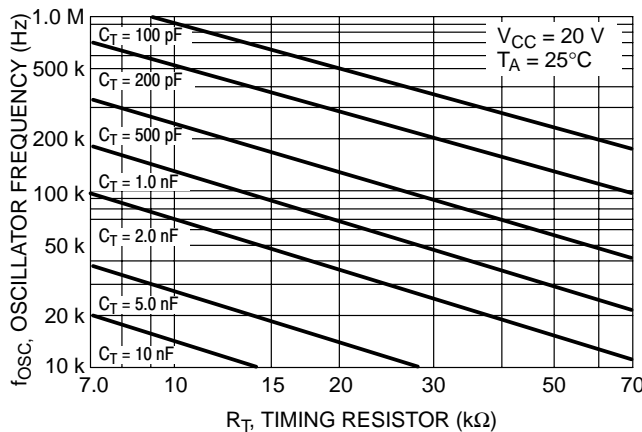


Figure 2. Oscillator Frequency versus Timing Resistor

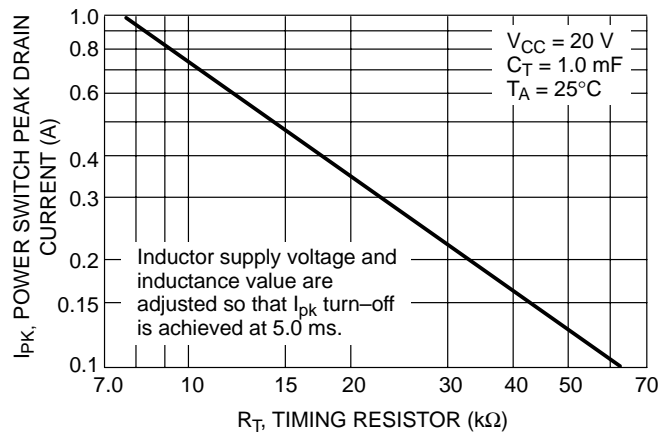


Figure 3. Power Switch Peak Drain Current versus Timing Resistor

A snubber is required across the primary winding because of the leakage inductance. An RCD snubber is implemented. The diode will be an MURS160 ultrafast diode. The resistor and capacitor are dependent upon the leakage inductance of the primary windings of the transformer. Preliminary values of the resistor and capacitor are derived from the following equations. The final values will be determined on the bench with an examination of the trade-offs between system efficiency and peak voltage.

$$V_{reflect} = (V_{out} + V_{diode}) \cdot \frac{N_p}{N_s} \quad (\text{eq. 10})$$

$$R = \frac{2 \cdot V_{max} \cdot (V_{max} - (V_{out} + V_{diode}) \cdot \frac{N_p}{N_s})}{L_{leak} \cdot I_p^2 \cdot \text{freq}} \quad (\text{eq. 11})$$

$$C = \frac{V_{max}}{V_{cr} \cdot R \cdot \text{freq}} \quad (\text{eq. 12})$$

See AND8023/D page 7 for further information on equations 11 and 12.(2)

V_{max} is the desired peak voltage on the power switch minus the bulk voltage, V_{cr} is the clamping ripple usually 20 V, and freq is the switching frequency of the MC33363B.

Most of the values necessary to calculate the secondary components are complete. To select the diode, the secondary current peak, the blocking voltage and the forward voltage drop are the main criteria. Ripple current from the transformer, the output voltage and the output ripple current are the criteria to select the output capacitors.

$$V_{block} = V_{out} + \left(V_{dc} \cdot \frac{N_s}{N_p} \right) \quad (\text{eq. 13})$$

$$C = \frac{1}{V_{ripple} \cdot \text{freq}} \quad (\text{eq. 14})$$

Utilizing the above equations and previous calculations, the diode needs to be rated at a minimum of 13 volts and 4.44 amperes. The MBRD835L is selected as the Schottky diode. The typical forward diode drop is 0.40 volts at 4.0 amperes. Rubycon capacitor YXG 6.3 volts 820 uF is chosen for the output filter. The voltage on the output capacitor needs to be rated twice the output voltage for safety. The rated maximum allowable ripple current of the capacitors is 0.865 amperes, so 2 capacitors in parallel will be used. A 1.0 uF ceramic capacitor will also be used in parallel with the aluminum electrolytics for fast transient response.

Equations 13 and 14 are used to calculate the auxiliary winding values. The auxiliary winding diode is an ultrafast recovery surface mount, MURS120. The diode is rated for 1 ampere and 200 volts blocking. The typical forward voltage drop is 0.6 volts. The higher forward voltage drop is acceptable due to the lower power requirements of the auxiliary versus the secondary output. The auxiliary winding filter is the capacitor on the V_{CC} pin.

Feedback Network

With the low voltage output requirement, a programmable precision reference with a low operating voltage is essential. Isolated feedback networks provide an additional challenge since optoisolator diode forward voltages remain at 1.25 volts. With a conventional circuit using the TL431 and TLV431, the lowest ideal power supply output voltages are 3.75 and 2.5 volts respectively. If the reference operating voltage range is reduced to 0.9 volts, the minimum output voltage is lowered to 2.15 volts. See Figure 4 for the circuit schematic.

This reduction is significant but will not satisfy a 1.8 volt output. By adding a resistor and PNP transistor, the reference voltage is reduced to the emitter-base voltage drop plus the programmable reference operating voltage. With the conventional TLV431 the minimum operating voltage is reduced to 1.95 volts. With the reduced reference voltage available in the NCP100, the minimum operating voltage is 1.6 volts. See Figure 5 for the circuit schematic. Figure 6 illustrates the output voltage ranges between the TL431, TLV431 and NCP100.

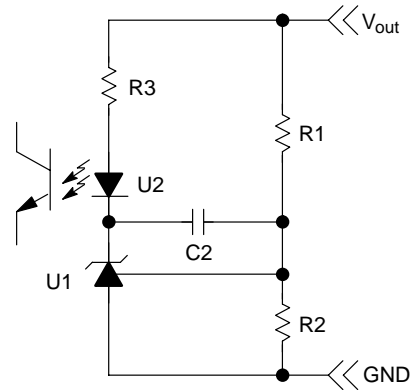


Figure 4. Traditional Isolated Feedback Network

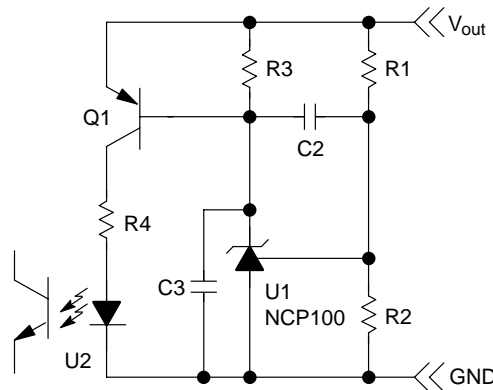


Figure 5. 1.8 Volt Isolated Feedback Network

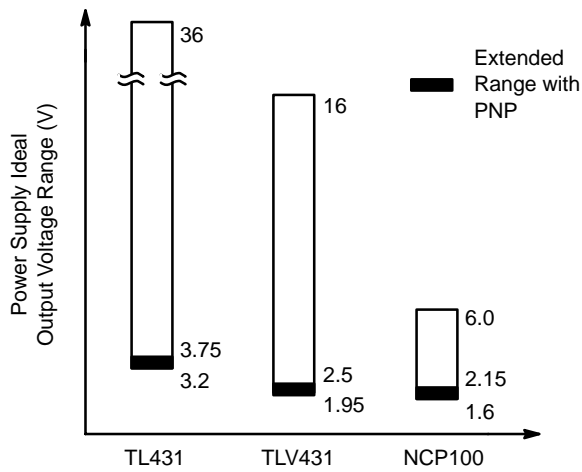


Figure 6. Power Supply Output Voltage Range Comparison for Programmable Precision References

Figure 5 illustrates an additional 1.0 uF capacitor, C3, which is required for normal operation of the NCP100. Bandwidth concerns are raised when a capacitor is placed from the cathode to anode of the precision reference. Since the PNP transistor is used to drive the optocoupler, the cathode current of the NCP100 is amplified by the beta of the transistor therefore minimizing the effects of C3. R3 is a factor in the overall gain of the feedback network and R4 limits the current into the diode of the optocoupler. The roll-off frequency of the feedback network is related to R1 and C2 only. Figure 7 shows the circuit schematic, the gain and the phase of the TLV431 and the NCP100 with and without the PNP.

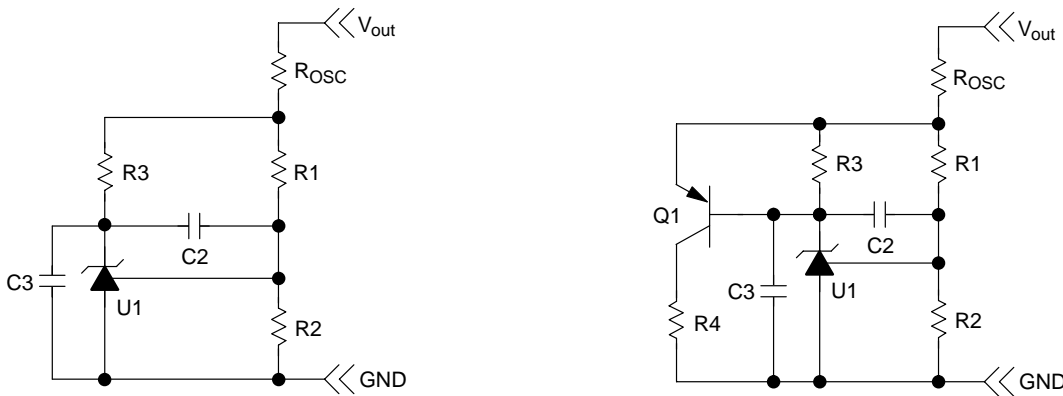


Figure 7. Schematics for Gain and Phase Comparison of TLV431 and NCP100
Note: C3 only for NCP100

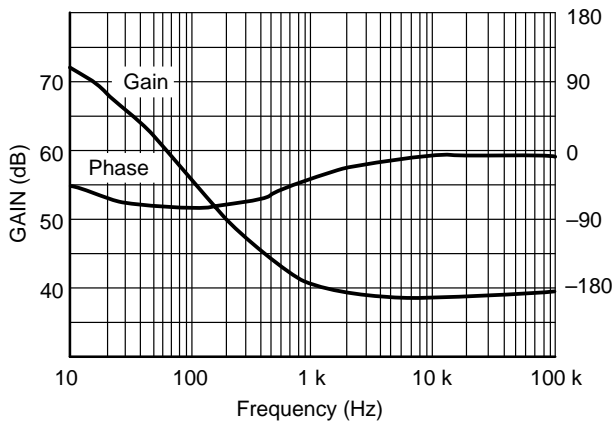


Figure 8. TLV431 with PNP

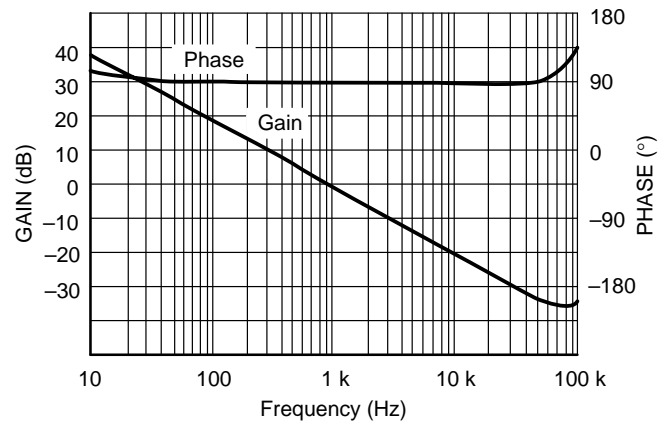


Figure 9. TLV431 without PNP

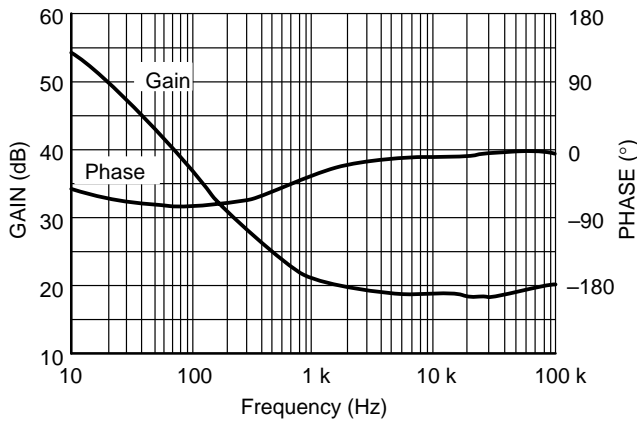


Figure 10. NCP100 with PNP

The values for the components in Figure 7 are as follows. $R_{osc} = 12$ Ohms, $R_1 = 12$ k Ohms, R_2 is adjustable dependant upon U_1 , R_3 and $R_4 = 100$ Ohms, $C_2 = 0.01$ μ F, and $C_3 = 1.0$ μ F.

The open loop gain and phase test injects the oscillator signal across R_{osc} . The reference voltage is measured between R_{osc} and R_1 to ground. The test voltage is measured at the collector of Q_1 for the PNP circuit or at the cathode of U_1 for the traditional circuit. V_{out} is set to 4.0 volts and R_2 is adjusted so the DC test voltage is at the center of its range. R_2 is the only component adjusted since it is neglected for AC analysis.

There are negligible differences for the open loop gain and phase of the NCP100 with C_3 compared to the TLV431 without C_3 in the PNP circuit. Therefore there is no penalty with the added C_3 in the NCP100 circuit.

If the PNP transistor is removed, the TLV431 open loop responses with and without C_3 are very similar with a single pole roll-off. The pole appears to be near 4 Hz.

The NCP100 without the PNP has flat gain until a pole at 1.3 kHz due to R_1 and C_2 . The plot remains at

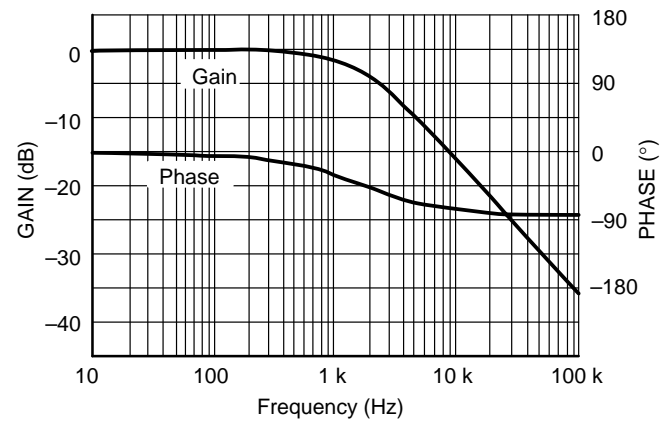


Figure 11. NCP100 without PNP

-20 dB/decade and -90 degree phase margin until well beyond 100 kHz.

The overall gain of the feedback network can be limited due to R_1 , R_3 and C_2 in Figure 7. Varying these components will modify the maximum gain of the system.

Since the NCP100 provides a single pole roll-off, the MC33363B compensation pin will not provide this function. The feedback pin of the MC33363B is connected directly to the reference pin. This will keep the output of the error amp low. A diode is in series with the output of the op-amp allowing the compensation pin to directly control the feedback to the oscillator ramp. The compensation pin is connected to the feedback pin via a 2.7 k Ohm resistor. The collector of the optocoupler is also connected to the compensation pin and the emitter is grounded. This completes the design of the system. See Figure 14 for the board layout of the schematic in Figure 1 and Table 1 for the component values.

Results

Table A lists the system parameters. Figure A illustrates the single pole feedback from the circuit in Figure 1 and the layout in Figure 13. The output current step response has a negligible effect on the output voltage. Figure B is the output

step response of the circuit from 20% to 100%. The circuit performance achieves the desired results for the off-line converter. The efficiency is low, but is expected for a low power off-line converter.

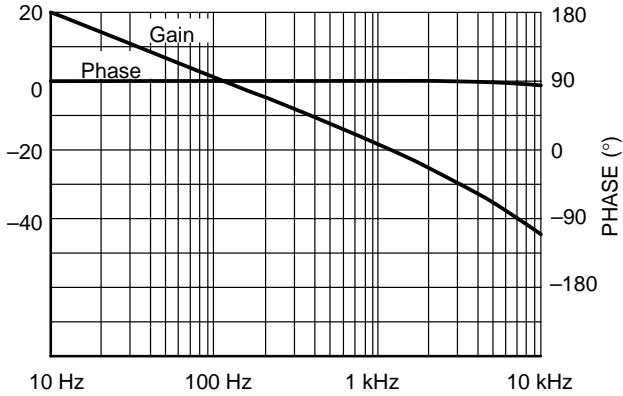


Figure 12. Gain and phase results from the 1.8 volt 1 ampere off-line converter.

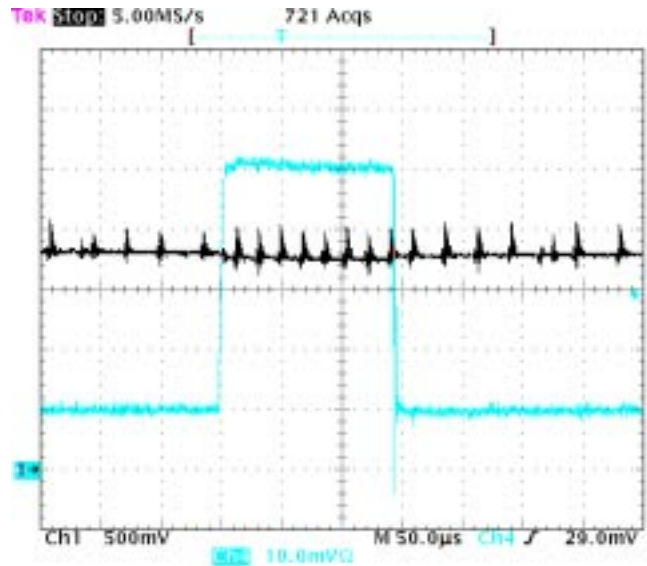


Figure 13. Output step response from 0.2 to 1.0 Amperes.

Table 1. System Parameters

Parameter	Value
Operating Frequency	110 kHz
Primary Peak Current Limit	230 mA
Efficiency at 1.0 Ampere Load, 120 VACin	58.6 %
Efficiency at 1.0 Ampere Load, 230 VACin	57.2 %
Load Regulation, 120 VACin, 0.1 to 1.0 Ampere Load	22 mV
Line Regulation, 1.0 Ampere Load, 90 to 265 VACin	1 mV
Low Frequency Output Ripple, 1.0 Ampere Load	43 mVpp
High Frequency Output Ripple, 1.0 Ampere Load	350 mVpp

AND8031/D

Table 2. Component Values of Off-Line Power Supply

Reference	Part	Description	Manufacturer
C1	1 n	X Capacitor	
C2	4.7 u	400 Volt Bulk YK Series	Rubycon
C3	390 p	SMT Ceramic 0805	TDK
C4,C10,C12	1 u	SMT Ceramic 1206	TDK
C5	47 p	1 k Volt Ceramic Disc	
C6	10 u	25 Volt MS5 Series	Rubycon
C7,C8	820 u	6.3 Volt YXG Series	Rubycon
C9,C11	0.1 u	Y Capacitor	TDK
D1,D2,D3,D4	1N4005	Standard Recovery Rectifiers	ON Semiconductor
D5	MURS160T3	600 Volt Ultra Fast Rectifier	ON Semiconductor
D6	MURS120T3	200 Volt Ultra Fast Rectifier	ON Semiconductor
D7	MBRD835L	35 Volt Schottky Rectifier	ON Semiconductor
JP1,JP2		3.5 mm 2 Pole Connector	Wieland
Q1	BC858ALT1	Small Signal PNP	ON Semiconductor
R1,R2	4.7	SMT 0805	
R3	30 k	SMT 0805	
R4	3.9 k	1/4 Watt	
R6	2.7 k	SMT 0805	
R9	12 k	SMT 1206	
R10	7.5 k	SMT 0805	
R11	10	SMT 0805	
R12	120	SMT 0805	
U1	MC33363BDW	SMPS Controller with FET	ON Semiconductor
U2	Transformer	Transformer EE16 CTX22-14966	Coiltronics
U3	SFH615-4	4 pin Optocoupler	
U4	NCP100	Programmable Precision Reference	ON Semiconductor

References

1. Brown, Marty, Power Supply Cookbook, Newton, MA, Butterworth-Heinemann, 1994.
2. Basso, Christophe, Application Note: "AND8023: Implementing the NCP1200 in Low-Cost AC/DC Converters," ON Semiconductor, August, 2000.

Notes

Notes

Notes

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