

Micro Family Isolated Remote Sense

Introduction

Vicor’s Micro DC-DC converters do not have remote sense pins. Nevertheless, remote sense can be achieved by employing external circuitry. A circuit is proposed that senses the voltage at the load and adjusts the converter output voltage to compensate for the voltage drop in the leads / traces. It also provides isolation of the sense leads / traces that is beneficial in high-noise applications.

Design Considerations

DC Considerations:

The Vicor Micro module’s output can be trimmed up to a maximum of 10% over the nominal output voltage. This limits the amount of lead / trace resistance the isolated remote sense circuit can correct for. Maximum round trip lead resistance at full load is governed by the following formula:

$$R_{\text{lead(max)}} = \frac{1.1 V_{\text{nom}} - V_{\text{POL}}}{0.9 I_{\text{max}}}$$

V_{nom} : the nominal output voltage of the converter

V_{POL} : the voltage at the point-of load

I_{max} : converter rated output power $\div V_{\text{nom}}$

Note: As the module output voltage is trimmed up, the output current drawn from the module must be reduced proportionally. This is why I_{max} is multiplied by 90% when the module is trimmed up 10%.

The proposed circuit may cause oscillation with large output capacitance at light loads. A minimum load of 10% of the maximum is recommended.

AC Considerations:

As the load is moved further away from the output of the converter, lead / trace impedance will increase. Since the Micro module has local sensing at the output terminals, this impedance causes the voltage that the converter regulates to be different from the voltage seen by the load. This remote sense circuit solves this problem by putting the lead / trace impedance inside the control loop of the converter. Figure 1 shows a very lossy distribution network between the converter and point-of-load.

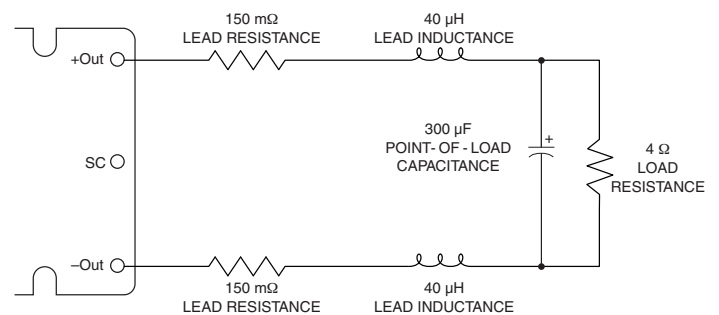


Figure 1: Example of long leads between converter and load (First-order model)

Figure 2 illustrates the effect of this parasitic network in the time domain. The step response shows the point-of-load voltage is delayed with respect to the converter output. This limits how fast the converter can correct for voltage changes at the load.

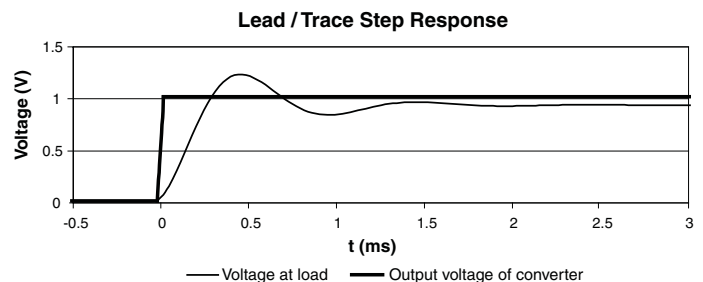


Figure 2: Lead / trace step response

In the frequency domain this delay results in phase shift that increases with frequency. This is demonstrated by the Bode plot of this network shown in Figure 3.

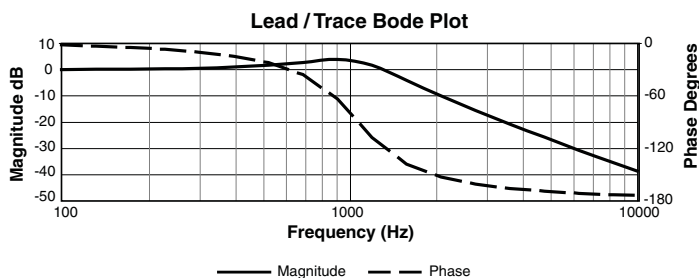


Figure 3: Lead / trace model Bode plot

Whenever phase shift is introduced into a control network, the stability of that loop will be degraded. This occurs because the control loop is trying to correct for conditions that have already occurred. As phase shift approaches -180° , the control loop falls so far behind that the system will begin to oscillate. This consideration will always be an issue when remote sense is used, regardless of whether it is realized with built-in sense leads or an external circuit.

To prevent instability the designer must reduce the loop gain below unity before phase shift becomes significant. This leads to a fundamental trade off in control system design because reducing the system bandwidth will degrade its transient response.

Remote Sense Circuit Functional Description

As shown in Figure 5, an op-amp with a built-in precision reference (U3) maintains voltage regulation at the point-of-load. The reference voltage is scaled up from 200 mV with the external resistive network formed by R7 and R8 to 1.245 V. The reference has a compensation capacitor (C2) that slows the ramp-up time to control turn-on overshoot.

The op-amp compares this reference voltage to the point-of-load voltage scaled by R9 and R10. The output of the op-amp drives the cathode of an optocoupler (U1). The optocoupler isolates noise voltages that are present at the negative output lead thereby keeping them from appearing on the SC pin. The speed at which the external loop can respond to load transients is determined by the compensation capacitor (C3) and the parallel combination of the voltage sensing resistors (R9 and R10). R11 allows C3 to fully discharge when power is removed from the circuit.

The optocoupler is connected to the SC pin via resistors R1 and R2. These resistors program the maximum output voltage at the converter (R1) and the minimum output voltage of the converter (R2). Figure 4 shows how the internal connections of the SC pin form a RC filter that will limit overall system bandwidth.

The op-amp and the optocoupler are both powered from the output of the converter via a shunt regulator (U2) that is programmed to provide a 2 V rail.

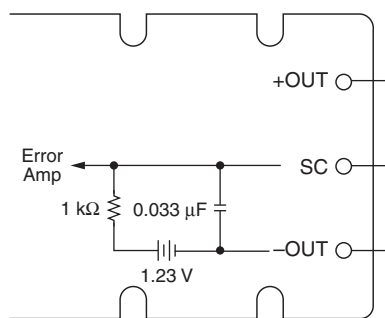


Figure 4: Model of the SC pin

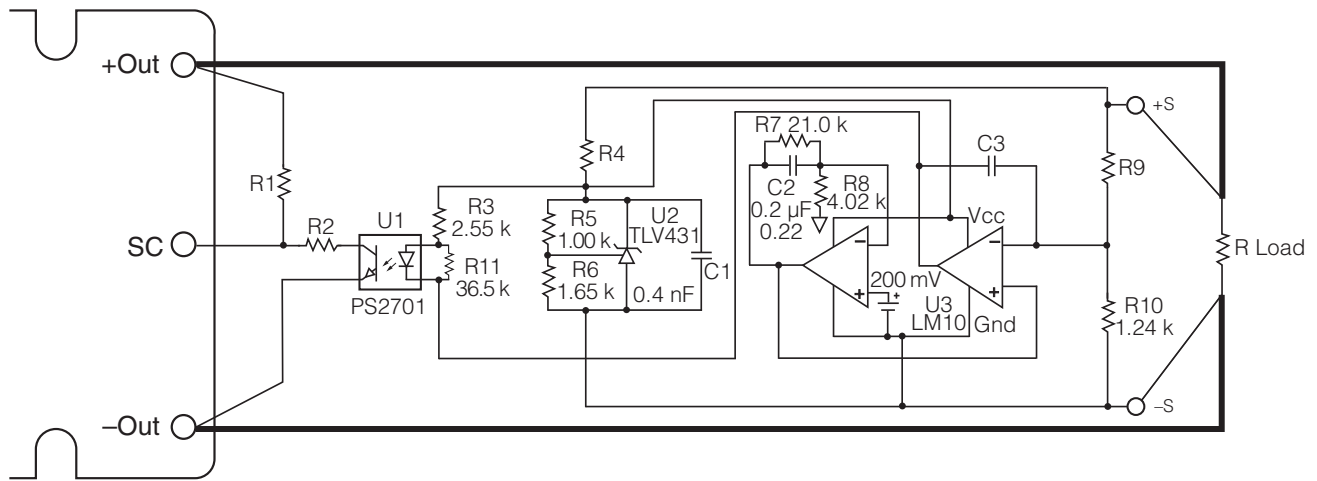


Figure 5: Circuit schematic

Component Selection

Table 1 gives resistor values for some common output voltages. For applications that are not listed formulas for calculating resistor values are also given.

V _{nom} (V)	V _{out(max)} (V)	V _{out(min)} (V)	R4 (kΩ)	R1 (kΩ)	R2 (kΩ)	R9 (kΩ)
3.3	3.63	2.97	0.091	18.7	3.57	2.05
5	5.5	4.5	0.2	34	3.57	3.74
8	8.8	7.2	0.39	60.4	3.57	6.65
12	13.2	10.8	0.68, 0.5W	95.3	3.57	10.7
15	16.5	13.5	0.82, 0.5W	124	3.57	13.7
24	26.4	21.6	1.5, 0.5W	205	3.57	22.6
28	30.8	25.2	1.8, 1.0W	237	3.57	26.7
36	39.6	32.4	2.2, 1.0W	309	3.57	34.8
48	52.8	43.2	3, 1.0W	422	3.57	46.4

Table 1: Resistor values

All resistors are ¼ W unless otherwise specified

V_{nom}: the nominal output voltage of the converter and point-of-load voltage

V_{out(max)}: the maximum output voltage of the converter that supplies the load and lead / trace loss

V_{out(min)}: the minimum output voltage of the converter typically set to 90% of V_{nom}

Solving for R1

The value of R1 is governed by the following formula:

$$R1 = 1 \text{ k}\Omega \frac{(V_{out(max)} - 1.23 \text{ V})V_{nom}}{1.23 \text{ V}(V_{out(max)} - V_{nom})} - 1 \text{ k}\Omega$$

Solving for R2

R2 can be calculated as follows, where V_{CEsat} is the saturation voltage of the optocoupler given by the manufacturer (0.3 V for the NEC PS2701):

$$R2 = \frac{\frac{V_{out(min)}1.23 \text{ V}}{V_{nom}} - V_{CEsat}}{\frac{V_{out(min)}}{R1} \left(1 - \frac{1.23 \text{ V}}{V_{nom}}\right) + \frac{1.23 \text{ V}}{1 \text{ k}\Omega} \left(1 - \frac{V_{out(min)}}{V_{nom}}\right)}$$

Solving for R4

With different module output voltages R4 will need to be changed such that current being fed into the TLV431 regulator is approximately 15 mA. The following equations can be used to find the appropriate value for R4 and its power dissipation P_{R4}:

$$R4 = \frac{V_{nom} - 2 \text{ V}}{15 \text{ mA}}$$

$$P_{R4} = (V_{nom} - 2 \text{ V})15 \text{ mA}$$

Solving for R9 and R10

The value of R10 should be 1.24 kΩ to obtain a reasonable value for the parallel combination of R9 and R10. R9 can be calculated as follows:

$$R9 = R10 \left(\frac{V_{nom}}{1.245 V} - 1 \right)$$

Solving for C3

For good stability the bandwidth of a remote sense circuit must be lower than the frequency where phase shift from the leads / traces becomes significant. The appropriate integrator crossover frequency can be estimated with a first-order model of the leads / traces. For many applications 200 Hz makes a good starting value (C3 = 0.68 μF). The optimal value will depend on design requirements. Figure 6 can be used to find C3 once the desired crossover frequency is known.

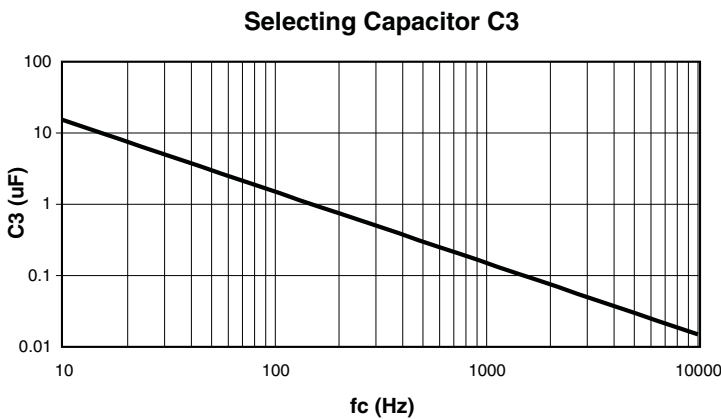


Figure 6: Crossover frequency vs. Capacitance

Solving for C2

C2 programs the ramp-up time of the remote sense circuit's reference. To minimize overshoot this should be longer than the 4 ms start up ramp of the converter (C2 = 0.22 μF). The effect of changing C2 is shown in the waveforms section.

The following table gives component values for a circuit configured for a point-of-load voltage of 3.3 V.

Ref Des	Part Description	Rating
R1	18.7 kΩ	1/4 W
R2	3.57 kΩ	1/4 W
R3	2.55 kΩ	1/4 W
R4	91 Ω	1/4 W
R5	1 kΩ	1/4 W
R6	1.65 kΩ	1/4 W
R7	21 kΩ	1/4 W
R8	4.02 kΩ	1/4 W
R9	2.05 kΩ	1/4 W
R10	1.24 kΩ	1/4 W
R11	36.5 kΩ	1/4 W
C1	470 pF	100 V
C2	0.22 μF	16 V
C3	0.68 μF	16 V
U1	NEC PS2701 (Digi-Key #PS2701-1-ND)	NA
U2	TI TLV431 (Digi-Key #296-10727-5-ND)	NA
U3	National LM10 (Digi-Key #LM10CWM-ND)	NA

Table 2: Parts list for 3.3 V module

Circuit Waveforms

Transient Response

Figures 7, 8, and 9 show the response to a load current step. C3 is sized for a 200 Hz crossover frequency giving a clean transient response with no oscillation.

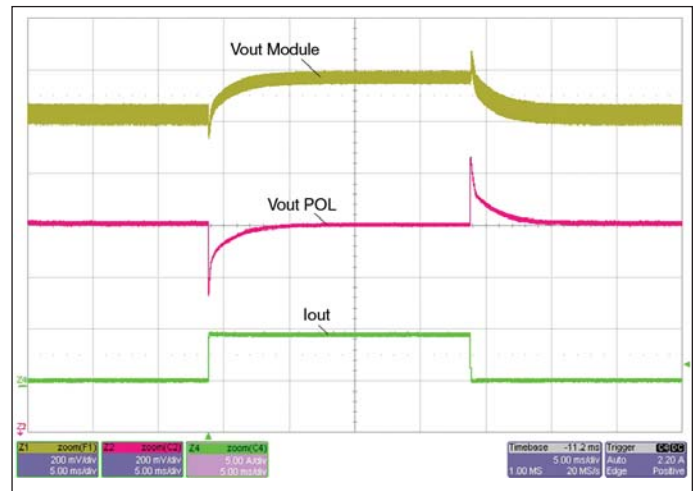


Figure 7: Load step 0.5 A to 5 A to 0.5 A, C3 = 0.64 μF, 300 μF at point-of-load (V48C3V3E75B)

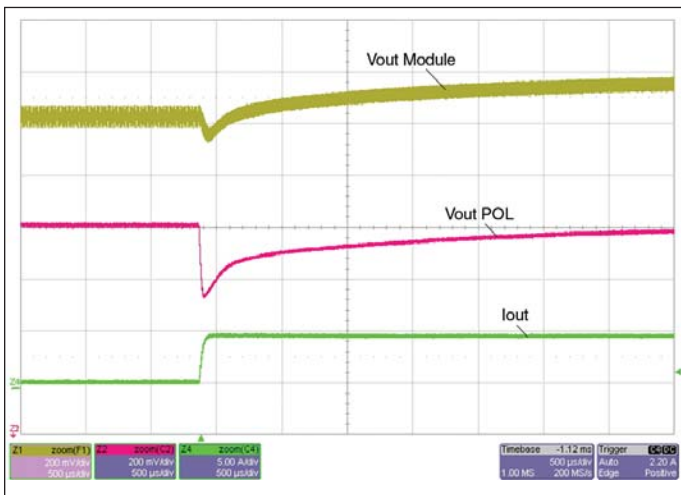


Figure 8: Load step 0.5 A to 5 A, $C_3 = 0.64 \mu\text{F}$, $300 \mu\text{F}$ at point-of-load (V48C3V3E75B)

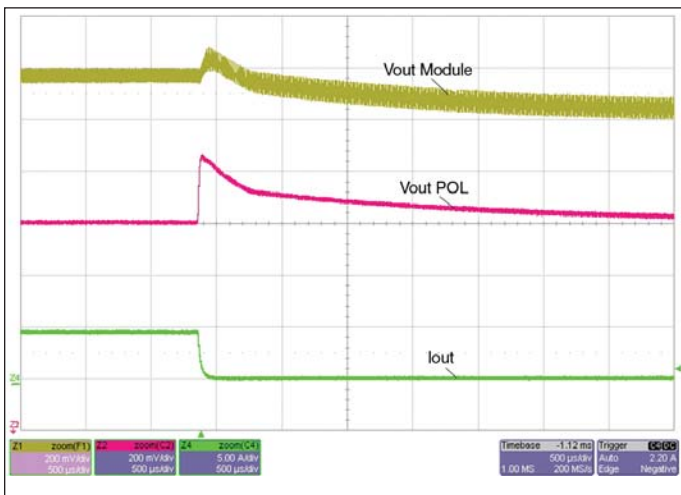


Figure 9: Load step 5 A to 0.5 A, $C_3 = 0.64 \mu\text{F}$, $300 \mu\text{F}$ at point-of-load (V48C3V3E75B)

Transient Response Improperly Sized C_3

Figures 10 and 11 show the response to a load step with C_3 undersized. They illustrate how both C_3 and the point-of-load capacitance contribute to the circuit's closed loop response.

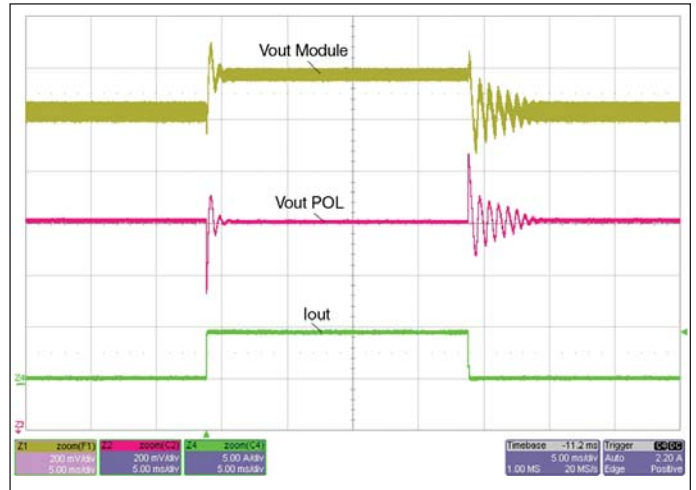


Figure 10: Load step 0.5 A to 5 A to 0.5 A, $C_3 = 0.033 \mu\text{F}$, $300 \mu\text{F}$ at point-of-load (V48C3V3E75B)

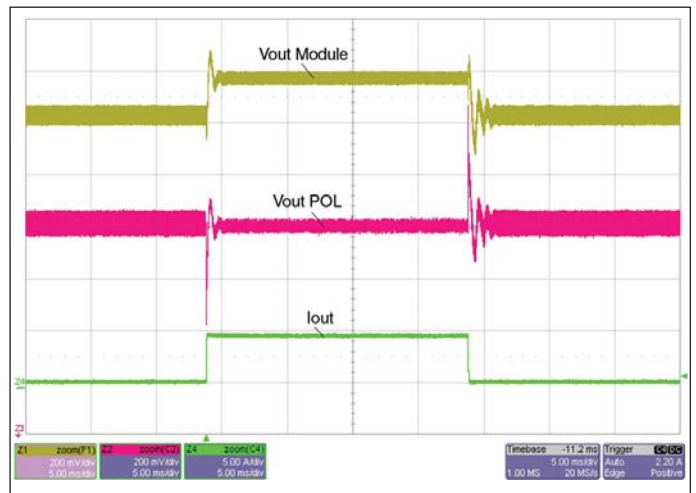


Figure 11: Load step 0.5 A to 5 A to 0.5 A, $C_3 = 0.033 \mu\text{F}$, no point-of-load capacitance (V48C3V3E75B)

Effect of C2 on Start up

Figure 12 shows a good intermediate value for C2 which results in a reasonable start up time and eliminates overshoot. Depending on the application C2 may need to be resized. In Figure 13, C2 is intentionally undersized causing the reference to come up too fast and the point-of-load voltage to overshoot. Figure 14 shows start up with C2 oversized resulting in a long delay before the nominal voltage is reached.

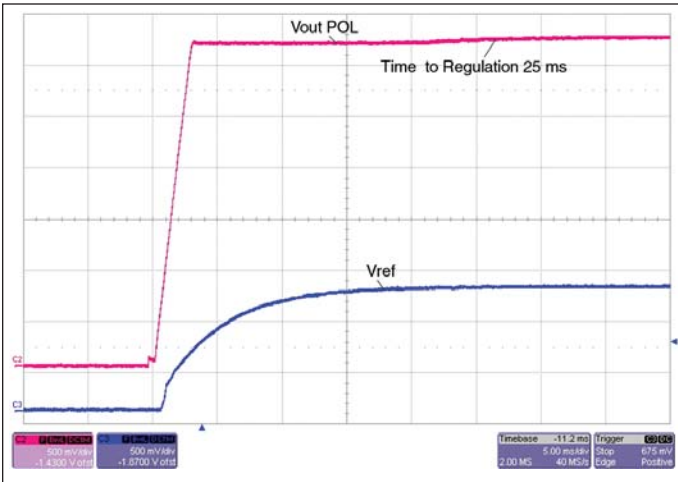


Figure 12: Start up C2 = 0.21 μ F (V48C3V3E75B)

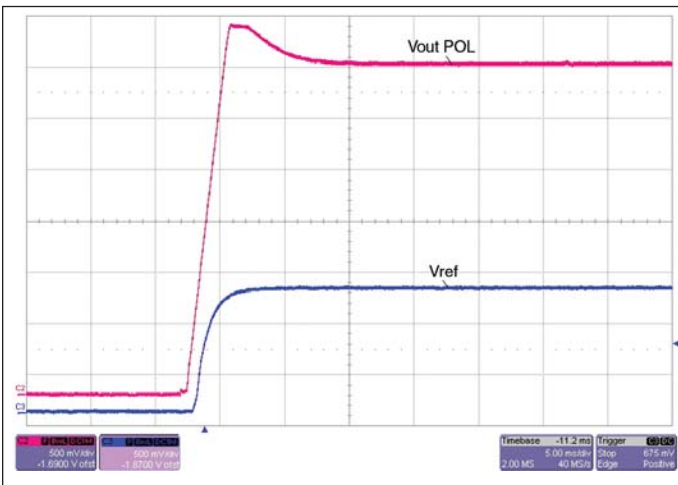


Figure 13: Start up C2 = 0.047 μ F (V48C3V3E75B)

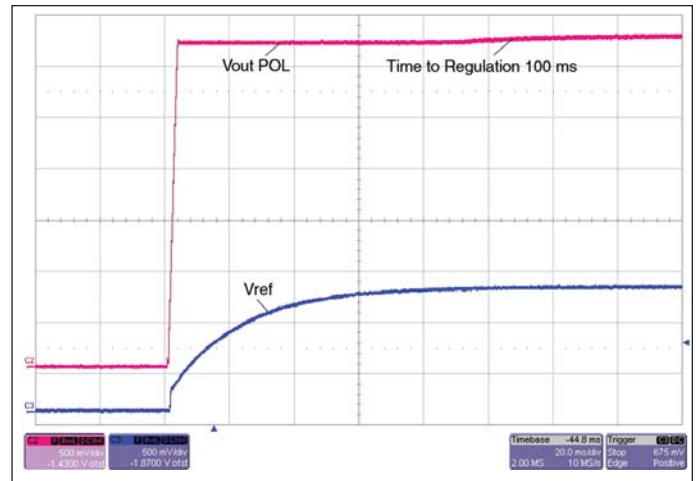


Figure 14: Start up C2 = 1.0 μ F (V48C3V3E75B)

For more information on remote sense capabilities, please contact Vicor's Applications Engineers at 1-800-927-9474 or vicorpower.com/support/ for worldwide assistance.