Positive Floating Hot-Swap Power Manager

Check for Samples: UCC2917, UCC3917

FEATURES

- **Manages Hot-Swap of 15 V and Above**
- **Precision Fault Threshold**
- **Programmable Average Power Limiting**
- **Programmable Linear Current Control**
- **Programmable Overcurrent Limit**
- **Programmable Fault Time**
- **Internal Charge Pump to Control External N-channel MOSFET Device**
- **Fault Output and Catastrophic Fault Indication**
- **Fault Mode Programmable to Latch or Retry**
- **Shutdown Control**
- **Undervoltage Lockout**

APPLICATIONS

- **390-V DC Distribution**
- **General High-Voltage Power Management**

DESCRIPTION

The UCCx917 family of positive-floating hot-swap managers provides complete power management, hot-swap, and fault handling capability. The voltage limitation of the application is only restricted by the external component voltage limitations. The device provides its own supply voltage via a charge pump referenced to VOUT. The onboard 10-V shunt regulator protects the device from excess voltage. The devices also have catastrophic fault indication to alert the user that the ability to shut off the output N-channel MOSFET has been bypassed. All control and housekeeping functions are integrated and externally programmable. These include the fault current level, maximum output sourcing current, maximum fault time, soft-start time, and average N-channel MOSFET power limiting.

The fault level across the current-sense amplifier is fixed at 50 mV to minimize total drop out. Once 50 mV is exceeded across the current-sense resistor, the fault timer starts. The maximum allowable sourcing current is programmed with a voltage divider from the VREF/CATFLT pin to generate a fixed voltage on the MAXI pin. The current level at which the output appears as a current source is equal to V_{MAXI} divided by the current-sense resistor. If desired, a controlled current startup can be programmed with a capacitor on MAXI.

When the output current is below the fault level, the output device is switched on with full gate drive. When the output current exceeds the fault level, but is less than maximum allowable sourcing level programmed by MAXI, the output remains switched on, and the fault timer starts charging the timing capacitor C_T. Once C_T charges to 2.5 V, the output device is turned off and attempts either a retry sometime later or waits for the state on the LATCH pin to change if in latch mode. When the output current reaches the maximum sourcing current level, the output device appears as a current source.

ORDERING INFORMATION

T_J	PACKAGED DEVICES		
	DIP (J)	DIP (N)	SOIC (D)
–40°C to 85°C	UCC2917J	UCC2917N	UCC2917D
0°C to 70°C	UCC3917J	UCC3917N	UCC3917D

These devices have limited built-in ESD protection. The leads should be shorted together or the device placed in conductive foam during storage or handling to prevent electrostatic damage to the MOS gates.

ABSOLUTE MAXIMUM RATINGS(1)

over operating free-air temperature range (unless otherwise noted)

		VALUE		UNIT
		MIN	MAX	
Output Current	Supply (2)		20	mA
	$\overline{\text{SHTDWN}}$, $\overline{\text{LATCH}}$, VREF (2)		–500	µA
Line Current	PLIM		10	mA
Input voltage	MAXI		V_{DD} + 0.3	V
Junction temperature, T_J		–55	150	°C
Storage temperature, T_{stg}		–65	150	°C
Lead temperature (Soldering, 10 sec.)			300	°C

(1) Stresses beyond those listed under "absolute maximum ratings" may cause permanent damage to the device. These are stress ratings only, and functional operation of the device at these or any other conditions beyond those indicated under "recommended operating conditions" is not implied. Exposure to absolute-maximum-rated conditions for extended periods may affect device reliability.colsep

(2) Currents are positive into, negative out of the specified terminal. Consult the *Packaging* section of the *Interface Products Data Book* (TI Literature Number SLUD002) for thermal limitations and considerations of package.

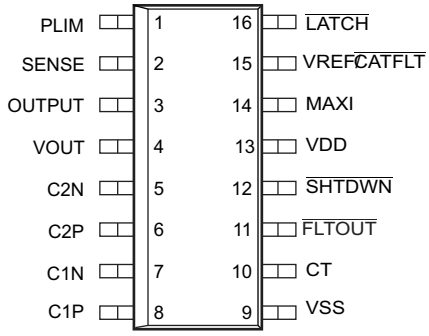
ELECTRICAL CHARACTERISTICS

0°C ≤ T_A ≤ 70°C for the UCC3917, –40°C to 85°C for the UCC2917, C_{CT} = 4.7 nF, T_A = T_J, all voltages are with respect to VOUT, current is positive into and negative out of the specified terminal, (unless otherwise noted)

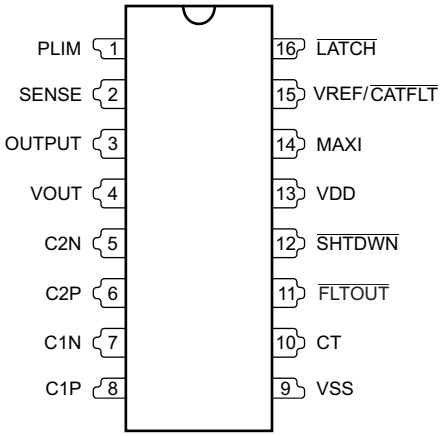
PARAMETER		TEST CONDITIONS	MIN	TYP	MAX	UNIT
INPUT SUPPLY						
I_{DD}	Supply current(1)	From VOUT	4.0	5	11	mA
	UVLO turn on threshold		7.9	8.8	9.7	V
	UVLO off voltage		5.5	6.5	7.5	V
	VSS regulator voltage		–6	–5	–4	V
FAULT TIMING						
	Overcurrent threshold	T_A = 25°C	47.5	50	53	mV
		Over operating temperature	46	50	54	mV
	Overcurrent input bias			50	500	
	CT charge current bias	V_{CT} = 1 V	–78	–50	–28	µA
	CT catastrophic fault threshold		3.4		4.5	V
	CT fault threshold		2.25	2.5	2.75	V
	CT reset threshold		0.32	0.5	0.62	V
D	Output duty cycle	Fault condition	1.7%	2.7%	3.7%	
OUTPUT						
V_{OH}	High-level output voltage	I_{OUT} = 0	6	8	10	V
		I_{OUT} = –100 µA	5	7	9	
V_{OL}	Output low voltage	I_{OUT} = 500 µA		0.03	0.50	V
		I_{OUT} = 1 mA		0.6	0.9	
LINEAR CURRENT						
	Sense control voltage	V_{MAXI} = 100 mV	85	100	115	mV
		V_{MAXI} = 400 mV	370	400	430	mV
I_{BIAS}	Input bias current	V_{MAXI} = 200 mV		50	500	nA
$\overline{\text{SHUTDOWN}}$						
	Shutdown threshold		2.0	2.4	2.8	V
	Input current	$V_{\overline{SHTDWN}}$ = 0 V	24	40	60	µA
	Shutdown delay			100	500	ns
$\overline{\text{LATCH}}$						
$V_{\overline{LATCH}}$	Latch threshold		1.7	2	2.3	V
	Input current	$V_{\overline{LATCH}}$ = 0 V	24	40	60	µA
$\overline{\text{FLTOUT}}$						
	Fault output high	V_{CT} = 0 V, I_{SOURCE} = 0 µA	6	8	10	V
	Fault output low	V_{CT} = 5 V, I_{SINK} = 200 µA		0.01	0.05	V
POWER LIMITING						
	V_{SENSE} regulator voltage	I_{PLIM} = 64 µA	4.5	5.	5.5	V
	Duty cycle control	I_{PLIM} = 64 µA	0.6%	1.2%	1.7%	
		I_{PLIM} = 1 mA	0.045%	0.1%	0.2%	
VREF/$\overline{\text{CATFLT}}$						
	V_{REF} regulator voltage		4.5	5.	5.5	V
	Fault output low	$I_{VREF/\overline{CATFLT}}$ = 5 mA		0.22	0.50	V
I_{SINK}	Output sink current	V_{CT} = 5 V, $V_{VREF/\overline{CATFLT}}$ = 5 V	15	40	70	mA
	Overload comparator threshold	Relative to MAXI	110	200	290	mV

(1) Set by user using the R_{SS} resistor.

D PACKAGE
16 PINS
(TOP VIEW)



J, N PACKAGE
16 PINS
(TOP VIEW)



DEVICE INFORMATION

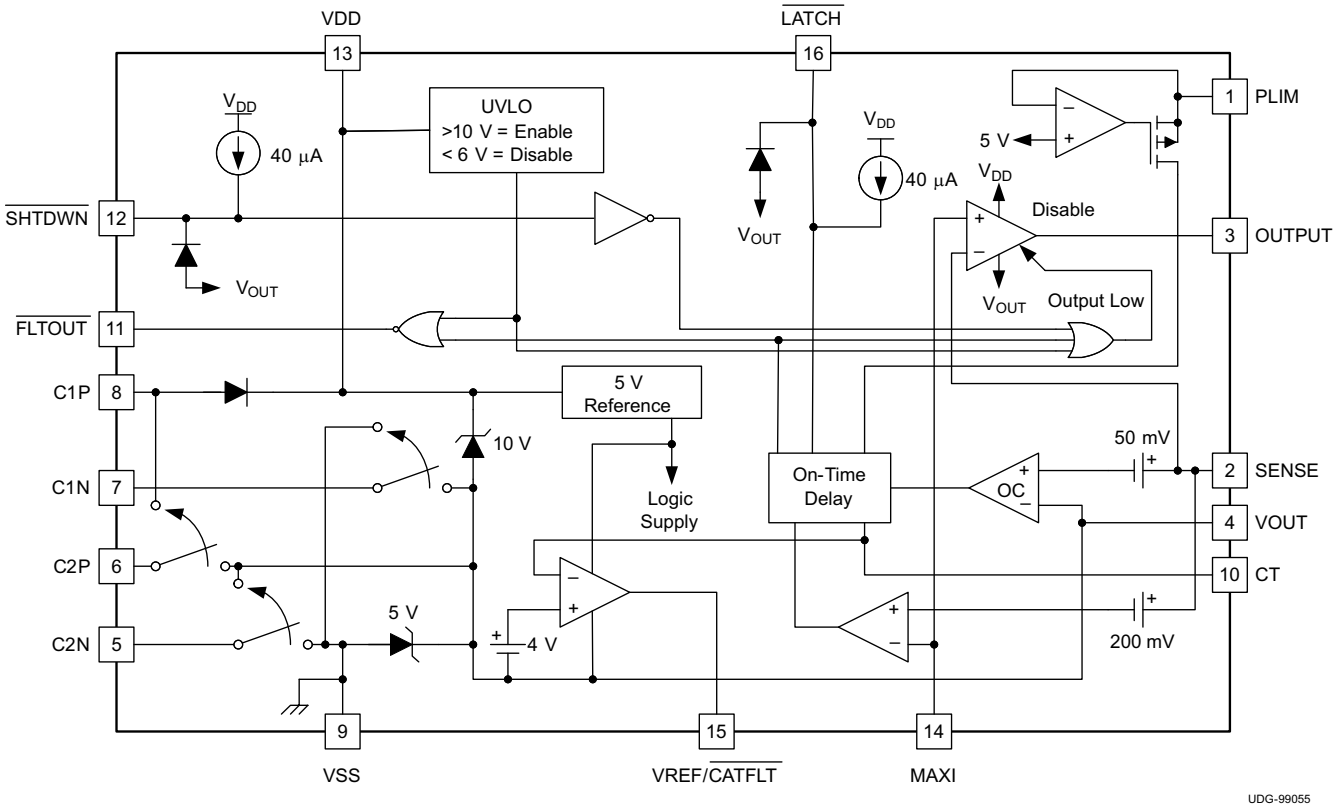
PIN DESCRIPTIONS

PIN		I/O	DESCRIPTION
NAME	NO.		
C1N	7	I	Negative side of the upper charge-pump capacitor.
C1P	8	I	Positive side of the upper charge-pump capacitor.
C2N	5	I	Negative side of the lower charge-pump capacitor.
C2P	6	I	Positive side of lower charge-pump capacitor.
CT	10	I	A capacitor is connected to this pin to set the fault time. The fault time must be more than the time to charge the external load capacitance (see application information).
FLTOUT	11	O	Provides fault output indication. Interface to this pin is usually performed through level-shift transistors. Under a non-fault condition, $\overline{\text{FLTOUT}}$ is pulled to a high state. When a fault is detected by the fault timer or the undervoltage lockout, this pin is driven to a low state, indicating the output N-channel MOSFET is in the off state.
$\overline{\text{LATCH}}$	16	I	Pulling this pin low causes a fault to latch until this pin is brought high or a power-on reset is attempted. However, pulling this pin high before the reset time is reached does not clear the fault until the reset time is reached. Keeping $\overline{\text{LATCH}}$ high results in normal operation of the fault timer. Users should note there is an R-C delay dependent upon the external capacitor at this pin.
MAXI	14	I	Programs the maximum-allowable sourcing current. Since VREF/$\overline{\text{CATFLT}}$ is a regulated voltage, a voltage divider can be derived to generate the program level for MAXI. The current level at which the output appears as a current source is equal to the voltage on MAXI divided by the current-sense resistor. If desired, a controlled current start-up can be programmed with a capacitor on MAXI (to VOUT), and a programmed start delay can be achieved by driving the shutdown with an open collector/drain device into an RC network.
OUTPUT	3	O	Gate drive to the N-channel MOSFET pass element.
PLIM	1	I	This feature ensures that the average external N-channel MOSFET power dissipation is controlled. A resistor is connected from this pin to the drain of the external N-channel MOSFET pass element. When the voltage across the N-channel MOSFET exceeds 5 V, current flows into PLIM, which adds to the fault timer charge current, reducing the duty cycle from the 3% level.
SENSE	2	I	Input voltage from the current-sense resistor. When there is greater than 50 mV across this pin with respect to VOUT, a fault is sensed, and the C_{CT} capacitor starts to charge.
$\overline{\text{SHTDWN}}$	12	I	This pin provides shutdown control. Interface to this pin is usually performed through level-shift transistors. When shutdown is driven low, the output disables the N-channel MOSFET pass device.
VOUT	4	I	Ground reference for the device.

PIN DESCRIPTIONS (continued)

PIN		I/O	DESCRIPTION
NAME	NO.		
VDD	13	I	Power to the device is supplied by an external current-limiting resistor on initial power up or if the load is shorted. As the load voltages rises (VOUT), a small amount of power is drawn from VOUT by an internal charge pump. The charge pump's input voltage is regulated by an on-device 5-V zener. Power to VDD is supplied by the charge pump under normal operation (i.e., external FET is on).
VREF/$\overline{\text{CATFLT}}$	15	O	This pin primarily provides an output reference for the programming of MAXI. Secondarily, it provides catastrophic fault indication. In a catastrophic fault, when the device unsuccessfully attempts to shutdown the N-channel MOSFET pass device, this pin pulls to a low state when CT charges above the catastrophic fault threshold. A possible application for this pin is to trigger the shutdown of an auxiliary FET in series with the main FET for redundancy.
VSS	9	I	Negative reference out of the device. This pin is normally current fed via a resistor to load ground.

FUNCTIONAL BLOCK DIAGRAM



APPLICATION INFORMATION

FAULT TIMING

Figure 1 shows the detailed circuitry for the fault timing function of the UCC3917. For simplicity, first consider a typical fault mode where the overload comparator and the current source I3 do not come into play. A typical fault occurs once the voltage across the current-sense resistor, R_S, exceeds 50 mV. This causes the overcurrent comparator to trip and the timing capacitor to charge with current source I1 plus the current from the power limiting amplifier, or PLIM amplifier. The PLIM amplifier is designed to only source current into the CT pin once the voltage across the output FET exceeds 5 V. The current I_{PL} is related to the voltage across the FET as shown in Equation 1.

$$I_{PL} = \frac{(V_{IN} - V_{VOUT})}{R_{PL}} \quad (1)$$

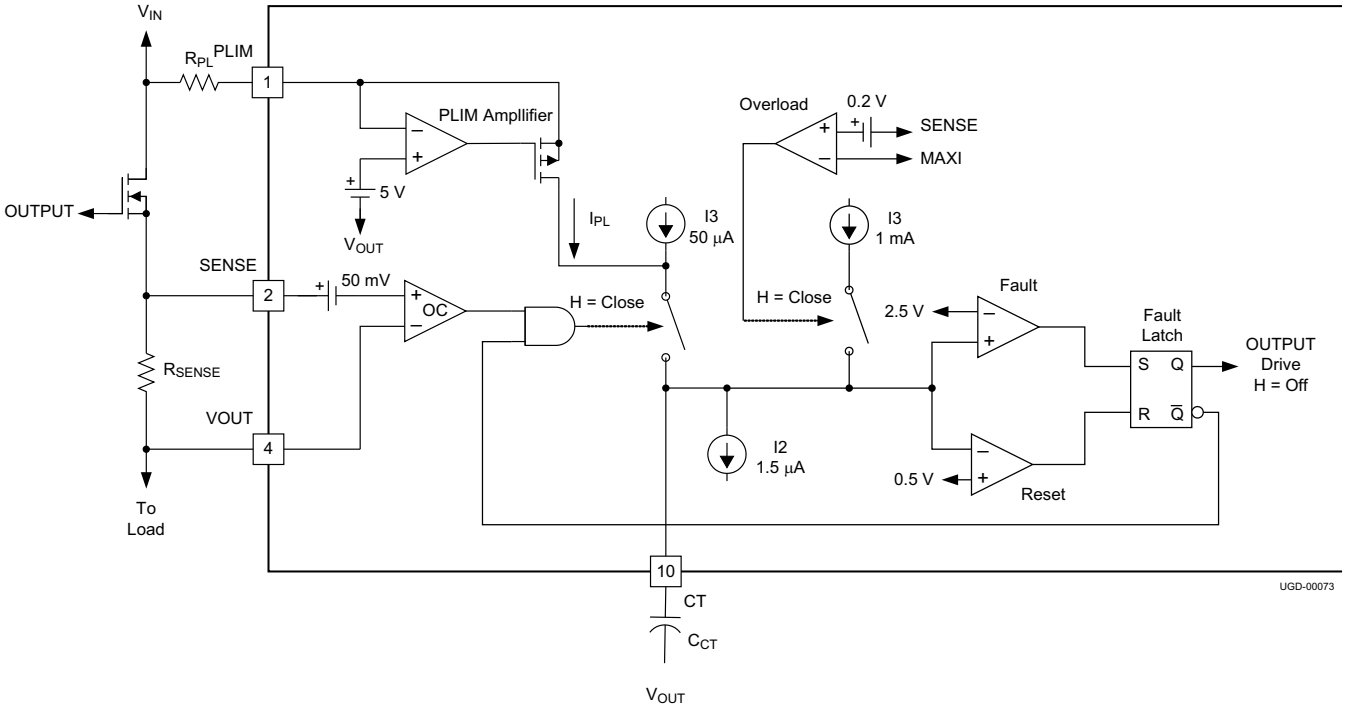


Figure 1. Fault Timing Circuitry for the UCC3917, Including Power Limit and Overload

NOTE

Under normal fault conditions where the output current is slightly above the fault level, $V_{VOUT} \cong V_{IN}$, $I_{PL} = 0$, and the C_{CT} charging current is I1.

During a fault, C_{CT} charges at a rate determined by the internal charging current and the external timing capacitor, CT. Once C_{CT} charges to 2.5 V, the fault comparator switches and sets the fault latch. Setting the fault latch causes both the output to switch off and the charging switch to open. CT must now discharge with current source I2 until 0.5 V is reached. Once the voltage at C_{CT} reaches 0.5 V, the fault latch resets (assuming $\overline{\text{LATCH}}$ is high, otherwise the fault latch does not reset until the $\overline{\text{LATCH}}$ pin is brought high or a power-on reset occurs). This re-enables the output and allows the fault circuitry to regain control of the charging switch. If a fault is still present, the overcurrent comparator closes the charging switch causing the cycle to repeat.

Under a constant fault the duty cycle is shown in Equation 2.

$$D = \frac{I_2}{I_{PL} - I_1} = \frac{1.5\,\mu A}{I_{PL} + 50\,\mu A} \quad (2)$$

where

- I_{PL} is 0 µA under normal operations (see Figure 2)

However, during large transients average power dissipations can be limited using the PLIM pin. The average dissipation in the pass element is shown in Equation 3.

$$P_{FET(avg)} = (V_{IN} - V_{VOUT}) \times I_{MAXI} \times D = (V_{IN} - V_{VOUT}) \times I_{MAXI} \times \frac{1.5\,\mu A}{I_{PL} + 50\,\mu A} \quad (3)$$

where

- both Equation 4 and Equation 5 are true

$$V_{IN} - V_{OUT} >> 5\,V \quad (4)$$

$$I_{PL} = \frac{(V_{IN} - V_{VOUT})}{R_{PL}} \quad (5)$$

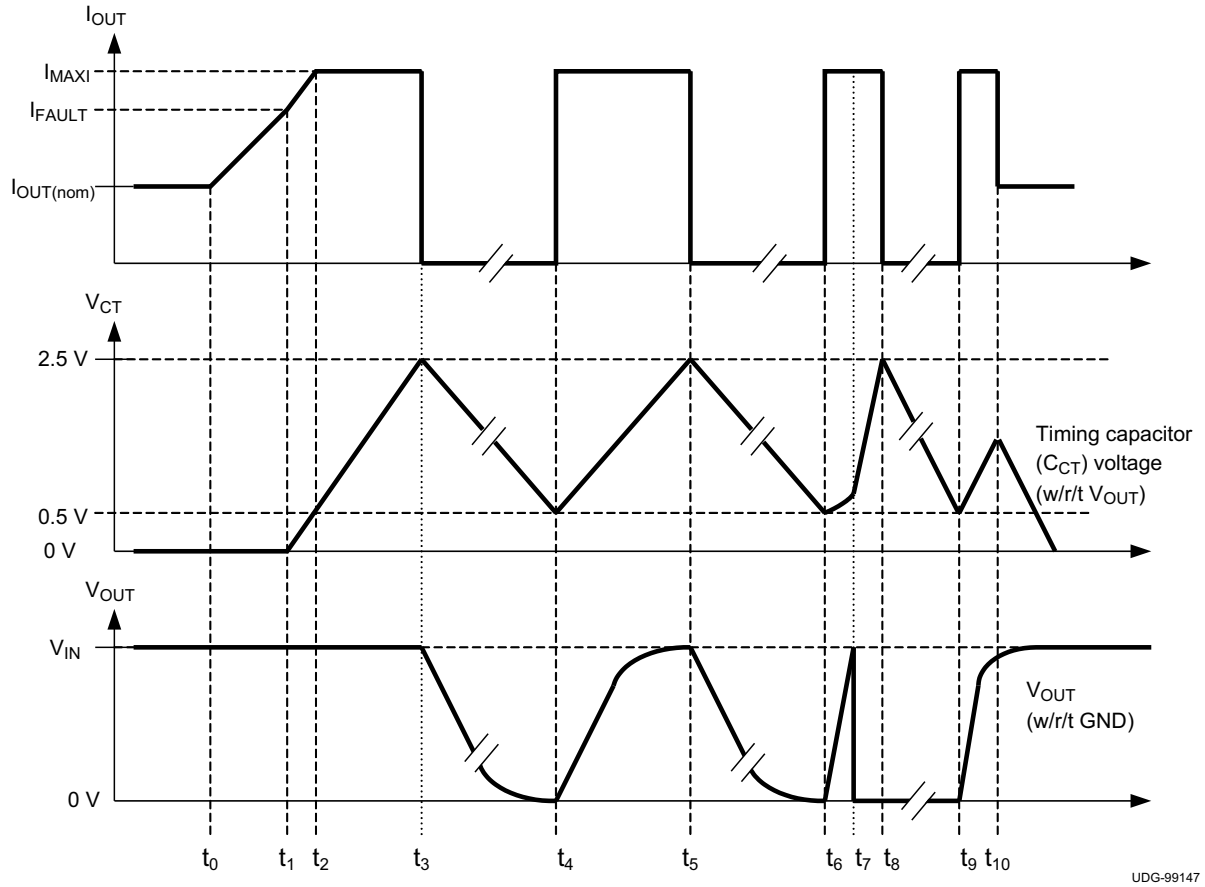


Figure 2. Typical Timing Diagram

Table 1. Timing Stages

TIME	CONDITION	DESCRIPTION
t0	**Safe condition**	Output current is nominal, output voltage is at the positive rail, V_{IN}
t1	**Fault control reached**	Output current rises above the programmed fault value, C_T begins to charge with approximately 50 µA
t2	**Maximum current reached**	**O**output current reaches the programmed maximum level and becomes a constant current with value I_{MAX}.
t3	**Fault occurs**	C_{CT} has charged to 2.5 V, fault output goes low, the FET turns off allowing no output current to flow, VOUT discharges to ground
t4	**Retry**	C_T has discharged to 0.5 V, but fault current is still exceeded, C_T begins charging again, FET is on, V_{VOUT} rises to V_{IN}.
t5	**t5 = t3**	This Illustrates 3% duty cycle.
t6	**t6 = t4**	
t7	**Output short circuit**	if VOUT is short circuited to ground, C_T charges at a higher rate depending upon the values for V_{IN} and R_{PL}.
t8	**Fault occurs**	**O**utput is still short circuited, but the occurrence of a fault turns the FET off so no current is conducted.
t9	**Fault remains**	Output short circuit released, still in fault mode.
t10	**t10 = t0**	Fault released, safe condition – return to normal operation of the circuit breaker.

Note that t6 – t5 ≅ 36 × (t5 – t4).

and where I_{PL} >> 50 µA, the duty cycle can be approximated in Equation 6.

$$D = \frac{1.5\,\mu A \times R_{PL}}{(V_{IN} - V_{VOUT})} \tag{6}$$

Therefore the average power dissipation in the MOSFET can be approximated by Equation 7.

$$P_{FET(avg)} = (V_{IN} - V_{VOUT}) \times I_{MAXI} \times \frac{1.5\,\mu A \times R_{PL}}{(V_{IN} - V_{VOUT})} = I_{MAXI} \times 1.5\,\mu A \times R_{PL} \tag{7}$$

Notice that since (V_{IN} – V_{VOUT}) cancels, average power dissipation is limited in the N-channel MOSFET pass element (see Figure 3). Also, a value for R_{PL} can be approximated by using Equation 8.

$$R_{PL} = \frac{P_{FET(avg)}}{I_{MAX} \times 1.5\,\mu A} \tag{8}$$

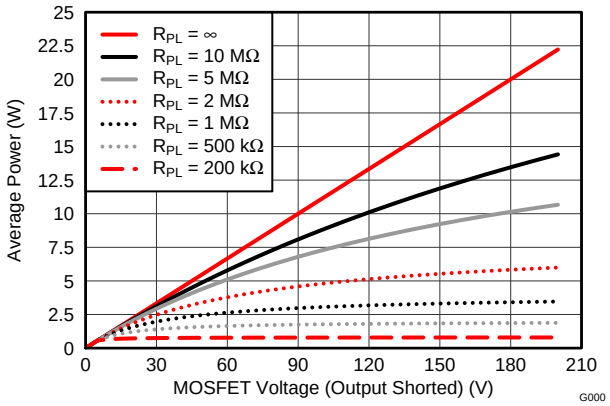


Figure 3.

OVERLOAD COMPARATOR

The overload comparator provides protection against a shorted load during normal operation when the external N-channel FET is fully enhanced. Once the FET is fully enhanced the linear current amplifier essentially saturates and the system is in effect operating open loop. Once the FET is fully enhanced the linear current amplifier requires a finite amount of time to respond to a shorted output possibly destroying the external FET. The overload comparator is provided to quickly shutdown the external MOSFET in the case of a shorted output (if the FET is fully enhanced). During an output short, C_T is charged by I3 at ≈ 1 mA. The current threshold for the overload comparator is a function of I_{MAX} and a fixed offset and is defined as:

$$I_{OVERLOAD} = I_{MAX} \times \frac{200\,mV}{R_S} \tag{9}$$

WHen the overcurrent comparator trips, the UCC3917 enters a programmed fault mode (hiccup or latched). It should be noted that on subsequent retries during hiccup mode or if a short should occur when the UCC3917 is actively limiting the current, the output current does not exceed I_{MAX}. In the event that the external FET does not respond during a fault the UCC3917 sets the VREF/CATFLT pin low to indicate a catastrophic failure.

SELECTING THE MINIMUM TIMING CAPACITANCE

To ensure that the device starts up correctly the designer must ensure that the fault time programmed by CT exceeds the startup time of the load. The startup time (t_{START}) is a function of several components; load resistance and load capacitance, soft-start components R1, R2 and C_{SS}, the power limit current contribution determined by R_{PL}, and C_{IN}.

Use Equation 10 to calculate the start time using a parallel capacitor-constant current load.

$$t_{START} = \frac{C_{LOAD} \times V_{IN}}{(I_{MAX} - I_{LOAD})} \tag{10}$$

Use Equation 11 to calculate calculate the start time using a parallel R-C load.

$$t_{START} = R_{LOAD} \times C_{LOAD} \times \ln \times \left(1 - \frac{V_{IN}}{I_{MAX} \times R_{LOAD}}\right) \tag{11}$$

If the power limit function is not be used, then $C_{CT(min)}$ can be found using Equation 12.

$$C_{T(min)} = \frac{I_{CH} \times t_{START}}{dV_{CT}}$$

where

- dV_{CT} is the hysteresis on the fault detection circuitry (12)

During operation in the latched fault mode configuration dV_{CT} = 2.5 V. When the UCC3917 is configured for the hiccup or retry mode of fault operation dV_{CT} = 2.0 V.

If the power limit function is used, the C_{CT} charging current becomes a function of I_{CH} + I_{PL}. $C_{CT(min)}$ is found by integrating Equation 13 with respect to V_{CT}.

$$C_{T(min)} = \left(I_{CH} + \frac{V_{IN} - (I_{MAX} \times R_{LOAD}) \times \left[1 - e^{\left(\frac{-t}{R_{LOAD} \times C_{LOAD}}\right)}\right]}{R_{PL}} \right) \times \frac{dt}{dV_{CT}} \tag{13}$$

The minimum timing capacitance is calculated in Equation 14.

$$C_{T(min)} = \frac{1}{R_{PL} \times dV_{CT}} \times \left[\left((I_{CH} \times R_{PL}) + V_{IN} - (I_{MAX} \times R_{LOAD}) \right) \times t_{START} + V_{IN} \times R_{LOAD} \times C_{LOAD} \right] \tag{14}$$

SELECTING OTHER EXTERNAL COMPONENTS

Other external components are necessary for correct operation of the device. Referring to Figure 13, resistors R_{SENSE}, R_{SS}, R_{DD}, R17, R18, and R19 and Equation 15 theough Equation 17 apply:

$$R_{SENSE} = \frac{50\,mV}{I_{FAULT}} \tag{15}$$

$$R_{SS} = \left(\frac{(V_{IN} - 5\,V)}{I_{DD}} \right) \tag{16}$$

$$R_{DD} = \left(\frac{(V_{IN} - 10\,V)}{I_{DD}} \right) \tag{17}$$

(R17 + R18 + R19) > 20 kΩ (current limit out of VREF)

Use a value of 0.1 µF for the external charge pump capacitors.

SOFT-START OPERATION

The soft-start circuits in Figure 4, and Figure 5 gradually ramp up the load current on power-up, retry, or if the SHTDWN pin is pulled high. Control circuitry (not shown) turns on Q1 to discharge C1 when FLTOUT or SHTDWN are low (i.e., external power MOSFET is off) so the load current always ramps from zero. The circuit in Figure 4 uses an inexpensive bipolar transistor for Q1 so the component cost is lower than the circuit in Figure 5.

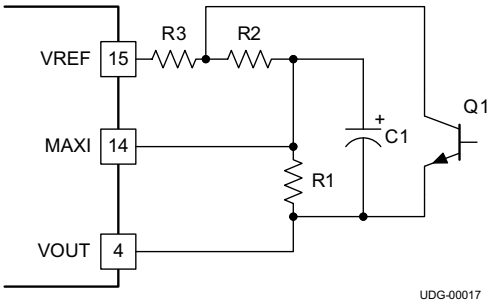


Figure 4. Soft-Start Circuit Using A Higher-Cost Bi-Polar Transistor

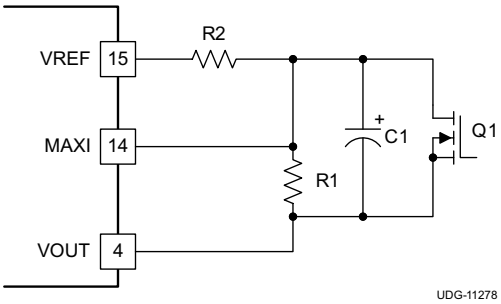


Figure 5. Soft-Start Circuit Using A Lower-Cost MOSFET

Soft-start operation minimizes the voltage disturbance on the power bus when a circuit card is inserted into a live back plane. This disturbance could reset a system, which is not desirable when high availability is required. A server is an example of a high availability system.

Soft-start operation is initiated with the SHTDWN pin in as shown in Figure 6. The anode of D2 is grounded when the card is in the back plane. R2 limits the SHTDWN pin current to between 60 µA and 500 µA (i.e., 60 µA < 0.65 V / R2 < 500 µA).

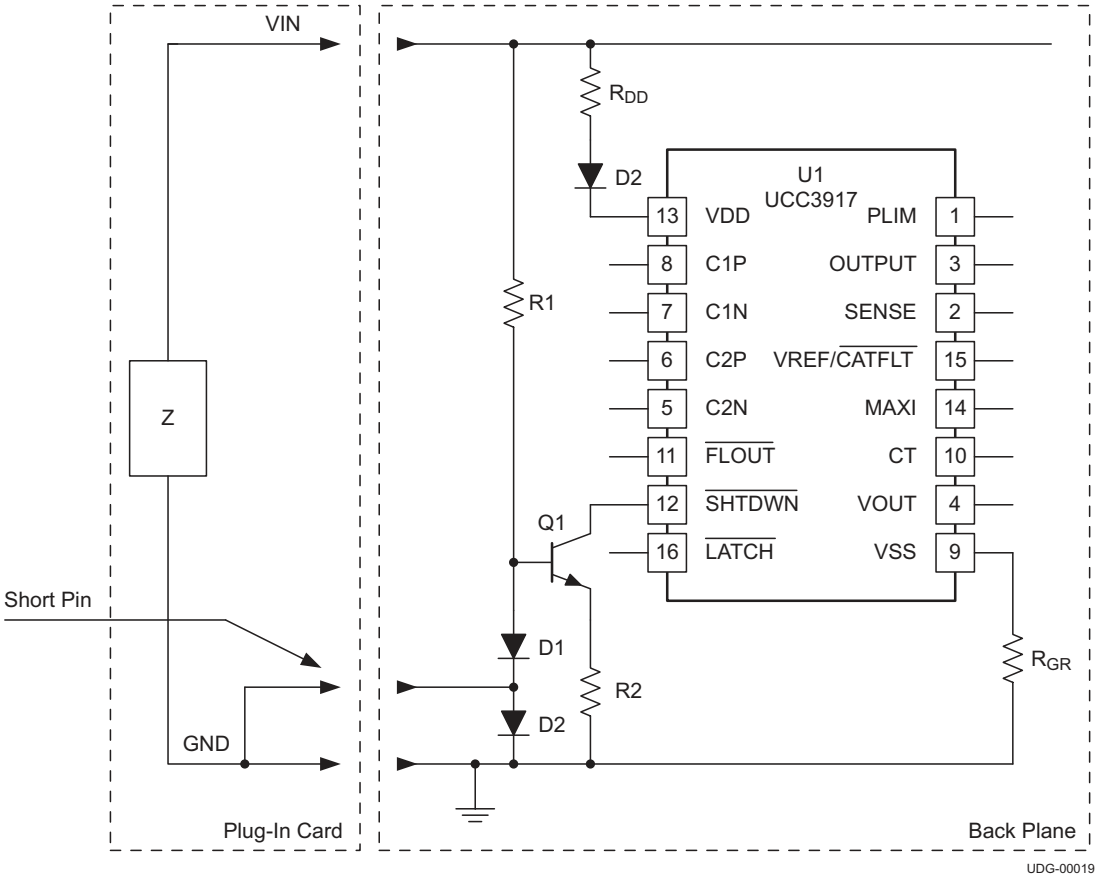


Figure 6. Soft-Start Operation with $\overline{\text{SHTDWN}}$

I/O INTERFACE

The $\overline{\text{SHTDWN}}$ and $\overline{\text{LATCH}}$ inputs and $\overline{\text{FLTOUT}}$ output are referenced to VOUT. Level-shifting circuits are needed if the device communicates with logic that is referenced to load/system ground.

INTERFACING TO $\overline{\text{LATCH}}$ AND $\overline{\text{SHTDWN}}$

Two level shift circuits for $\overline{\text{LATCH}}$ and $\overline{\text{SHTDWN}}$ are shown in Figure 7. The optocoupler (Figure 7) is simple, but the constant-current sink (Figure 8) is a low-cost solution.

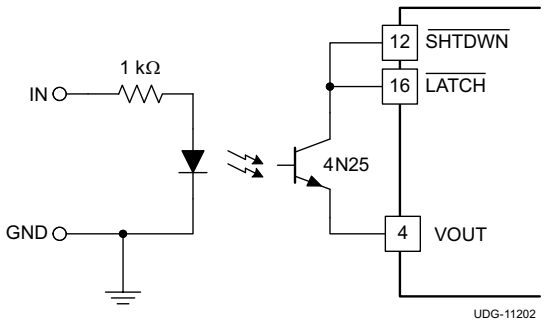


Figure 7. Optocoupler Interface

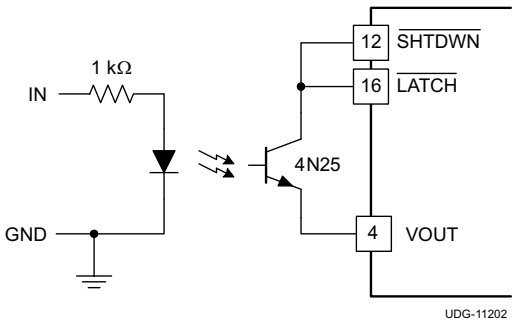


Figure 8. Constant-Current Sink Interface

Design Example 1: Using the TTL Signal to Control the $\overline{\text{LATCH}}$ Pin Input

A TTL signal controls the $\overline{\text{LATCH}}$ input of the UCC3917 using the circuit in Figure 8. Determine the component values if the maximum load voltage is 60 V.

The assumptions for this analysis are:

- $V_{BE} \approx 0.65$ V
- $V_{CE(sat)} \approx 0.1$ V
- R1||R2 << hfe × R3
- Voltage measurements are with respect to load ground

Calculation Steps

Step 1. Select Q1.

The LATCH input is internally pulled up to the charge pump voltage, which is 10 V above the load voltage. Q1 is therefore subjected to 70 V in a 60 V system. A FMMTA06 transistor, with a $V_{CEO(max)}$ of 80 V, is suitable for Q1 in this application.

Step 2. Determine R1, R2 and R3.

The interface circuit responds to a TTL input as follows.

- Logic "0" input: 0 V < V_{IL} < 0.8 V ⇒ 0 µA < I_C < 60 µA and V_C > 1.7 V
- Logic "1" input: 2 V < V_{IH} < 5 V ⇒ 60 mA < I_C < 500 µA and V_C < 1.7 V

This response establishes the relationship between R1, R2, and R3.

If $V_{IN} = V_{IL(max)} = 0.8$ V, then Q1 is off and $V_B - V_{IL(max)} \times \left(\frac{R_2}{R_1 + R_2}\right) < V_{BE} \longrightarrow \frac{R_1}{R_2} > 0.23$

If $V_{IN} = V_{IH(max)} = 5$ V, then:

$$V_B - V_{IL(max)} \times \left(\frac{R_2}{R_1 + R_2}\right) < V_{BE} \longrightarrow \frac{R_1}{R_2} > 0.23 \quad I_C = \left(\frac{1.7\,V - V_{CE(sat)}}{R_3}\right) < 500\,\mu A \longrightarrow R3 > 3.2\,k\Omega$$

$$V_C = V_{CE(sat)} + V_E < 1.7\,V \longrightarrow V_E < 1.6\,V \quad V_E = (V_B - V_{BE}) < 1.6\,V \longrightarrow V_B < 2.25\,V$$

$$\frac{(V_B - V_{IH(max)}) \times R_2}{R_1 + R_2} < 2.25\,V \longrightarrow \frac{R_1}{R_2} > 1.222$$

If $V_{IN} = V_{IH(max)} = 2$ V, then:

$$V_B = \frac{(V_{IH(min)} \times R_2)}{(R_1 + R_2)} \longrightarrow \frac{2\,V}{1 + \left(\frac{R_1}{R_2}\right)}$$

$$I_C = \frac{(V_B - V_{BE})}{R_3} > 60\,\mu A \longrightarrow R_3 < \frac{(V_B - V_{BE})}{60\,\mu A}$$

In summary, R1, R2, and R3 obey the inequalities:

$$\frac{R_1}{R_2} > 1.222 \quad \text{and} \quad 3.2\,k\Omega < R_3 < \frac{(V_B - 0.65)}{60\,\mu A}, \text{ where } V_B = \frac{2\,V}{1 + \left(\frac{R_1}{R_2}\right)}$$

If R1 / R2 = 1.3, then 3.2 kΩ < R3 < 3.66 kΩ. R1 = 4.64 kΩ for the case where R2 = R3 = 3 kΩ.

The same design can be used to control the UCC3917's $\overline{\text{SHTDWN}}$ input.

Interfacing to $\overline{\text{FLTOUT}}$

The level shift circuit in Figure 9 is a way to interface to $\overline{\text{FLTOUT}}$. The operation of this circuit and the $\overline{\text{SHTDWN}}$ / $\overline{\text{LATCH}}$ level shift circuit in Figure 8 are similar.

Design Example 2: A TTL-Compatible Output Level Shifter Using $\overline{\text{FLOUT}}$

This design example describes a TTL compatible output level shifter for $\overline{\text{FLTOUT}}$. The maximum system voltage is 60 V.

Use a level shift circuit as shown in Figure 9. The $\overline{\text{FLTOUT}}$ output can swing to the charge pump voltage, which is 10 V above the load voltage. In a 60-V application, the collector-emitter of Q1 can be as high as –70 V. A FMMT593 transistor, with a $V_{CEO(max)}$ rating of –100 V, is a suitable choice for Q1.

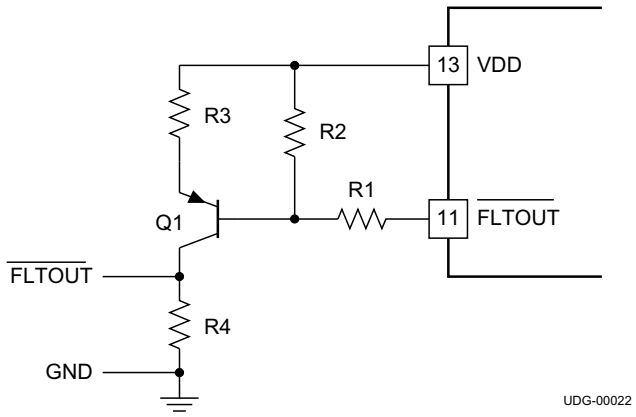


Figure 9. Interfacing to $\overline{\text{FLTOUT}}$

Calculation Steps

Step 1. Output saturation voltage constraint.

$$V_{C(on)} = V_E + V_{CE(sat)} > 2.4\,V\,(\text{TTL output high}) \tag{18}$$

$$\text{If } V_{C(on)} = 2.6\,V, \text{ then } V_E = \left(2.6\,V + \left(-0.1\,V\right)\right) = 2.5\,V \tag{19}$$

Step 2. Source current constraint.

I_C = 100 µA

Step 3. Calculate the value of R3.

$$R_3 = \frac{\left(6\,V - V_E\right)}{I_E} = \frac{\left(6\,V - V_E\right)}{I_C} = \frac{\left(6\,V - 2.5V\right)}{100\,\mu A} = 35\,k\Omega \tag{20}$$

Step 4. Calculate the base voltage.

$$V_B = V_E + V_{BE} = \left(2.5\,V - 0.65\,V\right) = 1.85\,V \tag{21}$$

Step 5. Calculate the voltage divider.

The voltage divider formula for R1 and R2 is shown in Equation 22

$$\frac{R_2}{\left(R_1 + R_2\right)} \times 6\,V \cong \left(6\,V - V_B\right) \text{ or } \frac{R_1}{R_2} = \frac{6\,V}{\left(V_B - 1\right)} \tag{22}$$

Equation 23 assumes negligible loading by Q1.

$$\frac{R_1}{R_2} \ll hfe \times R_3 \tag{23}$$

If hfe = 100, then:

$$\frac{R_1}{R_2} = \left(\frac{6}{(1.85-1)}\right) = 2.24 \text{ and } \frac{R_1}{R_2} << (100 \times 35k\Omega) = 3.5M \quad (24)$$

If R2 = R3 = 34.8 kΩ, then R1 = 15.4 kΩ

Step 6. Calculate the output voltage.

The output voltage is set by R4.

$$I_C \times R_4 > 2.4\,V, \quad R_4 > \frac{2.4\,V}{100\,\mu A} = 24\ k\Omega \quad (25)$$

Choose an R4 value of 49.9 kΩ.

PRELOADING THE OUTPUT

R_{DD} provides a sneak path for current between 3 mA and 11 mA (e.g., at 0 V output) to trickle into the load when the power FET is off (see Figure 10).

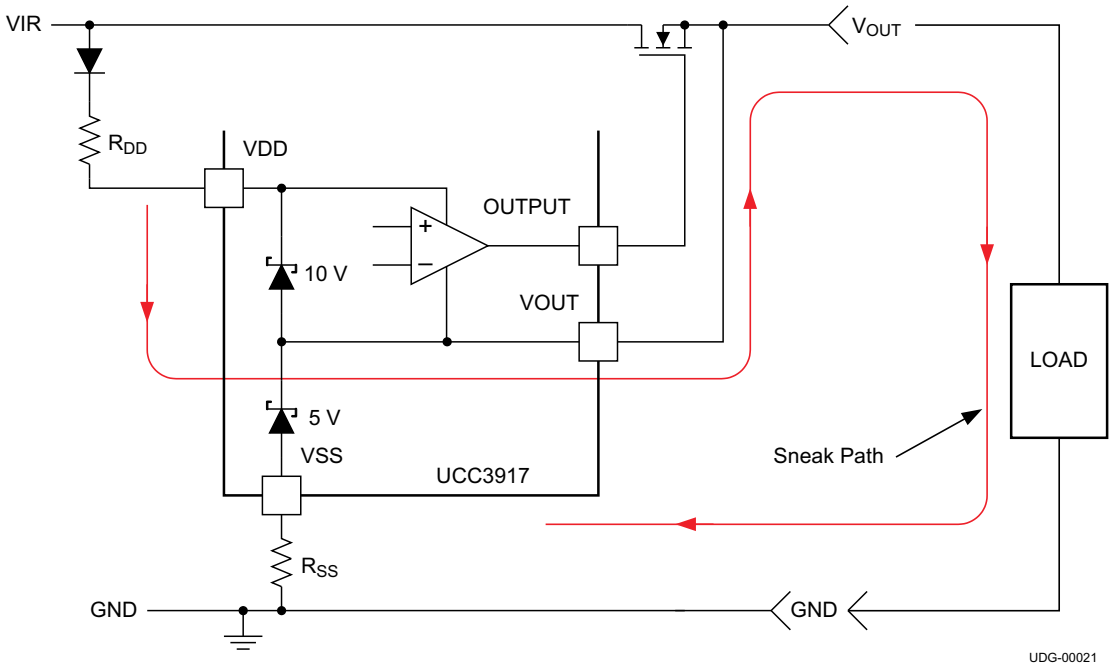


Figure 10. Simplified Schematic Illustrating I_{DD} Sneak Path

This current causes an unacceptably high output voltage at shutdown if the output is not adequately loaded. In this case, it is necessary to preload the HSPM output to keep the shutdown voltage level acceptable. The preload also insures reliable start-up of the UCC3917 by holding the output voltage low when power is first applied to the HSPM.

A resistor is usually an unacceptable preload because it creates a power dissipation problem when the FET turns on. For example, a 90.9-Ω preload (used to limit the shutdown voltage of a 48-V HSPM to less than 1 V) adds 25-W of power dissipation to the system. In a 100-V system, this dissipation increases to 110 W. The power dissipation overhead increases with the system voltage squared for a resistive preload.

Figure 11 shows how the active load limits the shutdown voltage without creating a power dissipation problem.

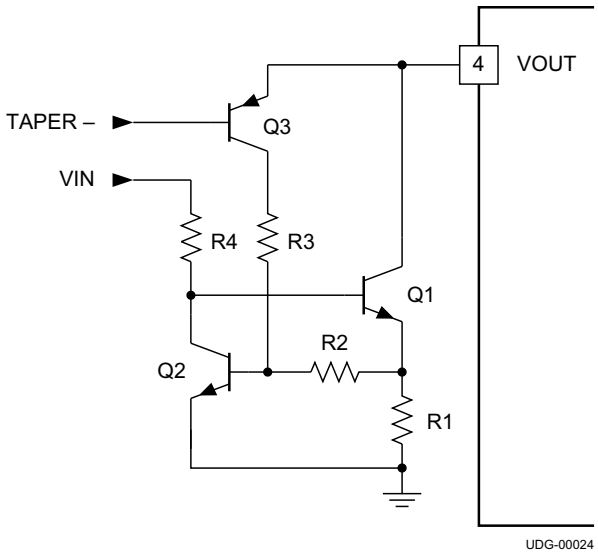


Figure 11. Active Preload

This load is a constant-current sink (i.e., Q3 is off) when the power FET is off. The shutdown voltage is less than 0.85 V if the sink current, set by R1, is greater than 11 mA.

$$I_{SNKFET(off)} = \frac{V_{BE}}{R_1} > 11A \tag{26}$$

The power dissipation of Q1 is kept to a minimum when the power FET turns on by tapering the sink current as the load voltage rises as shown in Equation 27 .

$$I_{SNKFET(on)} = \left(\left(\frac{V_{BE}}{R_1}\right) - V_{OUT}\right) \times \left(\frac{R_2}{(R_1 \times R_3)}\right) \tag{27}$$

For R1 << R2 << R3

Control circuitry turns on Q3 to activate current tapering. Tapering the current causes the power dissipation of Q1 to peak when the load voltage is calculated in Equation 28.

$$V_{OUT} = \frac{V_{BE}}{2} \times \frac{R_3}{R_2} \tag{28}$$

The power dissipated by Q1 at this voltage is shown in Equation 29.

$$P_{D(max)Q1} = \left(\frac{V_{BE}}{2}\right)^2 \times \frac{R_3}{(R_1 \times R_2)} \tag{29}$$

In the case of a brownout or if the input voltage rises slowly (e.g., adjustable lab power supply), it is possible for Q1 to remain in the maximum power dissipation region for a significant time. Limiting the power dissipation of Q1 below its maximum rating insures reliable operation in this case.

Design Example 3: A 14-mA Active Preload for a 60-V Hot Swap Power Manager (HSPM)

Calculation Steps

Step 1. Set the sink current.

$$R_1 = \frac{V_{BE}}{I_{SNKFET(off)}} = \frac{0.65\,V}{14\,mA} = 46.4\,\Omega \tag{30}$$

Use a BC846B transistor for Q1. This device has a collector breakdown voltage of 65 V and power dissipation rating of 225 mW.

Step 2. Select R2 and R3.

Select R2 and R3 to limit the power dissipation of Q1 to less than 225 mW, in this example 150 mW is chosen.

$$\frac{R_3}{R_2} = \left(\frac{2}{V_{BE}}\right)^2 \times R_1 \times P_{D(max)Q1} = \left(\frac{2}{0.65\,V}\right)^2 \times 46.4\,\Omega \times 0.15\,W = 65.9 \tag{31}$$

If R2 = 3.01 kΩ, then R3 = 198 kΩ.

The power dissipation of Q1 is shown in Figure 12.

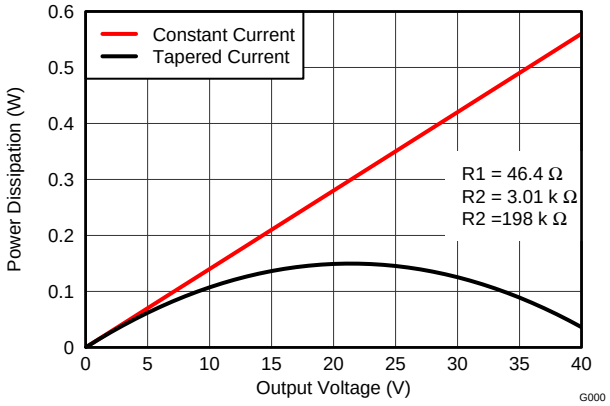


Figure 12. Output Voltage vs. Power Dissipation

PROTECTING THE 5-V REGULAOR

The UCC3917's 5-V regulator can overvoltage if VOUT is loaded with less than 11 mA (min) on power up. The overvoltage mechanism is best understood by recognizing that the 5-V Zener diode in the UCC3917 block diagram, is actually a feedback shunt regulator. This regulator turns on when the voltage across the UCC3917's 10-V Zener diode is greater than the UVLO threshold. If VOUT is unloaded and power is applied to the UCC3917, the UVLO threshold cannot be reached and the 5-V regulator impedance is infinite.

Consequently, the entire input voltage appears across the shunt regulator causing it to break down. Clamping its voltage with Zener diode to 5.6 V can protect the regulator.

NOTE

The Zener diode is unnecessary if the current drawn from VOUT is greater than 11 mA when power is *initially* applied to the UCC3917.

EVALUATION CIRCUIT EXAMPLE

A 28 V to 60 V at 1-A HSPM evaluation circuit is shown in Figure 13. Level translation circuitry allows communications with logic referenced to load ground. This circuit is available as a DV3917 Evaluation Board. Contact your local Texas Instruments sales representative for more information.

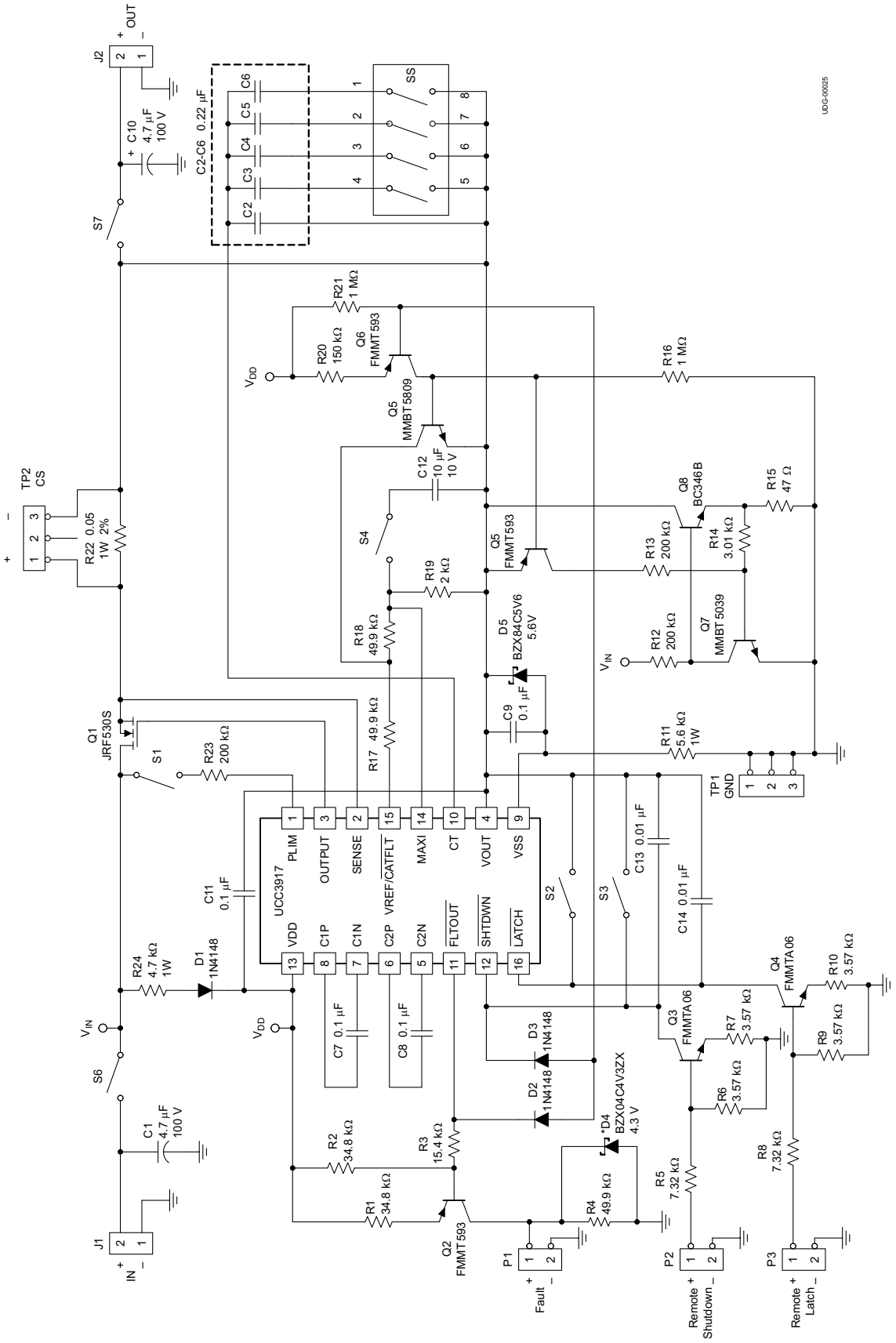


Figure 13. A 28 V to 60 V at 1-A Positive Floating HSPM Evaluation Circuit Using the UCC3917

SAFETY RECOMMENDATIONS

Although the UCC3917 is designed to provide system protection for all fault conditions, all integrated circuits can ultimately fail short. For this reason, if the UCC3917 is intended for use in safety critical applications where UL or some other safety rating is required, a redundant safety device such as a fuse should be placed in series with the power device. The UCC3917 prevents the fuse from blowing for virtually all fault conditions, increasing system reliability and reducing maintenance cost, in addition to providing the hot-swap benefits of the device.

REVISION HISTORY

Changes from Revision C (February, 2013) to Revision D **Page**

- Changed the UCC2917 and UCC3917 marketing status from Product Preview to Production Data. 1