



# -48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection

MAX5938

## General Description

The MAX5938 is a hot-swap controller for -10V to -80V rails. The MAX5938 allows circuit line cards to be safely hot-plugged into a live backplane without causing a glitch on the power supply. It integrates an adjustable circuit-breaker function requiring no RSENSE.

The MAX5938 provides a controlled turn-on for circuit cards, which limits inrush current and prevents both glitches on the power-supply rail and damage to board connectors and components. Before startup, the MAX5938 performs a Load Probe™ test to detect the presence of a short-circuit condition. If a short-circuit condition does not exist, the device limits the inrush current drawn by the load by gradually turning on the external MOSFET. Once the external MOSFET is fully enhanced, the MAX5938 provides overcurrent and short-circuit protection by monitoring the voltage drop across the R<sub>DS(ON)</sub> of the external power MOSFET. The MAX5938 integrates a 400mA fast GATE pulldown to guarantee that the external MOSFET is rapidly turned off in the event of an overcurrent or short-circuit condition.

The MAX5938 also protects the system against input voltage (V<sub>IN</sub>) steps. During an input voltage step, the device limits the current drawn by the load to a safe level without shutting down the load. The device also includes ON/OFF control, selectable PGOOD output polarity, undervoltage (UV) and overvoltage (OV) protection.

The device offers latched (MAX5938L) or autoretry (MAX5938A) fault management. Both the MAX5938A and MAX5938L are available in a 16-pin QSOP package and are specified for the extended (-40°C to +85°C) temperature range.

## Applications

Servers  
Telecom Line Cards  
Network Switches  
Solid-State Circuit Breakers  
Network Routers

*Typical Operating Circuit appears at end of data sheet.*

*Load Probe is a trademark of Maxim Integrated Products, Inc.*

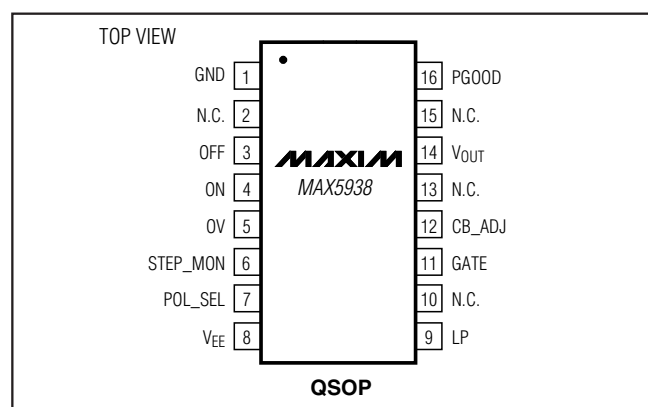
## Features

- ◆ -10V to -80V Operation
- ◆ No External RSENSE Required
- ◆ Drives Large Power MOSFETS
- ◆ Eliminates Inrush Current Spikes During Hot Plug into Powered Backplane
- ◆ Eliminates Inrush Current Spikes and Dropping of Load During Large V<sub>IN</sub> Steps
- ◆ Adjustable Circuit-Breaker Threshold with Temperature Compensation
- ◆ Circuit-Breaker Fault with Transient Rejection
- ◆ Shorted Load Detection (Load Probe) Before Power MOSFET Turn-On
- ◆ Programmable Load-Voltage Slew Rate Controls Inrush Current
- ◆ ±2.4% Accuracy, Programmable Turn-On/Off Voltage (UVLO)
- ◆ Overvoltage Fault Protection with Transient Rejection
- ◆ Autoretry and Latched Fault Management Available
- ◆ Low Quiescent Current (1mA)

## Ordering Information

PART	TEMP RANGE	PIN-PACKAGE
MAX5938AE4E	-40°C to +85°C	16 QSOP
MAX5938LE4E	-40°C to +85°C	16 QSOP

## Pin Configuration



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## ABSOLUTE MAXIMUM RATINGS

V <sub>EE</sub> , V <sub>OUT</sub> , PGOOD, LP, STEP_MON to GND	+0.3V to -85V	GATE (during 2V clamp, continuous)	50mA
PGOOD to V <sub>OUT</sub>	-0.3V to +85V	GATE (during gate pulldown, continuous)	50mA
V <sub>OUT</sub> , LP, STEP_MON to V <sub>EE</sub>	-0.3V to +85V	Continuous Power Dissipation (T <sub>A</sub> = +70°C)	
GATE to V <sub>EE</sub>	-0.3V to +20V	16-Pin QSOP (derate 8.3mW/°C above +70°C)	667mW
ON, OFF, OV, POL_SEL, CB_ADJ to V <sub>EE</sub>	-0.3V to +6V	Operating Temperature Range	-40°C to +85°C
Input Current		Junction Temperature	+150°C
LP (internally duty-cycle limited)	1A	Storage Temperature Range	-65°C to +150°C
PGOOD (continuous)	80mA	Lead Temperature (soldering, 10s)	+300°C
GATE (during 15V clamp, continuous)	30mA		

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 in the operational sections of the specifications is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

## ELECTRICAL CHARACTERISTICS

(V<sub>EE</sub> = -10V to -80V, V<sub>IN</sub> = (GND - V<sub>EE</sub>), V<sub>STEP\_MON</sub> = V<sub>EE</sub>, R<sub>LP</sub> = 200Ω, V<sub>ON</sub> = V<sub>OFF</sub> = 2V, V<sub>OV</sub> = V<sub>CB\_ADJ</sub> = V<sub>EE</sub>, POL\_SEL open, T<sub>A</sub> = -40°C to +85°C, unless otherwise noted. Typical values are at V<sub>EE</sub> = -48V, T<sub>A</sub> = +25°C.) (Notes 1, 2)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS	
Operating Voltage Range	V <sub>EE</sub>	Referenced to GND	-80		-10	V	
Operating Supply Current	I <sub>CC</sub>			0.95	1.4	mA	
<b>ON/OFF, OV</b>							
ON Reference Threshold Rising	V <sub>ON_REF,R</sub>	V <sub>ON</sub> increasing	1.219	1.25	1.281	V	
ON Reference Threshold Falling	V <sub>ON_REF,F</sub>	V <sub>ON</sub> decreasing	1.069	1.125	1.181	V	
ON Glitch Rejection (Note 3)	t <sub>REJ</sub>	V <sub>ON</sub> decreasing	0.80	1.5	2.25	ms	
OFF Reference Threshold	V <sub>OFF_REF</sub>		1.219	1.25	1.281	V	
ON/OFF/OV Input Bias Current	I <sub>BIAS</sub>		-25		+25	nA	
OV Reference Threshold, Rising	V <sub>OV_REF,R</sub>	V <sub>OV</sub> increasing	1.219	1.25	1.281	V	
OV Reference Threshold, Falling	V <sub>OV_REF,F</sub>	V <sub>OV</sub> decreasing	1.069	1.125	1.181	V	
OV Transient Rejection	t <sub>OVREJ</sub>	OV increasing	0.80	1.5	2.25	ms	
Power-Up Delay (Note 4)	t <sub>ONDLY</sub>		80	220	380	ms	
V <sub>OUT</sub> to V <sub>EE</sub> Leakage Current		V <sub>OFF</sub> = V <sub>EE</sub> = -80V, V <sub>OUT</sub> = GND, PGOOD open		0.01	1	μA	
LP to V <sub>EE</sub> Leakage Current		V <sub>OFF</sub> = V <sub>EE</sub> = -80V, LP = GND		0.01	1	μA	
POL_SEL to V <sub>EE</sub> Input Current		POL_SEL = V <sub>EE</sub>	-50	-34	-20	μA	
<b>GATE DRIVE</b>							
External Gate-Drive Voltage	V <sub>GS</sub>	V <sub>GATE</sub> - V <sub>EE</sub>	V <sub>IN</sub> = 10V	6.5	6.8	7.2	V
			14V < V <sub>IN</sub> < 80V	8.1	10	12.8	
GATE to V <sub>EE</sub> Clamp Voltage		MOSFET fully enhanced	I <sub>CLAMP</sub> = 9mA	13.5	16		V
			I <sub>CLAMP</sub> = 20mA		17	19.5	
		Power-off, V <sub>EE</sub> = GND	I <sub>CLAMP</sub> = 1mA		2.1	2.55	
			I <sub>CLAMP</sub> = 10mA		2.5	2.93	
Open-Loop Gate Charge Current	I <sub>G,ON</sub>	V <sub>GATE</sub> = V <sub>EE</sub> , V <sub>OUT</sub> = GND	-66	-52	-35	μA	
GATE Pulldown Switch On-Resistance	R <sub>G,OFF</sub>	V <sub>GATE</sub> - V <sub>EE</sub> = 500mV	V <sub>IN</sub> > 10V		9.0	14.1	mA
			V <sub>IN</sub> > 14V		7.5	12.5	
Output-Voltage Slew Rate	SR	dV <sub>OUT</sub> /dt , C <sub>SLEW</sub> = 0	2.4	9.0	14.8	V/ms	

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## ELECTRICAL CHARACTERISTICS (continued)

( $V_{EE} = -10V$  to  $-80V$ ,  $V_{IN} = (GND - V_{EE})$ ,  $V_{STEP\_MON} = V_{EE}$ ,  $R_{LP} = 200\Omega$ ,  $V_{ON} = V_{OFF} = 2V$ ,  $V_{OV} = V_{CB\_ADJ} = V_{EE}$ ,  $POL\_SEL$  open,  $T_A = -40^\circ C$  to  $+85^\circ C$ , unless otherwise noted. Typical values are at  $V_{EE} = -48V$ ,  $T_A = +25^\circ C$ .) (Notes 1, 2)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS	
<b>CIRCUIT BREAKER AND SHORT CIRCUIT</b>							
CB_ADJ Bias Current	I <sub>CB_ADJ</sub>	CB_ADJ = V <sub>EE</sub>	T <sub>A</sub> = +85°C	55	72	89	μA
			T <sub>A</sub> = +25°C	39	50	61	
			T <sub>A</sub> = -40°C		33		
Circuit-Breaker Threshold	V <sub>CB</sub>	CB_ADJ = V <sub>EE</sub>	T <sub>A</sub> = +85°C	59	72	85	mV
			T <sub>A</sub> = +25°C	41	50	59	
			T <sub>A</sub> = -40°C		33		
		R <sub>CB_ADJ</sub> = 2kΩ	T <sub>A</sub> = +85°C	123	144	165	
			T <sub>A</sub> = +25°C	85	100	115	
			T <sub>A</sub> = -10°C	66	82	98	
		T <sub>A</sub> = -40°C		66			
I <sub>CB_ADJ</sub> Temperature Coefficient		-40°C < T <sub>A</sub> < +85°C		6000		ppm/°C	
Circuit-Breaker Glitch Rejection	t <sub>CB_DLY</sub>		1.0	1.2	1.6	ms	
Short-Circuit Threshold (Note 5)	V <sub>SC</sub>	CB_ADJ = V <sub>EE</sub>	T <sub>A</sub> = +85°C	112	144	176	mV
			T <sub>A</sub> = +25°C	75	100	125	
			T <sub>A</sub> = -10°C	50	82	114	
			T <sub>A</sub> = -40°C		66		
		R <sub>CB_ADJ</sub> = 2kΩ	T <sub>A</sub> = +85°C	224	288	352	
			T <sub>A</sub> = +25°C	159	200	241	
			T <sub>A</sub> = -10°C	108	164	220	
			T <sub>A</sub> = -40°C		132		
Short-Circuit Response Time		150mV overdrive, C <sub>LOAD</sub> = 0, to GATE below 1V		330	500	ns	
<b>INPUT-VOLTAGE STEP PROTECTION</b>							
Input-Voltage-Step Detection Threshold	STEP <sub>TH</sub>		1.219	1.25	1.281	V	
Input-Voltage-Step Threshold Offset Current	I <sub>STEP_OS</sub>		-10.8	-10	-9.2	μA	
<b>LOAD-PROBE CIRCUIT</b>							
Load-Probe Switch On-Resistance		V <sub>LP</sub> - V <sub>EE</sub> = 1V		7.5	11	Ω	
Load-Probe Timeout	t <sub>LP</sub>		80	220	380	ms	
Load-Probe Retry Time	t <sub>LP_OFF</sub>			16 x t <sub>LP</sub>		s	
Shorted Load Detection Voltage Threshold	V <sub>TH_LP</sub>	Referenced to GND	-220	-200	-180	mV	

# -48V Hot-Swap Controller with $V_{IN}$ Step Immunity, No $R_{SENSE}$ , and Overvoltage Protection

## ELECTRICAL CHARACTERISTICS (continued)

( $V_{EE} = -10V$  to  $-80V$ ,  $V_{IN} = (GND - V_{EE})$ ,  $V_{STEP\_MON} = V_{EE}$ ,  $R_{LP} = 200\Omega$ ,  $V_{ON} = V_{OFF} = 2V$ ,  $V_{OV} = V_{CB\_ADJ} = V_{EE}$ ,  $POL\_SEL$  open,  $T_A = -40^\circ C$  to  $+85^\circ C$ , unless otherwise noted. Typical values are at  $V_{EE} = -48V$ ,  $T_A = +25^\circ C$ .) (Notes 1, 2)

PARAMETER	SYMBOL	CONDITIONS	MIN	TYP	MAX	UNITS
<b>LOGIC AND FAULT MANAGEMENT</b>						
Autoretry Delay	$t_{RETRY}$			$16 \times t_{LP}$		s
PGOOD Assertion Threshold		$ V_{OUT} - V_{EE} $ falling		$0.74 \times V_{CB}$		mV
		Hysteresis		$0.26 \times V_{CB}$		
PGOOD Assertion Delay Time (Note 6)	$t_{PGOOD}$		0.70	1.26	1.85	ms
PGOOD Low Voltage	$V_{OL}$	$I_{SINK} = 1mA$ , referenced to $V_{OUT}$ , $V_{OUT} < (GND - 5V)$		0.05	0.4	V
PGOOD Open-Drain Leakage	$I_L$	$V_{EE} = -80V$ , $PGOOD = GND$		0.01	1	$\mu A$

**Note 1:** All currents into pins are positive and all currents out of pins are negative. All voltages referenced to  $V_{EE}$ , unless otherwise specified.

**Note 2:** All limits are 100% tested at  $+25^\circ C$  and  $+85^\circ C$ . Limits at  $-40^\circ C$  and  $-10^\circ C$  are guaranteed by characterization.

**Note 3:**  $V_{ON}$  drops below the  $V_{ON\_REF\_F}$  threshold are ignored during this time.

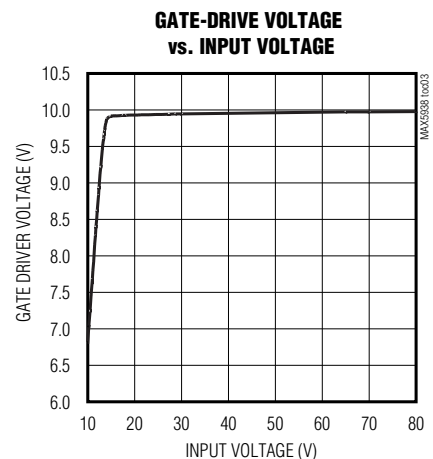
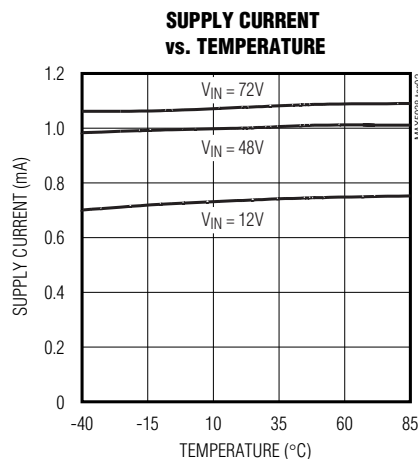
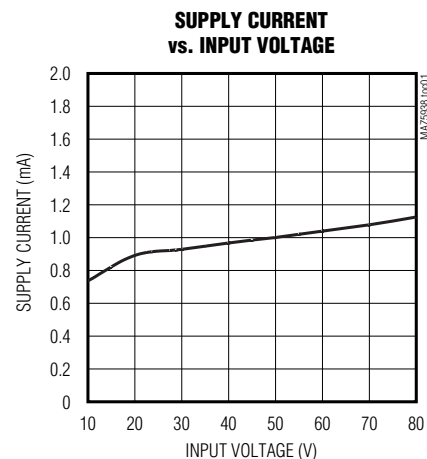
**Note 4:** Delay time from a valid on condition until the load-probe test begins.

**Note 5:** The short-circuit threshold is  $V_{SC} = 2 \times V_{CB}$ .

**Note 6:** The time when PGOOD condition is met until PGOOD signal is asserted.

## Typical Operating Characteristics

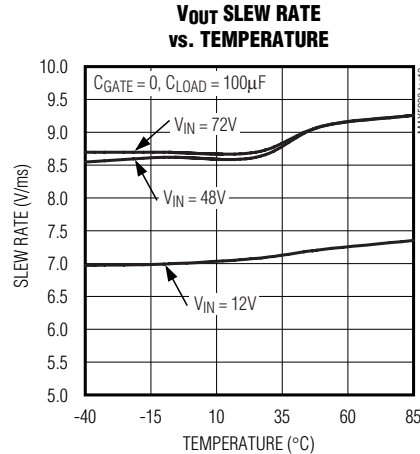
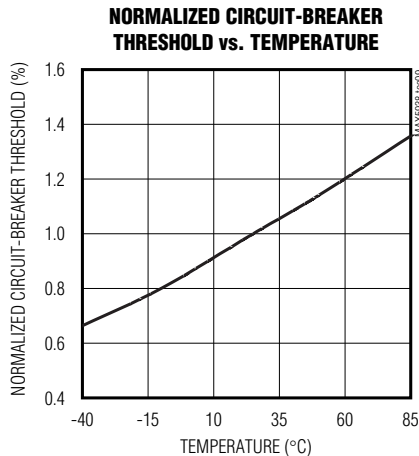
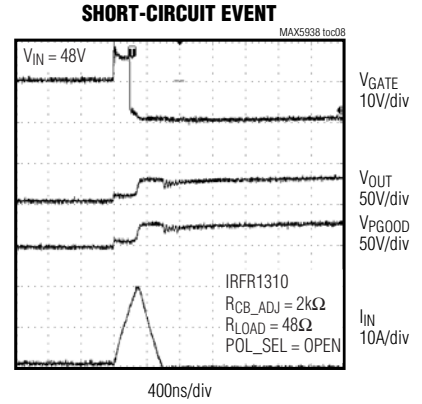
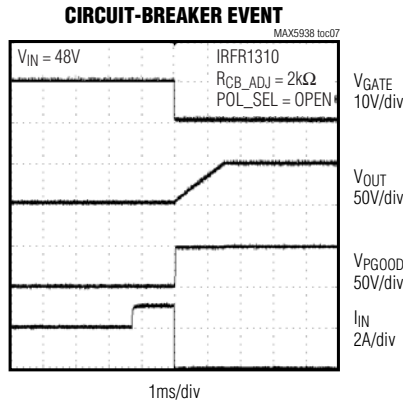
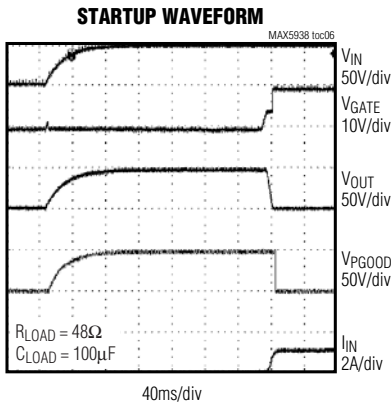
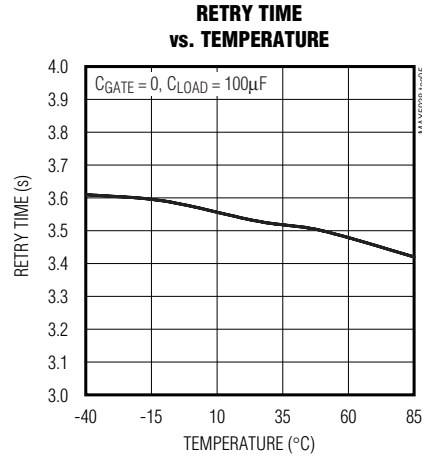
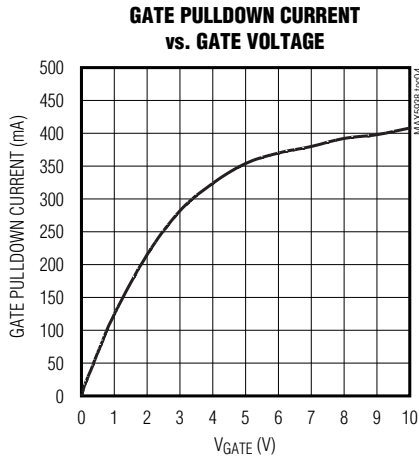
( $V_{EE} = -48V$ ,  $GND = 0V$ ,  $V_{IN} = GND - V_{EE}$ ,  $POL\_SEL =$  floating, all voltages are referenced to  $V_{EE}$ , unless otherwise noted.  $T_A = +25^\circ C$ , unless otherwise noted.)



# -48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection

## Typical Operating Characteristics (continued)

( $V_{EE} = -48V$ ,  $GND = 0V$ ,  $V_{IN} = GND - V_{EE}$ ,  $POL\_SEL = \text{floating}$ , all voltages are referenced to  $V_{EE}$ , unless otherwise noted.  $T_A = +25^\circ C$ , unless otherwise noted.)

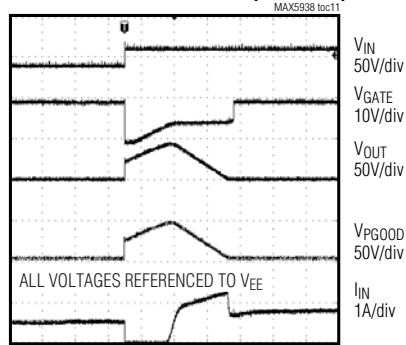


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## Typical Operating Characteristics (continued)

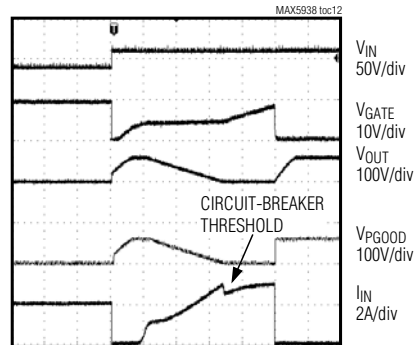
( $V_{EE} = -48V$ ,  $GND = 0V$ ,  $V_{IN} = GND - V_{EE}$ ,  $POL\_SEL =$  floating, all voltages are referenced to  $V_{EE}$ , unless otherwise noted.  $T_A = +25^\circ C$ , unless otherwise noted.)

**INPUT VOLTAGE STEP EVENT (NO FAULT)**



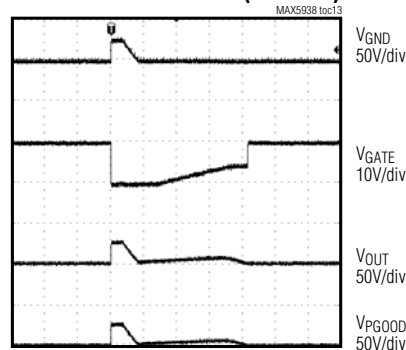
IRFR1310  
 $R_{CB\_ADJ} = 2k\Omega$   
 $R_{LOAD} = 80\Omega$

**INPUT VOLTAGE STEP TO FAULT MANAGEMENT**

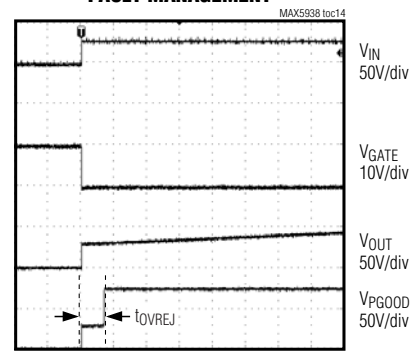


IRFR1310  
 $R_{CB\_ADJ} = 2k\Omega$   
 $R_{LOAD} = 20\Omega$

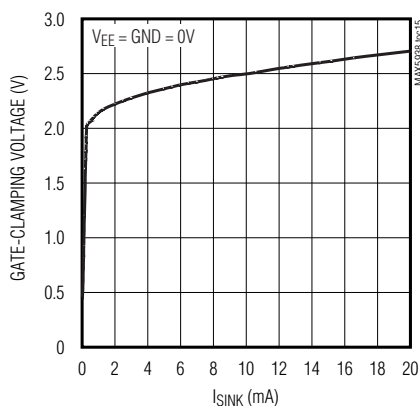
**OVERVOLTAGE TRANSIENT (NO FAULT)**



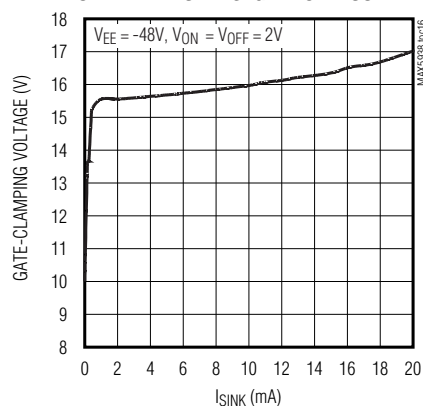
**OVERVOLTAGE TRANSIENT TO FAULT MANAGEMENT**



**GATE TO  $V_{EE}$  CLAMP VOLTAGE AT POWER-OFF vs. GATE SINK CURRENT**



**GATE TO  $V_{EE}$  CLAMP VOLTAGE MOSFET FULLY ENHANCED vs. GATE SINK CURRENT**



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MAX5938

## Pin Description

PIN	NAME	FUNCTION
1	GND	Ground. The high supply connection for a negative rail hot-swap controller.
2, 10, 13, 15	N.C.	No Connection. Not internally connected. Leave open.
3	OFF	Off Control Input. Referenced to V <sub>EE</sub> . Drive OFF and ON above 1.25V (typ) to turn on the MAX5938. When the ON input low requirements are met and OFF falls below 1.25V (typ), the MAX5938 turns off.
4	ON	On Control Input. Referenced to V <sub>EE</sub> . Drive ON and OFF above the 1.25V rising thresholds to turn on the MAX5938. When the voltage at OFF falls below its 1.25V (typ) threshold and the voltage at ON falls below 1.125V for longer than the ON 1.5ms glitch rejection period, the MAX5938 turns off.
5	OV	Overvoltage Control Input. Referenced to V <sub>EE</sub> . When the voltage at OV rises above the 1.25V rising threshold, GATE pulls to V <sub>EE</sub> until OV falls below the 1.125V falling threshold. If the overvoltage condition remains longer than 1.5ms, fault management initiates and PGOOD deasserts (see the <i>Detailed Description</i> ).
6	STEP_MON	Input Voltage Step Monitor. Connect a resistor between STEP_MON and V <sub>EE</sub> to set the step sensitivity. Connect a capacitor from GND to STEP_MON to adjust the step response relative to a negative step at V <sub>EE</sub> to eliminate false circuit-breaker and short-circuit faults. Connect to V <sub>EE</sub> to disable the step immunity function. See the <i>Selecting Resistor and Capacitor Values for Step Monitor</i> section in the <i>Applications Information</i> .
7	POL_SEL	PGOOD Output Polarity Select. Leave POL_SEL open for an active-low PGOOD assertion. Connect POL_SEL to V <sub>EE</sub> for an active-high open-drain PGOOD assertion.
8	V <sub>EE</sub>	Negative Input Voltage
9	LP	Load-Probe Detect. Connect a resistor from LP to V <sub>OUT</sub> to set the load-probe test current. <b>Limit load-probe test current to 1A.</b> Connect to V <sub>EE</sub> to disable load-probe function.
11	GATE	Gate-Drive Output. Connect to the gate of the external n-channel MOSFET.
12	CB_ADJ	Circuit-Breaker Adjust. Connect a resistor from CB_ADJ to V <sub>EE</sub> to adjust the circuit-breaker threshold. Short CB_ADJ to V <sub>EE</sub> for the default circuit-breaker 50mV (typ) threshold. Leave CB_ADJ open to disable circuit-breaker and short-circuit fault detection.
14	V <sub>OUT</sub>	Output Voltage Sense. V <sub>OUT</sub> is the negative rail of the load. Connect to the drain of the external n-channel MOSFET.
16	PGOOD	Power-Good Open-Drain Output. Referenced to V <sub>OUT</sub> . PGOOD asserts high (POL_SEL = V <sub>EE</sub> ) or low (POL_SEL open) when V <sub>OUT</sub> is within limits and there is no fault condition. PGOOD is deasserted when ON and OFF are cycled low.

## Detailed Description

The MAX5938 hot-swap controller incorporates over-current and overvoltage fault management and is intended for negative-supply-rail applications. The MAX5938 eliminates the need for an external RSENSE and includes V<sub>IN</sub> input step protection and load probe, which prevents powering up into a shorted load. It is intended for negative 48V telecom power systems where low cost, flexibility, multifault management, and compact size are required. The MAX5938 is ideal for the widest range of systems from those requiring low current with small MOSFETs to high-current systems requiring large power MOSFETs and low on-resistance.

The MAX5938 controls an external n-channel power MOSFET placed in the negative supply path of an external load. When no power is applied, the GATE output of the MAX5938 clamps the V<sub>GS</sub> of the MOSFET to 2V keeping the MOSFET turned off (Figure 2). When power is applied to the MAX5938, the 2V clamp at the GATE output is replaced by a strong pulldown device, which pulls GATE to V<sub>EE</sub> and the V<sub>GS</sub> of the MOSFET to 0. As shown in Figure 2, this transition enables the MAX5938 to keep the power MOSFET continually off during the board insertion phase when the circuit board first makes contact with the backplane. Without this clamp, the GATE output of a powered-down controller would be floating and the MOSFET reverse transfer

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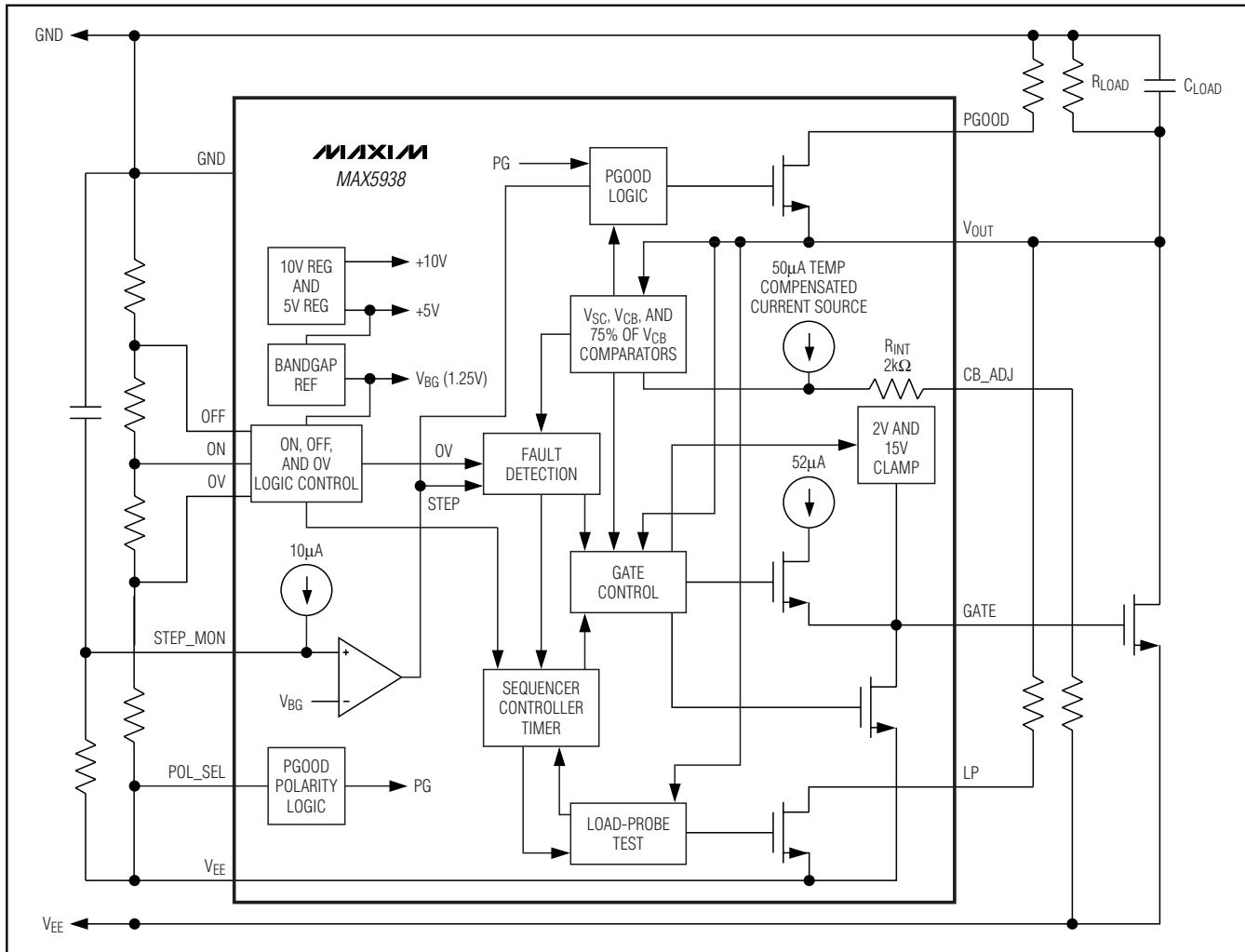


Figure 1. Functional Diagram

capacitance (gate-to-drain) would pull up and turn on the MOSFET gate when the MOSFET drain is rapidly pulled up by the  $V_{IN}$  step during backplane contact. The MAX5938 GATE clamp can overcome the gate-to-drain capacitance of large power MOSFETs with added slew-rate control ( $C_{SLEW}$ ) capacitors while eliminating the need for additional gate-to-source capacitance. The MAX5938 keeps the MOSFET off indefinitely if the supply voltage is below the user-set ON and OFF thresholds, if the supply voltage is above the user-set overvoltage (OV) threshold, or if a short circuit (user-defined) is detected in the load connected to the drain of the power MOSFET.

The MAX5938 conducts a load-probe test after contact transients from the hot plug-in have settled. This follows the MAX5938 power-up (when the ON, OFF, and OV conditions have been met for 220ms ( $t_{LP}$ )) and prior to the turn-on of the power MOSFET. This test pulls a user-programmable current through the load (1A, max) for up to 220ms ( $t_{LP}$ ) and tests for a voltage of 200mV across the load at  $V_{OUT}$  (Figure 3). This current is set by an external resistor,  $R_{LP}$  (Figure 17) between  $V_{OUT}$  and LP. When the voltage across the load exceeds 200mV, the test is truncated and the GATE turn-on sequence is started. If at the end of the 200ms ( $t_{LP}$ ) test period the voltage across the load has not reached 200mV, the load is assumed to be shorted and the current to the load from



# -48V Hot-Swap Controller with $V_{IN}$ Step Immunity, No $R_{SENSE}$ , and Overvoltage Protection

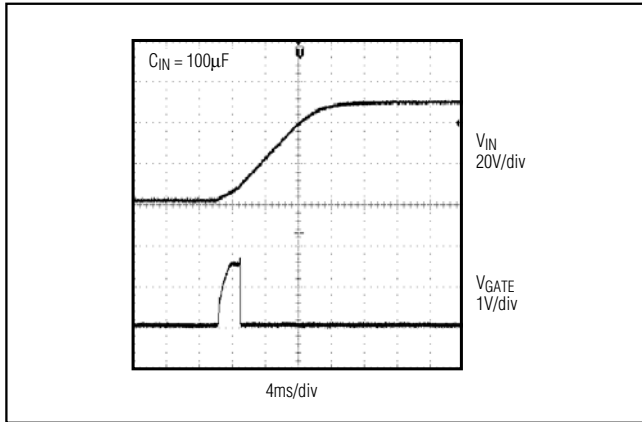


Figure 2. GATE Voltage Clamp During Power-Up

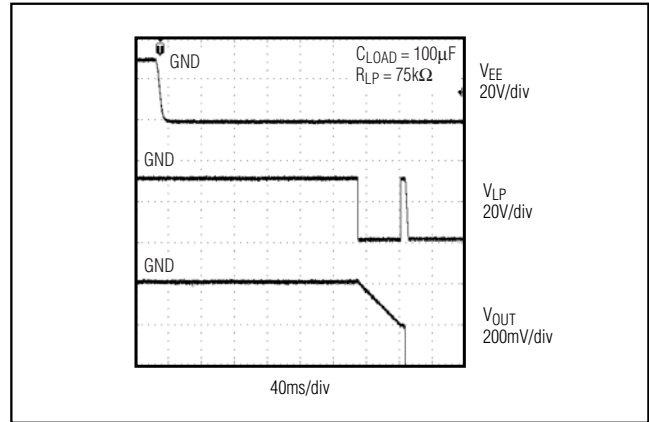


Figure 3. Load-Probe Test During Initial Power-Up

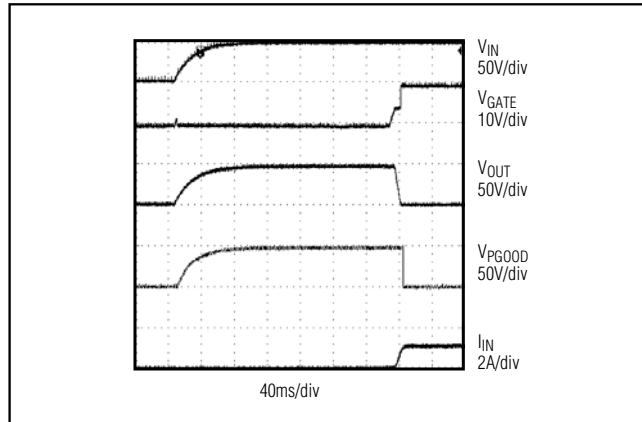


Figure 4. MAX5938 Normal Startup ( $POL\_SEL = \text{Floating}$ )

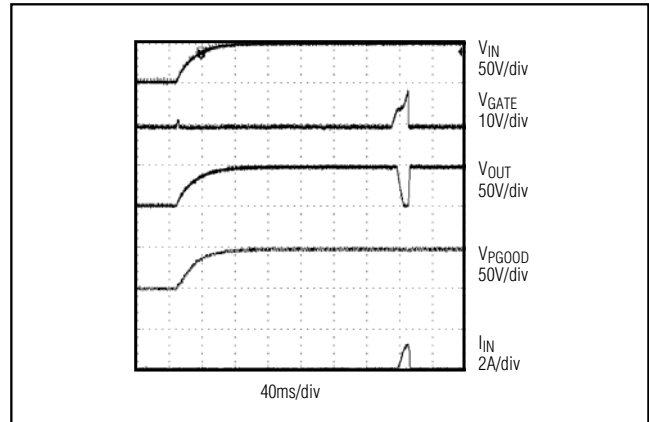


Figure 5. MAX5938 Startup into Fault Condition ( $POL\_SEL = \text{Floating}$ )

LP is shut off. The MAX5938A times out for  $16 \times t_{LP}$  then retry the load-probe test. The MAX5938L latches the fault condition indefinitely until ON and OFF are cycled low for 1.5ms or the power is recycled. See the *Applications Information* for recommendations on selecting  $R_{LP}$  to set the load-probe current level.

Upon successful completion of the load-probe test, the MAX5938 enters the power-up GATE cycle and begins ramping the GATE voltage with a  $52\mu\text{A}$  current source. This current source is restricted if  $V_{OUT}$  begins to ramp down faster than the default  $9\text{V/ms}$  slew rate. The  $V_{OUT}$  slew rate can be reduced to below  $9\text{V/ms}$  by adding  $C_{SLEW}$  from GATE to  $V_{OUT}$ . Charging up GATE enhances the power MOSFET in a controlled manner and ramping  $V_{OUT}$  at a user-settable rate controls the inrush current from the backplane. The MAX5938 continues to charge up the GATE until one of two events occurs: a normal power-up GATE cycle is completed or a power-up-to-fault-management fault is detected (see

the *GATE Cycles* section in *Appendix A*). In a normal power-up GATE cycle, the voltage at  $V_{OUT}$  (referenced to  $V_{EE}$ ) ramps to below 74% of the programmed circuit-breaker threshold voltage,  $V_{CB}$ . At this time, the remaining GATE voltage is rapidly pulled up to full enhancement. PGOOD is asserted 1.26ms after GATE is fully enhanced (see Figure 4). If the voltage at  $V_{OUT}$  remains above 74% of the programmed  $V_{CB}$  (when GATE reaches 90% of full enhancement), then a power-up-to-fault-management fault has occurred). GATE is rapidly pulled to  $V_{EE}$ , turning off the power MOSFET and disconnecting the load. PGOOD remains deasserted and the MAX5938 enters the fault management mode (Figure 5).

When the power MOSFET is fully enhanced, the MAX5938 monitors the drain voltage ( $V_{OUT}$ ) for circuit-breaker and short-circuit faults. The MAX5938 makes use of the power MOSFET's  $R_{DS(ON)}$  as the current-sense resistance to detect excessive current through

## -48V Hot-Swap Controller with $V_{IN}$ Step Immunity, No $R_{SENSE}$ , and Overvoltage Protection

the load. The short-circuit threshold voltage,  $V_{SC}$ , is twice  $V_{CB}$  ( $V_{SC} = 2 \times V_{CB}$ ) and is set by adjusting the resistance between  $CB\_ADJ$  and  $V_{EE}$ . There is an internal  $2k\Omega$  precision-trimmed resistor and an internal  $50\mu A$  current source at  $CB\_ADJ$ , which results in the minimum or default  $V_{SC}$  of  $100mV$  when  $CB\_ADJ$  is connected to  $V_{EE}$ . The current source is temperature compensated (increasing with temperature) to track the normalized temperature coefficient of  $R_{DS(ON)}$  for typical power MOSFETs.

When the load current is increased during full enhancement, this causes  $V_{OUT}$  to exceed  $V_{CB}$  but remains less than  $V_{SC}$ , and starts the  $1.2ms$  circuit-breaker glitch rejection timer. At the end of the glitch rejection period, if  $V_{OUT}$  still exceeds  $V_{CB}$ , the  $GATE$  is immediately pulled to  $V_{EE}$  ( $330ns$ ),  $PGOOD$  is deasserted, and the part enters fault management. Alternatively, during full enhancement when  $V_{OUT}$  exceeds  $V_{SC}$ , there is no glitch rejection timer.  $GATE$  is immediately pulled to  $V_{EE}$ ,  $PGOOD$  is deasserted, and the part enters fault management.

The  $V_{IN}$  step immunity provides a means for transitioning through a large step increase in  $V_{IN}$  with minimal backplane inrush current and without shutting down the load. Without  $V_{IN}$  step immunity (when the power MOSFET is fully enhanced), a step increase in  $V_{IN}$  will result in a high inrush current and a large step in  $V_{OUT}$ , which can trip the circuit breaker.

With  $V_{IN}$  step immunity, the  $STEP\_MON$  input detects the step before a short circuit is detected at  $V_{OUT}$  and alters the MAX5938 response to  $V_{OUT}$  exceeding  $V_{SC}$  due to the step. The  $1.25V$  voltage threshold at  $STEP\_MON$  and a  $10\mu A$  current source at  $STEP\_MON$

allow the user to set the sensitivity of the step detection with an external resistor to  $V_{EE}$ . A capacitor is placed between  $GND$  and the  $STEP\_MON$  input, which in conjunction with the resistor, sets the  $STEP\_MON$  time constant.

When a step is detected by the  $STEP\_MON$  input rising above its threshold ( $STEP_{TH}$ ), the overcurrent fault management is blocked and remains blocked as long as  $STEP_{TH}$  is exceeded. When  $STEP_{TH}$  is exceeded, the MAX5938 takes no action until  $V_{OUT}$  rises above  $V_{SC}$  or above  $V_{CB}$  for the  $1.2ms$  circuit-breaker glitch rejection period. When either of these conditions occurs, a step  $GATE$  cycle begins and the  $GATE$  is immediately brought to  $V_{EE}$ , which turns off the power MOSFET to minimize the resulting inrush current surge from the backplane.  $PGOOD$  remains asserted.  $GATE$  is held at  $V_{EE}$  for  $350\mu s$ , and after about  $1ms$ , begins to ramp up, enhancing the power MOSFET in a controlled manner as in the power-up  $GATE$  cycle. This provides a controlled inrush current to charge the load capacitance to the new supply voltage (see the *GATE Cycles* section in *Appendix A*).

As in the case of the power-up  $GATE$  cycle, if  $V_{OUT}$  drops to less than  $74\%$  of the programmed  $V_{CB}$ , independent of the state of  $STEP\_MON$ , the  $GATE$  voltage is rapidly pulled to full enhancement.  $PGOOD$  remains asserted throughout the step (Figure 6). Otherwise, if the  $STEP\_MON$  input has decayed below its threshold but  $V_{OUT}$  remains above  $74\%$  of the programmed  $V_{CB}$  (when  $GATE$  reaches  $90\%$  of full enhancement), a step-to-fault-management fault has occurred.  $GATE$  is rapidly pulled to  $V_{EE}$ , turning off the power MOSFET and disconnecting the load;  $PGOOD$  is deasserted and the MAX5938 enters the fault management mode (Figure 7).

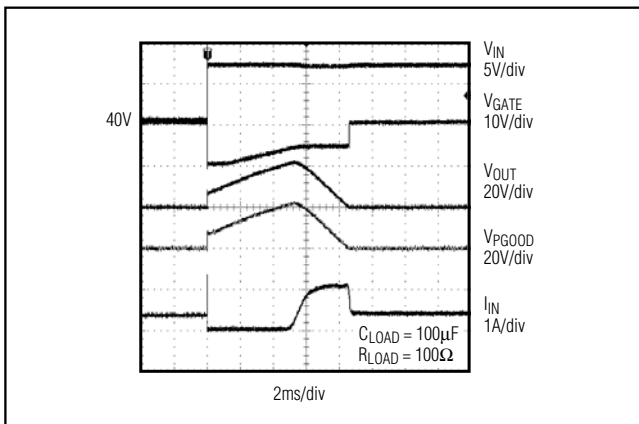


Figure 6. MAX5938 Response to a Step Input with No Fault ( $V_{OUT} < 0.75V_{CB}$ )

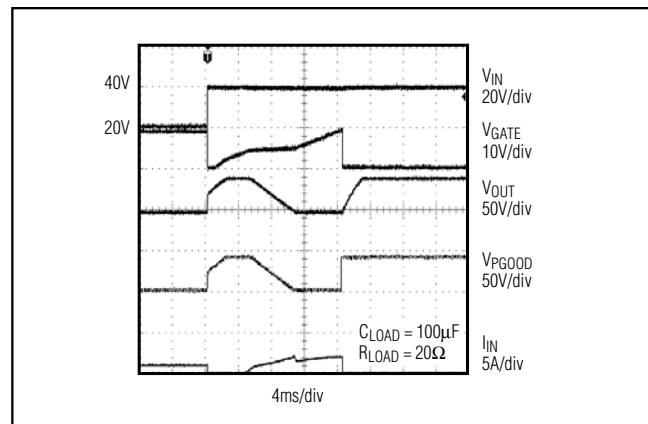


Figure 7. MAX5938 Response to a Step Input Ending in a Fault ( $V_{OUT} > 0.75V_{CB}$ )

# -48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection

## Fault Management

Fault management can be triggered by the following conditions:

- $V_{OUT}$  exceeds 74% of  $V_{CB}$  during GATE ramp at 90% of full enhancement,
- $V_{OUT}$  exceeds the  $V_{CB}$  for longer than 1.2ms during full enhancement,
- $V_{OUT}$  exceeds the  $V_{SC}$  during full enhancement,
- Load-probe test fails,
- $V_{IN}$  exceeds the programmed overvoltage (OV) limit for more than 1.5ms.

Once in the fault management mode, GATE will always be pulled to  $V_{EE}$ , which turns off the external MOSFET and always deasserts PGOOD. If CB\_ADJ is left open, short-circuit and circuit-breaker faults are ignored. The MAX5938A version has automatic retry following a fault while the MAX5938L remains latched in the fault condition.

### Autoretry Fault Management (MAX5938A)

If the MAX5938A entered fault management due to an OV fault, it will start the autoretry timer when the OV fault is removed. For circuit-breaker and short-circuit faults, the autoretry timer starts immediately. The timer times out in 3.5s (typ) after which the sequencer initiates a load-probe test and if successful, initiates a normal power-up GATE cycle.

### Latched Fault Management (MAX5938L)

When the MAX5938L enters fault management it remains in this condition indefinitely until the power is recycled or until OFF is brought below 1.25V (no time dependence) and ON is brought below 1.125V for 1.5ms (typ). In addition, if the MAX5938L enters fault management due to an overvoltage fault, the overvoltage fault must be removed. When the last of these conditions has been met, the sequencer initiates a load-probe test and if successful, a normal power-up GATE cycle begins. A manual reset circuit as in Figure 2 can be used to clear the latch.

### Circuit-Breaker Threshold

The MAX5938 has a minimum circuit-breaker threshold voltage of 50mV when CB\_ADJ is connected to  $V_{EE}$ . The  $V_{CB}$  is half  $V_{SC}$  and can be increased by placing a resistor between CB\_ADJ and  $V_{EE}$  according to the following:

$$V_{CB}(mV) = \frac{1}{2} \times V_{SC}(mV) = \frac{1}{2} \times I_{CB\_ADJ}(\mu A) \times [R_{INT}(k\Omega) + R_{CB\_ADJ}(k\Omega)]$$

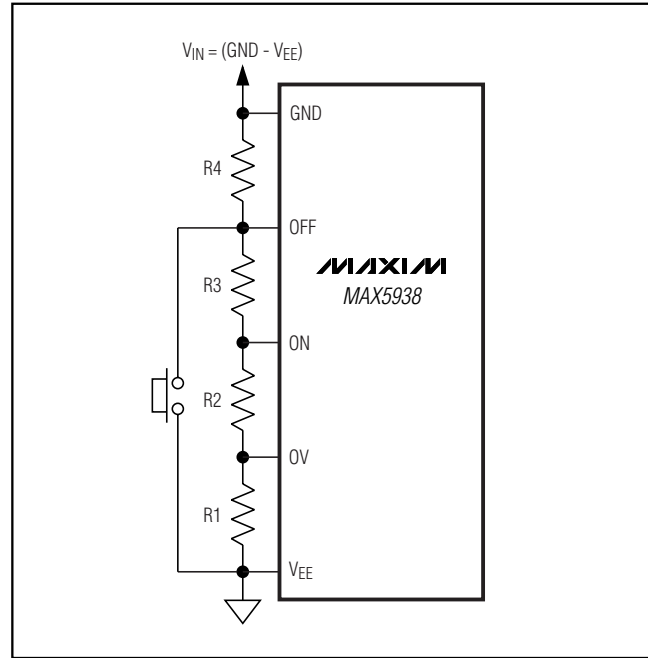


Figure 8. Resetting the MAX5938L after a Fault Condition Using a Push-Button Switch

where  $I_{CB\_ADJ} = 50\mu A$  (typ at  $+25^\circ C$ ),  $R_{INT}$  is an internal precision,  $\pm 0.5\%$ ,  $2k\Omega$  resistor at CB\_ADJ and  $R_{CB\_ADJ}$  is the external resistor between CB\_ADJ and  $V_{EE}$ . The current source  $I_{CB\_ADJ}$  is temperature-compensated (increasing with temperature) to track the normalized temperature coefficient of typical power MOSFETs.

The proper circuit-breaker threshold for an application depends on the  $R_{DS(ON)}$  of the external power MOSFET and the maximum current the load is expected to draw. To avoid false fault indication and dropping of the load, the designer must take into account the load response to voltage ripples and noise from the backplane power supply as well as switching currents in the downstream DC-DC converter that is loading the circuit. While the circuit-breaker threshold has glitch rejection that ignores ripples and noise lasting less than 1.2ms, the short-circuit detection is designed to respond very quickly (less than 330ns) to a short circuit. For this reason, set  $V_{SC}$  and  $V_{CB}$  with an adequate margin to cover all possible ripples, noise, and system current transients (see the *Setting the Circuit-Breaker and Short-Circuit Thresholds* section in the *Applications Information*).

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## Disabling Circuit-Breaker and Short-Circuit Functions

In the MAX5938, the circuit-breaker and short-circuit functions can be disabled, if desired, although this is not recommended. (See Warning note in the *PGOOD Open-Drain Output* section). This can be accomplished by leaving  $CB\_ADJ$  open. In this case,  $PGOOD$  asserts 1.26ms after  $GATE$  has ramped to 90% of full enhancement, after which  $V_{OUT}$  is ignored, resulting in the circuit-breaker and short-circuit faults being ignored.

## PGOOD Open-Drain Output

The power-good output,  $PGOOD$ , is open drain and is referenced to  $V_{OUT}$ . It asserts and latches if  $V_{OUT}$  ramps below 74% of  $V_{CB}$ , and with the built-in delay, this occurs 1.26ms after the external MOSFET becomes fully enhanced.  $PGOOD$  deasserts any time the part enters fault management.  $PGOOD$  has a delayed response to ON and OFF. The  $GATE$  will go to  $V_{EE}$  when OFF is brought below 1.25V (no time dependence) while ON is brought below 1.125V for 1.5ms. This turns off the power MOSFET and allows  $V_{OUT}$  to rise depending on the RC time constant of the load.  $PGOOD$ , in this situation, deasserts when  $V_{OUT}$  rises above  $V_{CB}$  for more than 1.4ms or above  $V_{SC}$ , whichever occurs first (see Figure 9b).

Since  $PGOOD$  is open drain, it requires an external pullup resistor to GND. Due to this external pullup,  $PGOOD$  does not follow positive  $V_{IN}$  steps as well as if it were driven by an active pullup. As a result, when  $PGOOD$  is asserted high, an apparent negative glitch appears at  $PGOOD$  during a positive  $V_{IN}$  step. This negative glitch is a result of the RC time constant of the external resistor and the  $PGOOD$  pin capacitance lagging the  $V_{IN}$  step. It is not due to switching of the internal logic. To minimize this negative transient, it may be necessary to increase the pullup current and/or to add a small amount of capacitance from  $PGOOD$  to GND to compensate for the pin capacitance.

The  $PGOOD$  output logic polarity is selected using  $POL\_SEL$  input. For an active-high output, connect  $POL\_SEL$  to  $V_{EE}$ . Leave  $POL\_SEL$  open for an active-low output.

**WARNING:** When disabling the circuit-breaker and short-circuit functions ( $CB\_ADJ$  open),  $PGOOD$  asserts 1.26ms after the power MOSFET is fully enhanced independent of  $V_{OUT}$ . Once the MOSFET is fully enhanced and ON and OFF are pulled below their respective thresholds, the  $GATE$  will be pulled to  $V_{EE}$  to turn off the power MOSFET and disconnect the load. When the cir-

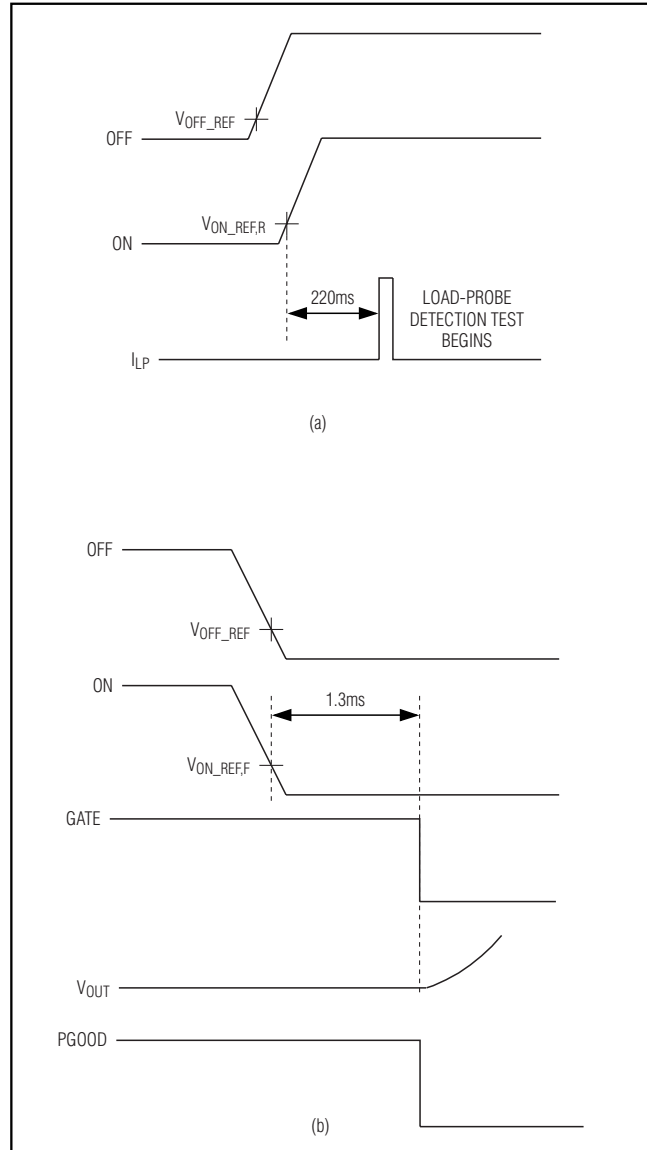


Figure 9. ON and OFF Timing Diagram

cuit-breaker and short-circuit functions are disabled and ON and OFF are cycled low,  $PGOOD$  is deasserted. In summary, when  $CB\_ADJ$  is open (once the MOSFET is fully enhanced), the MAX5938 ignores  $V_{OUT}$  and deasserts  $PGOOD$  only for an overvoltage fault, when ON and OFF are cycled low or when the power to the MAX5938 is fully recycled.

# -48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection

## Undervoltage Lockout (OFF and ON) and OV Functions

OV, ON, and OFF provide an accurate means to set the overvoltage, turn-on, and turn-off voltage levels. All three are high-impedance inputs and by use of a 4-element resistor-divider from GND to VEE, the user can set an upper VEE threshold for triggering an overvoltage fault, a middle threshold for turning the part on, and a lower threshold for turning the part off.

The input voltage threshold at OFF is 1.25V. ON has hysteresis with a rising threshold of 1.25V and a falling threshold of 1.125V. The logic of the inputs is such that both OFF and ON must be above their thresholds to latch the part on. Both OFF and ON must be below their respective thresholds to latch the part off, otherwise the part stays in its current state. There is glitch rejection on the ON input going low, which additionally requires that ON remain below its falling threshold for 1.5ms to turn off the part. A startup delay of 220ms allows contacts and voltages to settle prior to initiating the startup sequence. This startup delay is from a valid ON condition until the start of the load-probe test.

The OV input has hysteresis with a rising threshold of 1.25V and a falling threshold of 1.125V. The OV input also has a rising fault transient delay of 1.5ms. When OV rises above its threshold, an OV GATE cycle is immediately initiated (see the *GATE Cycles* section in *Appendix A*). The GATE output is brought to VEE with about 300ns of propagation delay. If the OV input drops below its falling threshold before the fault transient delay of about 1.5ms, the device will not enter fault management mode and the GATE output will ramp up to fully enhance the external MOSFET (Figure 10). Otherwise, an OV fault occurs (Figure 11). See the *Setting ON, OFF, and OV Voltage Levels* section in the *Applications Information*.

## Output Voltage (VOUT) Slew-Rate Control

The VOUT slew rate controls the inrush current required to charge the load capacitor. The MAX5938 has a default internal slew rate set for 9V/ms. The internal circuit establishing this slew rate accommodates up to about 1000pF of reverse transfer capacitance (Miller Capacitance) in the external power MOSFET without effecting the default slew rate. Using the default slew rate, the inrush current required to charge the load capacitance is given by:

$$I_{NRUSH} \text{ (mA)} = C_{LOAD} \text{ (}\mu\text{F)} \times SR \text{ (V/ms)}$$

where SR = 9V/ms (default, typ).

The slew rate can be reduced by adding an external slew-rate control capacitor (CSLEW) from VOUT (the drain of the power MOSFET) to the GATE output of the MAX5938 (Figure 19). Values of CSLEW < 4700pF have little effect on the slew rate because of the default slew-rate control circuit. For CSLEW > 4700pF, the combination of CSLEW and reverse transfer capacitance of the external power MOSFET dominate the slew rate. When CSLEW > 4700pF, SR and CSLEW are inversely related as follows (Figure 18):

$$SR \text{ (V/ms)} = 23 / C_{SLEW} \text{ (nF)}$$

If the reverse transfer capacitance of the external power MOSFET is large compared to the externally added CSLEW, then it should be added to CSLEW in the equation above.

See the *Adjusting the VOUT Slew Rate* section in the *Applications Information* and Figure 18, which graphically displays the relation between CSLEW and slew rate. This section discusses specific recommendations for compensating power MOSFET parasitics that may lead to oscillation when an external CSLEW is added.

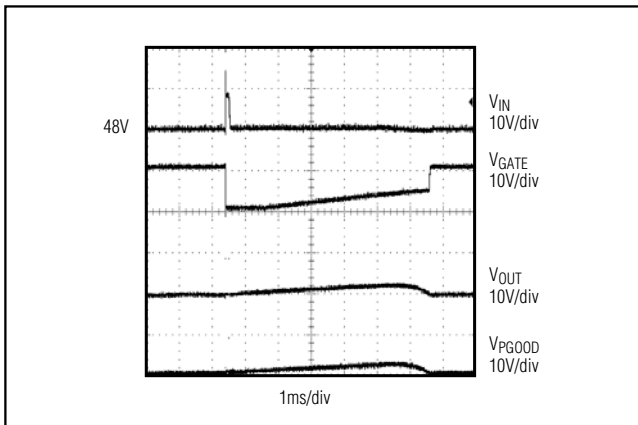


Figure 10. Overvoltage Gate Cycle Without a Fault (toV < 1.3ms)

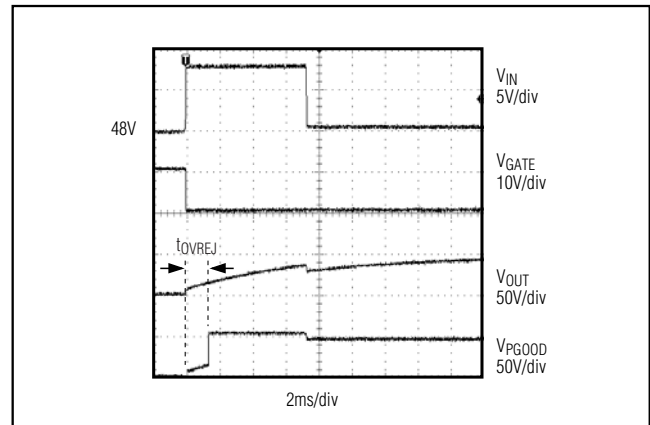


Figure 11. Overvoltage Fault (toV > 1.3ms)

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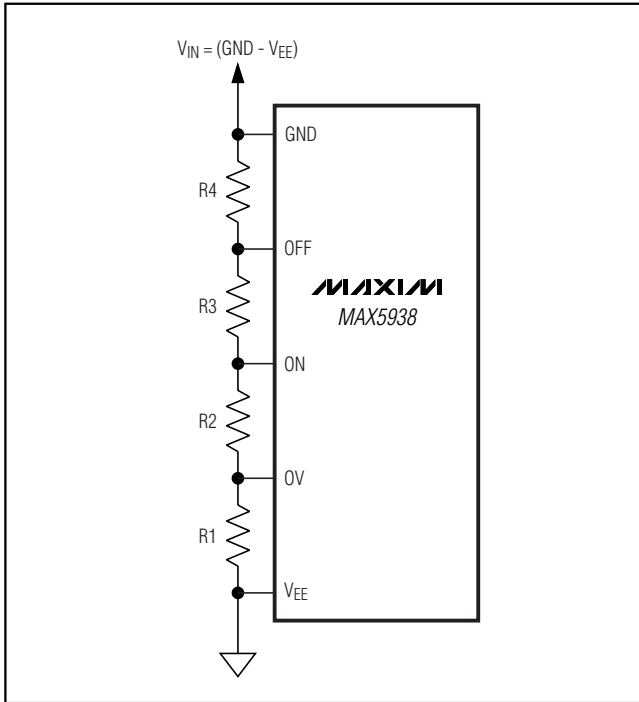


Figure 12. Programming the MAX5938's ON, OFF, and OV Thresholds

## Applications Information

### Setting ON, OFF, and OV Voltage Levels

The trip levels for ON, OFF, and OV can readily be set with a 4-element resistor-divider. Total resistance is a trade off of quiescent current, threshold tolerance due to pin input bias current (25nA), and the ability to follow very fast supply transients. Both ON and OV have hysteresis on the reference threshold voltage: the rising reference threshold is 1.25V and the falling threshold is 1.125V. The reference threshold voltage for OFF is 1.25V. In determining a set of resistors, use  $V_{REF} = 1.25V$  for ON, OFF, and OV and an  $R_{TOT} = 100k\Omega$  in this example. See Figure 12 for nomenclature. For this example, use  $V_{OV} = 80V$ ,  $V_{ON} = 42V$ , and  $V_{OFF} = 38V$  as the desired voltage trip levels.

- 1)  $R_4 = R_{TOT} \times V_{REF} / V_{OV}$
- 2)  $R_3 = R_{TOT} \times V_{REF} / V_{ON} - R_4$
- 3)  $R_2 = R_{TOT} \times V_{REF} / V_{OFF} - R_3 - R_4$
- 4)  $R_1 = R_{TOT} - R_2 - R_3 - R_4$

The exact result to three decimal places is  $R_1 = 96.711k\Omega$ ,  $R_2 = 313\Omega$ ,  $R_3 = 1.414k\Omega$ , and  $R_4 = 1.563k\Omega$ . When converted to the nearest 1% standard resistor, the values become  $R_1 = 97.6k\Omega$ ,  $R_2 = 316\Omega$ ,  $R_3 = 1.40k\Omega$ , and  $R_4 = 1.58k\Omega$ .

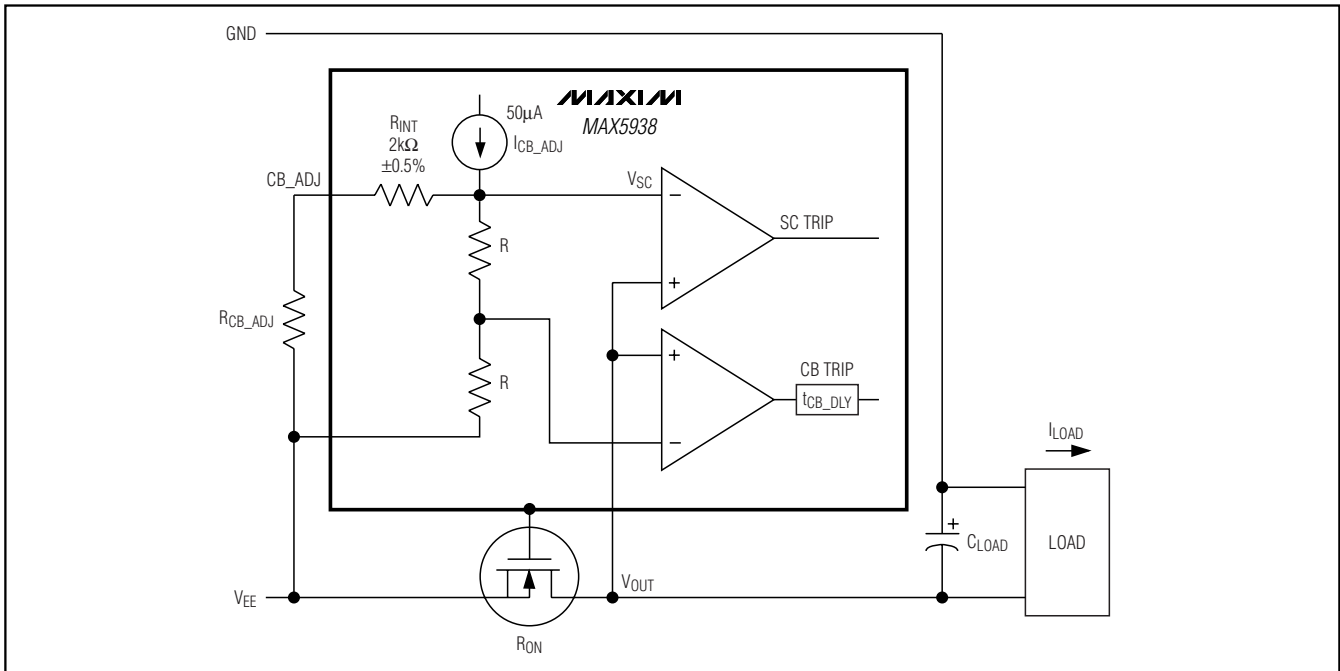


Figure 13. MAX5938 Circuit-Breaker Threshold Adjustment



## -48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection

Determine the trip voltages these values will actually yield for rising and falling voltages. Rising voltages use the  $V_{REF} = 1.25V$  reference threshold, while falling voltages use  $V_{LO} = 1.125V$  reference threshold.

- 1)  $R_{TOT} = R_1 + R_2 + R_3 + R_4$
- 2)  $V_{OV,RISING} = V_{REF} \times R_{TOT} / R_4$
- 3)  $V_{OV,FALLING} = V_{LO} \times R_{TOT} / R_4$
- 4)  $V_{ON,RISING} = V_{REF} \times R_{TOT} / (R_3 + R_4)$
- 5)  $V_{ON,FALLING} = V_{LO} \times R_{TOT} / (R_3 + R_4)$
- 6)  $V_{OFF} = V_{REF} \times R_{TOT} / (R_2 + R_3 + R_4)$

The resulting voltage levels are  $V_{OV,RISING} = 79.82V \pm 2.5\%$ ,  $V_{OV,FALLING} = 71.84V \pm 5\%$ ,  $V_{ON,RISING} = 42.32V \pm 2.5\%$ ,  $V_{ON,FALLING} = 38.09V \pm 5\%$ , and  $V_{OFF} = 38.26V \pm 2.5\%$ . The voltage tolerance does not account for the tolerance of the resistors.

### Setting the Circuit-Breaker and Short-Circuit Thresholds

The MAX5938 can operate with a wide range of power MOSFETs to meet the requirements of almost any application. MOSFETs mentioned here are done to demonstrate certain capabilities and features of the MAX5938. They should not be construed as a recommendation or a limitation of the interoperability of the MAX5938.

In the implementation of the circuit-breaker and short-circuit functions, the MAX5938 eliminates the need for an external current-sense resistor at the source of the power MOSFET. As in any other hot-swap controller, the proper circuit-breaker threshold for an application must take into account the DC level of  $V_{OUT}$ , while at the same time accommodating the AC response of  $V_{OUT}$  to the modulation of  $V_{IN}$ . The AC response from  $V_{IN}$  to  $V_{OUT}$  is dependent on the parasitics of the load, especially the load capacitor, in conjunction with the  $R_{DS(ON)}$  of the power MOSFET. It behaves as a highly dampened second-order system. As such, this system functions as a bandpass filter from  $V_{IN}$  to  $V_{OUT}$ . The response of  $V_{OUT}$  to load-switching currents and voltage ripple and noise from the backplane power supply must be taken into account. Adequate margin must be provided between  $V_{CB}$ ,  $V_{SC}$ , and the DC level of  $V_{OUT}$ , which depends on the  $R_{DS(ON)}$  of the power MOSFET (with  $V_{GS}$  at 10V) and the maximum current the load is expected to draw. While the circuit-breaker threshold has glitch rejection for  $V_{OUT}$  excursions lasting less than 1.4ms, the short-circuit detection is designed to respond very quickly (less than 330ns) to a short circuit. In the application, select a value for  $R_{CB\_ADJ}$  resulting in a  $V_{CB}$  that exceeds the product of  $R_{DS(ON)}$

and the maximum load current plus one half the peak-to-peak AC response of  $V_{OUT}$  to load-switching currents and the noise and ripple at  $V_{IN}$ :

$$R_{DS(ON)} (\Omega) \times I_{LOAD,MAX}(mA) + 1/2 \times V_{OUT,AC} < V_{CB}(mV)$$

where

$$V_{CB}(mV) = 1/2 \times I_{CB\_ADJ}(\mu A) \times [R_{INT}(k\Omega) + R_{CB\_ADJ}(k\Omega)]$$

$R_{DS(ON)}$  in a power MOSFET has a positive temperature coefficient and the MAX5938, when placed adjacent to the power MOSFET, tracks and compensates for this temperature coefficient. In the MAX5938,  $V_{CB}$  is half of  $V_{SC}$ , which is set by placing an external resistance between  $CB\_ADJ$  and  $V_{EE}$ . The minimum (default) short-circuit threshold voltage,  $V_{SC}$ , is set by an internal  $2k\Omega$  precision-trimmed ( $\pm 0.5\%$ ) resistor providing a minimum series resistance and a temperature-compensated  $50\mu A$  ( $+25^\circ C$ ) current source. When  $CB\_ADJ$  is connected to  $V_{EE}$  this gives a 50mV circuit-breaker threshold. When an external resistor,  $R_{CB\_ADJ}$ , is placed between  $CB\_ADJ$  and  $V_{EE}$ , the new circuit-breaker threshold becomes:

$$V_{CB} (mV) = 1/2 \times V_{SC} (mV) = 1/2 \times I_{CB\_ADJ} (\mu A) \times (2k\Omega + R_{CB\_ADJ})$$

and at  $+25^\circ C$ , it becomes:

$$V_{CB} (mV) = 1/2 V_{SC} (mV) = 1/2 \times 50\mu A \times (2k\Omega + R_{CB\_ADJ})$$

The short-circuit and circuit-breaker voltages are sensed at  $V_{OUT}$ , which is the drain of the power MOSFET. The  $R_{DS(ON)}$  of the MOSFET is the current-sense resistance and so the total current through the load and load capacitance is the drain current of the power MOSFET. Accordingly, the voltage at  $V_{OUT}$  as a function of MOSFET drain current is:

$$V_{OUT} = I_{D,MOSFET} \times R_{DS(ON)}$$

The temperature compensation of the MAX5938 is designed to track the  $R_{DS(ON)}$  of the typical power MOSFET. Figure 14 shows the typical normalized tempco of the circuit-breaker threshold along with the normalized tempco of  $R_{DS(ON)}$  for several typical power MOSFETs. When determining the circuit-breaker threshold in an application go to the power MOSFET manufacturer's data sheet and locate the maximum  $R_{DS(ON)}$  at  $+25^\circ C$  with a  $V_{GS}$  of 10V. Next, find the figure presenting the tempco of normalized  $R_{DS(ON)}$  or on-resistance vs. temperature. Since this curve is in

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normalized units, typically with a value of 1 at +25°C, it is possible to multiply the curve by the drain voltage at +25°C and convert the curve to drain voltage. Now compare this curve to that of the MAX5938 normalized tempco of the circuit-breaker threshold to make a determination of the tracking error in mV between the power MOSFET [ $I_{D,MOSFET} \times R_{DS(ON)}$ ] and the MAX5938 [ $I_{CB\_ADJ} (\mu A) \times (2k\Omega + R_{CB\_ADJ})$ ] over the operating temperature range of the application.

If the tempco of the power MOSFET is greater than the MAX5938's, then additional margin in setting the circuit-breaker and short-circuit voltages will be required at higher temperatures as compared to +25°C (Figure 15). When dissipation in the power MOSFET is expected to lead to local temperature elevation relative to ambient

conditions, it becomes imperative that the MAX5938 be located as close as possible to the power MOSFET. The marginal effect of temperature differences on circuit-breaker and short-circuit voltages can be estimated from a comparative plot such as Figure 14.

## Selecting a Resistor and Capacitor for Step Monitor

When a positive  $V_{IN}$  step or ramp occurs, the  $V_{IN}$  increase results in a voltage rise at both STEP\_MON and  $V_{OUT}$  relative to  $V_{EE}$ . When the voltage at STEP\_MON is above STEP\_TH, the MAX5938 blocks short-circuit and circuit-breaker faults. During this STEP\_MON high condition, if  $V_{OUT}$  rises above  $V_{SC}$ , the MAX5938 will immediately and very rapidly pull GATE to  $V_{EE}$ . This turns off the power MOSFET to avoid inrush current spiking. GATE is held low for 350 $\mu s$ . About 1ms after the start of GATE pulldown, the MAX5938 begins to ramp GATE up to turn on the MOSFET in a controlled manner that results in ramping  $V_{OUT}$  down to the new supply level (see the GATE Cycles section in Appendix A). This occurs with the least possible disturbance to  $V_{OUT}$ , although during the brief period that the MOSFET is off, the voltage across the load droops slightly depending on the load current and load storage capacitance. PGOOD remains asserted throughout the  $V_{IN}$  step event.

The objective in selecting the resistor and capacitor for the step monitor function is to ensure that the  $V_{IN}$  steps of all anticipated slopes and magnitudes will be properly detected and blocked, which otherwise would result in a circuit-breaker or short-circuit fault. The following is a brief analysis for finding the resistor and capacitor. For a more complete analysis, see Appendix B.

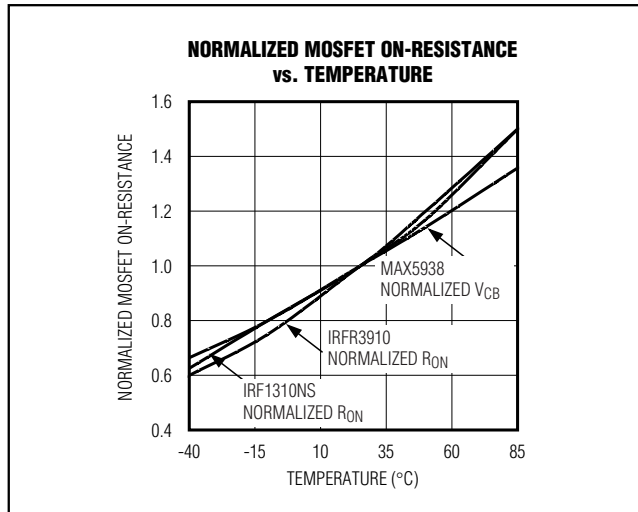


Figure 14. MAX5938 Normalized Circuit Breaker

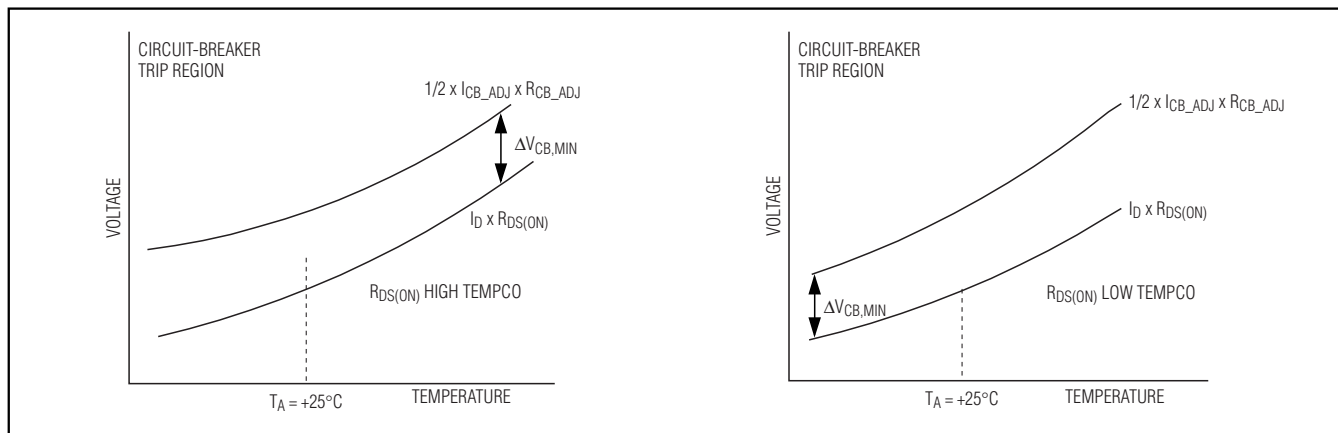


Figure 15. Circuit-Breaker Voltage Margin For High and Low Tempco Power MOSFETS



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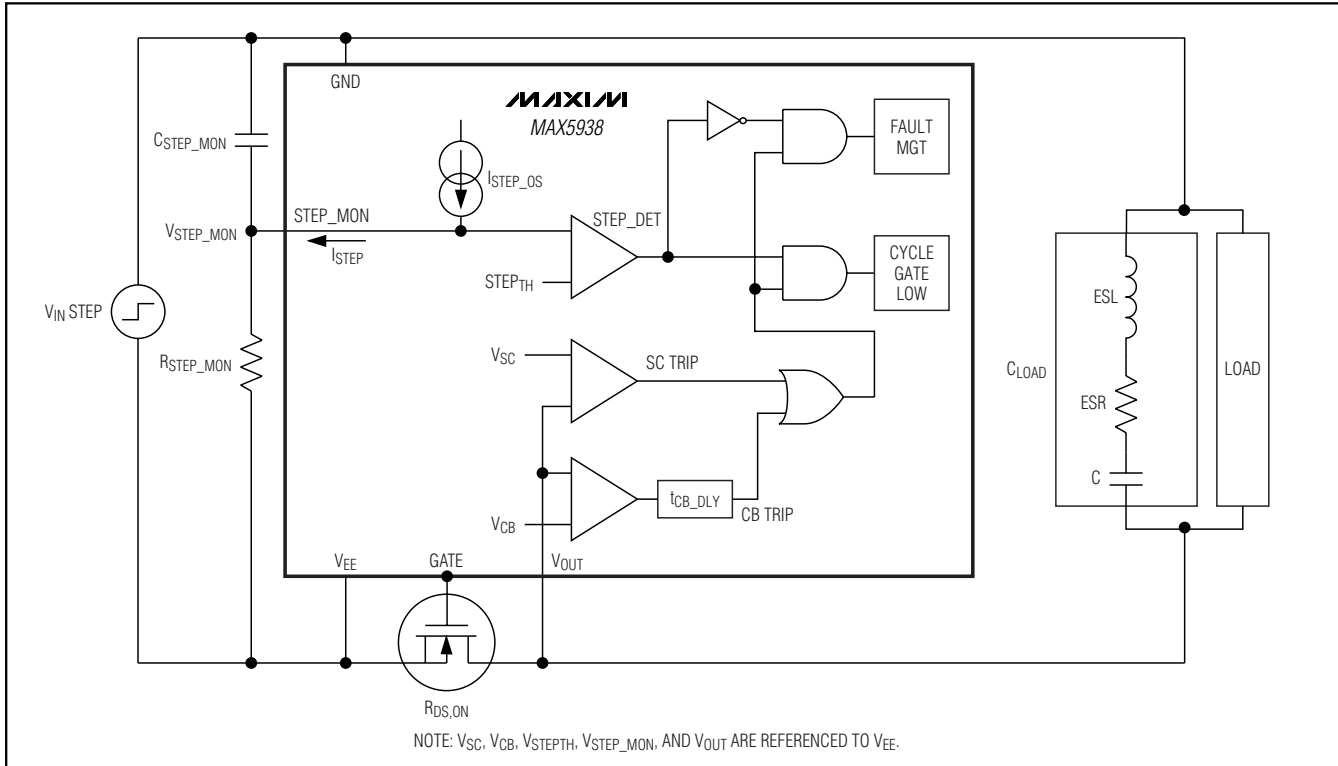


Figure 16. MAX5938 Step Immunity Functional Diagram

Figure 16 is a functional diagram exhibiting the elements of the MAX5938 involved in the step immunity function. This functional diagram shows the parallel relationship between  $V_{OUT}$  and  $V_{STEP\_MON}$ . Each has an  $I \cdot R$  component establishing the DC level prior to a step. While it is referred to as a  $V_{IN}$  step, it is the dynamic response to a finite voltage ramp that is of interest.

Given a positive  $V_{IN}$  ramp with a ramp rate of  $dV/dt$ , the approximate response of  $V_{OUT}$  to  $V_{IN}$  is:

$$V_{OUT}(t) = \frac{(dV/dt) \times \tau_C \times (1 - e^{-t/\tau_{L,eqv}})}{I_{LOAD}} + R_{DS(ON)}$$

where  $\tau_C = C_{LOAD} \times R_{DS(ON)}$  and  $\tau_{L,eqv}$  is the equivalent time constant of the load that must be found empirically (see *Appendix B*).

Similarly, the response of  $STEP\_MON$  to a  $V_{IN}$  ramp is:

$$V_{STEP\_MON}(t) = \frac{(dV/dt) \times \tau_{STEP} \times (1 - e^{-t/\tau_{STEP}})}{R_{STEP\_MON}} + 10\mu A$$

where  $\tau_{STEP} = R_{STEP\_MON} \times C_{STEP\_MON}$ .

For proper step detection,  $V_{STEP\_MON}$  must exceed  $STEP_{TH}$  prior to  $V_{OUT}$  reaching  $V_{SC}$  or within 1.4ms of  $V_{OUT}$  reaching  $V_{CB}$  (over all  $V_{IN}$  ramp rates anticipated in the application).  $V_{STEP\_MON}$  must be set below  $STEP_{TH}$

with an adequate margin,  $\Delta V_{STEP\_MON}$ , to accommodate the tolerance of both  $I_{STEP\_OS}$  ( $\pm 8\%$ ) and  $R_{STEP\_MON}$ .  $R_{STEP\_MON}$  is typically set to  $100k\Omega$ , which gives a  $\Delta V_{STEP\_MON}$  for a worst-case high of 0.36V.

The margin of  $V_{OUT}$ , with respect to  $V_{SC}$  and  $V_{CB}$ , was set when  $R_{CB\_ADJ}$  was selected as described in the *Setting the Circuit-Breaker and Short-Circuit Thresholds* section. This margin may be lower at one of the temperature extremes and if so, that value should be used in the following discussion. These margins will be called  $\Delta V_{CB}$  and  $\Delta V_{SC}$  and they represent the minimum  $V_{OUT}$  excursion required to trip the respective fault.

To set  $\tau_{STEP}$  to block all  $V_{CB}$  and  $V_{SC}$  faults for any ramp rate, find the ratio of  $\Delta V_{STEP\_MON}$  to  $\Delta V_{CB}$  and choose  $\tau_{STEP}$  so:

$$\tau_{STEP} = 1.2 \times \tau_C \times \Delta V_{STEP\_MON} / \Delta V_{CB}$$

And since  $R_{STEP\_MON} = 100k\Omega$ . This results in  $C_{STEP\_MON} = \tau_{STEP} / 100k\Omega$ .

After the first-pass component selection, if sufficient timing margin exists (see *Appendix B*), it is possible but not necessary to lower  $R_{STEP\_MON}$  below  $100k\Omega$  to reduce the sensitivity of  $STEP\_MON$  to  $V_{IN}$  noise.

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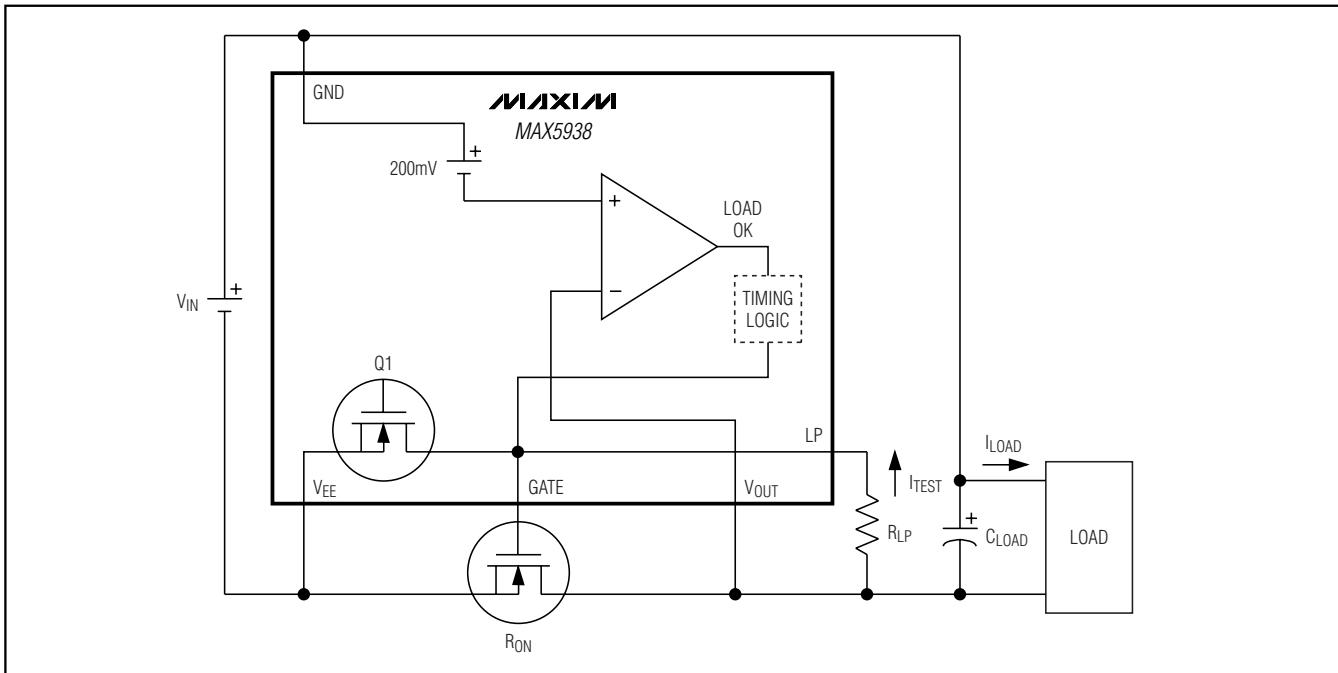


Figure 17. Load Probe Functional Diagram

Appendix B gives a more complete analysis and discussion of the step monitor function. It provides methods for the characterization of the load response to a  $V_{IN}$  ramp and graphical verification of the step monitor timing margins for a set of design parameters.

### Selecting the PGOOD Pullup Resistor

Due to the open-drain driver, PGOOD requires an external pullup resistor to GND. This resistor should be selected to minimize the current load while PGOOD is low. The PGOOD output specification for  $V_{OL}$  is 0.4V at 1mA. As described in the *Detailed Description*, the external pullup interferes with the ability of PGOOD to follow positive  $V_{IN}$  steps as well as if it were driven by an active pullup. When PGOOD is asserted high, an apparent negative glitch appears at PGOOD during a positive  $V_{IN}$  step. To minimize this negative transient it may be necessary to increase the pullup current and/or to add a small amount of capacitance from PGOOD to GND to compensate for the pin capacitance.

### Setting the Test Current Level for Load-Probe Test

The load-probe test is a current test of the load that avoids turning on the power MOSFET. The MAX5938 has an internal switch (Q1 in Figure 17) that pulls current through the load and through an external current-

limiting resistor,  $R_{LP}$ . During the test, this switch is pulsed on for up to 220ms (typ). Current is pulled through the load, which should charge up the load capacitance unless there is a short. If the voltage across the load exceeds 200mV, the test is truncated and normal power-up is allowed to proceed. If the voltage across the load does not reach 200mV in the 220ms period that the current is on, the load is assumed to be shorted and the current to the load from LP is shut off. The MAX5938A will timeout for  $16 \times t_{LP}$  then retry the load-probe test. The MAX5938L will latch the fault condition indefinitely until ON and OFF are cycled low for 1.5ms or the power is recycled.

In the application, the current-limiting resistor should be selected to minimize the current pulled through the load while guaranteeing that it will charge the maximum expected load capacitance to 220mV in 90ms. These parameters are the maximum load-probe test voltage and the minimum load-probe current pulse period, respectively. The maximum current possible is 1A, which is adequate to test a load capacitance as large as 190,000 $\mu$ F over the typical telecom operating voltage range.

$$I_{TEST} (A) = C_{LOAD,MAX} (F) \times 220mV / 90ms$$

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Since the minimum intended  $V_{IN}$  for the application will result in the lowest  $I_{TEST}$  during the load-probe test, this  $V_{IN,MIN}$  should be used to set the  $R_{LP}$ . This voltage will likely be near  $V_{ON,FALLING}$  or  $V_{OFF}$  for the application.

$$R_{TEST}(\Omega) = V_{IN,MIN} / I_{TEST} = V_{IN,MIN} \times 90ms / (C_{LOAD(MAX)} \times 220mV)$$

Example:  $V_{IN}$  operating range = 36V to 72V,  $C_{LOAD} = 10,000\mu F$ . First, find the  $R_{LP}$  that will guarantee a successful test of the load.

$$R_{LP} = 36V \times 90ms / (10,000\mu F \times 220mV) = 1.472\Omega \Rightarrow 1.47k\Omega \pm 1\%$$

Next, evaluate the  $R_{LP}$  at the maximum operating voltage to verify that it will not exceed the 1A current limit for the load-probe test.

$$I_{TEST,MAX} = V_{IN,MAX} / R_{LP} = 72V / 1.47k\Omega = 49.0mA$$

If the  $C_{LOAD(MAX)}$  is increased to 190,000 $\mu F$ , the test current will approach the limit. In this case,  $R_{LP}$  will be a much lower value and must include the internal switch resistance. To find the external series resistor value that will guarantee a successful test at the lowest supply voltage, the maximum value for the load-probe switch on-resistance of 11 $\Omega$  should be used:

$$R_{LP,TOT} = 36V \times 90ms / (190,000\mu F \times 220mV) = 90\Omega = 11\Omega + R_{LP}$$

$$R_{LP} = 77.51\Omega - 11\Omega = 66.51\Omega \Rightarrow 66.5\Omega \pm 1\%$$

Again  $R_{LP}$  must be evaluated at the maximum operating voltage to verify that it will not exceed the 1A current limit for the load-probe test. In this case, the minimum value for the load-probe switch on-resistance of 6 $\Omega$  should be used:

$$I_{TEST,MAX} = V_{IN,MAX} / R_{LP,TOT} = 72V / (66.5\Omega + 6\Omega) = 993mA$$

### **Adjusting the VOUT Slew Rate**

The default slew rate is set internally for 9V/ms. The slew rate can be reduced by placing an external capacitor from the drain of the power MOSFET to the GATE output of the MAX5938. Figure 18 shows a graph of Slew Rate vs.  $C_{SLEW}$ . This graph shows that for  $C_{SLEW} < 4700pF$  there is very little effect to the addition of external slew-rate control capacitance. This is intended so the GATE output can drive large MOSFETs with significant gate capacitance and still achieve the default slew rate. To select a slew-rate control capacitor, go into the graph with the desired slew rate and find the value of the Miller Capacitance. When  $C_{SLEW} > 4700pF$ , SR and  $C_{SLEW}$  are inversely related. Given the desired slew rate, the required  $C_{SLEW}$  is found as follows:

$$C_{SLEW}(nF) = 23 / SR (V/ms)$$

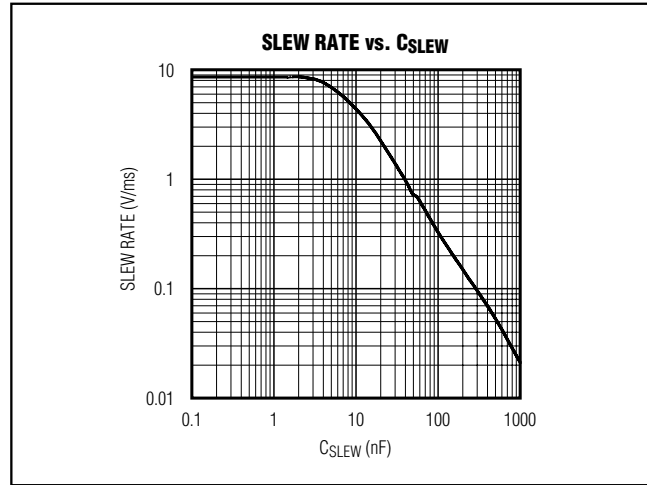


Figure 18. MAX5938 Slew Rate vs.  $C_{SLEW}$

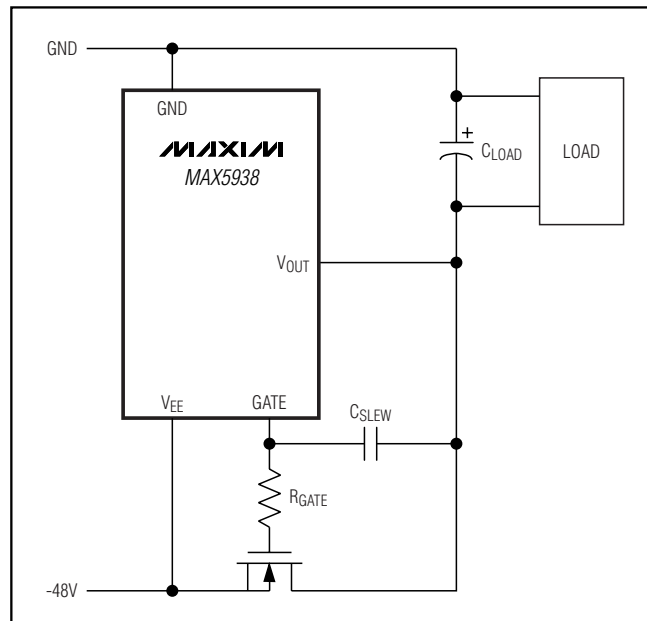


Figure 19. Adjusting the MAX5938 Slew Rate

From the data sheet of the power MOSFET find the reverse transfer capacitance (gate-to-drain capacitance) above 10V. If the reverse transfer capacitance of the external power MOSFET is 5% or more of  $C_{SLEW}$ , then it should be subtracted from  $C_{SLEW}$  in the equation above.

Figure 19 gives an example of the external circuit for controlling slew rate. Depending on the parasitics associated with the selected power MOSFET, the addition of  $C_{SLEW}$  may lead to oscillation while the MOSFET and GATE con-

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control are in the linear range. If this is an issue, an external resistor,  $R_{GATE}$ , in series with gate of the MOSFET is recommended to prevent possible oscillation. It should be as small as possible, e.g.,  $5\Omega$  to  $10\Omega$ , to avoid impacting the MOSFET turn-off performance of the MAX5938.

## Layout Guidelines

To benefit from the temperature compensation designed into the MAX5938, the part should be placed as close as possible to the power MOSFET that it is controlling. The  $V_{EE}$  pin of the MAX5938 should be placed close to the source pin of the power MOSFET and they should share a wide trace. A common top layer plane would service both the thermal and electrical requirements. The load-probe current must be taken into account. If this current is high, the layout traces and current-limiting resistor must be sized appropriately. Stray inductance must be minimized in the traces of the overall layout of the hot-swap controller, the power MOSFET and the load capacitor. Starting from the board contacts, all high-current traces should be short, wide, and direct. The potentially high pulse current pins of the MAX5938 are GATE (when pulling GATE low), load probe, and  $V_{EE}$ . Because of the nature of the hot-swap requirement no decoupling capacitor is recommended for the MAX5938. Because there is no decoupling capacitor, stray inductance may result in excessive ringing at the GND pin during power-up or during very rapid  $V_{IN}$  steps. This should be examined in every application design since ringing at the GND pin may exceed the absolute maximum supply rating for the part.

## Input Transient Protection

During hot plug-in/unplug and fast  $V_{IN}$  steps, stray inductance in the power path may cause voltage ringing above the normal input DC value, which may exceed the absolute maximum supply rating. An input transient such as that caused by lightning can also put a severe transient peak voltage on the input rail. The following techniques are recommended to reduce the effect of transients:

- 1) Minimize stray inductance in the power path using wide traces and minimize loop area including the power traces and the return ground path.
- 2) Add a high-frequency (ceramic) bypass capacitor on the backplane as close as possible to the plug-in connector (Figure 20).
- 3) Add a  $1k\Omega$  resistor in series with the MAX5938's GND pin and a  $0.1\mu F$  capacitor from GND to  $V_{EE}$  to limit transient current going into this pin.

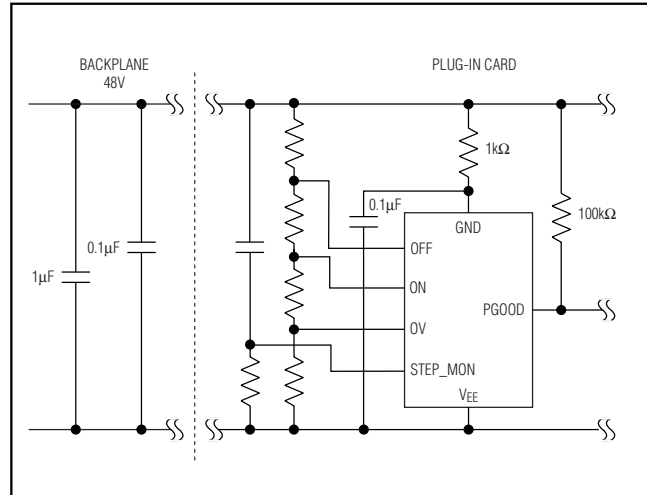


Figure 20. Protecting the MAX5938 Input from High-Voltage Transients

## Appendix A

### GATE Cycles

The power-up GATE cycle, step GATE cycle, and the OV GATE cycle are quite similar but have distinct differences. Understanding these differences may clarify application issues.

### GATE Cycle During Power-Up

The power-up GATE cycle occurs during the initial power-up of the MAX5938 and the associated power MOSFET and load. The power-up GATE cycle can result in full enhancement or in a fault (all voltages are relative to  $V_{EE}$ ).

### Power-Up-to-Full-Enhancement Fault:

- 1) At the beginning of the power-up sequence to the start of the power-up GATE cycle, the GATE is held at  $V_{EE}$ . Following a successful completion of the load-probe test, GATE is held at  $V_{EE}$  for an additional  $350\mu s$  and then is allowed to float for  $650\mu s$ . At this point, the GATE begins to ramp with  $52\mu A$  charging the gate of the power MOSFET. [GATE turn-on]
- 2) When GATE reaches the gate threshold voltage of the power MOSFET,  $V_{OUT}$  begins to ramp down toward  $V_{EE}$ . [ $V_{OUT}$  ramp]
- 3) When  $V_{OUT}$  ramps below  $74\% V_{CB}$ , the GATE is rapidly pulled to full enhancement and the power-up GATE cycle is complete.  $1.26ms$  after GATE is pulled to full enhancement, PGOOD asserts. [Full enhancement]

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## **Power-Up-to-Fault-Management Fault:**

- 1) Same as step 1 above. [GATE turn-on]
- 2) Same as step 2 above. [ $V_{OUT}$  ramp]
- 3) GATE ramps to 90% of full enhancement while  $V_{OUT}$  remains above 74%  $V_{CB}$ , at which point the GATE is rapidly pulled to  $V_{EE}$  and fault management is initiated. [Fault management]

### ***GATE Cycle During $V_{IN}$ Step***

A step GATE cycle occurs only after a successful power-up GATE cycle to full enhancement occurs and as a result of a positive  $V_{IN}$  step (all voltages are relative to  $V_{EE}$ ).

## **Step-to-Full-Enhancement Fault:**

- 1) A  $V_{IN}$  step occurs resulting in  $STEP\_MON$  rising above  $STEP_{TH}$  before  $V_{OUT}$  rises above  $V_{SC}$ . [Step detection]
- 2) After a step is detected,  $V_{OUT}$  rises above  $V_{SC}$  in response to the step. When  $V_{OUT}$  rises above  $V_{SC}$ , GATE is immediately pulled to  $V_{EE}$ , rapidly turning off the power MOSFET. GATE is held at  $V_{EE}$  for 350 $\mu$ s to damp any ringing. Once GATE is pulled to  $V_{EE}$ , the gate cycle has begun and  $STEP\_MON$  can safely drop below  $STEP_{TH}$  and successfully complete a step GATE cycle to full enhancement without initiating fault management. [GATE pulldown]
- 3) Following the 350 $\mu$ s of GATE pulldown, GATE is allowed to float for 650 $\mu$ s. At this point, the GATE begins to ramp with 52 $\mu$ A charging the gate of the power MOSFET. [GATE turn-on]
- 4) When GATE reaches the gate threshold voltage of the power MOSFET,  $V_{OUT}$  begins to ramp down toward the new lower  $V_{EE}$ . In the interval where GATE is below the MOSFET threshold, the MOSFET is off and  $V_{OUT}$  will droop depending on the RC time constant of the load. [ $V_{OUT}$  ramp]
- 5) When  $V_{OUT}$  ramps below 74%  $V_{CB}$ , the GATE pulls rapidly to full enhancement and the step GATE cycle is complete. If  $STEP\_MON$  remains above  $STEP_{TH}$  when GATE has ramped to 90% of full enhancement and  $V_{OUT}$  remains above 74% of  $V_{CB}$ , GATE remains at 90% and is not pulled to full enhancement. In this condition, if  $V_{OUT}$  drops below 74% of  $V_{CB}$  before  $STEP\_MON$  drops below  $STEP_{TH}$ , GATE is rapidly pulled to full enhancement and the step GATE cycle is complete. PGOOD remains asserted throughout the step GATE cycle. [Full enhancement]

## **Step-to-Fault-Management Fault:**

- 1) Same as step 1 above. [Step detection]
- 2) Same as step 2 above. [GATE pulldown]
- 3) Same as step 3 above. [GATE turn-on]
- 4) Same as step 4 above. [ $V_{OUT}$  ramp]
- 5) If  $STEP\_MON$  is below  $STEP_{TH}$  when GATE ramps to 90% of full enhancement and  $V_{OUT}$  remains above 74%  $V_{CB}$ , GATE is rapidly pulled to  $V_{EE}$ . Fault management is initiated and PGOOD is deasserted. If  $STEP\_MON$  is above  $STEP_{TH}$  when GATE ramps to 90% of full enhancement and  $V_{OUT}$  remains above 74% of  $V_{CB}$ , GATE remains at 90%. It is not pulled to full enhancement nor is it pulled to  $V_{EE}$ . In this condition, if  $V_{OUT}$  drops below 74% of  $V_{CB}$  before  $STEP\_MON$  drops below  $STEP_{TH}$ , GATE is rapidly pulled to full enhancement and a fault is avoided. Conversely, if  $STEP\_MON$  drops below  $STEP_{TH}$  first, the GATE is rapidly pulled to  $V_{EE}$ , fault management is initiated, and PGOOD is deasserted. [Fault management]

It should be emphasized that while  $STEP\_MON$  remains above  $STEP_{TH}$  the current fault management is blocked. During this time it is possible for there to be multiple events involving  $V_{OUT}$  rising above  $V_{SC}$  then falling below 74%  $V_{CB}$ . In each of these events, when  $V_{OUT}$  rises above  $V_{SC}$ , a full GATE cycle is initiated where GATE is first pulled low then allowed to ramp up. Then finally, when  $V_{OUT}$  conditions are met, it will be fully enhanced.

### ***GATE Cycle During Momentary Overvoltage***

An OV GATE cycle occurs only after a successful power-up GATE cycle to full enhancement and as a result of a momentary excursion of OV above the OV threshold voltage. An OV GATE cycle does not result in an OV fault unless OV remains above the threshold for more than 1.5ms (all voltages are relative to  $V_{EE}$ ).

## **OV GATE Cycle to Full enhancement:**

- 1) When OV rises above the OV threshold voltage, GATE is immediately pulled to  $V_{EE}$ , rapidly turning off the power MOSFET. GATE is held at  $V_{EE}$  indefinitely while OV is above the OV threshold voltage. It is held for an additional 350 $\mu$ s to damp any ringing. [GATE pulldown]
- 2) Following the GATE pulldown, GATE is allowed to float for 650 $\mu$ s. At this point, the GATE begins to ramp with 52 $\mu$ A charging the gate of the power MOSFET. [GATE turn-on]

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- 3) When GATE reaches the gate threshold voltage of the power MOSFET,  $V_{OUT}$  begins to ramp back down toward  $V_{EE}$ . In the interval where GATE is below the MOSFET threshold, the MOSFET is off and  $V_{OUT}$  will droop depending on the RC time constant of the load. [ $V_{OUT}$  ramp]
- 4) When  $V_{OUT}$  ramps below 74%  $V_{CB}$ , the GATE is rapidly pulled to full enhancement and the OV GATE cycle is complete. [Full enhancement]

## OV GATE Cycle to Fault management:

- 1) Same as step 1 above. [GATE pulldown]
- 2) Same as step 2 above. [GATE turn-on]
- 3) Same as step 3 above. [ $V_{OUT}$  ramp]
- 4) If GATE ramps to 90% of full enhancement and  $V_{OUT}$  remains above 74%  $V_{CB}$ , GATE is rapidly pulled to  $V_{EE}$ , fault management is initiated, and PGOOD is deasserted. [Fault management]

## GATE Output

GATE is a complex output structure and its condition at any moment is dependent on various timing sequences in response to multiple inputs. A diode to  $V_{EE}$  prevents negative excursions. For positive excursions, the states are:

- 1) Power-off with 2V clamp.
- 2)  $8\Omega$  pulldown to  $V_{EE}$ .
  - a. Continuous during startup delay and during fault conditions.
  - b. Pulsed following detected step or OV condition.
- 3) Floating with 16V clamp [prior to GATE ramp].
- 4)  $52\mu\text{A}$  current source with 16V clamp [GATE ramp].
- 5) Pullup to internal 10V supply with 16V clamp [full enhancement].

## Appendix B

### Step Monitor Component Selection Analysis

As mentioned previously in the *Setting the Circuit-Breaker and Short-Circuit Thresholds* section, the AC response from  $V_{IN}$  to  $V_{OUT}$  is dependent on the parasitics of the load. This is especially true for the load capacitor in conjunction with the power MOSFET's  $R_{DS(ON)}$ . The load capacitor (with parasitic ESR and LSR) and the power MOSFET's  $R_{DS(ON)}$  can be modeled as a heavily damped second-order system. As such, this system functions as a bandpass filter from

$V_{IN}$  to  $V_{OUT}$  limiting the ability of  $V_{OUT}$  to follow the  $V_{IN}$  ramp. STEP\_MON lags the  $V_{IN}$  ramp with a first-order RC response, while  $V_{OUT}$  lags with an overdamped second-order response.

Given a positive  $V_{IN}$  ramp with a ramp rate of  $dV/dt$ , the approximate response of  $V_{OUT}$  to  $V_{IN}$  is:

$$V_{OUT}(t) = (dV/dt) \times \tau_C \times (1 - e^{-t/\tau_{L,eqv}}) + R_{DS(ON)} \times I_{LOAD} \quad (\text{Equation 1})$$

where  $\tau_C = C_{LOAD} \times R_{DS(ON)}$ .

Equation 1 is a simplification for the overdamped second-order response of the load to a ramp input,  $\tau_C = C_{LOAD} \times R_{DS(ON)}$  and corresponds to the ability of the load capacitor to transfer  $dV/dt$  current to the fully enhanced power MOSFET's  $R_{DS(ON)}$ . The equivalent time constant of the load ( $\tau_{L,eqv}$ ) accounts for the parasitic series inductance and resistance of the capacitor and board interconnect. To characterize the load dynamic response to  $V_{IN}$  ramps, determine  $\tau_{L,eqv}$  empirically with a few tests.

Similarly, the response of STEP\_MON to a  $V_{IN}$  ramp is:

$$V_{STEP\_MON}(t) = (dV/dt) \times \tau_{STEP} \times (1 - e^{-t/\tau_{STEP}}) + 10\mu\text{A} \times R_{STEP\_MON} \quad (\text{Equation 2})$$

where  $\tau_{STEP} = R_{STEP\_MON} \times C_{STEP\_MON}$ .

For proper step detection,  $V_{STEP\_MON}$  must exceed  $STEP_{TH}$  prior to  $V_{OUT}$  reaching  $V_{SC}$  or within 1.4ms of  $V_{OUT}$  reaching  $V_{CB}$  (or overall  $V_{IN}$  ramp rates anticipated in the application). It is impossible to give a fixed set of design guidelines that rigidly apply over the wide array of applications using the MAX5938. There are, however, limiting conditions and recommendations that should be observed.

One limiting condition that must be observed is to ensure that the STEP\_MON time constant,  $\tau_{STEP}$ , is not so low that at the lowest ramp rate, the anticipated  $STEP_{TH}$  cannot be obtained. The product  $(dV/dt) \times \tau_{STEP} = \tau_{STEP\_MON,MAX}$ , is the maximum differential voltage at STEP\_MON if the  $V_{IN}$  ramp were to continue indefinitely. A related condition is setting the STEP\_MON voltage below  $STEP_{TH}$  with adequate margin,  $\Delta V_{STEP\_MON}$ , to accommodate the tolerance of both  $I_{STEP\_OS}$  ( $\pm 8\%$ ) and  $R_{STEP\_MON}$ . In determining  $\tau_{STEP\_MON}$ , use the  $9.2\mu\text{A}$  limit to ensure sufficient margin with worst-case  $I_{STEP\_OS}$ .

The margin of  $V_{OUT}$  (with respect to  $V_{SC}$  and  $V_{CB}$ ) is set when  $R_{CB\_ADJ}$  is selected as described in the *Setting the Circuit-Breaker and Short-Circuit Thresholds* section. This margin may be lower at one of the temperature extremes and if so, that value should be used in the following discussion. These margins will be called

# -48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection

$\Delta V_{CB}$  and  $\Delta V_{SC}$  and they represent the minimum  $V_{OUT}$  excursion required to trip the respective fault.  $R_{STEP\_MON}$  will typically be set to  $100k\Omega \pm 1\%$ . This gives a  $\Delta V_{STEP\_MON}$  of 0.25V, a worst-case low of 0.13V, and a worst-case high of 0.37V. In finding  $\tau_{STEP}$  in the equation below, use  $\Delta V_{STEP\_MON} = 0.37V$  to ensure sufficient margin with worst-case  $I_{STEP\_OS}$ .

To set  $\tau_{STEP}$  to block all  $V_{CB}$  and  $V_{SC}$  faults for any ramp rate, find the ratio of  $\Delta V_{STEP\_MON}$  to  $\Delta V_{CB}$  and choose  $\tau_{STEP}$  so:

$$\tau_{STEP} = 1.2 \times \tau_C \times \Delta V_{STEP\_MON} / \Delta V_{CB}$$

and since  $R_{STEP\_MON} = 100k\Omega$ :

$$C_{STEP\_MON} = \tau_{STEP} / R_{STEP\_MON} = \tau_{STEP} / 100k\Omega$$

After the first-pass component selection, if sufficient timing margin exists, it is possible (but not necessary) to lower  $R_{STEP}$  below  $100k\Omega$  to reduce the sensitivity of  $STEP\_MON$  to  $V_{IN}$  noise.

### Verification of the Step Monitor Timing

It is prudent to verify conclusively that all circuit-breaker and short-circuit faults will be blocked for all ramp rates. To do this, some form of graphical analysis is recommended but first, find the value of  $\tau_{L,eqv}$  of the load by a series of ramp tests as indicated earlier. These tests include evaluating the load with a series of  $V_{IN}$  ramps of increasing ramp rates and monitoring the rate of rise of  $V_{OUT}$  during the ramp. Each  $V_{IN}$  ramp should have a constant slope. The  $V_{OUT}$  response data must be taken only during the positive ramp. Data taken after  $V_{IN}$  has leveled off at the new higher value must not be used.

Figure 21 shows the load in parallel with the load capacitor,  $C_{LOAD}$ , and the parallel connection in series with the power MOSFET, which is fully enhanced with  $V_{GS} = 10V$ . The objective is to determine  $\tau_{L,eqv}$  from the  $V_{OUT}$  response.

Figure 22 shows the general response of  $V_{OUT}$  to a  $V_{IN}$  ramp over time  $t$ . Equation 1 gives the response of  $V_{OUT}$  to a ramp of  $dV/dt$ . The product  $(dV/dt) \times \tau_C = \Delta V_{OUT(max)}$  or the maximum  $V_{OUT}$  voltage differential if the  $V_{IN}$  ramp were to continue indefinitely. The parameter of interest is  $\Delta V_{OUT}$  due to the ramp  $dV/dt$ , thus it is necessary to subtract the DC shift in  $V_{OUT}$  due to the load resistance. For some loads, which are relatively independent of supply voltage, this may be insignificant.

$$V_{OUT}(t) = V_{OUT}(t) - R_{DS(ON)} \times I_{LOAD}$$

where  $I_{LOAD}$  is a function of the  $V_{OUT}$  level that should be determined separately with DC tests.

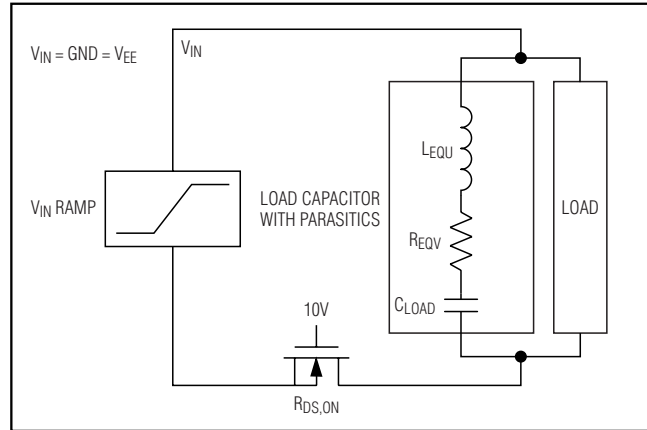


Figure 21.  $V_{IN}$  Ramp Test Of Load

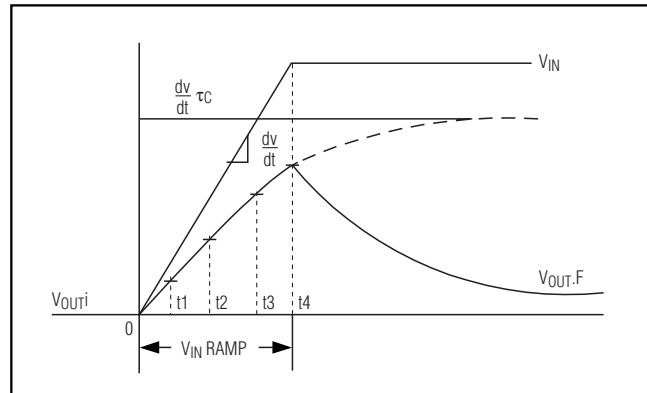


Figure 22. General Response of  $V_{OUT}$  to a  $V_{IN}$  Ramp

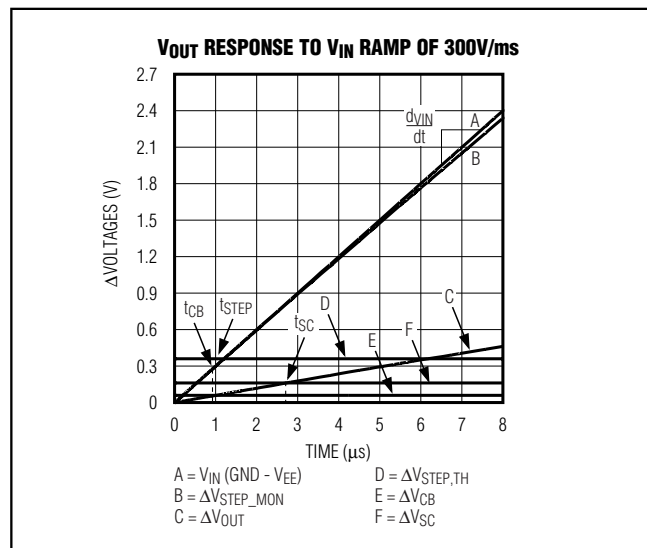


Figure 23.  $V_{OUT}$  Response to  $V_{IN}$  Ramp of 300V/ms



# **-48V Hot-Swap Controller with $V_{IN}$ Step Immunity, No $R_{SENSE}$ , and Overvoltage Protection**

At any time (t) the  $\Delta V_{OUT}$  fraction of  $\Delta V_{OUT(max)}$  is:

$$\Delta V_{OUT}(t) / [(dV/dt) \times \tau_C] = (1 - e^{-t / \tau_{L,eqv}})$$

If  $V_{OUT}(t)$  is measured at time t, then the equivalent time constant of the load is found from:

$$\tau_{L,eqv} = -t / \ln(1 - \Delta V_{OUT} / [(dV/dt) \times \tau_C])$$

As mentioned earlier, several measurements of  $\Delta V_{OUT}$  at times t1, t2, t3, and t4 should be made during the ramp. Each of these may result in slightly different values of  $\tau_{L,eqv}$  and all values should then be averaged. In making the measurements, the  $V_{IN}$  ramp duration should be such that  $\Delta V_{OUT}$  reaches 2 or 3 times the selected  $\Delta V_{SC}$ . The ramp tests should include three ramp rates:  $\Delta V_{SC} / \tau_C$ ,  $2 \times \Delta V_{SC} / \tau_C$ , and  $4 \times \Delta V_{SC} / \tau_C$ . The values of  $\tau_{L,eqv}$  may vary over the range of slew rates due to measurement error, nonlinear dynamics in the load, and due to the fact that Equation 1 is a simplification from a higher order dynamic system. The resulting range of  $\tau_{L,eqv}$  values should be used to validate the performance of the final design.

Having  $\tau_C$ ,  $\tau_{L,eqv}$ ,  $R_{STEP}$ , and  $C_{STEP}$  in a graphical analysis using Equation 1 and Equation 2 can verify the step monitor function by displaying the relative timing of  $t_{CB}$ ,  $t_{STEP}$ , and  $t_{SC}$ , which are the times when  $V_{CB}$ ,  $V_{STEP\_MON}$ , and  $V_{SC}$  voltage thresholds are exceeded. A simple Excel spreadsheet for this purpose can be supplied by Maxim upon request. Figures 23, 24, and 25 graphically verify a particular solution over 3 decades of  $V_{IN}$  ramp rates. In addition, Figure 25 verifies that this solution will block all circuit-breaker and short-circuit faults for even the lowest  $V_{IN}$  ramp that will cause  $V_{OUT}$  to exceed  $V_{CB}$ .

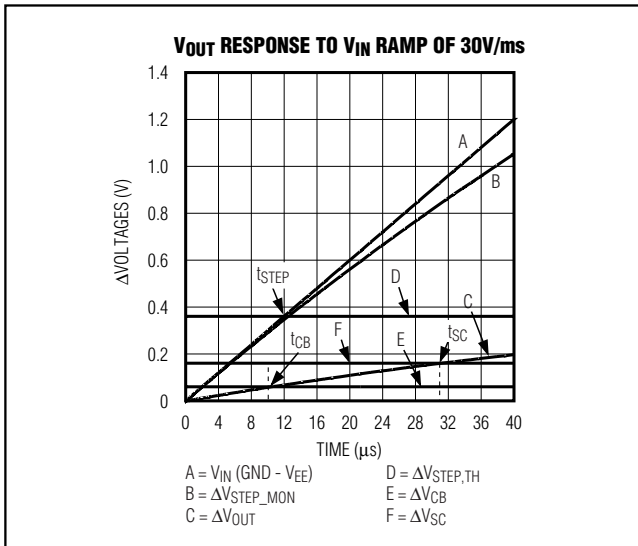


Figure 24.  $V_{OUT}$  Response to  $V_{IN}$  Ramp of 30V/ms

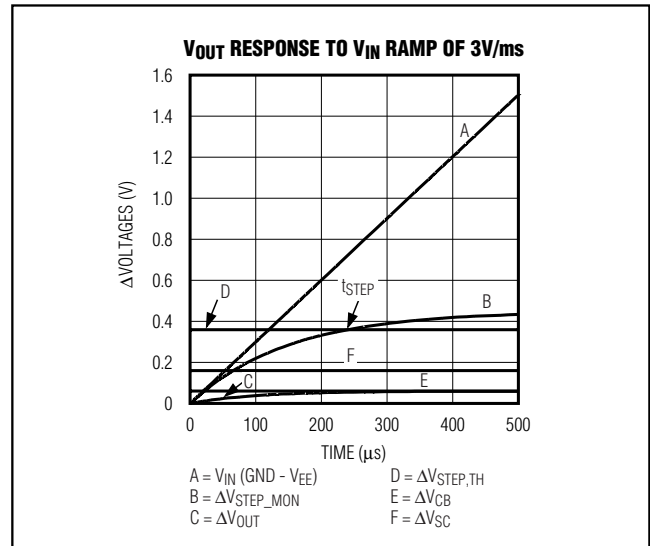


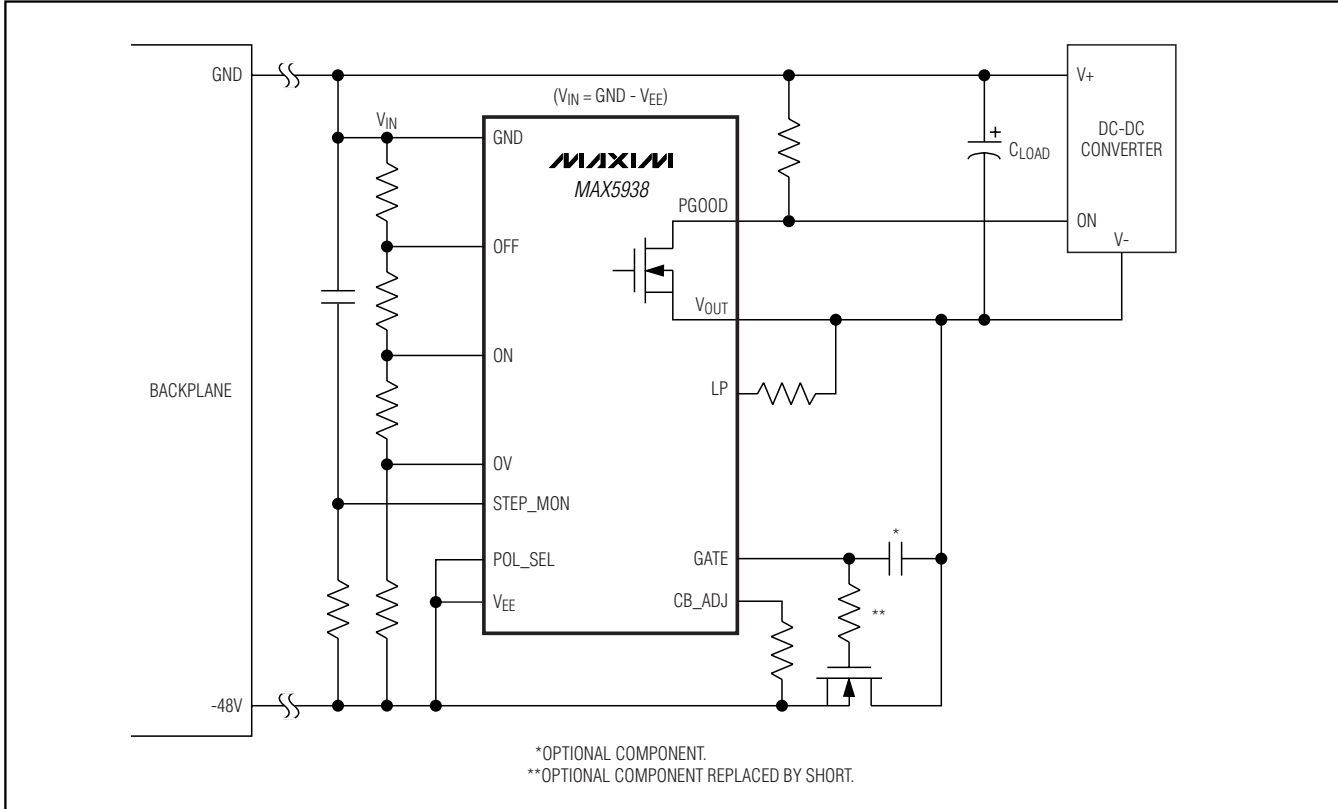
Figure 25.  $V_{OUT}$  Response to  $V_{IN}$  Ramp of 3V/ms



# -48V Hot-Swap Controller with $V_{IN}$ Step Immunity, No RSENSE, and Overvoltage Protection

## Typical Operating Circuit

MAX5938



## Timing Table

NAME	SYMBOL	TYPICAL TIME (s)
Power-Up Delay	$t_{\text{ONDLY}}$	220m
Load-Probe Test Timeout	$t_{\text{LP}}$	220m
Load-Probe Retry Time	$t_{\text{LP\_OFF}}$	3.5
PGOOD Assertion Delay Time	$t_{\text{PGOOD}}$	1.26m
Autoretry Delay	$t_{\text{RETRY}}$	3.5
Circuit-Breaker Glitch Rejection	$t_{\text{CB\_DLY}}$	1.4m
ON Glitch Rejection	$t_{\text{REJ}}$	1.5m
OV Transient Rejection	$t_{\text{OVREJ}}$	1.5m
GATE Pulldown Pulse Following a $V_{IN}$ Step	—	350 $\mu$
GATE Low after a $V_{IN}$ Step, Prior to Ramp	—	1m

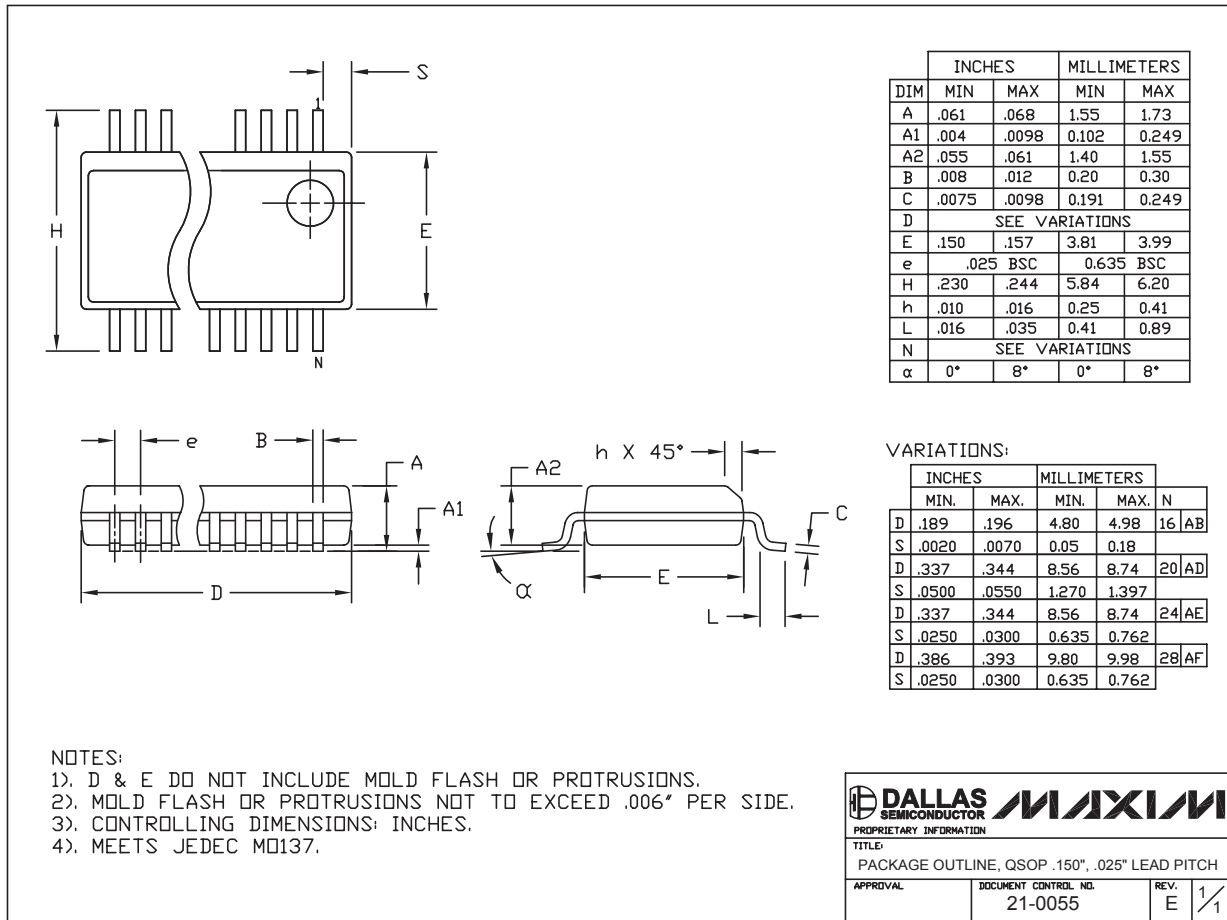
## Chip Information

TRANSISTOR COUNT: 2320  
PROCESS: BiCMOS

# **-48V Hot-Swap Controller with VIN Step Immunity, No RSENSE, and Overvoltage Protection**

## **Package Information**

(The package drawing(s) in this data sheet may not reflect the most current specifications. For the latest package outline information, go to [www.maxim-ic.com/packages](http://www.maxim-ic.com/packages).)



QSOP-EPS

**DALLAS SEMICONDUCTOR** **MAXIM**

PROPRIETARY INFORMATION

TITLE:  
PACKAGE OUTLINE, QSOP .150", .025" LEAD PITCH

APPROVAL	DOCUMENT CONTROL NO. 21-0055	REV. E	1/1
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