

## Description

The CS51033 is available in 8L SO and 8L PDIP plastic packages.

## Features

- 1A Totem Pole Output Driver
- High Speed Oscillator (700kHz max)
- No Stability Compensation Required
- Lossless Short Circuit Protection
- 2% Precision Reference
- Programmable Soft Start

**NOTE:** Capacitors C<sub>2</sub>, C<sub>3</sub> and C<sub>4</sub> are low ESR tantalum caps used for noise reduction.



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## Absolute Maximum Ratings

Power Supply Voltage, $V_{CC}$ .....	5V
Driver Supply Voltage, $V_C$ .....	20V
Driver Output Voltage, $V_{GATE}$ .....	20V
$C_{OSC}$ , CS, $V_{FB}$ (Logic Pins) .....	5V
Peak Output Current .....	1.0A
Steady State Output Current .....	200mA
Operating Junction Temperature, $T_J$ .....	150°C
Storage Temperature Range, $T_S$ .....	-65 to 150°C
ESD (Human Body Model).....	2kV
Lead Temperature Soldering	
Wave Solder (through hole styles only) .....	10 sec. max, 260°C peak
Reflow (SMD styles only) .....	60 sec. max above 183°C, 230°C peak

**Electrical Characteristics: Specifications apply for  $3.135 \leq V_{CC} \leq 3.465V$ ,  $3V \leq V_C \leq 16V$ ,  
 $-40^\circ C \leq T_A \leq 125^\circ C$ ,  $-40^\circ C \leq T_J \leq 125^\circ C$ , unless otherwise specified.**

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT
<b>■ Oscillator</b>					
	$V_{FB} = 1.2V$				
Frequency	$C_{OSC} = 470pF$	160	200	240	kHz
Charge Current	$1.4V < V_{COSC} < 2V$		110		$\mu A$
Discharge Current	$2.7V > V_{COSC} > 2V$		660		$\mu A$
Maximum Duty Cycle	$1 - (t_{OFF} / t_{ON})$	80.0	83.3		%
<b>■ Short Circuit Timer</b>					
	$V_{FB} = 1.0V$ ; $CS = 0.1\mu F$ ; $V_{COSC} = 2V$				
Charge Current	$1V < V_{CS} < 2V$	175	264	325	$\mu A$
Fast Discharge Current	$2.55V > V_{CS} > 2.4V$	40	66	80	$\mu A$
Slow Discharge Current	$2.4V > V_{CS} > 1.5V$	4	6	10	$\mu A$
Start Fault Inhibit Time		0.70	0.85	1.40	ms
Valid Fault Time	$2.6V > V_{CS} > 2.4V$	0.2	0.3	0.45	ms
GATE Inhibit Time	$2.4V > V_{CS} > 1.5V$	9	15	23	ms
Duty Cycle		2.5	3.1	4.6	%
<b>■ CS Comparator</b>					
	$V_{FB} = 1V$				
Fault Enable CS Voltage			2.5		V
Max. CS Voltage	$V_{FB} = 1.5V$		2.6		V
Fault Detect Voltage	$V_{CS}$ when GATE goes high		2.4		V
Fault Inhibit Voltage	Minimum $V_{CS}$		1.5		V
Hold Off Release Voltage	$V_{FB} = 0V$	0.4	0.7	1.0	V
Regulator Threshold Voltage Clamp	$V_{CS} = 1.5V$	0.725	0.866	1.035	V
<b>■ <math>V_{FB}</math> Comparator</b>					
	$V_{COSC} = V_{CS} = 2V$				
Regulator Threshold Voltage	$T_J = 25^\circ C$ (Note 1)	1.225	1.250	1.275	V
	$T_J = -40$ to $125^\circ C$	1.210	1.250	1.290	V
Fault Threshold Voltage	$T_J = 25^\circ C$ (Note 1)	1.12	1.15	1.17	V
	$T_J = -40$ to $125^\circ C$	1.10	1.15	1.19	V
Threshold Line Regulation	$3.135V \leq V_{CC} \leq 3.465$		6	15	mV
Input Bias Current	$V_{FB} = 0V$		1	4	$\mu A$
Voltage Tracking	(Regulator Threshold Voltage - Fault Threshold Voltage)	70	100	120	mV
Input Hysteresis Voltage			4	20	mV

Electrical Characteristics: Specifications apply for  $3.135 \leq V_{CC} \leq 3.465V$ ,  $3V \leq V_C \leq 16V$ ,  
 $-40^{\circ}C \leq T_J \leq 125^{\circ}C$ , unless otherwise specified.

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT
<b>■ Power Stage</b>					
	$V_C = 10V$ ; $V_{FB} = 1.2V$				
GATE DC Low Saturation Voltage	$V_{COSC} = 1V$ ; 200mA Sink		1.2	1.5	V
GATE DC High Saturation Voltage	$V_{COSC} = 2.7V$ ; 200mA Source; $V_C = V_{GATE}$		1.5	2.1	V
Rise Time	$C_{GATE} = 1nF$ ; $1.5V < V_{GATE} < 9V$		25	60	ns
Fall Time	$C_{GATE} = 1nF$ ; $9V > V_{GATE} > 1.5V$		25	60	ns
<b>■ Current Drain</b>					
$I_{CC}$	$3.135V < V_{CC} < 3.465V$ , Gate switching		3.5	6.0	mA
$I_C$	$3V < V_C < 16V$ , Gate non-switching		2.7	4.0	mA

Note1: Guaranteed by design not 100% tested in production.

Package Pin Description		
PACKAGE PIN #	PIN SYMBOL	FUNCTION
<b>8L SO Narrow &amp; PDIP</b>		
1	$V_{GATE}$	Driver pin to gate of external PFET.
2	PGnd	Output power stage ground connection.
3	$C_{OSC}$	Oscillator frequency programming capacitor.
4	Gnd	Logic ground.
5	$V_{FB}$	Feedback voltage input.
6	$V_{CC}$	Logic supply voltage.
7	CS	Soft start and fault timing capacitor.
8	$V_C$	Driver supply voltage.

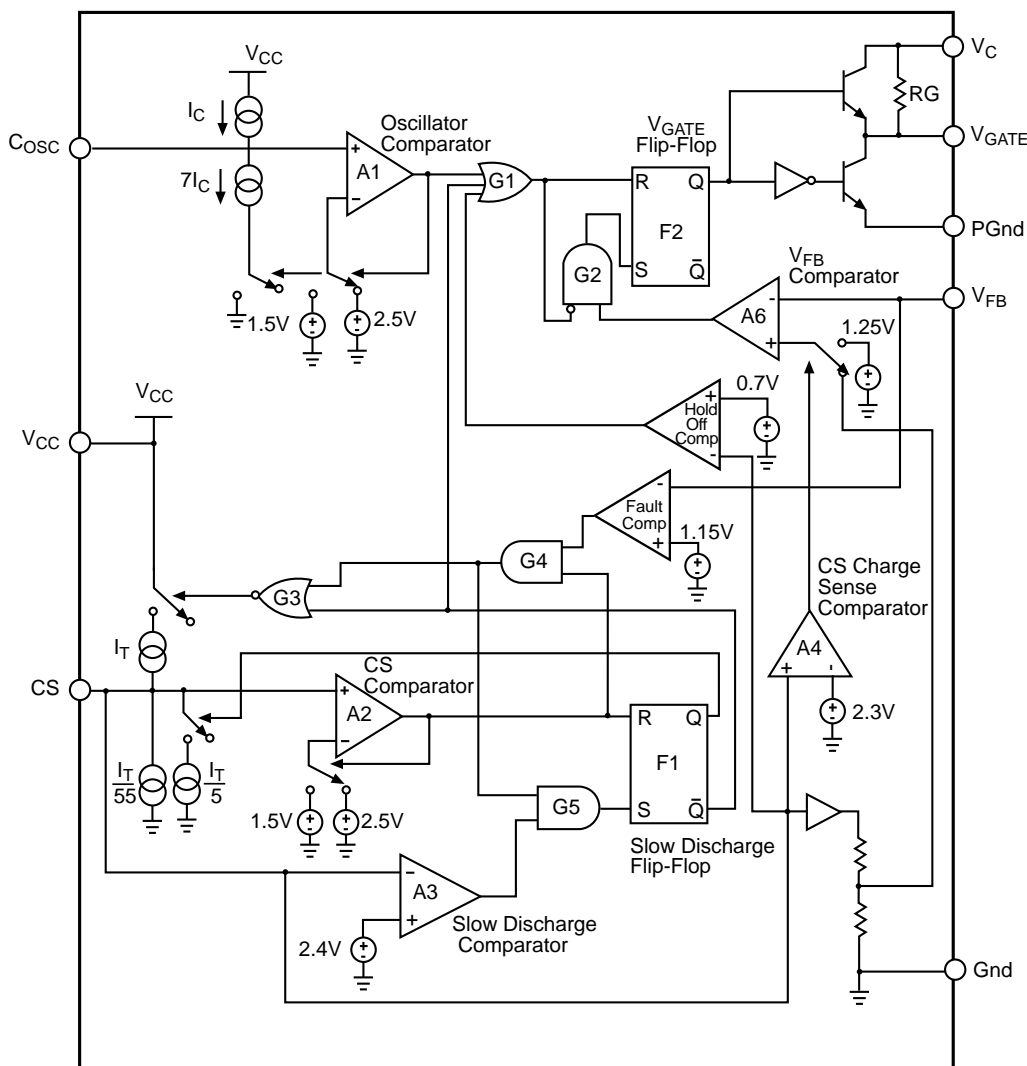


Figure 1: Block Diagram for CS51033

## Circuit Description

### Theory of Operation

#### Control Scheme

The CS51033 monitors the output voltage to determine when to turn on the PFET. If  $V_{FB}$  falls below the internal reference voltage of  $1.25V$  during the oscillator's charge cycle, the PFET is turned on and remains on for the duration of the charge time. The PFET gets turned off and remains off during the oscillator's discharge cycle time with the maximum duty cycle to 80%. It requires  $7mV$  typical, and  $20mV$  maximum ripple on the  $V_{FB}$  pin is required to operate. This method of control does not require any loop stability compensation.

#### Startup

The CS51033 has an externally programmable soft start feature that allows the output voltage to come up slowly, preventing voltage overshoot on the output.

At startup, the voltage on all pins is zero. As  $V_{CC}$  rises, the  $V_C$  voltage along with the internal resistor  $R_G$  keeps the PFET off. As  $V_{CC}$  and  $V_C$  continue to rise, the oscillator capacitor ( $C_{OSC}$ ) and the Soft start/Fault Timing capacitor ( $CS$ ) charges via internal current sources.  $C_{OSC}$  gets charged by the current source  $I_C$  and  $CS$  gets charged by the  $I_T$  source combination described by:

$$I_{CS} = I_T - \left( \frac{I_T}{55} + \frac{I_T}{5} \right)$$

The internal Holdoff Comparator ensures that the external PFET is off until  $V_{CS} > 0.7V$  preventing the GATE flip-flop (F2) from being set. This allows the oscillator to reach its operating frequency before enabling the drive output. Soft start is obtained by clamping the  $V_{FB}$  comparator's (A6) reference input to approximately  $1/2$  of the voltage at the  $CS$  pin during startup, permitting the control loop and the output voltage to slowly increase. Once the  $CS$  pin charges above the Holdoff Comparator trip point of  $0.7V$ , the low

feedback to the  $V_{FB}$  Comparator sets the GATE flip-flop during  $C_{OSC}$ 's charge cycle. Once the GATE flip-flop is set,  $V_{GATE}$  goes low and turns on the PFET. When  $V_{CS}$  exceeds 2.4V, the CS charge sense comparator (A4) sets the  $V_{FB}$  comparator reference to 1.25V completing the startup cycle.

### Lossless Short Circuit Protection

The CS51033 has "Lossless" short circuit protection since there is no current sense resistor required. When the voltage at the CS pin (the fault timing capacitor voltage) reaches 2.5V, the fault timing circuitry is enabled. During normal operation the CS voltage is 2.6V. During a short circuit condition or a transient condition, the output voltage moves lower and the voltage at  $V_{FB}$  drops. If  $V_{FB}$  drops below 1.15V, the output of the fault comparator goes high and the CS51033 goes into a fast discharge mode. The fault timing capacitor, CS, discharges to 2.4V. If the  $V_{FB}$  voltage is still below 1.15V when the CS pin reaches 2.4V, a valid fault condition has been detected. The slow discharge comparator output goes high and enables gate G5 which sets the slow discharge flip flop. The  $V_{gate}$  flip flop resets and the output switch is turned off. The fault timing capacitor is slowly discharged to 1.5V. The CS51033 then enters a normal startup routine. If the fault is still present when the fault timing capacitor voltage reaches 2.5V, the fast and slow discharge cycles repeat as shown in figure 2.

If the  $V_{FB}$  voltage is above 1.15V when CS reaches 2.4V a fault condition is not detected, normal operation resumes and CS charges back to 2.6V. This reduces the chance of erroneously detecting a load transient as a fault condition.

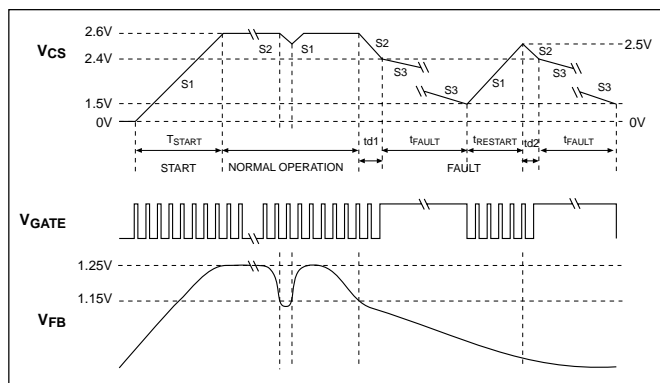


Figure 2. Voltage on start capacitor ( $V_{CS}$ ), the gate ( $V_{GATE}$ ), and in the feedback loop ( $V_{FB}$ ), during startup, normal and fault conditions.

### Buck Regulator Operation

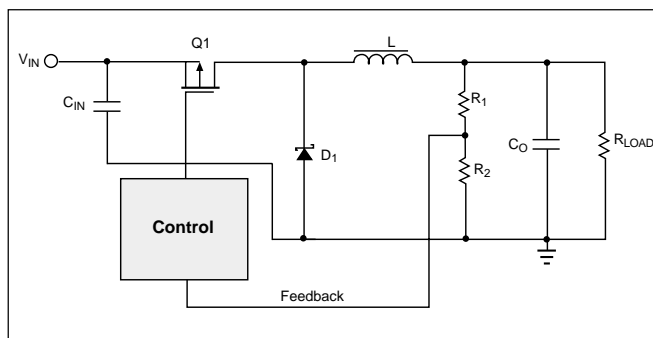


Figure 3. Buck regulator block diagram.

A block diagram of a typical buck regulator is shown in Figure 3. If we assume that the output transistor is initially off, and the system is in discontinuous operation, the inductor current  $I_L$  is zero and the output voltage is at its nominal value. The current drawn by the load is supplied by the output capacitor  $C_O$ . When the voltage across  $C_O$  drops below the threshold established by the feedback resistors  $R_1$  and  $R_2$  and the reference voltage  $V_{REF}$ , the power transistor  $Q_1$  switches on and current flows through the inductor to the output. The inductor current rises at a rate determined by  $(V_{IN} - V_{OUT}) / \text{Load}$ . The duty cycle (or "on" time) for the CS51033 is limited to 80%. If the output voltage remains higher than nominal during the entire  $C_{OSC}$  charge time, the  $Q_1$  does not turn on, skipping the pulse.

### CHARGE PUMP CIRCUIT

(Refer to the CS51033 Application Diagram)

An external charge pump circuit is necessary when the input voltage is below 5V to ensure that there is sufficient gate drive voltage for the external FET. When  $V_{IN}$  is applied, capacitors  $C_1$  and  $C_2$  will be charged to a diodes drop below  $V_{IN}$  via diodes  $D_2$  and  $D_4$ , respectively. When the PFET turns on, its drain voltage will be approximately equal to  $V_{IN}$ . Since the voltage across  $C_1$  can not change instantaneously,  $D_2$  is reverse biased and the anode voltage rises to approximately  $2 \times 3.3V - V_{D2}$ .  $C_1$  transfers some of its stored charge  $C_2$  via  $D_3$ . After several cycles there is sufficient gate drive voltage.

## Applications Information

### Designing a Power Supply with the CS51033

#### Specifications

$V_{IN} = 3.3V \pm 10\%$  (i.e. 3.63V max., 2.97V min.)

$V_{OUT} = 1.5V \pm 2\%$

$I_{OUT} = 0.3A$  to  $3A$

Output ripple voltage  $< 33mV$ .

$F_{SW} = 200kHz$ .

#### 1) Duty Cycle Estimates

Since the maximum duty cycle,  $D$ , of the CS51033 is limited

to 80% min. it is best to estimate the duty cycle for the various input conditions to see that the design will work over the complete operating range.

The duty cycle for a buck regulator operating in a continuous conduction mode is given by:

$$D = \frac{V_{OUT} + V_D}{V_{IN} - V_{SAT}}$$

Where  $V_{SAT}$  is  $R_{ds(on)} \times I_{OUT}$  Max.

In this case we can assume that  $V_D = 0.6V$  and  $V_{SAT} = 0.6V$  so the equation reduces to:

$$D = \frac{V_{OUT}}{V_{IN}}$$

From this, the maximum duty cycle  $D_{MAX}$  is 53%, this occurs when  $V_{IN}$  is at its minimum while the minimum duty cycle  $D_{MIN}$  is 0.35%.

## 2) Switching Frequency and on and off time calculations.

$F_{SW} = 200\text{KHz}$ . The switching frequency is determined by  $C_{OSC}$ , whose value is determined by :

$$C_{OSC} = \frac{95}{F_{SW} \times \left( 1 - \left( \frac{F_{SW}}{3 \times 10^6} \right) - \left( \frac{30 \times 10^3}{F_{SW}} \right)^2 \right)} \approx 470\text{pF}$$

$$T = \frac{1}{F_{SW}} = 5\mu\text{s}$$

$$T_{ON(MAX)} = 5\mu\text{s} \times 0.53 = 2.65\mu\text{s}$$

$$T_{ON(MIN)} = 5\mu\text{s} \times 0.35 = 1.75\mu\text{s}$$

$$T_{OFF(MAX)} = 5\mu\text{s} - 0.7\mu\text{s} = 4.3\mu\text{s}$$

## 3) Inductor selection

Pick the inductor value to maintain continuous mode operation down to 0.3 Amps.

The ripple current  $\Delta I = 2 \times I_{OUT(MIN)} = 2 \times 0.3\text{A} = 0.6\text{A}$ .

$$L_{MIN} = \frac{V_{OUT} + V_D \times T_{OFF(MAX)}}{\Delta I} = \frac{2.1\text{V} \times 4.3\mu\text{s}}{0.6\text{A}} \approx 15\mu\text{H}$$

The CS51033 will operate with almost any value of inductor. With larger inductors the ripple current is reduced and the regulator will remain in a continuous conduction mode for lower values of load current. A smaller inductor will result in larger ripple current. The core must not saturate with the maximum expected current, here given by:

$$I_{MAX} = \frac{I_{OUT} + \Delta I}{2} = 3\text{A} + 0.6\text{A}/2 = 3.3\text{A}$$

## 4) Output Capacitor

The output capacitor limits the output ripple voltage. The CS51033 needs a maximum of 15mV of output ripple for the feedback comparator to change state. If we assume that all the inductor ripple current flows through the output capacitor and that it is an ideal capacitor (i.e. zero ESR), the minimum capacitance needed to limit the output ripple to 50mV peak to peak is given by:

$$C_O = \frac{\Delta I}{8 \times F_{SW} \times \Delta V} = \frac{0.6\text{A}}{8 \times (200 \times 10^3 \text{ Hz}) \times (33 \times 10^{-3} \text{ V})} \approx 11.4\mu\text{F}$$

The minimum ESR needed to limit the output voltage ripple

to 50mV peak to peak is:

$$ESR = \frac{\Delta V}{\Delta I} = \frac{50 \times 10^{-3}}{0.6\text{A}} = 55\text{m}\Omega$$

The output capacitor should be chosen so that its ESR is at least half of the calculated value and the capacitance is at least ten times the calculated value. It is often advisable to use several capacitors in parallel to reduce the ESR.

Low impedance aluminum electrolytic, tantalum or organic semiconductor capacitors are a good choice for an output capacitor. Low impedance aluminum are the cheapest but are not available in surface mount at present. Solid tantalum chip capacitors are available from a number of suppliers and offer the best choice for surface mount applications. The capacitor working voltage should be greater than the output voltage in all cases.

## 5) $V_{FB}$ Divider

$$V_{OUT} = 1.25\text{V} \left( \frac{R1 + R2}{R2} \right) = 1.25\text{V} \left( \frac{R1}{R2} + 1 \right)$$

The input bias current to the comparator is  $4\mu\text{A}$ . The resistor divider current should be considerably higher than this to ensure that there is sufficient bias current. If we choose the divider current to be at least 250 times the bias current this gives a divider current of 1mA and simplifies the calculations.

$$\frac{1.5\text{V}}{1\text{mA}} = R1 + R2 = 1.5\text{K}\Omega$$

Let  $R2 = 1\text{K}$

Rearranging the divider equation gives:

$$R1 = R2 \left( \frac{V_{OUT}}{1.25} - 1 \right) = 1\text{K}\Omega \left( \frac{1.5\text{V}}{1.25} - 1 \right) = 200\Omega$$

## 6) Divider bypass capacitor $C_{rr}$

Since the feedback resistors divide the output voltage by a factor of 4, i.e.  $5\text{V}/1.25\text{V} = 4$  it follows that the output ripple is also divided by four. This would require that the output ripple be at least 60mV ( $4 \times 15\text{mV}$ ) to trip the feedback comparator. We use a capacitor  $C_{rr}$  to act as an ac short so that the output ripple is not attenuated by the divider network. The ripple voltage frequency is equal to the switching frequency so we choose  $C_{rr}$  so that:

$$X_C = \frac{1}{2\pi f C}$$

is negligible at the switching frequency.

In this case  $F_{SW}$  is 200kHz if we allow  $X_C = 3\Omega$  then:

$$C = \frac{1}{2\pi f 3} \approx 0.265\mu\text{F}$$

## 7) Soft start and Fault timing capacitor $C_S$ .

$C_S$  performs several important functions. First it provides a dead time for load transients so that the IC does not enter a fault mode every time the load changes abruptly. Secondly it disables the fault circuitry during startup, it also provides soft start by clamping the reference voltage during startup

to rise slowly and finally it controls the Hiccup short circuit protection circuitry. This function reduces the PFET's duty cycle to 2% of the  $C_S$  period.

The most important consideration in calculating  $C_S$  is that it's voltage does not reach 2.5V (the voltage at which the fault detect circuitry is enabled) before  $V_{FB}$  reaches 1.15V otherwise the power supply will never start.

If the  $V_{FB}$  pin reaches 1.15V the fault timing comparator will discharge  $C_S$  and the supply will not start. For the  $V_{FB}$  voltage to reach 1.15V the output voltage must be at least  $4 \times 1.15 = 4.6V$ .

If we choose an arbitrary startup time of  $200\mu s$  we calculate the value of  $C_S$  from:

$$T = \frac{C_S \times 2.5V}{I_{CHARGE}}$$

$$C_{S(min)} = \frac{200\mu s \times 264\mu A}{2.5V} = 0.02\mu F$$

Use  $0.1\mu f$ .

The fault time out time is the sum of the slow discharge time the fast discharge time and the recharge time and is obviously dominated by the slow discharge time. The first parameter is the slow discharge time, it is the time for the  $C_S$  capacitor to discharge from 2.4V to 1.5V and is given by:

$$T_{SLOWDISCHARGE} = \frac{C_S \times (2.4V - 1.5V)}{I_{DISCHARGE}}$$

Where  $I_{DISCHARGE}$  is  $6\mu A$  typical.

$$T_{SLOWDISCHARGE} = C_S \times 1.5V \times 10^5$$

The fast discharge time occurs when a fault is first detected. The  $C_S$  capacitor is discharged from 2.5V to 2.4V.

$$T_{FASTDISCHARGE} = \frac{C_S \times (2.5V - 2.4V)}{I_{FASTDISCHARGE}}$$

Where  $I_{FASTDISCHARGE}$  is  $66\mu A$  typical.

$$T_{FASTDISCHARGE} = C_S \times 1515$$

The recharge time is the time for  $C_S$  to charge from 1.5V to 2.5V.

$$T_{CHARGE} = \frac{C_S \times (2.5V - 1.5V)}{I_{CHARGE}}$$

Where  $I_{CHARGE}$  is  $264\mu A$  typical.

$$T_{CHARGE} = C_S \times 3787$$

The fault time out time is given by:

$$T_{FAULT} = C_S \times (3787 + 1515 + 1.5 \times 10^5)$$

$$T_{FAULT} = C_S \times 1.55 \times 10^5$$

For this circuit

$$T_{FAULT} = 0.1 \times 10^{-6} \times 1.55 \times 10^5 = 0.0155$$

A larger value of  $C_S$  will increase the fault time out time but will also increase the soft start time.

### 8) Input Capacitor.

The input capacitor reduces the peak currents drawn from the input supply and reduces the noise and ripple voltage on the  $V_{CC}$  and  $V_C$  pins. This capacitor must also ensure that the  $V_{CC}$  remains above the UVLO voltage in the event of an output short circuit.  $C_{IN}$  should be a low ESR capacitor of at least  $100\mu f$ . A ceramic surface mount capacitor should also be connected between  $V_{CC}$  and ground to prevent spikes.

### 9) MOSFET Selection

The CS51033 drive a P-channel MOSFET. The  $V_{GATE}$  pin swings from Gnd to  $V_C$ . The type of PFET used depends on the operating conditions but for input voltages below 7V a logic level FET should be used.

Choose a PFET with a continuous drain current ( $I_D$ ) rating greater than the maximum output current.  $R_{DS(on)}$  should be less than

$$R_{DS} < = \frac{0.6V}{I_{OUT(max)}} 167m\Omega$$

The Gate-to-Source voltage  $V_{GS}$  and the Drain-to Source Breakdown Voltage should be chosen based on the input supply voltage.

The power dissipation due to the conduction losses is given by:

$$P_D = I_{OUT}^2 \times R_{DS(on)} \times D$$

The power dissipation due to the switching losses is given by:

$$P_D = 0.5 \times V_{IN} \times I_{OUT} \times (T_R^r + T_F) \times F_{SW}$$

Where  $t_r$  = Rise Time and  $t_f$  = Fall Time.

### 10) Diode Selection.

The flyback or catch diode should be a Schottky diode because of it's fast switching ability and low forward voltage drop. The current rating must be at least equal to the maximum output current. The breakdown voltage should be at least 20V for this 12V application.

The diode power dissipation is given by:

$$P_D = I_{OUT} \times V_D \times (1 - D_{MIN})$$

# Package Specification

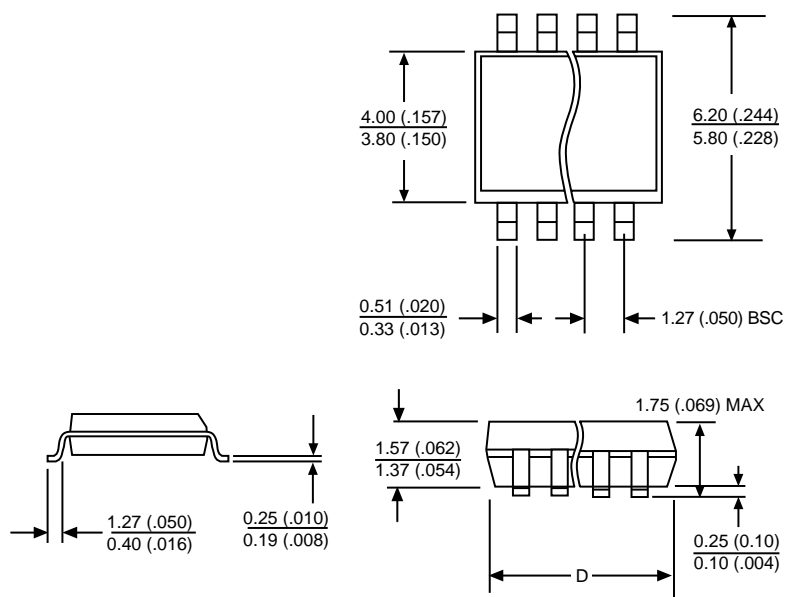
## PACKAGE DIMENSIONS IN mm (INCHES)

Lead Count	D			
	Metric		English	
	Max	Min	Max	Min
8L SO Narrow	5.00	4.80	.197	.189
8L PDIP	10.16	9.02	.400	.355

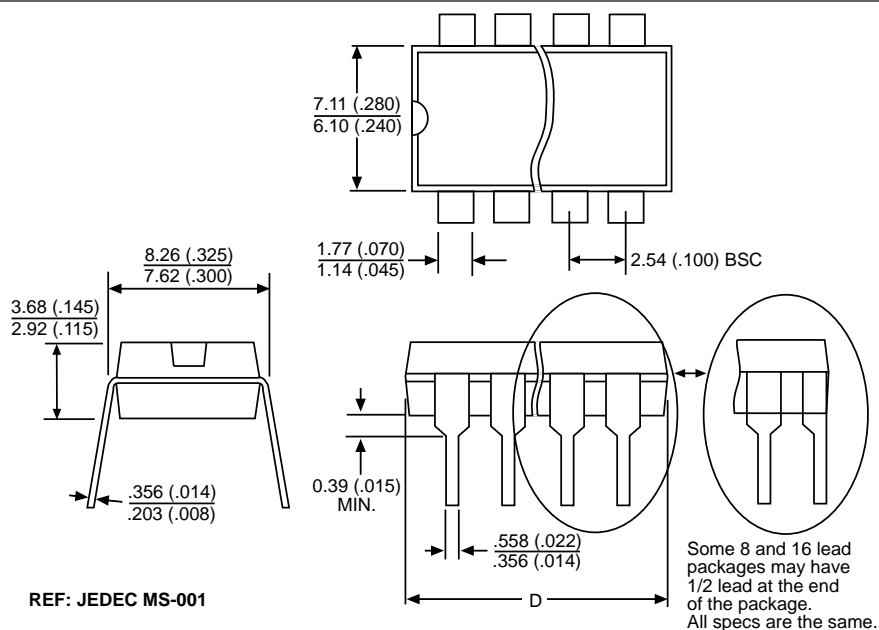
## PACKAGE THERMAL DATA

Thermal Data		8L SO Narrow	8L PDIP	
$R_{\theta JC}$	typ	45	52	$^{\circ}\text{C/W}$
$R_{\theta JA}$	typ	165	100	$^{\circ}\text{C/W}$

## Surface Mount Narrow Body (D); 150 mil wide



## Plastic DIP (N); 300 mil wide



## Ordering Information

Part Number	Description
CS51033YD8	8L SO Narrow
CS51033YDR8	8L SO Narrow ( <i>tape &amp; reel</i> )
CS51033YN8	8L PDIP

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