



MOTOROLA

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# MC34023 MC33023

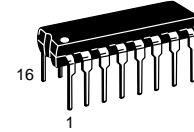
## High Speed Single-Ended PWM Controller

The MC34023 series are high speed, fixed frequency, single-ended pulse width modulator controllers optimized for high frequency operation. They are specifically designed for Off-Line and DC-to-DC converter applications offering the designer a cost-effective solution with minimal external components. These integrated circuits feature an oscillator, a temperature compensated reference, a wide bandwidth error amplifier, a high speed current sensing comparator, and a high current totem pole output ideally suited for driving a power MOSFET.

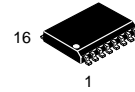
Also included are protective features consisting of input and reference undervoltage lockouts each with hysteresis, cycle-by-cycle current limiting, and a latch for single pulse metering.

The flexibility of this series allows it to be easily configured for either current mode or voltage mode control.

- 50 ns Propagation Delay to Output
- High Current Totem Pole Output
- Wide Bandwidth Error Amplifier
- Fully-Latched Logic with Double Pulse Suppression
- Latching PWM for Cycle-By-Cycle Current Limiting
- Soft-Start Control with Latched Overcurrent Reset
- Input Undervoltage Lockout with Hysteresis
- Low Start-Up Current (500  $\mu$ A Typ)
- Internally Trimmed Reference with Undervoltage Lockout
- 90% Maximum Duty Cycle (Externally Adjustable)
- Precision Trimmed Oscillator
- Voltage or Current Mode Operation to 1.0 MHz
- Functionally Similar to the UC3823

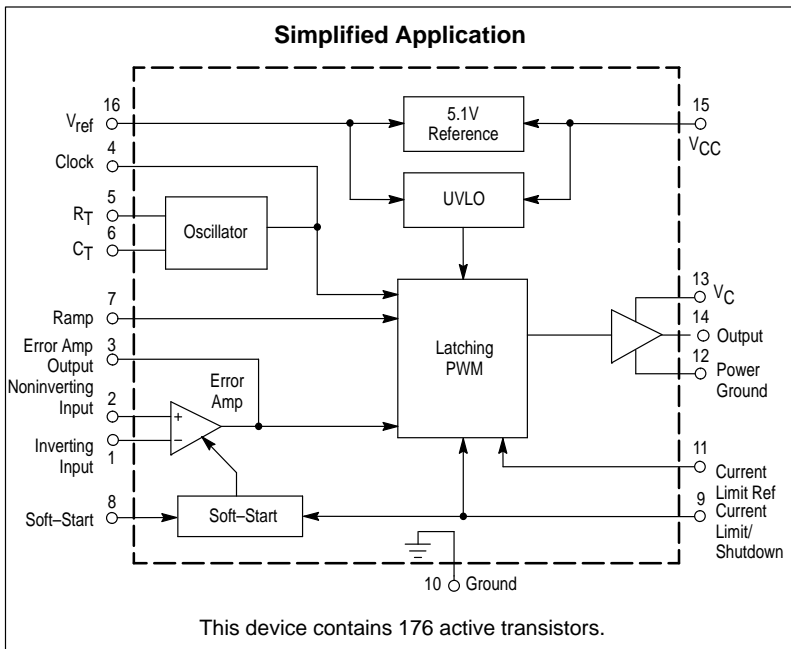
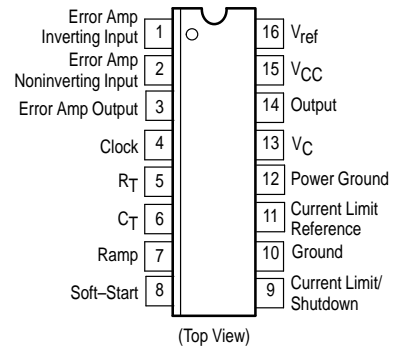


**P SUFFIX**  
PLASTIC PACKAGE  
CASE 648



**DW SUFFIX**  
PLASTIC PACKAGE  
CASE 751G  
(SO-16L)

### PIN CONNECTIONS



### ORDERING INFORMATION

Device	Operating Temperature Range	Package
MC33023P	$T_A = -40^\circ$ to $+105^\circ\text{C}$	Plastic DIP
MC33023DW		SO-16L
MC34023P	$T_A = 0^\circ$ to $+70^\circ\text{C}$	Plastic DIP



# MC34023 MC33023

**ELECTRICAL CHARACTERISTICS** ( $V_{CC} = 15\text{ V}$ ,  $R_T = 3.65\text{ k}\Omega$ ,  $C_T = 1.0\text{ nF}$ , for typical values  $T_A = +25^\circ\text{C}$ , for min/max values  $T_A$  is the operating ambient temperature range that applies [Note 2], unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
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## ERROR AMPLIFIER SECTION

Input Offset Voltage	$V_{IO}$	–	–	15	mV
Input Bias Current	$I_{IB}$	–	0.6	3.0	$\mu\text{A}$
Input Offset Current	$I_{IO}$	–	0.1	1.0	$\mu\text{A}$
Open-Loop Voltage Gain ( $V_O = 1.0\text{ V to }4.0\text{ V}$ )	$A_{VOL}$	60	95	–	dB
Gain Bandwidth Product ( $T_J = +25^\circ\text{C}$ )	GBW	4.0	8.3	–	MHz
Common Mode Rejection Ratio ( $V_{CM} = 1.5\text{ V to }5.5\text{ V}$ )	CMRR	75	95	–	dB
Power Supply Rejection Ratio ( $V_{CC} = 10\text{ V to }30\text{ V}$ )	PSRR	85	110	–	dB
Output Current, Source ( $V_O = 4.0\text{ V}$ ) Sink ( $V_O = 1.0\text{ V}$ )	$I_{Source}$ $I_{Sink}$	0.5 1.0	3.0 3.6	– –	mA
Output Voltage Swing, High State ( $I_O = -0.5\text{ mA}$ ) Low State ( $I_O = 1\text{ mA}$ )	$V_{OH}$ $V_{OL}$	4.5 0	4.75 0.4	5.0 1.0	V
Slew Rate	SR	6.0	12	–	V/ $\mu\text{s}$

## PWM COMPARATOR SECTION

Ramp Input Bias Current	$I_{IB}$	–	–0.5	–5.0	$\mu\text{A}$
Duty Cycle, Maximum Minimum	$DC_{(max)}$ $DC_{(min)}$	80 –	90 –	– 0	%
Zero Duty Cycle Threshold Voltage Pin 3(4) (Pin 7(9) = 0 V)	$V_{th}$	1.1	1.25	1.4	V
Propagation Delay (Ramp Input to Output, $T_J = +25^\circ\text{C}$ )	$t_{PLH(in/out)}$	–	60	100	ns

## SOFT-START SECTION

Charge Current ( $V_{Soft-Start} = 0.5\text{ V}$ )	$I_{chg}$	3.0	9.0	20	$\mu\text{A}$
Discharge Current ( $V_{Soft-Start} = 1.5\text{ V}$ )	$I_{dischg}$	1.0	4.0	–	mA

## CURRENT SENSE SECTION

Input Bias Current (Pin 9(12) = 0 V to 4.0 V)	$I_{IB}$	–	–	15	$\mu\text{A}$
Current Limit Comparator Input Offset Voltage (Pin 11(14) = 1.1 V)	$V_{IO}$	–	–	45	mV
Current Limit Reference Input Common Mode Range (Pin 11(14))	$V_{CMR}$	1.0	–	1.25	V
Shutdown Comparator Threshold	$V_{th}$	1.25	1.40	1.55	V
Propagation Delay (Current Limit/Shutdown to Output, $T_J = +25^\circ\text{C}$ )	$t_{PLH(in/out)}$	–	50	80	ns

## OUTPUT SECTION

Output Voltage Low State ( $I_{Sink} = 20\text{ mA}$ ) ( $I_{Sink} = 200\text{ mA}$ ) High State ( $I_{Source} = 20\text{ mA}$ ) ( $I_{Source} = 200\text{ mA}$ )	$V_{OL}$ $V_{OH}$	– – 13 12	0.25 1.2 13.5 13	0.4 2.2 – –	V
Output Voltage with UVLO Activated ( $V_{CC} = 6.0\text{ V}$ , $I_{Sink} = 0.5\text{ mA}$ )	$V_{OL(UVLO)}$	–	0.25	1.0	V
Output Leakage Current ( $V_C = 20\text{ V}$ )	$I_L$	–	100	500	$\mu\text{A}$
Output Voltage Rise Time ( $C_L = 1.0\text{ nF}$ , $T_J = +25^\circ\text{C}$ )	$t_r$	–	30	60	ns
Output Voltage Fall Time ( $C_L = 1.0\text{ nF}$ , $T_J = +25^\circ\text{C}$ )	$t_f$	–	30	60	ns

## UNDERVOLTAGE LOCKOUT SECTION

Start-Up Threshold ( $V_{CC}$ Increasing)	$V_{th(on)}$	8.8	9.2	9.6	V
UVLO Hysteresis Voltage ( $V_{CC}$ Decreasing After Turn-On)	$V_H$	0.4	0.8	1.2	V

## TOTAL DEVICE

Power Supply Current Start-Up ( $V_{CC} = 8.0\text{ V}$ ) Operating	$I_{CC}$	– –	0.5 20	1.2 30	mA
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**NOTES:** 1. Maximum package power dissipation limits must be observed.

2. Low duty cycle pulse techniques are used during test to maintain junction temperature as close to ambient as possible.

$T_{low} = 0^\circ\text{C}$  for MC34023       $T_{high} = +70^\circ\text{C}$  for MC34023  
 $= -40^\circ\text{C}$  for MC33023       $= +105^\circ\text{C}$  for MC33023

Figure 1. Timing Resistor versus Oscillator Frequency

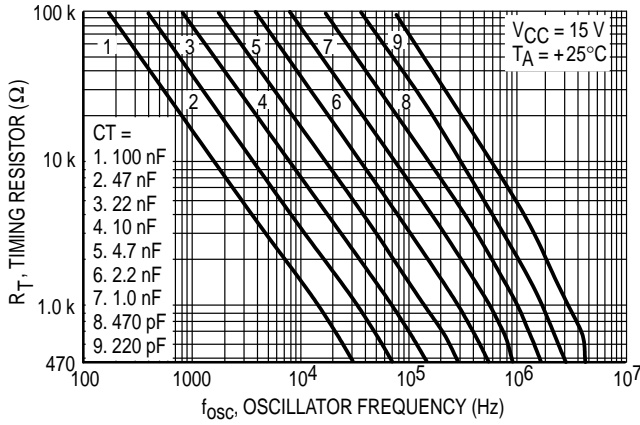


Figure 2. Oscillator Frequency versus Temperature

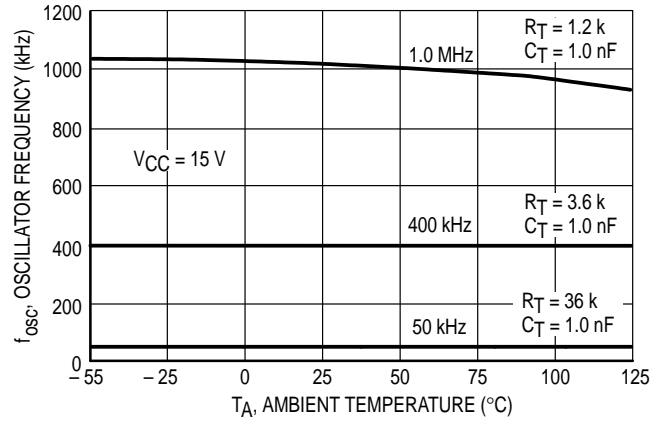


Figure 3. Error Amp Open Loop Gain and Phase versus Frequency

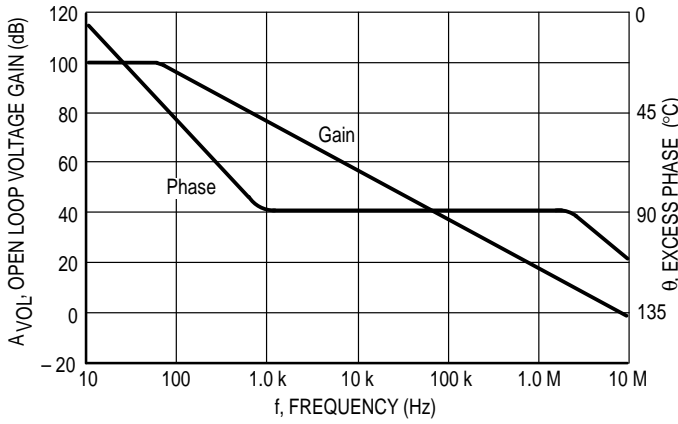


Figure 4. PWM Comparator Zero Duty Cycle Threshold Voltage versus Temperature

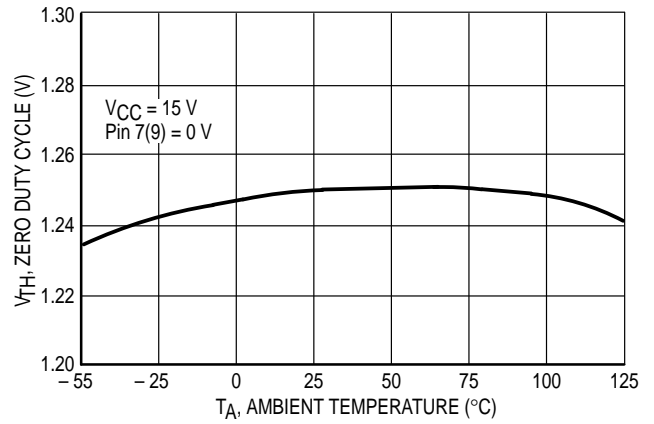


Figure 5. Error Amp Small Signal Transient Response

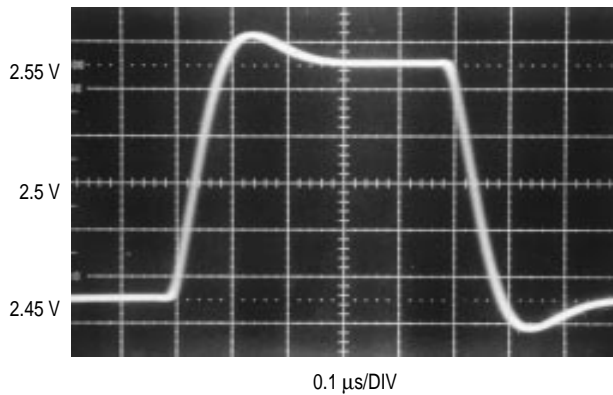


Figure 6. Error Amp Large Signal Transient Response

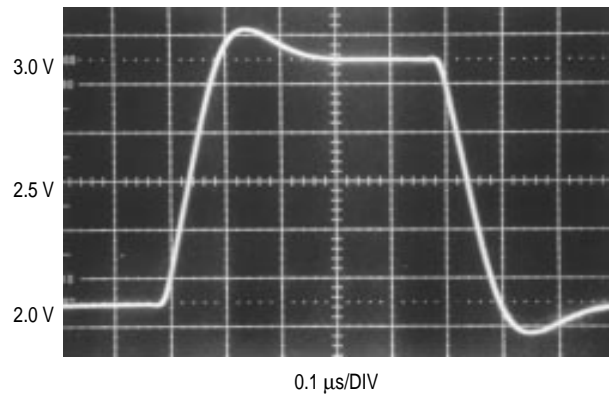


Figure 7. Reference Voltage Change versus Source Current

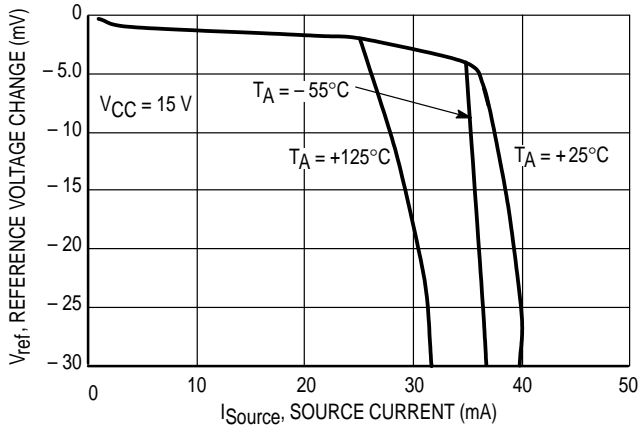


Figure 8. Reference Short Circuit Current versus Temperature

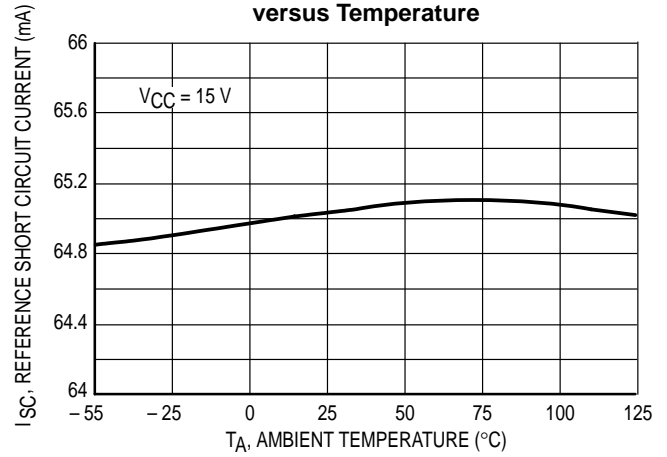
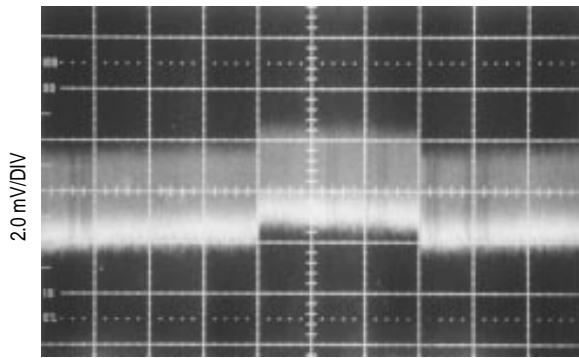
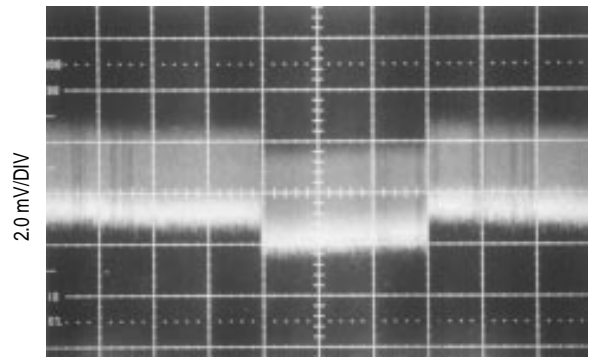


Figure 9. Reference Line Regulation



$V_{ref}$  LINE REGULATION 10 V to 24 V  
(2.0 ms/DIV)

Figure 10. Reference Load Regulation



$V_{ref}$  LOAD REGULATION 1.0 mA to 10 mA  
(2.0 ms/DIV)

Figure 11. Current Limit Comparator Input Offset Voltage versus Temperature

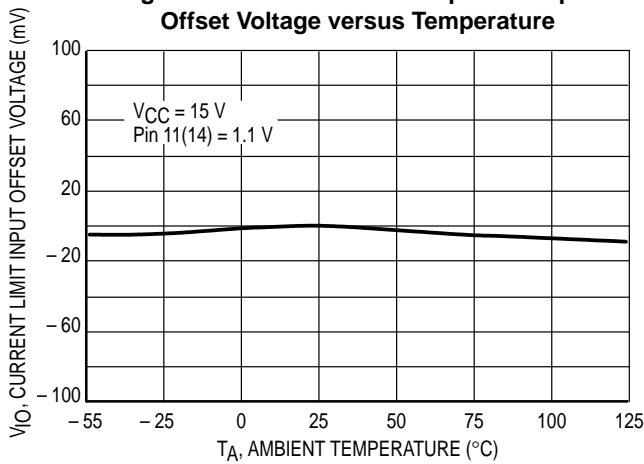


Figure 12. Shutdown Comparator Threshold Voltage versus Temperature

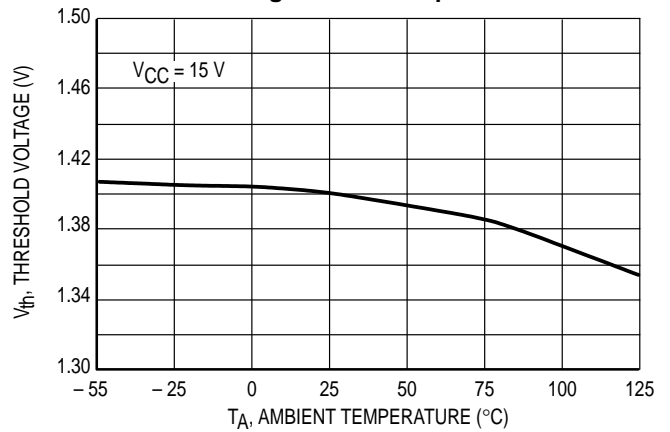


Figure 13. Soft-Start Charge Current versus Temperature

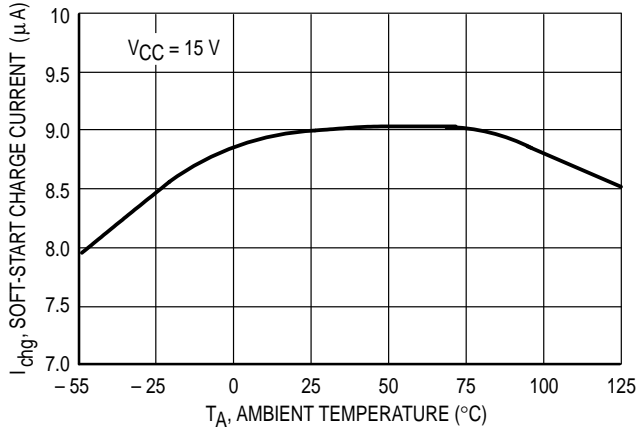


Figure 14. Output Saturation Voltage versus Load Current

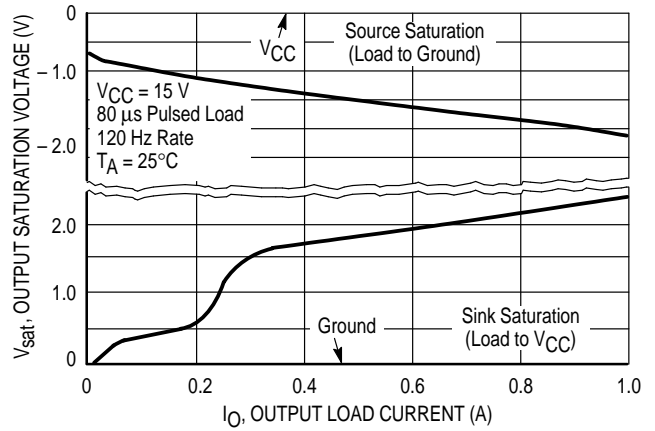
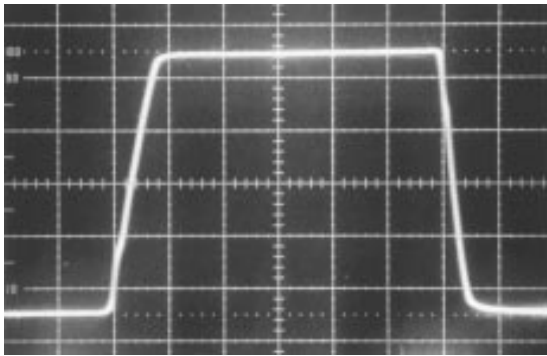
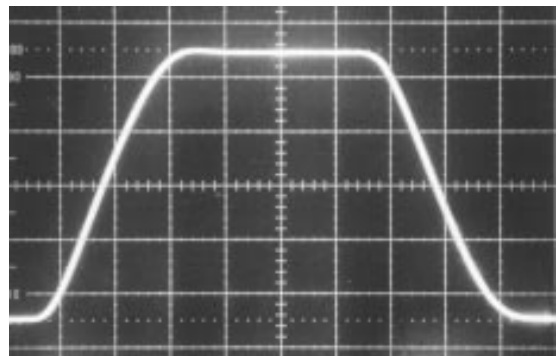


Figure 15. Drive Output Rise and Fall Time



OUTPUT RISE & FALL TIME 1.0 nF LOAD  
50 ns/DIV

Figure 16. Drive Output Rise and Fall Time



OUTPUT RISE & FALL TIME 10 nF LOAD  
50 ns/DIV

Figure 17. Supply Voltage versus Supply Current

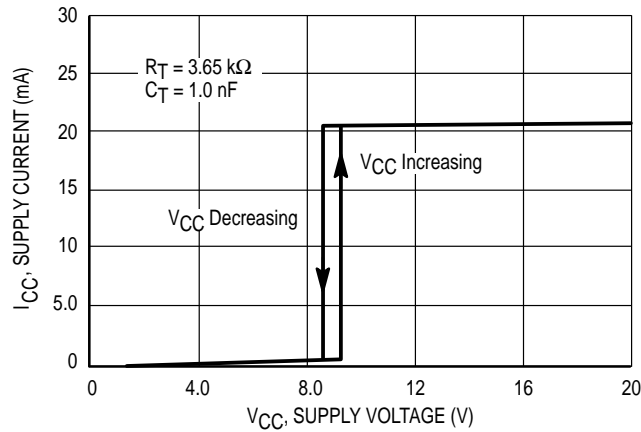


Figure 18. Representative Block Diagram

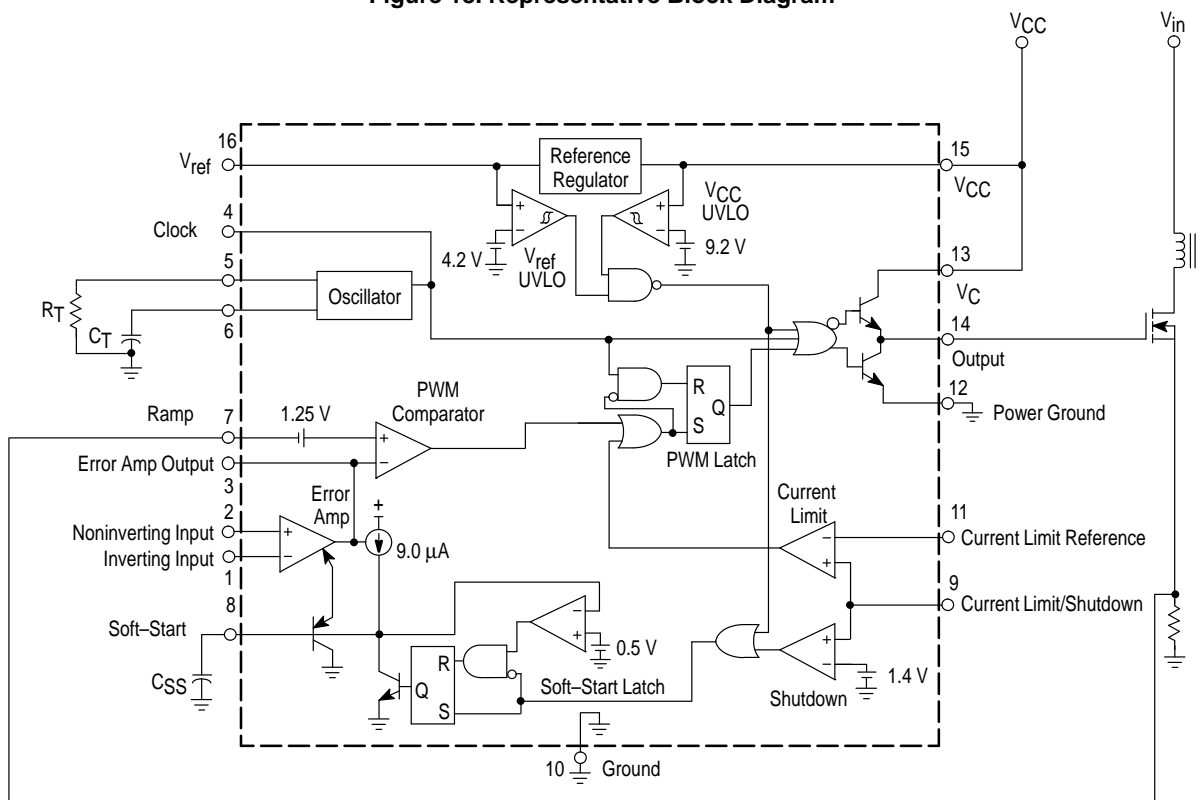
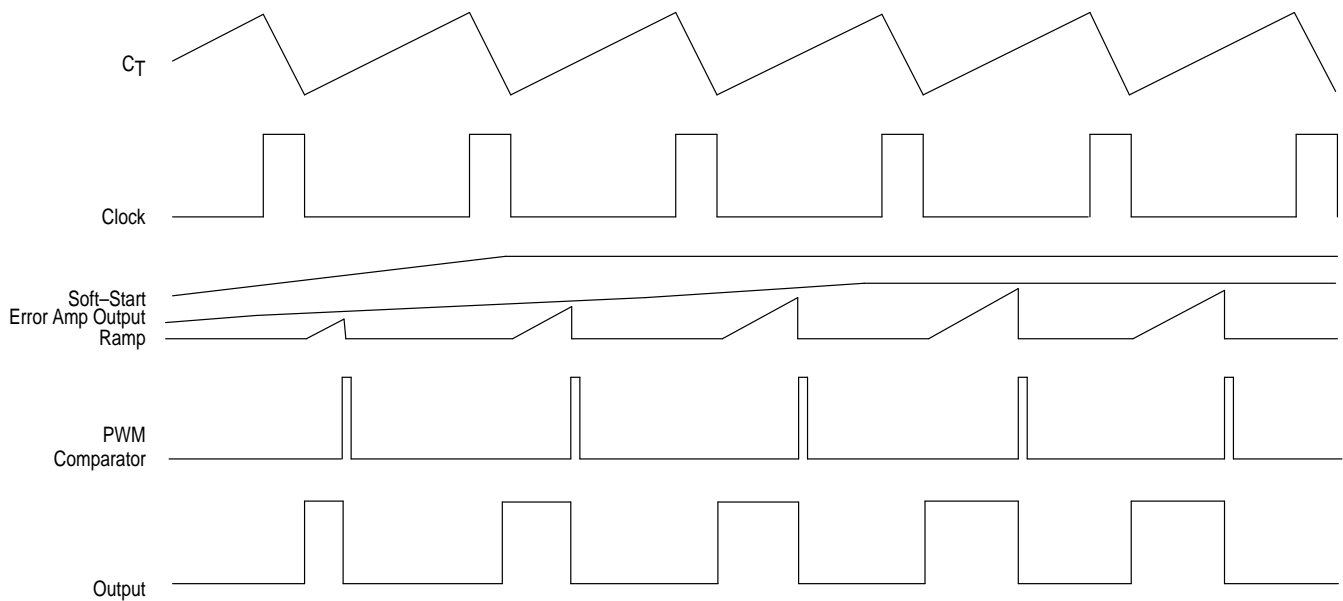


Figure 19. Current Limit Operating Waveforms



## OPERATING DESCRIPTION

The MC33023 and MC34023 series are high speed, fixed frequency, single-ended pulse width modulator controllers optimized for high frequency operation. They are specifically designed for Off-Line and DC-to-DC converter applications offering the designer a cost effective solution with minimal external components. A representative block diagram is shown in Figure 18.

**Oscillator**

The oscillator frequency is programmed by the values selected for the timing components  $R_T$  and  $C_T$ . The  $R_T$  pin is set to a temperature compensated 3.0 V. By selecting the value of  $R_T$ , the charge current is set through a current mirror for the timing capacitor  $C_T$ . This charge current runs continuously through  $C_T$ . The discharge current is ratioed to be 10 times the charge current, which yields the maximum duty cycle of 90%.  $C_T$  is charged to 2.8 V and discharged to 1.0 V. During the discharge of  $C_T$ , the oscillator generates an internal blanking pulse that resets the PWM Latch and, inhibits the outputs. The threshold voltage on the oscillator comparator is trimmed to guarantee an oscillator accuracy of 5.0% at 25°C.

Additional dead time can be added by externally increasing the charge current to  $C_T$  as shown in Figure 23. This changes the charge to discharge ratio of  $C_T$  which is set internally to  $I_{charge}/10 I_{charge}$ . The new charge to discharge ratio will be:

$$\% \text{ Deadtime} = \frac{I_{\text{additional}} + I_{\text{charge}}}{10 (I_{\text{charge}})}$$

A bidirectional clock pin is provided for synchronization or for master/slave operation. As a master, the clock pin provides a positive output pulse during the discharge of  $C_T$ . As a slave, the clock pin is an input that resets the PWM latch and blanks the drive output, but does not discharge  $C_T$ . Therefore, the oscillator is not synchronized by driving the clock pin alone. Figures 27, 28 and 29 provide suggested synchronization.

**Error Amplifier**

A fully compensated Error Amplifier is provided. It features a typical DC voltage gain of 95 dB and a gain bandwidth product of 8.3 MHz with 75 degrees of phase margin (Figure 3). Typical application circuits will have the noninverting input tied to the reference. The inverting input will typically be connected to a feedback voltage generated from the output of the switching power supply. Both inputs have a common mode voltage ( $V_{CM}$ ) input range of 1.5 V to 5.5 V. The Error Amplifier Output is provided for external loop compensation.

**Soft-Start Latch**

Soft-Start is accomplished in conjunction with an external capacitor. The Soft-Start capacitor is charged by an internal 9.0  $\mu\text{A}$  current source. This capacitor clamps the output of the error amplifier to less than its normal output voltage, thus

limiting the duty cycle. The time it takes for a capacitor to reach full charge is given by:

$$t \approx (4.5 \cdot 10^5) C_{\text{Soft-Start}}$$

A Soft-Start latch is incorporated to prevent erratic operation of this circuitry. Two conditions can cause the Soft-Start circuit to latch so that the Soft-Start capacitor stays discharged. The first condition is activation of an undervoltage lockout of either  $V_{CC}$  or  $V_{ref}$ . The second condition is when current sense input exceeds 1.4 V. Since this latch is "set dominant", it cannot be reset until either of these signals is removed and, the voltage at  $C_{\text{Soft-Start}}$  is less than 0.5 V.

**PWM Comparator and Latch**

A PWM circuit typically compares an error voltage with a ramp signal. The outcome of this comparison determines the state of the output. In voltage mode operation the ramp signal is the voltage ramp of the timing capacitor. In current mode operation the ramp signal is the voltage ramp induced in a current sensing element. The ramp input of the PWM comparator is pinned out so that the user can decide which mode of operation best suits the application requirements. The ramp input has a 1.25 V offset such that whenever the voltage at this pin exceeds the error amplifier output voltage minus 1.25 V, the PWM comparator will cause the PWM latch to set, disabling the outputs. Once the PWM latch is set, only a blanking pulse by the oscillator can reset it, thus initiating the next cycle.

**Current Limiting and Shutdown**

A pin is provided to perform current limiting and shutdown operations. Two comparators are connected to the input of this pin. The reference voltage for the current limit comparator is not set internally. A pin is provided so the user can set the voltage. When the voltage at the current limit input pin exceeds the externally set voltage, the PWM latch is set, disabling the output. In this way cycle-by-cycle current limiting is accomplished. If a current limit resistor is used in series with the power devices, the value of the resistor is found by:

$$R_{\text{Sense}} = \frac{I_{\text{Limit Reference Voltage}}}{I_{\text{pk (switch)}}$$

If the voltage at this pin exceeds 1.4 V, the second comparator is activated. This comparator sets a latch which, in turn, causes the soft start capacitor to be discharged. In this way a "hiccup" mode of recovery is possible in the case of output short circuits. If a current limit resistor is used in series with the output devices, the peak current at which the controller will enter a "hiccup" mode is given by:

$$I_{\text{shutdown}} = \frac{1.4 \text{ V}}{R_{\text{Sense}}}$$



**Undervoltage Lockout**

There are two undervoltage lockout circuits within the IC. The first senses  $V_{CC}$  and the second  $V_{ref}$ . During power-up,  $V_{CC}$  must exceed 9.2 V and  $V_{ref}$  must exceed 4.2 V before the outputs can be enabled and the Soft-Start latch released. If  $V_{CC}$  falls below 8.4 V or  $V_{ref}$  falls below 3.6 V, the outputs are disabled and the Soft-Start latch is activated. When the UVLO is active, the part is in a low current standby mode allowing the IC to have an off-line bootstrap start-up circuit. Typical start-up current is 500  $\mu$ A.

**Output**

The MC34023 has a high current totem pole output specifically designed for direct drive of power MOSFETs. It is capable of up to  $\pm 2.0$  A peak drive current with a typical rise and fall time of 30 ns driving a 1.0 nF load.

Separate pins for  $V_C$  and Power Ground are provided. With proper implementation, a significant reduction of switching transient noise imposed on the control circuitry is possible. The separate  $V_C$  supply input also allows the designer added flexibility in tailoring the drive voltage independent of  $V_{CC}$ .

**Reference**

A 5.1 V bandgap reference is pinned out and is trimmed to an initial accuracy of  $\pm 1.0\%$  at 25°C. This reference has short circuit protection and can source in excess of 10 mA for powering additional control system circuitry.

**Design Considerations**

**Do not attempt to construct the converter on wire-wrap or plug-in prototype boards.** With high frequency, high power, switching power supplies it is imperative to have separate current loops for the signal paths and for the power paths. The printed circuit layout should contain a ground plane with low current signal and high current switch and output grounds returning on separate paths back to the input filter capacitor. Shown in Figure 35 is a printed circuit layout of the application circuit. Note how the power and ground traces are run. All bypass capacitors and snubbers should be connected as close as possible to the specific part in question. The PC board lead lengths must be less than 0.5 inches for effective bypassing for snubbing.

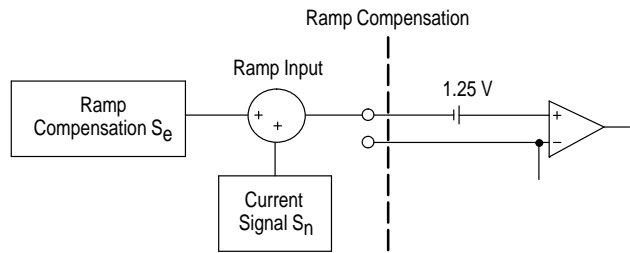
**Instabilities**

In current mode control, an instability can be encountered at any given duty cycle. The instability is caused by the

current feedback loop. It has been shown that the instability is caused by a double pole at half the switching frequency. If an external ramp ( $S_e$ ) is added to the on-time ramp ( $S_n$ ) of the current-sense waveform, stability can be achieved.

One must be careful not to add too much ramp compensation. If too much is added the system will start to perform like a voltage mode regulator. All benefits of current mode control will be lost. Figure 25 is an example of one way in which external ramp compensation can be implemented.

**Figure 20. Ramp Compensation**



A simple equation can be used to calculate the amount of external ramp slope necessary to add that will achieve stability in the current loop. For the following equations, the calculated values for the application circuit in Figure 34 are also shown.

$$S_e = \frac{V_O}{L} \left( \frac{N_S}{N_P} \right) (R_S) A_i$$

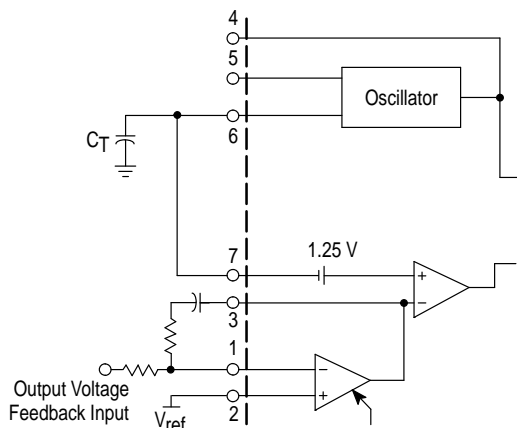
- where:
- $V_O$  = DC output voltage
  - $N_P, N_S$  = number of power transformer primary or secondary turns
  - $A_i$  = gain of the current sense network (see Figures 23 and 24)
  - $L$  = output inductor
  - $R_S$  = current sense resistance

For the application circuit:  $S_e = \frac{5}{1.8 \mu} \left( \frac{2}{8} \right) (0.3)(0.55)$   
 $= 0.115$  V/ms

PIN FUNCTION DESCRIPTION

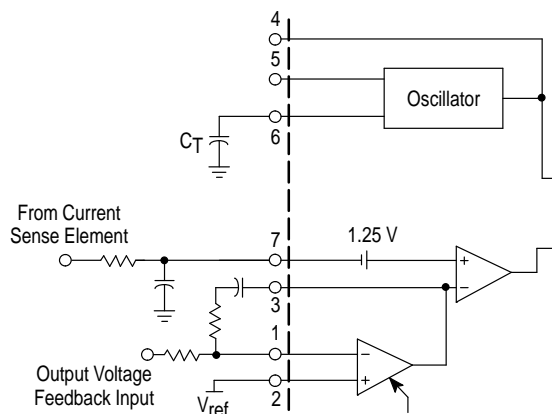
Pin	Function	Description
DIP/SOIC		
1	Error Amp Inverting Input	This pin is usually used for feedback from the output of the power supply.
2	Error Amp Noninverting Input	This pin is used to provide a reference in which an error signal can be produced on the output of the error amp. Usually this is connected to $V_{ref}$ , however an external reference can also be used.
3	Error Amp Output	This pin is provided for compensating the error amp for poles and zeros encountered in the power supply system, mostly the output LC filter.
4	Clock	This is a bidirectional pin used for synchronization.
5	$R_T$	The value of $R_T$ sets the charge current through timing Capacitor, $C_T$ .
6	$C_T$	In conjunction with $R_T$ , the timing Capacitor sets the switching frequency.
7	Ramp Input	For voltage mode operation this pin is connected to $C_T$ . For current mode operation this pin is connected through a filter to the current sensing element.
8	Soft-Start	A capacitor at this pin sets the Soft-Start time.
9	Current Limit/Shutdown	This pin has two functions. First, it provides cycle-by-cycle current limiting. Second, if the current is excessive, this pin will reinitiate a Soft-Start cycle.
10	Ground	This pin is the ground for the control circuitry.
11	Current Limit Reference Input	This pin voltage sets the threshold for cycle-by-cycle current limiting.
12	Power Ground	This is a separate power ground return that is connected back to the power source. It is used to reduce the effects of switching transient noise on the control circuitry.
13	$V_C$	This is a separate power source connection for the outputs that is connected back to the power source input. With a separate power source connection, it can reduce the effects of switching transient noise on the control circuitry.
14	Output	This is a high current totem pole output.
15	$V_{CC}$	This pin is the positive supply of the control IC.
16	$V_{ref}$	This is a 5.1 V reference. It is usually connected to the noninverting input of the error amplifier.

Figure 21. Voltage Mode Operation



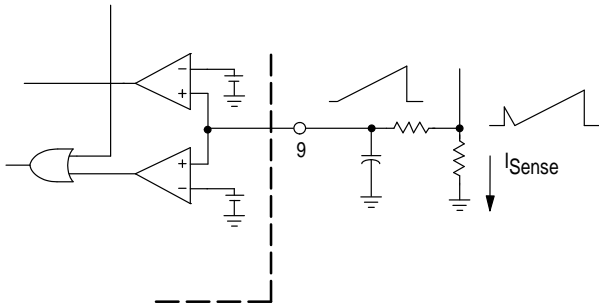
In voltage mode operation, the control range on the output of the Error Amplifier from 0% to 90% duty cycle is from 2.25 V to 4.05 V.

Figure 22. Current Mode Operation



In current mode control, an RC filter should be placed at the ramp input to filter the leading edge spike caused by turn-on of a power MOSFET.

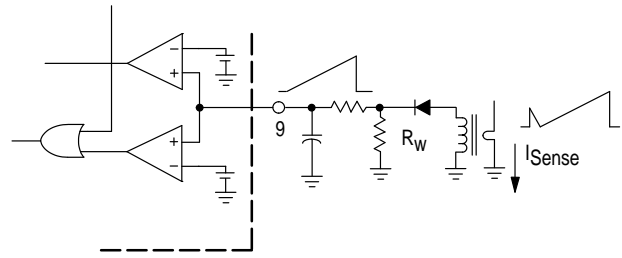
Figure 23. Resistive Current Sensing



The addition of an RC filter will eliminate instability caused by the leading edge spike on the current waveform. This sense signal can also be used at the ramp input pin for current mode control. For ramp compensation it is necessary to know the gain of the current feedback loop. If a transformer is used, the gain can be calculated by:

$$A_i = \frac{R_{Sense}}{\text{turns ratio}}$$

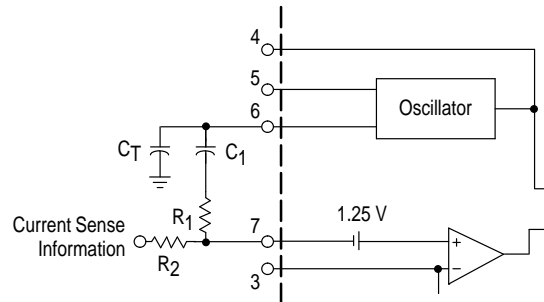
Figure 24. Primary Side Current Sensing



The addition of an RC filter will eliminate instability caused by the leading edge spike on the current waveform. This sense signal can also be used at the ramp input pin for current mode control. For ramp compensation it is necessary to know the gain of the current feedback loop. The gain can be calculated by:

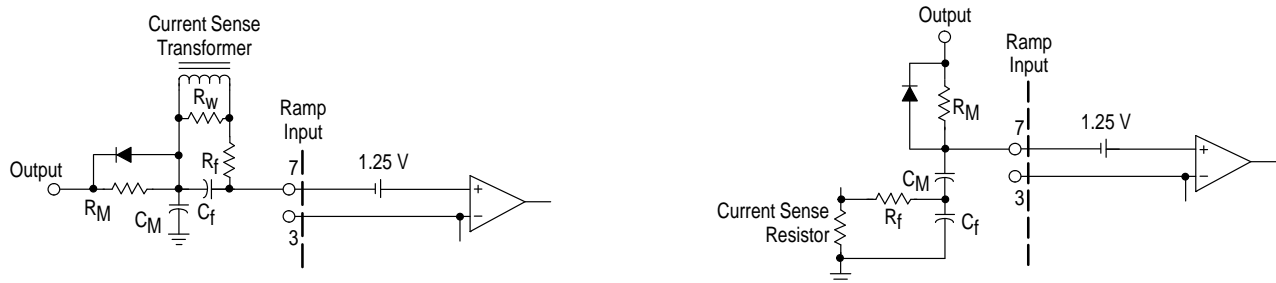
$$A_i = \frac{R_w}{\text{turns ratio}}$$

Figure 25A. Slope Compensation (Noise Sensitive)



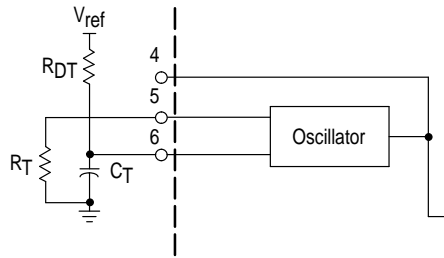
This method of slope compensation is easy to implement, however, it is noise sensitive. Capacitor  $C_1$  provides AC coupling. The oscillator signal is added to the current signal by a voltage divider consisting of resistors  $R_1$  and  $R_2$ .

Figure 25B. Slope Compensation (Noise Immune)



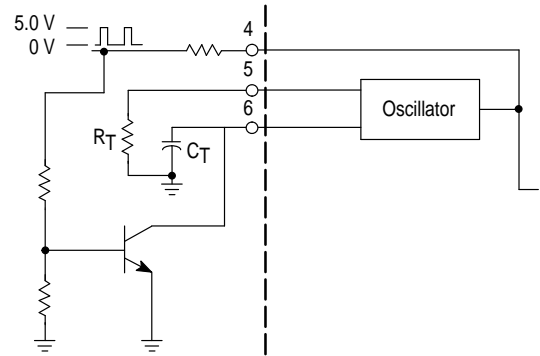
When only one output is used, this method of slope compensation can be used and it is relatively noise immune. Resistor  $R_M$  and capacitor  $C_M$  provide the added slope necessary. By choosing  $R_M$  and  $C_M$  with a larger time constant than the switching frequency, you can assume that its charge is linear. First choose  $C_M$ , then  $R_M$  can be adjusted to achieve the required slope. The diode provides a reset pulse at the ramp input at the end of every cycle. The charge current  $I_M$  can be calculated by  $I_M = C_M S_e$ . Then  $R_M$  can be calculated by  $R_M = V_{CC}/I_M$ .

Figure 26. Dead Time Addition



Additional dead time can be added by the addition of a dead time resistor from  $V_{ref}$  to  $C_T$ . See text on Oscillator section for more information.

Figure 27. External Clock Synchronization



The sync pulse fed into the clock pin must be at least 3.9 V.  $R_T$  and  $C_T$  need to be set 10% slower than the sync frequency. This circuit is also used in Voltage Mode operation for master/slave operation. The clock signal would be coming from the master which is set at the desired operating frequency, while the slave is set 10% slower.

Figure 28. Current Mode Master/Slave Operation Over Short Distances

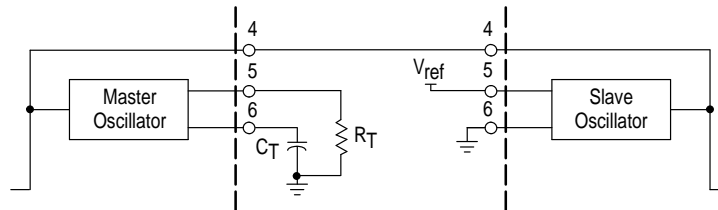


Figure 29. Synchronization Over Long Distances

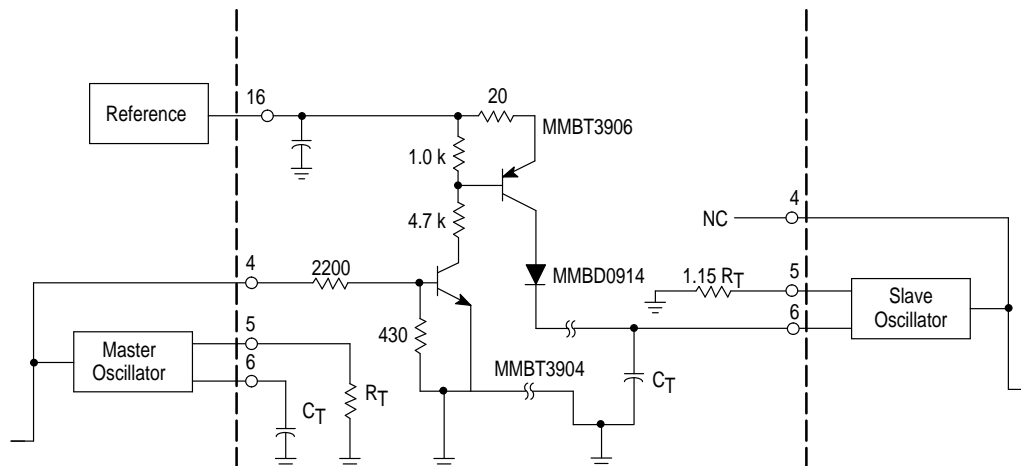
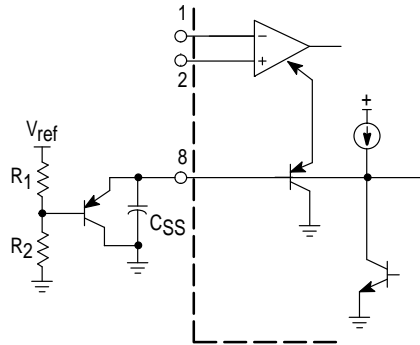


Figure 30. Buffered Maximum Clamp Level

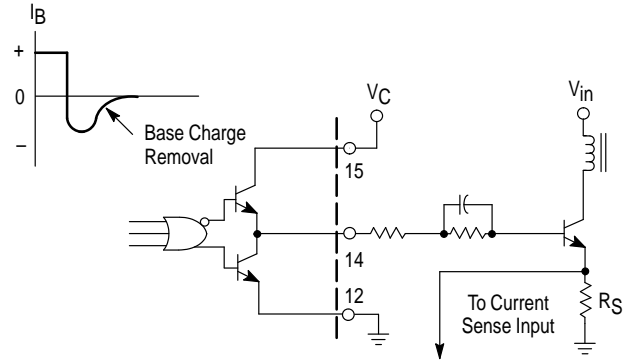


In voltage mode operation, the maximum duty cycle can be clamped. By the addition of a PNP transistor to buffer the clamp voltage, the Soft-Start current is not affected by  $R_1$ .

The new equation for Soft-Start is 
$$t \approx \frac{V_{\text{clamp}} + 0.6}{9.0 \mu\text{A}} (C_{\text{SS}})$$

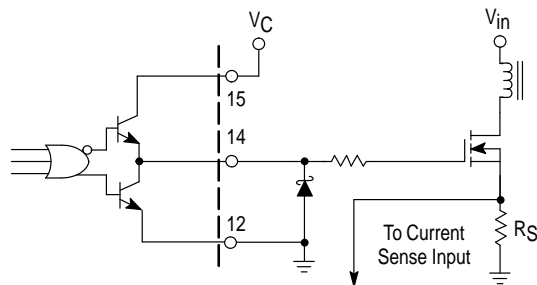
In current mode operation, this circuit will limit the maximum voltage allowed at the ramp input to end a cycle.

Figure 31. Bipolar Transistor Drive



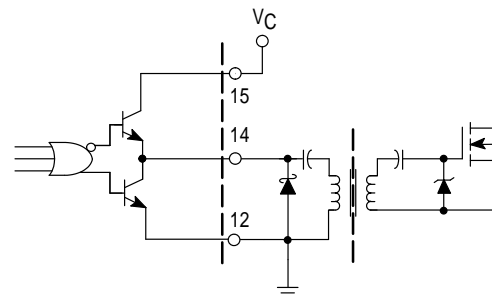
The totem pole output can furnish negative base current for enhanced transistor turn-off, with the addition of the capacitor in series with the base.

Figure 32. MOSFET Parasitic Oscillations



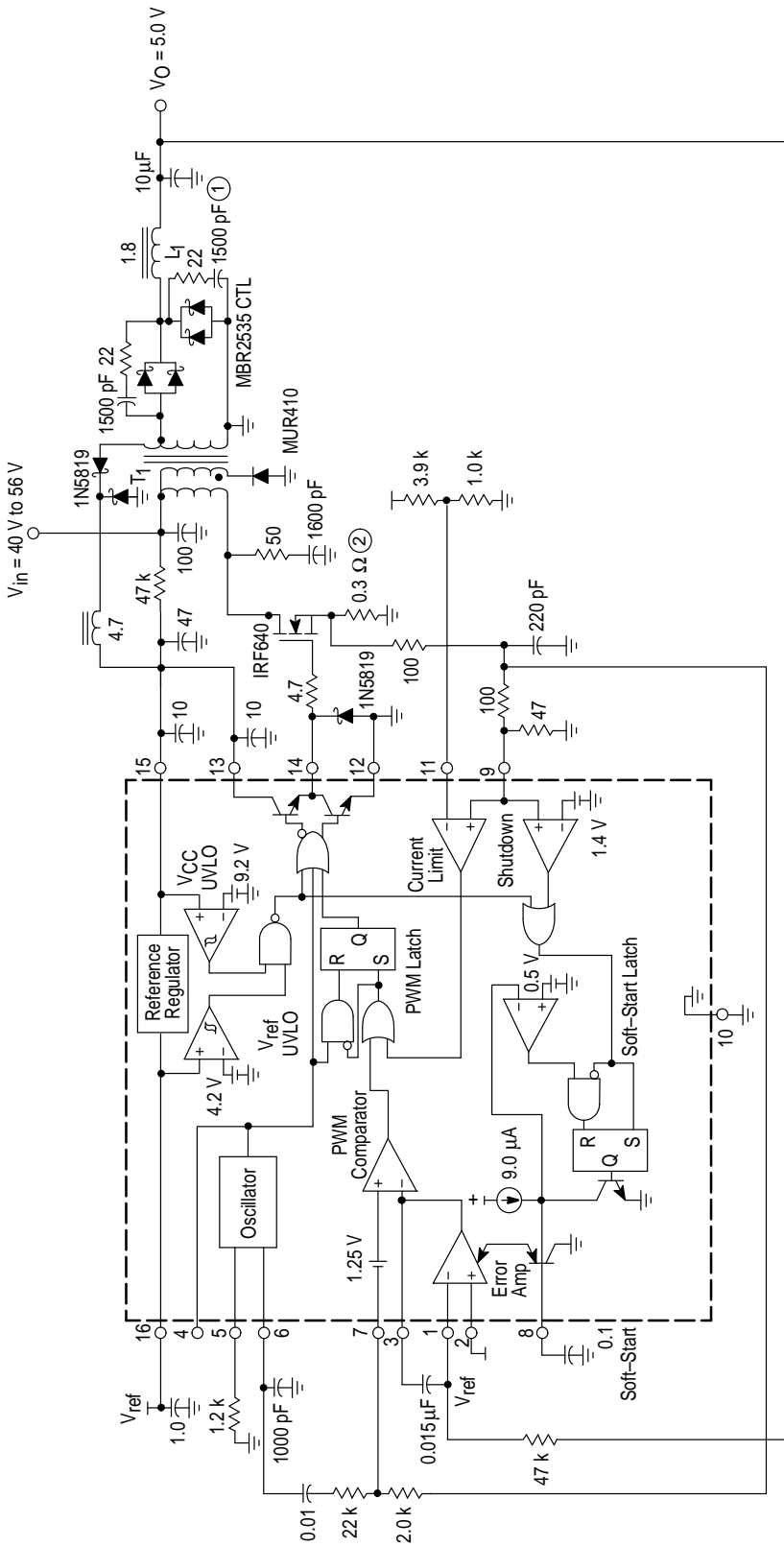
A series gate resistor may be needed to dampen high frequency parasitic oscillation caused by the MOSFET's input capacitance and any series wiring inductance in the gate-source circuit. The series resistor will also decrease the MOSFET switching speed. A Schottky diode can reduce the driver's power dissipation due to excessive ringing, by preventing the output pin from being driven below ground. The Schottky diode also prevents substrate injection when the output pin is driven below ground.

Figure 33. Isolated MOSFET Drive



The totem pole output can easily drive pulse transformers. A Schottky diode is recommended when driving inductive loads at high frequencies. The diode can reduce the driver's power dissipation due to excessive ringing, by preventing the output pin from being driven below ground.

Figure 34. Application Circuit



- T<sub>1</sub> - Primary: 8 turns #48 AWG (1300 strands litz wire)  
 Secondary: 2 turns 0.003" (2 layers) copper foil  
 Bootstrap: 1 turn added to secondary #56 AWG  
 Core: Philips 3F3, part #4312 020 4124  
 Bobbin: Philips part #4322 021 3525  
 Colicraft P3269-A
- L<sub>1</sub> - 2 turns #48 AWG (1300 strands litz wire)  
 Core: Philips 3F3, part #EP10-3F3  
 Bobbin: Philips part #EP10PCB1-8  
 L = 1.8 µH  
 Colicraft P3270-A

Heatsinks - Power FET: AAVID Heatsink #533902B02552 with clip  
 Output Rectifiers: AAVID Heatsink #533402B02552 with clip  
 Insulators - All power devices are insulated with Berquist Sil-Pad 150

- ① - 10(1.0 µF) ceramic capacitors in parallel
- ② - 5(1.5 Ω) resistors in parallel

Test	Condition	Result
Line Regulation	V <sub>in</sub> = 40 V to 56 V, I <sub>O</sub> = 7.5 A	14 mV = ±0.275%
Load Regulation	V <sub>in</sub> = 48 V, I <sub>O</sub> = 4.0 A to 7.5 A	54 mV = ± 1.0%
Output Ripple	V <sub>in</sub> = 48 V, I <sub>O</sub> = 7.5 A	10 mVp-p
Efficiency	V <sub>in</sub> = 48 V, I <sub>O</sub> = 7.5 A	69.8%

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Figure 35. PC Board With Components

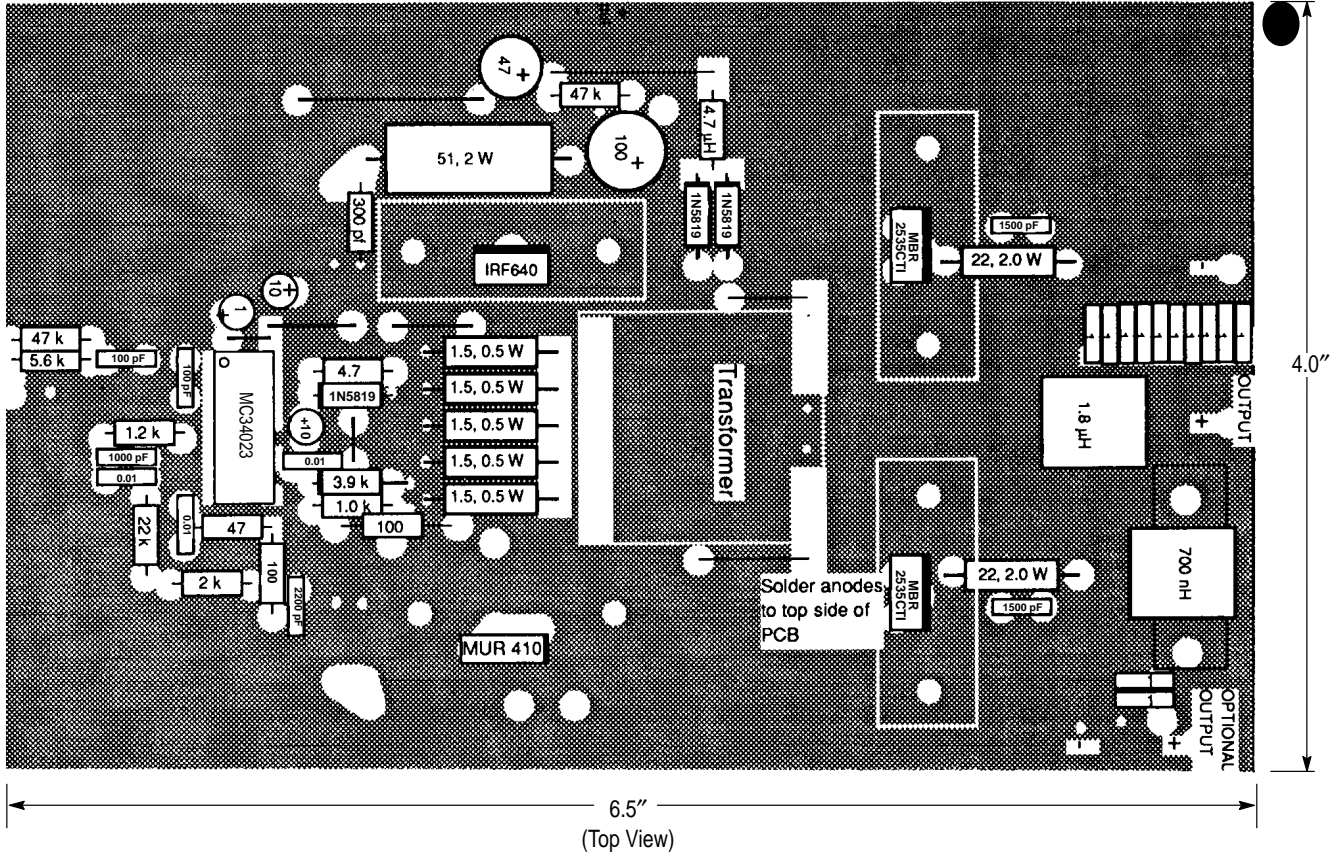
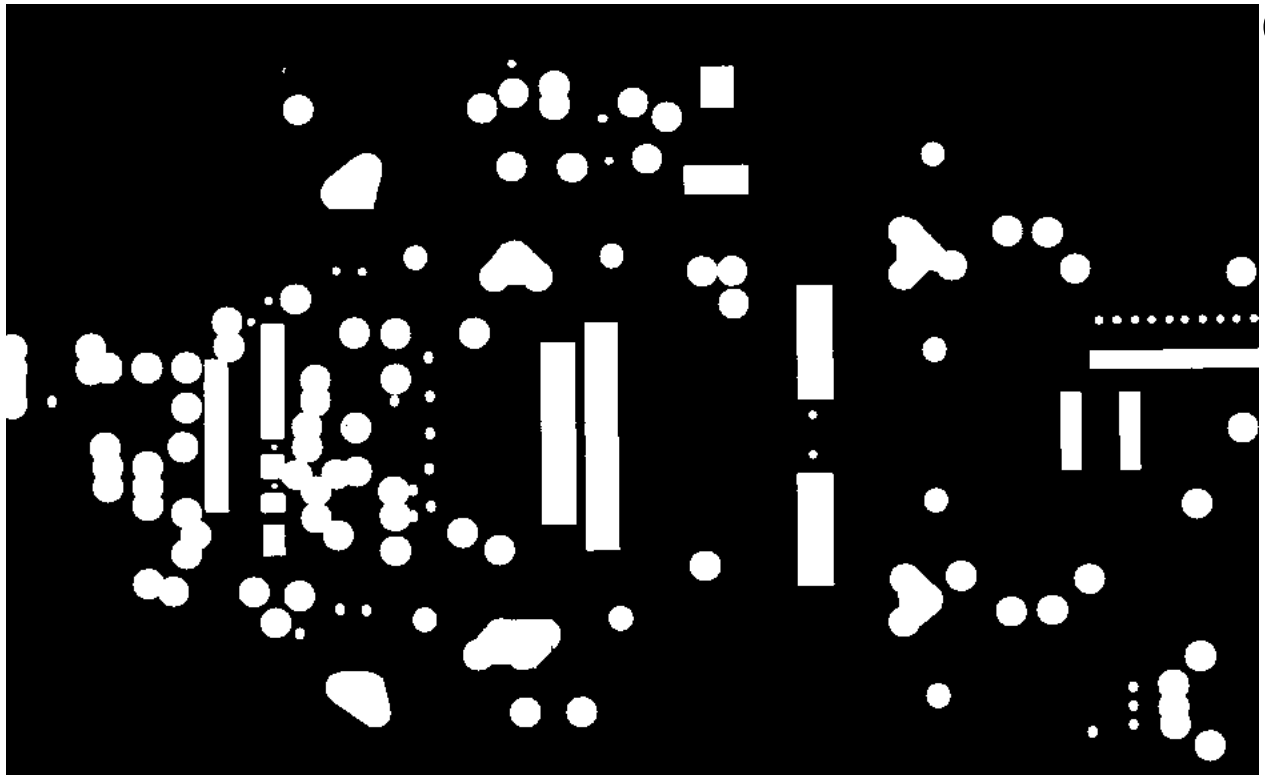
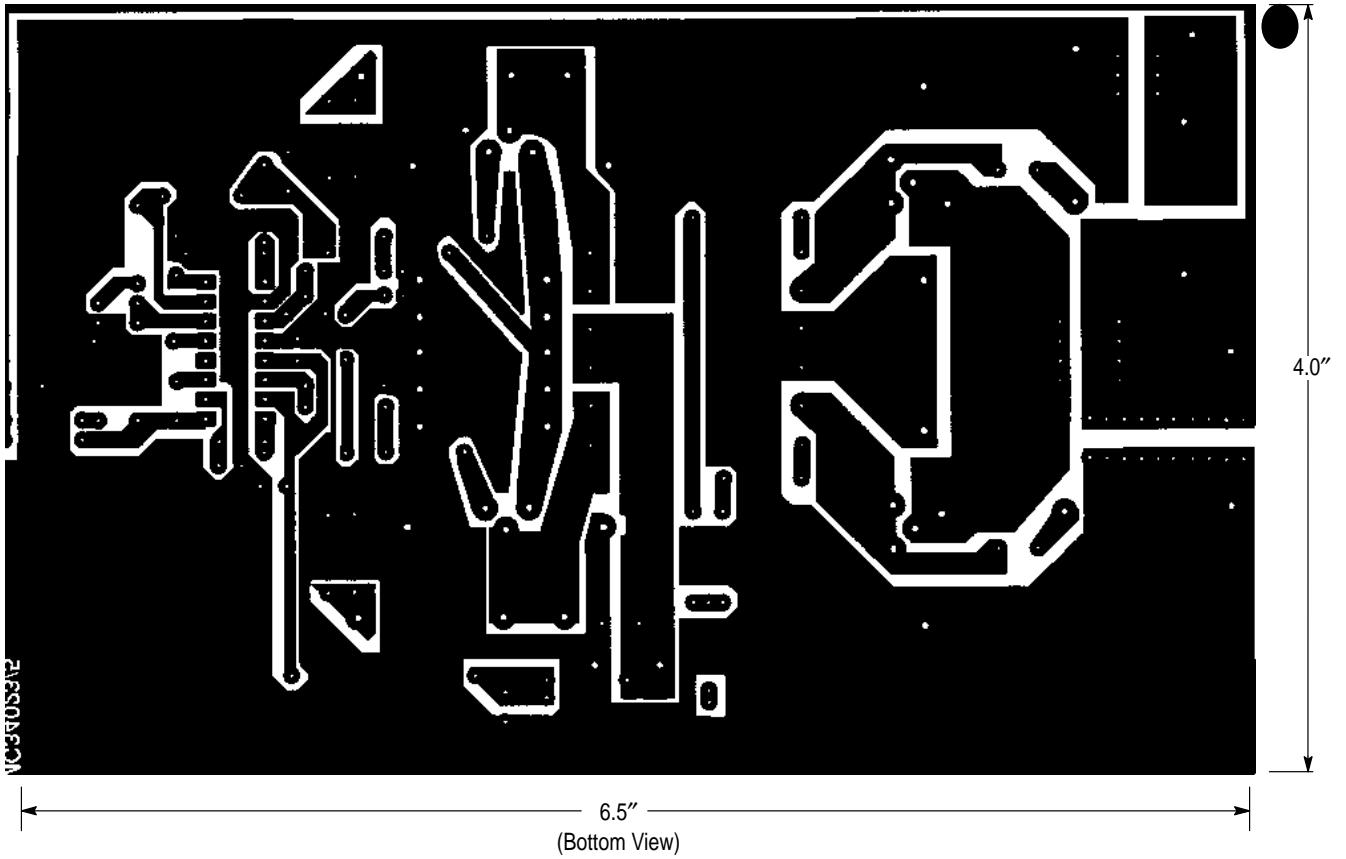


Figure 36. PC Board Without Components



(Top View)



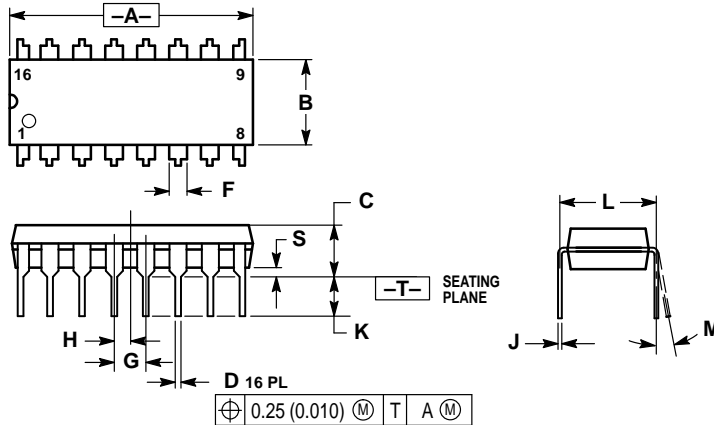
6.5"  
(Bottom View)



# MC34023 MC33023

## OUTLINE DIMENSIONS

### P SUFFIX PLASTIC PACKAGE CASE 648-08

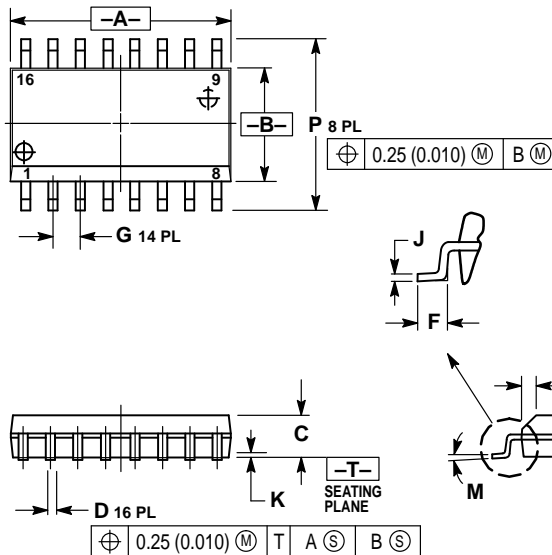


#### NOTES:

1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. DIMENSION L TO CENTER OF LEADS WHEN FORMED PARALLEL.
4. DIMENSION B DOES NOT INCLUDE MOLD FLASH.
5. ROUNDED CORNERS OPTIONAL.

DIM	INCHES		MILLIMETERS	
	MIN	MAX	MIN	MAX
A	0.740	0.770	18.80	19.55
B	0.250	0.270	6.35	6.85
C	0.145	0.175	3.69	4.44
D	0.015	0.021	0.39	0.53
F	0.040	0.070	1.02	1.77
G	0.100 BSC		2.54 BSC	
H	0.050 BSC		1.27 BSC	
J	0.008	0.015	0.21	0.38
K	0.110	0.130	2.80	3.30
L	0.295	0.305	7.50	7.74
M	0°	10°	0°	10°
S	0.020	0.040	0.51	1.01

### DW SUFFIX PLASTIC PACKAGE CASE 751G-02 (SO-16L)



#### NOTES:

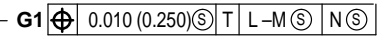
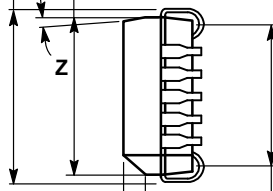
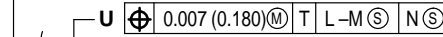
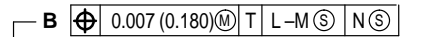
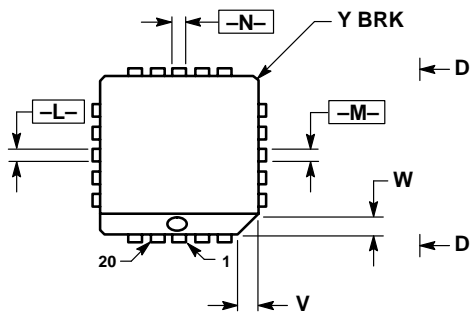
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: MILLIMETER.
3. DIMENSIONS A AND B DO NOT INCLUDE MOLD PROTRUSION.
4. MAXIMUM MOLD PROTRUSION 0.15 (0.006) PER SIDE.
5. DIMENSION D DOES NOT INCLUDE DAMBAR PROTRUSION. ALLOWABLE DAMBAR PROTRUSION SHALL BE 0.13 (0.005) TOTAL IN EXCESS OF D DIMENSION AT MAXIMUM MATERIAL CONDITION.

DIM	INCHES		MILLIMETERS	
	MIN	MAX	MIN	MAX
A	0.400	0.411	10.15	10.45
B	0.292	0.299	7.40	7.60
C	0.093	0.104	2.35	2.65
D	0.014	0.019	0.35	0.49
F	0.020	0.035	0.50	0.90
G	0.050 BSC		1.27 BSC	
J	0.010	0.012	0.25	0.32
K	0.004	0.009	0.10	0.25
M	0°	7°	0°	7°
P	0.395	0.415	10.05	10.55
R	0.010	0.029	0.25	0.75

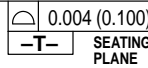
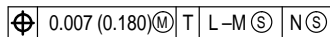
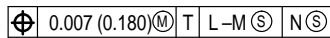
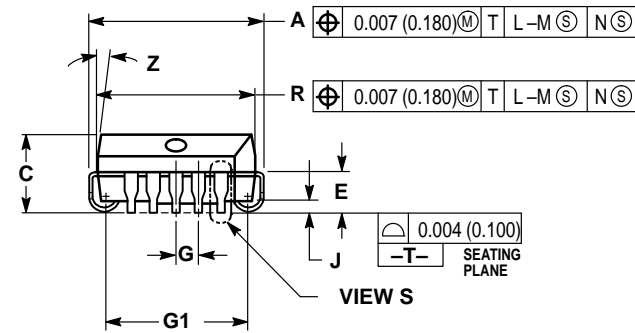
# MC34023 MC33023

## OUTLINE DIMENSIONS

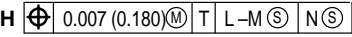
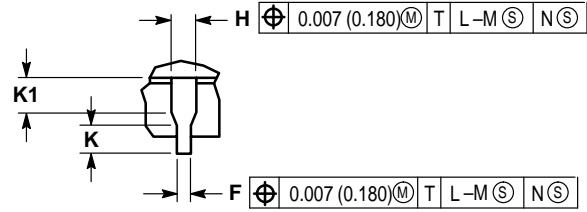
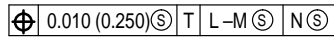
FN SUFFIX  
PLASTIC PACKAGE  
CASE 775-02  
(PLCC)



VIEW D-D



VIEW S




VIEW S

NOTES:

- DATUMS -L-, -M-, AND -N- DETERMINED WHERE TOP OF LEAD SHOULDER EXITS PLASTIC BODY AT MOLD PARTING LINE.
- DIM G1, TRUE POSITION TO BE MEASURED AT DATUM -T-, SEATING PLANE.
- DIM R AND U DO NOT INCLUDE MOLD FLASH. ALLOWABLE MOLD FLASH IS 0.010 (0.250) PER SIDE.
- DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
- CONTROLLING DIMENSION: INCH.
- THE PACKAGE TOP MAY BE SMALLER THAN THE PACKAGE BOTTOM BY UP TO 0.012 (0.300). DIMENSIONS R AND U ARE DETERMINED AT THE OUTERMOST EXTREMES OF THE PLASTIC BODY EXCLUSIVE OF MOLD FLASH, TIE BAR BURRS, GATE BURRS AND INTERLEAD FLASH, BUT INCLUDING ANY MISMATCH BETWEEN THE TOP AND BOTTOM OF THE PLASTIC BODY.
- DIMENSION H DOES NOT INCLUDE DAMBAR PROTRUSION OR INTRUSION. THE DAMBAR PROTRUSION(S) SHALL NOT CAUSE THE H DIMENSION TO BE GREATER THAN 0.037 (0.940). THE DAMBAR INTRUSION(S) SHALL NOT CAUSE THE H DIMENSION TO BE SMALLER THAN 0.025 (0.635).

DIM	INCHES		MILLIMETERS	
	MIN	MAX	MIN	MAX
A	0.385	0.395	9.78	10.03
B	0.385	0.395	9.78	10.03
C	0.165	0.180	4.20	4.57
E	0.090	0.110	2.29	2.79
F	0.013	0.019	0.33	0.48
G	0.050 BSC		1.27 BSC	
H	0.026	0.032	0.66	0.81
J	0.020	-	0.51	-
K	0.025	-	0.64	-
R	0.350	0.356	8.89	9.04
U	0.350	0.356	8.89	9.04
V	0.042	0.048	1.07	1.21
W	0.042	0.048	1.07	1.21
X	0.042	0.056	1.07	1.42
Y	-	0.020	-	0.50
Z	2°	10°	2°	10°
G1	0.310	0.330	7.88	8.38
K1	0.040	-	1.02	-

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