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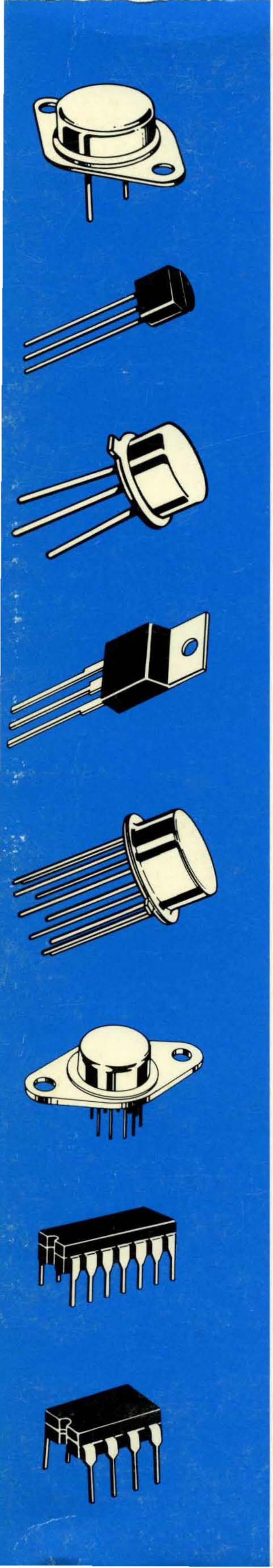

LINEAR/SWITCHMODE VOLTAGE REGULATOR HANDBOOK

THEORY AND PRACTICE

MOTOROLA

LINEAR/SWITCHMODE VOLTAGE REGULATOR HANDBOOK

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Second U.S. Edition
First Printing

Printed in U.S.A.

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PREFACE

In most electronic systems, voltage regulation is required for various functions. Today's complex electronic systems are requiring greater regulating performance, higher efficiency and lower parts count. Present integrated circuit and power package technology has produced IC voltage regulators which can ease the task of regulated power supply design, provide the performance required and remain cost effective. Available in a growing variety, Motorola offers a wide range of regulator products from fixed and adjustable voltage types to special-function and switching regulator control ICs.

This handbook describes Motorola's voltage regulator products and provides information on applying these products. Basic Linear regulator theory and switching regulator topologies has been included along with practical design examples. Other relevant topics include: trade-offs of Linear versus switching regulators, series pass elements for Linear regulators, switching regulator component design considerations, heatsinking, construction and layout, power supply supervisory and protection, and reliability. A Motorola regulator selector guide along with data sheets and an industry cross-reference are also contained in this handbook. A transistor and rectifier selector guide for switching regulators of various configurations and power levels is provided in Appendix A and B.

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SECTION 1
BASIC LINEAR REGULATOR THEORY

A. THE IC VOLTAGE REGULATOR

The basic functional block diagram of an integrated circuit voltage regulator is shown in Figure 1-1. It consists of a stable reference, whose output voltage is V_{REF}, and a high gain error amplifier. The output voltage, V_O, is equal to, or a multiple of, V_{REF}. The regulator will tend to keep V_O constant by sensing any changes in V_O and trying to return it to its original value. Therefore, the ideal voltage regulator could be considered a voltage source with a constant output voltage. However, in practice the IC regulator is better represented by the model shown in Figure 1-2.

In this figure, the regulator is modeled as a voltage source with a positive output impedance, Z_O. The value of the voltage source, V, is not constant; instead, it varies with changes in supply voltage, V_{CC}, and with changes in IC junction temperature, T_J, induced by changes in ambient temperature and power dissipation. Also, the regulator output voltage, V_O, is affected by the voltage drop across Z_O, caused by the output current, I_O. In the following text, the reference and amplifier sections will be described, and their contributions to the changes in the output voltage analyzed.

B. THE VOLTAGE REFERENCE

Naturally, the major requirement for the reference is that it be stable; variations in supply voltage or junction temperature should have little or no effect on the value of the reference voltage, V_{REF}.

The Zener Diode Reference

The simplest form of a voltage reference is shown in Figure 1-3a. It consists of a resistor and a zener diode. The zener voltage, V_Z, is used as the reference voltage. In order to determine V_Z, consider Figure 1-3b. The zener diode, VR1, of Figure 1-3a has been replaced with its equivalent circuit model and the value of V_Z is therefore given by (at a constant junction temperature):

$$V_Z = V_{BZ} + I_Z Z_Z = V_{BZ} + \left(\frac{V_{CC} - V_{BZ}}{R + Z_Z}\right) Z_Z \qquad (1)$$

where V_{BZ} = zener breakdown voltage

I_Z = zener current

Z_Z = zener impedance at I_Z

Note that changes in the supply voltage give rise to changes in the zener current, thereby changing the value of V_Z, the reference voltage.

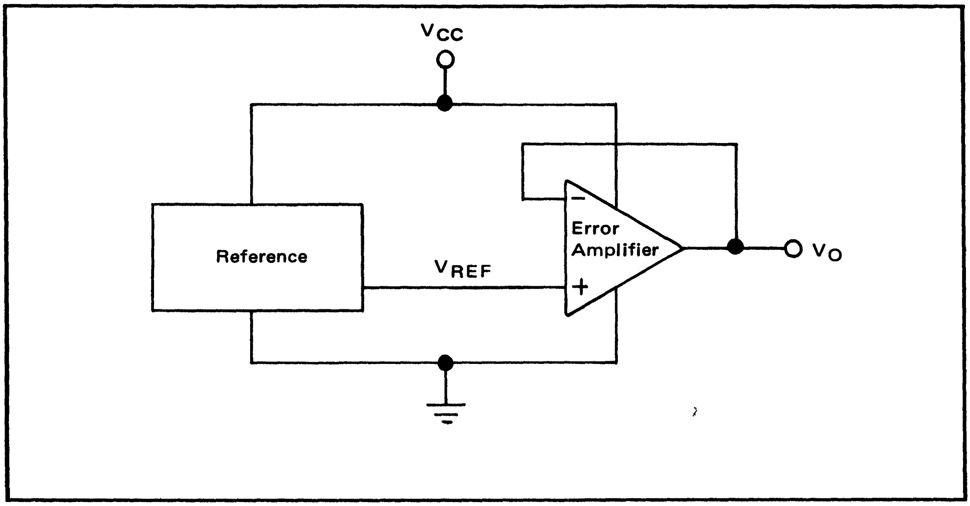

Figure 1-1. Voltage Regulator Functional Block Diagram

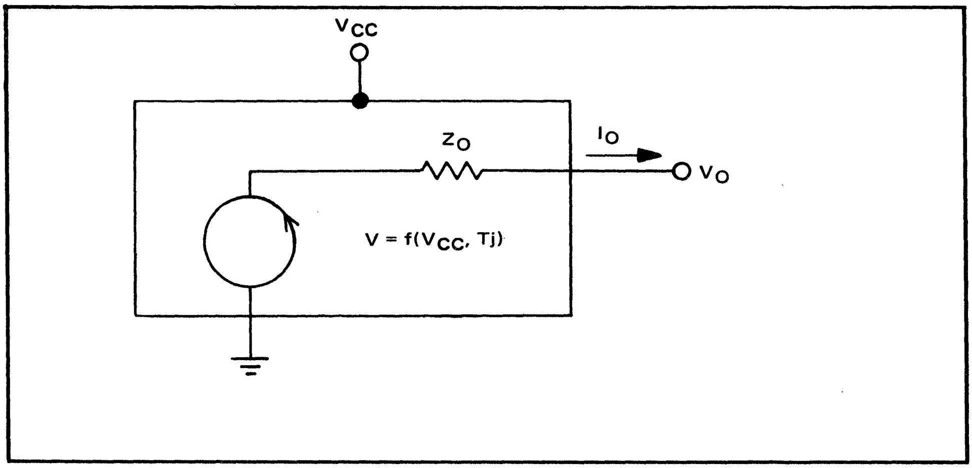

Figure 1-2. Voltage Regulator Equivalent Circuit Model

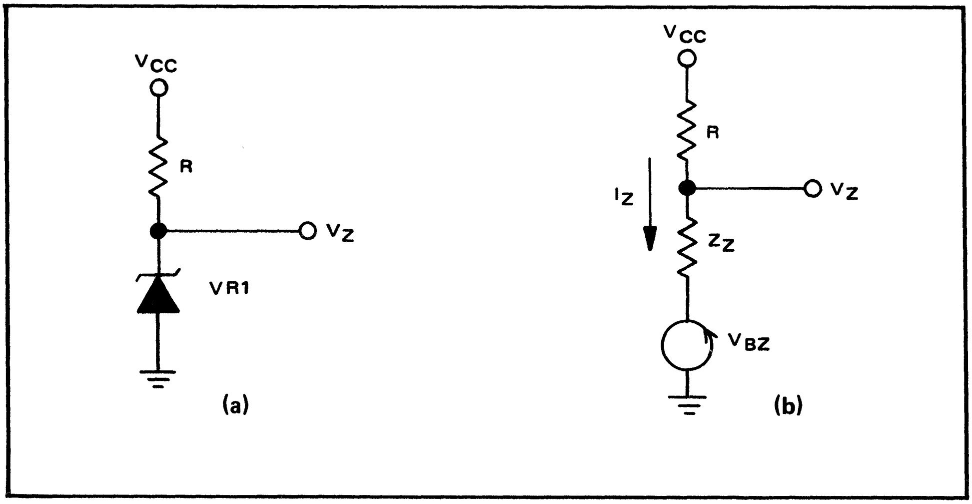

Figure 1-3. Zener Diode Reference

The Constant Current — Zener Reference

The effect of zener impedance can be minimized by driving the zener diode with a constant current as shown in Figure 1-4. The value of the zener current is largely independent of Vcc and is given by:

$$I_Z = \frac{V_{BEQ1}}{R_{SC}} \qquad (2)$$

where V_{BEQ1} = base-emitter voltage of Q1

This gives a reference voltage of:

$$V_{REF} = V_Z + V_{BEQ1} = V_{BZ} + I_Z Z_Z + V_{BEQ1} \qquad (3)$$

where Iz is constant and given by equation 2.

The reference voltage (about 7 V) of this configuration is therefore largely independent of supply voltage variations. This configuration has the additional benefit of better temperature stability than that of a simple resistor-zener reference.

Referring back to Figure 1-3a, it can be seen that the reference voltage temperature stability is equal to that of the zener diode, VR1. The stability of zener diodes used in most integrated circuitry is about +2.2 mV/°C or ≃ .04%/°C(for a 6.2 V zener). If the junction temperature varies 100°C, the zener, or reference, voltage would vary 4%. A variation this large is usually unacceptable.

However, the circuit of Figure 1-4 does not have this drawback. Here the positive 2.2 mV/°C temperature coefficient (TC) of the zener diode is offset by the negative 2.2 mV/°C TC of the V_{BE} of Q1. This results in a reference voltage with very stable temperature characteristics.

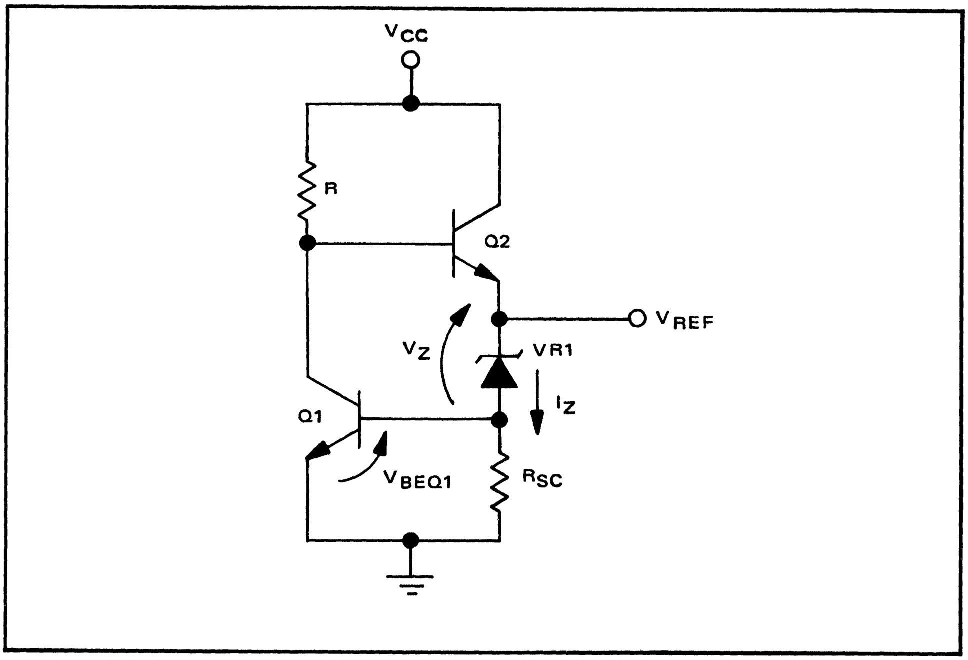

Figure 1-4. Constant Current – Zener Reference

The Bandgap Reference

Although very stable, the circuit of Figure 1-4 does have a disadvantage in that it requires a supply voltage of 9 volts or more. Another type of stable reference which requires only a few volts to operate was described by Widlar[1] and is shown in Figure 1-5. In this circuit V_{REF} is given by:

$$V_{REF} = V_{BEQ3} + I_2R_2 \tag{4}$$

where $I_2 = \frac{V_{BEQ1} - V_{BEQ2}}{R_1}$ (neglecting base currents)

The change in V_{REF} with junction temperature is given by:

$$\Delta V_{REF} = \Delta V_{BE3} + \{\frac{\Delta V_{BEQ1} - \Delta V_{BEQ2}}{R_1}\} R_2 \tag{5}$$

It can be shown that,

$$\Delta V_{BEQ1} = \Delta T_j \; K \; 1n \; I_1 \tag{6}$$

and $$\Delta V_{BEQ2} = \Delta T_j \; K \; 1n \; I_2 \tag{7}$$

where K = a constant

ΔT_j = change in junction temperature

and $I_1 > I_2$

Combining (5), (6), and (7)

$$\Delta V_{REF} = \Delta V_{BEQ3} + \Delta T_jK \left(\frac{R_2}{R_1}\right) 1n\frac{I_1}{I_2} \tag{8}$$

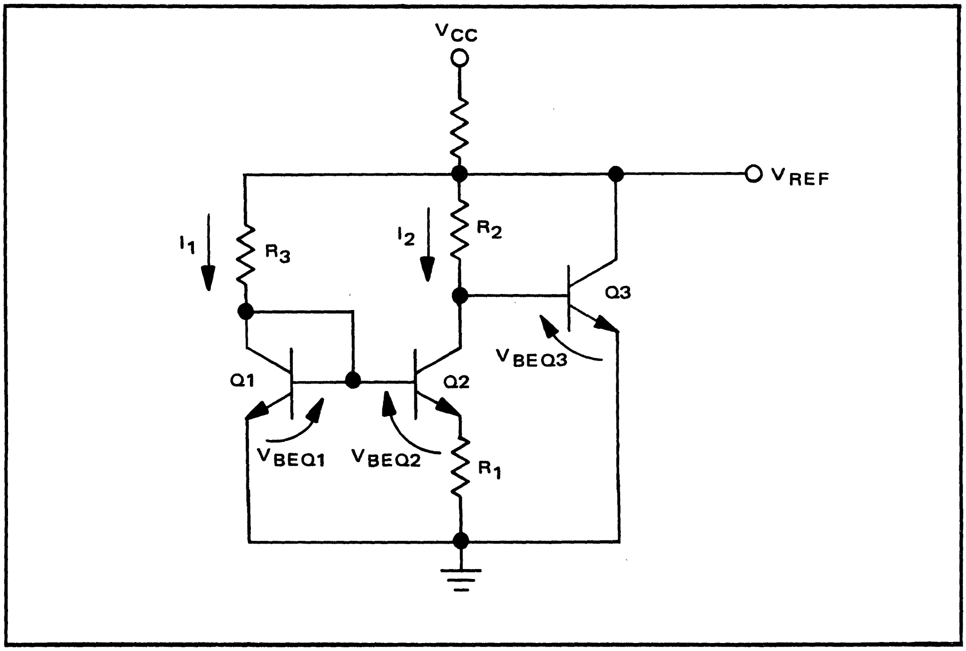

Figure 1-5. Bandgap Reference

Since ΔV_{BEQ3} is negative, and with $I_1 > I_2$, $\ln I_1/I_2$ is positive, the net change in V_{REF} with temperature variations can be made to equal zero by appropriately selecting the values of I_1, R_1, and R_2.

C. THE ERROR AMPLIFIER

Given a stable reference, the error amplifier becomes the determining factor in integrated circuit voltage regulator performance. Figure 1-6 shows a typical differential error amplifier in a voltage regulator configuration. With a constant supply voltage, V_{CC}, and junction temperature, the output voltage is given by:

$$V_O = A_{VOL}\, v_i - Z_{OL}\, I_O = A_{VOL}\{(V_{REF} \pm V_{IO}) - V_O\, \beta\} - Z_{OL}\, I_O \qquad (9)$$

where A_{VOL} = amplifier open loop gain

V_{IO} = input offset voltage

Z_{OL} = open loop output impedance

$\beta = \dfrac{R_1}{R_1 + R_2}$ = feedback ratio (β is always $\leqslant 1$)

I_O = output current

v_i = true differential input voltage

Manipulating (9)

$$V_O = \frac{(V_{REF} \pm V_{IO}) - \dfrac{Z_{OL}}{A_{VOL}} I_O}{\beta + \dfrac{1}{A_{VOL}}} \qquad (10)$$

Note that if the amplifier open loop gain is infinite, this expression reduces to:

$$V_O = \frac{1}{\beta}(V_{REF} \pm V_{IO}) = (V_{REF} \pm V_{IO})\,(1 + \frac{R_2}{R_1}) \qquad (11)$$

The output voltage can thus be set any value equal to or greater than ($V_{REF} \pm V_{IO}$). Note also that if A_{VOL} is not infinite, with constant output current (a non-varying output load), the output voltage can still be "tweaked in" by varying R_1 and R_2, even though V_O will not exactly equal that given by equation 11.

Assuming a stable reference and a finite value of A_{VOL}, inaccuracy of the output voltage can be traced to the following amplifier characteristics:

1. **Amplifier input offset voltage drift —**

The input transistors of integrated circuit amplifiers are usually not perfectly matched. As in operational amplifiers, this is expressed in terms of an input offset voltage, V_{IO}. At a given temperature, this effect can be nulled out of the desired output voltage by adjusting V_{REF} or $1/\beta$. However, V_{IO} drifts with temperature, typically ±5 to 15 μV/°C, causing a proportional change in the output voltage. Closer matching of the internal amplifier input transistors, minimizes this effect, as does selecting a feedback ratio, β, to be close to unity.

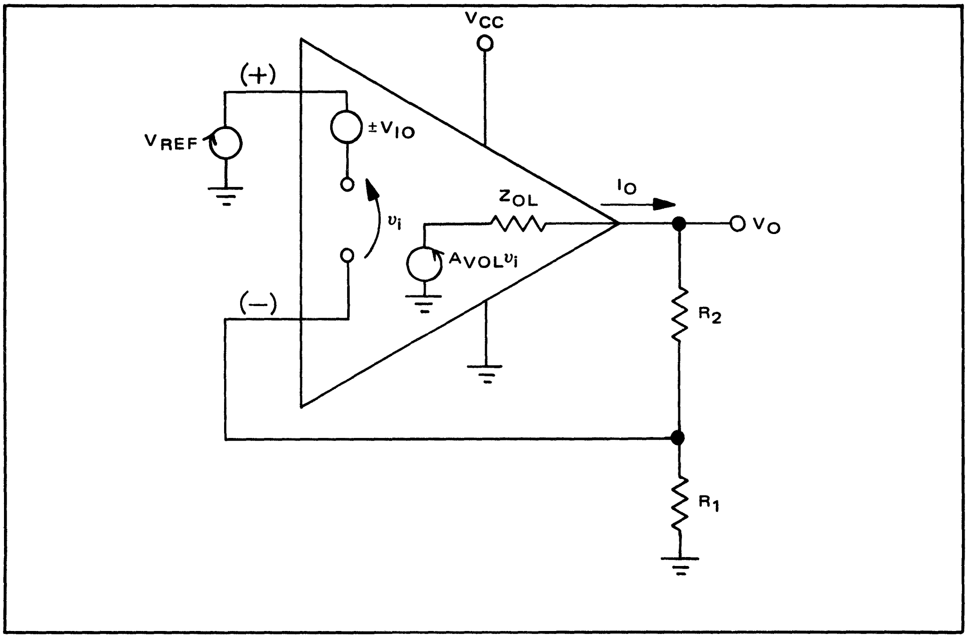

Figure 1-6. Typical Voltage Regulator Configuration

2. **Amplifier power supply sensitivity —**
Changes in regulator output voltage due to power supply voltage variations can be attributed to two amplifier performance parameters: power supply rejection ratio (PSRR) and common-mode rejection ratio (CMRR). In modern integrated circuit regulator amplifiers, the utilization of constant current sources gives such large values of PSRR that this effect on V_O can usually be neglected. However, supply voltage changes can affect the output voltage since these changes appear as common mode voltage changes, and they are best measured by the CMRR.

The definition of common mode voltage, V_{CM}, illustrated by Figure 1-7a, is:

$$V_{CM} = \left(\frac{V_1 + V_2}{2}\right) - \left(\frac{V_+ + V_-}{2}\right) \qquad (12)$$

where V_1 = voltage on amplifier non-inverting input

V_2 = voltage on amplifier inverting input

V_+ = positive supply voltage

V_- = negative supply voltage

In an ideal amplifier, only the differential input voltage ($V_1 - V_2$) has any effect on the output voltage; the value of V_{CM} would not effect the output. In fact, V_{CM} does influence the amplifier output voltage. This effect can be modeled as an additional voltage offset at the amplifier input equal to V_{CM}/CMRR as shown in Figures 1-7b and1-8. The latter figure is the same configuration as Figure 1-6, with amplifier input offset voltage and output impedance deleted for clarity and common-mode voltage effects added. The output voltage of this configuration is given by:

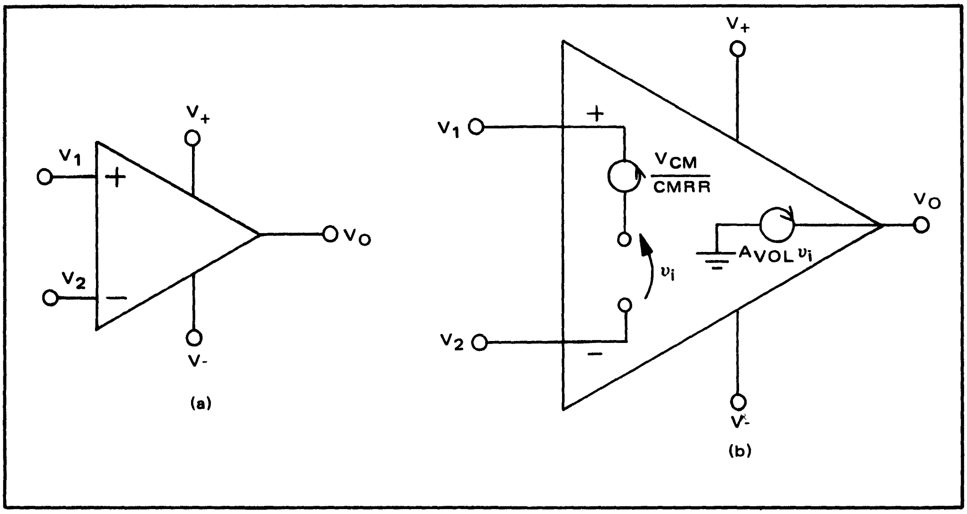

Figure 1-7. Definition of Common-mode Voltage Error

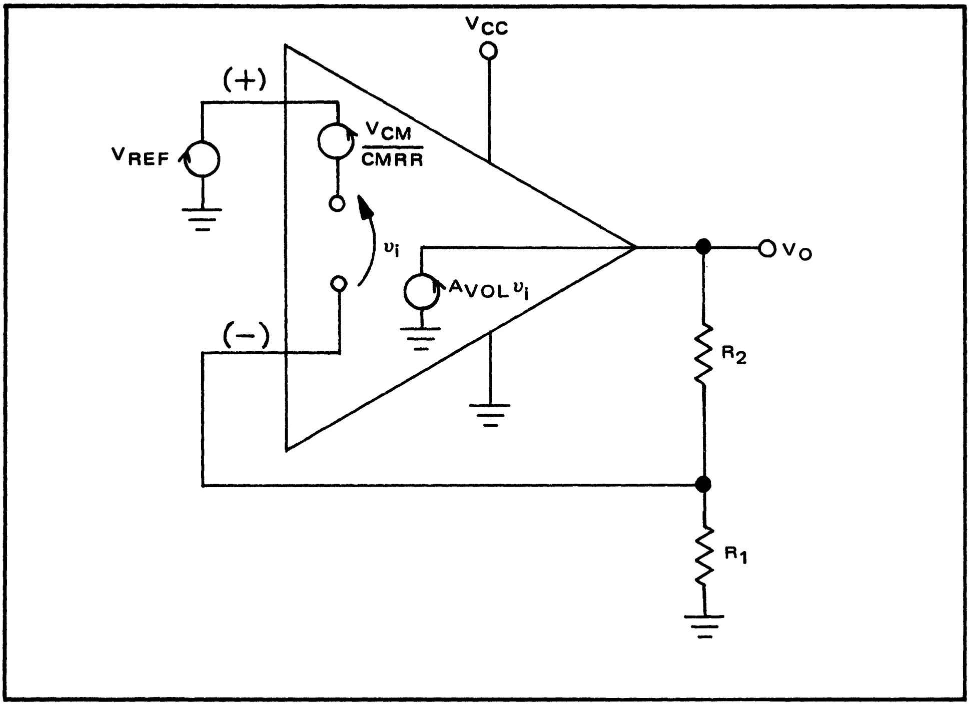

Figure 1-8. Common-mode Regulator Effects

$$V_O = A_{VOL} v_i = A_{VOL}(V_{REF} - \frac{V_{CM}}{CMRR} - \beta V_O) \quad (13)$$

Manipulating,

$$V_O = \frac{(V_{REF} - \frac{V_{CM}}{CMRR})}{\beta + \frac{1}{A_{VOL}}} \quad (14)$$

where $V_{CM} = V_{REF} - \frac{V_{CC}}{2}$ (15)

and CMRR = common-mode rejection ratio

It can be seen from equations (14) and (15) that the output can vary when V_{CC} varies. This can be reduced by designing the amplifier to have a high A_{VOL}, a high CMRR, and by choosing the feedback ratio, β, to be unity.

3. Amplifier Output Impedance —

Referring back to equation (9), it can be seen that the equivalent regulator output impedance, Z_O, is given by:

$$Z_O = \frac{\Delta V_O}{\Delta I_O} \simeq \frac{Z_{OL}}{\beta A_{VOL}} \quad (16)$$

This impedance must be as low as possible, in order to minimize load current effects on the output voltage. This can be accomplished by lowering Z_{OL}, choosing an amplifier with high A_{VOL}, and by selecting the feedback ratio, β, to be unity.

A simple way of lowering the effective value of Z_{OL} is to make an impedance transformation with an emitter follower, as shown in Figure 1-9. Given a change in output current, ΔI_O, the amplifier will see a change of only $\Delta I_O/h_{FEQ1}$ in its output current, I_O'. Therefore Z_{OL} in equation (16) has been effectively reduced to Z_{OL}/h_{FEQ1}, reducing the overall regulator output impedance, Z_O.

D. THE REGULATOR WITHIN A REGULATOR APPROACH

In the preceding text, we have analyzed the sections of an integrated circuit voltage regulator and determined how they contribute to its non-ideal performance characteristics. These are shown in Table 1-1 along with procedures which minimize their effects.

It can be seen that in all cases regulator performance can be improved by selecting A_{VOL} as high as possible and $\beta = 1$. Since a limit is soon approached in how much A_{VOL} can be practically obtained in an integrated circuit amplifier, selecting a feedback ratio, β, equal to unity is the only viable way of improving total regulator performance, especially in reducing regulator output impedance. However, this method presents a basic problem to the regulator designer. If the configuration of Figure 1-6 is used, the output voltage cannot be adjusted to a value other than V_{REF}. The solution is to utilize a different regulator configuration known as the "regulator within a regulator approach."[2] Its greatest benefit is in reducing total regulator output impedance.

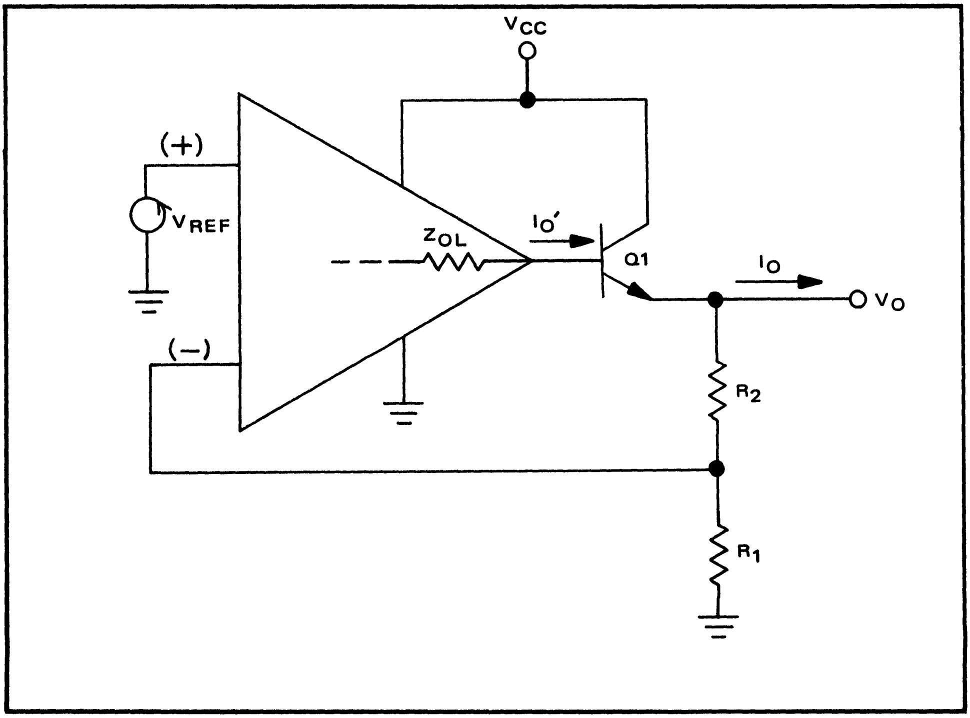

Figure 1-9. Emitter Follower Output

TABLE 1-1

Vo CHANGES SECTION	EFFECT CAN BE INDUCED BY	MINIMIZED BY SELECTING
Reference	Vcc	1. Constant current-zener method 2. Bandgap reference
	Tj	1. Bandgap reference 2. TC compensated zener method
Amplifier	Vcc	1. High CMRR amplifier 2. High AVOL amplifier 3. $\beta = 1$
	Tj	1. Low VIO drift amplifier 2. High AVOL amplifier 3. $\beta = 1$
	Io	1. Low ZOL amplifier 2. High AVOL amplifier 3. Additional emitter follower output 4. $\beta = 1$

As shown in Figure 1-10, amplifier A1 sets up a voltage, V_1, given by:

$$V_1 \simeq V_{REF}\,(1 + \frac{R_2}{R_1}) \tag{17}$$

V_1 now serves as the reference voltage for amplifier A2, whose output voltage, V_O, is given by:

$$V_O \simeq V_1 \simeq V_{REF}\,(1 + \frac{R_2}{R_1}) \tag{18}$$

Note that the output impedance of A2, and therefore the regulator output impedance, has been minimized by selecting A2's feedback factor to be unity; and that output voltage can still be set at voltages greater than V_{REF} by adjusting R_1 and R_2.

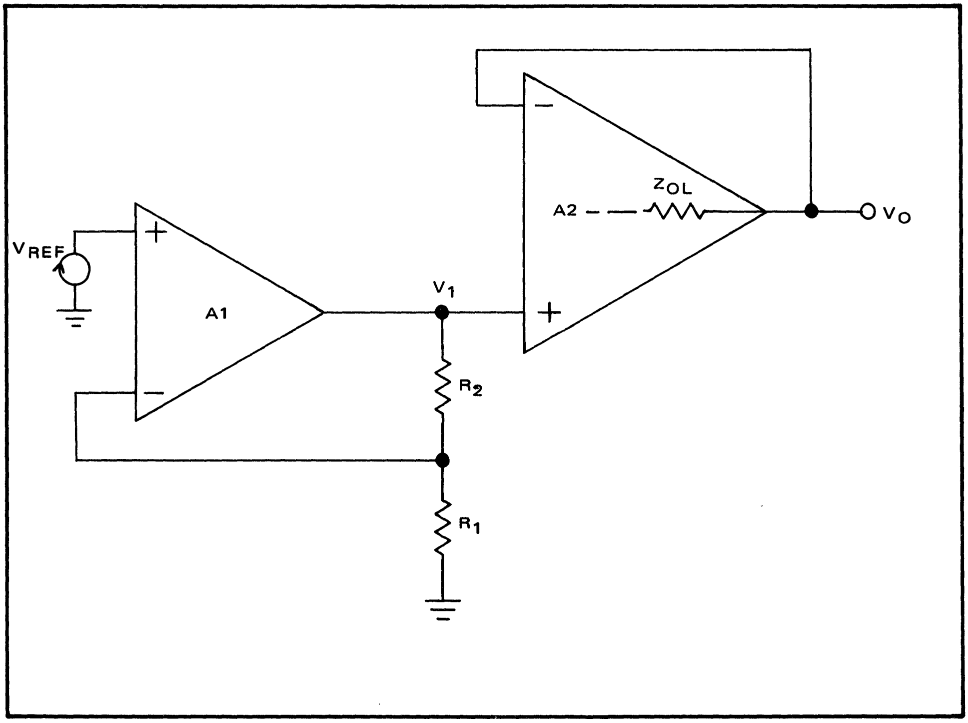

Figure 1-10. The "Regulator within a Regulator" Configuration

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[1] Widlar, R. J., "New Developments in IC Voltage Regulators," IEEE Journal of Solid State Circuits, Feb. 1971, Vol. SC-6, pgs. 2-7.

[2] Tom Fredericksen, IEEE Journal of Solid State Circuits, Vol. SC-3, Number 4, Dec. 1968, "A Monolithic High Power Series Voltage Regulator."

SECTION 2

SELECTING A LINEAR IC VOLTAGE REGULATOR

A. SELECTING THE TYPE OF REGULATOR

There are five basic linear regulator types; these are the positive, negative, fixed output, tracking and floating regulators. Each has its own particular characteristics and best uses, and selection depends on the designer's needs and trade-offs in performance and cost.

1. **Positive Versus Negative Regulators.**

In most cases, a positive regulator is used to regulate positive voltages and a negative regulator negative voltages. However, depending on the system's grounding requirements, each regulator type may be used to regulate the ''opposite'' voltage.

Figures 2-1a and 2-1b show the regulators used in the conventional and obvious mode. Note that the ground reference for each (indicated by the heavy line) is continuous. Several positive regulators could be used with the same input supply to deliver several voltages with common grounds; negative regulators may be utilized in a similar manner.

If no other common supplies or system components operate off the input supply to the regulator, the circuits of Figures 2-1c and 2-1d may be used to regulate positive voltages with a negative regulator and vice versa. In these configurations, the input supply is essentially floated, i.e., neither side of the input is tied to the system ground.

There are methods of utilizing positive regulators to obtain negative output voltages without sacrificing ground bus continuity; however, these methods are only possible at the expense of increased circuit complexity and cost. An example of this technique is shown in Section 3.

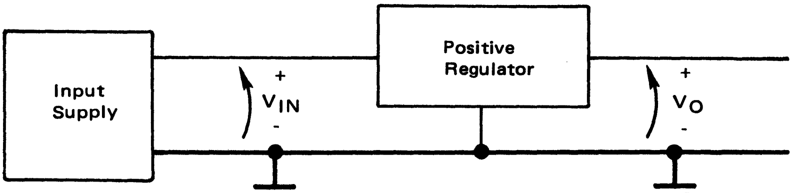
2. **Three Terminal, Fixed Output Regulators**

These regulators offer the designer a simple, inexpensive way to obtain a source of regulated voltage. They are available in a variety of positive or negative output voltages and current ranges. The advantages of these regulators are:

a) Easy to use.
b) Internal overcurrent and thermal protection.
c) No circuit adjustments necessary.
d) Low cost.

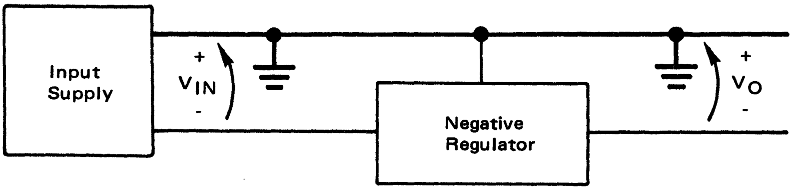
Their disadvantages are:

a) Output voltage cannot be precisely adjusted. (Methods for obtaining adjustable outputs are shown in Section 3).
b) Available only in certain output voltages and currents.
c) Obtaining greater current capability is more difficult than with other regulators. (Methods for obtaining greater output currents are shown in Section 3.)



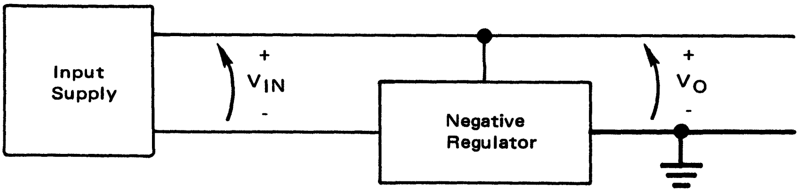
(a)

Positive Output Using Positive Regulator



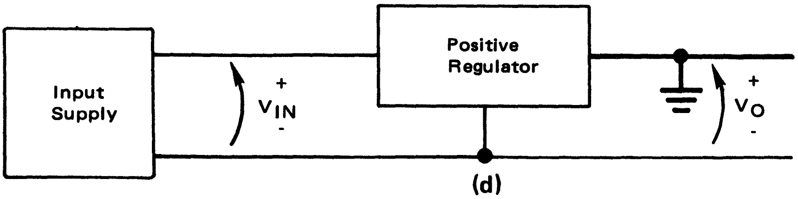
(b)

Negative Output Using Negative Regulator



(c)

Positive Output Using Negative Regulator



(d)

Negative Output Using Positive Regulator

Figure 2-1. Regulator Configurations

3. Three Terminal, Adjustable Output Regulators

Like the three terminal fixed regulators, the three terminal adjustable regulators are easy and inexpensive to use. These devices provide added flexibility with output voltage adjustable over a wide range, from 1.2 V to nearly 40 V, by means of an external, two-resistor voltage divider. A variety of current ranges from 100 mA to 3.0 Amperes are available.

4. Tracking Regulators

Often a regulated source of symmetrical positive and negative voltage is required for supplying op amps, etc. In these cases, a tracking regulator is required. In addition to supplying regulated positive and negative output voltages, the tracking regulator assures that these voltages are balanced; in other words, the midpoint of the positive and negative output voltages is at ground potential.

This function can be implemented using a positive output regulator together with an op amp or negative output regulator. However, this method results in the use of two IC packages and a multitude of external components. To minimize component count, an IC is offered which performs this function in a single package: the MC1568/MC1468 ±15V tracking regulator.

5. Floating Regulators

If the desired output voltage is in excess of 40 volts, a floating regulator such as the MC1566/MC1466 should be considered. The output voltage of this regulator can be any magnitude and is limited only by the capabilities of an external transistor. However, an additional floating low voltage input supply is required.

B. SELECTING AN IC REGULATOR

Once the type of regulator is decided upon, the next step is to choose a specific device. As an aid in choosing an appropriate IC regulator, a Selection Guide is contained in Section 17.

To provide higher currents than are available from monolithic technologies, an IC regulator will often be used as a driver to a boost transistor. This complicates the selection and design task, as there are now several overlapping solutions to many of the design problems.

Unfortunately, there is no exact step-by-step procedure that can be followed which will lead to the ideal regulator and circuit configuration for a specific application. The regulating circuit that is finally accepted will be a compromise between such factors as performance, cost, size and complexity.

Because of this, the following general design procedure is suggested:

1. Select the regulators which meet or exceed the requirements for line regulation, load regulation, TC of the output voltage and operating ambient temperature range. At this point, do not be overly concerned with the regulator capabilities in terms of output voltage, output current, SOA and special features.
2. Next, select application circuits from Section 3 which meet the requirements for output current, output voltage, special features, etc. Preliminary designs using the chosen regulators and circuit configurations are then possible. From these designs a judgement can be made by the designer as to which regulator — circuit configuration combination best meets his requirements in terms of cost, size and complexity.

SECTION 3

LINEAR REGULATOR CIRCUIT CONFIGURATION AND DESIGN CONSIDERATIONS

Once the IC regulators, which meet the designer's performance requirements, have been selected, the next step is to determine suitable circuit configurations. Initial designs are devised and compared to determine the IC regulator/circuit configuration that best meets the designer's requirements. In this section, several circuit configurations and design equations are given for the various regulator ICs. Additional circuit configurations can be found on the device data sheets (see Section 18). Organization is first by regulator type and then by variants, such as current boost. Each circuit diagram has component values for a particular voltage and current regulator design.

A. Positive, Adjustable
B. Negative, Adjustable
C. Positive, Fixed
D. Negative, Fixed
E. Tracking
F. Floating
G. Special
    1. Obtaining Extended Output Voltage Range
    2. Electronic Shutdown
H. General Design Considerations

It should be noted that all circuit configurations shown have constant current limiting; if foldback limiting is desired, see Section 4C for techniques and design equations.

A. POSITIVE, ADJUSTABLE OUTPUT IC REGULATOR CONFIGURATIONS

1. Basic Regulator Configurations

Positive Three-Terminal Adjustables

These adjustables, comprised of the LM117L, LM117M, LM117, and LM150 series devices range in output currents of 100mA, 500mA, 1.5A, and 3.0A respectively. All of these devices utilize the same basic circuit configuration as shown in Figure 3-1A.

MC1723(C)

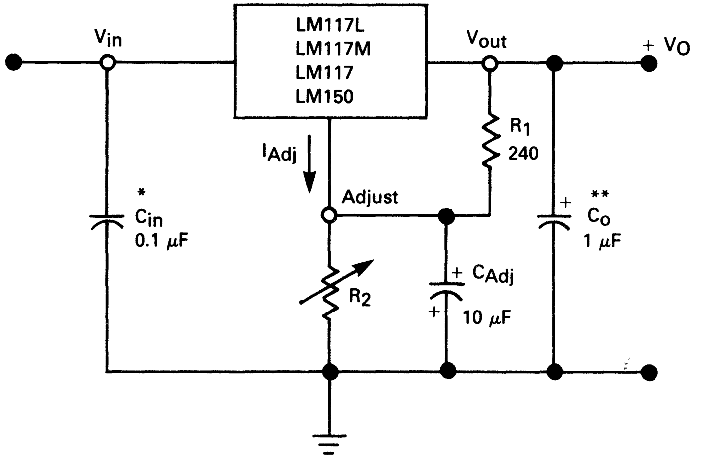
The basic circuit configurations for the MC1723(C) regulator are shown in Figures 3-3A and 3-2A. For output voltages from ≃ 7 V to 37 V the configuration of Figure 3-2A can be used, while Figure 3-3A can be used to obtain output voltages from 2 V to ≃ 7 V.

MC1569, MC1469

Figure 3-4A shows the basic circuit configuration for the MC1569, MC1469 regulator IC. Depending on V_{IN}, T_A, heatsinking and package utilized, output currents in excess of 500 mA can be obtained with this configuration.

2. Output Current Boosting

If output currents greater than those available from the basic circuit configurations are desired, the current boost circuits shown in this section can be used. The output currents which can be obtained with these configurations are limited only by the capabilities of the external pass element(s).



* C_{in} is required if regulator is located an appreciable distance from power supply filter.

** C_O is not needed for stability, however it does improve transient response.

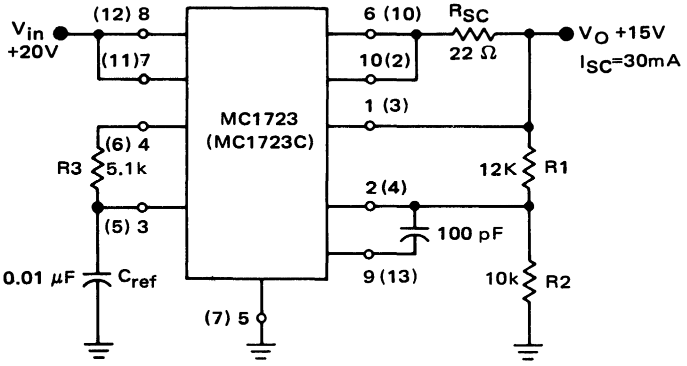
† C_{Adj} is not required; however, it does improve Ripple Rejection

$$V_{out} = 1.25\,V\left(1 + \frac{R_2}{R_1}\right) + I_{Adj}\,R_2$$

Since I_{Adj} is controlled to less than 100 μA, the error associated with this term is negligible in most applications.

Figure 3-1A — Basic Configuration for Positive, Adjustable Ouput Three-Terminal Regulators

Pin Numbers Adjacent to Terminals are for the Metal Package.
Pin Numbers in Parenthesis are for the Dual In-Line Package.



$R_{SC} \cong \frac{0.66V}{I_{SC}}$; $10\,k\Omega < R1+R2 < 100\,k\Omega$

$R3 \cong R1 \parallel R2$; $0 \leqslant C_{REF} \leqslant 0.1\mu F$

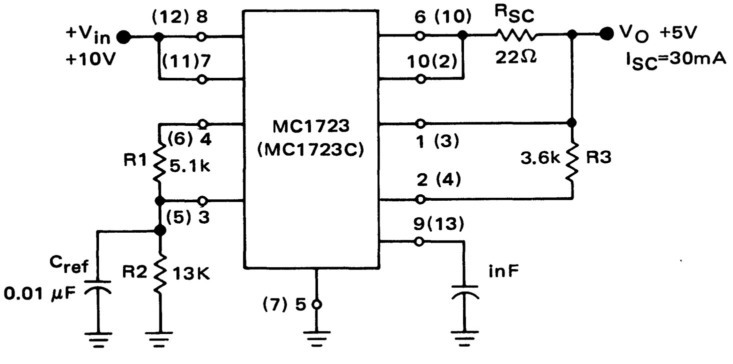
$R2 = \frac{V_{REF}}{V_O}(R1+R2) \approx \frac{7V}{V_O}(R1+R2)$

See Section 3H for General Design Considerations

Values shown are for a $\boxed{15V, 30mA}$ regulator using an MC1723CL for a $T_{A_{MAX}} = 25^{\circ}C$

Figure 3-2A. MC1723 Basic Circuit Configuration for $V_{REF} \leqslant V_O \leqslant 37V$

Pin Numbers Adjacent to Terminals are for the Metal Package.
Pin Numbers in Parenthesis are for the Dual In-Line Package.



$R_{SC} \simeq \frac{0.66V}{I_{SC}}$; $10k\Omega < R1+R2 < 100k\Omega$

$R2 = \frac{V_O}{V_{REF}}(R1+R2) \simeq \frac{V_O}{7V}(R1+R2)$; $R3=R1 \parallel R2$

$0 \leqslant C_{REF} \leqslant 0.1\mu F$

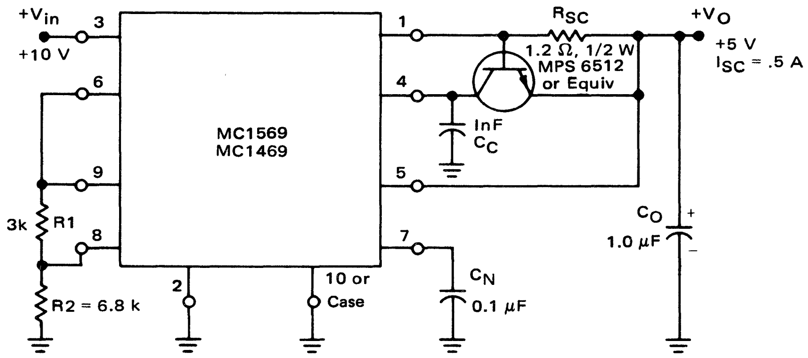
See Section 3H for General Design Considerations

Values shown are for a $\boxed{5V, 30mA}$ regulator using an MC1723CL for a $T_{A_{MAX}} = 70^{\circ}C$

Figure 3-3A. MC1723 Basic Circuit Configuration for $2v \leqslant V_O \leqslant V_{REF}$

MC1723(C)

To obtain greater output currents with the MC1723 the configurations shown in Figures 3-5A and 3-6A can be used. Figure 3-5A uses an NPN external pass element, while a PNP is used in Figure 3-6A.



R2 = 6.8 k; R1 =($\frac{V_O}{V_{REF}}$ -1) R2 ≅ ($2V_O$ -7) kΩ

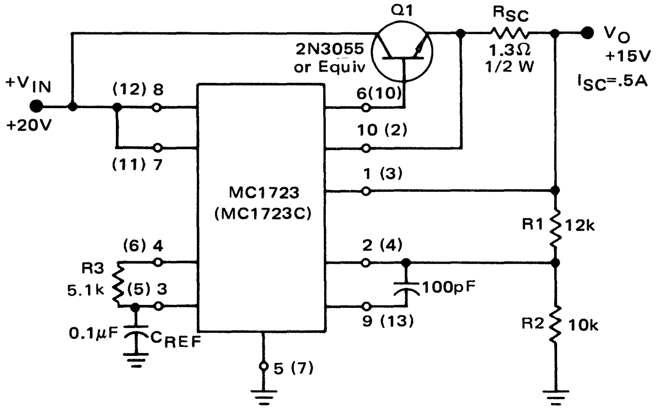
C_N = 0.1 μF; $C_C \geqslant$ InF; $R_{SC} \cong \frac{0.6}{I_{SC}}$

See Section 3H for General Design Considerations

Values shown are for a $\boxed{\text{5V, 500mA}}$ regulator using a MC1569R on a 3°C/W heatsink at T_A up to +70°C

Figure 3-4A. MC1569, MC1469 Basic Circuit Configuration

Pin Numbers Adjacent to Terminals are for the Metal Package.
Pin Numbers in Parenthesis are for the Dual In-Line Package.



$R_{SC} \cong \frac{0.66V}{I_{SC}}$; 10kΩ<R1+R2<100kΩ

$R2 = \frac{V_{REF}}{V_O}(R1+R2) \cong \frac{7V}{V_O}(R1+R2)$

$0 \leqslant C_{REF} \leqslant 0.1\mu F$; $R3 \cong R1 \parallel R2$

Selection of Q1 based on considerations of Section 4

See Section 3H for General Design Considerations

Values shown are for a $\boxed{\text{15V, 500mA}}$ regulator using an MC1723CL (unheatsinked) and a 2N3055 on a 6°C/W heatsink for T_A up to +70°C

Figure 3-5A. MC1723(C) NPN Boost Configuration

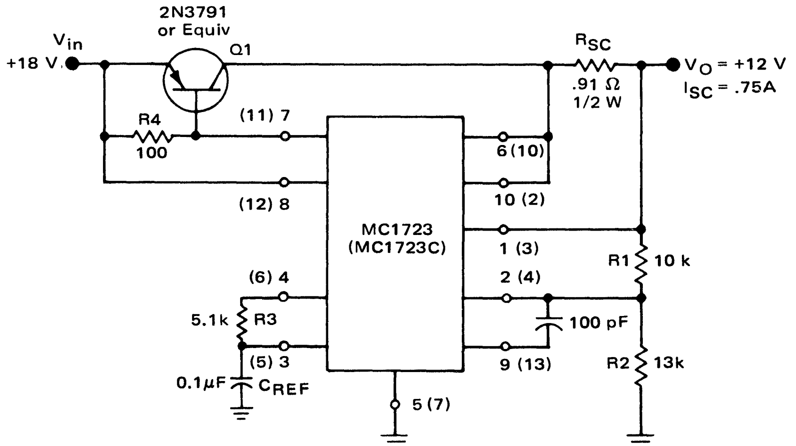
MC1569, MC1469

Figures 3-7A and 3-8A show typical current boosting configurations for the MC1569, MC1469 using an NPN and a PNP series pass element, respectively.

3. High Efficiency Regulator Configurations

When large output currents at voltages under approximately 9 volts are desired, the configurations of Figures 3-9A and 3-10A can be utilized to obtain increased operating efficiency. This is accomplished by providing a separate low voltage input supply for the pass element. This method, however, usually necessitates that separate short circuit protection be provided for the IC regulator and external pass element. Figure 3-9A shows a high efficiency regulator configuration for the MC1723(C), while Figure 3-10A is for the MC1469, MC1469.

Pin Numbers Adjacent to Terminals are for the Metal Package.
Pin Numbers in Parenthesis are for the Dual In-Line Package.



$R_{SC} \cong \frac{0.66v}{I_{SC}}$; $10k\Omega < R1+R2 < 100k\Omega$; $0 \leq C_{REF} \leq 0.1\mu F$

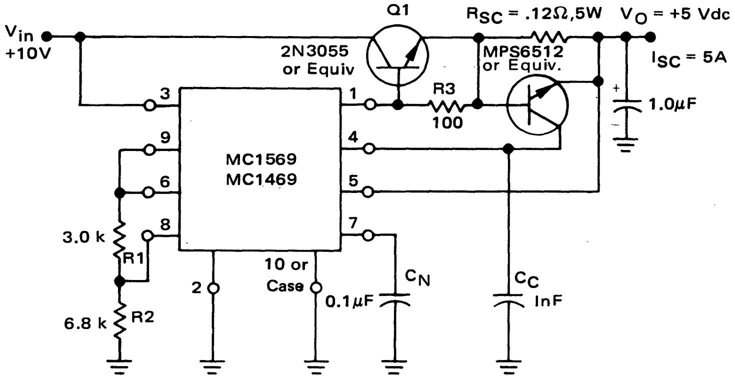
$R2 = \frac{V_{REF}}{V_O}(R1+R2) \cong \frac{7V}{V_O}(R1+R2)$; $R3 = R1 \parallel R2$

$0 < R4 \leq V_{BE_{ON(Q1)}}/5mA$

Selection of Q1 based on considerations of Section 4
See Section 3H for General Design Considerations

Values shown are for a $\boxed{\text{12V, 750mA}}$ regulator using an unheatsinked MC1723CL and a 2N3791 on a 4°C/W heatsink for T_A up to +70°C

Figure 3-6A. MC1723(C) PNP Boost Configuration



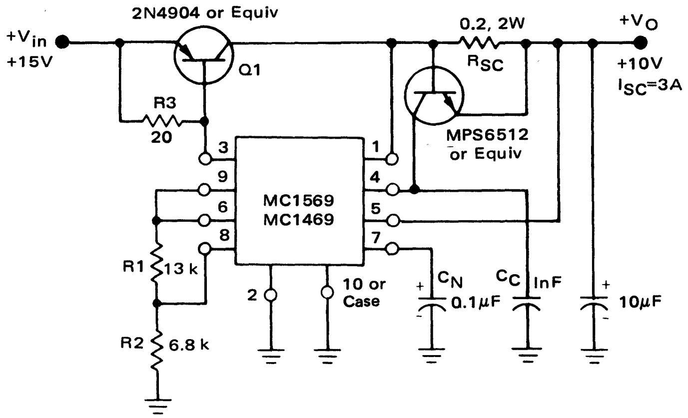
R2=6.8k ; R1=$(\frac{V_O}{V_{REF}}-1)$xR2 ≅ $(2V_O-7)$ kΩ

$C_N=0.1\mu F$; $C_C \geq$ InF ; $R_{SC} \cong \frac{0.6V}{I_{SC}}$; $0<R3\leq\frac{V_{BE_{on}}(Q1)}{I_{CEO}(Q1)}$

Selection of Q1 based on considerations of Section 4
See Section 3H for General Design Considerations

Values shown are for a $\boxed{5V, 5A}$ regulator using an MC1469R regulator on a 20° C/W heatsink with Q1 mounted on a 0.9° C/W heatsink for $T_{A_{MAX}}$=+70° C

Figure 3-7A. MC1569, MC1469 NPN Boost Configuration



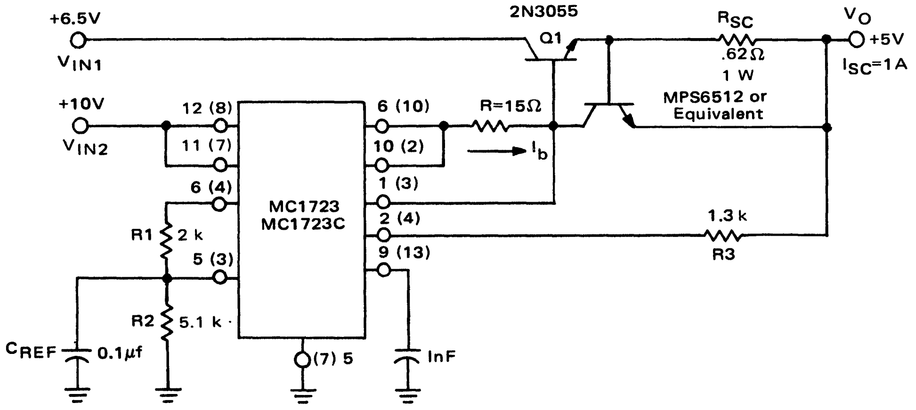
R2=6.8 k ; R1=$(\frac{V_O}{V_{REF}}-1)$xR2≅$(2V_O-7)$kΩ;C_N=0.1μF;$C_C\geq$InF

$R_{SC}\cong\frac{0.6V}{I_{SC}}$;$0<R3\leq\frac{V_{BE_{on}}(Q1)}{12mA}$

Selection of Q1 based on considerations of Section 4
See Section 3H for General Design Considerations

Values shown are for $\boxed{10V, 3A}$ regulator using an MC1469R unheatsinked with Q1 on a 0.5° C/W heatsink for $T_{A_{MAX}}$ = +70°C

Figure 3-8A. MC1569, MC1469 PNP Boost Configuration

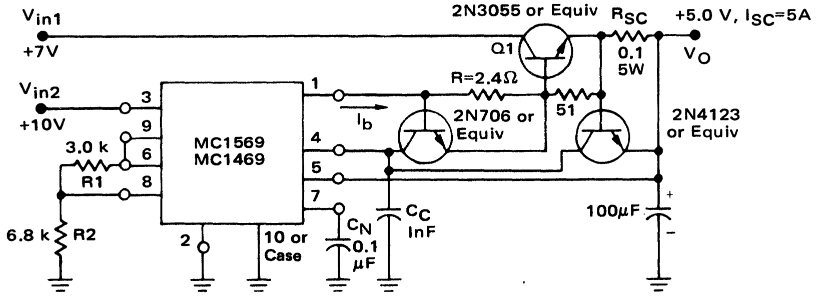


Pin Numbers Adjacent to Terminals are for the Metal Package.
Pin Numbers in Parenthesis are for the Dual In-Line Package.

$R_{SC} \cong \frac{0.6V}{I_{SC}}$; $R \cong \frac{0.66V}{I_{bMAX}}$;10 kΩ<R1+R2<100 kΩ; $R2 = \frac{V_O}{V_{REF}}(R1+R2) \cong \frac{V_O}{7V}(R1+R2)$
$0 \leqslant C_{REF} \leqslant 0.1\mu F$; R3≅R1 ‖ R2; Selection of Q1 based on considerations of Section 4; See Section 3H for general design considerations

Values shown for a $\boxed{5V, 1A}$ regulator using an MC1723CL unheatsinked and a 2N3055 on a 10°C/W heatsink for T_A up to +70°C

Figure 3-9A. MC1723(C) High Efficiency Regulator Configuration



$R2 = 6.8k$; $R1 = (\frac{V_O}{V_{REF}} - 1) \times R2 \cong (2V_O - 7)k\Omega$; $C_N = 0.1\mu F$; $C_C \geqslant InF$, $R_{SC} \cong \frac{0.6V}{I_{SC}}$
$R \approx \frac{0.6V}{I_{bMAX}}$; Selection of Q1 based on considerations of Section 4

See Section 3H for General Design Considerations

Values shown are for a $\boxed{5V, 5A}$ regulator using an MC1469R on a 26°C/W heatsink and Q1 mounted on a 1°C/W heatsink for $T_{A_{MAX}} = +70°C$

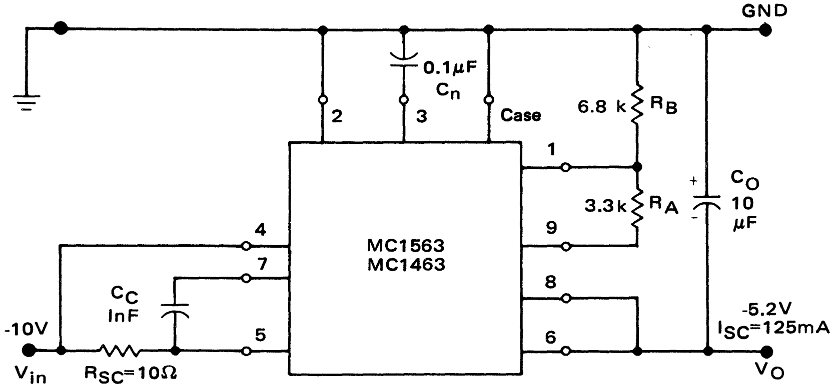
Figure 3-10A. MC1569, MC1469 High Efficiency Regulator Configuration

B. NEGATIVE, ADJUSTABLE OUTPUT IC REGULATOR CONFIGURATIONS

1. Basic Regulator Configurations

MC1563, MC1463

Figure 3-1B illustrates the basic circuit configuration for the MC1563, MC1463 negative regulator IC. Output currents in excess of 500 mA can be obtained depending on input voltage, heatsinking and maximum ambient temperature.



$$R_B=6.8k\Omega ; R_A=R_B\times(\frac{V_O}{V_{REF}}-1)\cong(2|V_O|-7)k\Omega$$

$$C_N=0.1\mu F ; C_C \geqslant InF ; R_{SC} \approx \frac{1.4V}{I_{SC}}$$

See Section 3H for General Design Considerations

Values shown are for a $\boxed{-5.2V, 125mA}$ regulator using a MC1463R at T_A up to +70°C

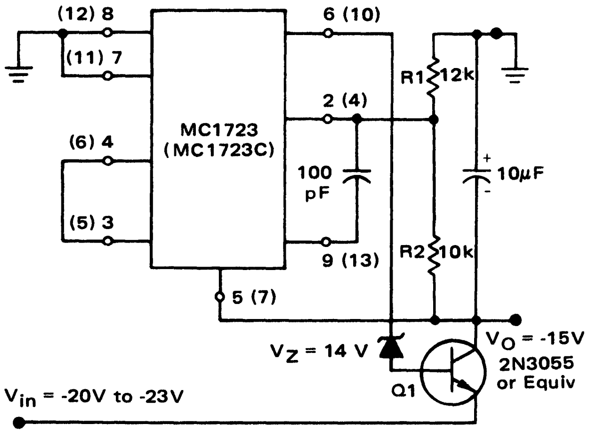
Figure 3-1B. MC1563, MC1463 Basic Regulator Configuration

MC1723(C)

Although a positive regulator, the MC1723(C) can be used in a negative regulator circuit configuration if the superior regulation and performance capabilities of the MC1563 are not needed. This is done by using an external pass element and a zener level shifter as shown in Figure 3-2B. It should be noted that for proper operation, the input supply must not vary over a wide range, since the correct value for V_z depends directly on this voltage. In addition, it should be noted that this circuit will not operate with a shorted output.

2. Output Current Boosting

Figure 3-3B shows a configuration for obtaining increased output current capability from the MC1563, MC1463 regulator by the use of an external series pass element(s).



$|V_O| \geqslant 10V$; $10k\Omega \leqslant R1+R2 \leqslant 100k\Omega$; $R2 = \frac{V_{REF}}{|V_O|}(R1+R2) \cong \frac{7V}{|V_O|}(R1+R2)$

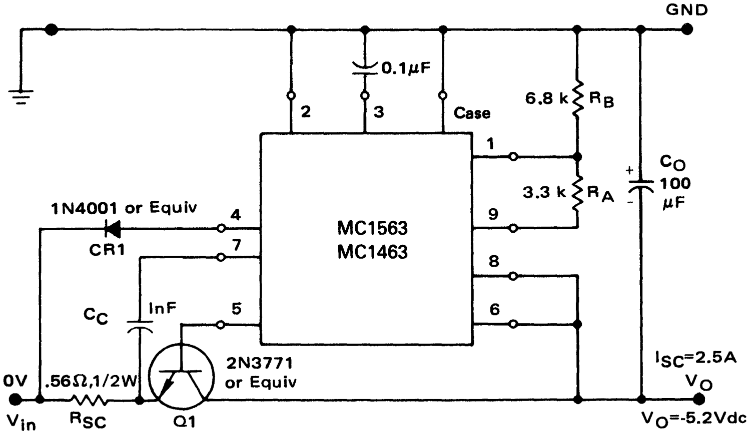
$V_Z \leqslant |V_{IN}| - V_{BE(Q1)} - 3V$; $V_Z \geqslant |V_{IN}| - |V_O| - V_{BE(Q1)} + 6V$

Selection of Q1 based on considerations of Section 4
See Section 3H for General Design Considerations

Warning: Do not short circuit output

Values shown are for a $\boxed{-15V, 750mA}$ regulator using the MC1723CL with Q1 mounted on a 20° C/W heatsink at T_A up to +70° C. (DO NOT SHORT CIRCUIT OUTPUT)

Figure 3-2B. MC1723(C) Negative Regulator Configuration



$R_B = 6.8 k\Omega$; $R_A = R_B \times (\frac{V_O}{V_{REF}} - 1) \cong (2|V_O| - 7) k\Omega$; $C_C \geqslant InF$; $R_{SC} \cong \frac{1.4V}{I_{SC}}$

CRI: add one diode in series with CRI for each additional base emitter junction in composite Q1

Selection of Q1 based on considerations of Section 4
See Section 3H for General Design Considerations

Values shown are for a $\boxed{-5.2V, 2.5A}$ regulator using an MC1463R (unheatsinked) with Q1 mounted on a 1° C/W heatsink for T_A up to +70° C

Figure 3-3B. MC1563, MC1463 Current Boost Configuration

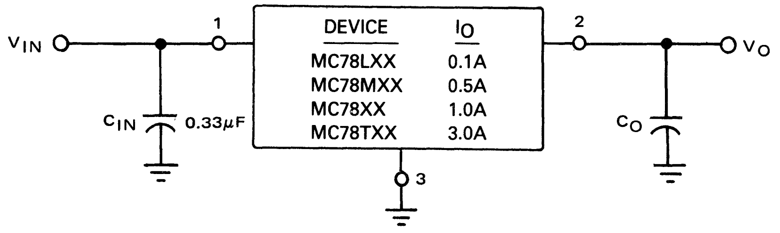
C. POSITIVE, FIXED OUTPUT IC REGULATOR CONFIGURATIONS

1. Basic Regulator Configurations

The basic current configuration for the positive three terminal regulators is shown in Figure 3-1C. Depending on which regulator type is used, this configuration can provide output currents in excess of 3A.

2. Output Current Boosting

Figure 3-2C illustrates a method for obtaining greater output currents with the three terminal positive regulators. Although any of these regulators may be used, usually it is most economical to use the 1 ampere MC7800C in this configuration.



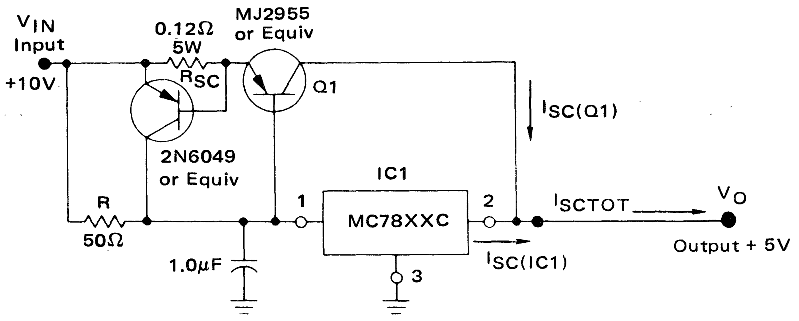
C_{IN}: required if regulator is located more than a few (≈2" to 4") inches away from input supply capacitor; for long input leads to regulator, up to 1μF may be needed for C_{IN}. C_{IN} should be a high frequency type capacitor

C_O: improves transient response

XX: these two digits of the type number indicate nominal output voltage.

See Section 3H for General Design Considerations
See Section 17 for available device output voltages
See Section 15 for heatsinking

Figure 3-1C. Basic Circuit Configuration for the Positive, Fixed Output Three Terminal Regulators



XX = 2 digits of type number indicating voltage. See Section 17 for available device output voltages

R: used to divert IC regulator bias current and determines at what output current level Q1 begins conducting. $0<R\leq\frac{V_{BEON(Q1)}}{I_{BIAS(IC1)}}$; $R_{SC}\approx\frac{0.6V}{I_{SC(Q1)}}$; $I_{SCTOT}=I_{SC(Q1)}+I_{SC(IC1)}$

Selection of Q1 based on considerations of Section 4
See Section 3H for General Design Considerations

Values shown are for a $\boxed{5V, 5A}$ regulator using an MC7805CK on a 2.5°C/W heatsink and Q1 on a 1°C/W heatsink for T_A up to 70°C.

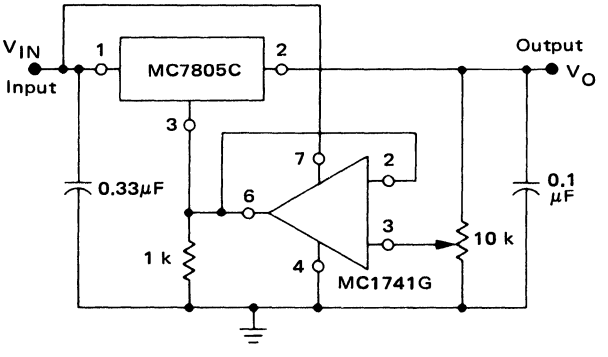
Figure 3-2C. Current Boost Configuration for Positive Three Terminal Regulators

3. Obtaining an Adjustable Output Voltage

With the addition of an op amp, an adjustable output voltage supply can be obtained with the MC7805C. Regulation characteristics of the three terminal regulators are retained in this configuration, shown in Figure 3-3C. If lower output currents are required, an MC78M05C (0.5A) could be used in place of the MC7805C.

4. Current Regulator

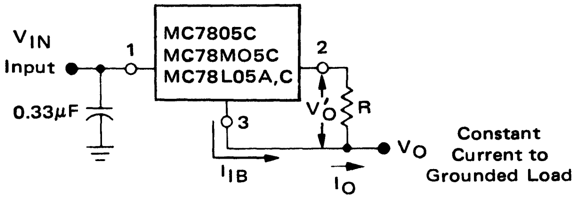
In addition to providing voltage regulation, the three terminal positive regulators can also be used as current regulators to provide a constant current source. Figure 3-4C shows this configuration. The output current can be adjusted to any value from ≈ 8 mA (IQ, the regulator bias current) up to the available output current of the regulator. Five volt regulators should be used to obtain the greatest output voltage compliance range for a given input voltage.



V_O=7V to 33V; V_{IN}-$V_O \geqslant$ 2V; $V_{IN} \geqslant$35V

See Section 3H for General Design Considerations

Figure 3-3C. Adjustable Ouput Voltage Configuration Using a Three Terminal Positive Regulator



$I_O = \frac{V'_O}{R} + I_{IB}$; Current Reg $\Delta I_O = \frac{\Delta V'_O}{R} + \Delta I_{IB}$

$V_O + V'_O + 2V \leqslant V_{IN} \leqslant 35V$

See Section 3H for General Design Considerations

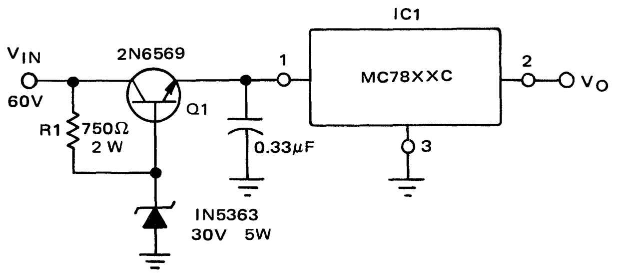
Figure 3-4C. Current Regulator Configuration

5. High Input Voltage

Occasionally, it may be necessary to power a three terminal regulator from a supply voltage greater than $V_{IN(MAX)}$ (35V or 40V). In these cases a preregulator circuit, as shown in Figure 3-5C may be used.

6. High Output Voltage

If output voltages above 24 V are desired, the circuit configuration of Figure 3-6C may be used. Zener diode Z1 sets the output voltage, while Q1, Z2, & D1 assure that the MC7824C does not have more than 30 V across it during short circuit conditions.



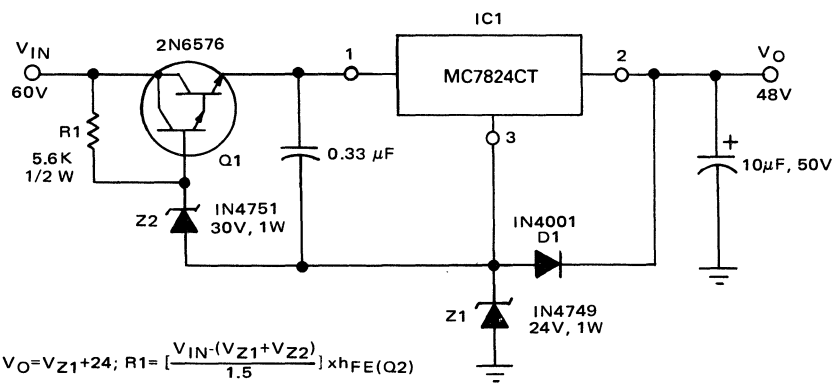
XX=2 digits of type number indicating voltage

$R1=(\frac{V_{IN}-30}{1.5})xh_{FEQ1};\ V_{CEO_{Q1}} \geqslant V_{IN}$

See Section 3H for General Design Considerations

Values shown for V_{IN}=60V; Q1 should be mounted on a 2°C/W heatsink for operation at T_A up to +70°C. IC1 should be appropriately heatsinked for the package type used.

Figure 3-5C. Preregulator for Input Voltages Above V_{INMAX}



$V_O=V_{Z1}+24;\ R1=[\frac{V_{IN}-(V_{Z1}+V_{Z2})}{1.5}]xh_{FE(Q2)}$

See Section 3H for General Design Considerations

Values shown are for a 48V, 1A regulator; Q1 mounted on a 10°C/W heatsink and IC1 mounted on a 2°C/W heatsink for T_A up to +70°C.

Figure 3-6C. High Output Voltage Configuration for Three Terminal Positive Regulators

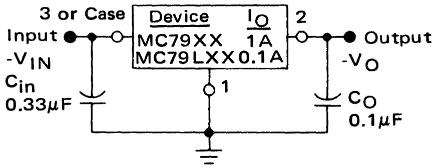
D. NEGATIVE, FIXED OUTPUT IC REGULATOR CONFIGURATIONS

1. Basic Regulator Configurations

Figure 3-1D gives the basic circuit configuration for the MC79XX and MC79LXX three terminal negative regulators.

Output Current Boosting

In order to obtain increased output current capability from the negative three terminal regulators, the current boost configuration of Figure 3-2D may be used. Currents which can be obtained with this configuration are limited only by the capabilities of the external pass transistor(s).



C_{IN}: required if regulator is located more than a few inches (≈2" to 4") away from input supply capacitor; for long input leads to regulator, up to 1μF may be required. C_{IN} should be a high frequency type capacitor.

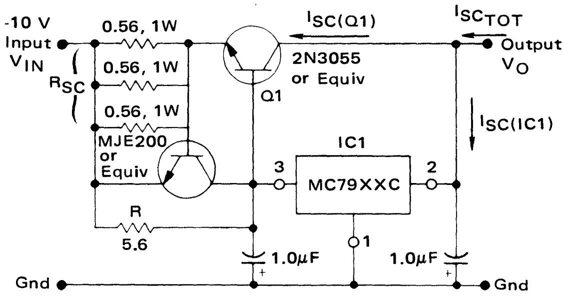
C_O: improves stability and transient response

XX: these two digits of the type number indicate nominal output voltage. See Section 17 for available device output voltages

See Section 3H for General Design Considerations

See Section 15 for heatsinking

Figure 3-1D. Basic Circuit Configuration for the Negative Three Terminal Regulators



XX= 2 digits of type number indicating output voltage. See Section 2 for voltages available

R: used to divert regulator bias current and determines at what output current level Q1 begins conducting. $O < R \leqslant \frac{V_{BEON(Q1)}}{I_{BIAS(IC1)}}$

$I_{SC\,TOT} = I_{SC(Q1)} + I_{SC(IC1)}$

$R_{SC} \approx \frac{0.6V}{I_{SC(Q1)}}$

Selection of Q1 based on considerations of Section 4

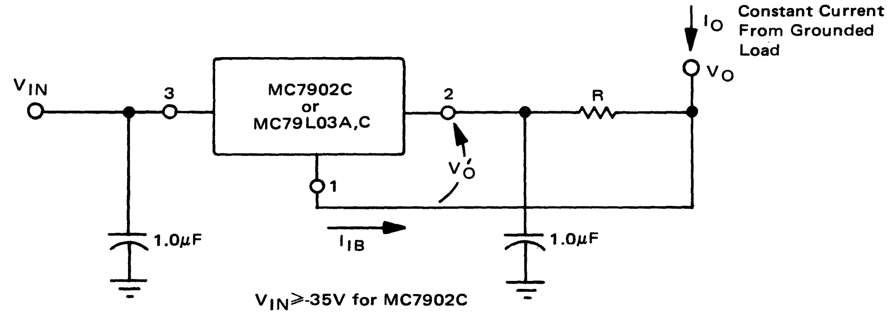
See Section 3H for General Design Considerations

Values shown are for a -5V, 4A regulator using an MC7805CK on a 1.5°C/W heatsink with Q1 mounted on a 1°C/W heatsink for T_A up to 70°C.

Figure 3-2D. Output Current Boost Configuration for Three Terminal Negative Regulators

2. Current Regulator

The three terminal negative regulators may also be used to provide a constant current sink, as shown in Figure 3-3D. In order to obtain the greatest output voltage compliance range at a given input voltage, the MC7902 or MC79L03 should be used in this configuration.



$V_{IN} \geqslant -35V$ for MC7902C
$V_{IN} \geqslant -30V$ for MC79L03C
$V_{IN} \leqslant V_O + V_O - 2V$
$I_O = \frac{V_O'}{R} + I_{IB}$

Current regulation, $\Delta I_O = \frac{\Delta V_O'}{R} + \Delta I_{IB}$

See Section 3H for General Design Considerations

Figure 3-3D. Current Regulator Configuration for the Three Terminal Negative Regulators

E. TRACKING IC REGULATOR CONFIGURATIONS

MC1568, MC1468

Figure 3-1E shows the basic circuit configuration for the MC1568, MC1468 Dual Tracking Regulator. The outputs of this device are internally set at ± 15V. (The output voltage can be externally adjusted with some accompanying loss of temperature performance; see device data sheet, Section 18.) This configuration is capable of providing up to ± 100mA of load current, depending on operating conditions and package style chosen. If greater output currents are desired, the current boost configuration shown in Figure 3-2E can be used.

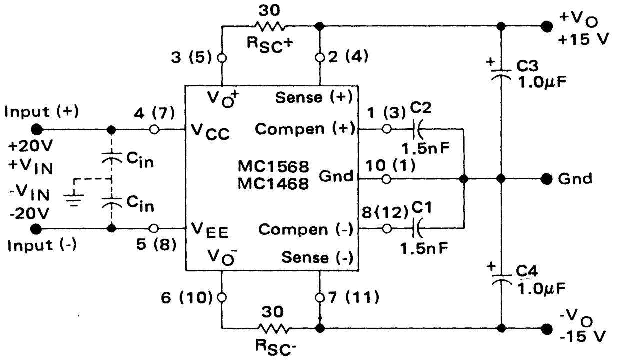
It should be noted that in this configuration, when the positive output of the MC1568, MC1468 drops below approximately 14.5V, e.g. during a short circuit, the negative output will not drop proportionally. Instead, it collapses to ≈ 0V. This can create a latch condition, depending on the type of load.

MC1563/MC1569

If a "true" tracking regulator configuration is desired, the MC1569, MC1469 can be used in conjunction with the MC1563, MC1463 as shown in Figure 3-3E.

In this circuit, the MC1563, MC1463 sets and regulates the negative output voltage, while the MC1569, MC1469 acts as a balancing amplifier to regulate the positive output voltage. The magnitude of the positive output voltage is equal to and tracks the negative output voltage. Since the MC1569's amplifier inputs are at ground potential, its case (or pin 10) is connected to a negative voltage to allow sufficient amplifier common-mode operating range.

Pin numbers adjacent to terminals are for the G and R suffix packages only. Pin numbers in parenthesis are for the L suffix package only. Pin 10 is ground for the G suffix package only. For the R package, the case is ground.



C1 and C2 should be located as close to the device as possible. A 0.1 µF ceramic capacitor (C_{in}) may be required on the input lines if the device is located an appreciable distance from the rectifier filter capacitors. C3 and C4 may be increased to improve load transient response and to reduce the output noise voltage. At low temperature operation, it may be necessary to bypass C4 with a 0.1µF ceramic disc capacitor.

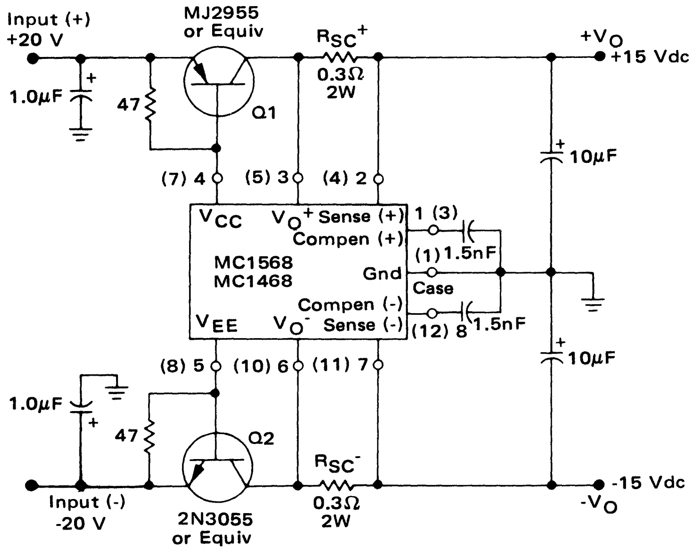
See Section 3H for General Design Considerations

$$R_{SC}^{+} \cong \frac{0.6V}{I_{SC}^{+}} \;;\; R_{SC}^{-} \cong \frac{0.6V}{I_{SC}^{-}}$$

Values shown are for a $\boxed{\pm 15V,\ 20mA}$ regulator using an MC1468R regulator for $T_A \leqslant 75^\circ C$.

Figure 3-1E. MC1568, MC1468 Basic Regulator Configuration

Pin numbers adjacent to terminals are for the G and R suffix packages only. Pin numbers in parenthesis are for the L suffix package only. Pin 10 is ground for the G suffix package only. For the R package, the case is ground.



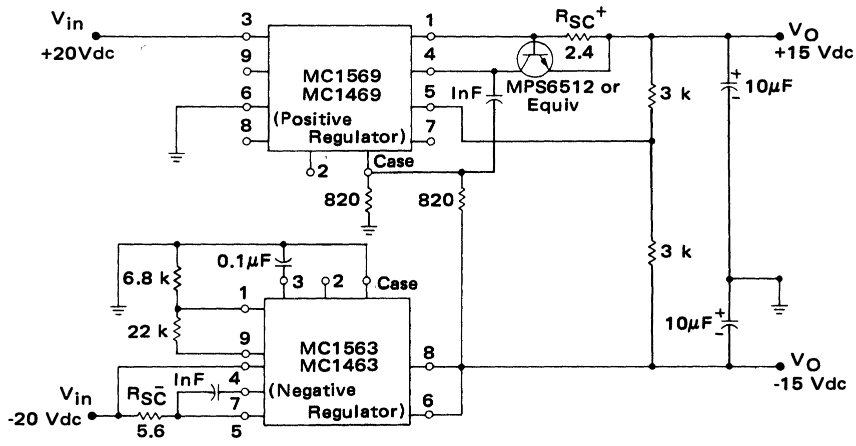
$$R_{SC}^{+} \approx \frac{0.6V}{I_{SC}^{+}} \;;\; R_{SC}^{-} \approx \frac{0.6V}{I_{SC}^{-}}$$

Selection of Q1 based on considerations of Section 4

See Section 3H for General Design Considerations

Values shown are for a $\boxed{\pm 15V\ \pm 2A}$ regulator using an MC1468R on a 2°C/W heatsink with Q1 & Q2 mounted on a 1°C/W heatsink for $T_A \leqslant 70^\circ C$.

Figure 3-2E. MC1568, MC1468 Current Boost Configuration



$$R_{SC}^{+} \approx \frac{0.6V}{I_{SC}^{+}} \;;\; R_{SC}^{-} \approx \frac{1.4V}{I_{SC}^{-}}$$

See Section 3H for General Design Considerations

Values shown are for a $\boxed{\pm 15V, \pm 250mA}$ regulator using an MC1569R and MC1563R mounted on a 3°C/W heatsink for $T_A \leqslant 70°C$

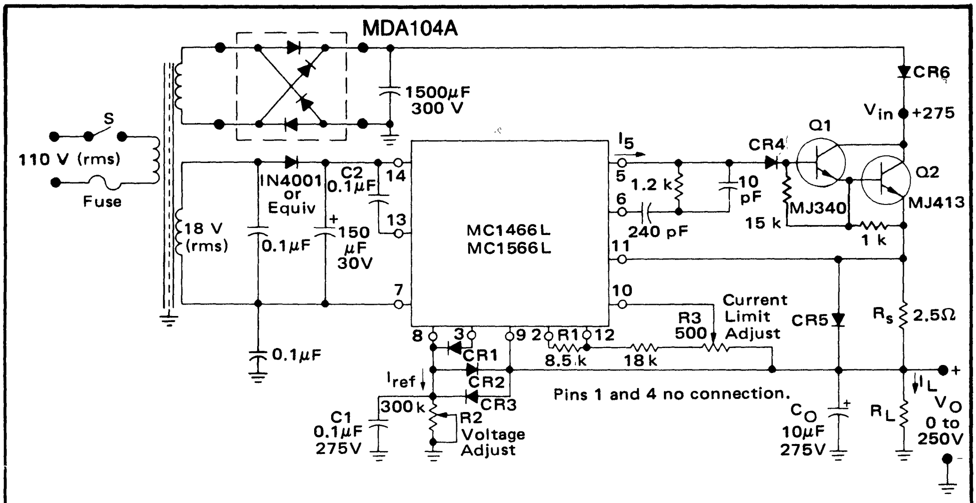
Figure 3-3E. Tracking Regulator Configuration Using the MC1569 & MC1563

F. FLOATING REGULATOR CONFIGURATIONS

If an output voltage exceeding 40V is required, the MC1566L, MC1466L floating regulator can be used, as shown in Figure 3-1F. Although a standard regulator (MC1569, MC1723, etc.) can be used to regulate output voltages above 40V, by the use of level shifting techniques (see Section 3G), the output voltage of these configuration is not adjustable over a wide range, as is the output voltage of the MC1566L. In addition, the MC1566L has several features which are not available elsewhere:

1. Output voltage adjustable to zero volts.
2. Output voltage and current capabilities limited only by choice of external series pass element.
3. Internal current limit amplifier for excellent current regulation and sharp cross-over between constant voltage and constant current regulation modes.

Note that an auxiliary supply is used to power the MC1566, MC1466. This supply must be isolated from the main supply voltage since the MC1566 "floats" on the output voltage. (For a complete description of the MC1566's operation, consult its data sheet, in Section 18.)



DESIGN CONSIDERATIONS

1. Constant Voltage:
   For constant voltage operation, output voltage V_O is given by:
   $V_O = (I_{ref})(R_2)$
   where R2 is the resistance from pin 8 to ground and I_{ref} is the output current of pin 3.
   The recommended value of I_{ref} is 1.0 mAdc. Resistor R1 sets the value of I_{ref}:
   $$I_{ref} = \frac{8.5}{R_1}$$
   where R1 is the resistance between pins 2 and 12.
2. Constant Current:
   For constant current operation:
   (a) Select R_s for a 250 mV drop at the maximum desired regulated output current, I_{max}.
   (b) Adjust potentiometer R3 to set constant current output at desired value between zero and I_{max}.
3. If V_{in} is greater than 20 Vdc, CR2, CR3, and CR4 are necessary to protect the MC1466/MC1566 during short-circuit or transient conditions.
4. In applications where very low output noise is desired, R2 may be bypassed with C1 (0.1μF to 2.0μF). When R2 is bypassed, CR1 is necessary for protection during short-circuit conditions.
5. CR5 is recommended to protect the MC1466/MC1566 from simultaneous pass transistor failure and output short-circuit.
6. The RC network (10 pF, 240 pF, 1.2 k ohms) is used for compensation. The values shown are valid for all applications. However, the 10 pF capacitor may be omitted if f_T of Q1 and Q2 is greater than 0.5 MHz.
7. For remote sense applications, the positive voltage sense terminal (pin 9) is connected to the positive load terminal through a separate sense lead; and the negative sense terminal (the ground side of R2 is connected to the negative load terminal through a separate sense lead.
8. C_O may be selected by using the relationship:
   C_O = (100 μF) $I_{L(max)}$, where $I_{L(max)}$ is the maximum load current in amperes.
9. C2 is necessary for the internal compensation of the MC1466/MC1566.
10. For optimum regulation, current out of pin 5, I_5, should not exceed 0.5 mAdc. Therefore select Q1 and Q2 such that:
    $$\frac{I_{max}}{\beta_1\beta_2} \leq 0.5 \text{ mAdc}$$
    where: I_{max} = maximum short-circuit load current (mAdc)
    $\beta 1$ = minimum beta of Q1
    $\beta 2$ = minimum beta of Q2
    Although Pin 5 will source up to 1.5 mAdc, $I_5 > 0.5$ mAdc will result in a degradation in regulation.
11. CR6 is recommended when $V_O > 150$ Vdc and should be rated such that Peak Inverse Voltage $> V_O$.

Q1 & Q2 selected on the basis of considerations given in Section 3
See Section 3H for General Design Considerations

Values shown are for a 0 to 250V, 100 mA regulator using an MC1466L with Q1 & Q2 mounted on a 1°C/W heatsink for $T_A \leq 70°C$.

Figure 3-1F. MC1566, MC1466 Floating Regulator Configuration

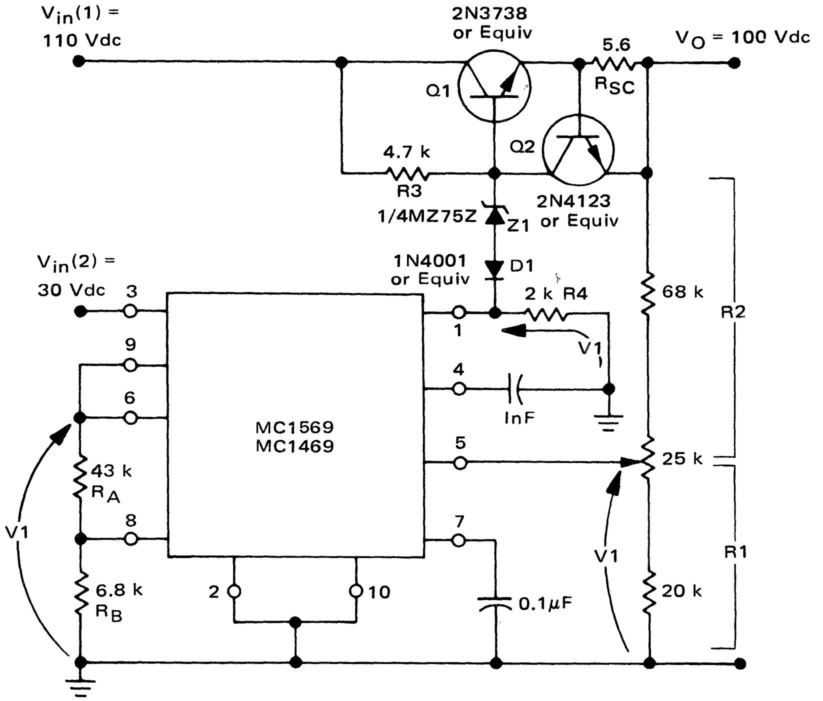
G. SPECIAL REGULATOR CONFIGURATIONS

1. Obtaining Extended Output Voltage Range

As mentioned in the previous section, the output voltage capability of an IC regulator can be increased by using a level shifting technique. In these circuit configurations, the IC regulator is powered from a low voltage supply and its output is shifted by a zener diode to control the base of an external pass element which regulates the high voltage output. A typical configuration is shown in Figure 3-1G for an MC1569, MC1469. This technique can be used with any adjustable output regulator so long as the IC pin voltages, currents, and differentials do not exceed device data sheet specifications.

2. Electronic Shutdown

Occasionally, it is desired that the regulator have an electronic shutdown feature with which the output voltage can be reduced to zero by an external signal.



$R_{SC} \approx \frac{0.6V}{I_{SC}}$; $R3 \approx \frac{h_{FE(Q1)} \times [V_{IN}(1) - V_O]}{I_{SC}}$; $V_{Z1} = V_O - V1$;

$3.5V < V1 < (V_{IN}(2) - 3.0V)$; $R4 < \frac{V1 \times h_{FE(Q1)}}{I_{SC}}$

$R_B = 6.8\,k$; $R_A \cong \left[(2 \times V1) - 7\right] k\Omega$; $V_O \approx V1 \left[1 + \frac{R2}{R1}\right]$

Selection of Q1 based on considerations of Section 4

See Section 3H for General Design Considerations

Values shown are for a 100V, 80mA regulator using an MC1469G on a 30°C/W heatsink with Q1 mounted on a 1°C/W heatsink for $T_A \leqslant 70°C$.

Figure 3-1G. MC1569, MC1469 Output Voltage Boosting Configuration

MC1569 and MC1563

These regulators have internal electronic shutdown circuitry. To activate the shutdown feature, a 1mA minimum, 10mA maximum current is applied to pin 2 of these regulators. This current may be the output of a logic gate or buffer or other external circuitry. This feature can be used to obtain thermal shutdown when the regulator's junction temperature limit is exceeded, as shown in Figures 3-2G and 3-3G; to latch the output when a short circuit occurs, as shown in Figure 3-4G; or to remotely shut down the regulator during standby periods in battery operated equipment.

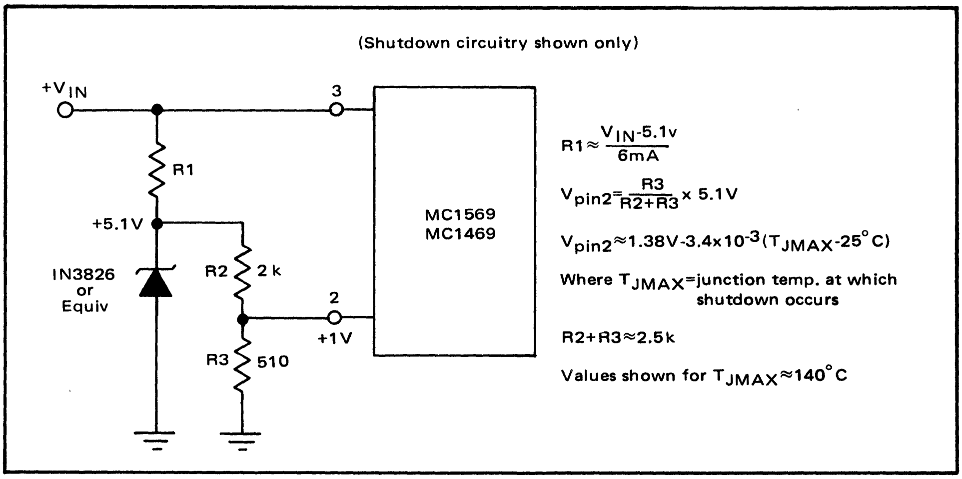


Figure 3-2G. MC1569 Thermal Shutdown Configuration

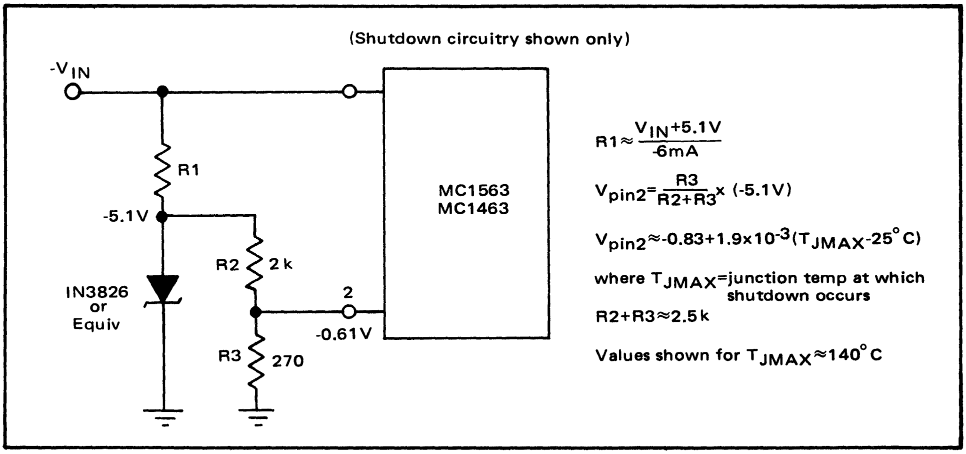
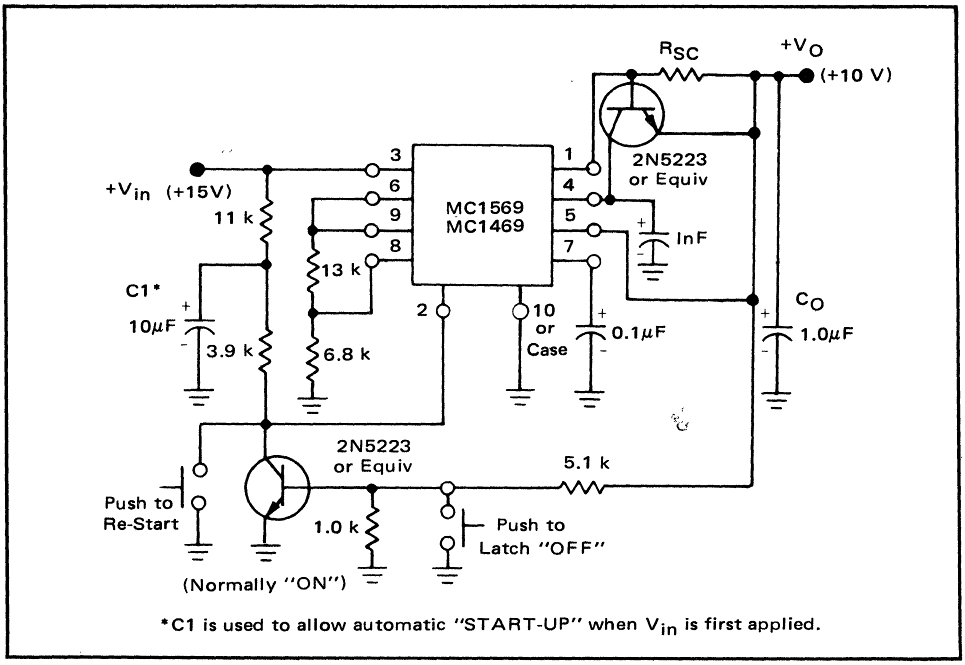


Figure 3-3G. MC1563 Thermal Shutdown Configuration



*C1 is used to allow automatic "START-UP" when V_{in} is first applied.

Figure 3-4G. MC1569 Automatic Latch into Shut-Down When Output is Short Circuited with Manual Reset

MC1723

Although the MC1723 does not have internal electronic shutdown circuitry, this feature can be added externally, as shown in Figure 3-5G. This technique can be used with any externally compensated regulator IC.

H. GENERAL DESIGN CONSIDERATIONS

In addition to the design equations given in the regulator circuit configuration panels of Sections 3A–G, there are a few general design considerations which apply to all regulator circuits. These considerations are given below:

1. Regulator voltages — for any circuit configuration, the worse-case voltages present on each pin of the IC regulator must be within the maximum and/or minimum limits specified on the device data sheets. These limits are instantaneous values, not averages. They include:

a. VIN MIN
b. VIN MAX
c. (VIN − VOUT) MIN
d. VO MIN
e. VO MAX

For example, the voltage between pins 8 and 5 (VIN) of an MC1723CG must never fall below 9.5V, even instantaneously, or the regulator will not function properly.

2. Regulator Power Dissipation, Junction Temperature and Safe Operating Area

The junction temperature, power dissipation output current or safe operating area limits of the IC regulator must never be exceeded.

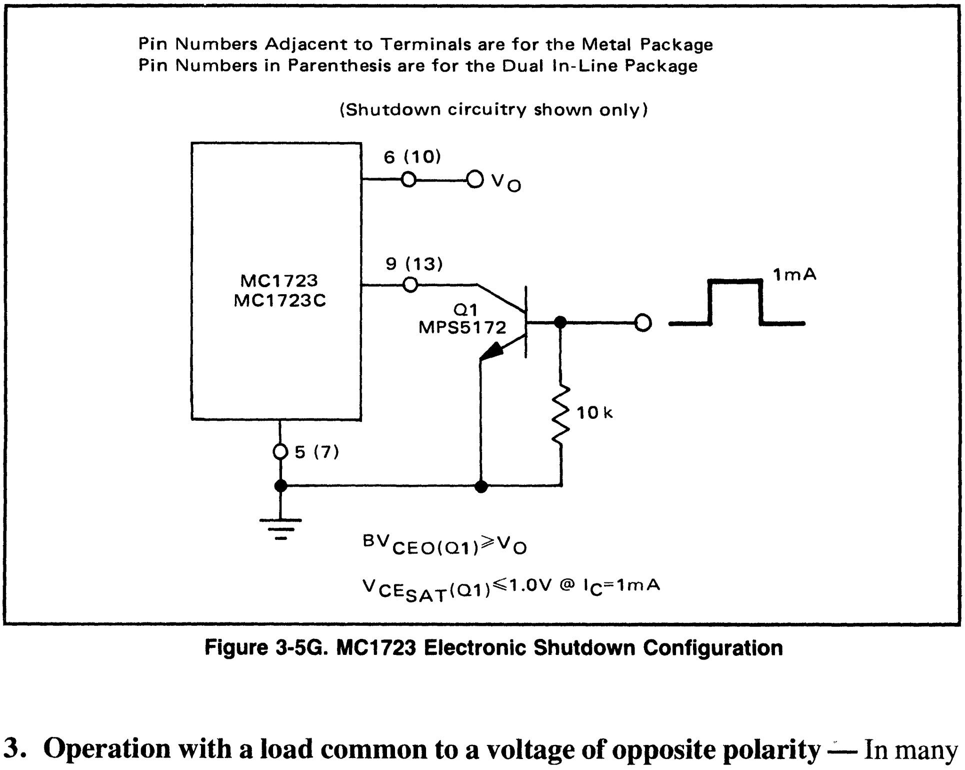


Figure 3-5G. MC1723 Electronic Shutdown Configuration

3. Operation with a load common to a voltage of opposite polarity — In many cases, a regulator powers a load which is not connected to ground but instead is connected to a voltage source of opposite polarity (e.g. op amps, level shifting circuits, etc.). In these cases, a clamp diode should be connected to the regulator output as shown in Figure 3-1H. This protects the regulator, during startup and short-circuit operation, from output polarity reversals.

4. Reverse Bias Protection — Occasionally, there exists the possibility that the input voltage to the regulator can collapse faster than the output voltage. This could occur, for example, if the input supply is "crowbarred" during an output overvoltage condition. If the output voltage is greater $\approx$ 7V, the emitter-base junction of the series pass element (internal or external) could break down and be damaged. To prevent this, a diode shunt can be employed, as shown in Figure 3-2H.

Figure 3-3H shows a three-terminal positive-adjustable regulator with the recommended protection diodes for output voltages in excess of 25 volts, or high-output capacitance values ($C_O > 25\ \mu F$, $C_{Adj} > 10\ \mu F$). Diode D_1 prevents C_O from discharging through the regulator during an input short-circuit. Diode D_2 protects against capacitor C_{Adj} from discharging through the regulator during an output short circuit. The combination of diodes D_1 and D_2 prevents C_{Adj} from discharging through the regulator during an input short circuit.

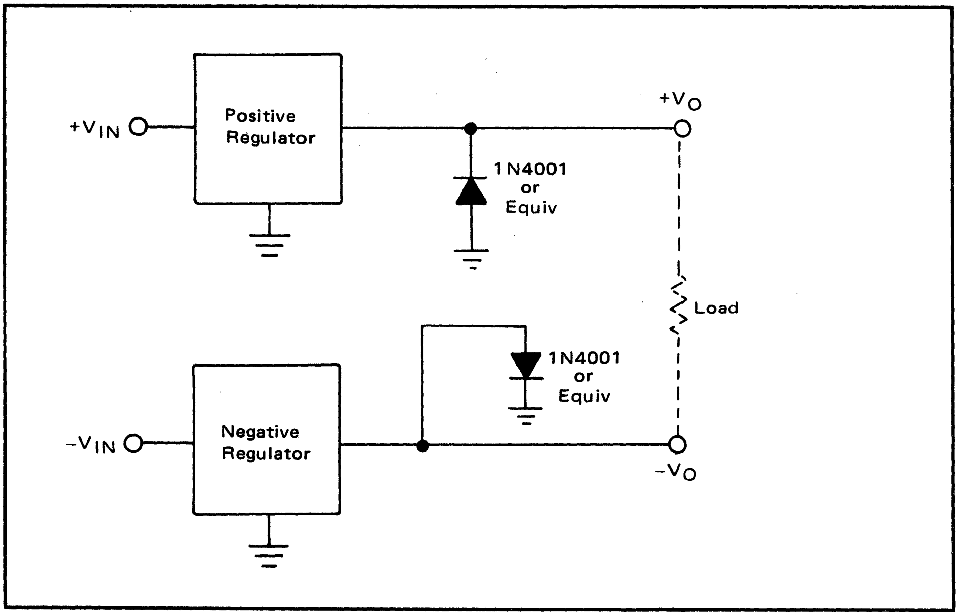


Figure 3-1H. Output Polarity Reversal Protection

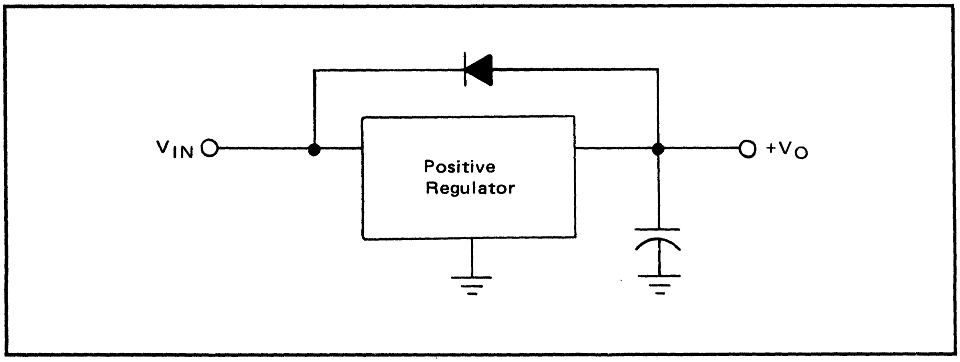


Figure 3-2H. Reverse Bias Protection

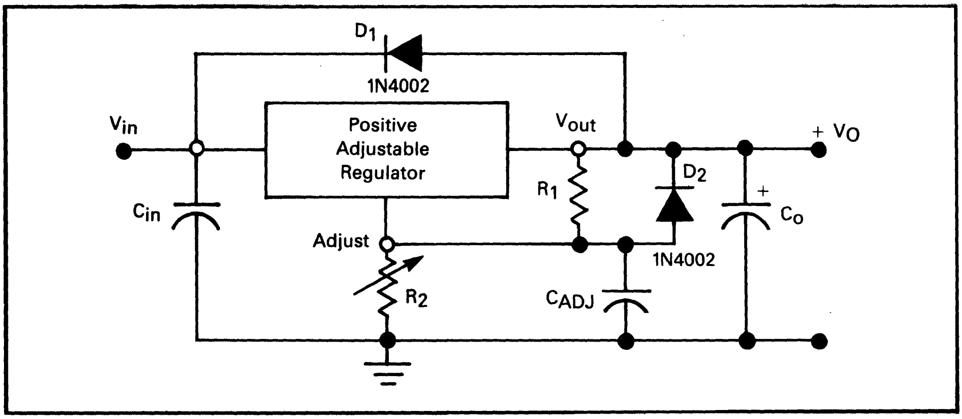


Figure 3-3H. Reverse Bias Protection for Three Terminal Adjustable Regulators

SECTION 4

SERIES PASS ELEMENT CONSIDERATIONS FOR LINEAR REGULATORS

Presently, most monolithic IC voltage regulators that are available have output current capabilities from 100 mA to 3.0 A. If greater current capability is required, or if the IC regulator does not possess sufficient safe-operating-area (SOA), the addition of an external series pass element is necessary.

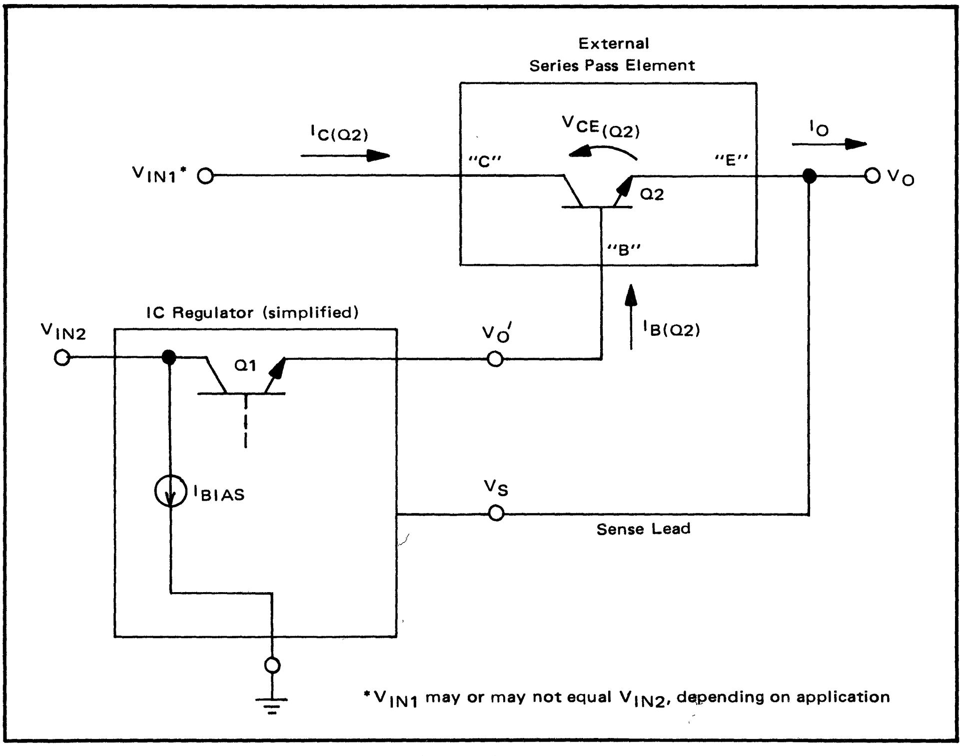
In this section, configurations, specifications and current limit techniques for external series pass elements will be considered. For illustrative purposes, pass elements for only positive regulator types will be discussed. However, the same considerations apply for pass elements used with negative regulators.

A. SERIES PASS ELEMENT CONFIGURATIONS

Using an NPN Type Transistor

If the IC regulator has an external sense lead, an NPN type series pass element may be used, as shown in Figure 4-1A. This pass element could be a single transistor or multiple transistors arranged in darlington and/or paralleled configurations.

In this configuration, the IC regulator supplies the base current (I_B) to the pass element, Q2, which acts as a current amplifier and provides the increased output current (Io) capability.



*V_{IN1} may or may not equal V_{IN2}, depending on application

Figure 4-1A. NPN Type Series Pass Element Configuration

Using a PNP Type Transistor

If the IC regulator does not have an external sense lead, as in the case of the three terminal, fixed output regulators, the configuration of Figure 4-1B can be used. (Regulators which possess an external sense lead may also be used with this configuration.) As before, the PNP type pass element can be a single transistor or multiple transistors.

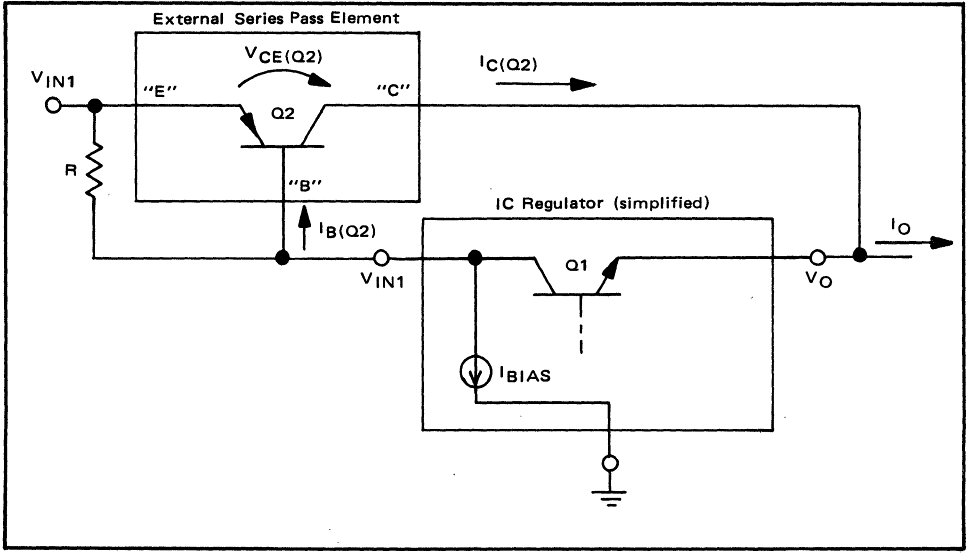


Figure 4-1B. PNP Type Series Pass Element Configuration

This configuration functions in a similar manner to that of Figure 4-1A, in that the regulator supplies base current to pass element. The resistor, R, serves to route the IC regulator bias current, I_{BIAS}, away from the base of Q2. If not included, regulation would be lost at low output currents. The value of R is low enough to prevent Q2 from turning on when I_{BIAS} flows through this resistor, and is given by:

$$0 < R \leqslant \frac{V_{BE\ ON}(Q2)}{I_{BIAS}} \qquad (4.0)$$

B. SERIES PASS ELEMENT SPECIFICATIONS

Independent of which configuration is utilized, the transistor or transistors that compose the pass element must have adequate ratings for I_{CMAX}, V_{CEO}, h_{FE}, power dissipation, and safe-operating-area.

1. I_{CMAX} — for the pass element of Figure 4-1A, I_{CMAX} is given by:

$$I_{CMAX(Q2)} \geqslant I_{OMAX} - I_{BMAX(Q2)} = I_{OMAX} - \frac{I_{CMAX(Q2)}}{h_{FE(Q2)}} \qquad (4.1)$$

$$\geqslant I_{OMAX} \qquad (4.2)$$

For the configuration of Figure 4-1B:

$$I_{CMAX(Q2)} \geqslant I_{OMAX} + I_{BMAX(Q2)} \qquad (4.3)$$

$$\geqslant I_{OMAX} \qquad (4.4)$$

2. V_{CEO} — since $V_{CE(Q2)}$ is equal to $V_{IN1(MAX)}$ when the output is shorted or during start up:

$$V_{CEO(Q2)} \geqslant V_{IN1(MAX)} \tag{4.5}$$

3. hFE — the minimum DC current gain for Q2 in Figures 4-1A and 4-1B is given by:

$$h_{FEMIN(Q2)} \geqslant \frac{I_{CMAX(Q2)}}{I_{BMAX(Q2)}} \ @ \ V_{CE} = (V_{IN1(MIN)} - V_O) \tag{4.6}$$

4. Maximum Power Dissipation, $P_{D(MAX)}$ and Safe-Operating Area (SOA) — for any transistor there are certain combinations of I_C and V_{CE} at which it may safely be operated. When plotted on a graph, whose axes are V_{CE} and I_C, a safe-operating region is formed.

As an example, the safe-operating-area (SOA) curve for the well known 2N3055 NPN silicon power transistor is shown in Figure 4-2. The boundaries of the SOA curve are formed by the I_{CMAX}, power dissipation, second breakdown and V_{CEO} ratings of the transistor. Notice, that the power dissipation and second breakdown ratings are given for a case temperature of +25°C, and must be derated at higher case temperatures. (Derating factors may be found in the transistors' data sheets.) These boundaries must never be exceeded during operation, or destruction of the transistor or transistors which constitute the pass element may result. (In addition, the maximum operating junction temperature must not be exceeded. See Section 15.)

C. CURRENT LIMITING TECHNIQUES

In order to select a transistor or transistors with adequate SOA, the locus of pass element I_C and V_{CE} operating points must be known. This locus of points is determined by the input voltage (V_{IN1}), output voltage (V_O), output current (I_O) and the type of output current limiting technique employed.

In most cases, V_{IN1}, V_O, and the required output current are already known. All that is left to determine is how the chosen current limit scheme affects required pass element SOA.

NOTE: Since the external pass element is merely an extension of the IC regulator, the following discussions apply equally well to IC regulators not using an external pass element.

1. Constant Current Limiting

This method is the simplest to implement and is extensively used, especially at the lower output current levels. The basic curcuit configuration is shown in Figure 4-3A, and operates in the following manner:

As the output current increases, the voltage drop across R_{SC} increases, proportionately. When the output current has increased to the point that the voltage drop across R_{SC} is equal to the base-emitter "on" voltage of Q3 (V_{BEON}(Q3)), Q3 conducts. This diverts base current (I_{DRIVE}) away from Q1, the IC regulator's internal series pass element. Base drive ($I_{B(Q2)}$) of Q2 is therefore reduced and its collector-emitter voltage increases, thereby reducing the output voltage below its regulated value, V_{OUT}. The resulting output voltage-current characteristic is shown in Figure 4-3B. The value of I_{SC} is given by:

$$I_{SC} = \frac{V_{BEON(Q3)}}{R_{SC}} \tag{4.7}$$

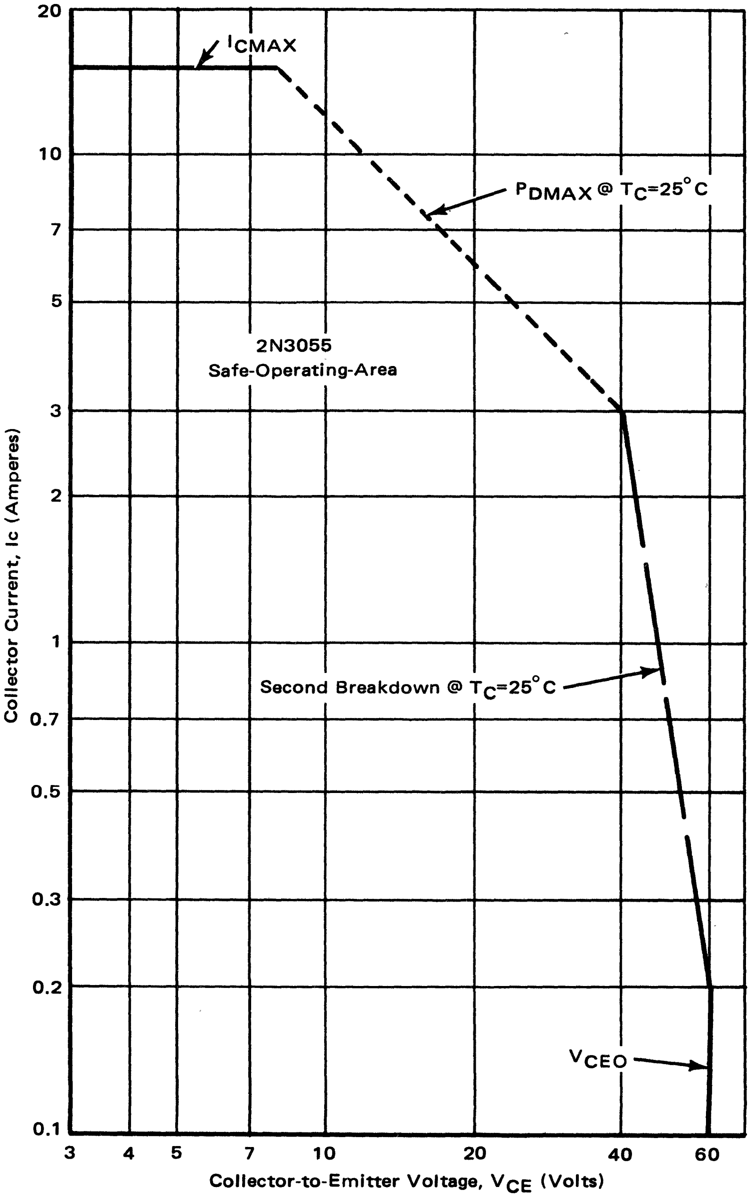


Figure 4-2. 2N3055 Safe-Operating-Area

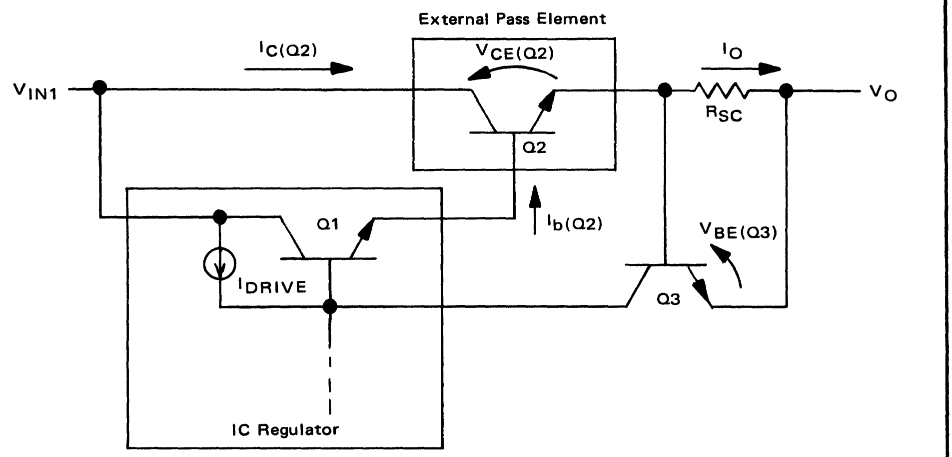


Figure 4-3A. Constant Current Limiting

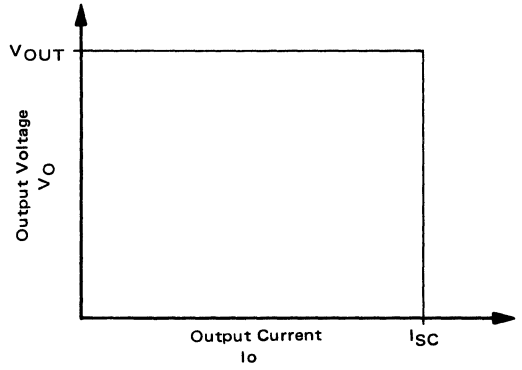


Figure 4-3B. Constant Current Limiting

By using the base of Q1 in the IC regulator as a control point, this configuration has the added benefit of limiting the IC regulator output current ($I_{B(Q2)}$) to $I_{SC}/h_{FE(Q2)}$, as well as limiting the collector current of Q2 to I_{SC}. Of course, access to this point is necessary. Fortunately, it is usually available in the form of a separate pin or as the regulator's compensation terminal.*

The required safe-operating-area for Q2 can be obtained by plotting the V_{CE} and I_C of Q2 given by:

$$V_{CE(Q2)} = V_{IN1} - V_O - I_O R_{SC} \simeq V_{IN1} - V_O \tag{4.8}$$

$$I_{C(Q2)} \simeq I_O \tag{4.9}$$

where $$V_O = V_{OUT} \text{ for } 0 \leqslant I_O \leqslant I_{SC} \tag{4.10}$$

and $$I_O = I_{SC} \text{ for } 0 \leqslant V_O \leqslant V_{OUT} \tag{4.11}$$

---

*The three terminal regulators have internal current limiting and therefore do not provide access to this point. If an external pass element is used with these regulators, constant current limiting can still be accomplished by diverting pass element drive. See Section 3 for circuit techniques.

The resulting plot is shown in Figure 4-4. The transistor chosen for Q2 must have an SOA which encloses this plot, as shown in this Figure.

Note that the greatest demand on the transistors SOA capability occurs when the output of the regulator is short circuited and the pass element must support the full input voltage and short circuit current simultaneously.

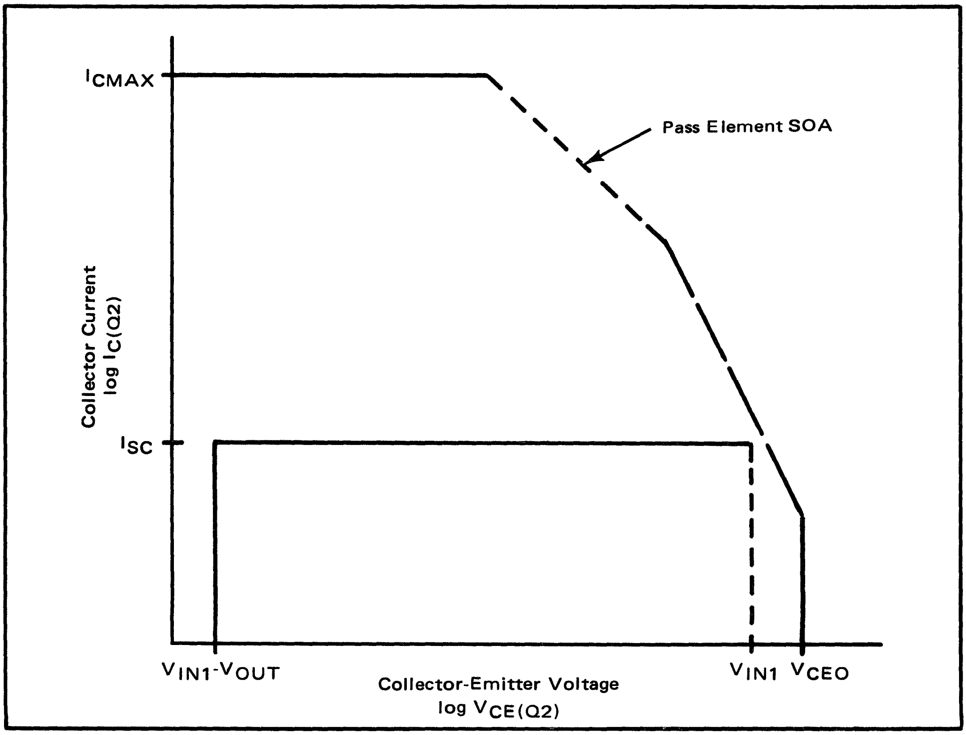


Figure 4-4. Constant Current Limit SOA Requirements

2. Foldback Current Limiting

A disadvantage of the constant current limit technique is that in order to obtain sufficient SOA the pass element must have a much greater collector current capability than is actually needed. If the short circuit current could be reduced, while still allowing full output current to be obtained during normal regulator operation, more efficient utilization of the pass elements SOA capability would result. This can be done by using a ''foldback'' current limiting technique instead of constant current limiting.

The basic circuit configuration for this method is shown in Figure 4-5A. The circuit operates in a manner similar to that of the constant current limiting circuit, in that output current control is obtained by diverting base drive away from Q1 with Q3.

At low output currents, V_A approximately equals V_O and V_{R2} is less than than V_O. Q3 is therefore non-conducting and the output voltage remains constant. As the output current increases, the voltage drop across R_{SC} increases until V_A and V_{R2} are great enough to bias Q3 on. The output current at which this occurs is I_K, the ''knee'' current.

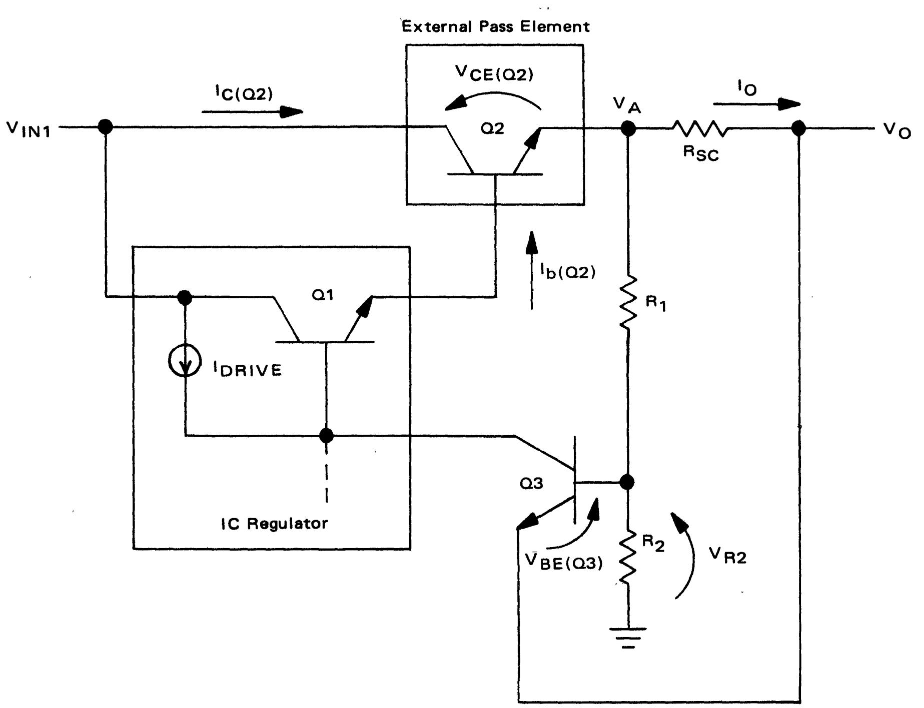


Figure 4-5A. Foldback Current Limiting

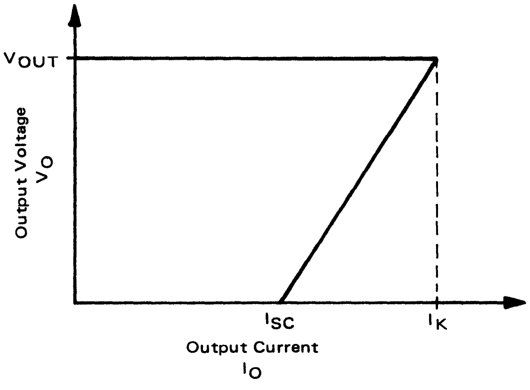


Figure 4-5B. Foldback Current Limiting

The output voltage will now decrease. Less output current is now required to keep V_A and V_{R2} at a level sufficient to bias Q3 on since the voltage at its emitter has the tendency to decrease faster than that at its base. The output current will continue to "foldback" as the output voltage decreases, until an output short circuit current level, I_{SC}, is reached when the output voltage is zero. The resulting output current-voltage characteristic is shown in Figure 4-5B. The values for R1, R2, and R_{SC} (neglecting base current of Q3) are given by:

$$R_{SC} = \frac{V_{OUT}/I_{SC}}{(1 + \frac{V_{OUT}}{V_{BEON(Q3)}}) - \frac{I_K}{I_{SC}}} \qquad (4.12)$$

$$\frac{R2}{R1 + R2} = \frac{V_{BEON(Q3)}}{I_{SC} R_{SC}} \qquad (4.13)$$

and $$R1 + R2 \leqslant \frac{V_{OUT}}{I_{DRIVE}} \qquad (4.14)$$

where V_{OUT} = normal regulator output voltage

I_K = knee current

I_{SC} = short circuit current

I_{DRIVE} = base drive to regulator's internal pass element(s)

A plot of Q2 operating points which result when using this technique are shown in Figure 4-6. Note that the pass element is required to operate with a collector current of only Isc during short circuit conditions, not the full output current, IK. This resuts in a more efficient utilization of the SOA of Q2 allowing the use of a smaller transistor than if constant current limiting were used. Although foldback current limiting allows use of smaller pass element transistors for a given regulator output current than does constant current limiting, it does have a few disadvantages.

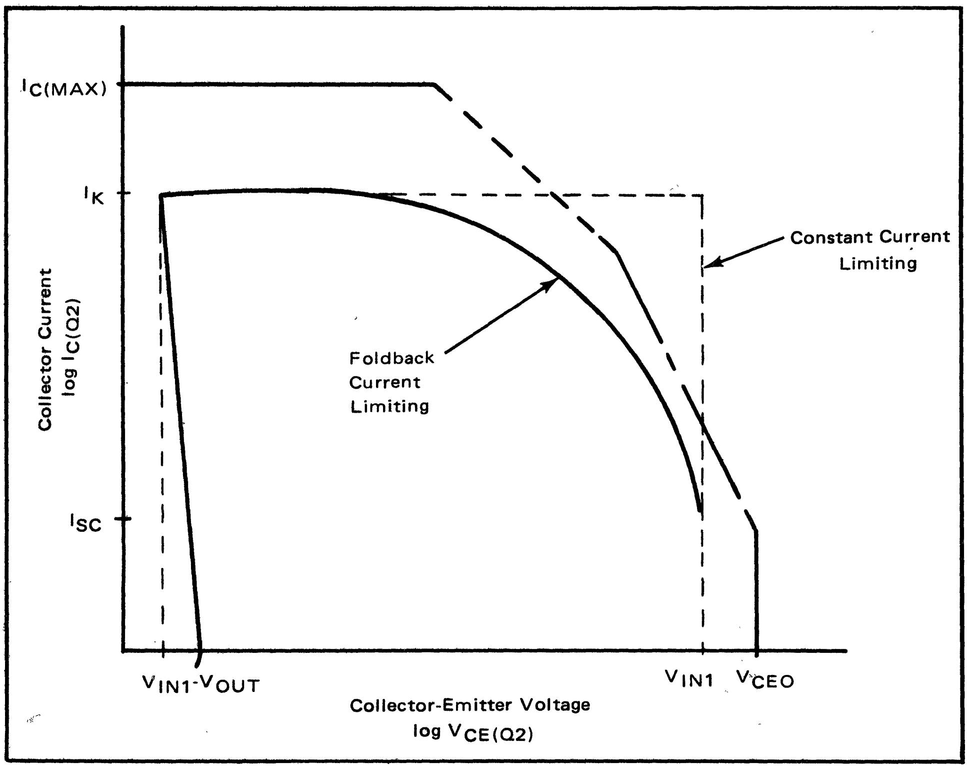


Figure 4-6. Foldback Current Limit SOA Requirements

Referring to Equation (4.12), as the foldback ratio, I_K/I_{SC}, is increased, the required value of R_{SC} increases. This results in a greater input voltage at higher foldback ratios. In addition, it can be seen for Equation (4.12) that there exists an absolute limit to the foldback ratio equal to:

$$\left(\frac{I_K}{I_{SC}}\right)_{MAX} = 1 + \frac{V_{OUT}}{V_{BEON(Q3)}} \text{ for } R_{SC} = \infty \qquad (4.15)$$

For these reasons, foldback ratios greater than 2:1 or 3:1 are not usually practical for the lower output voltage regulators.

D. PARALLELING PASS ELEMENT TRANSISTORS

Occasionally, it will not be possible to obtain a transistor with sufficient safe-operating-area. In these cases it is necessary to parallel two or more transistors. Even if a single transistor with sufficient capability is available, it is possible that paralleling two smaller transistors is more economical.

In order to insure that the collector currents of the paralleled transistors are approximately equal, the configuration of Figure 4-7 can be used. Emitter ballasting resistors are used to force collector current sharing between Q1 and Q2. The collector current mismatch can be determined by considering the following:

From Figure 4-7,

$$V_{BE1} + V_1 = V_{BE2} + V_2 \qquad (4.16)$$

and $\Delta V_{BE} = \Delta V$ (4.17)

where $\Delta V_{BE} = V_{BE1} - V_{BE2}$

and $\Delta V = V_2 - V_1$

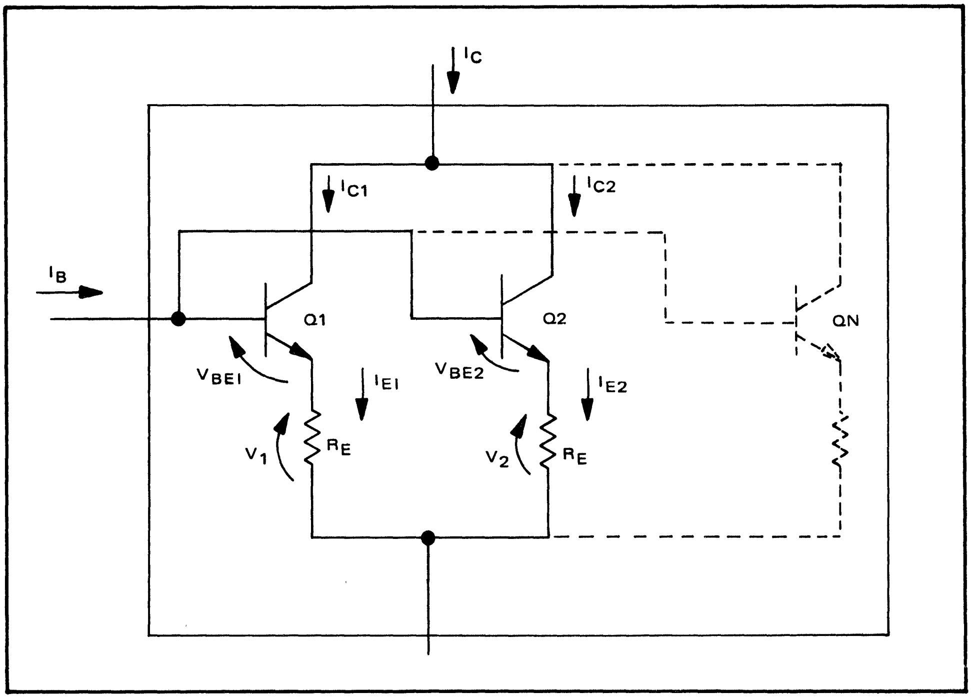


Figure 4-7. Paralleling Pass Element Transistors

Assuming $I_{E1} \simeq I_{C1}$ and $I_{E2} \simeq I_{C2}$, the collector current mismatch is given by,

$$\frac{I_{C2} - I_{C1}}{I_{C2}} = \frac{\left(\frac{V_2}{R_E}\right) - \left(\frac{V_1}{R_E}\right)}{\left(\frac{V_2}{R_E}\right)} = \frac{V_2 - V_1}{V_2} = \frac{\Delta V}{V_2} \qquad (4.18)$$

$$= \frac{\Delta V_{BE}}{V_2} \qquad (4.19)$$

and,

$$\text{percent collector current mismatch} = \frac{\Delta V_{BE}}{V_2} \times 100\% \qquad (4.20)$$

From Equation (4.20), the collector current mismatch is dependent on ΔV_{BE} and V_2. Since ΔV_{BE} is usually acceptable, V_2 should be 1.0 V to 0.5 V, respectively. R_E is therefore given by:

$$R_E = \frac{0.5 \text{ to } 1.0 \text{ V}}{I_{C1}} = \frac{0.5 \text{ V to } 1.0 \text{ V}}{I_{C2}} = \frac{0.5 \text{ V to } 1.0 \text{ V}}{I_C/2} \qquad (4.21)$$

E. TRANSISTOR SELECTION GUIDE

As an aid in selecting an appropriate series pass element, the following selection guide has been included.

Device and Polarity NPN	 PNP	V_{CEO} Volts Min	h_{FE} Min/Max	I_C Amps	$V_{ce(sat)}$ Volts Max	I_C Amps	f_T MHz Min	P_D Watts Max	Case
0.3 Amp									
MJE3440		250	40/160	0.02	0.5	0.05	15	15	77
MJE3439		350	40/160	0.02	0.5	0.05	15	15	77
0.5 Amp									
2N5655		250	30/250	0.1	1.0	0.1	10	20	77
2N5656		300	30/250	0.1	1.0	0.1	10	20	77
MJE340	MJE350	300	30/240	0.05				20	77
2N5657		350	30/250	0.1	1.0	0.1	10	20	77
1.0 Amp									
TIP29	TIP30	40	15/75	1.0	0.7	1.0	3.0	30	221A
2N4921	2N4918	40	30/150	0.5	0.6	1.0	3.0	30	77
TIP29A	TIP30A	60	15/75	1.0	0.7	1.0	3.0	30	221A
2N4922	2N4919	60	30/150	0.5	0.6	1.0	3.0	30	77
TIP29B	TIP30B	80	15/75	1.0	0.7	1.0	3.0	30	221A
2N4923	2N4920	80	30/150	0.5	0.6	1.0	3.0	30	77
TIP29C	TIP30C	100	15/75	1.0	0.7	1.0	3.0	30	221A
2N3738	2N6424	225	40/200	0.1	2.5	0.25	10	20	80
TIP47		250	30/150	0.3	1.0	1.0	10	40	221A
TIP48		300	30/150	0.3	1.0	1.0	10	40	221A
2N3739	2N6425	300	40/200	0.1	2.5	0.25	10	20	80
TIP49		350	30/150	0.3	1.0	1.0	10	40	221A
2.0 Amp									
2N3583	2N6420	175	40/200	0.5	5.0	1.0	10	35	80
2N3584	2N6421	250	8/80	1.0	0.75	1.0	10	35	80
2N3585	2N6422	300	8/80	1.0	0.75	1.0	10	35	80
2N4240	2N6423	300	30/150	0.75	1.0	0.75	15	35	80
2.5 Amps									
BU205		750	2/	2.5	5.0	2.5	7.5	10	01

PREFERRED SILICON POWER TRANSISTORS (continued)

Device and Polarity NPN	PNP	V_{CEO} Volts Min	h_{FE} Min/Max	I_C Amps	$V_{ce(sat)}$ Volts Max	I_C Amps	f_T MHz Min	P_D Watts Max	Case
3.0 Amps									
MJE520		30	25/	1.0				25	77
MJE31	MJE32	40	25/	1.0	1.2	3.0	3.0	40	77
	2N3867	40	40/200	1.5	0.75	1.5	60	6.0	31
	2N3868	60	30/150	1.5	0.75	1.5	60	6.0	31
MJE31A	MJE32A	60	25/	1.0	1.2	3.0	3.0	40	77
MJE31B	MJE32B	80	25/	1.0	1.2	3.0	3.0	40	77
MJE181	MJE171	80	50/250	0.1	0.9	1.5	50	1.5	77
MJE31C	MJE32C	100	25/	1.0	1.2	3.0	3.0	40	77
3.5 Amp									
2N3902		400	30/90	1.0	0.8	1.0	2.8	100	01
4.0 Amp									
2N5190	2N5193	40	25/100	1.5	0.6	1.5	2.0	40	77
2N6037	2N6034	40	750/15K	2.0	2.0	2.0	1.0	40	77
MJE3300	MJE3310	40	1000/	1.0	1.5	1.5	20	15	77
2N6121	2N6124	45	25/100	1.5	0.6	1.5	2.5	40	221A
2N3054A	2N6049	55	25/250	0.5	1.0	0.5	3.0	75	80
2N6122	2N6125	60	25/100	1.5	0.6	1.5	2.5	40	221A
2N6413	2N6415	60	40/250	0.2	2.5	4.0	50	15	77
2N5191	2N5194	60	25/100	1.5	0.6	1.5	2.0	40	77
	2N3740	60	30/100	0.25	0.6	1.0	3.0	25	80
2N6294	2N6296	60	750/18K	2.0	2.0	4.0	50	80	
2N6038	2N6035	60	750/15K	2.0	2.0	2.0	1.0	40	77
MJE3301	MJE3311	60	1000	1.0	1.5	1.5	20	15	77
MJE800	MJE700	60	750/	1.5	2.5	1.5	1.0	40	77
2N6123	2N6126	80	20/80	1.5	0.6	1.5	2.5	40	221A
MJE3302	MJE3312	80	1000/	1.0	1.5	1.5	20	15	77
2N5192	2N5195	80	20/80	1.5	0.6	1.5	2.0	40	77
	2N3741	80	30/100	0.25	0.6	1.0	3.0	25	80
2N6295	2N6297	80	750/18K	2.0	2.0	2.0	4.0	50	80
2N6039	2N6036	80	750/15K	2.0	2.0	2.0	1.0	40	77
5.0 Amp									
MJE200	MJE210	40	45/180	2.0	0.75	2.0	65	15	77
2N4232A	2N6313	60	25/100	1.5	0.7	1.5	4.0	75	80
MJE1100	MJE1090	60	750/	3.0	2.5	3.0		70	90
2N4233A	2N6314	80	25/100	1.5	0.7	1.5	4.0	75	80
2N6233		225	25/125	1.0	0.5	1.0	20	50	80
2N6497		250	10/75	2.5	1.0	2.5	5.0	80	221A
MJE51T		250	5/	5.0	2.0	5.0	2.5	80	221A
2N6234		275	25/125	1.0	0.5	1.0	20	50	80
2N6498		300	10/75	2.5	1.25	2.5	80	5.0	221A
MJE52T		300	5/	5.0	2.0	5.0	2.5	80	221A
2N6235		325	25/125	1.0	0.5	1.0	20	50	80
MJ3030		325			2.0	3.0		125	01
2N6499		350	10/75	2.5	1.5	2.5	5.0	80	221A
MJE53T		350	5/	5.0	2.0	5.0	2.5	80	221A
BU208		700	2.25/	4.5	5.0	4.5	4.0	1.25	01

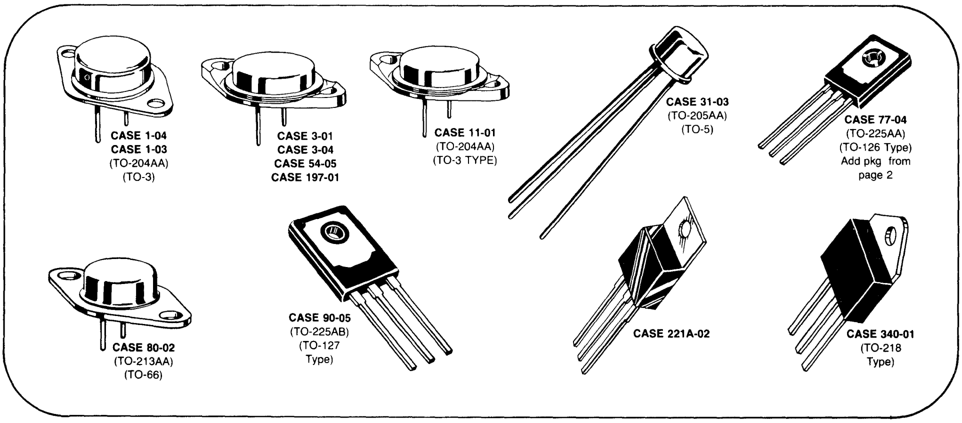
PREFERRED SILICON POWER TRANSISTORS (continued)

Device and Polarity NPN	PNP	V_{CEO} Volts Min	h_{FE} Min/Max	I_C Amps	$V_{ce(sat)}$ Volts Max	I_C Amps	f_T MHz Min	P_D Watts Max	Case
6.0 Amp									
TIP41	TIP42	40	15/75	3.0	1.5	6.0	3.0	2.0	221A
TIP41A	TIP42A	60	15/75	3.0	1.5	6.0	3.0	2.0	221A
TIP41B	TIP42B	80	15/75	3.0	1.5	6.0	3.0	2.0	221A
TIP41C	TIP42C	100	15/75	3.0	1.5	6.0	3.0	2.0	221A
2N5758	2N6226	100	25/100	3.0	1.0	3.0	1.0	150	11
2N5959	2N6227	120	20/80	3.0	1.0	3.0	1.0	150	11
2N5760	2N6228	140	15/60	3.0	1.0	3.0	1.0	150	11
8.0 Amp									
2N6300	2N6298	60	750/18K	4.0	2.0	4.0	4.0	75	80
2N6055	2N6053	60	750/18K	4.0	2.0	4.0	4.0	100	11
2N6043	2N6040	60	1K/10K	4.0	2.0	4.0	4.0	75	221A
MJ1000	MJ900	60	1000/	3.0	2.0	3.0		90	11
2N6301	2N6299	80	750/18K	4.0	2.0	4.0	4.0	75	80
2N6056	2N6054	80	750/18K	4.0	2.0	4.0	4.0	100	11
2N6044	2N6041	80	1K/10K	4.0	2.0	4.0	4.0	75	221A
2N6045	2N6042	100	1K/10K	3.0	2.0	3.0	4.0	75	221A
2N6306		250	15/75	3.0	0.8	3.0	5.0	125	01
2N6307		300	15/75	3.0	1.0	3.0	5.0	125	01
2N6308		350	12/60	3.0	1.5	3.0	5.0	125	01
10.0 Amp									
2N6383	2N6648	40	1K/20K	5.0	2.0	5.0	20	100	11
2N6384	2N6649	60	1K/20K	5.0	2.0	5.0	20	100	11
MJE3055	MJE2955	60	20/100	4.0	1.1	4.0	2.0	90	90
MJE3055T	MJE2955T	60	20/100	4.0	1.1	4.0	2.0	90	221A
MJE4340	MJE4350	100	50/	10.0	0.5	5.0	1.0	125	340
MJE4341	MJE4351	120	50/	10.0	0.5	5.0	1.0	125	340
MJE4342	MJE4352	140	50/	10.0	0.5	5.0	1.0	125	340
MJE4343	MJE4353	160	50/	10.0	0.5	5.0	1.0	125	340
2N5877	2N5875	60	20/100	4.0	1.0	5.0	4.0	150	11
2N3715	2N3791	60	50/150	1.0	0.8	5.0	4.0	150	11
2N5878	2N5876	80	20/100	4.0	1.0	5.0	4.0	150	11
2N6385	2N6650	80	1K/20K	5.0	2.0	5.0	20	100	11
2N3716	2N3792	80	50/150	1.0	0.8	5.0	4.0	150	11
2N5632	2N6229	100	25/100	5.0	1.0	7.5	1.0	150	11
2N5633	2N6230	120	20/80	5.0	1.0	7.5	1.0	150	11
2N5634	2N6231	140	15/60	5.0	1.0	7.5	1.0	150	11
MJ413		325	20/80	0.5	0.8	0.5	2.5	125	11
MJ423		325	30/90	1.0	0.8	1.0	2.5	125	11
12.0 Amp									
2N6569		40	15/200	4.0	1.5	4.0	1.5	100	11
2N5989	2N5986	40	20/120	6.0	0.7	6.0	2.0	100	90
2N5990	2N5987	60	20/120	6.0	0.7	6.0	2.0	100	90
2N6057	2N6050	60	750/18K	6.0	2.0	6.0	4.0	150	01
2N5991	2N5988	80	20/120	6.0	0.7	6.0	2.0	100	90
2N6058	2N6051	80	750/18K	6.0	2.0	6.0	4.0	150	01
2N6059	2N6052	100	750/18K	6.0	2.0	6.0	4.0	150	01

PREFERRED SILICON POWER TRANSISTORS (continued)

Device and Polarity NPN	PNP	V_{CEO} Volts Min	h_{FE} Min/Max	I_C Amps	$V_{ce(sat)}$ Volts Max	I_C Amps	f_T MHz Min	P_D Watts Max	Case
15.0 Amp									
2N6486	2N6489	40	20/150	5.0	1.3	5.0	5.0	75	221A
2N6487	2N6490	60	20/150	5.0	1.3	5.0	5.0	75	221A
2N3055	MJ2955	60	20/70	4.0	1.1	4.0	2.5	115	11
2N5881	2N5879	60	20/100	6.0	1.0	7.0	4.0	160	11
2N6576		60	500/5K	10.0	4.0	15		120	11
2N6488	2N6491	80	20/150	5.0	1.3	5.0	5.0	75	221A
2N5882	2N5880	80	20/100	6.0	1.0	7.0	4.0	160	11
2N6577		90	500/5K	10.0	4.0	15		120	11
2N6578		120	500/5K	10.0	4.0	15		120	11
2N6249		200	10/50	10.0	1.5	10	2.5	175	01
2N6250		275	8/50	10.0	1.5	10	2.5	175	01
2N6251		350	6/50	10.0	1.5	10	2.5	175	01
16.0 Amp									
2N5629	2N6029	100	25/100	8.0	1.0	10	1.0	200	11
2N5630	2N6030	120	20/80	8.0	1.0	10	1.0	200	11
2N5631	2N6031	140	15/60	8.0	1.0	10	1.0	200	11
20.0 Amp									
2N6282	2N6285	60	750/18K	10.0	2.0	10	4.0	160	01
2N5303	2N5745	80	15/160	10.0	1.0	10	2.0	200	11
2N6283	2N6286	80	750/18K	10.0	2.0	10	4.0	160	01
2N6284	2N6287	100	750/18K	10.0	2.0	10	4.0	160	01
25.0 Amp									
2N5885	2N5883	60	20/100	10.0	1.0	15	4.0	200	11
2N5886	2N5884	80	20/100	10.0	1.0	15	4.0	200	11
2N6338		100	30/120	10.0	1.0	10	40	200	01
2N6339		120	30/120	10.0	1.0	10	40	200	01
2N6340		140	30/120	10.0	1.0	10	40	200	01
2N6341		150	30/120	10.0	1.0	10	40	200	01
30.0 Amp									
2N5301	2N4398	40	15/60	15.0	0.75	10	2.0	200	11
2N5302	2N4399	60	15/60	15.0	0.75	10	2.0	200	11
MJ802	MJ4502	90	25/100	7.5	0.8	7.5	2.0	200	11
50.0 Amp									
2N5685	2N5683	60	15/60	25.0	1.0	25	2.0	300	197
2N5686	2N5684	80	15/60	25.0	1.0	25	2.0	300	197
2N6274		100	30/120	20.0	1.0	20	30	250	197
2N6275		120	30/120	20.0	1.0	20	30	250	197
2N6276		140	30/120	20.0	1.0	20	30	250	197
2N6277		150	20/120	20.0	1.0	20	30	250	197

SILICON POWER DEVICE PACKAGES



SECTION 5
LINEAR REGULATOR CONSTRUCTION AND LAYOUT

An important, and often neglected, aspect of the total regulator circuit design is the actual layout and component placement of the circuit. In order to obtain excellent transient response performance, high frequency transistors are used in modern integrated circuit voltage regulators. Proper attention to circuit layout is therefore necessary in order to prevent regulator instability or oscillations, or degraded performance.

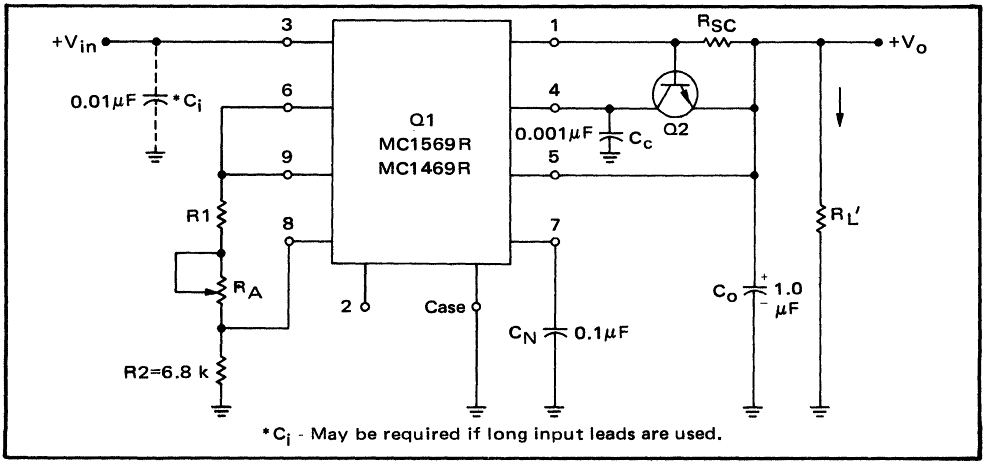
In this section, guidelines will be given on proper regulator layout and placement of circuit components. In addition, topics such as remote voltage sensing and semiconductor mounting techniques will also be considered.

1. General Layout and Component Placement Considerations

As mentioned previously, modern integrated circuit regulators are necessarily high bandwidth devices in order to obtain good transient response characteristics. To insure stable closed loop operation, all these devices are frequency compensated, either internally or externally. This compensation can easily be upset by unwanted stray circuit capacitances and lead inductances, resulting in spurious oscillations. Therefore, it is important that the circuit lead lengths be short and the layout as tight as possible. Particular attention should be paid to locating the compensation and bypass capacitors as close to the IC as possible. Lead lengths associated with the external pass element(s), if used, should also be minimized.

Often overlooked is the stray inductance associated with the input leads to the regulator circuit. If the lead length from the input supply filter capacitor to the regulator input is more than a couple of inches, a 0.01-1.0μF high frequency type capacitor (tantalum, ceramic, etc.) should be used to bypass the supply leads close to the regulator input pins.

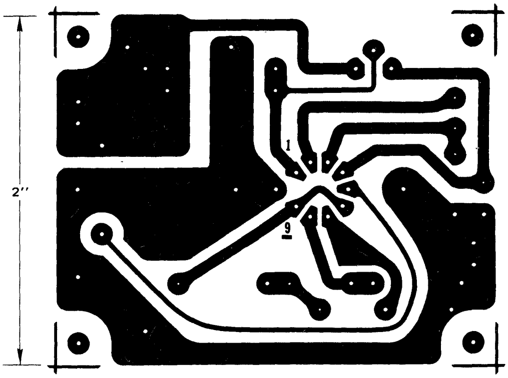
A typical good circuit layout is shown in Figure 5-1 for an MC1569R regulator circuit configuration.



*Ci - May be required if long input leads are used.

Figure 5-1. Typical Regulator Circuit Layout

Typical Printed Circuit Board Layout



Location of Components

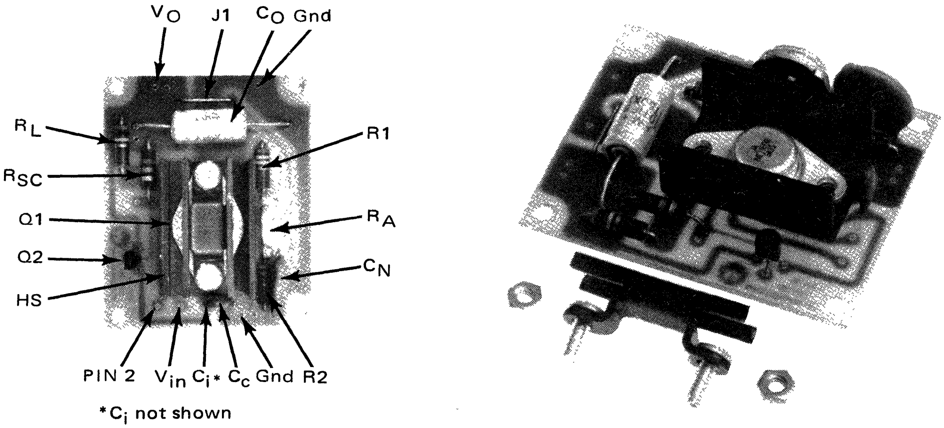


Figure 5-1. Typical Regulator Circuit Layout (cont.)

2. Ground Loops and Remote Voltage Sensing

Ground Loops

Regulator performance can also suffer if ground loops in the circuit wiring are not avoided. The most common ground loop problem occurs when the return lead of the input supply filter capacitor is improperly located, as shown in Figure 5-2. If this return lead is physically connected between the load return and the regulator circuit ground point (''B''), a ripple voltage component (60 or 120 Hz) can be induced on the load voltage, V_L. This is due to the high peaks of the filter capacitor ripple current, i_{ripple}, flowing through the lead resistance between the load and regulator. These peaks can be 5 to 15 times the value of load current. Since the regulator will only keep constant the voltage between its sense lead and ground point, points ''A'' and ''B'' in Figure 5-2, this additional ripple voltage, V_{LEAD}, will appear at the load.

This problem can be avoided by proper placement and connection of the filter capacitor return load as shown in Figure 5-3.

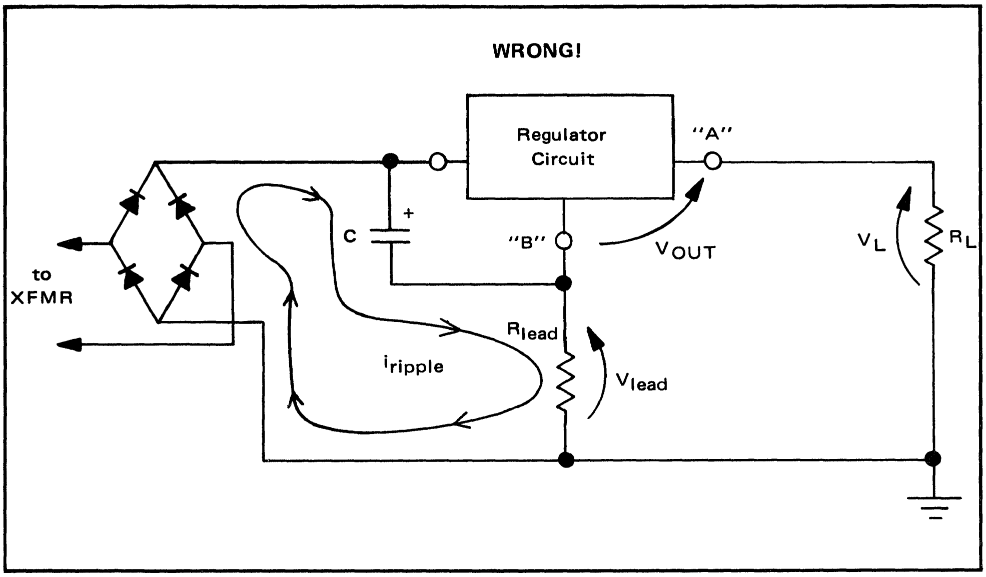


Figure 5-2. Filter Capacitor Ground Loop

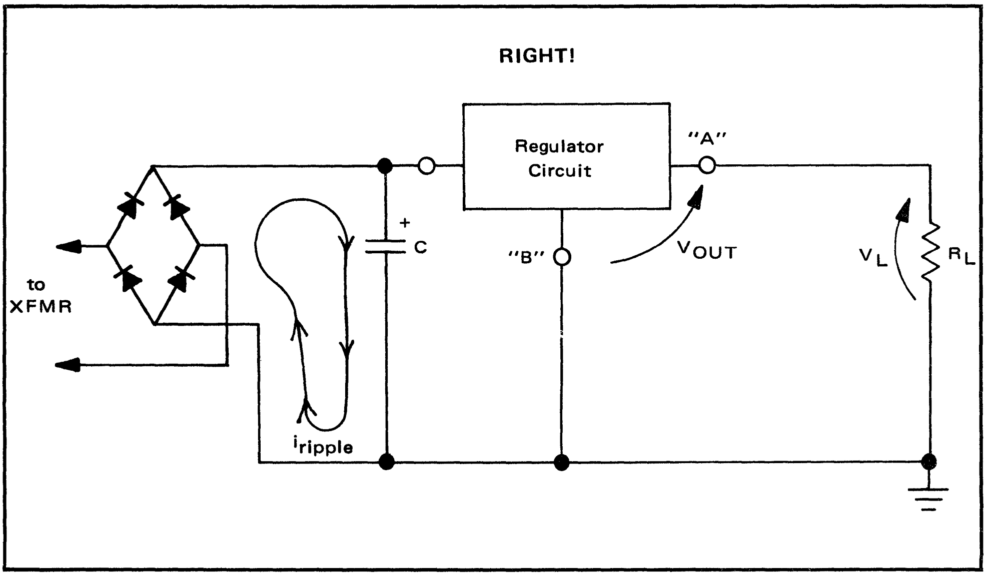


Figure 5-3.

Remote Voltage Sensing

Closely related to the above ground loop problem, is resistance in the current carrying leads to the load. This can cause poorer than expected load regulation in cases where the load currents are large or where the load is located some distance from the regulator. This is illustrated in Figure 5-4. As stated previously, the regulator circuit will keep the voltage present between its sense and ground pins constant. From Figure 5-4 we can see that any lead resistance between these points and the load will cause the load voltage, VL, to vary with varying load current, iL. This effectively lowers the load regulation of the circuit.

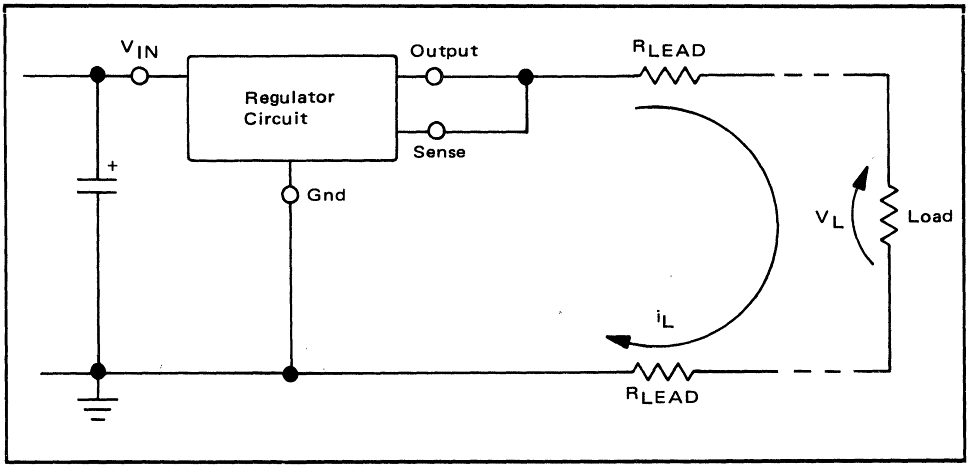


Figure 5-4. Effects of Resistance in Output Leads

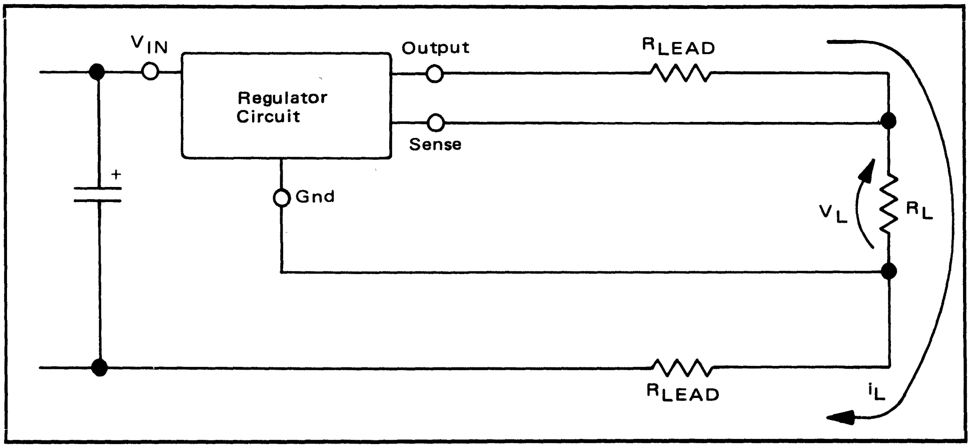


Figure 5-5. Remote Voltage Sensing

This problem can be avoided by use of remote sense leads, as shown in Figure 5-5. The voltage drops in the high current carrying leads now have no effect on the load voltage, VL. However, since the sense and ground leads are usually rather long, care must be exercised that their associated lead inductance is minimized, or loop instability may result. The ground and sense leads should be formed into a twisted pair lead to minimize their lead inductance and noise pickup.

3. Semiconductor Mounting Considerations

An area of regulator construction which frequently does not receive proper attention is the mounting of the semiconductor power devices. Improper mounting of the external series pass transistor(s) and/or IC regulator, if in a power type package (TO-3, TO-66, TO-220, etc.), can result in higher than expected case to heatsink thermal resistances (for thermal information see Section 15) or worse, mechanical damage to the package.

Most problems associated with mounting can be avoided if the following rules are observed:

1. The mounting surface should be flat, smooth, free of deep scratches or burrs, and free of paint, varnish, anodization, or oxidation.
2. Always use a thermal joint compound at the mounting interface (Dow-Corning 340, etc.)
3. Mounting holes should be no larger than those on the semiconductor package; and should be free of burrs or chamfers.
4. TO-3 and TO-66 style packages can be torqued down to the torque limit of the mounting hardware.

Examples of TO-3/TO-66 and TO-220 (Case 221A) mounting techniques are shown in Figures 5-6 and 5-7, respectively.

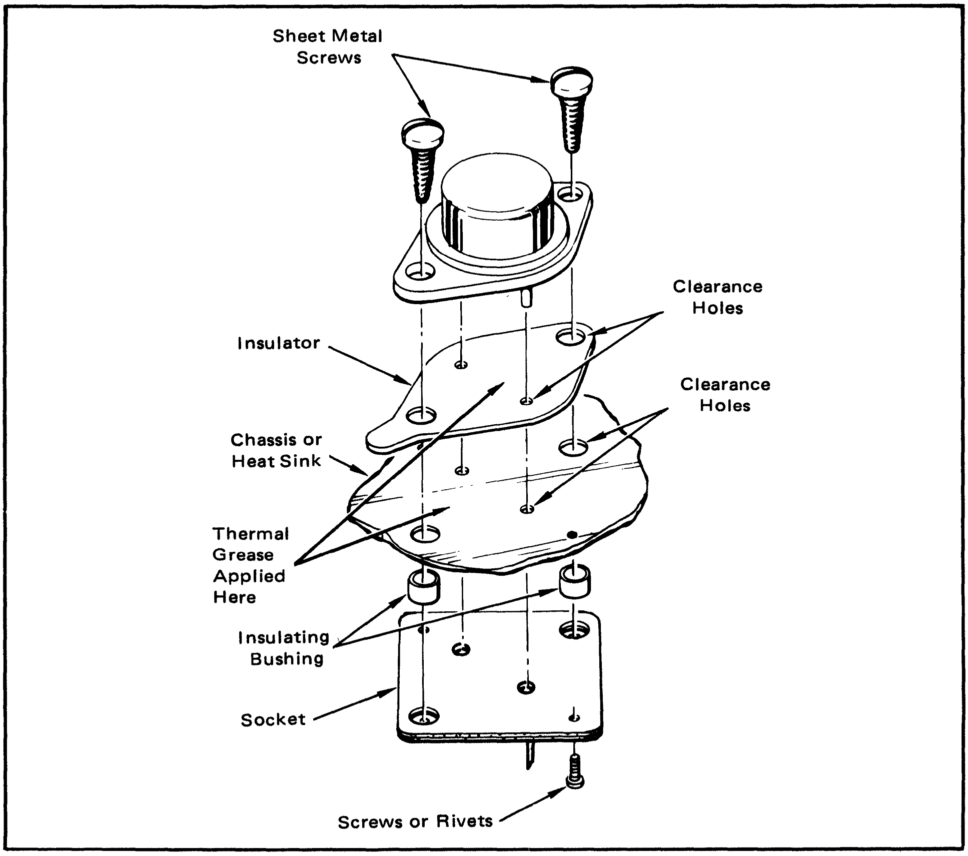
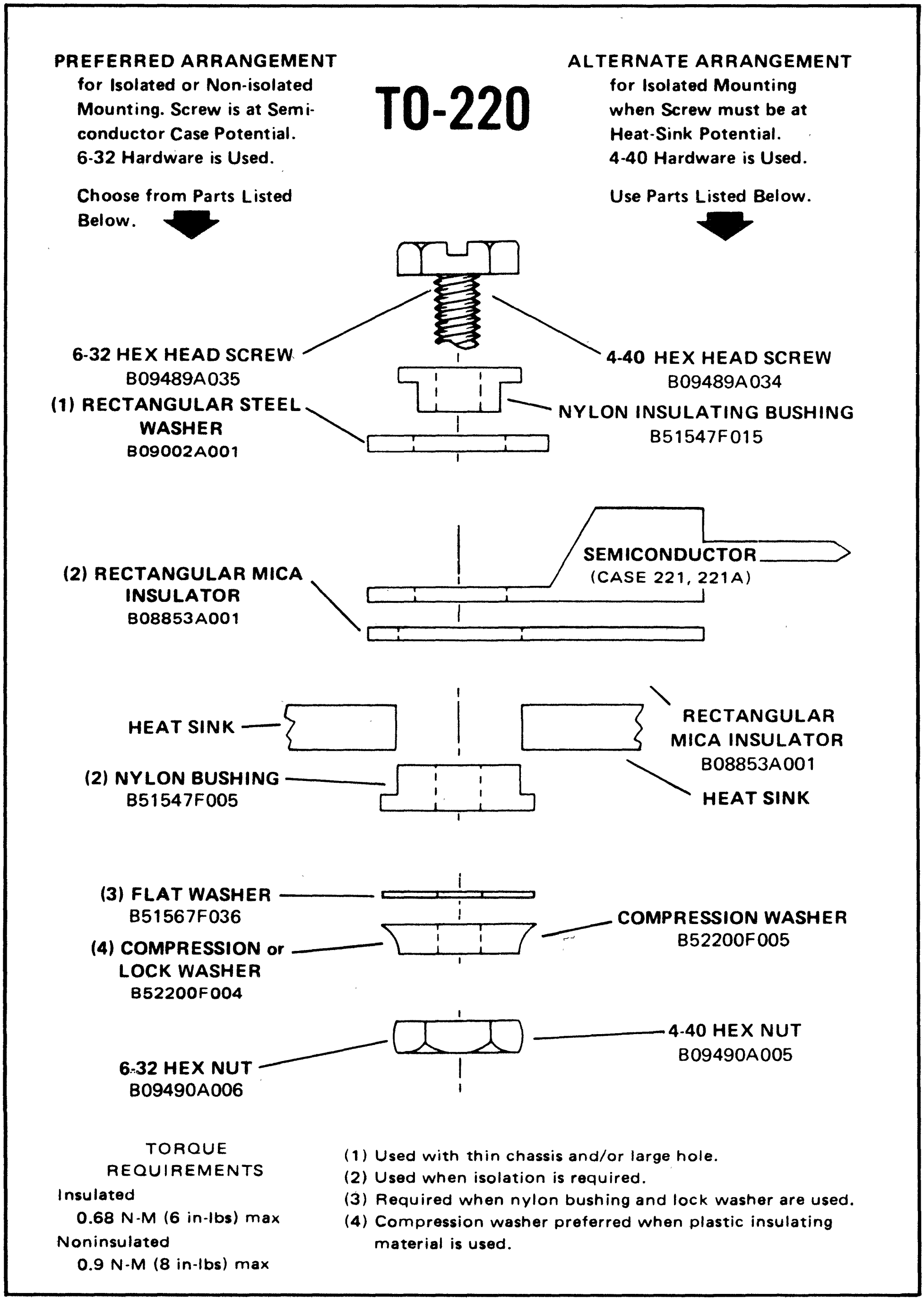


Figure 5-6. Mounting Details for Flat-Base Mounted Semiconductors (TO-66 Shown). When not using a socket, machine screws tightened to their torque limits will produce lowest thermal resistance.



TORQUE
REQUIREMENTS
Insulated
0.68 N-M (6 in-lbs) max
Noninsulated
0.9 N-M (8 in-lbs) max

(1) Used with thin chassis and/or large hole.
(2) Used when isolation is required.
(3) Required when nylon bushing and lock washer are used.
(4) Compression washer preferred when plastic insulating material is used.

Figure 5-7. Mounting Scheme for the TO-220 (Case 221A)

SECTION 6

LINEAR REGULATOR DESIGN EXAMPLE

As an illustration of the use of the material contained in the preceeding sections, the following regulator design example is given.

Regulator Performance Requirements

Output Voltage, Vo = +10V ± .1V

Output Current, Io = 1A, current limited

Load Regulation, $\leqslant$ .1% for Io = 10mA to 750mA

Line Regulation, $\leqslant$ .1%

Output ripple, $\leqslant$ 2mV p-p

Max Ambient Temperature, $T_A \leqslant +70°C$

Supply will have common loads to a negative supply

1. IC Regulator Selection: Study of the available regulators given in the selection guide of Section 17 reveals that both the MC1723C and MC1469 would meet the regulation performance requirements. Both regulators must be current boosted to obtain the required 1A output current A rough cost estimate shows that an MC1723C/ series pass element combination is the most economical approach.

2. Circuit Configuration: In Section 3, an appropriate circuit configuration is found. This is the MC1723 NPN boost configuration of Figure 3-5A.

3. Determination of Component Values: Using the equations given in Figure 3-5A, the values of C_{REF}, R1, R2, R3 and R_{SC} are determined:

a. C_{REF} is chosen to be $0.1\mu F$ for low noise operation.

b. R1 + R2 is chosen to be ≈ 10K.

c. R2 is then given by: $R2 \approx \frac{7v}{V_o}(R1 + R2) = .7\ (10K) = 7K$

d. Since V_{REF} can vary by as much as ± 5% for the MC1723C, R2 should be made variable by at least that much, so that Vo can be set to the required value of +10V ± .1V. R2 is therefore chosen to consist of a 62K resistor and a 2K trimpot.

e. R1 = 10K − R2 = 10K − 7K = 3K

f. $R_{SC} \approx \frac{0.6V}{I_{SC}} = \frac{0.6V}{1A} = .6\Omega$; .56Ω, 1W chosen for R_{SC}.

g. $R3 = R1 \parallel R2 \cong 2.2K$

4. Determination of Input Voltage, V_{IN}: There are two basic constraints on the input voltage: (1) the device limits for minimum and maximum V_{IN} and (2) the minimum input-output voltage differential. These limits are found on the device data sheet (Section 18.) to be:

$$9.5V \leqslant V_{IN} \leqslant 40V \text{ and } (V_{IN} - V_O) \geqslant 3V$$

For the configuration of Figure 3-5A, $(V_{IN} - V_O)$ is given by:

$$(V_{IN} - V_O) = [V_{IN} - (V_O + 2\phi)] \geqslant 3V \text{ where } \phi = V_{BEON} \approx 0.6V$$

Note that $(V_{IN} - V_O)$ is defined on the device data sheet to be the differential between the input and output pins. Since the base-emitter junction drops of Q1 and R_{SC} have been added to the circuit, they must be added to the minimum value of $(V_{IN} - V_O)$. Therefore,

$$V_{IN} \geqslant V_O + 2\phi + 3V = 10 + 1.2 + 3$$

$$V_{IN} \geqslant 14.2V$$

This condition also satisfies the requirement for a minimum V_{IN} of 9.5V.

b. In order to simplify the design of the input supply (see Section 8), V_{IN} is chosen to be 16V average with a 3V p-p ripple at full load and up to 25V at no load. This assures that the input voltage is always above the required minimum value of 14.2V. Now, the output ripple can be determined. The MC1723C has a typical ripple rejection ratio of −74 db, as given on its data sheet. With an input ripple of 3V p-p, the output ripple would be less than 1m V p-p, which meets the regulator output ripple requirements.

5. Determination of regulator package and available output current: Referring to the MC1723 data sheet (Section 18), there are two package styles to choose from. Since the two packages have different thermal characteristics, the amount of available output current will be different for each.

This can be found from:

$$T_J = T_A + \theta_{JA} P_D \text{ (Eq. 6.1 from Section 15)}$$

where θ_{JA} = heatsink and/or pkg total junction-to-ambient thermal resistance

$P_D = V_{IN} \times (I_O + I_{IB})$

I_{IB} = quiescent current of IC regulator

I_O = IC regulator output current

solving for I_O:

$$I_O = \left[\frac{(T_J - T_A)}{\theta_{JA} V_{IN}}\right] - I_{IB} \qquad (6.1)$$

From the device data sheet, we can find the values of T_J, θ_{JA}, and I_{IB}. Eq 6.1 can then be solved. The results are summarized below for an unheatsinked MC1723CL (ceramic DIP), an unheatsinked MC1723CG (metal can), and an infinitely heatsinked MC1723CG packages.

TABLE 6-1

	MC1723CL	MC1723CG	MC1723CG
Heatsink	**None**	**None**	**Infinite**
T_J	175°C	150°C	150°C
T_A	70°C	70°C	70°C
θ_{JA}	150°C/W	184°C/W	70°C/W
I_{IB}	4mA	4mA	4mA
I_O	40mA	23mA	67mA

A choice must now be made. Since it is desirable to have as much available current as possible to drive Q1 (thereby lowering its gain (h_{fe}) requirements), an infinitely heatsinked MC1723CG is the most desirable choice. However, the construction of an infinite heatsink is hardly practical. Therefore, the choice is between an unheatsinked MC1723CL and an MC1723CG with some form of heatsinking. The unheatsinked MC1723CL is chosen since this approach is the least complex.

6. Selection of the Series Pass Element, Q1: The transistor type chosen for Q1 must have the following characteristics (see Section 4):

a. $V_{CEO} \geqslant V_{INMAX}$

b. $I_{CMAX} \geqslant I_{SC}$

c. $h_{fe} \geqslant \frac{I_{SC}}{I_O}$ @ $V_{CE} = V_{IN} - V_O - \phi$

where $\phi = V_{BEON} \approx 0.6V$

d. $P_{DMAX} \geqslant V_{IN}, \times I_{SC}$

e. θ_{JC} such to allow practical heatsinking

f. SOA such that it can withstand

$$V_{CE} = V_{IN} \text{ @ } I_C = I_{SC}$$

for this example:

$$V_{CEO} \geqslant 25V$$

$$I_{CMAX} \geqslant 1A$$

$$h_{fe} \geqslant 25 \text{ @ } V_{CE} = 5V \text{ @ } I_C = 1A$$

$$P_{DMAX} \geqslant 16W$$

$$\theta_{JC} = 1.52°C/W$$

$$\text{SOA: } 1A \text{ @ } 16V$$

A 2N3055 transistor is chosen as a suitable device for Q1 using the selection guide of Section 4 and the transistor data sheets (available from device manufacturer).

7. Q1 Heatsink Calculation

$$T_J = T_A + \theta_{JA} P_D \text{ (Eq 15.1 from Section 15)}$$

where $P_D = V_{IN} \times I_{SC}$

$$\theta_{JA} = \theta_{JC} + \theta_{CS} + \theta_{SA} \text{ (Eq 6.2)}$$

solving for θ_{SA}:

$$\theta_{SA} = \left[\frac{T_J - T_A}{P_D}\right] - (\theta_{JC} + \theta_{CS}) \tag{6.2}$$

From the 2N3055 data sheet, $T_J = 200°C$ and $\theta_{JC} = 1.52°C/W$. The transistor will be mounted with thermal grease directly to the heatsink. Therefore, θ_{CS} is found to be 0.1°C/W from Table 15-1.

Solving 6.2:

$$\theta_{SA} = \left[\frac{200°C - 70°C}{16V \times 1A}\right] - (1.52 + 0.1)\ °C/W$$

$$\leqslant 6.6°C/W$$

A commercial heatsink is now chosen from Table 15-2 or a custom designed using the methods given in Section 15. For this example, a thermalloy 6003 heatsink having a θ_{CS} of 6.2°C/W was used.

8. Clamp Diode: Since the regulator can power a load which is also connected to a negative supply, a 1N4001 diode is connected to the output for protection. (See general design considerations, Section 3H.) The complete circuit schematic is shown in Figure 6-1.

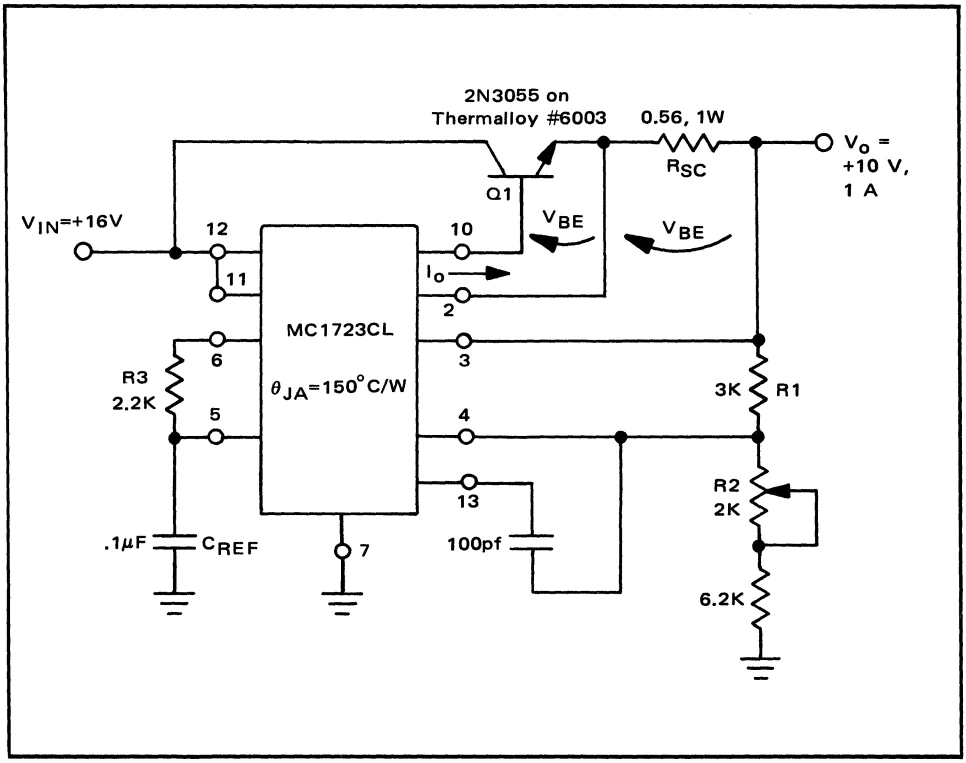


Figure 6-1. +10V, 1A Design Example

9. Construction Input Supply Design: The input supply is now designed using the information contained in Section 8 and the regulator circuit is constructed using the guidelines given in Section 5.

SECTION 7

LINEAR REGULATOR CIRCUIT TROUBLESHOOTING CHECKLIST

Occasionally the designer's prototype regulator circuit will not operate properly. If problems do occur, the trouble can be traced to a design error in 99.9% of the cases. As a troubleshooting aid to the designer, the following guide is presented.

Of course, it would be difficult, if not impossible, to devise a troubleshooting guide which would cover all possible situations. However, the checklist provided will help the designer pinpoint the problem in the majority of cases. To use the guide, first locate the problem's symptom(s) and then carefully recheck the regulator design in the area indicated using the information contained in the referenced handbook section.

SYMPTOM	DESIGN AREA TO CHECK	REFER TO SECTION
Regulator Oscillates	1. Layout 2. Compensation capacitor too small 3. Input leads not bypassed 4. External pass element parasitically oscillating	5 3, 18 5 5
Loss of Regulation at Light Loads	1. Emitter-Base resistor in "PNP" type boost configuration too large 2. Absence of 1 mA "minimum" load (see load regulation test spec on device data sheet) 3. Improper circuit configuration	4 18 3
Loss of Regulation at Heavy Loads	1. Input Voltage too low (V_{INMIN}, $\lvert V_{IN} - V_O \rvert_{MIN}$) 2. External pass element gain too low 3. Current limit too low 4. Line resistance between sense points and load 5. Inadequate heatsinking	2, 3, 18 17 4 3 5 15
IC Regulator or Pass Element Fails after Warm-Up or at High T_A	1. Inaequate heatsinking 2. Input Voltage Transient (V_{INMAX}, V_{CEO})	15 2, 4, 5, 17, 18
Pass Element Fails During Short Circuit	1. Insufficient pass element ratings (SOA, I_{CMAX}) 2. Inadequate heatsinking	4 15

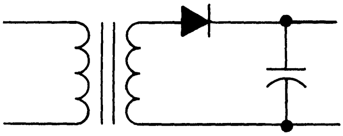
TROUBLESHOOTING CHECKLIST

SYMPTOM	DESIGN AREA TO CHECK	REFER TO SECTION
IC Regulator Fails During Short Circuit	1. IC current or SOA capability exceeded 2. Inadequate heatsinking	2, 18
IC Regulator Fails During Power Up	1. Input voltage transient (V_{INMAX}) 2. IC current or SOA capability exceeded as load (capacitor) is charged up.	2, 18 2, 18
IC Regulator Fails During Power-Down	1. Regulator reverse biased	3.H
Output Voltage Does Not Come Up During Power-Up or After Short Circuit	1. Output polarity reversal 2. Load has ''latched-up'' in some manner (usually seen with op amps, current sources, etc.)	3.H
Excessive 60 or 120 Hz Output Ripple	1. Input supply filter capacitor ground loop	5

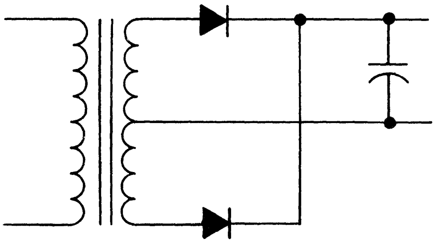
If, after carefully rechecking the circuit, the designer is not successful in resolving the problem, seek assistance from the factory by contacting the nearest Motorola Sales office.

SECTION 8
DESIGNING THE INPUT SUPPLY

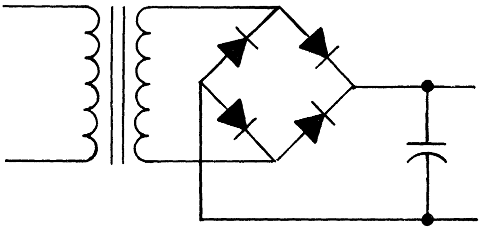
Most input supplies used to power series pass regulator circuits consist of a 60 Hz, single phase step-down transformer followed by a rectifier circuit whose output is smoothed by a choke or capacitor input filter. The type of rectifier circuit used can be either a half-wave, full-wave, or full-wave bridge type, as shown in Figure 8-1. The half-wave circuit is used in low current applications, while the full-wave is preferrable in high-current, low output voltage cases. The full-wave bridge is usually used in all other high-current applications.



Half-Wave



Full-Wave or Full-Wave Center Tap



Full-Wave Bridge

Figure 8-1. Rectification Schemes

In this section, specification of the filter capacitor, rectifier and transformer ratings will be discussed. The specifications for the choke input filter will not be considered since the simpler capacitor input type is more commonly used in series regulated circuits. A detailed description of this type of filter can be found in the reference listed at the end of this section.

1. Design of Capacitor-Input Filters

The best practical procedure for the design of capacitor-input filters still remains based on the graphical data presented by Schade[1] in 1943. The curves shown in Figures 8-2 through 8-5 give all the required design information for half-wave and full-wave rectifier circuits. Whereas Schade originally also gave curves for the impedance of vacuum-tube rectifiers, the equivalent values for semiconductor diodes must be substituted. However, the rectifier forward drop often assumes more significance than the dynamic resistance in low-voltage supply applications, as the dynamic resistance can generally be neglected when compared with the sum of the transformer secondary-winding resistance plus the reflected primary-winding resistance. The forward drop may be of considerable importance, however, since it is about 1 V, which clearly cannot be ignored in supplies of 12 V or less.

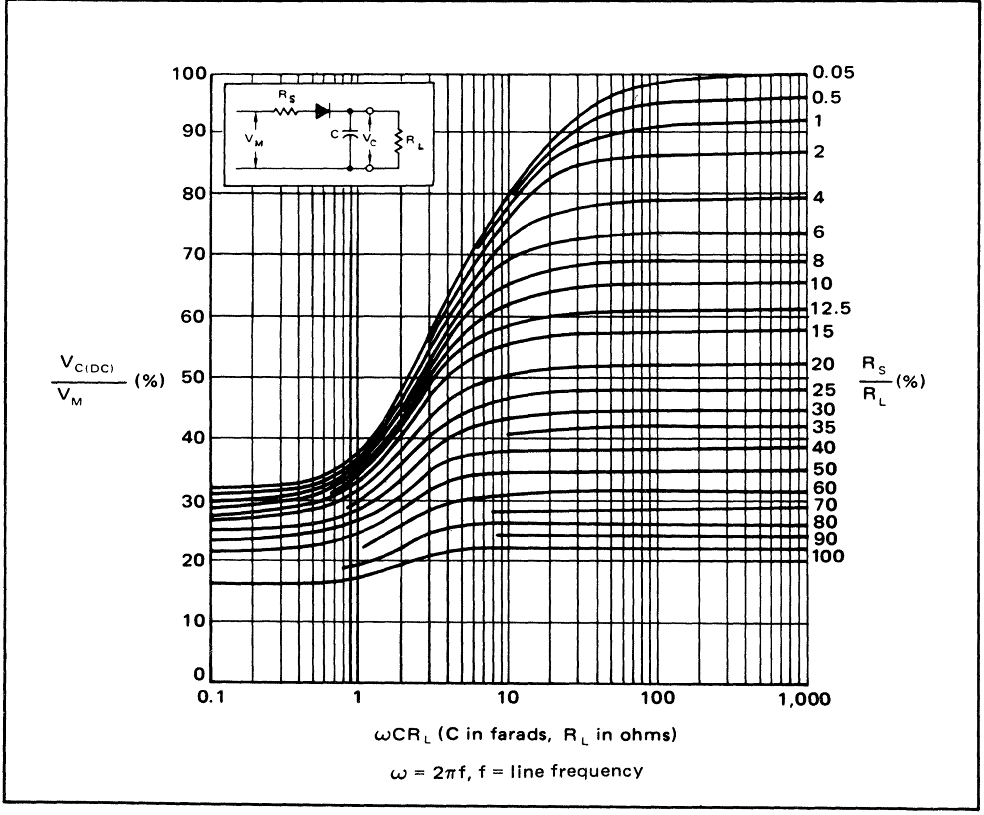


Figure 8-2. Relation of applied alternating peak voltage to direct output voltage in half-wave capacitor-input circuits. *(From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)*

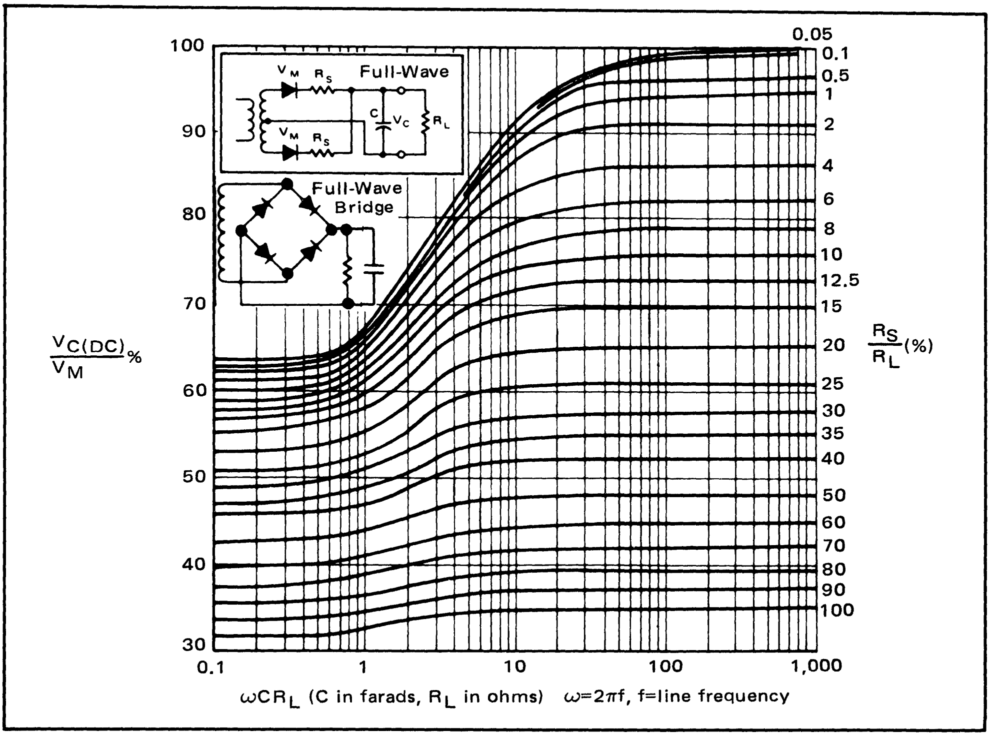
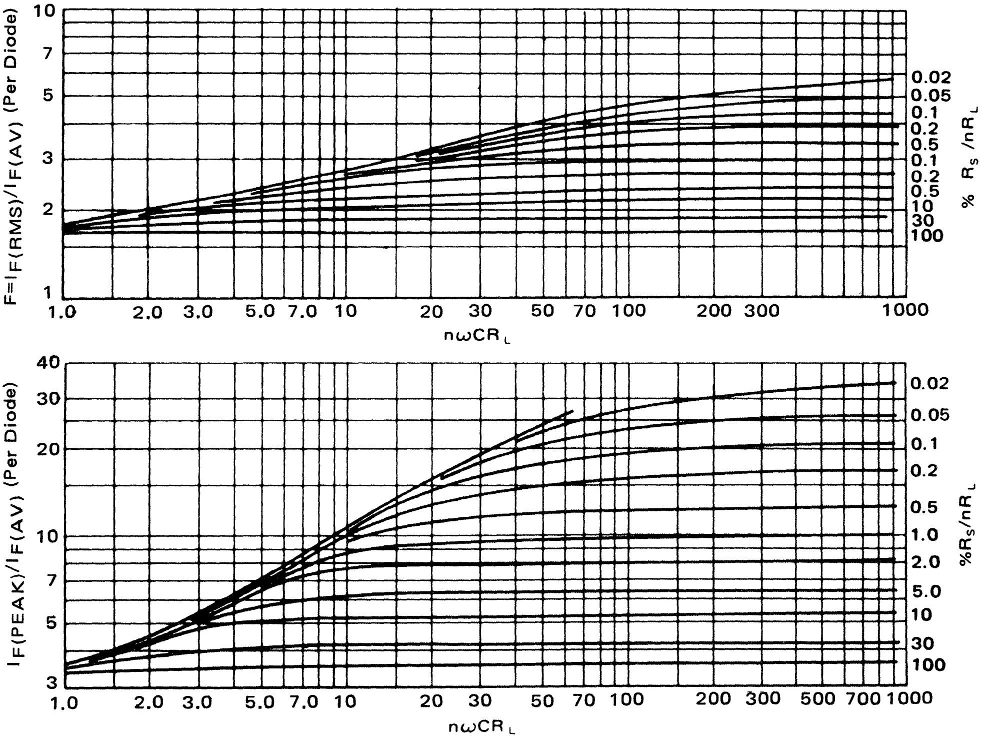


Figure 8-3. Relation of applied alternating peak voltage to direct output voltage in full-wave capacitor-input circuits. *(From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)*



$n = \begin{cases} 1 & \text{For Half-Wave Single-Phase Rectifier Circuits} \\ 2 & \text{For Full-Wave Single-Phase Rectifier Circuits} \end{cases}$

C in Farads

R_L in Ohms

$\omega = 2\pi f$ where f = Line Frequency

R_S = RMS Equivalent Source Resistance

Figure 8-4. Relation of RMS and peak to average diode current in capacitor-input circuits.
(From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)

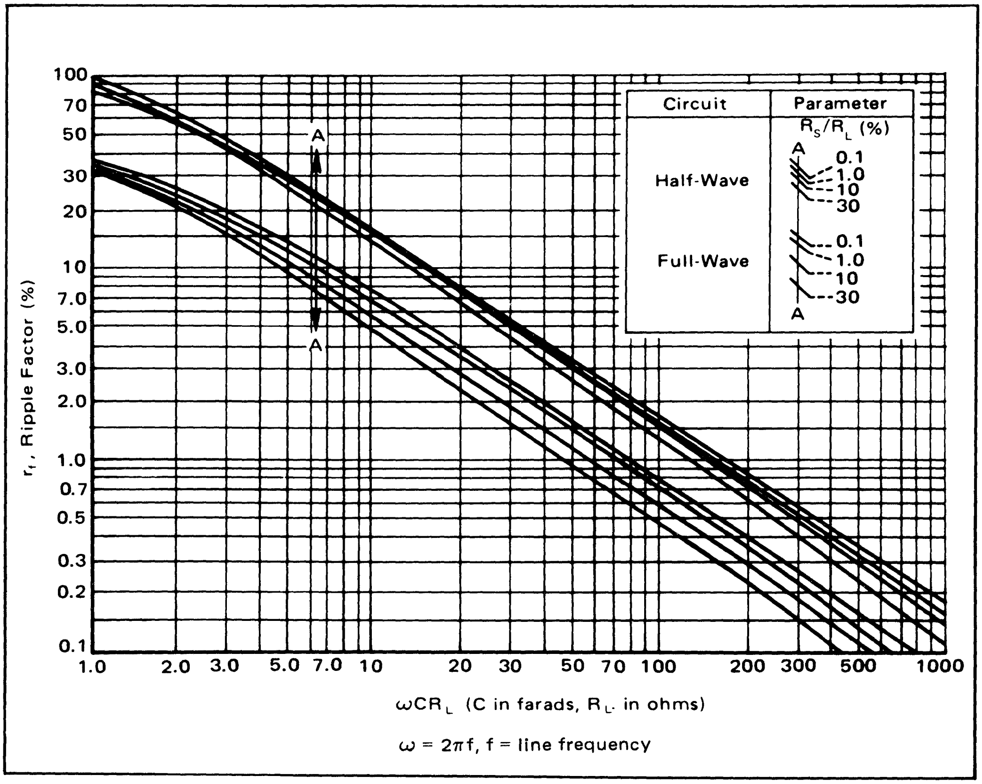


Figure 8-5. Root-mean-square ripple voltage for capacitor-input circuits. *(From O. H. Schade, Proc. IRE, vol. 31, p. 356, 1943.)*

Returning to the above curves, the full-wave circuit will be considered. Figure 8-3 shows that a circuit must operate with ωCRL ≥ 10 in order to hold the voltage reduction to less than 10 percent and ωCRL ≥ 40 to obtain less than 2 percent reduction. However, it will also be seen that these voltage-reduction figures require Rs/RL, where Rs is now the total series resistance, to be about 0.1% which, if attainable, causes repetitive peak-to-average current ratios from 10 to 17 respectively, as can be seen from Figure 8-4. These ratios can be satisfied by many diodes; however, they may not be able to tolerate the turn-on surge current generated when the input-filter capacitor is discharged and the transformer primary is energized at the peak of the input waveform. The rectifier is then required to pass a surge current determined by the peak secondary voltage less the rectifier forward drop and limited only by the series resistance Rs. In order to control this turn-on surge, additional resistance must often be provided in series with each rectifier. It becomes evident, then, that a compromise must be made between voltage reduction on the one hand and diode surge rating and hence average current-carrying capacity on the other hand. If small voltage reduction, that is good voltage regulation, is required, a much larger diode is necessary than that demanded by the average current rating.

Surge Current

The capacitor-input filter allows a large surge to develop, because the reactance of the transformer leakage inductance is rather small. The maximum instantaneous surge current is approximately V_M/R_S and the capacitor charges with a time constant $\tau \approx R_S C_1$. As a rough — but conservative — check, the surge will not damage the diode if V_M/R_S is less than the diode I_{FSM} rating and τ is less than 8.3 ms. It is wise to make R_S as large as possible and not pursue tight voltage regulation; therefore, not only will the surge be reduced but rectifier and transformer ratings will more nearly approach the dc power requirements of the supply.

As an aid in the selection of a suitable rectifier or bridge, the brief selection guide of Table 8-1 is included.

TABLE 8-1
RECTIFIERS

$I_{F(AVG)}$	I_{FSM}	SERIES
1.0A	30A	1N4000
1.5A	50A	1N5391
3.0A	100A	MR500
3.0A	200A	1N5400
6.0A	400A	MR750
12A	300A	MR1120
20A	400A	MR2000S
25A	600A	MR2500S
40A	800A	1N1183A
	BRIDGES	
1.0A	30A	MDA100A
2.0A	50A	MDA200
4.0A	100A	MDA400
8.0A	400A	MDA800
12A	400A	MDA1200
25A	400A	MDA2500
35A	400A	MDA3500

2. Design Procedure

A. From the regulator circuit design (see Section 6), we know:

$V_{C(DC)}$ = The required full load average DC output voltage of the capacitor input filter

$V_{Ripple(p-p)}$ = the maximum full load peak-to-peak ripple voltage

V_m = the maximum no load output voltage

I_o = the full-load filter output current

f = the input AC line frequency

B. From Figure 8-5, we can determine a range of minimum capacitor values to obtain sufficient ripple attenuation. First determine r_f:

$$r_f = \frac{V_{Ripple(p-p)}}{2\sqrt{2}\,V_{C(DC)}} \times 100\% \tag{8.1}$$

a range for $\omega C R_L$ can now be found from Figure 8-5.

C. Next, determine the range of R_S/R_L from Figure 8-2 or 8-3 using $V_{C(DC)}$ and the values for $\omega C R_L$ found in part B. If the range of $\omega C R_L$ values initially determined from Figure 8-5 is above $\simeq$ 10, R_S/R_L can be found from Figures 8-2 and 8-3 using the lowest ωCR_L value. Otherwise, several iterations between Figures 8-2 or 8-3 and 8-5 may be necessary before an exact solution for R_S/R_L and ωCR_L for a given r_f and $V_{C(DC)}/V_m$ can be found.

D. Once ωCR_L is found, the value of the filter capacitor, C, can be determined from:

$$C = \frac{\omega CR_L}{2\pi\left(\frac{V_{C(DC)}}{I_O}\right)} \tag{8.2}$$

E. The rectifier requirements may now be determined:

1. Average Current

$$I_{F(AVG)} = I_O \text{ for half-wave rectification} \tag{8.3}$$

$$= I_O/2 \text{ for full-wave rectification}$$

2. RMS and Peak repetitive rectifier current ratings can be determined from Figure 8-4.

3. The rectifier PIV rating is 2 V_m for the half-wave and full wave circuits, V_m for the full-wave bridge circuit. In addition, a safety margin of 20% to 50% is advisable due to the possibility of line transients.

4. Maximum Surge Current

$$I_{SURGE} = V_m/(R_S + ESR) \tag{8.4}$$

where ESR = minimum equivalent series resistance of filter capacitor from its data sheet

F. Transformer Specification

1. Secondary Leg RMS Voltage

$$V_S = \{V_m + (n)\ 1.0\}/\sqrt{2} \tag{8.5}$$

where n = 1 for half-wave and full-wave

= 2 for full-wave bridge

2. Total resistance of secondary and any external resistors to be equal to R_S found from Figures 8-2, -3, and -4 (see Part C).

3. Secondary RMS Current

Half-Wave = I_{rms}

Full-Wave = I_{rms}

Full-Wave Bridge = $\sqrt{2}\ I_{rms}$

where I_{rms} = rms rectifier current (from part E.1 and E.2).

4. Transformer VA rating

Half-Wave = $V_S\ I_{rms}$

Full-Wave = $2\ V_S\ I_{rms}$ (8.7)

Full-Wave Bridge = $V_S\ I_{rms}\ (\sqrt{2})$

where I_{rms} = rms rectifier current (from part E.1 and E.2)

and V_S = Secondary Leg RMS Voltage

3. Design Example

A. Find the values for the filter capacitor, transformer rectifier ratings, given:

Full-Wave Bridge Rectification

$$V_{C(DC)} = 16V$$

$$V_{RIPPLE\ (p\text{-}p)} = 3\ V$$

$$V_M = 25\ V$$

$$I_O = 1\ A$$

$$f = 60\ Hz$$

B. Using Equation (8.1)

$$r_f = \frac{3}{2\sqrt{2}\,(16)} \times 100\% = 6.6\%$$

from Figure 8-5, $\omega CR_L \simeq 7$ to 10

C. Using $\omega CR_L = 10$, R_S/R_L is found from Figure 8-3 using:

$$\frac{V_{C(DC)}}{V_M} = \frac{16}{25} = .64 = .64\%$$

$$. .\ R_S/R_L = 20\% \text{ or } R_S = .2 \times R_L = .2(\frac{V_{C(DC)}}{I_O}) = .2\ (16)$$

$$R_S = 3.2\ \Omega$$

D. From Equation (8.2), the filter capacitor size is found:

$$C = \frac{\omega CR_L}{2\pi f(\frac{V_{C(DC)}}{I_O})} = \frac{10}{2\pi(60)16} = 1657\ \mu F$$

E. The rectifier ratings are now specified:

1. $I_{F(AVG)} = I_O/2 = 0.5$ A from Eq (8.3)
2. $I_{F(RMS)} = 2 \times I_{F(AVG)} = 1$ A from Fig. 8-4
3. $I_{F(PEAK)} = 5.2 \times I_{F(AVG)} = 2.6$ A from Fig. 8-4
4. PIV $= V_M = 25$ V (use 50 V for safety margin)
5. $I_{SURGE} = V_M/(R_S + ESR) \simeq 25/3.2 = 7.8$ A from Eq (8.4) (neglecting capacitor ESR)

F. The transformer should have the following ratings:

1. $V_S = \{V_M + n(1.0)\}/\sqrt{2} = (25 + 2)/\sqrt{2} = 19$ VRMS {from Eq (8.5)}
2. Secondary Resistance should be 3.2 Ω.
3. Secondary RMS current rating should be 1.4 A {from Eq (8.6)}
4. From Eq. (8.7), the transformer should have a 27 VA rating.

It should be noted that, in order to simplify the procedure, the above design does not allow for line voltage variations or component tolerances. The designer should take these factors into account when designing his input supply. Typical tolerances would be: Line Voltage – +10%, –15% and Capacitors – +75%, –10%.

REFERENCES

1. O. H. Schaade, Proc. IRE, Vol. 31, 1943.
2. Motorola Silicon Rectifier Manual, 1980.

SECTION 9

SWITCHING REGULATORS VERSUS LINEAR REGULATORS

A. THE MARKET

A switching power supply or switcher is a high frequency power conversion circuit. It uses the ac power line to produce one or more regulated dc voltages. Switchers became practical in the early 60's with the advent of fast, high voltage transistors that made it possible for designers to operate directly off the rectified high voltage (120/220 V) ac lines. By 1970 almost every power supply company had a switcher or line of switchers in their catalog. And today, it is estimated that 20% of the regulated AC-DC power supply market belongs to switchers (See Figure 9-1). The chart indicates that this market will enjoy a compound growth rate (CGR) of about 15% annually but that switchers will average a 30% CGR and will capture 40% of the market by 1985. At this time, the fastest growing market segment is the small, single transistor, switchers (50 to 150 watts). These supplies are benefitting from the current boom in microprocessor and minicomputer equipment such as bank auto tellers and point-of-sale terminals.

B. COMPARISON WITH LINEAR REGULATORS

Switching power supplies offer advantages of efficiency, size, and weight, but also require a more complex design, cannot meet some of the performance capabilities of linear supplies, and can generate a considerable amount of electrical noise. Even with some of the disadvantages, switchers are being accepted in the industry, particularly where size and efficiency are of prime importance. In most applications performance is adequate, and they are cost competitive in the 50 W power level and above. Figure 9-2 illustrates the trends in cost as a function of

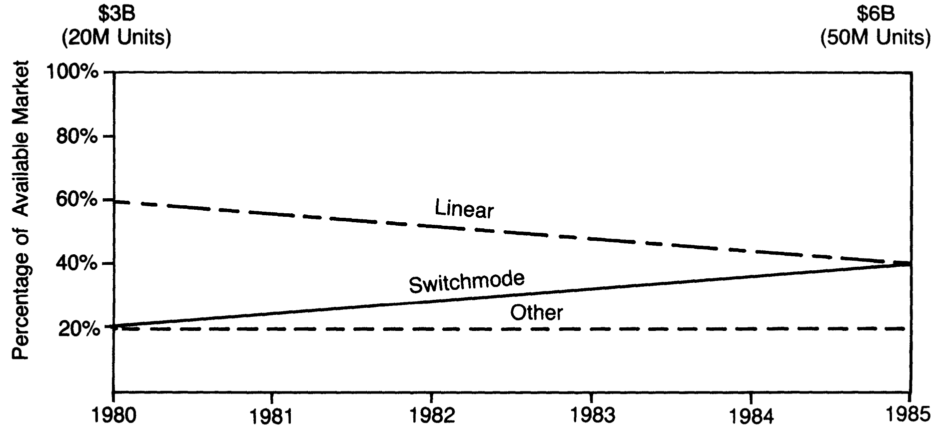


Figure 9-1. Market Trends for Power Supplies

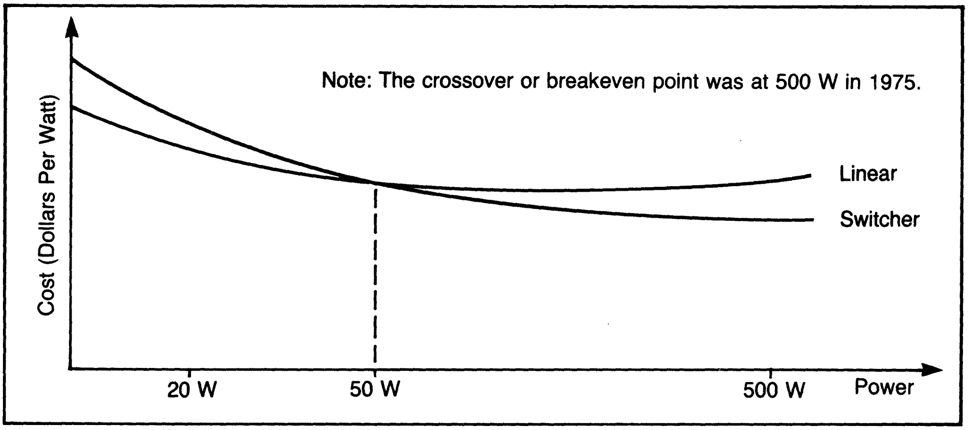


Figure 9-2. 1980 Cost Comparison

output power. Because the switcher's passive components such as transformers and filters are smaller, they are almost always lower in cost than the high power (100 W) linear regulators. However, active component count is high (70 to 140 devices) and remains high regardless of the output power rating. This makes it less cost effective at the lower power levels. Switchers have been significantly cost reduced in the past five years because designers have been able to simplify the control circuits and have found even lower cost alternatives in the passive component area. The 500 W break even point (switcher versus linear) was broken five years ago, and the present 50 W break even point is expected to drop to 20 W in the next couple of years. An example of present parts cost in a 50 W switcher is shown in Table 9-1. The active component semiconductor cost is a somewhat higher percentage of the total at this power level. The average cost of semiconductors for switchers tends to be about 10% of the selling price. This can be subdivided into 5% for rectifiers and about 2% each for transistors and IC's.

Finally, the actual performance comparison chart is shown in Table 9-2. Single output switcher efficiencies run from 70 to 80% but occasionally fall to 60–65% with post regulated auxiliary outputs. Some linear power supplies on the other hand, are operated with up to 50% efficiency, but these are areas where line variations and short hold-up time problems are minimal. Most linear supplies

TABLE 9-1

Approximate Parts Cost of Similar 50 W Power Supplies (1980)

Component	20 kHz Switcher ($)	Linear ($)
Magnetics	8	10
Capacitors	7	7
*Rectifiers	5	3
*Transistors	3	2
*IC's	2	2
Misc. (Line/Heat Sinks)	5	8
TOTAL	30	32

*Semiconductors account for 22% of the total cost in linear power supplies and 33.4% for switchers.

TABLE 9-2

20 kHz Switcher versus Linear Performance

Parameter	Switcher	Linear
Efficiency	75%	30%
Size	2.0 W/IN^3	0.5 W/IN^3
Weight	40 W/lb.	10 W/lb.
Cost 200–500 W*	$1.00/W	$1.25/W
Cost 50–150 W*	$1.50/W	$1.50/W
Line and Load Regulation	0.1%	0.1%
Output Ripple V_{P-P}	50 mV	5.0 mV
Noise V_{P-P}	50–200 mV	—
Transient Response	1 ms	20 μs
Hold-Up Time	20–30 ms	1–2 ms

*Based on 1980 Cost Figures

operate with typical efficiencies of only 30%. The overall size reduction of a 20 kHz switcher is about 4:1 over an equivalent linear supply. Newer designs in the 100 to 200 kHz region end up at about 6:1. Other characteristics such as static regulation specs are comparable, while ripple and load transient response are usually worse. Output noise specs can be somewhat misleading. Very often a 200 mV switching spike at the output may be attenuated considerably at the load itself due to the series inductance of the connecting cables and the additional filter capacitors found in many logic circuits. In the future, noise generated at higher switching frequencies (100–500 kHz) will probably be easier to filter and the transient response will be faster. Switchers also exhibit long hold-up time due to their inherent ability to regulate over wide variations in input voltage. It is easier to store the required energy in high voltage input filter capacitors (200–400 V) than in lower voltage (20–50 V) capacitors common to linear power supplies. This is because the physical size of a capacitor is dependent on its CV product, while energy storage is proportional to CV^2.

SECTION 10
SWITCHING REGULATOR TOPOLOGIES

A switching power supply is a relatively complex circuit as is shown by the four basic building blocks of Figure 10-1. It is apparent here that the heart of the supply is really the high frequency inverter. It is here that the work of chopping the rectified line at a high frequency (≥20 kHz) is done. It is here also that the line voltage is transformed down to the correct output level for use by logic or other electronic circuits. The remaining blocks support this basic function. The 60 Hz input line is rectified and filtered by one block, and after the inverter steps this voltage down, the output is again rectified and filtered. The task of regulating the output voltage is left to the control circuit which closes the loop from the output to the inverter. Most control circuits generate a fixed frequency internally and utilize pulse width modulation techniques to implement the desired regulation. Basically, the on-time of the square wave drive to the inverter is controlled by the output voltage. As the load is removed or input voltage increases, a slight rise in output voltage will signal the control circuit to deliver narrower pulses to the inverter, and conversely, as the load is increased or input voltage decreases, wider pulses will be fed to the inverter.

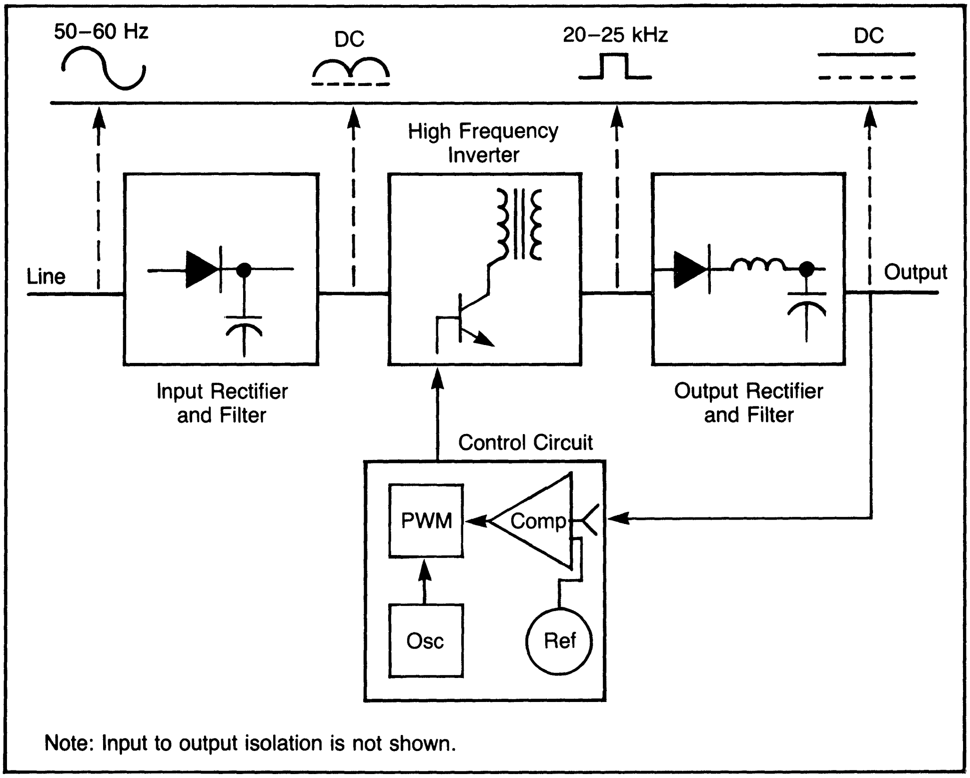


Figure 10-1. Functional Block Diagram — Switching Power Supply

A. BUCK AND BOOST

The inverter topologies used in today's switchers actually evolved from the buck and boost circuits shown in Figure 10-2A & 10-2B. In each case, the regulating means and loop analysis will remain similar, but a transformer is added in order to provide electrical isolation between the line and load. The forward converter family which includes the push-pull and half bridge circuits evolved from the buck regulator (Figure 10-2A). And the newest switcher, the flyback converter, actually evolved from the boost regulator. The buck circuit interrupts the line and provides a variable pulse width square wave to a simple averaging LC filter. In this case, the first order approximation of the output voltage is $V_{out} = V_{in} \times$ duty cycle, and regulation is accomplished by simply varying the duty cycle. This is satisfactory for most analysis work, and only the transformer turns ratio will have to be adjusted slightly to compensate for IR drops, diode drops, and transistor saturation voltages.

Operation of the boost circuit (Figure 10-2B) is more subtle in that it first stores energy in a choke and then delivers this plus energy from the input line to the load. However, the flyback regulators which evolved from this configuration deliver only the inductive energy stored in the choke to the load. This method of operation is actually based on the boost variation model shown in Figure 10-2C. Here, when the switch is opened, only the stored inductive energy is delivered to the load. The true boost circuit can also regulate by stepping up (or boosting) the input voltage, whereas the variation or flyback regulator can step the input voltage up or down. Analysis of the boost regulator begins by dealing with the choke as an energy storage element which delivers a fixed amount of power to the load:

$$P_O = 1/2\, L\, I^2 f_o$$

where I = the peak choke current

f_o = the operating frequency

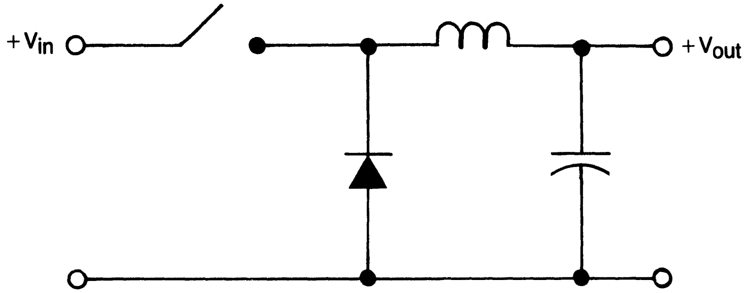
and L = the inductance

Because it delivers a fixed amount of power to the load regardless of load impedance (except for short circuits), the boost regulator is the designer's first choice in photo-flash and capacitive-discharge (CD) automotive ignition circuits to recharge the capacitive load. It also makes a good battery charger. For an electronic circuit load, however, the load resistance must be known in order to determine the output voltage:

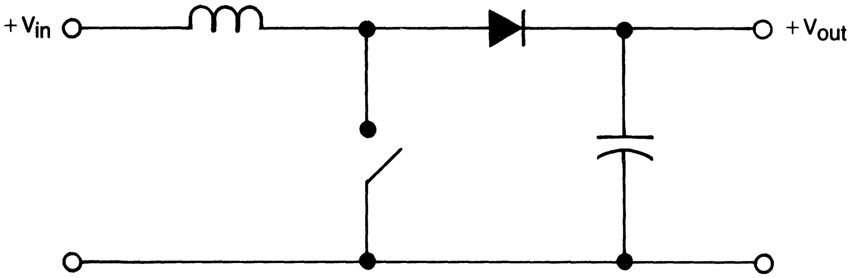
$$V_O = \sqrt{P_o R_L} = I \sqrt{\frac{L f_o R_L}{2}}$$

where R_L = The load resistance

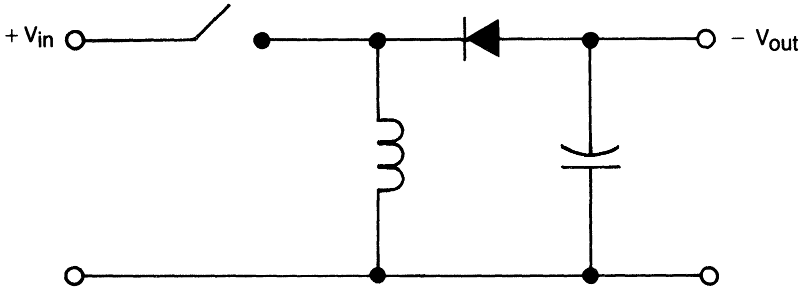
In this case, the choke current is proportional to the on time or duty cycle of the switch, and regulation for fixed loads simply involves varying the duty cycle as before. However, the output also depends on the load (which was not the case with buck regulators) and results in a variation of loop gain with load.



A. Buck (Step-Down)
$V_{out} \leq V_{in}$



B. Boost (Step-Up)
$V_{out} > V_{in}$



C. Boost variation which
resembles the flyback
regulator (step up or
down)

Figure 10-2. Non-Isolated DC-DC Converters

For both regulators, transient response or responses to step changes in load are very difficult to analyze. They lead to what is termed a "load dump" problem. This requires that energy already stored in the choke or filter be provided with a place to go when load is abruptly removed. Practical solutions to this problem include limiting the minimum load and using the right amount of filter capacitance to give the regulator time to respond to this change.

B. FLYBACK AND FORWARD CONVERTERS

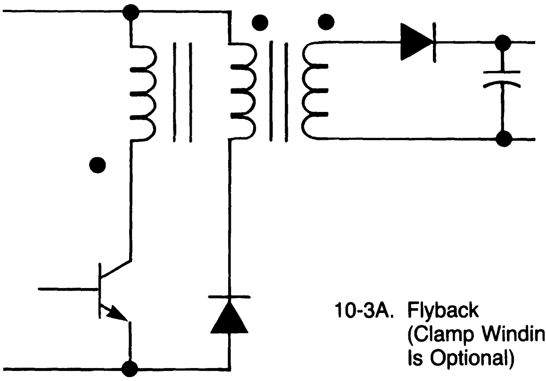
To take advantage of the regulating techniques just discussed, and also provide isolation, a total of five popular topologies have evolved and are illustrated in figures 10-3 and 10-6. Each circuit has a practical power range or capability associated with it as follows:

Circuit	Power Range	Motorola Reference
Flyback	50 to 100 watts	EB87
Forward	100 to 200 watts	Power Leader
Push-Pull	200 to 500 watts	EB88, AN-737A
Half Bridge	200 to 500 watts	EB's 86 & 100, AN-767
Full Bridge	500 to 2000 watts	EB-85

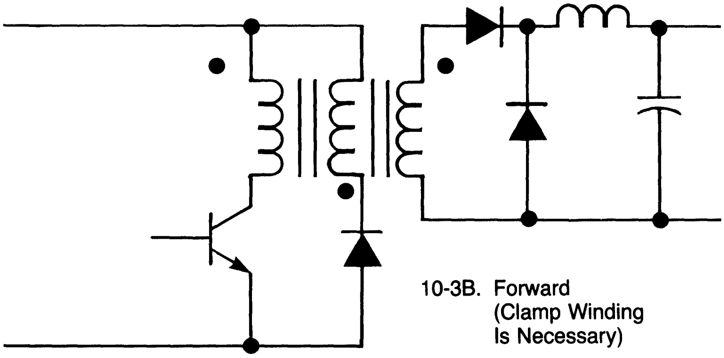
First to be discussed will be the low power (20–200 W) converters which are dominated by the single transistor circuits shown in Figure 10-3. All of these circuits operate the magnetic element in the unipolar rather than bipolar mode. This means that transformer size is sacrificed for circuit simplicity.

1. Flyback — The flyback (alternately known as the "ringing choke") regulator stores energy in the primary winding and dumps it into the secondary windings (Figure 10-3A). A clamp winding is usually present to allow energy stored in the leakage reactance to return safely to the line instead of avalanching the switching transistor. The operating model for this circuit is the boost circuit variation discussed earlier. The flyback is the lowest cost regulator (except at high power levels) because output filter chokes are not required, since the output capacitors feed from a current source rather than a voltage source. Because of this, the flyback will have higher output ripple than the forward converter. However, the flyback is an excellent choice when multiple output voltages are required and does tend to provide better cross regulation than the other types. In other words, changing the load on one winding will have little effect on the output voltage of the others.

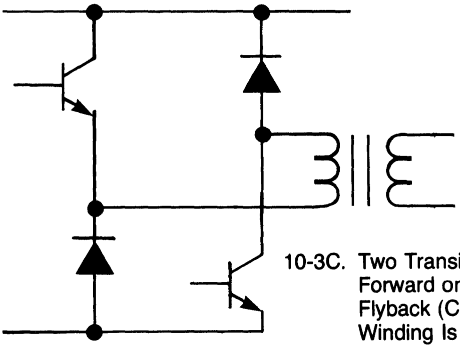
A 120/220 Vac flyback design requires transistors that block twice the peak line plus transients or about 1.0 kV. Presently, variations of 1200 to 1500 V horizontal deflection transistors are used here. These bipolar devices are relatively slow (t_f = 200–500 ns) and tend to limit efficient operating frequencies to 20–40 kHz. Introduction of 1000 V TMOS FET will soon permit operation at much higher frequencies. Faster 1.0 kV bipolar transistors are also anticipated in the near future and will provide a lower cost alternative. The two transistor variation of this circuit (Figure 10-3C) eliminates the clamp winding and adds



10-3A. Flyback (Clamp Winding Is Optional)



10-3B. Forward (Clamp Winding Is Necessary)



10-3C. Two Transistor Forward or Flyback (Clamp Winding Is Not Needed)

Figure 10-3. Low Power Popular (20–200 W) Converter Topologies

a transistor and diode to effectively clamp peak transistor voltages to the line. With this circuit a designer can safely use the faster 400 V to 500 V bipolar or FET Switchmode transistors and push operating frequencies considerably higher. There is a cost penalty here over the single transistor circuit due to the extra transistor, diode and floating base drive requirement of the upper switch transistor.

A subtle variation in the method of operation can be applied to either of these circuits. The difference is referred to as operation in the discontinuous or continuous mode, and the waveform diagrams are shown in Figure 10-4. The analysis given in the earlier section on boost regulators dealt strictly with the discontinuous mode where all the energy is dumped from the choke before the transistor turns on again. If the transistor is turned on while energy is still being dumped into the load, the circuit is operating in the continuous mode. This is generally an advantage for the transistor in that it needs to switch only half as much peak current in order to deliver the same power to the load. In many instances, the same transformer may be used with only the gap reduced to provide more inductance. Sometimes the core size will need to be increased to support the higher LI product (2 to 4 times) now required, because the inductance must increase by almost 10 times to effectively reduce the peak current by two. In dealing with the continuous mode, it should also be noted that the transistor must now turn-on from 500 to 600 V rather than 400 V level, because there no longer is any dead time to allow the flyback voltage to settle back down to the input voltage level. Generally it is advisable to have $V_{CEO\ (sus)}$ ratings comparable to the turn-on requirements.

The flyback converter stands out from the others in its need for a low inductance, high current primary. Conventional E and pot core ferrites are difficult to work with because their permeability is too high even with relatively large gaps (50 to 100 mili-inches). The industry needs something better (like powered iron) that will provide permeabilities of 60 to 120 instead of 2000 to 3000 for this application.

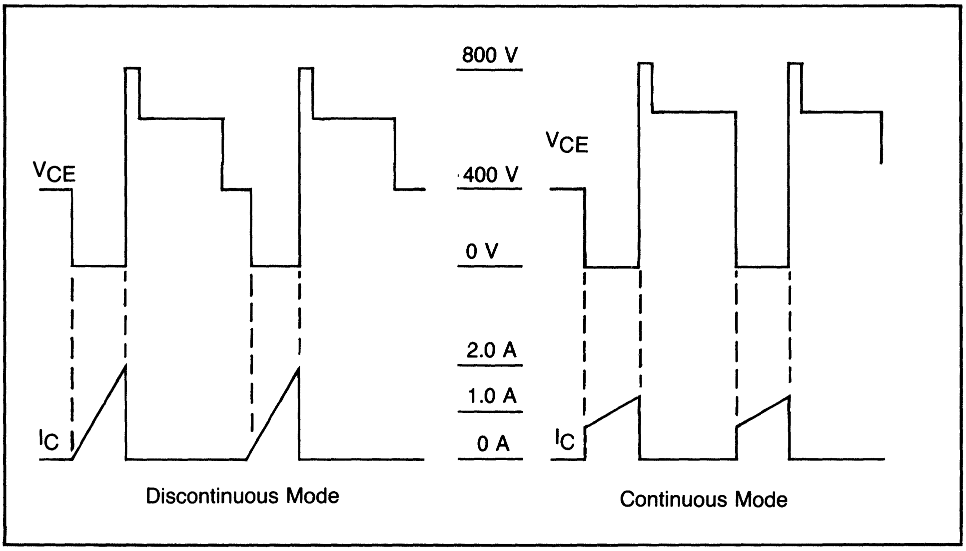


Figure 10-4. Flyback Transistor Waveforms

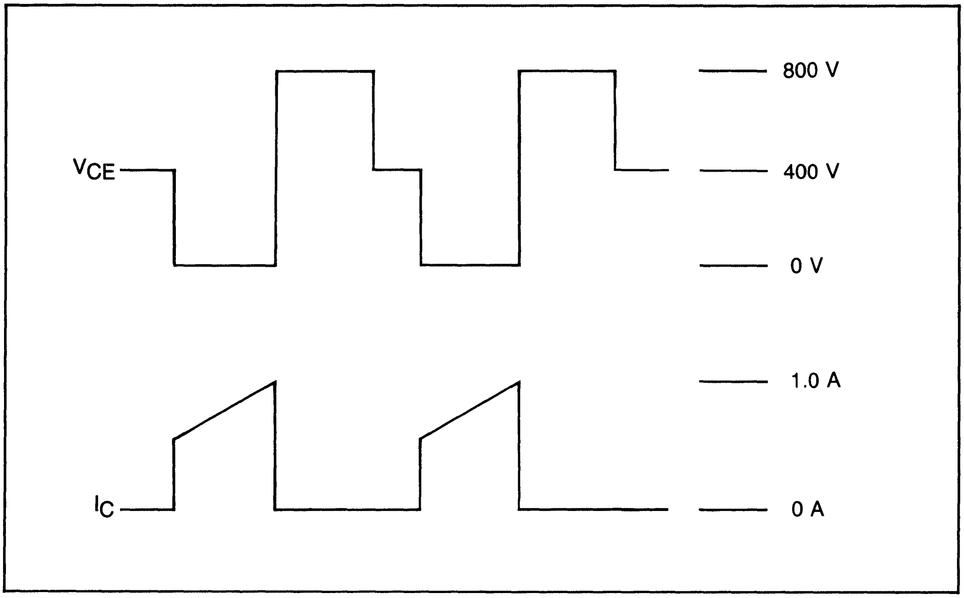


Figure 10-5. Forward Converter Transistor Waveforms

2. Forward — The single transistor forward converter is shown in Figure 10-3B. Although it initially appears very similar to the flyback, it is not. The operating model for this circuit is actually the buck regulator discussed earlier. Instead of storing energy in the transformer and then delivering it to the load, this circuit uses the transformer in the active or forward mode and delivers power to the load while the transistor is on. The additional output rectifier is used as a freewheeling diode from the LC filter, and the third winding is actually a reset winding. It generally has the same turns as the primary (is usually bifilar wound) and clamps the reset voltage to twice the line. However, its main function is to return energy stored in the magnetizing inductance to the line and thereby reset the core after each cycle of operation. Because it takes the same time to set and reset the core, the duty cycle of this circuit cannot exceed 50%. This also is a very popular low power converter, and like the flyback, is practically immune from transformer saturation problems. Transistor waveforms shown in Figure 10-5 illustrate that the voltage requirements are identical to the flyback. For the single transistor versions, 400 V turn-on and 1.0 kV blocking devices like the 1200 to 1500 V deflection transistors are required. The two transistor circuit variation shown in Figure 10-3C again adds a cost penalty, but allows a designer to use the faster 400 to 500 V devices. With this circuit, operation in the discontinuous mode refers to the time when the load is reduced to a point where the filter choke runs "dry." This means that choke current starts at and returns to zero during each cycle of operation. Even though there are no adverse effects on the components themselves, most designers prefer to avoid this type of mode because of higher ripple and noise. Standard ferrite cores work fine here and in the high power converters as well. In these applications, no gap is used as the high permeability (3000) results in a desirable effect of very low magnetizing current levels.

C. PUSH-PULL AND BRIDGE CONVERTERS

The high power circuits shown in Figure 10-6 all operate the magnetic element in the bipolar or push-pull mode and require 2 to 4 inverter transistors. Because the transformers operate in this mode, they tend to be almost half the size of the equivalent single transistor converters and thereby provide a cost advantage over their counterparts at power levels of 100 watts to 1.0 kW.

1. Push-Pull — The push-pull converter shown in Figure 10-6A is one of the oldest converter circuits around. Its early use was in low voltage inverters such as the 12 Vdc to 120 Vdc power source for recreational vehicles and in dc to dc converters. Because these converters are free running rather than driven and operate from low voltages, transformer saturation problems are minimal. In the high voltage off line switchers, saturation problems are common and difficult to solve. The transistors are also subjected to twice the peak line voltage which requires the use of relatively slow 1.0 kV transistors. Both of these drawbacks have tended to discourage designers of off line switchers from using this topology.

2. Half and Full Bridge — The most popular high power converter today is the half bridge (Figure 10-6B). It has two clear advantages over the push-pull type. First, the transistors never see more than the peak line voltage and standard 400 V fast Switchmode transistors that are now readily available may be used. Second, and probably even more important, transformer saturation problems are easily minimized by use of a small coupling capacitor (2.0 $\mu F \leq C_C \leq 5.0 \mu F$) as shown. Because the primary winding is driven in both directions, a full wave output filter, rather than half, is now used, and the core is actually utilized more effectively. Another more subtle advantage of this circuit is that the input filter capacitors are placed in series across the rectified 220 Vac line which allows them to be used as the voltage doubler elements on a 120 Vac line. This allows the inverter transformer to operate from a nominal 320 Vdc bus when the circuit is connected to either 120 Vac or 220 Vac. Finally, this topology allows diode clamps across each transistor to contain destructive switching transients. The designers dream, of course, is for fast transistors that can handle a clamped inductive load line at rated current. And a few (like the Switchmode III and TMOS FET series from Motorola) are beginning to appear on the market. However, the older designs in this area still end up using snubbers to protect the transistor which sacrifices both cost and efficiency.

The effective current limit of today's low cost TO-3 transistors (300 mil die) is somewhere in the 10 to 20 A area. Once this limit is reached, the designer generally changes to the full bridge topology shown in Figure 10-6C. Because full line rather than half is applied to the primary winding, the power output can almost double that of the half bridge with the same switching transistors.

Another variation of the half bridge is the split winding circuit shown in Figure 10-6D. A diode clamp can protect the lower transistor but a snubber or zener clamp must still be used to protect the top transistor from switching transients. Because both emitters are at an ac ground point, expensive drive transformers can now be replaced by lower cost capacitively coupled drive circuits.

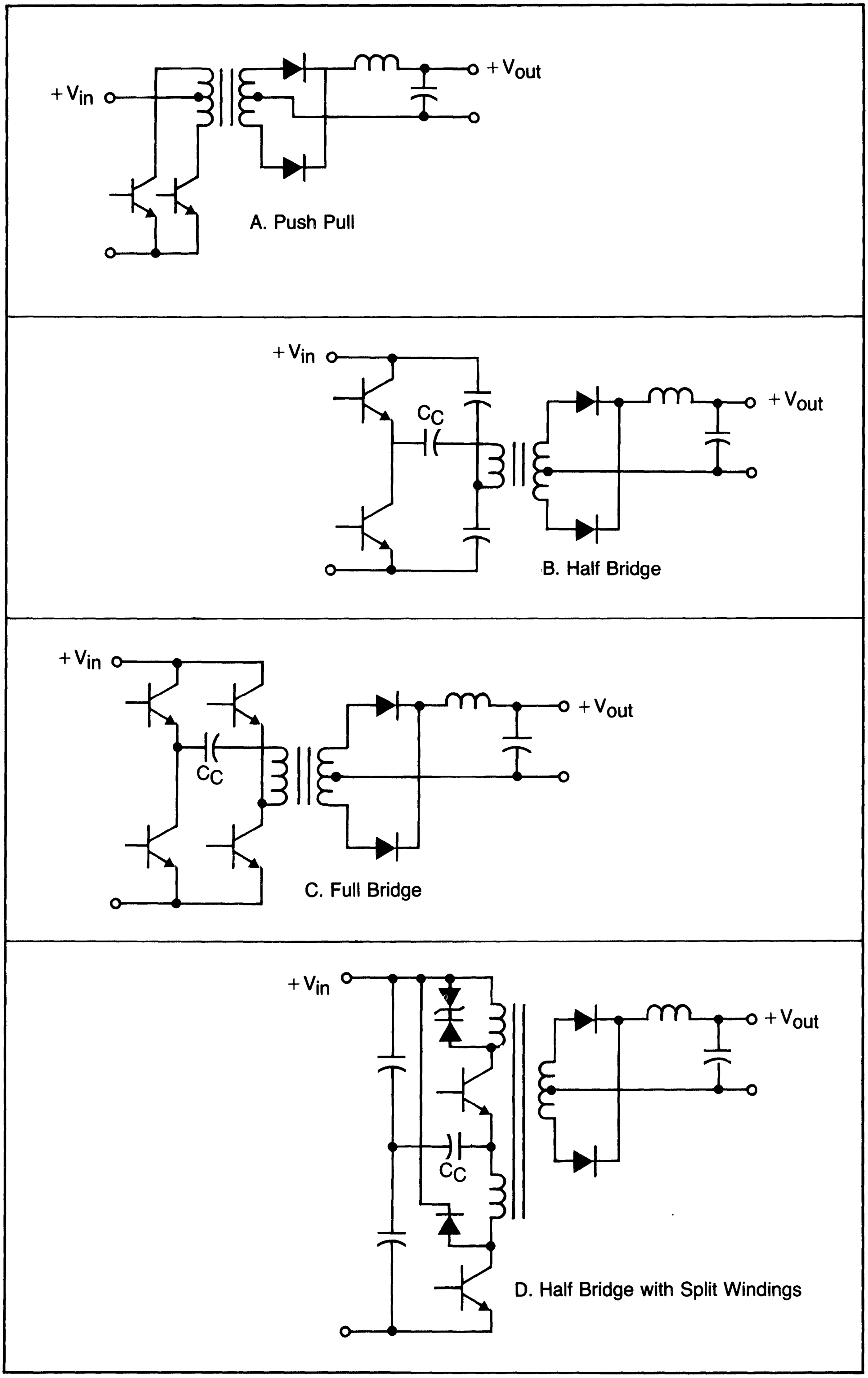


Figure 10-6. High Power Popular Converter Topologies (100 W–1.0 kW)

SECTION 11

SWITCHING REGULATOR COMPONENT DESIGN TIPS

A. TRANSFORMERS

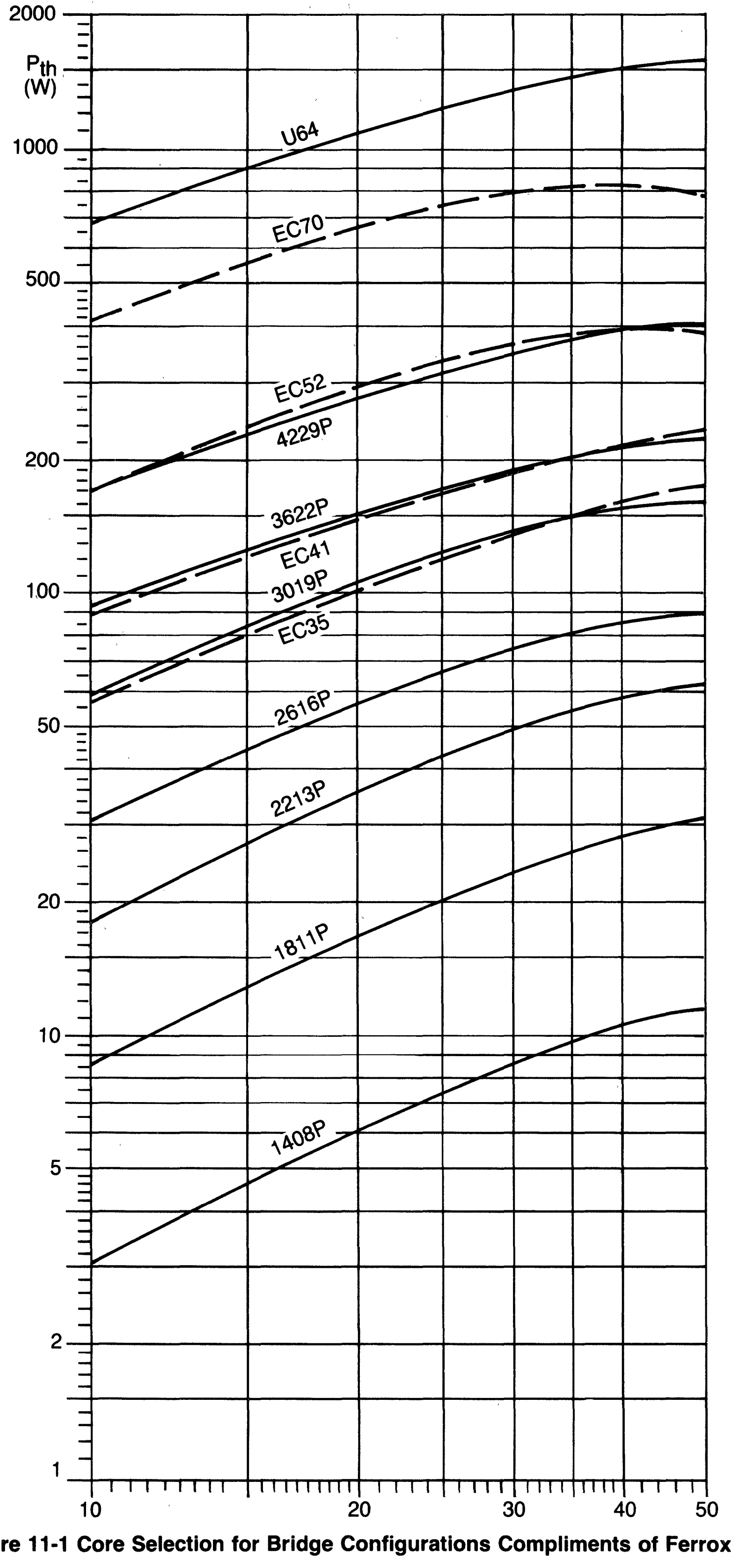
With respect to transformer design, many of today's designers would say don't try it. They'd advise using a consultant or winding house to perform this task, and with good reason. It takes quite a bit of time to develop a "feel" for this craft and be able to use both experience and intuition to find solutions to second and third order problems. Because of these subtle problems, most designers find that after the first paper design is done, as many as four or five lab iterations may be necessary before the transformer meets the design goals. However, there is a considerable design challenge in this area and a great deal of satisfaction can be obtained by mastering it.

As do all others, this component design begins by requesting all available literature from the appropriate manufacturers, and then following up with phone calls when specific questions arise. A partial list of companies is shown in Table 11-1. Designs below 50 W generally use pot cores, but for 50 W and above E cores are preferred. E cores expose the windings to air so that heat is not trapped inside. The exposure also makes it easier to bring out connections for tapped windings. Remember that flyback designs require lower permeability cores than the others. The classic approach is to consult manufacturers charts like the one shown in Figure 11-1 and then pick a core with the required power handling ability. Both E and E-C (E cores with a round center leg) are popular now, and they are available from several manufacturers. E-C cores offer a performance advantage (better coupling) but standard E cores cost less and are also used in these applications. Another approach that seems to work equally as well is to do a paper design of the estimated windings and number of turns required. Size the wire for 500 circular mils (CM) per ampere and then find a core that has the required window area for this design. Now, before the windings are put on, it is a good idea to modify the turns so that they fit on the bobbin in an integer number of layers. This involves checking the turns per inch of wire against the bobbin length. The primary generally goes on first and then the secondary windings. If the primary hangs over an extra half layer, try reducing the turns or the wire size. Conversely, if the secondary does not take up a full layer, try bifilar winding (parallel) using wire half the size originally chosen; i.e., 3 wire sizes smaller like 23 versus 20. This technique ultimately results in the use of foil for the higher current (20 A) low voltage windings. Most windings can be separated with 3 mil mylar (usually yellow) tape, but for good isolation, cloth is recommended between primary and secondary.

TABLE 11-1

Partial List of Core (C) and Transformer (T) Manufacturers

Company	Location	Code
Ferroxcube Inc.	Sauggerties, N.Y.	C
Indiana General	Keasby, N.J.	C
Stackpole	St. Marys, PA.	C
TDK	El Segundo, CA.	C
Pulse Engineering	San Diego, CA.	T
Coilcraft	Cary, IL.	T



Note: Power handling decreases by a factor of 2 in forward and 4 in flyback configurations.

Figure 11-1 Core Selection for Bridge Configurations Compliments of Ferroxcube

Finally, once a mechanical fit has been obtained, it is time for the circuit tests. The voltage rating is strictly a mechanical problem and is one of the reasons why U.L. normally does not allow high voltage bifilar windings. The inductance and saturating current level of the primary are inherent to the design, and should be checked in the circuit or other suitable test fixture. Such a fixture is shown in Figure 11-2 where the transistor and diode are sized to handle the anticipated currents. The pulse generator is run at a low enough duty cycle to allow the core to reset. Pulse width is increased until the start of saturation is observed (I_{sat}). Inductance is found using

$$L = V\frac{di}{dt}$$

In forward converters, the transformer generally has no gap in order to minimize the magnetizing current (I_M). For these applications the core should be chosen to be large enough so that the resulting LI product insures that I_M at operating voltages is less than I_{sat}. For flyback designs, a gap is necessary and the test circuit is useful again to evaluate the effect of the gap. The gap will normally be quite large where:

$L_g >> L_m/\mu$

$L_g =$ gap length

$L_m =$ magnetic path length

$\mu =$ permeability

Under this stipulation, the gap directly controls the LI parameters. Doubling it will decrease L by two and increase I_{sat} by two. Again, the anticipated switching currents must be less than I_{sat} when the core is gapped to ensure correct inductance.

Transformer tests in the actual supply are usually done with a high voltage dc power supply on the primary and with a pulse generator or other manual control for the pulse width drive such as using the control IC in an open loop configuration.

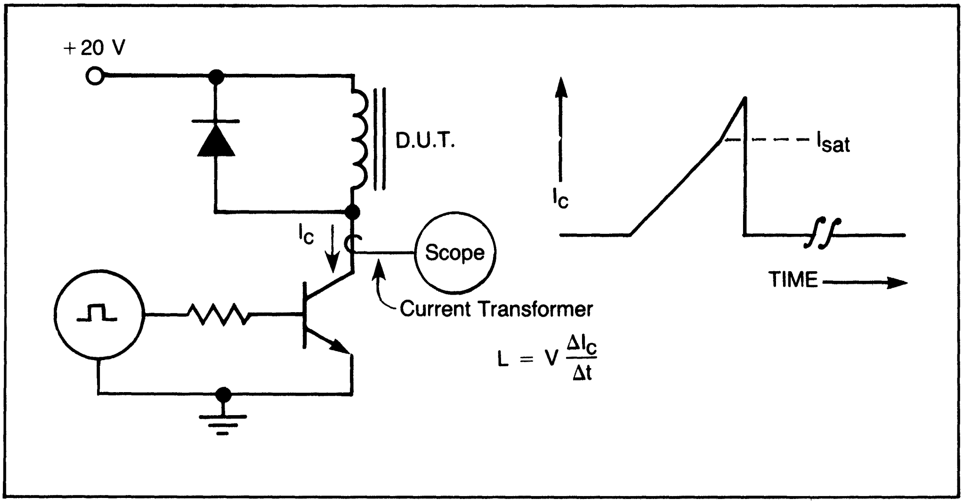


Figure 11-2. Simple Coil Tester

Here the designer must recheck three areas:

1. No evidence of core saturation
2. Correct amount of secondary voltage
3. Minimum core or winding heat rise

If problems are detected in any of these areas, one possible solution is to redesign using the next larger core size. However, if problems are minimal, or none exist, it is possible to stay with the same core or even consider using the next smaller size.

B. TRANSISTORS

The initial selection of a transistor(s) for a switcher is basically a problem of finding the one with voltage and current capabilities that are compatible with the application. For the final choice, performance and cost tradeoffs among devices from the same or several manufacturers have to be weighed. Before these devices can be put in the circuit, both protective and drive circuits will have to be designed.

Motorola's first line of devices for switchers were trademarked ''Switchmode'' transistors and introduced in the early 70's. Data sheets were provided with all the information that a designer would need, including reverse bias safe operating area (RBSOA) and performance at elevated temperature (100°C). The first series was the 2N6542 through 6547, TO-3 devices which were followed by the MJE13004 series in a plastic TO-220 package. Finally, high voltage (1.0 kV) requirements were met by the metal MJ12002 and MJ8500 series and the plastic MJE12007. Just recently, Motorola introduced three new families of ''Switchmode'' transistors shown in Table 11-2. The Switchmode II series is basically a faster switching version of Switchmode I. Switchmode III is the Cadillac of today's industry with both exceptional speed and RBSOA. Here, device cost is up but system costs may be lowered because of reduced snubber requirements and higher operating frequencies. A similar argument applies to Motorola T-MOS FET's. These devices make it possible to switch efficiently at higher frequencies (200 to 500 kHz), but the main selling point is that they are easier to drive. This latter point is the one most often made to show that systems savings are again quite possible even though the initial device cost is higher.

TABLE 11-2

Motorola High Voltage Switching Transistor Technologies

Family	Typical Device	Typical Fall Time	Approximate Switching Frequency
SWITCHMODE I	2N6545 MJE13005 MJE12007	200–500 ns	20K
SWITCHMODE II	MJ12010	100 ns	100K
SWITCHMODE III	MJ13010	50 ns	200K
T-FET'S	MTP565	20 ns	500K

TABLE 11-3

Power Transistor Voltage Chart

Line Voltage	Circuit			
	Flyback, Forward or Push-Pull		Half or Full Bridge	
	V_{CEV}	$V_{CEO(sus)}$	$V_{CEO(sus)}$	V_{CEV}
220	850	400	400	400
120	450	200	200	200

Table 11-3 is a review of the transistor voltage requirements for the various off line converter circuits. As illustrated, the most stringent requirement for single transistor circuits (flyback and forward) is the blocking or V_{CEV} rating. Bridge circuits, on the other hand, turn on and off from the dc bus and their most critical voltage is the turn on or $V_{CEO\ (sus)}$ rating. To help designers select parts for these applications, Motorola has provided the selection charts in Appendix A. Each table lists devices that are appropriate for a given line voltage and circuit configuration and various power handling capabilities. Table 1 contains devices listed by their current (power handling) rating and $200 < V_{CEO} < 400$ V for use in 120 Vac bridge circuits. Tables 2 and 3 list the remaining devices ($V_{CEO} \geqslant 400$ V) which would be appropriate for 220 Vac and 380 Vac bridge circuits. Tables 4 and 5 list devices by their V_{CEV} rating. These tables can therefore be used to select devices for either 120 or 220 Vac single transistor circuits (flyback and forward converters).

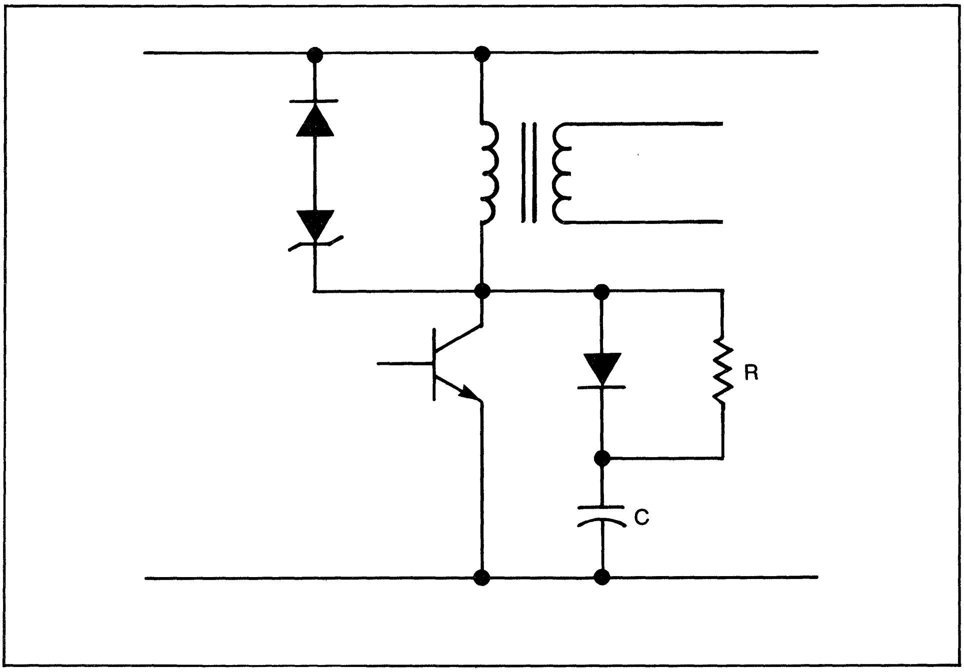


Figure 11-3. Zener Clamp and Snubber for Single Transistor Converters

Most Switchmode transistor load lines are inductive during turn on and turn off. Turn on is generally inductive because the short circuit created by output rectifier reverse recovery times is isolated by leakage inductance in the transformer. This inductance effectively snubs most turn-on load lines so that the rectifier recovery (or short circuit) current and the input voltage are not applied simultaneously to the transistor. Sometimes primary interwinding capacitance presents a small current spike, but usually turn-on transients are not a problem. Turn-off transients due to this same leakage inductance, however, are almost always a problem. In bridge circuits, clamp diodes can be used to limit these voltage spikes. If the resulting inductive load line exceeds the transistor's reverse bias switching capability (RBSOA) then an RC network may also be added across the primary to absorb some of this transient energy. The time constant of this network should equal the anticipated switching time of the transistor (100 ns to 1 μs). Resistance values of 100 to 1000 ohms in this RC network are generally appropriate. Trial and error will indicate how low the resistor has to be to provide the correct amount of snubbing. For single transistor converters, the snubber shown in Figure 11-3 is generally used. Here slightly different criteria are used to define the R and C values:

$$C = \frac{I\ t_f}{V}$$

where I = The peak switching current

t_f = The transistor fall time

V = The peak switching voltage

(Approximately twice the dc bus)

also R = t_{on}/C (it is not necessary to completely discharge this capacitor to obtain the desired effects of this circuit)

where t_{on} = The minimum on time or pulse width

and $$P_R = \frac{CV^2f}{2}$$

where P_R = The power rating of the resistor

and f = The operating frequency

Most of today's transistors that are used in 20 kHz converters switch slow enough so that most of the energy stored in the leakage inductance is dissipated by the snubber or transistor, causing very little voltage overshoot. Higher speed converters and transistors present a slightly different problem. In these newer designs, snubber elements are smaller and voltage spikes from energy left in the leakage inductance may be a more critical problem depending on how good the coupling is between the primary and clamp windings. If necessary, protection from these spikes may be obtained by adding a zener and rectifier across the primary as shown in Figure 11-3. Motorola's 1.0 W and 5.0 W zener devices with ratings up to 200 V can provide the clamping or spike limiting function. If the zener must handle most of the power, its size can be estimated using:

$$P_Z = \frac{L_L I^2 f}{2}$$

where P_Z = The zener power rating

and L_L = The leakage inductance (measured with the clamp winding or secondary shorted)

There are probably as many base drive circuits for bipolars as there are designers. Ideally, the transistor should have just enough forward drive (current) to stay in or near saturation and reverse drive that varies with the amount of

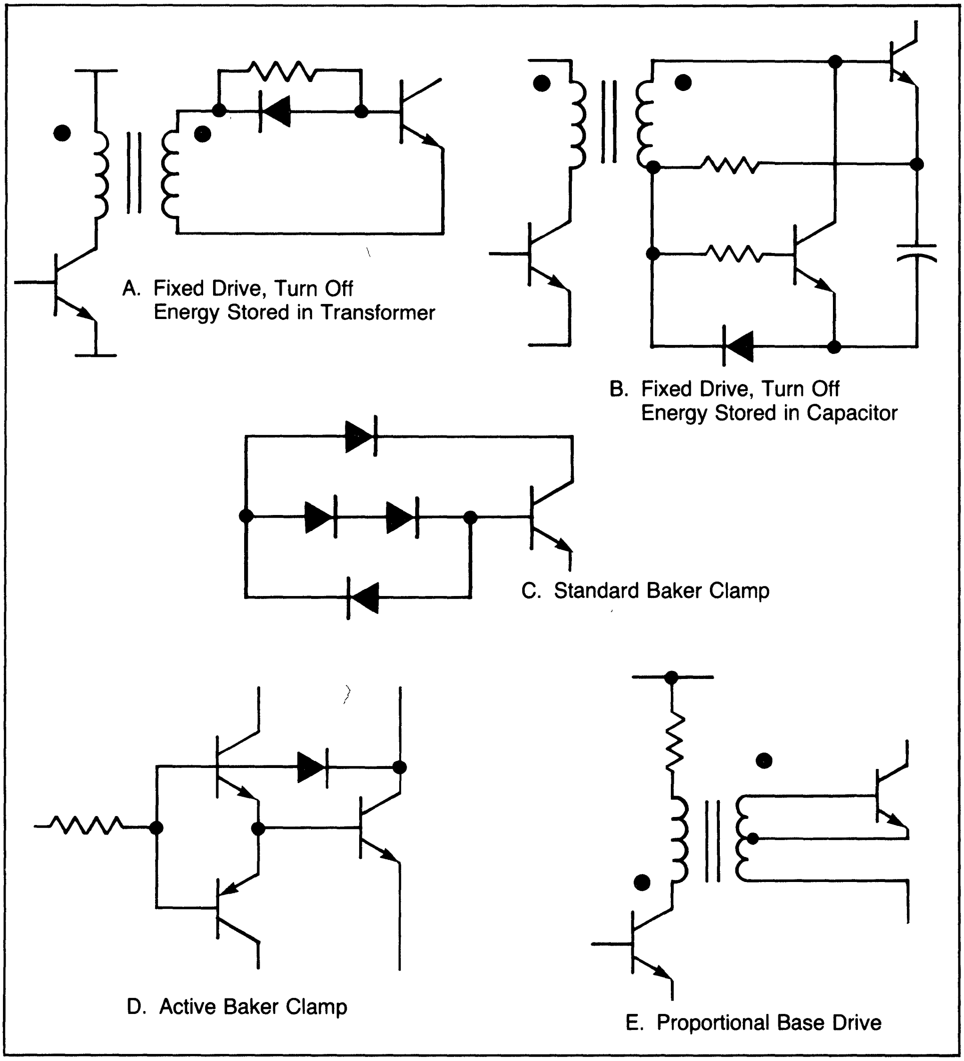


Figure 11-4. Typical Bipolar Base Drive Circuits

stored base charge such as a low impedance reverse voltage. Many of today's common drive circuits are shown in Figure 11-4. The fixed drive circuits of 11-4A and 11-4B tend to emphasize economy, while the Baker clamp and proportional drive circuits of 11-4C, 11-4D and 11-4E emphasize performance over cost.

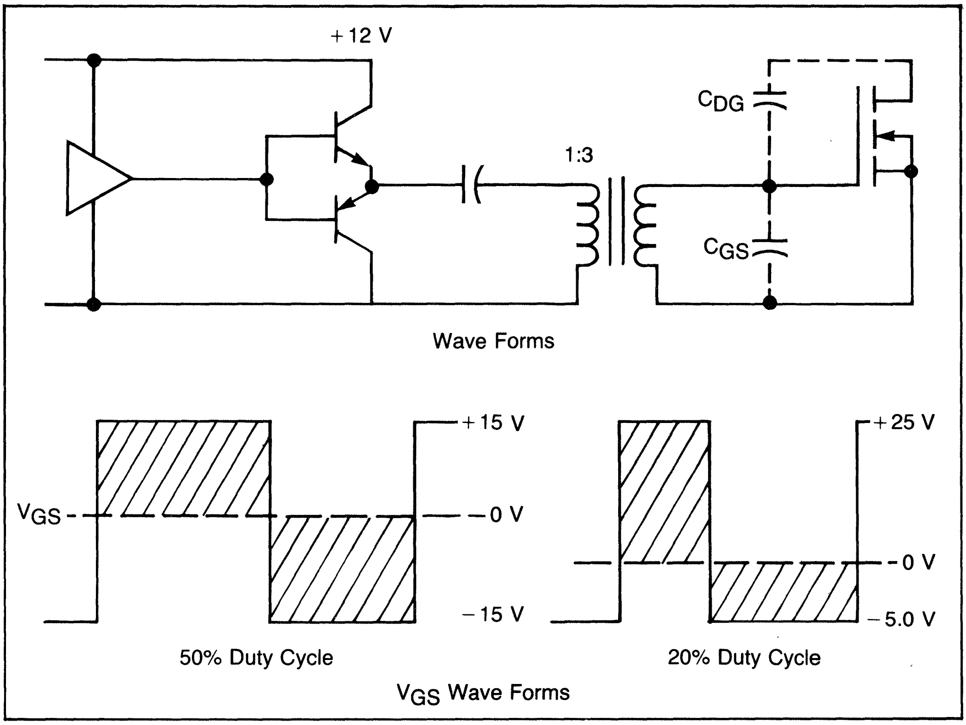


Figure 11-5A. Typical Transformer Coupled FET Drive

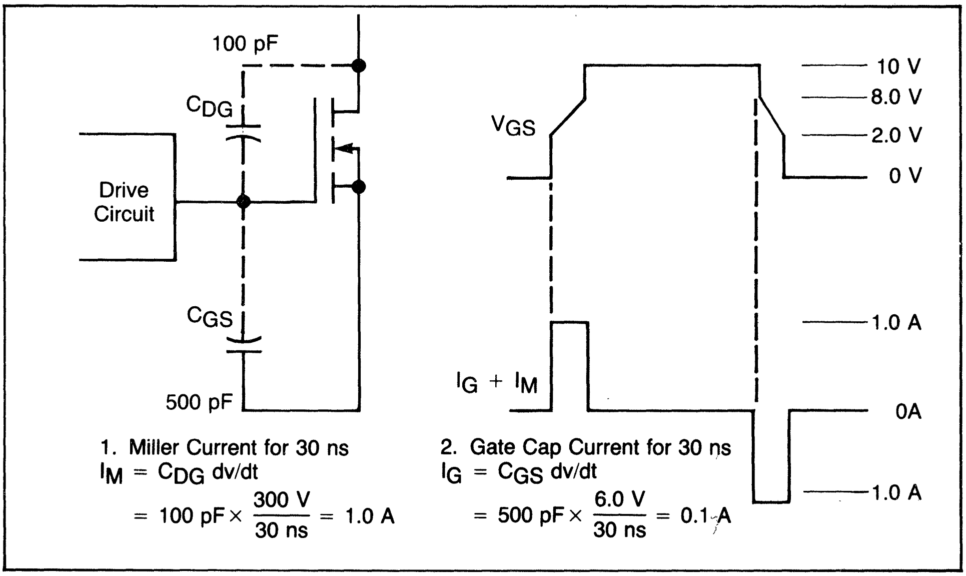


Figure 11-5B. FET Drive Current Requirements

FET drive circuits are just beginning to appear. The standard that has evolved at this time is shown in Figure 11-5A. This transformer coupled circuit will produce forward and reverse voltages applied to the FET gate which vary with the duty cycle as shown. For this example, a V_{GS} rating of 20 V would be adequate for one condition, but not the other. Higher V_{GS} ratings would solve the problem, but at this time it is advisable to use a regulated logic supply and provide only the minimum gate drive required for these situations. Finally, there is one point that is not obvious when looking at the circuit. It turns out that FET's can be directly coupled to many IC's with only to 100 mA of sink and source output capability and still switch efficiently at 20 kHz. However, to switch efficiently at higher frequencies, several amperes of drive may be required on a pulsed basis in order to quickly charge and discharge the gate capacitances. A simple example will serve to illustrate this point and also show that the Miller effect, produced by C_{DG}, is the predominant speed limitation when switching high voltages (see Figure 11-5B). A FET responds instantaneously to changes in gate voltage and will begin to conduct when the threshold is reached ($V_{GS} = 2.0$ to 3.0 V) and be fully on with $V_{GS} = 7.0$ to 8.0 V. Gate waveforms will show a step at a point just above the threshold voltage which varies in duration depending on the amount of drive current available. The drive current determines both the rise and fall times for the drain current. To estimate drive current requirements, two simple calculations with gate capacitances can be made:

1. $I_M = C_{DG}dv/dt$

and 2. $I_G = C_{GS}dv/dt$

where I_M is the current required by the Miller effect to charge the drain to gate capacitance at the rate it is desired to move the drain voltage (and current). And I_G is usually the lesser amount of current required to charge the gate to source capacitance through the linear region (2.0 to 8.0 V). As an example, if 30 ns switching times are desired at 300 V where $C_{DG} = 100$ pF and $C_{GS} = 500$ pF, then

$$I_M = 100 \text{ pF} \times 300 \text{ V/30 ns} = 1.0 \text{ A and}$$

$$I_G = 500 \text{ pF} \times 6.0 \text{ V/30 ns} = 0.1 \text{ A}$$

This example shows the direct proportion of drive current capability to speed. It also illustrates that for most devices, C_{DG} will have the greatest effect on switching speed and that C_{GS} is important only in estimating turn on and turn off delays.

Aside from rather unique drive requirements, a FET is very similar to a bipolar transistor. Today's 400 V FET's compete with bipolar transistors in many switching applications. They are faster and easier to drive, but do cost more and have higher saturation, or more precisely, on voltages. The performance or efficiency tradeoffs are best analyzed using Figure 11-6. Here, typical power losses for 5.0 A switching transistors versus frequency are shown. The FET and bipolar losses were calculated at $T_J = 100°C$ rather than 25°C because on resistance and switching times are highest here, and 100°C is typical of many applications. These curves are asymptotes of the actual device performance, but are useful in establishing the "break point" of various devices, which is the point where

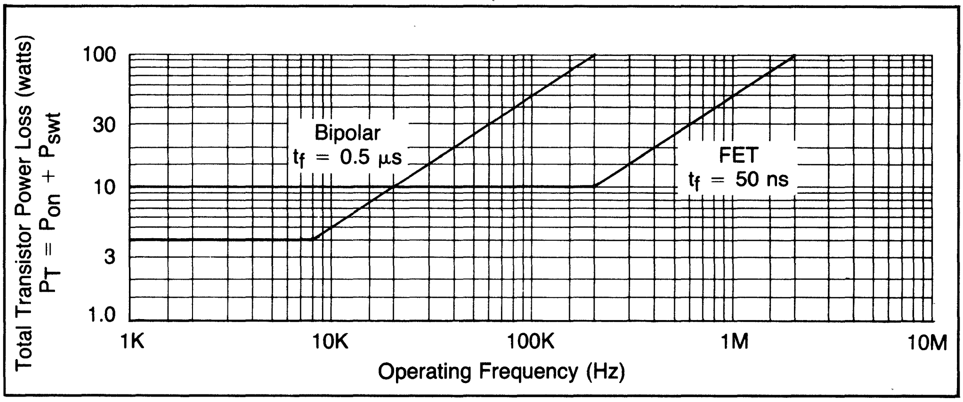


Figure 11-6. Typical Switching Losses at 5.0 A and T_J = 100°C

saturation and switching losses are equal. Since this is as low as 10 kHz for some bipolars, it is possible that a FET even with high on voltages can be competitive efficiency-wise at 20 kHz. The faster Switchmode II and III bipolar products fall somewhere between the curves shown and therefore are more competitive with FET's at the higher operating frequencies.

C. RECTIFIERS

Once components for the inverter section of a switcher have been chosen, it is time to determine how to get power into and out of this section. This is where the all important rectifier comes into play. The input rectifier is generally a bridge that operates off the ac line and into a capacitive filter. For the output section, most designers use Schottkys for efficient rectification of the low voltage, 5.0 V output windings, and for the higher voltage (12 to 15 V) outputs, the more economical fast recovery diodes are used. A guide to Motorola's rectifier products is given in Appendix B. Here devices that would normally be used in switchers from 10 to 2000 watts are listed next to circuits in which they would generally be used.

For the process of choosing an input rectifier, it is useful to visualize the circuit shown in Figure 11-7. To reduce cost, most earlier approaches of using choke input filters, soft start relays (Triacs), or SCR's to bypass a large limiting resistor have been abandoned in favor of using small limiting resistors or NTC thermistors, and a large bridge. The bridge must be able to withstand the surge currents that exist from repetitive starts at peak line. The procedure for finding the right component and checking its fit is as follows:

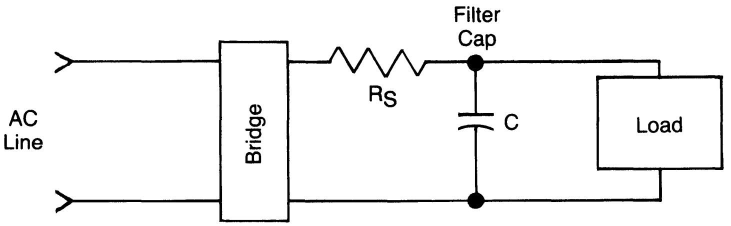
1. Choose a rectifier with 2 to 5 times the average I_O required.
2. Estimate the peak surge current (I_p) and time (t) using:

$$I_p = \frac{1.4V_{in}}{R_S} \qquad t = R_S C$$

Where V_{in} is The RMS input voltage

R_S = the total limiting resistance, and

C = the filter capacitance



Steps:

1. Choose a rectifier with an I_O rating of 2 to 5 times the actual load.
2. Measure or calculate the inrush current at peak line voltage.
3. Compare to the equivalent diode rating using I_{FSM} and $I^2\sqrt{t} = K$.
4. If line 3 is less than line 2, use a larger rectifier or increase R_S.

Figure 11-7. Choosing Input Rectifiers

3. Compare this current pulse to the sub cycle surge current rating (I_S) of the diode itself. If the curve of I_S versus time is not given on the data sheet, the approximate value for I_S at a particular pulse width (t) may be calculated knowing:

   - I_{FSM} — the single cycle (8.3 ms) surge current rating.
   - $I^2 \sqrt{t} = K$ which applies when the thermal response, r(t), is proportional to $\sqrt{t}$ (for t<8.3 ms). This gives:

   $$I_s^2 \sqrt{t} = I^2_{FSM}\sqrt{8.3 \text{ ms}} \text{ or}$$

   $$I_s = I_{FSM}\left(\frac{8.3 \text{ ms}}{t}\right)^{1/4} \quad \text{(t is in milliseconds)}$$

4. If $I_S < I_P$, consider either increasing the limiting resistor (R_S) or utilizing a larger diode.

In the output section where high frequency rectifiers are needed, there are several types available to the designer. In addition to the Schottky (SBR) and fast recovery (FR), there is also an ultra fast recovery (UFR) which fills the gap between the 50 V Schottky and the 600 V fast recovery lines. Comparative performance and cost data for devices with similar current ratings is shown in Table 11-4. The obvious point here is that lower forward voltage improves efficiency and faster recovery times reduces turn-on losses in the switching transistors, but the tradeoff is higher cost. As stated earlier, Schottkys are generally used for 5.0 V outputs and fast recovery devices for ≥12 V outputs. The ultra fast is competing primarily with the Schottky in those applications where cost is more important than efficiency. Of these devices, only the Schottky may need special handling. Ten years ago Schottkys were very fragile and could fail short from either excessive dv/dt (1.0 to 5.0 volts per nano-second) or reverse avalanche. Present day devices, however, all have something similar to Motorola's "guard ring" and internal zener, which minimizes these earlier problems and reduces the need for RC snubbers and other external protective networks.

TABLE 11-4

Output Rectifier Type Comparisons

	SBR	UFR	FR
V_F	0.5–0.6	0.9–1.0	1.2–1.4
t_{rr}	10 ns	25 ns	150 ns
t_{rr} FORM	"SOFT"	"ABRUPT"	"EITHER"
V_R	30–50 V	50–150 V	50–600 V

NOTES: 1. Low V_F improves efficiency
2. Low t_{rr} reduces transistor switching losses
3. Soft (verses abrupt) recovery reduces noise

D. CAPACITORS AND FILTERS

In today's 20 kHz switchers, aluminum electrolytics are still predominate. The good news is that most have been characterized, improved, and cost reduced for this application. The input filter requires a voltage rating that depends on the peak line voltage; i.e., 400 to 450 V for a 220 Vac switcher. If voltage is increased beyond this point, the capacitor will begin to act like a zener and be thermally destroyed from high leakage currents if the rating is exceeded for enough time. When filter capacitors are placed in series across the rectified line, as in a doubler circuit, voltage sharing can be a problem. Here extra voltage capability may be needed to make up for the imbalances caused by different values of capacitance and leakage current. A bleeder resistor is normally used here not only for safety but to mask the differences in leakage current. The RMS current rating is also an important consideration for input capacitors and is an example of improvements offered by today's manufacturers. Earlier "lytics" usually lacked this rating and often overheated. Large capacitors that were not needed for performance were used just to reduce this heating. However, today's devices, like the swedged variety from Mepco-Electra offer lower thermal resistance, improved connection to the foil and good RMS ratings. A partial list of manufacturers that supply both high voltage input and the lower voltage output capacitors for switchers is shown in Table 11-5. Most of the companies offer not only the standard 85°C components, but devices with up to 125°C ratings, which are required because of the high ambient temperatures (55 to 85°C) in which switchers must operate, many times without the benefit of fans.

TABLE 11-5

Partial List of Capacitor Companies

Company (U.S.)	Location
Sprague	North Adams, MA
Mepco/Electra	Columbia, SC
Cornell-Dublier	Sanford, NC
Sangamo	Pickens, SC
Mallory	Indianapolis, IN

For output capacitors the buzz word is low ESR (equivalent series resistance). It turns out that for most capacitors even in the so-called "low ESR" series, the output ripple depends more on this resistance than on the capacitor value itself. Although typical and maximum ESR ratings are now available on most capacitors designed for switchers, the lead inductance generally is not specified except for the ultra-high frequency four-terminal capacitors from some vendors. This parameter is responsible for the relatively high switching spikes that appear at the output. However, at present, most designers find it less costly and more effective to add a high frequency noise filter rather than use a relatively expensive capacitor with low equivalent series inductance (ESL).

High frequency noise or spike filters are made using small powdered iron toroids (1/2 to 1" OD) with distributed windings to minimize interwinding capacitance. The output is bypassed using a small 0.1 μF ceramic or a 10 to 50 μF tantalum or both. Larger powered iron toroids are often used in the main LC output filter, although the higher permeability ferrite C and E cores with relatively large gaps can also be used. Calculations for the size of this component should take into account the minimum load so that the choke will not run "dry" as stated earlier.

E. CONTROL CIRCUITS

Ten years ago, discrete control circuits were in use and very few IC's could be found. Since that time, various semiconductor companies recognized the designer's needs for a dedicated control IC. Now a variety of these circuits are on the market and widely used. They provide the designer with a cost incentive over the discrete, or a simpler control circuit, or both. Internally, most of these resemble the functional configuration shown in Figure 11-8. The basic regulating function is performed in the pulse width modulator (PWM) section. Here, the dc feedback signal is compared to a fixed frequency sawtooth (or triangular) waveform. The result is a variable duty cycle pulse train which, with suitable buffer or interface circuits, can be used to drive the power switching transistor(s). Some IC's provide only a single output while others provide the phase splitter shown to alternately pulse two output channels. In this latter case, provisions are usually made either internally or by wire "OR"-ing the outputs to convert the dual output to a single output channel. Additionally, most IC's provide the error amplifier section shown as a means to process, compare and amplify the feedback signal.

TABLE 11-6

Desirable Features of Switchmode Control IC's

- PROGRAMMABLE (TO 500 kHz) FIXED FREQUENCY OSCILLATOR
- LINEAR PWM SECTION WITH DUTY CYCLE FROM 0 TO 100 %
- ON BOARD ERROR AMPLIFIERS
- ON BOARD REFERENCE REGULATOR
- ADJUSTABLE DEAD TIME
- UNDERVOLTAGE (LOW V_{CC}) INHIBIT
- GOOD OUTPUT DRIVE (100 TO 200 mA)
- OPTION OF SINGLE OR DUAL CHANNEL OUTPUT
- UN-COMMITTED OUTPUT COLLECTOR AND EMITTER OR TOTEM POLE DRIVE CONFIGURATION
- SOFT START
- CURRENT LIMITING WITH "HICCUP MODE" AS BACKUP
- SYNC CAPABILITY

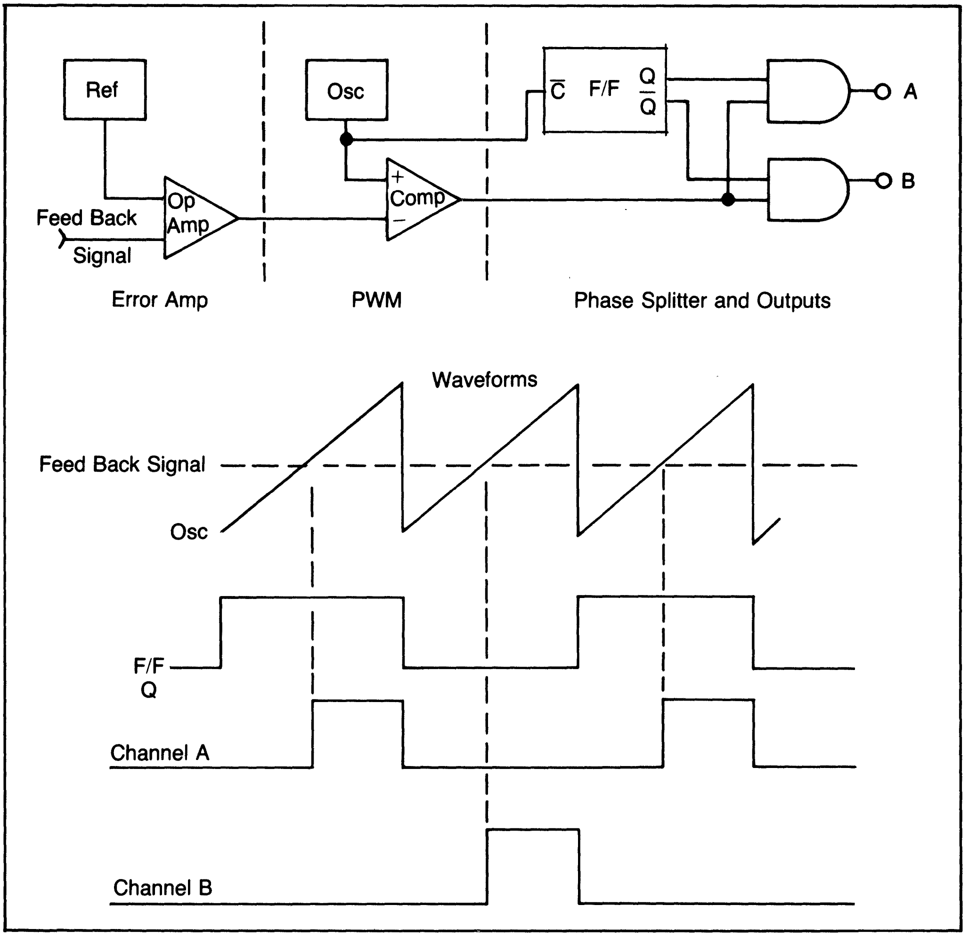


Figure 11-8. Basic Pulse Width Modulator Control IC

Features required by a control IC vary to some extent because of the particular needs of a designer and on the circuit topology chosen. However, most of today's current generation IC's have evolved with the capabilities or features listed in Table 11-6. It is primarily the cost differences in these parts that determines whether all or only part of these features will be incorporated. Most of these are evident to the designer who has already started comparing data sheets. A selector guide of control IC's available from Motorola is shown in Table 17-4 on page 160.

Because low cost and second sources are important, parts like the TL494 (available from Motorola) have already captured a large share of the market. New products such as the SG1525A/27A and SG1526 are quickly gaining popularity. These devices offer additional features like totem pole outputs and digital current limiting and are available from Motorola.

To satisfy the need for a low cost control IC for low power (20 to 100 W) applications, Motorola has introduced a single channel Control IC known as the MC34060.

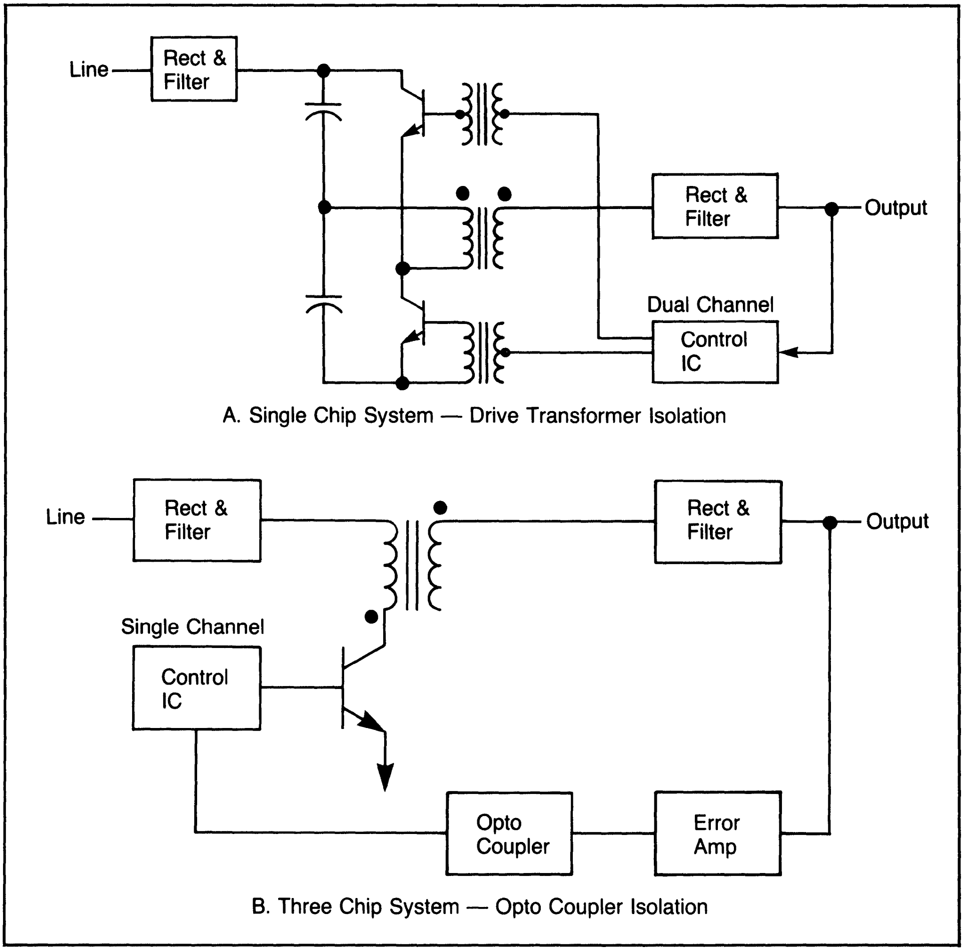


Figure 11-9. Control Circuit Topologies

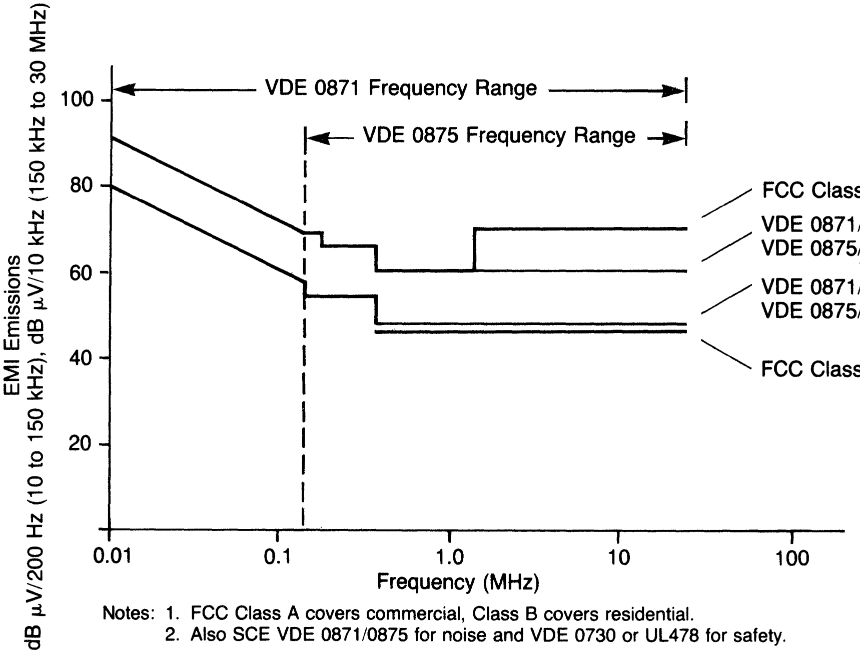
When it is necessary to drive two or more power transistors, drive transformers are a practical interface element and are driven by the conventional dual channel IC just discussed (Figure 11-9A). In the case of a single transistor converter, however, it is usually more cost effective to directly drive the transistor from the IC (Figure 11-9B). In this situation, an opto coupler is commonly used to couple the feedback signal from the output back to the control IC. And the error amplifier in this case is nothing more than an op amp, and reference such as the TL431 from Motorola.

SECTION 12

THE FUTURE FOR SWITCHING REGULATORS

The future offers a lot of growth potential for switchers in general — and low power switchers (50–200 watts) in particular. The latter are responding to the growth in microprocessor-based equipment, as well as computer peripherals. Today's topologies have already been challenged by the sine wave inverter, which reduces noise and improves transistor reliability, but results in a cost penalty. Also, a trend has begun toward higher switching frequencies to further reduce size and cost. The latest bipolar transistor can operate efficiently up to 100 kHz, and the FET seems destined to own the 200 to 500 kHz range.

The growth pattern predicted at this time can possibly be impacted by noise problems. Originally governed only by MIL specs and the VDE in Europe, the FCC (effective October 1981) has released a set of specifications that apply to electronic systems which often include switchers (see FCC Class A in Figure 12-1). It seems probable, however, that system engineers or power supply designers will be able to add the necessary line filters and EMI shields without adding a significant cost.



Notes: 1. FCC Class A covers commercial, Class B covers residential.
2. Also SCE VDE 0871/0875 for noise and VDE 0730 or UL478 for safety.

Figure 12-1. Noise Limits

The most optimistic note concerning switchers is in the components area. Switching power supply components have actually evolved from components used in similar applications. And it is very likely that newer and more mature products specifically for switchers will continue to appear over the next several years. The ultimate effect of this evolution will be to further simplify and cost reduce these designs. Because the designer and component manufacturer must work as a team to bring this about, companies like Motorola that are looking to the future will continue a dialogue with designers to keep abreast with their current and future product needs.

SECTION 13

SWITCHING REGULATOR DESIGN EXAMPLES

Three switching regulator power supply designs are covered in this section. Part A describes a 400 W half bridge and a 1000 W full bridge configuration in which the TL494 control I.C. is utilized. Part B describes a 60 W flyback regulator where a MC34060 control I.C. is used. All three design examples are off-line supplies which can operate from either 115 or 230 Vac.

A. A SIMPLIFIED POWER-SUPPLY DESIGN USING THE TL494 CONTROL CIRCUIT

The TL494 is a fixed-frequency pulse width modulation control circuit, incorporating the primary building blocks required for the control of a switching power supply. (See Figure 13-1.) An internal-linear sawtooth oscillator is frequency-programmable by two external components, R_T and C_T. The oscillator frequency is determined by:

$$f_{osc} \cong \frac{1.1}{R_T C_T}$$

Output pulse width modulation is accomplished by comparison of the positive sawtooth waveform across capacitor C_T to either of two control signals. The NOR

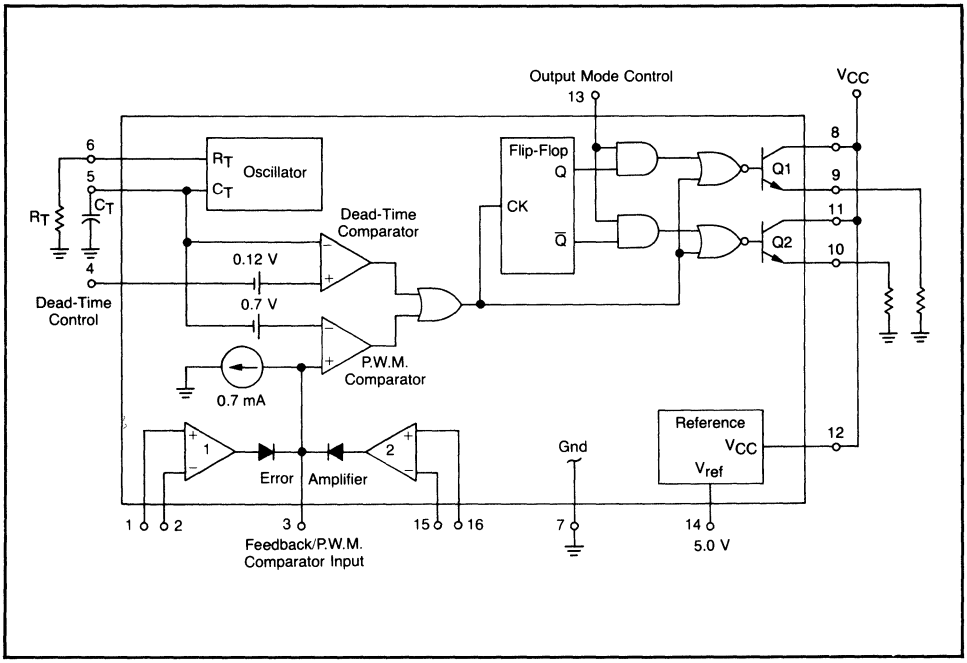


Figure 13-1. TL494 Block Diagram

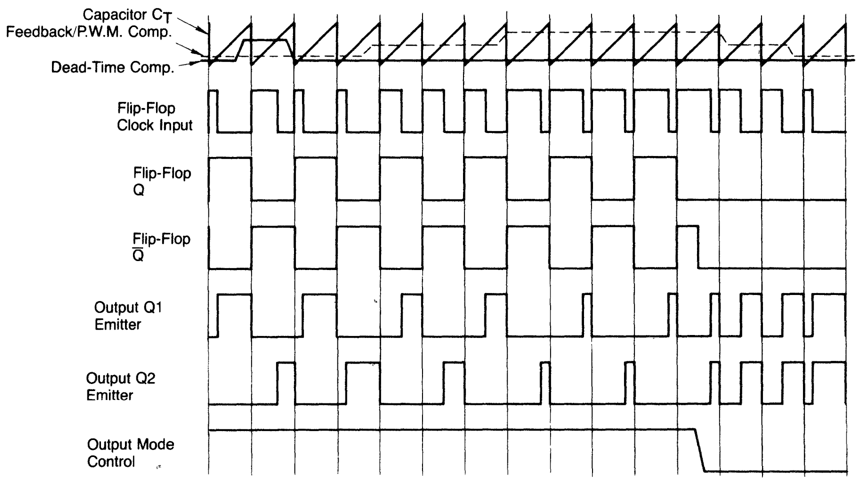


Figure 13-2. TL494 Timing Diagram

gates, which drive output transistors Q1 and Q2, are enabled only when the flip-flop clock-input line is in its low state. This happens only during that portion of time when the sawtooth voltage is greater than the control signals. Therefore, an increase in control-signal amplitude causes a corresponding linear decrease of output pulse width. (Refer to the timing diagram shown in Figure 13-2.)

The control signals are external inputs that can be fed into the dead-time control (Figure 13-1, Pin 4), the error amplifier inputs (pins 1, 2, 15, 16), or the feedback input (Pin 3). The dead-time control comparator has an effective 120 mV input offset which limits the minimum output dead time to approximately the first 4% of the sawtooth-cycle time. This would result in a maximum duty cycle of 96% with the output mode control (Pin 13) grounded, and 48% with it connected to the reference line. Additional dead time may be imposed on the output by setting the dead time-control input to a fixed voltage, ranging between 0 to 3.3 V.

The pulse width modulator comparator provides a means for the error amplifiers to adjust the output pulse width from the maximum percent on-time, established by the dead time control input, down to zero, as the voltage at the feedback pin varies from 0.5 to 3.5 V. Both error amplifiers have a common-mode input range from -0.3 V to (V_{CC} -2.0 V), and may be used to sense power-supply output voltage and current. The error-amplifier outputs are active high and are ORed together at the non-inverting input of the pulse-width modulator comparator. With this configuration, the amplifier that demands minimum output on time, dominates control of the loop.

When capacitor C_T is discharged, a positive pulse is generated on the output of the dead-time comparator, which clocks the pulse-steering flip-flop and inhibits the output transistors, Q1 and Q2. With the output-mode control connected to the reference line, the pulse-steering flip-flop directs the modulated pulses to each of the two output transistors alternately for push-pull operation. The output frequency is equal to half that of the oscillator. Output drive can also be taken from Q1 or Q2, when single-ended operation with a maximum on time of less than 50% is required. This is desirable when the output transformer has a ringback winding with a catch diode used for snubbing. When higher output drive currents are required for single-ended operation, Q1 and Q2 may be connected in parallel, and the output mode control pin must be tied to ground to disable the flip-flop. The output frequency will now be equal to that of the oscillator.

The TL494 has an internal 5.0 V reference capable of sourcing up to 10 mA of load currents for external bias circuits. The reference has an accuracy of ±5% over an operating temperature range of 0 to 70°C.

Application of The TL494 in a 400 W and 1000 Watt Off-Line Power Supply

A 5 V, 80 A line operated 25 kHz switching power supply, designed around the TL494, is shown in Figure 13-3, and the performance data is shown in Table 13-1. The explanation of each section of the power supply, which follows, applies not only to this model but to the higher power (12 V, 84 A) model shown in Figure 13-4, as well. In comparing the two, note that the 400-watt design is a half-bridge, while the 1,000 watt is a full bridge. The 1,000 watt power supply components switching transistors, transformers, and output rectifiers have been beefed up.

1. AC Input Section

The operating ac line voltage is selectable for a nominal of 115 or 230 volts by moving the jumper links to their appropriate positions. The input circuit is a full wave voltage doubler when connected for 115 Vac operation with both halves of the bridge connected in parallel for added line surge capability. When connected for 230 Vac operation, the input circuit forms a standard full wave bridge.

The line voltage tolerance for proper operation is −10, +20% of nominal. The ac line inrush current, during power-up, is limited by resistor R1. It is shorted out of the circuit by triac Q1, only after capacitors C1 and C2 are fully charged, and the high frequency output transformer T1, commences operation.

2. Power Section

The high frequency output transformer is driven in a half-bridge configuration by transistors Q3 and Q5. Each transistor is protected from inductive turn-off voltage transients by an R-C snubber and a fast recovery clamp rectifier. Transistors Q2 and Q4 provide turn-off drive to Q3 and Q5, respectively. In order to describe the operation of Q2, consider that Q6 and Q3 are turned on. Energy is coupled from the primary to the secondary of T3, forward biasing the base-emitter of Q3, and charging C3 through CR1. Resistor R3 provides a dc path for the 'on' drive after C3 is fully charged. Note that the emitter-base of Q2 is reverse biased during this time. Turn-off drive to Q3 commences during the dead-time period, when both Q6 and Q7 are off. During this time, capacitor C3 will forward bias the base-emitter of Q2 through R3 and R2 causing it to turn-on. The base-emitter of Q3 will now be reverse biased by the charge stored in C3 coupled through the collector-emitter of Q2.

TABLE 13-1

400 Watt Switcher Performance Data

Test	Conditions		Results
	Input	Output	
Line Regulation	103.5 to 138 VAC	5 volts and 80 amps	8 mV 0.16%
Load Regulation	115 VAC	5 volts, 0 to 80 amps	20 mV 0.4%
Output Ripple	115 VAC	5 volts and 80 amps	P.A.R.D. 50 mV P-P
Efficiency	115 VAC	5 volts and 80 amps	73%
Line Inrush Current	115 VAC	5 volts and 80 amps	24 amps peak

3. Output Section

The ac voltage present at the secondaries of T1 is rectified by four MBR6035 Schottky devices connected in a full wave center tapped configuration. Each device is protected from excessive switching voltage spikes by an R-C snubber, and output current sharing is aided by having separate secondary windings. Output current limit protection is achieved by incorporating a current sense transformer T4. The out-of-phase secondary halves of T1 are cross connected through the core of T4, forming a 1-turn primary. The 50 kHz output is filtered by inductor L1, and capacitor C4. Resistor R4 is used to guarantee that the power supply will have a minimum output load current of 1.0 ampere. This prevents the output transistors Q3 and/or Q5 from cycle skipping, as the required on-time to maintain regulation into an open circuit load is less than that of the devices' storage time. Transformer T5 is used to reduce output switching spikes by providing common mode noise rejection, and its use is optional.

The MC3423, U1, is used to sense an overvoltage condition at the output, and will trigger the crowbar S.C.R., Q8. The trip voltage is centered at 6.4 V with a programmed delay of 40 μs. In the event that a fault condition has caused the crowbar to fire, a signal is sent to the control section via jumper 'A' or 'B.' This signal is needed to shut down the output, which will prevent the crowbar S.C.R. from destruction due to over dissipation. Automatic over voltage reset is achieved by connecting jumper 'A.' The control section will cycle the power supply output every 2 seconds until the fault has cleared. If jumper 'B' is connected, S.C.R. Q12 will inhibit the output until the ac line is disconnected.

4. Low Voltage Supply Section

A low current internal power supply is used to keep the control circuitry active and independent from external loading of the output section. Transformer T2, Q9 and CR2 form a simple 14.3 V series pass regulator.

5. Control Section

The TL494 provides the pulse-width modulation control for the power supply. The minimum output dead-time is set to approximately 4% by grounding Pin 4 through R5. The soft start is controlled by C5 and R5. Transistor Q11 is used to discharge C5 and to inhibit the operation of the power supply if a low ac line voltage condition is sensed indirectly by Q10, or the output inhibit line is grounded.

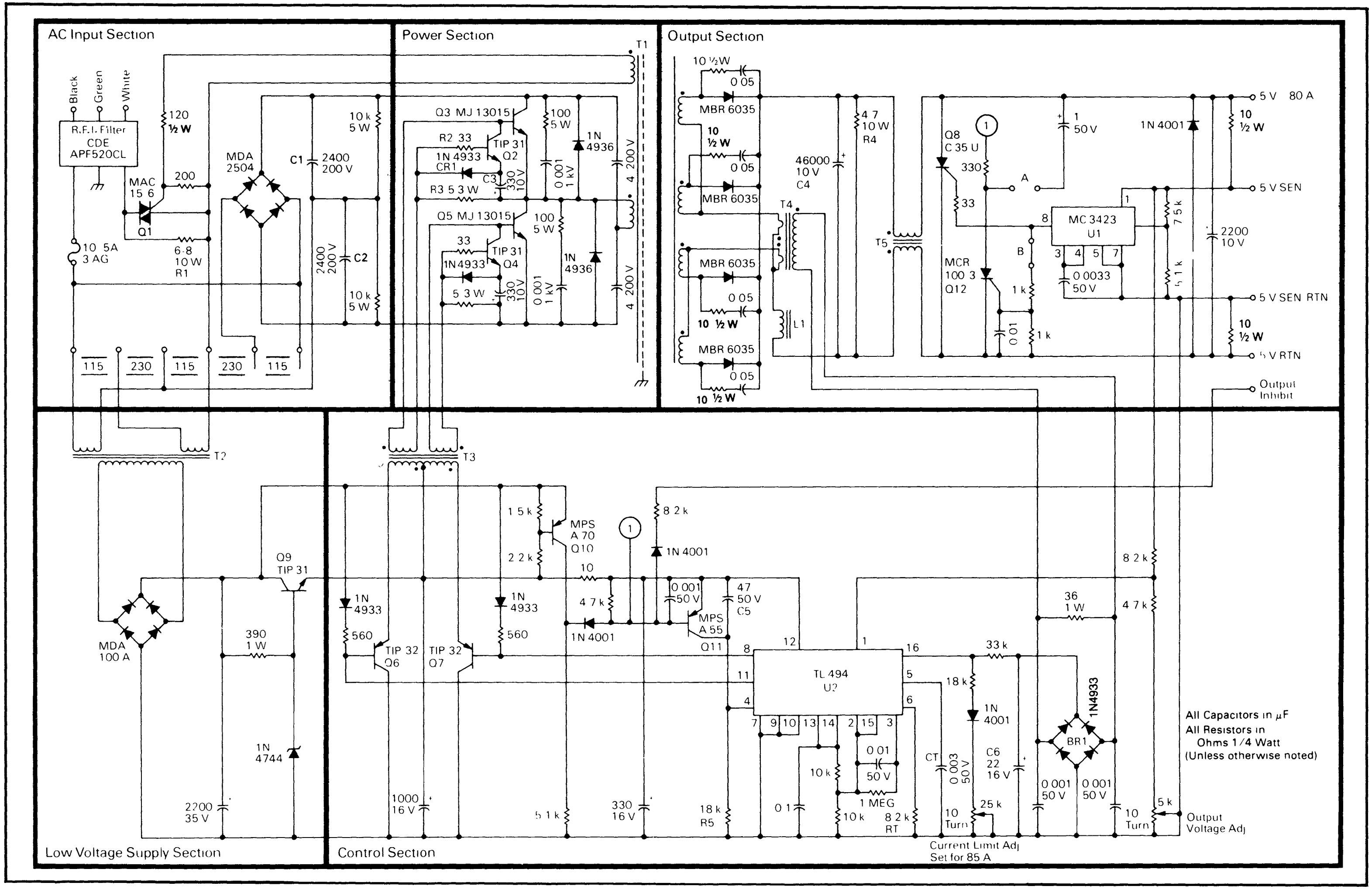


Figure 13-3. 400 Watt SWITCHMODE Power Supply

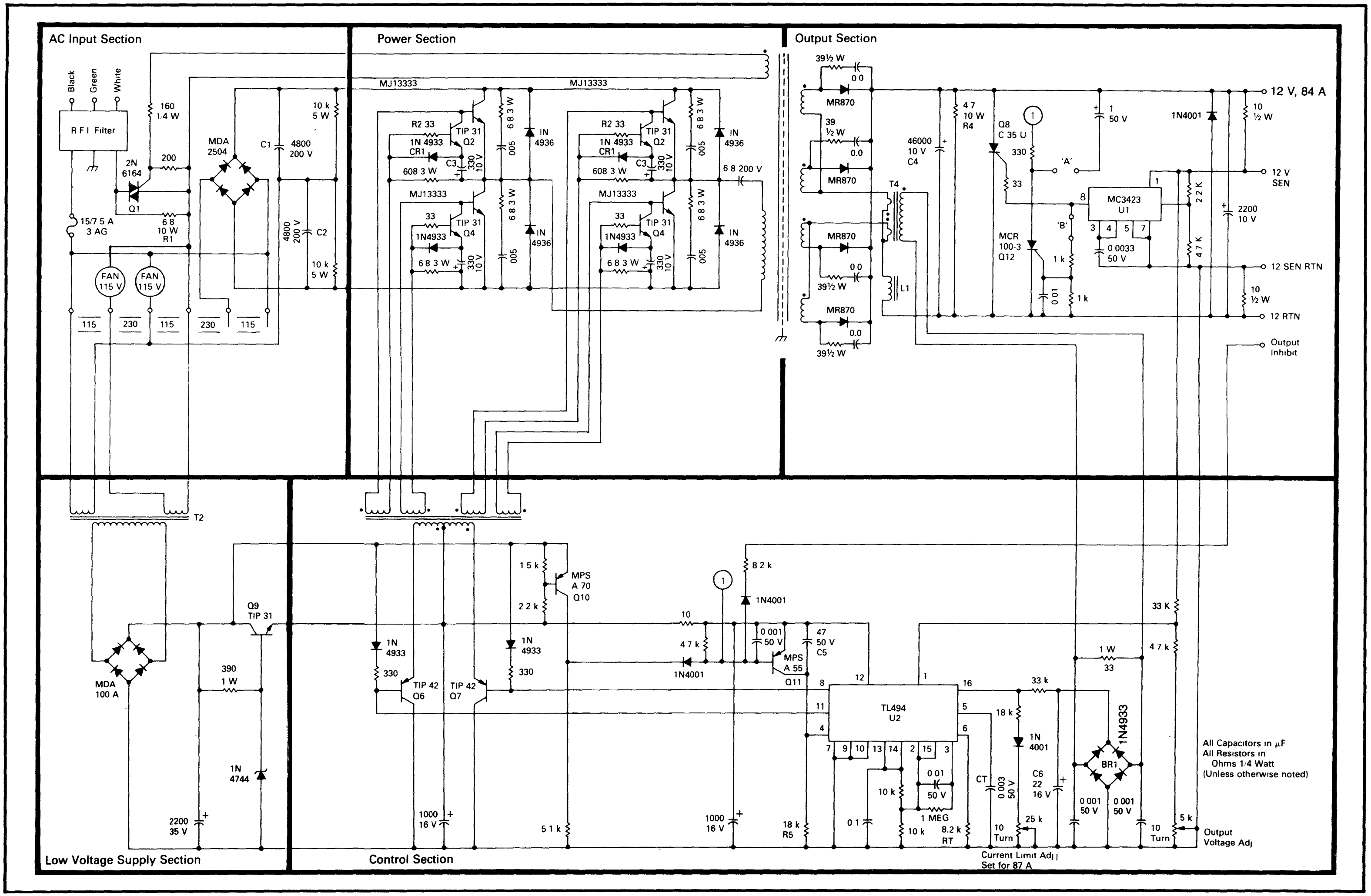


Figure 13-4. 1000 Watt SWITCHMODE Power Supply

Error amplifiers 1 and 2 are used for output voltage and current-level sensing, respectively. The inverting inputs of both amplifiers are connected together to a 2.5 V reference derived from Pin 14. By connecting the two inputs together, only one R-C feedback network is needed to set the voltage gain and roll-off characteristics for both amplifiers. Remote output voltage sensing capability is provided, and the supply will compensate for a combined total of 0.5 V drop in the power busses to the load. The secondary of the output current sense transformer T4, is terminated into 36 Ω and peak detected by BR1 and C6. The current limit adjust is set for a maximum output current of 85 amperes.

The oscillator frequency is set to 50 kHz by the timing components R_T and C_T. This results in a 25 kHz two phase output drive signal, when the output mode (Pin 13) is connected to the reference output (Pin 14).

TABLE 13-2

Transformer Data for 400 Watt SWITCHMODE Power Supply

T1	Core:	Ferroxcube EC 70-3C8, 0.002" gap in each leg.	
	Bobbin:	Ferroxcube 70 PTB.	
	Windings:	Primary (Q3, Q5):	50 turns total, #17 AWG Split wound about secondary.
		Primary (Q1):	4 turns, #17 AWG.
		Secondary, 4 each:	3 turns, #14 AWG Quad Filar wound.
		Shield, 2 each:	Made from soft allow copper 0.002" thick.
T2	Core:	Allegheny Ludlum EI-75-M6, 29 gauge.	
	Bobbin:	Bobbin Cosmo EI75.	
	Windings:	Primary, 2 each:	1000 turns, #36 AWG.
		Secondary:	200 turns, #24 AWG.
T3	Core:	Ferroxcube 846T250-3C8.	
	Windings:	Primary, 2 each:	30 turns, #30 AWG Bifilar wound.
		Secondary, 4 each:	12 turns, #20 AWG Bifilar wound.
T4	Core:	Magnetics Inc. 55059-A2	
	Windings:	Primary, 2 each:	1 turn, #14 AWG Quad Filar wound. Taken from secondary to T1.
		Secondary:	500 turns, #30 AWG.
T5	Core:	Magnetics Inc. 55071-A2	
	Windings:	Primary:	4 turns, #16 AWG Hex Filar wound.
		Secondary:	4 turns, #16 AWG Hex Filar wound.
L1	Core:	TDK H7C2DR56 x 35	
	Winding:	5 turns, soft alloy copper strap, 0.9" wide x 0.020" thick, 6.0 μH.	

TABLE 13-3

Transformer Data for 1,000 Watt Switching Power Supply

T1	Core:	Ferroxcube EC70-3C8, 0.002″ gap in each leg.	
	Bobbin:	Ferroxcube 70 PTB.	
	Windings:	Primary (Q3, Q5):	44 turns total, #18 AWG Bifilar Split wound about secondary.
		Primary (Q1):	3 turns, #18 AWG.
		Secondary, 4 each:	4 turns, #16 AWG Septe Filar wound.
		Shield, 2 each:	Made from soft alloy copper 0.002″ thick.
T2	Core:	Allegheny Ludlum EI-75-M6, 29 gauge.	
	Bobbin:	Bobbin Cosmo EI75.	
	Windings:	Primary, 2 each:	1000 turns, #36 AWG.
		Secondary:	200 turns, #24 AWG.
T3	Core:	Ferroxcube 846 T250-3C8.	
	Windings:	Primary, 2 each:	30 turns, #30 AWG Bifilar wound.
		Secondary, 4 each:	12 turns, #20 AWG Bifilar wound.
T4	Core:	Magnetics Inc. 55071-A2	
	Windings:	Primary, 2 each:	1 turn, #14 AWG Quad Filar wound. Taken from secondary to T1.
		Secondary:	500 turns, #30 AWG.
L1	Core:	TDK H7C2 DR 56 x 35	5 turns, soft alloy copper strap,
	Winding:		0.9″ wide x 0.020″ thick, 6.0 μH

B. 60-WATT FLYBACK SWITCHING POWER SUPPLY DESIGN

The flyback-regulator circuit (Figures 13-5 and 13-6) with a single drive transistor needs only a few main parts:

A unique flyback transformer

A single control IC (MC34060)

A fast-switching high-voltage transistor

Single output filters in each of the four outputs

The flyback base-drive circuit

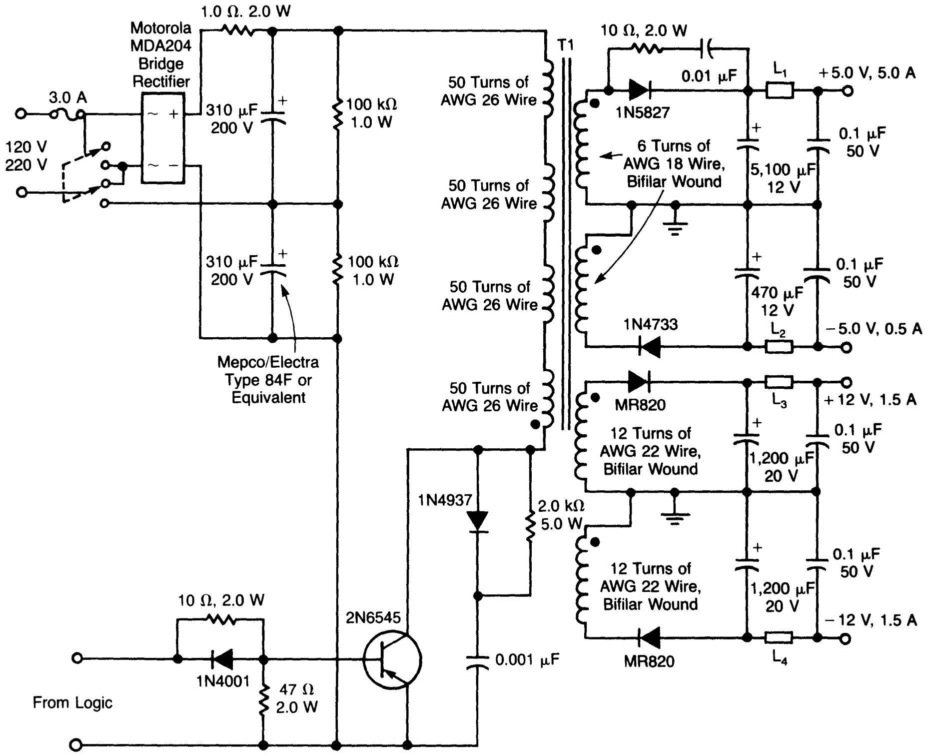
AC-line input voltage doublers.

In the power stage of Figure 13-5, a single 2N6545 transistor blocks 800 V and switches 1.0 A in 40 ns. The control section utilizes a low cost MC34060 Pulse Width Modulator control IC to minimize parts count.

The following paragraphs provide useful information and performance results regarding this Flyback design.

1. Sandwiching The Windings

The flyback transformer uses an EC-41 ferrite core made by the Ferroxcube Corp. It has a 40 : 1 turns ratio and is wound by a sandwich technique that improves the coupling between its primary and secondary windings.



T1 Specifications
Ferroxcube E Core EC41-3C8. Gap Poles of 50 mils (100 mils Total). Primary Inductance Is 4.0 mH at 3.0 A. Transformer Also Available from Pulse Engineering. #PE62792.

Winding Locations
- ▭ Primary
- ▭ +12 V
- ▭ Primary
- ▭ ±5 V
- ▭ Primary
- ▭ −12 V
- ▭ Primary

Seven Layers

L_1 through L_4
Ferrite Beads
0.3 in. Outside Diameter
0.09 in. Inside Diameter
0.3 in. Length
μ of 3,000

Output Capacitors
Sangamo Type 301, Mepco/Electra Type 3428, or Equivalent.

Figure 13-5. Flyback Power Stage Provides Output Voltages of +5, −5, +12 and −12 V

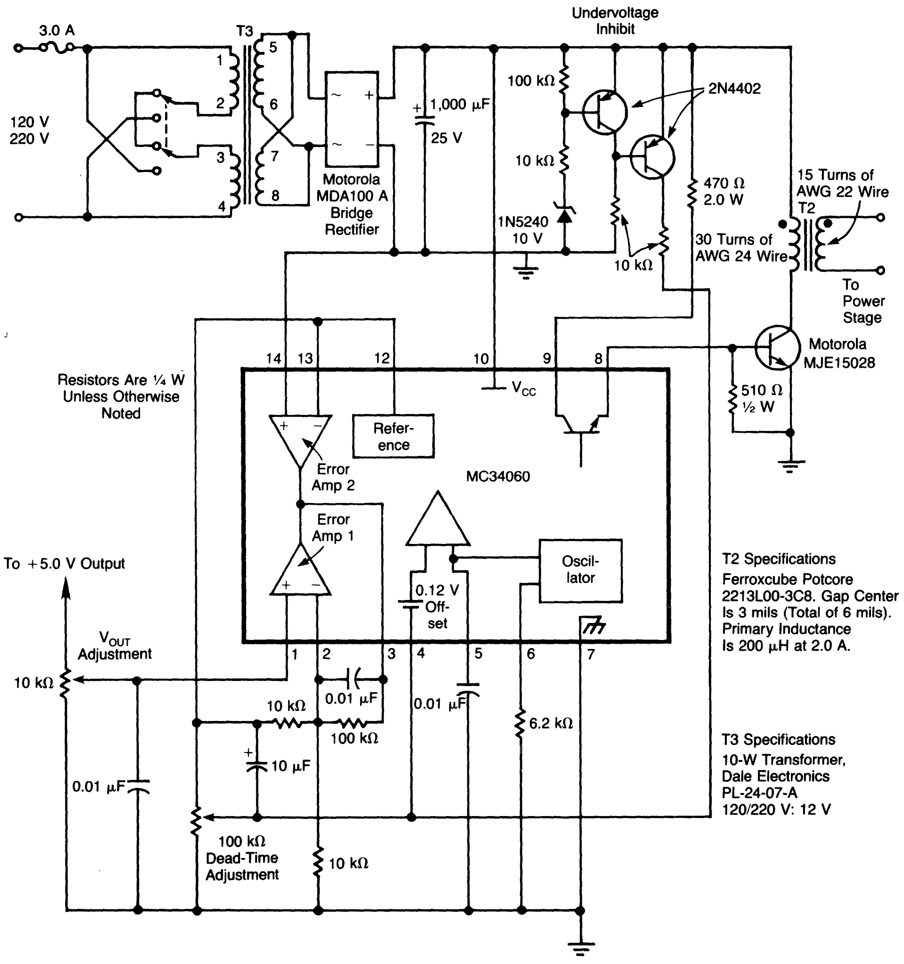


Figure 13-6. The Power Supply's Control Functions Are Obtained from The MC34060

The primary winding consists of four split windings in series with each other. The four windings of the secondary alternate in a sandwich construction with the four primary windings. Total core gap is 100 mils, and primary-winding inductance is 4.5 millinenries at 2.5 amperes. Transformer performance can be gauged from the fact that although the output current ratings for the secondary transformer windings are specified as 5.0, 1.5, and 0.5 A for 5.0 ± 12, and −5.0 V, respectively, actual respective current values are 8, 3, and 4 A (Figure 13-7). The flyback transformer can be hand-wound over an EC-41 ferrite core obtainable from Ferroxcube Corp. The four secondary windings alternate in a sandwich construction with four split primary windings that are connected in series with each other. All of the power-supply control functions reside in the MC34060 pulse width modulation control I.C. It includes a 20-kilohertz oscillator, a dead-time adjustment (50% maximum) for preventing transformer saturation, two error amplifiers to process both current and voltage feedback signals, and an output stage that produces 200 milliampere pulses to drive the power transistor. An undervoltage-inhibiting circuit is added externally to the control IC. Consisting of two transistors and a zener diode, it inhibits output pulses when the drive voltage is less than 10 V.

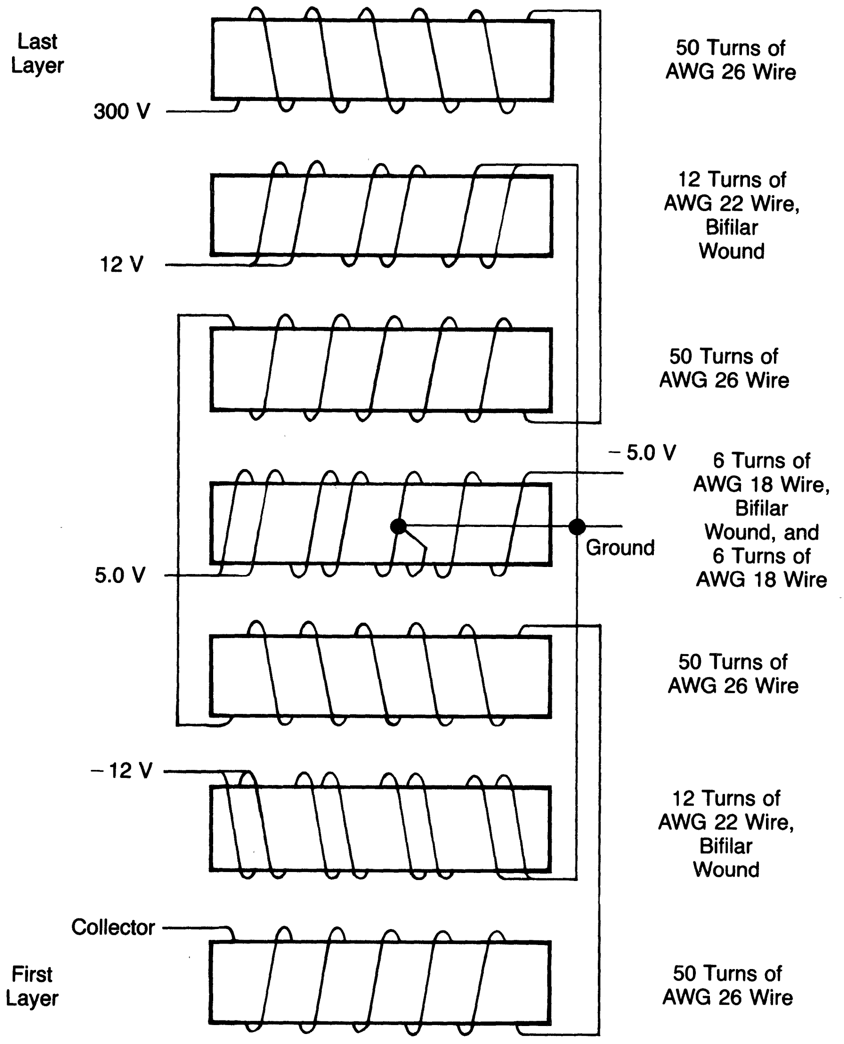
For fast switching, a Motorola type 2N6545 transistor is used. It is capable of switching 2.0 A in just 40 nanoseconds and can block up to 800 V under worst-case conditions. Because of the transistor's high speed, losses due to the snubber (the RC network in the collector circuit) are low — typically 2.0 W, or less than 2% of the total delivered power. Output Transistor current and voltage waveforms, along with load lines, are shown in Figures 13-8 and 13-9.

Each of the four output stages employs one filter capacitor and one diode. The capacitors (series 301 from Sangamo, 3428 from Mepco/Electra, or UPT from Cornell-Dubilier), exhibit low equivalent series resistance, typically 10 to 100 milliohms. Noise spikes are reduced dramatically (by as much as a factor of four) by the addition of a ferrite bead and ceramic capacitor across each of the output filter capacitors. Ripple test data for various types of capacitors is shown in Table 13-4.

TABLE 13-4. Ripple Test Data for Various Capacitors

Output	Test	Sangamo 301	Mepco/Electra 3428	CDE UPT	Mallory VPR	Sprague 432D
+5.0 V	Capacitance/volts	5,100 µF, 12 V	800 µF, 7.5 V	5,000 µF, 12 V	5,300 µF, 20 V	5,600 µF, 10 V
	Ripple (P-P)	200 mV	360 mV	170 mV	250 mV	200 mV
	Spikes (P-P)	660 mV	640 mV	980 mV	880 mV	580 mV
+12 V	Capacitance/volts	1,200 µF, 20 V	1,400 µF, 20 V	1,000 µF, 20 V	1,200 µF, 12 V	1,200 µF, 20 V
	Ripple	210 mV	260 mV	200 mV	200 mV	n.a.
	Spikes	740 mV	1,100 mV	1,800 mV	1,440 mV	n.a.
−5.0 V	Capacitance/volts	470 µF, 12 V	2,100 µF, 10 V	680 µF, 12 V	1,200 µF, 12 V	560 µF, 40 V
	Ripple	160 mV	160 mV	180 mV	140 mV	180 mV
	Spikes	540 mV	1,300 mV	680 mV	360 mV	440 mV

Winding	+ 5.0 V	±12 V	−5.0 V	300 V
Specifications	5.0 A	1.5 A	0.5 A	0.3 A
Actual	8.0 A	3.0 A	4.0 A	0.6 A



Ferroxcube E Core Type EC-41
Total Gap = 100 mils
Primary Inductance = 4.5 mH at 2.5 A

Figure 13-7. Flyback Transformer

The use of a flyback transformer for base drive greatly simplifies the drive circuit. Besides the transformer, only three other components are employed: a drive transistor capable of handling 2.0 A, a resistor, and a diode. The flyback transformer turns on the transistor with a 5.0 V drive pulse while simultaneously storing the energy from the 2.0 A current drawn by the transistor. This stored energy becomes the reverse bias drive when the pulse from the transformer is terminated. The reverse bias drive removes stored charge quickly — within 2 μs — and then causes the transistor's base to avalanche for the short while it takes to reset the transformer. Typically, if the transistor is initially turned on for 20 μs with a 5.0 V pulse, a 10 μs 10 V pulse is needed to reset if after it has been turned off.

At the ac line input, two axial-lead 310 μF, 200 V capacitors (Mepco/Electra series 84F) are connected in series with each other across the bridge rectifier output, thus acting as a voltage doubler when operating from 120 Vac line. A nominal 320 V bus is thus provided across the transformer's primary winding, regardless of whether it operates from a 120 Vac or a 220 Vac line input.

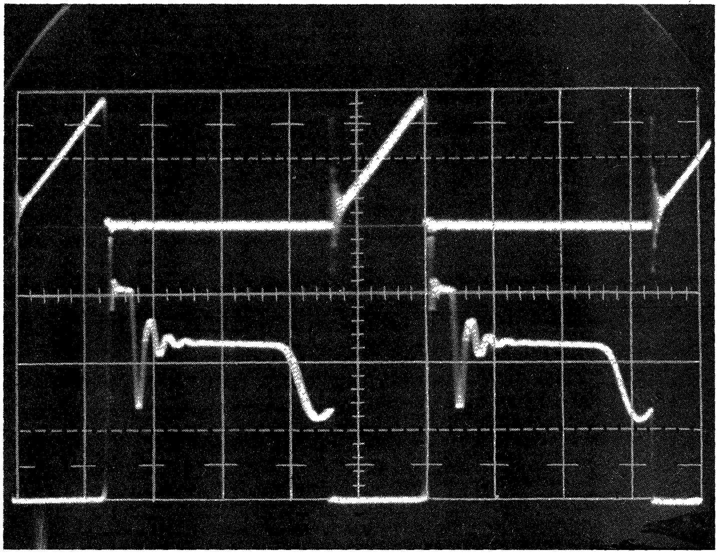
2. Advantages of Flyback — One of the most popular low wattage switching-regulator power supply circuits is the forward converter. The transformer, having only a 15 : 1 ratio of primary to secondary turns, is simpler than the flyback type approach, but requires four expensive filtering chokes. In addition, the secondary windings are unregulated, so output voltages vary with line and load variations more than they do in the case of a flyback transformer.

A flyback regulator with a control IC isolated from the primary side has a number of advantages. Feedback signals can be coupled directly to the transformer. Also, current-limiting protection on any or all of the output windings is simplified. Since the control IC has an extra amplifier, the addition of a sense resistor and simple divider network to the high-current 5.0 V output makes it easy to protect that output against short circuits (Figure 13-10). The addition of three more similar networks and a quad operational amplifier makes it a simple matter to protect all four outputs against short circuits.

This approach breaks with convention. Other switching-regulator schemes place the control IC at the primary side of the transformer, where the transistor emitter current is sensed for overcurrent protection. Optocouplers then have to be inserted in the feedback loop for proper isolation. Moreover, optocouplers drift over temperature.

3. Final Results — The output power stage can be checked out by using a pulse generator to energize the drive transistor and transformer; and subsequently, to calculate the snubber values. To improve coupling and reduce the 13 to 14 V nominal output to 12 V, the 5.0 V secondary winding can be increased from an initial five turns to six.

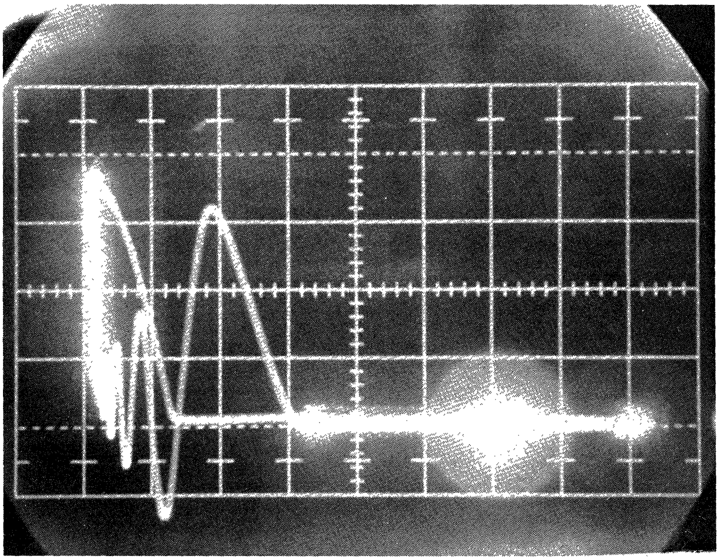
Adding control logic involves designing the base drive transformer and finding values for the feedback network that will provide optimum performance without creating instability. An operational amplifier gain of 20 with a rolloff at 160 Hz is sufficient. A dead-time limit of 50% keeps the drive transformer from saturation without interfering with low-line-voltage performance. An undervoltage-inhibiting circuit keeps the control circuit disabled at voltages under 10 V to prevent output pulses from occurring before sufficient drive is available to the output stage.



Vertical Scale:
I_C = 0.5 A/cm

Horizontal Scale:
t = 10 μs/cm

Figure 13-8. 2N6545 Current and Voltage Wave Forms



Vertical Scale:
I = 0.2 A/cm

Horizontal Scale:
V = 100 V/cm

Figure 13-9. 2N6545 Load Line

60-W FLYBACK REGULATOR WAVEFORMS

Despite the power supply's low parts count and simplicity of design, it has an impressive level of performance. For a nominal input of 120 Vac, it maintains regulation over an input range of 90 to 140 Vac and load range of 2 : 1 (half load to full load). For example, line and load regulation for the 5.0 V output are 2.5% and 1%, respectively. At an input of 90 Vac, full-load output voltages are 4.848, −4.930, −12.78 and 12.68 V, respectively, for the 5.0, −5.0, −12 and 12 V outputs. At 120 Vac, full-load output voltages are 5.001, −4.977, −12.98 and 12.94 V. At 140 Vac, full-load voltages are 5.983, −5.061, −13.16 and 13.10 V.

Half-load regulation is equally impressive. At a 90 Vac input, output voltages are 5.040, −5.075, −13.13 and 13.07 V. At a 120-V input, they are 5.098, −5.162, −13.30 and 13.20 V. At a 140-V input, they are 5.114, −5.191, −13.35, and 13.28 V.

Should it become necessary to work over a wider load range, such as from full to no load, the power transformer would have to be redesigned to protect the drive transistor from load dump conditions. This can be done by increasing the transformer's core size from the present EC-41 to EC-52 and by adding a primary bifilar winding coupled through a diode to the dc bus.

The power supply is also very efficient. At 120 Vac in and a full-load condition, its efficiency was an impressive 80%. The only noticeable heat rise is in the small components like the snubber resistor and Schottky diode. All other components remain cool to the touch.

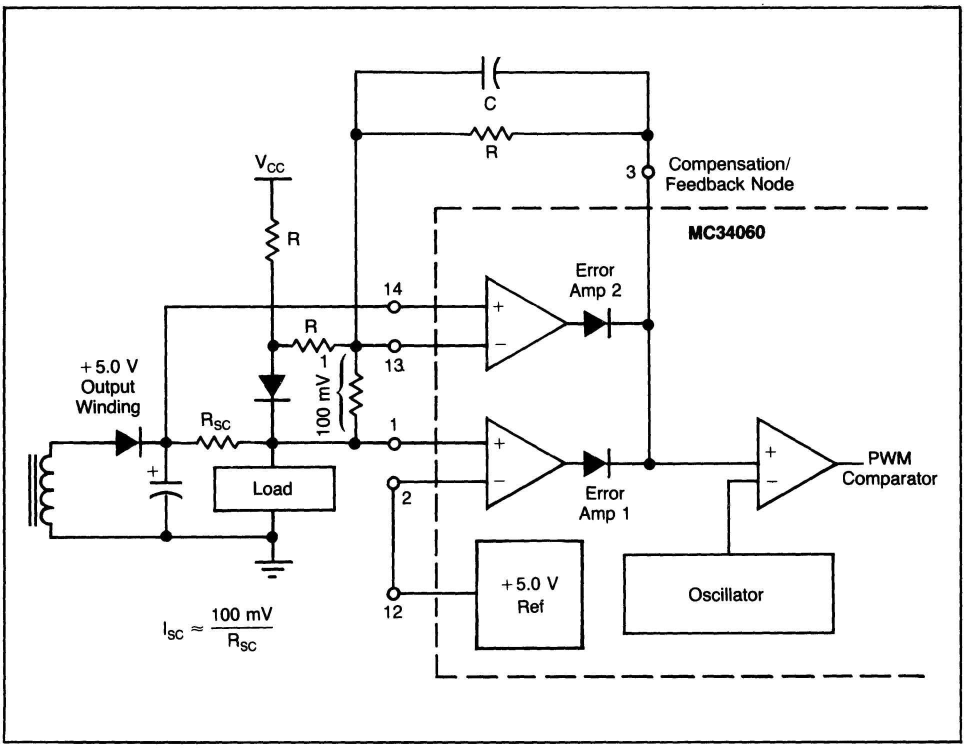


Figure 13-10. Current Limiting with the MC34060

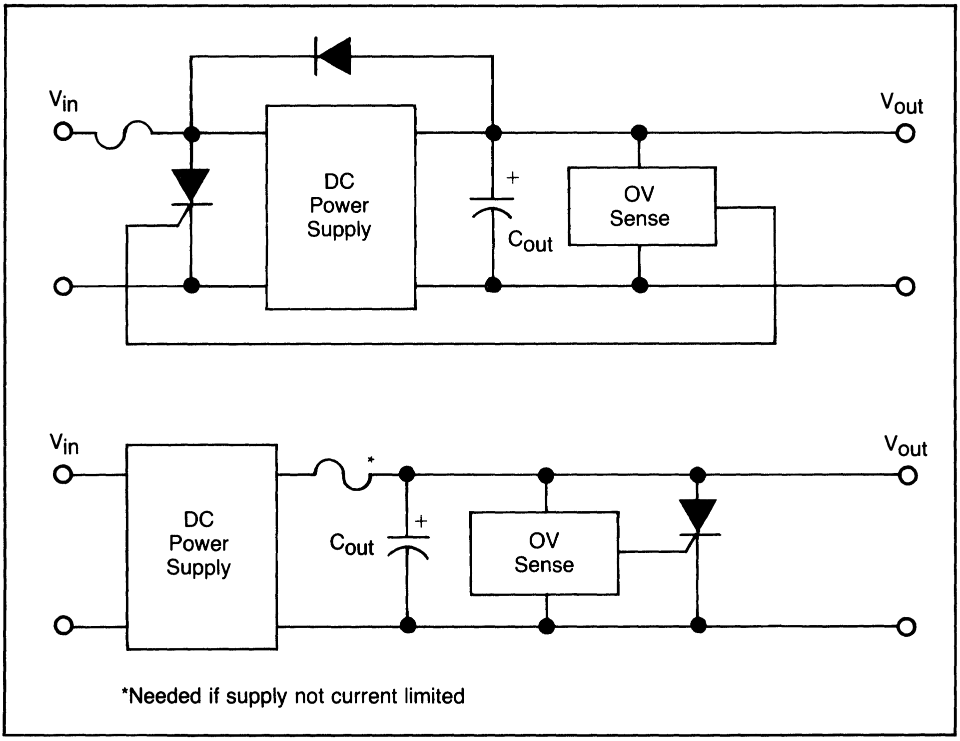
SECTION 14

POWER SUPPLY SUPERVISORY AND PROTECTION CONSIDERATIONS

The use of SCR crowbar overvoltage protection (OVP) circuits has been, for many years, a popular method of providing protection from accidental overvoltage stress for the load. In light of the recent advances in LSI circuitry, this technique has taken on added importance. It is not uncommon to have several hundred dollars worth of electronics supplied from a single low voltage supply. If this supply were to fail due to component failure or other accidental shorting of higher voltage supply busses to the low voltage bus, several hundred dollars worth of circuitry could literally go up in smoke. The small additional investment in protection circuitry can easily be justified in such applications.

A. THE CROWBAR TECHNIQUE

One of the simplest and most effective methods of obtaining overvoltage protection is to use a ''crowbar'' SCR placed across the equipment's dc power supply bus. As the name implies, the SCR is used much like a crowbar would be, to short the dc supply when an overvoltage condition is detected. Typical circuit configurations for this circuit are shown on Figure 14-1. This method is



*Needed if supply not current limited

Figure 14-1. Typical Crowbar OVP Circuit Configurations

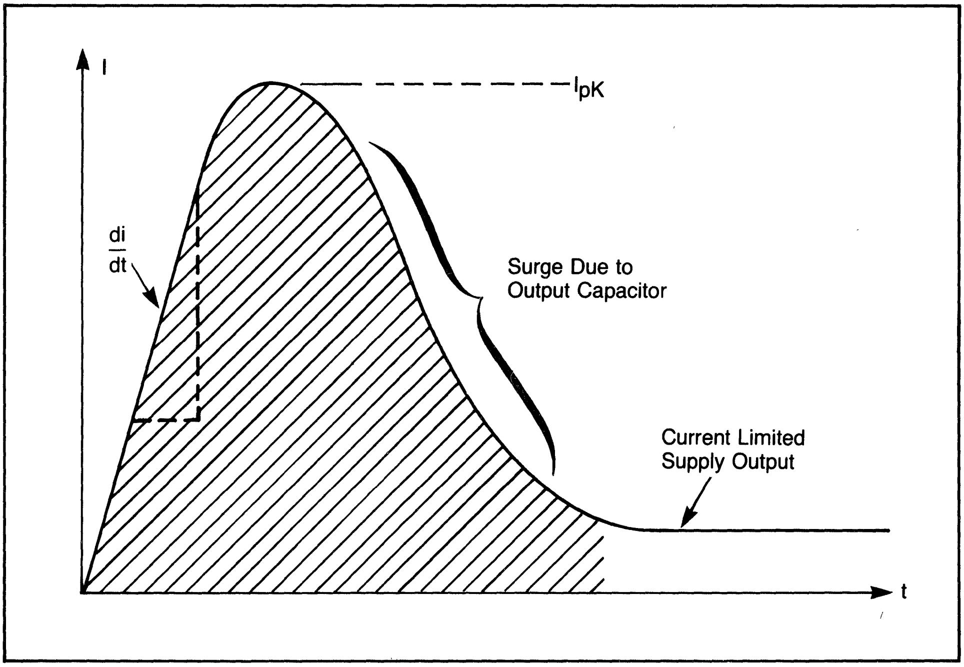


Figure 14-2. Crowbar SCR Surge Current Waveform

very effective in eliminating the destructive overvoltage condition. However, the effectiveness is lost if the OVP circuitry is not reliable.

B. SCR CONSIDERATIONS

Referring to Figure 14-1, it can easily been seen that, when activated, the crowbar SCR is subjected to a large current surge from the filter and output capacitors. This large current surge, illustrated in Figure 14-2, can cause SCR failure or degradation by any one of three mechanisms: di/dt, peak surge current, or I^2t. In many instances the designer must empirically determine the SCR and circuit elements which will result in reliable and effective OVP operation. To aid in the selection of devices for this application, Motorola has characterized several devices specifically for crowbar applications. A summary of these specifications and a selection guide for this application is shown in Table 14-1. This significantly reduces the amount of empirical testing that must be done by the designer. A good understanding of the factors that influence the SCR's di/dt and surge current capability will greatly simplify the total circuit design task.

TABLE 14-1

Crowbar SCRs

Device Type**	Peak Discharge Current*	di/dt*
MCR67	300 A	75 A/μs
MCR68	300 A	75 A/μs
MCR69	750 A	100 A/μs
MCR70	850 A	100 A/μs
MCR71	1700 A	200 A/μs

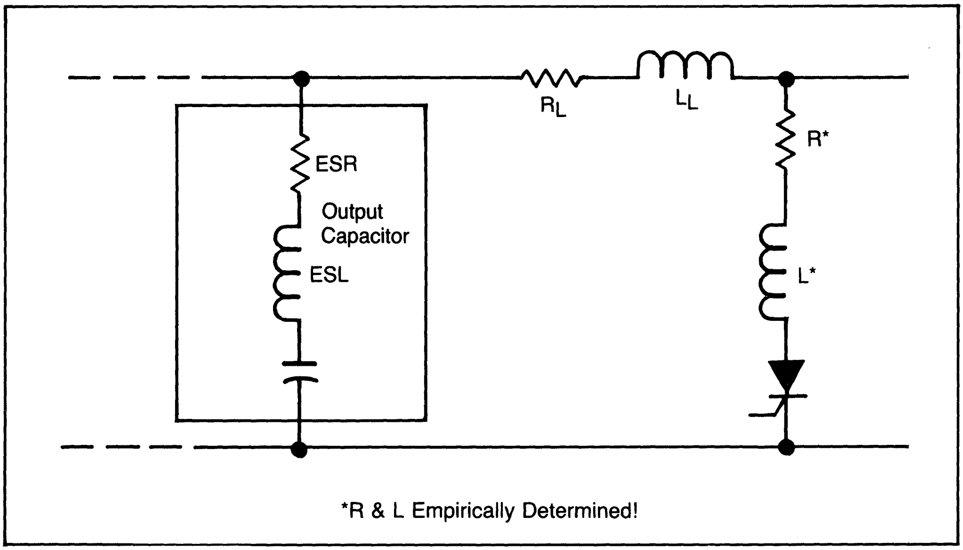
* tw = 1 μs, exponentially decaying
** All devices available with 25, 50, and 100 V ratings

1. di/dt — As the gate region of the SCR is driven on, its area of conduction takes a finite amount of time to grow, starting as a very small region and gradually spreading. Since the anode current flows through this turned-on gate region, very high current densities can occur in the gate region if high anode currents appear quickly (di/dt). This can result in immediate destruction of the SCR or gradual degradation of its forward blocking voltage capabilities, depending upon the severity of the occasion.

The value of di/dt that an SCR can safely handle is influenced by its construction and the characteristics of the gate drive signal. A center-gate-fire SCR has more di/dt capability than a corner-gate-fire type, and heavily overdriving (3 to 5 times I_{GT}) the SCR gate with a fast (< 1 μs) rise time signal will maximize its di/dt capability. A typical maximum di/dt in phase control SCRs of less than 50 A rms rating might be 200 A/μs, assuming a gate current of five times I_{GT} and < 1 μs rise time. If having done this, a di/dt problem still exists, the designer can also decrease the di/dt of the current waveform by adding inductance in series with the SCR, as shown in Figure 14-3. Of course, this reduces the circuit's ability to rapidly reduce the dc bus voltage, and a tradeoff must be made between speedy voltage reduction and di/dt.

2. Surge Current — If the peak current and/or the duration of the surge is excessive, immediate destruction due to device overheating will result. The surge capability of the SCR is directly proportional to its die area. If the surge current cannot be reduced (by adding series resistance — see Figure 14-3) to a safe level which is consistent with the system's requirements for speedy bus voltage reduction, the designer must use a higher current SCR. This may result in the average current capability of the SCR exceeding the steady state current requirements imposed by the dc power supply.

(For additional information on SCRs in crowbar applications refer to "Characterizing the SCR for Crowbar Applications," Al Pshaenich, Motorola AN-789).



*R & L Empirically Determined!

Figure 14-3. Circuit Elements Affecting SCR Surge & di/dt

C. THE SENSE AND DRIVE CIRCUIT

In order to maximize the crowbar SCR's di/dt capability, it should receive a fast rise time high-amplitude gate-drive signal. This must be one of the primary factors considered when selecting the sensing and drive circuitry. Also important is the sense circuitry's noise immunity.

Noise immunity can be a major factor in the selection of the sense circuitry employed. If the sensing circuit has low immunity and is operated in a noisy environment, nuisance tripping of the OVP circuit can occur on short localized noise spikes, which would not normally damage the load. This results in excessive system down time. There are several types of sense circuits presently being used in OVP applications. These can be classified into three types: zener, discrete, and "723."

1. The Zener Sense Circuit — Figure 14-4 shows the use of a zener to trigger the crowbar SCR. This method is NOT recommended since it provides very poor gate drive and greatly decreases the SCR's di/dt handling capability, especially since the SCR steals its own very necessary gate drive as it turns on. Additionally, this method does not allow the trip point to be adjusted except by zener replacement.

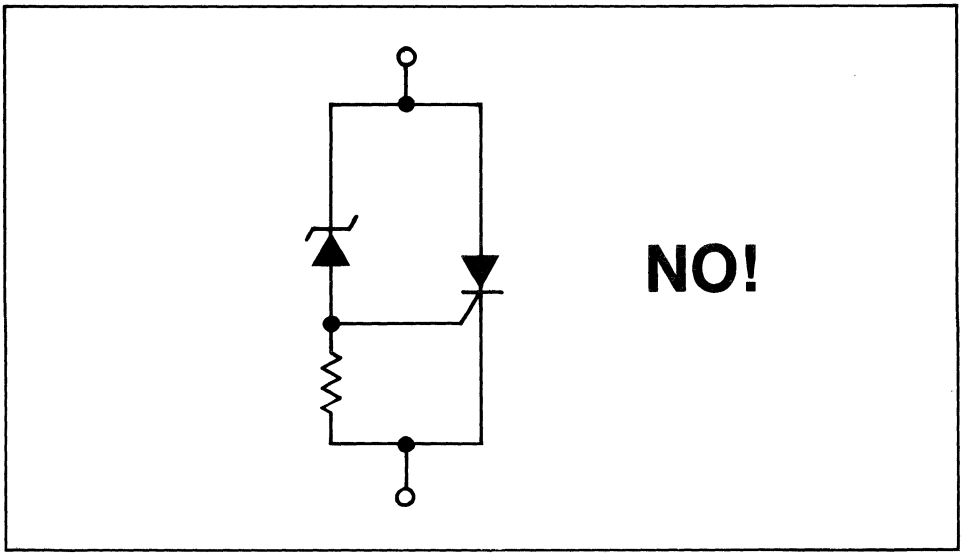


Figure 14-4. The Zener Sense Circuit

2. The Discrete Sense Circuit — A technique which can provide adequate gate drive and an adjustable, low temperature coefficient trip point is shown in Figure 14-5. While overcoming the disadvantages of the zener sense circuit, this technique requires many components and is more costly. In addition, this method is not particularly noise immune and often suffers from nuisance tripping.

3. The "723" Sense Circuit — By using an integrated circuit voltage regulator, such as the industry standard "723" type, a considerable reduction in component count can be achieved. This is illustrated in Figure 14-6. Unfortunately, this technique is not noise immune, and suffers an additional disadvantage in that it must be operated at voltages above 9.5 volts.

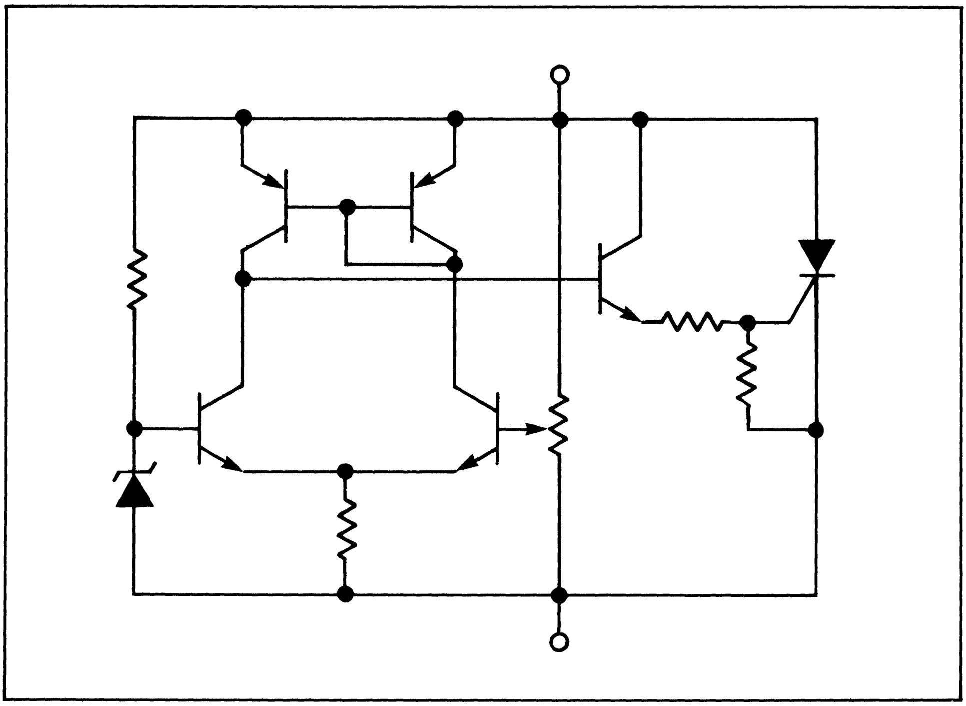


Figure 14-5. The Discrete Sense Circuit

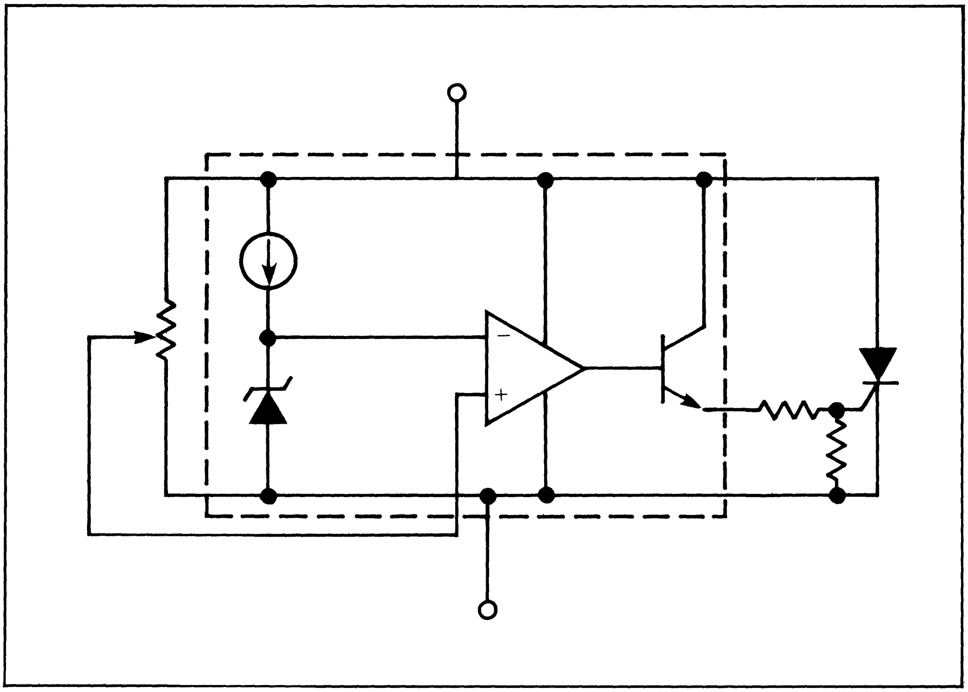


Figure 14-6. The "723" Sense Circuit

4. The MC3423 — To fill the need for a low cost, low complexity method of implementing crowbar overvoltage protection which does not suffer the disadvantages of previous techniques, an IC has been developed for use as an OVP sense and drive circuit, the MC3423.

The MC3423 was designed to provide output currents of up to 300 mA with a 400 mA/µs rise time in order to maximize the di/dt capabilities of the crowbar SCR. In addition, its features include:

1. Operation off 4.5 V to 40 V supply voltages.
2. Adustable, low temperature coefficient trip point.
3. Adjustable minimum overvoltage duration before actuation to reduce nuisance tripping in noisy environments.
4. Remote activation input.
5. Indication output.

5. Block Diagram — The block diagram of the MC3423 is shown in Figure 14-7. It consists of a stable 2.6 V reference, two comparators and a high current output. This output, together with the indication output transistor, is activated either by a voltage greater than 2.6 V on Pin 3 or by a TTL/5 volt CMOS high logic level on the remote activation input, Pin 5.

The circuit also has a comparator-controlled current source which can be used in conjunction with and external timing capacitor to set a minimum overvoltage duration (0.5 µs to 1.0 ms) before actuation occurs. This feature allows the OVP circuit to operate in noisy environments without nuisance tripping.

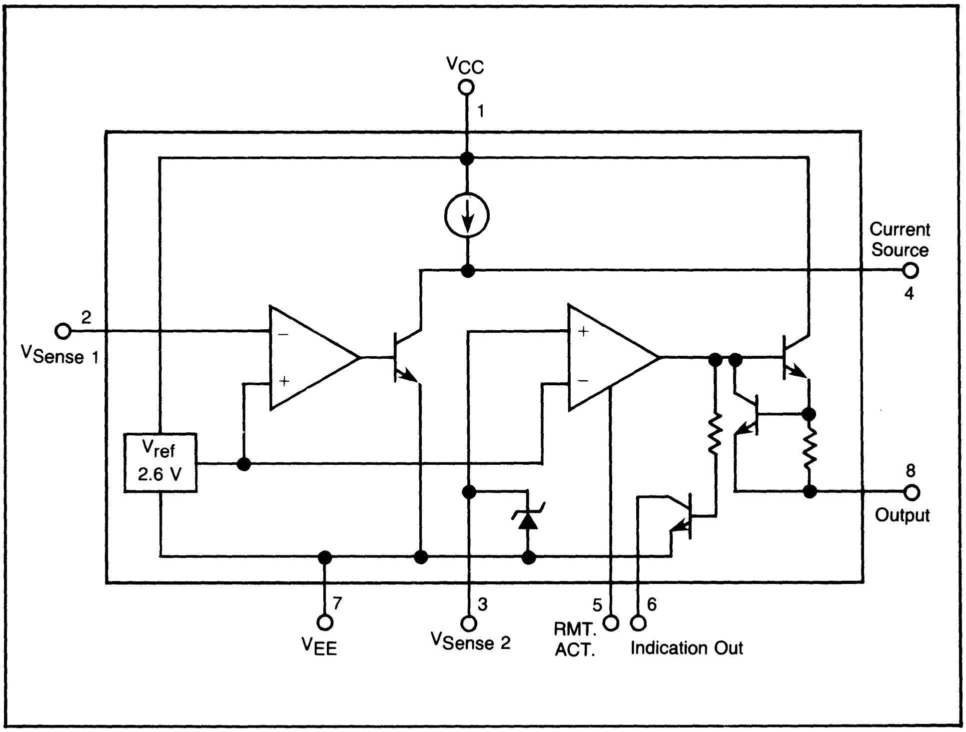
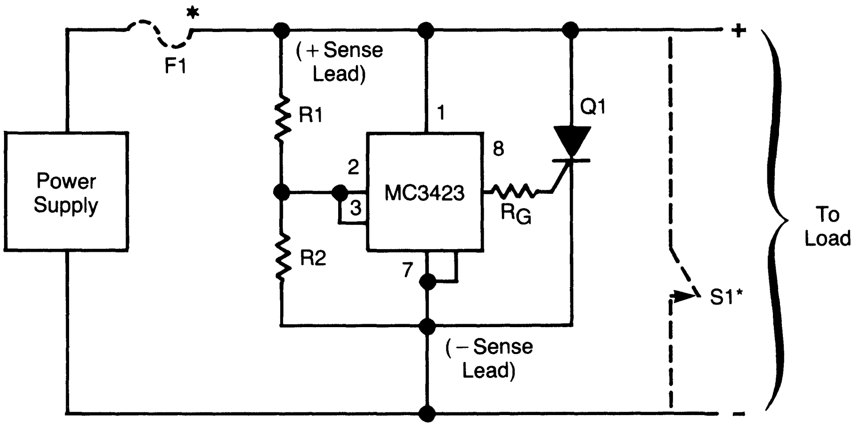


Figure 14-7. MC3423 Block Diagram

6. Basic Circuit Configuration — The basic circuit configuration of the MC3423 OVP is shown in Figure 14-8. In this circuit the voltage sensing inputs of both the internal amplifiers are tied together for sensing the overvoltage condition. The shortest possible propagation delay is thus obtained. The threshold or trip voltage at which the MC3423 will trigger and supply gate drive to the crowbar SCR, Q1, is determined by the selection of R1 and R2. Their values can be determined by the equations given in Figure 14-8 or by the graph shown in Figure 14-9. The switch, S1, shown in Figure 14-8 may be used to reset the SCR crowbar. Otherwise, the power supply, across which the SCR is connected, must be shut down to reset the crowbar. If a non current-limited supply is used a fuse or circuit breaker, F1, should be used to protect the SCR and/or the load.



$V_{TRIP} = V_{ref}\,(1+\frac{R1}{R2}) \approx 2.6\text{ V}\,(1+\frac{R1}{R2})$

R2 ≤ 10 kΩ for Minimum Drift

*Needed if supply is not current limited

Figure 14-8. MC3423 Basic Circuit Configuration

7. MC3423 Programmable Configuration — In many instances, MC3423 OVP will be used in a noisy environment. To prevent false tripping of the OVP circuit by noise which would not normally harm the load, MC3423 has a programmable delay feature. To implement this feature, the circuit configuration of Figure 14-10 is used.

Here a capacitor is connected from Pin 3 and Pin 4 to V_{EE}. The value of this capacitor determines the minimum duration of the overvoltage condition (t_D) which is necessary to trip the OVP. The value of C_D can be found from Figure 14-11. The circuit operates in the following manner: when V_{CC} rises above the trip point set by R1 and R2, the internal current source begins charging the capacitor, C_D, connected to pins 3 and 4. If the overvoltage condition remains present long enough for the capacitor voltage, V_{CD} to reach V_{ref}, the ouput is activated. If the overvoltage condition disappears before this occurs, the capacitor is discharged at a rate 10 times faster than the charging rate, resetting the timing feature until the next over-voltage condition occurs.

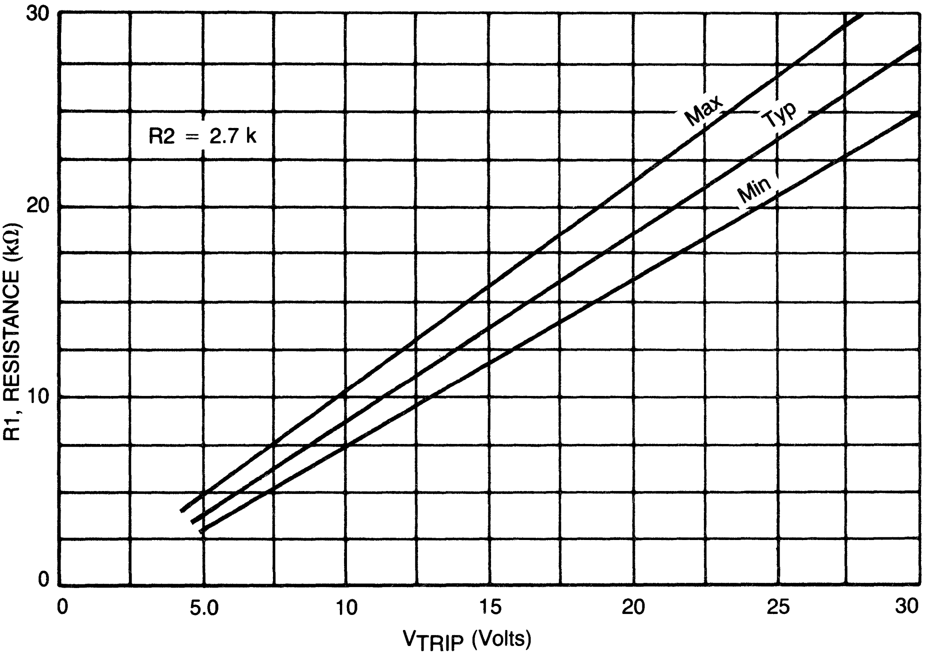
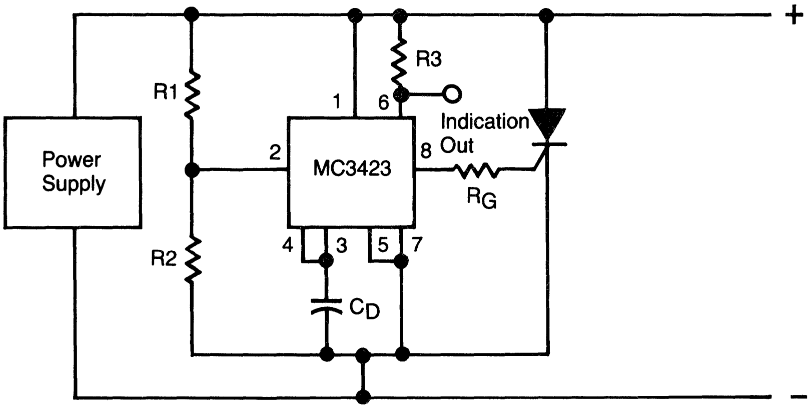


Figure 14-9. R1 versus Trip Voltage for The MC3423 OVP



$$R3 \geqslant \frac{V_{TRIP}}{10 \text{ mA}}$$

Figure 14-10. MC3423 Configuration for Programmable Minimum Duration of Overvoltage Condition before Tripping

8. Indication Output — An additional output for use as an indicator of OVP activation is provided by the MC3423. This output (Pin 6) is an open collector transistor which saturates when the MC3423 OVP is activated. It will remain in a saturated state until the SCR crowbar pulls the supply voltage, V_{CC}, below 4.5 V as in Figure 14-10. This output can be used to clock an edge triggered flop-flop whose output inhibits or shuts down the power supply when the OVP trips. This reduces or eliminates the heatsinking requirements for the crowbar SCR.

9. Remote Activation Input — Another feature of the MC3423 is its Remote Activation Input, Pin 5. If the voltage on this CMOS/TTL compatible input is held below 0.7 V, the MC3423 operates normally. However, if it is raised to a voltage above 2.0 V, the OVP output is activated independent of whether or not an overvoltage condition is present.

This feature can be used to accomplish an orderly and sequenced shutdown of system power supplies during a system fault condition. In addition, the Indication Output of one MC3423 can be used to activate another MC3423, if a single transistor inverter is used to interface the former's Indication Output to the latter's Remote Activation Input.

D. THE MC3424

In addition to the MC3423 a second IC, the MC3424, has been developed for overvoltage protection and power supply supervision. Similar in many respects to the MC3423, the MC3424 may also be programmed for under voltage detection

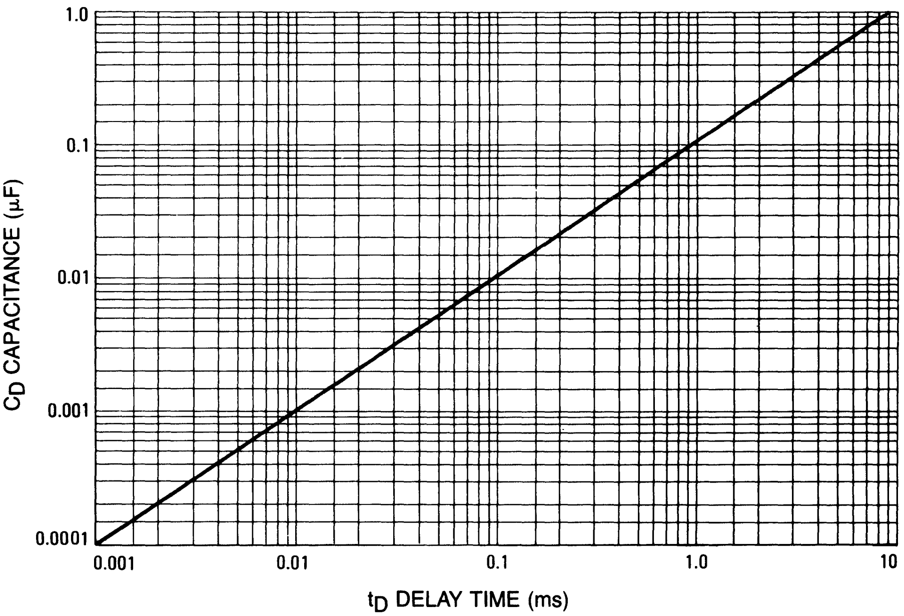


Figure 14-11. C_D versus Minimum Overvoltage Duration, t_D for The MC3423 OVP

or line loss monitoring. With a few passive components the MC3424 is able to perform all of the monitoring required for a power supply.

The block diagram of the MC3424 is shown in Figure 14-12. Notice that both inputs to the two sensing comparators (C1+, C1−, C2+, and C2−) are pinned out to provide additional flexibility. In addition the "−" inputs to the comparators are tied to controlled current sinks which may be used to provide hysteresis in the sensing function. The hysteresis voltage (V_H) at the comparator input can be calculated using the equation:

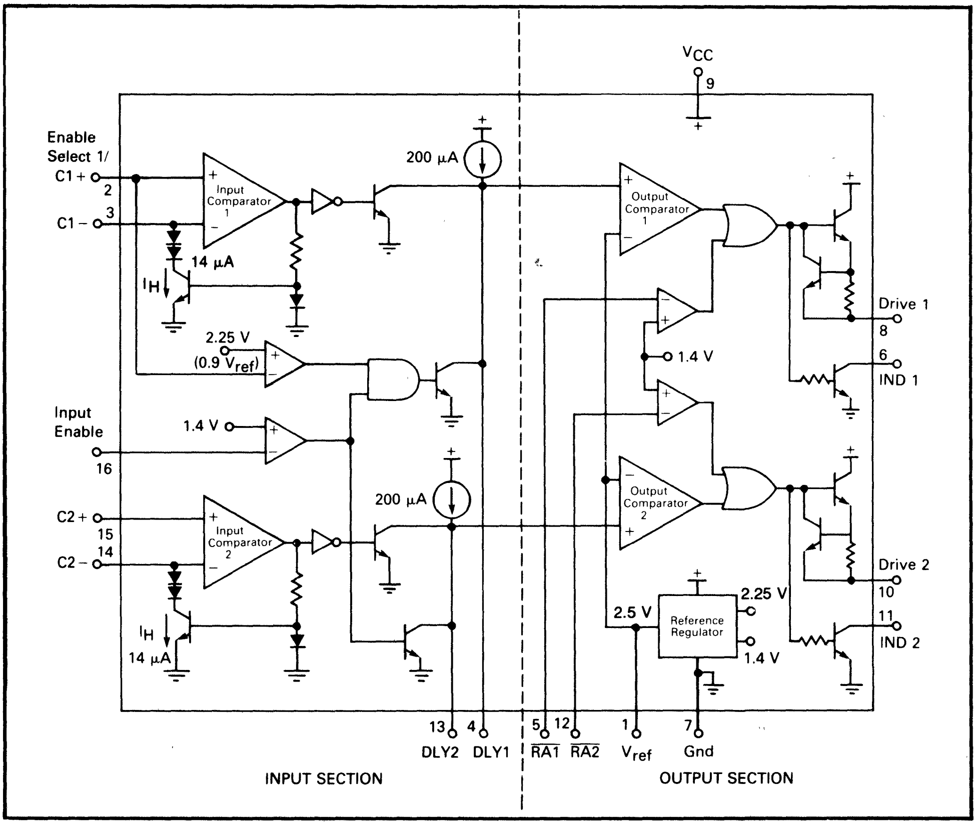
$$V_H = R_H I_H$$

Where R_H = equivalent resistance

I_H = comparator hysteresis current

If hysteresis is not required, it can be eliminated by making the equivalent resistance in series with the C− input (R_H) equal to zero or by configuring the device such that the quiescent operating point for the C− input is below 1.2 volts.

Both channels of the MC3424 may be operated independently, and both have high current drive outputs and open collector indicator outputs.



Note: All voltages and currents are nominal.

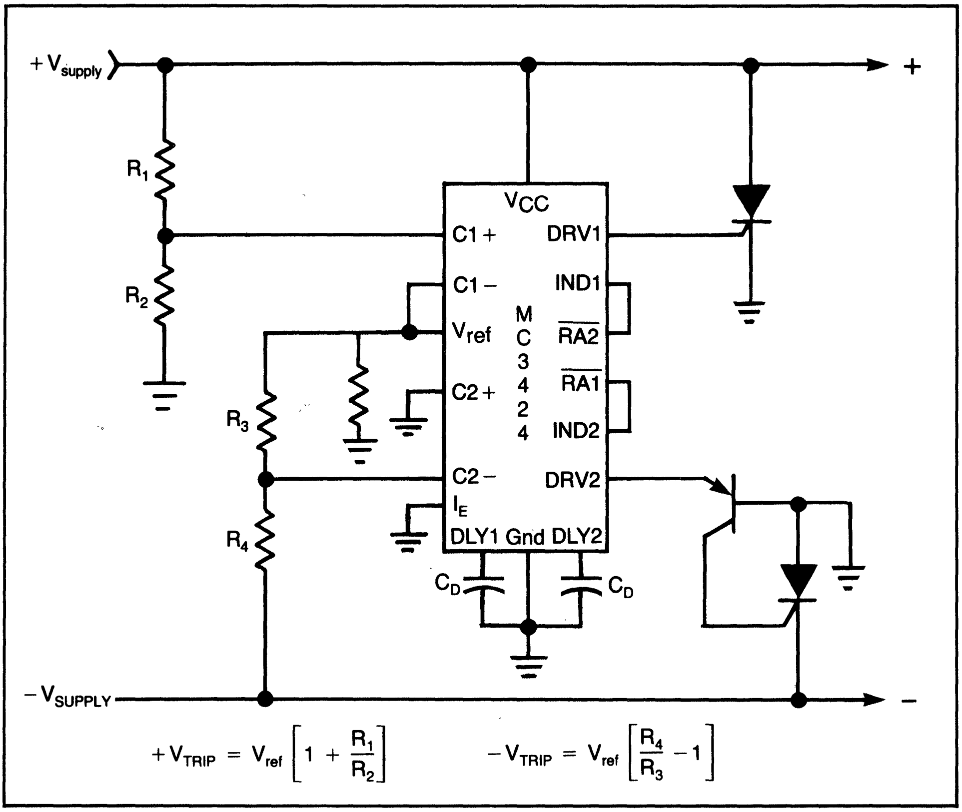
Figure 14-12. MC3424/MC3524 Block Diagram

1. Dual Overvoltage Protection — The circuit shown in Figure 14-13 uses the MC3424 to provide overvoltage sensing for a split supply. In this application the MC3424 is powered from the positive supply but senses both the positive and negative supplies, and will crowbar both supplies if a overvoltage condition is detected on either of the supplies.

To cause the MC3424 to crowbar both supplies, the indicator outputs from each half of the device are connected to the remote activation inputs of the other half of the device. With this arrangement, if either side of the device detects an overvoltage condition it will cause one of the SCRs to crowbar, and at the same time, activate the other half of the circuit, which will in turn cause the second SCR to crowbar.

If more than two supplies were to be protected, a similar arrangement could be used to cause all of the supplies to be crowbarred if any fault occurred. To accomplish this, simply connect all of the remote activation inputs and all of the indicator outputs together. Since the indicator outputs of the MC3424 are open collector devices, any one of the indicator outputs can activate all of the crowbars without any interference.

2. Line Loss Detection — In addition to providing overvoltage protection, the MC3424 can also be used to detect line loss or brownout conditions which will soon cause the power supply to fail. This is particularly important in many small



$$+V_{TRIP} = V_{ref}\left[1 + \frac{R_1}{R_2}\right] \qquad -V_{TRIP} = V_{ref}\left[\frac{R_4}{R_3} - 1\right]$$

Figure 14-13. OVP for Split Supply Operation

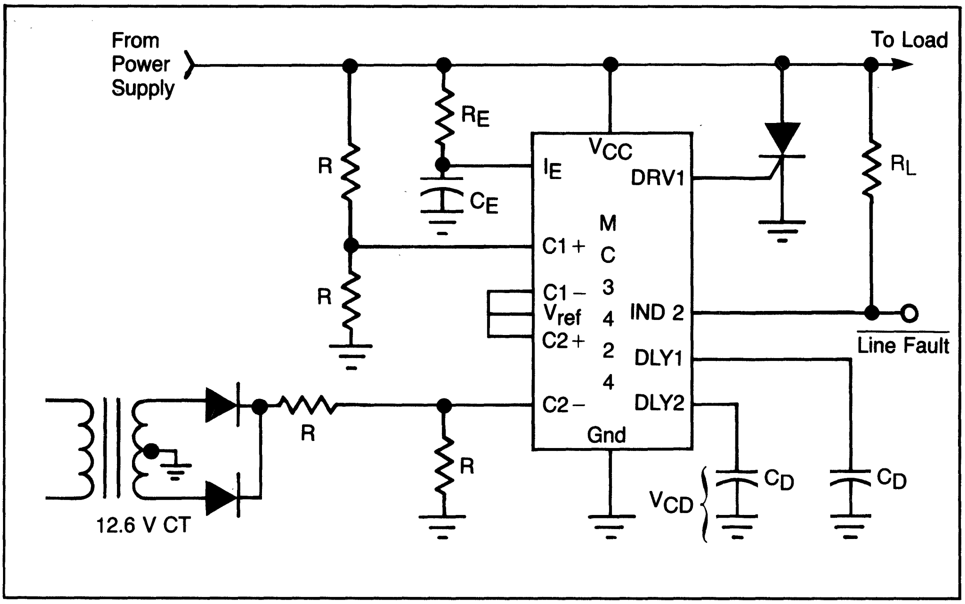


Figure 14-14. Sensing Line Fault and Over Voltage Conditions for Linear and Switching Power Supplies

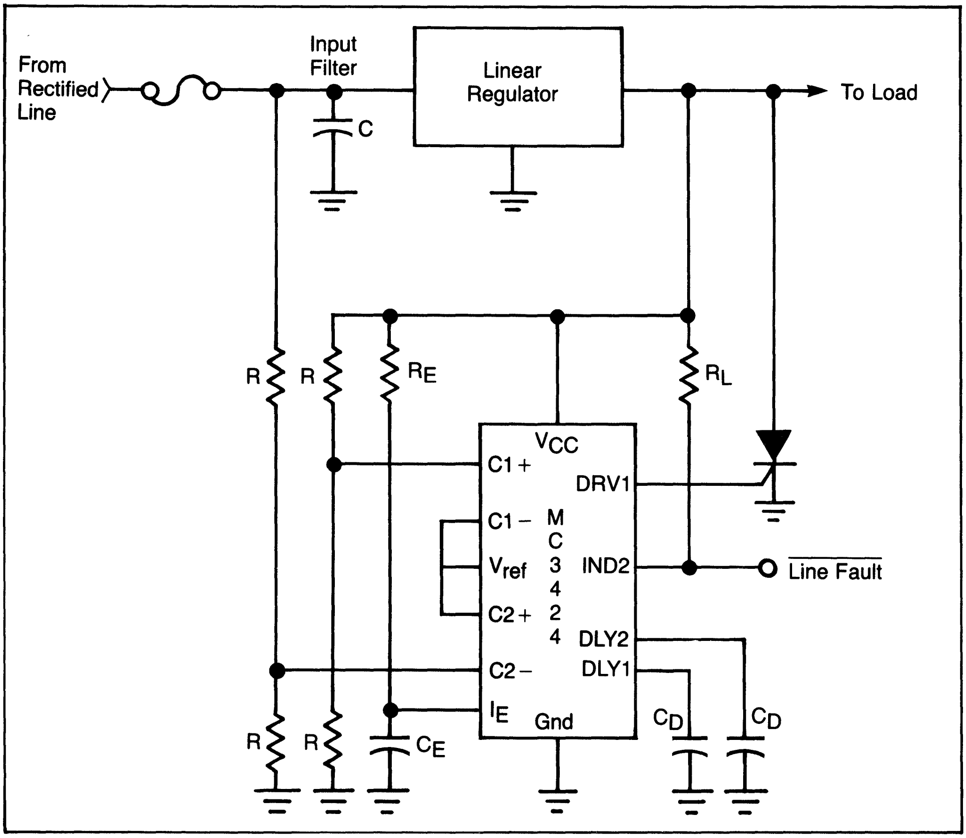


Figure 14-15. An Alternate Method of Sensing Line Fault and Overvoltage Conditions for Linear Power Supplies

and medium sized computer systems which must store part or all of the data currently being processed before the power failure. The use of circuits such as these will allow such systems to ''die with dignity.''

The circuits shown in Figures 14-14 and 14-15 both perform essentially the same function. The circuit shown in Figure 14-14 may be used with almost any type of regulator circuitry; however, the circuit shown in Figure 14-15 should only be used in linear type supplies where the filter capacitor is isolated from the line. Using the circuit in Figure 14-15 on switching supplies where the filter capacitors are not isolated from the line would defeat the isolation in the switching transformer.

The circuit shown in Figure 14-14 utilizes half of the MC3424 as an overvoltage protection circuit in a configuration like the programmable configuration discussed earlier for the MC3423. The remaining half of the device is configured for line loss and brownout detection. The C2+ and C2− inputs are connected as an undervoltage sensing circuit, and sense the center tap of a voltage divider driven with a full wave rectified signal proportional to the line voltage. At each peak of the line the output of the comparator discharges the delay capacitor (C_D). If a half cycle is missing from the line voltage, or if a brownout occurs reducing the peak line voltage, the delay capacitor will not be discharged and will continue to be charged as shown in Figure 14-16. If a sufficient number of half cycles are missing, or if the brownout continues for a sufficient time, the circuit will detect an ac line fault and output a line fault indication on the indicator output. The delay capacitor is used to provide some noise immunity and to prevent the loss of a single half cycle from triggering the line fault signal. The minimum time the fault condition must occur can be adjusted by changing the value of the delay capacitor.

The circuit shown in Figure 14-15 senses the voltage on the power supply filter capacitors to predict the imminent power supply failure. Since the voltage on the capacitor is proportional to the remaining charge, the remaining time the power supply will function can be calculated by the equation:

$$t = \frac{C\,(V_C - V_{min})}{I_{max}}$$

Where C = filter capacitance

t = time to power supply failure

I_{max} = maximum load current

V_C = filter capacitor voltage

V_{min} = minimum regulator input voltage

By setting t equal to the maximum time for the system to store all required data, and solving the equation for V_C, the minimum capacitor voltage can be calculated that will allow the supply to remain functional, while the system executes the power down sequence. The MC3424 is then configured as an undervoltage detector, as shown in Figure 14-15, and programmed to detect the minimum capacitor voltage V_C.

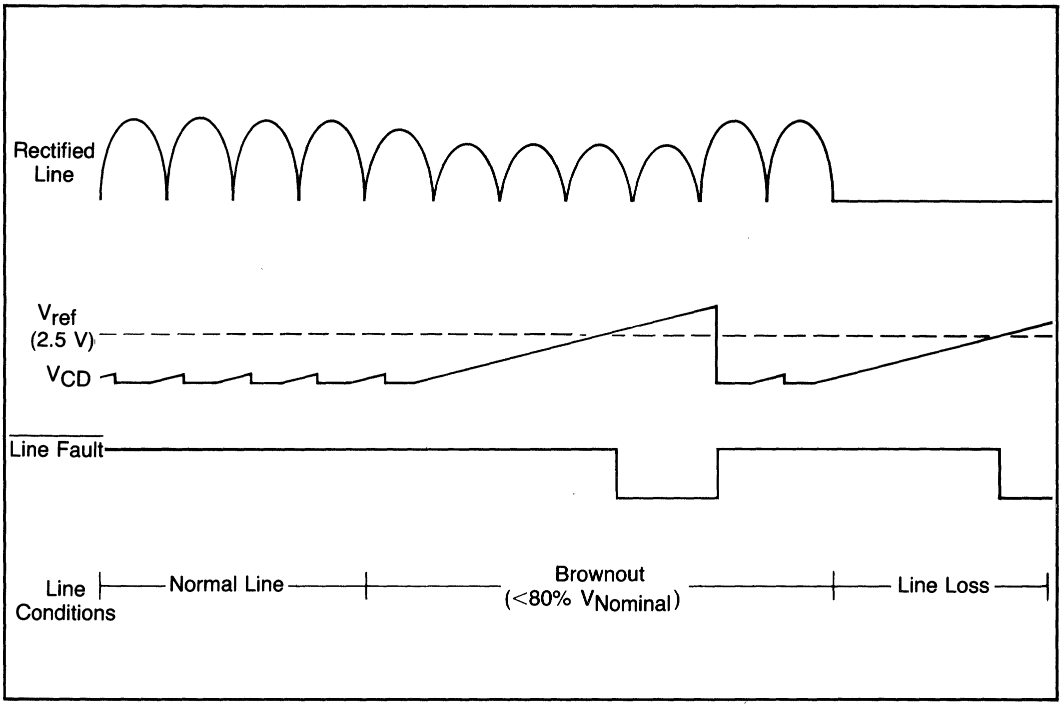


Figure 14-16. Waveforms Illustrating Brownout and Line Loss Detection for the Circuit of Figure 14-14.

REFERENCES

1. ''Characterizing the SCR for Crowbar Applications,'' Al Pshaenich, Motorola AN-789.

2. ''Semiconductor Considerations for DC Power Supply SCR Crowbar Circuits,'' Henry Wurzburg, Third National Sold-State Power Conversion Conference, June 25, 1976.

3. ''Is a Crowbar Enough?'' Willis C. Pierce Jr., Hewlett-Packard, Electronic Design 20, Sept. 27, 1974.

4. ''Transient Thermal Response-General Data and Its Use,'' Bill Roehr and Brice Shiner, Motorola AN-569.

SECTION 15
HEATSINKING

A. THE THERMAL EQUATION

A necessary and primary requirement for the safe operation of any semiconductor device, whether it be an IC or a transistor, is that its junction temperature be kept below the specified maximum value given on its data sheet. The operating junction temperature is given by:

$$T_j = T_A + P_D \theta_{JA} \qquad (15.1)$$

where T_j = junction temperature (°C)

T_A = ambient air temperature (°C)

P_D = power dissipated by device (watts)

θ_{JA} = thermal resistance from junction to ambient air (°C/W)

The junction-to-ambient thermal resistance, θ_{JA}, in Equation (15.1) can be expressed as a sum of thermal resistances as shown below:

$$\theta_{JA} = \theta_{JC} + \theta_{CS} + \theta_{SA} \qquad (15.2)$$

where θ_{JC} = junction-to-case thermal resistance

θ_{CS} = case-to-heatsink thermal resistance

θ_{SA} = heatsink-to-ambient thermal resistance

(Equation (15.2) applies only when an external heatsink is used. If no heatsink is used. θ_{JA} is equal to the device package θ_{JA} given on the data sheet.)

θ_{JC} depends on the device and its package (case) type, while θ_{SA} is a property of the heatsink and θ_{CS} depends on the type of package/heatsink interface employed. Values for θ_{JC} and θ_{SA} are found on the device and heatsink data sheets, while θ_{CS} is given in Table 15-1.

TABLE 15-1

θ_{CS} For Various Packages & Mounting Arrangements

	θ_{CS}			
	METAL-TO-METAL*		**USING AN INSULATOR***	
CASE	**DRY**	**With Heatsink Compound**	**With Heatsink Compound**	**Type**
TO-3	0.2°C/W	0.1°C/W	0.36°C/W	3 mil MICA
			0.28°C/W	Anodized Aluminum
TO-66	1.5°C/W	0.5°C/W	0.9°C/W	2 mil MICA
TO-220	1.2°C/W	1.0°C/W	1.6°C/W	2 mil MICA

*Typical values; heatsink surface should be free of oxidation, paint, and anodization

Examples showing the use of Equations 15.1 and 15.2 in thermal calculations are as follows:

Example 1: Find required heatsink θ_{SA} for an MC7805CT; given:

T_{jmax} (desired) = +125°C

T_{Amax} = +70°C

P_D = 2 watts

Mounted directly to heatsink with silicon thermal grease at interface

1. From MC7805CT data sheet, $\theta_{JC} = 5°C/W$
2. From Table 15-1, $\theta_{CS} = 2.6°C/W$
3. Using Equation 15.1 and 15.2, solve for θ_{SA}:

$$\theta_{SA} = \frac{(T_j - T_A)}{P_D} - \theta_{CS} - \theta_{JC}$$

$$\theta_{SA} = \frac{(125 - 70)}{2} - 5.0 - 2.6$$

$$\leqslant 19.9°C/W \text{ required}$$

Example 2: Find the maximum allowable T_A for an unheatsinked MC78L15CT, given:

T_jmax (desired) = +125°C

P_D = .25 watt

1. From MC78L15CT data sheet, $\theta_{JA} = 200°C/W$
2. Using Equation 15.1 find T_A:

$$\begin{aligned} T_A &= T_j - P_D\,\theta_{JA} \\ &= 125 - .25\,(200) \\ &= +75°C \end{aligned}$$

B. SELECTING A HEATSINK

Usually, the maximum ambient temperature, power being dissipated, the T_{jmax}, and θ_{JC} for the device being used are known. The required θ_{SA} for the heatsink is then determined using Equations 15.1 and 15.2, as in Example 1. The designer may elect to use a commercially available heatsink, or if packaging or economy demands it, design his own.

1. Commercial Heatsinks

As an aid in selecting a heatsink, a representative listing is shown in Table 15-2. This listing is by no means complete and is only included to give the designer an idea of what is available.

TABLE 15-2

Commercial Heatsink Selection Guide

No attempt has been made to provide a complete list of all heatsink manufacturers. This list is only representative.

TO-3 & TO-66	
θ_{SA}*(°C/W)	**Manufacturer/Series or Part Number**
0.3-1.0	Thermalloy — 6441, 6443, 6450, 6470, 6560, 6590, 6660, 6690
1.0-3.0	Wakefield — 641 Thermalloy — 6123, 6135, 6169, 6306, 6401, 6403, 6421, 6423, 6427, 6442, 6463, 6500
3.0-5.0	Wakefield — 621, 623 Thermalloy — 6606, 6129, 6141, 6303 IERC — HP Staver — V3-3-2
5.0-7.0	Wakefield — 690 Thermalloy — 6002, 6003, 6004, 6005, 6052, 6053, 6054, 6176, 6301 IERC — LB Staver — V3-5-2
7.0-10.0	Wakefield — 672 Thermalloy — 6001, 6016, 6051, 6105, 6601 IERC — LA, uP Staver — V1-3, V1-5, V3-3, V3-5, V3-7
10.0-25.0	Thermalloy — 6013, 6014, 6015, 6103, 6104, 6105, 6117

*All values are typical as given by mfgr. or as determined from characteristic curves supplied by manufacturer.

TO-5	
θ_{SA}*(°C/W)	**Manufacturer/Series or Part Number**
12.0-20.0	Wakefield — 260 Thermalloy — 1101, 1103 Staver — V3A-5
20.0-30.0	Wakefield — 209 Thermalloy — 1116, 1121, 1123, 1130, 1131, 1132, 2227, 3005 IERC — LP Staver — F5-5
30.0-50.0	Wakefield — 207 Thermalloy — 2212, 2215, 225, 2228, 2259, 2263, 2264 Staver — F5-5, F6-5
	Wakefield — 204, 205, 208 Thermalloy — 1115, 1129, 2205, 2207, 2209, 2210, 2211, 2226, 2230, 2257, 2260, 2262 Staver — F1-5, F5-5
θ_{SA}*(°C/W)	**CASE TO-220**
5.0-10.0	IERC H P3 Series Staver — V3-7-225, V3-7-96
10.0-15.0	Thermalloy — 6030, 6032, 6034 Staver — V4-3-192, V-5-1
15.0-20.0	Thermalloy — 6106 Staver — V4-3-128, V6-2
20.0-30.0	Wakefield — 295 Thermalloy — 6025, 6107

*All values are typical as given by mfgr. or as determined from characteristic curves supplied by manufacturer.

TO-92	
θ_{SA}*(°C/W)	**Manufacturer/Series or Part Number**
46	Staver F5-7A, F5-8
50	IERC RUR
57	Staver F5-7D
65	IERC RU
72	Staver F1-8, F2-7
80-90	Wakefield 292
85	Thermalloy 2224
DUAL-INLINE-PIN ICS	
20	Thermalloy — 6007
30	Thermalloy — 6010
32	Thermalloy — 6011
34	Thermalloy — 6012
45	IERC — LIC
60	Wakefield — 650, 651

*All values are typical as given by mfgr. or as determined from characteristic curves supplied by manufacturer.

Staver Co., Inc.: 41-51 N. Saxon Ave., Bay Shore, NY 11706
IERC: 135 W. Magnolia Blvd., Burbank, CA 91502
Thermalloy: P.O. Box 34829, 2021 W. Valley View Ln. Dallas, TX
Wakefield Engin Ind: Wakefield, MA 01880

2. Custom Heat Sink Design

Custom heatsinks are usually either forced air cooled or convection cooled. The design of forced air cooled heatsinks is usually done empirically, since it is difficult to obtain accurate air flow measurements. On the other hand, convection cooled heatsinks can be designed with fairly predictable characteristics. It must be emphasized, however, that any custom heatsink design should be thoroughly tested in the actual equipment configuration to be certain of its performance. In the following sections, a design procedure for convection cooled heatsinks is given.

Obviously, the basic goal of any heatsink design is to produce a heatsink with an adequately low thermal resistance, θ_{SA}. Therefore, a means of determining θ_{SA} is necessary in the design. Unfortunately, a precise calculation method for θ_{SA} is beyond the scope of this book.* However, a first order approximation can be calculated for a convection cooled heatsink if the following conditions are met:

1. The heatsink is a flat rectangular or circular plate whose thickness is much smaller than its length or width.
2. The heatsink will not be located near other heat radiating surfaces.
3. The aspect ratio of a rectangular heatsink (length:width) is not greater than 2:1.
4. Unrestricted convective air flow.

For the above conditions, the heatsink thermal resistance can be approximated by:

$$\theta_{SA} \simeq \frac{1}{A\eta\,(F_c h_c + \epsilon H_r)}\ (\text{°C/W}) \qquad (15\text{-}3)$$

where A = area of the heatsink surface

η = heatsink effectiveness

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*If greater precision is desired, or more information on heat flow and heatsinking is sought, consult the references list at the end of this section.

F_c = convective correction factor

h_c = convection heat transfer coefficient

ϵ = emissivity

H_r = normalized radiation heat transfer coefficient

The convective heat transfer coefficient, h_c, can be found from Figure 15-1. Note that it is a function of the heatsink fin temperature rise, $T_S - T_A$, and the heatsink significant dimension, L. The fin temperature rise, $T_S - T_A$, is given by:

$$T_S - T_A = \theta_{SA} P_D \qquad (15.4)$$

where T_S = heatsink temperature

T_A = ambient temperature

θ_{SA} = heatsink-to-ambient thermal resistance

P_D = power dissipated

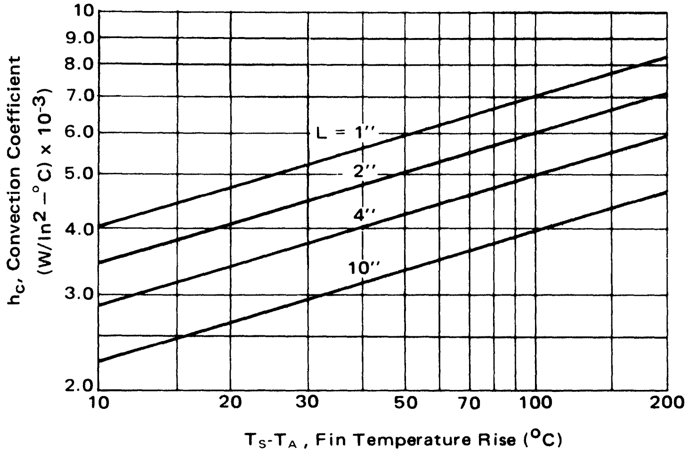


Figure 15-1. Convection Coefficient, h_C

The significant heatsink dimension, L, is dependent on the heatsink shape and mounting place and is given in Table 15-3.

The convective correction factor, F_c, is likewise dependent on shape and mounting plane of the heatsink and is also given in Table 15-3.

TABLE 15-3

Significant Dimension L and Correction Factor F_c for Convection Thermal Resistance

Surface	Significant Dimension L		Correction Factor Fc	
	Position	L	Position	Fc
Rectangular Plane	vertical	height — (max 2 ft)	Vertical Plane	1.0
	horizontal	$\frac{\text{length x width}}{\text{length + width}}$	Horizontal Plane both surfaces exposed	1.35
Circular Plane	vertical	π / 1 x diameter	top only exposed	0.9

The normalized radiation heat transfer coefficient, H_r, is dependent on the ambient temperature, T_A, and the heatsink temperature rise, $T_S - T_A$, given by Equation (15.4). H_r can be determined from Figure 15-2.

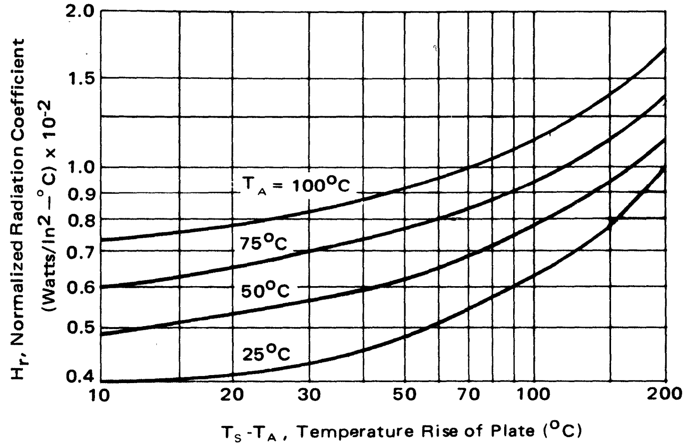


Figure 15-2. Normalized Radiation Coefficient, H_r

The emissivity, ϵ, can be found in Table 15-4 for various heatsink surfaces.

TABLE 15-4.

Typical Emissivities of Common Surfaces

Surface	Emissivity, ϵ
Aluminum, Anodized	0.7 — 0.9
Alodine on Aluminum	0.15
Aluminum, Polished	0.05
Copper, Polished	0.07
Copper, Oxidized	0.70
Rolled Sheet Steel	0.66
Air Drying Enamel (any color)	0.85 — 0.91
Oil Paints (any color)	0.92 — 0.96
Varnish	0.89 — 0.93

Finally, the heatsink efficient, η, can be found from the nomograph of Figure 15-3. Use of the nomograph is as follows:

a. Find $h_T = F_c h_c + \epsilon H_r$ from Figures 15-1, 15-2 and Tables 15-3 and 15-4, and locate this point on the nomograph.

b. Draw a line from h_T through chosen heatsink fin thickness, x, to find α.

c. Determine D for the heatsink shape as given in Figure 15-4 and draw a line from this point through α, which was found in (b), to determine η.

d. If power dissipating element is not located at heatsink's center of symmetry, multiply η by 0.7 (for vertically mounted plates only).

Note that in order to calculate θ_{SA} from Equation (15.3), it is necessary to know the heatsink size. Therefore, in order to arrive at a suitable heatsink design, a trial size is selected, its θ_{SA} evaluated, and the original size reduced or enlarged as necessary. This process is iterated until the smallest heatsink is obtained that has the required θ_{SA}. The following design example is given to illustrate this procedure:

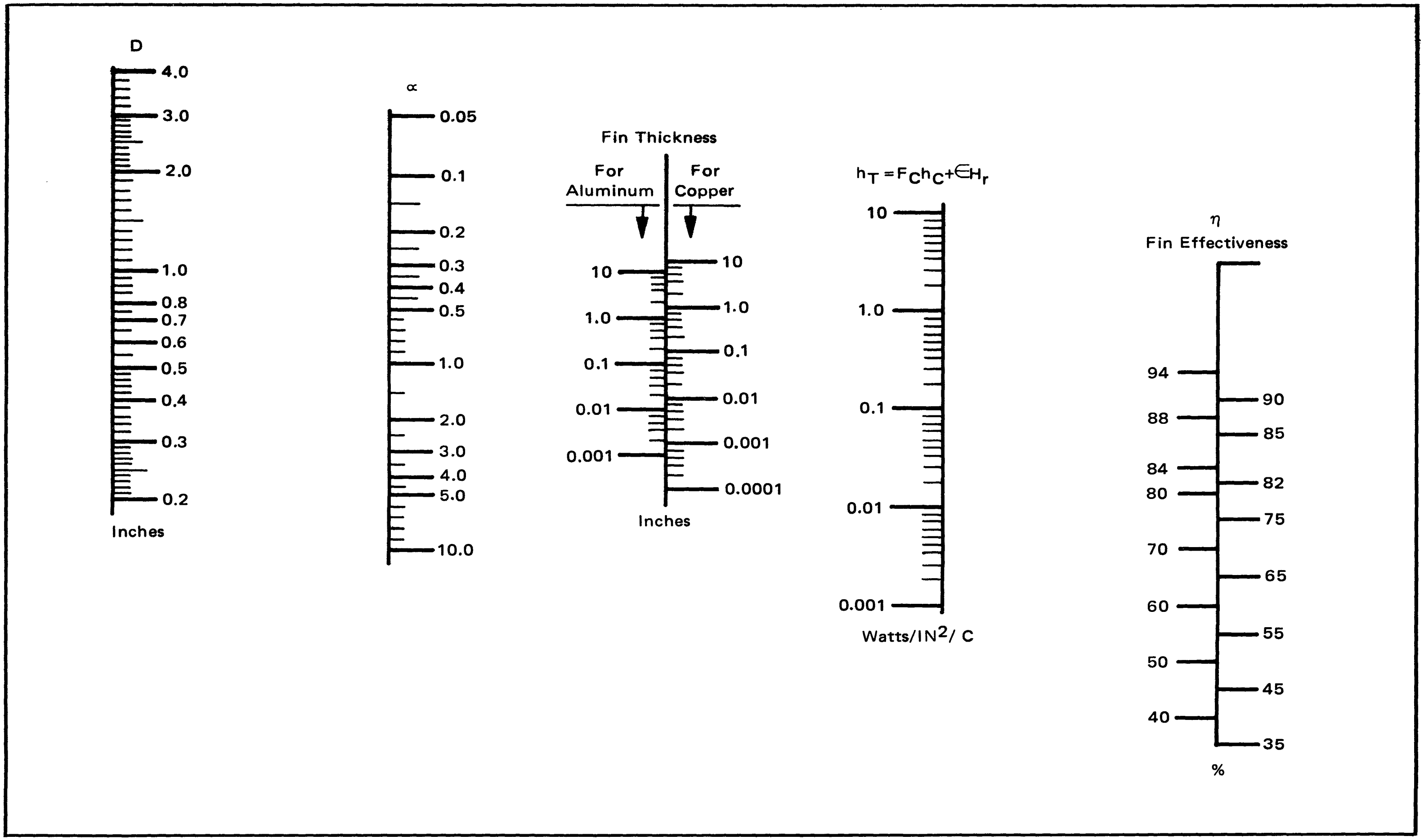


Figure 15-3. Fin Effectiveness Nomogram for Symmetrical Flat, Uniformly Thick Fins

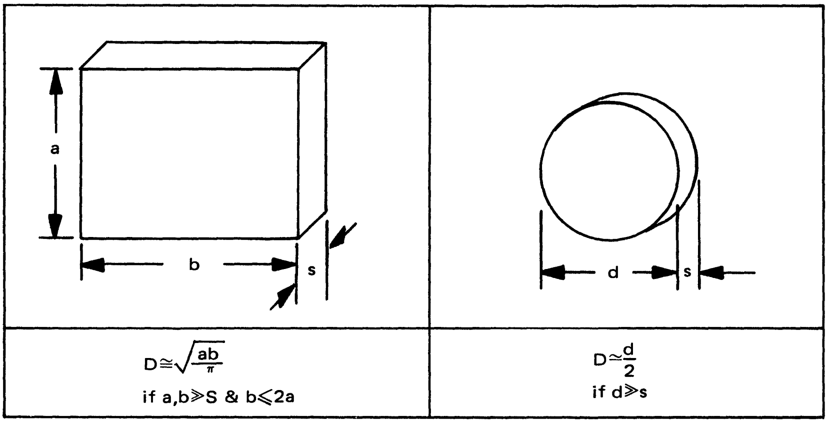


Figure 15-4. Determination of D for Use in η Nomograph of Figure 15-3

Heatsink Design Example

Design a flat rectangular heatsink for use with a horizontally mounted power device on a PC card, given the following:

1. Heatsink $\theta_{SA} = 25°C/W$
2. Power to be dissipated, $P_D = 2W$
3. Maximum ambient temperature, $T_A = 50°C$
4. Heatsink to be constructed from ⅛″ (0.125″) thick anodized aluminum.

a. First, a trial heatsink is chosen: 2″ x 3″ (experience will simplify this selection and reduce the number of necessary iterations.)

b. The factors in Equation (15.3) are evaluated by using the Figures and Tables given.

$A = 2'' \times 3'' = 6$ sq. in.

$L = 6/5'' = 1.2$ in. (from Table 15-3)

$T_S - T_A = 50°C$ (from Equation 15.4)

$h_c = 5.8 \times 10^{-3}$ W/in^2 − °C from Figure 15-1)

$F_c = 0.9$ (from Table 15-3)

$H_r = 6.1 \times 10^{-3}$ W/in^2 − °C (from Figure 15-2)

$\epsilon = 0.9$ (from Table 15-4)

$h_T = F_c h_c + H_r \epsilon = 10.7 \times 10^{-3}$ w/in^2 − °C

$\alpha = 0.13$ (from Figure 15-3)

$D = 1.77$ (from Figure 15-4)

$\eta > 0.94 \simeq 1$ (from Figure 15-3)

c. Using Equation 15.3, find θ_{SA}

$$\theta_{SA} \simeq \frac{1}{A\eta\,(F_c h_c + \epsilon H_r)} = 16.66°C/W < 25°C/W$$

d. Since 2″ x 3″ is too large, try 2″ x 2″. Following the same procedure, θ_{SA} is found to be 25°C/W, which exactly meets the design requirements.

REFERENCES

1. Bill Roehr, ''Motorola Silicon Rectifier Handbook,'' Chapter 10, Motorola Inc., 1973.

2. Werner Luft, ''Taking the Heat Off Semiconductor Devices,'' *Electronics,* June 12, 1959.

3. Frank Kreith, *Principles of Heat Transfer,* International Textbook Co., 1958.

SECTION 16
REGULATOR RELIABILITY

A. QUALITY CONCEPTS

The *quality* of a regulator, from a production line, is a measure that expresses the conformance of the device to a set of specifications. Such a measure is the percent rejects out of a collection of devices (lot, population). *One hundred percent inspection* has to be used to determine the quality of the lot. One characteristic of this approach is that it is expensive, and therefore, is used only where necessary. In addition, it may not be as accurate as it first appears because of operator errors due to fatigue and of course, it cannot be used where the inspection (test) is destructive. An alternative to this is *scientific acceptance sampling*. Acceptance sampling is a method by which a portion of the total population is examined. On the basis of the sample quality, (number of rejects out of a total sample that fail to conform to specifications) and by using the mathematics of probability and statistics, an estimate of the lot quality is made and the risk of an improper decision is specified. For example, a lot may be rejected because the sample quality was less than that prescribed by the mathematics of sampling and our original goal (maximum percent rejects allowed in a lot). Yet, if the lot was one hundred percent inspected, we may find that the actual percent rejects in the lot was less than the maximum percent rejects established as a goal (*Type I improper decision*). In a similar way, the reverse may happen: a lot may be accepted on the basis of the sample quality (sample rejects are fewer than those prescribed by the mathematics of sampling and our goal) and yet, if a 100% inspection was performed, the actual percent rejects in the lot could be more than our established goal (*Type II improper decision*). A sampling plan is specified by the sample size and the maximum allowable defectives (known as the acceptance number (ACCN)).

The risks involved in sampling are described by the operating characteristic (O.C.) curve of the sampling plan. As illustrated by Figure 16-1, this curve shows the probability of acceptance, on the vertical axis, vs the lot quality (percent rejects), on the horizontal axis. Each particular sampling plan will have its own O.C. curve.

Two points on the curve are of interest. The *AQL*, (acceptable quality level), signifies the quality level that will be accepted most of the time (usually this is set at 95%). In other words, the AQL specifies the risk of making the Type I improper decision, that is why it is often referred to as Producer's Risk. The other point on the curve is the *LTPD* (lot tolerance percent defective) which signifies the level of rejects in a lot that is unsatisfactory and should be rejected by the plan most of the time (usually this is set at 10%). This is also known as Consumer's Risk.

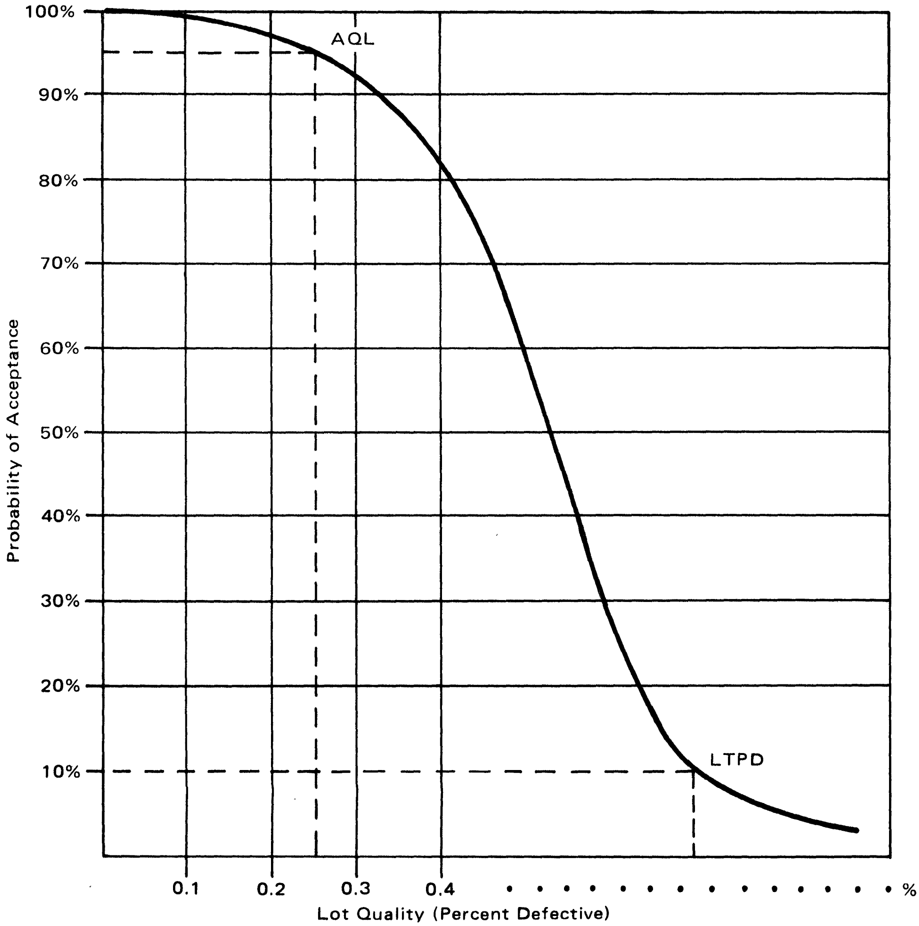


Figure 16-1. Typical Operating Characteristic (O.C.) Curve

Regulators can be produced to a variety of quality levels by combining different 100% and sample inspections and varying the criteria of acceptance and rejection. Thus, a customer can negotiate his own custom quality level if he wishes; however, this can become quite expensive in terms of time and money. That is why Motorola, in addition to the standard product level, produces regulators to four different levels of quality that are similar to those found in the MIL-M-38510 JAN Program processed in accordance with MIL-STD-883. The Motorola program is called MIL-M-38510 JAN Processed Product; a description of the program is beyond the scope of this section, however, Table 16-1 gives the outgoing quality assurance sampling plan for standard quality level regulators. It is important to discern the effects of the different quality levels. This can be done by noting the typical field removal rates (verified rejects plus removed devices verified good) for different classes of 38510 integrated circuits listed below.

	Field Removal Rate/1000 hours
Commercial (no burn-in)	0.1%
Class C	0.04%
Class B	0.004%
Class A	0.002%

TABLE 16-1

Outgoing Quality Assurance Sampling Plan for Regulators Standard Product			
Subgroups (Per Mil-Std-883, Method 5005)	**LTPD**	**ACCN**	**AQL**
A-1: Static Tests, 25°C	2.3	0	
A-2: Static Tests, Max. Temp.	3.8	1	
A-3: Static Tests, Min. Temp.	3.8	1	
A-4: Dynamic Tests, 25°C	2.3	0	
A-5: Dynamic Tests, Max. Temp.	3.8	1	
A-6: Dynamic Tests, Min. Temp.	3.8	1	
A-7: Funct. Test, 25°C	2.3	0	0.11
A-8: Funct. Test, Min/Max Temps.	2.3	0	0.11
A-9: Switching Tests, 25°C	2.3	0	
A-21: Key Parameters, 25°C	2.3	0	0.11

Although the above removal rates are not specifically for regulators, because these products are relatively new with respect to other integrated circuits, nevertheless, it is expected that regulators will have similar removal rates. Burn-in can be used to improve the failure rate of regulators. As a rule of thumb, a 10 to 1 improvement may be realized. This is because regulators are state-of-the-art devices, handling high voltages and currents.

B. RELIABILITY CONCEPTS

Reliability is the probability that a regulator will perform its specified function in a given environment for a specified period of time.The most frequently used reliability measure for regulators is the *failure rate,* expressed in percent per thousand hours. The number of rejects observed, taken over the number of device hours accumulated at the end of the observation period and expressed as a percent, is called the *point estimate* failure rate. This, however, is a number obtained from observations from a portion of all the regulators; if we are to use this number to estimate the failure rate of all regulators (total population), we need to say something about the risk we are taking by using this estimate. This statement is provided by the confidence level expressed together with the failure rate. For example, a 0.1% per 1000 hours failure rate at 90% confidence level means that 90% of the regulators will have a failure rate below 0.1%/1000 hrs — mathematically, the failure rate at a given confidence level is obtained from the point estimate and the CHI square (X^2) distribution. (The X^2 is a statistical distribution used to relate the observed and expected frequencies of an event). In practice, a reliability calculator rule is used that gives the failure rate at the confidence level desired for the number of rejects and device hours under question.

It is also important to note that, as the number of device hours increases, our confidence in the estimate increases. In integrated circuits, it is preferred to make estimates on the basis of 1,000,000,000 device hours or more. If such large numbers of device hours are not available for a particular device, then the point estimate is obtained from devices that are similar in process, voltage, construction, design, etc., and for which we expect to see the same failure modes in the field.

Finally, the *environment* is specified in terms of the *junction temperature* of the regulator by using one of the following two expressions:

$$\text{(A)}\quad T_J = T_A + \theta_{JA} P_D$$

or

$$\text{(B)}\quad T_J = T_C + \theta_{JC} P_D$$

where T_J = Junction Temperature

T_A = Ambient Temperature

T_C = Case Temperature

θ_{JA} = Junction to Ambient Thermal Resistance

θ_{JC} = Junction to Case Thermal Resistance

P_D = Power Dissipation

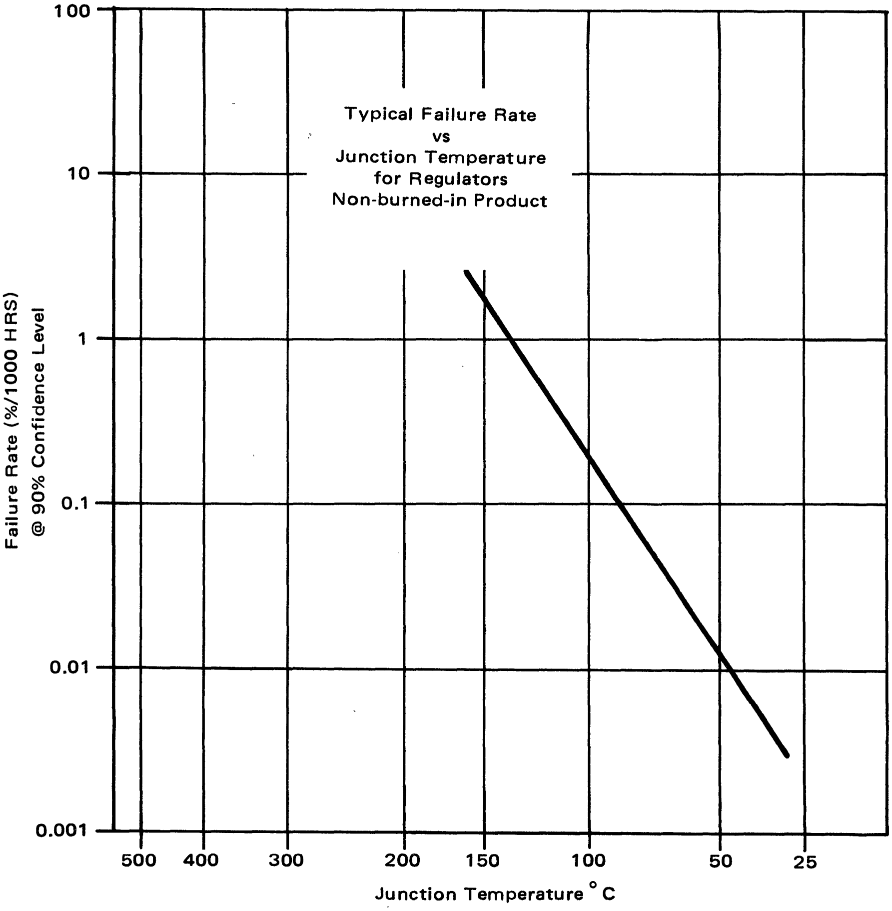


Figure 16-2

One other point worth remembering is that the failure rate for integrated circuits increases as the junction temperature increases while the causes of failure generally remain the same. Thus, we can test devices near their maximum junction temperatures, analyze the failures to assure that they are the types that are accelerated by temperature and then by applying known acceleration factors, estimate the failure rates for lower junction temperatures. Figure 16-2 shows a curve that gives estimates of typical failure rates vs temperature for regulators. To assure that the reliability level does not change over a period of time, Motorola performs a number of periodic audits such as EPIIC. These audit programs, besides monitoring the current reliability level, provide information on what will be required to achieve higher levels of reliability.

Frequently a question is raised about the reliability differences between *plastic vs hermetic regulators*. In general, for all Linear integrated Circuits, including regulators, the field removal rates for plastic and hermetic I/C's are the same for environments where there is no high humidity. In cases where the environment contains high humidity, higher failure rates are to be expected from plastic encapsulated devices. On the other hand, some users have reported favorable results in moderate humidity environments when boards with plastic I/C's (including regulators) are coated with protective materials, provided that the coating is done properly (adhering properly) and no new contaminants are introduced.

SECTION 17

IC REGULATOR SELECTION GUIDES

The selection guides in this section are included as an aid to choosing an appropriate IC regulator. These guides are organized according to regulator type and list all the IC voltage regulators presently offered by Motorola.

A. ADJUSTABLE OUTPUT REGULATORS

When an adjustable output voltage is required, use of the regulators shown in Table 17-1 is recommended. Output voltage is set by adjusting the value of an external resistor or resistors. More complete data on individual devices can be found in the data sheets of Section 18. An explanation of the column headings shown in Table 17-1 follows:

Maximum Output Current ($I_{O\ max}$)

Maximum output current in which key device parameters are specified.

Device

Motorola part number for the IC regulator.

Suffix

Designator for case type; and, in some products, includes temperature range.

Output Voltage (V_{out})

The range of output voltages that can be obtained with the regulator basic circuit configuration. (Methods for extending output voltage range are shown in Section 3.)

Input Voltage (V_{in})

Range of allowable DC input voltages. These are instantaneous values. Exceeding maximum input voltage could result in regulator damage, while dropping below minimum value will cause loss of regulation.

Input-Output Differential (V_{in}–V_{out})

This is the minimum voltage across the regulator for proper operation.

Maximum Power Dissipation ($P_{D\ max}$)

Maximum power the device can dissipate in free air at $T_A = 25°C$ without a heatsink; and with case temperature held constant at $T_C = 25°C$.

Line Regulation (Reg_{line})

The percent change of output voltage for a change in input supply voltage. Given by:

$$\text{Reg}_{\text{line}}\,(\%) = \frac{\Delta V_{out}}{V_{out}} \times \frac{1}{\Delta V_{in}} \times 100$$

where ΔV_{out} = change in V_{out}

ΔV_{in} = change in V_{in}

This performance figure applies for the entire output and input voltage range for the regulator. For actual test conditions, consult data sheets in Section 18.

Load Regulation (Reg_{load})

The percent change of output voltage for a change in output current. For actual test conditions, consult data sheets in Section 18.

Typical Temperature Coefficient of Output Voltage (T_C of V_{out})

Percent change in output voltage per degree Celsius rise in junction temperature.

Maximum Operating Junction Temperature ($T_{J\,max}$)

Maximum junction temperature allowed before damage occurs. For complete thermal information consult data sheets in Section 18. See Section 15 for heatsinking techniques.

Packages

Case 1: "TO-3" metal can
Case 29: "TO-92" plastic package
Case 79: "TO-39" metal can
Case 80-02: "TO-66" metal can
Case 221A: "TO-220" plastic package
Case 603: 10-pin "TO-5" metal can
Case 614: 9-pin "TO-66" metal can
Case 632: 14-pin ceramic dual-in-line package
Case 646: 14-pin plastic dual-in-line package
Case 751A: 14-pin plastic dual-in-line SOIC package

For detailed outline drawings of these case styles, consult Section 19.

TABLE 17-1

ADJUSTABLE OUTPUT REGULATORS

POSITIVE OUTPUT REGULATORS

I_O mA Max	Device Type	SUFFIX	V_{out} Volts		V_{in} Volts		V_{in} — V_{out} Differential Volts Min	P_D Watts Max		Regulation % V_{out} @ T_A = 25°C Typ		TC V_{out} Typ %/°C	T_J = °C Max	Case
			Min	Max	Min	Max		T_A = 25°C	T_C = 25°C	Line	Load			
100	LM317L	H,Z	1.2	37	5.0	40	3.0	Internally Limited		0.04	0.5	0.006	125	29, 79
	LM217L									0.02	0.3	0.004	150	
	LM117L*											0.003		
150	MC1723	CP	2.0	37	9.5	40	3.0	1.25	—	0.1	0.3	0.003	150	646
		CG						1.0	2.1	0.1		0.003		603C
		G								0.2		0.002		
		CL						1.5	—	0.1		0.003	175	632
		L							—	0.2		0.002		
		CD						1.25	—	0.1		0.003	150	751A
250	MC1469	G	2.5	32	9.0	35	3.0	0.68	1.8	0.03	0.13	0.002	150	603
	MC1569			37	8.5	40	2.7			0.015				
500	LM317M	T	1.2	37	5.0	40	3.0	Internally Limited		0.02	0.1	0.0056	125	221A
	LM317M	R												80
	LM217M											0.004	150	
	LM117M*											0.0036		
600	MC1469	R	2.5	32	9.0	35	3.0	3.0	14.0	0.03	0.05	0.002	150	614
	MC1569			37	8.5	40	2.7			0.015				
1500	LM317	T	1.2	37	5.0	40	3.0	Internally Limited		0.07	1.5	0.006	125	221A
	LM317	H, K												79, 1
	LM217											0.004		
	LM117*									0.05	1.0	0.003	150	
3000	LM350	T	1.2	33	5.0	36	3.0	Internally Limited		0.02	0.1	0.008	125	221A
	LM350	K												1
	LM250											0.0057	150	
	LM150*											0.0051		

#T_J = −40 to +125°C

*T_J = −55 to +150°C

†Output Voltage Tolerance for Worst Case

NEGATIVE OUTPUT REGULATORS

I_O mA Max	Device Type	SUFFIX	V_{out} Volts		V_{in} Volts		V_{in} — V_{out} Differential Volts Min	P_D Watts Max		Regulation % V_{out} @ T_A = 25°C Typ		TC V_{out} Typ %/°C	T_J = °C Max	Case
			Min	Max	Min	Max		T_A = 25°C	T_C = 25°C	Line	Load			
250	MC1463	G	−3.8	−32	9.0	35	3.0	0.68	1.8	0.03	0.05	0.002	150	603
	MC1563		−3.6	−33	8.5	40	2.7			0.015	0.13			
600	MC1463	R	−3.8	−34	9.0	35	3.0	2.4	9.0	0.03	0.05	0.002	175	614
	MC1563		−3.6	−37	8.5	40	2.7			0.015				
1500	LM337	T	−1.2	−37	5.0	40	3.0	Internally Limited		0.02	0.3	0.0048	125	221A
	LM337	H, K												79, 1
	LM237											0.0034	150	
	LM137*											0.0031		

*T_J = −55 to +150°C

B. FIXED OUTPUT REGULATORS

If low cost and easy implementation are prime regulator design considerations, the fixed output, three terminal regulators shown in Table 17-2 are recommended. These are available with output current capabilities from 100 mA to 3.0 A. All have internal overcurrent, safe-operating area, and thermal protection circuitry. Complete device specifications are given in the data sheets of Section 18. An explanation of the column headings shown in Table 17-2 follows:

Output Voltage (V_{out})

Nominal output voltage for positive and negative regulators. The adjacent column indicates worst case tolerance (Volts). (Methods for adjusting output voltage are shown in Section 3.)

Maximum Output Current ($I_{O\ max}$)

Maximum output current available from regulator under normal operating conditions. (Methods for obtaining greater output currents are shown in Section 3.)

Device

Two columns are provided listing Motorola part numbers for positive and negative voltage outputs.

Input Voltage min/max (V_{in})

Range of allowable instantaneous dc input voltage. Exceeding maximum V_{in} could result in regulator damage, while dropping below minimum value will cause loss of regulation.

Line Regulation (Reg_{line})

Change in output voltage for a given change in input voltage. Test specifications are given in the data sheets of Section 18.

Load Regulation (Reg_{load})

Change in output voltage for a given change in output current. Test specifications are given in the data sheets of Section 18.

Typical Temperature Coefficient of Output Voltage ($\Delta V/\Delta T$)

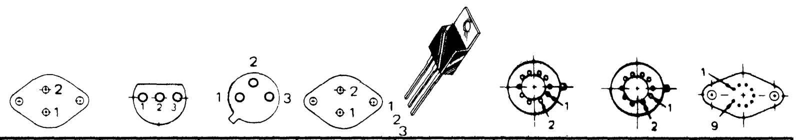
Typical change in output voltage per degree celsius change in junction temperature.

Packages

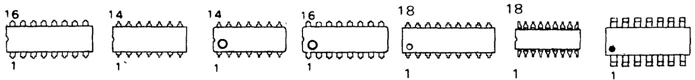
Case 1: "TO-3" metal can
Case 29: "TO-92" plastic package
Case 79: "TO-39" metal can
Case 221A: "TO-220" plastic package

For detailed outline drawings of these case styles, consult Section 19.

Package Styles



CASE	1 (TO-3)	29 (TO-92)	79 (TO-39)	80 (TO-66)	221A (TO-220)	603 (TO-5 Type)	603C	614 (TO-66)
MATERIAL	Metal	Plastic	Metal	Metal	Plastic	Metal	Metal	Metal
SUFFIX	K	P, Z	G, H	R	T	G	G	R



CASE	620	632 (TO-116)	646	648	707	726	751A
MATERIAL	Ceramic	Ceramic	Plastic	Plastic	Plastic	Ceramic	Plastic
SUFFIX	J, L	L	P	N, P	N	J	D

TABLE 17-2

FIXED OUTPUT VOLTAGE REGULATORS

FIXED/VOLTAGE, 3-TERMINAL REGULATORS FOR POSITIVE OR NEGATIVE POLARITY POWER SUPPLIES.

V_{out} Volts	Tol.† Volts	I_O mA Max	Device Type Positive Output	Device Type Negative Output	V_{in} Min/Max	Reg_{line} mV	Reg_{load} mV	$\Delta V_O/\Delta T$ mV/°C Typ	Case
2	±0.1	1500	—	MC7902C	5.5/35	40	120	1.0	1, 221A
3	±0.15	100	—	MC79L03AC	4.7/30	60	72	—	29, 79
	±0.3			MC79L03C		80			
5	±0.5	100	MC78L05C	MC79L05C	6.7/30	200	60	—	29, 79
	±0.25		MC78L05AC	MC79L05AC		150			
		500	MC78M05C	—	7/35	100	100	1.0	79, 221A
	±0.4	1500	LM109	—				1.1	1, 79
			LM209	—					
	±0.25		LM309	—		50		1.0	
	±0.35		MC7805*	—	8.0/35			0.6	1
	±0.25		MC7805B#	—	8/35	100		1.0	1, 221A
			MC7805C	MC7905C	7/35				
	±0.2		MC7805A*	—	7.5/35	10	50	0.6	1
			MC7805AC	MC7905AC			100		1, 221A
	±0.25		LM140-5*	—	7/35	50	50		1
			LM340-5	—					
		3000	MC78T05*	—	7.3/35	10	25	0.1	1
			MC78T05C	—					
	±0.2								1, 221A
			MC78T05A*	—					1
			MC78T05AC	—					1, 221A
	±0.4		LM123*	—	7.5/20	5.0	25	—	1
			LM223	—					
	±0.25		LM323	—					

(continued)

Fixed Output Voltage Regulators (continued)

V_{out} Volts	Tol.† Volts	I_O mA Max	Device Type Positive Output	Device Type Negative Output	V_{in} Min/Max	Regline mV	Regload mV	$\Delta V_O/\Delta T$ mV/°C Typ	Case
5.2	±0.26	1500	—	MC7905.2C	7.2/35	105	105	1.0	1, 221A
6	±0.3	500	MC78M06C	—	8/35	100	120	1.0	79, 221A
	±0.35	1500	MC7806*	—	9/35	60	100	0.7	1
	±0.3		MC7806B#	—	9/35	120	120		1,221A
			MC7806C	MC7906C	8/35				
	±0.24		MC7806A*	—	8.6/35	11	50		1
			MC7806AC	—			100		1, 221A
	±0.3		LM140-6*	—	8/35	60	60		1
			LM340-6	—					
		3000	MC78T06*	—	8.3/35	11	25	0.12	1
			MC78T06C	—					1, 221A
8	±0.8	100	MC78L08C	—	9.7/30	200	80	—	29, 79
			MC78L08AC	—		175			
	±0.4	500	MC79M08C	—	10/35	100	160	1.0	79, 221A
		1500	MC7808*	—	11.5/35	80	100		1
			MC7808B#	—	11.5/35	160	160		1, 221A
			MC7808C	MC7908C	10.5/35				
	±0.3		MC7808A*	—	10.6/35	13	50		1
			MC7808AC	—			100		1, 221A
	±0.4		LM140-8*	—	10.5/35	80	80		1
			LM340-8	—					
		3000	MC78T08*	—	10.4/35	13	25	0.16	1
			MC78T08C	—					1, 221A
12	±1.2	100	MC78L12C	MC79L12C	13.7/35	250	100	—	29, 79
	±0.6		MC78L12AC	MC79L12AC					
		500	MC78M12C	—	14/35	100	240	1.0	79, 221A
		1500	MC7812*	—	15.5/35	120	120	1.5	1
			MC7812B#	—		240	240		1, 221A
			MC7812C	MC7912C	14.5/35				
	±0.5		MC7812A*	—	14.8/35	18	50		1
			MC7812AC	—			100		1, 221A
	±0.6		LM140-12*	—	14.5/35	120	120	1.5	1
			LM340-12	—					
		3000	MC78T12*	—	14.5/35	18	25	0.24	1
			MC78T12C	—					1, 221A
	±0.5		MC78T12A*	—					1
			MC78T12AC	—					1, 221A
15	±1.5	100	MC78L15C	MC78L15C	16.7/35	300	150	—	29, 79
	±0.75		MC78L15AC	MC78L15A					
		500	MC78M15C	—	17/35	100	300	1.0	79, 221A
		1500	MC7815*	—	18.5/35	150	150	1.8	1
			MC7815B#	—		300	300		1, 221A
			MC7815C	MC7915C	17.5/35				
	±0.6		MC7815A*	—	17.9/35	22	50		1
			MC7815AC	—			100		1, 221A
	±0.75		LM140-15*	—	17.5/35	150	150		1
			LM340-15	—					
		3000	MC78T15*	—	17.5/40	22	25	0.3	1
			MC78T15C	—					1, 221A
	±0.6		MC78T15A*	—					1
			MC78T15AC	—					1, 221A

(continued)

Fixed Output Voltage Regulators (continued)

V_{out} Volts	Tol.† Volts	I_O mA Max	Device Type Positive Output	Device Type Negative Output	V_{in} Min/Max	Reg_{line} mV	Reg_{load} mV	$\Delta V_O/\Delta T$ mV/°C Typ	Case
18	±1.8	100	MC78L18C	MC79L18C	19.7/35	325	170	—	29, 79
	±0.9		MC78L18AC	MC79L18AC					
		500	MC78M18C	—	20/35	100	360	1.0	79, 221A
		1500	MC7818*	—	22/35	180	180	2.3	1
			MC7818B#	—		360	360		1, 221A
			MC7818C	MC7918C	21/35				
	±0.7		MC7818A*	—		31	50		1
			MC7818AC	—			100		1, 221A
	±0.9		LM140-18*	—		180	180		1
			LM340-18	—					
		3000	MC78T18*	—	20.6/40	31	25	0.36	1
			MC78T18C	—					1, 221A
20	±1.0	500	MC78M20C	—	22/40	10	400	1.1	79, 221A
24	±2.4	100	MC78L24C	MC79L24C	25.7/40	350	200	—	29, 79
	±1.2		MC78L24AC	MC79L24AC		300			
		500	MC78M24C	—	26/40	100	480	1.2	79, 221A
		1500	MC7824*	—	28/40	240	240	3.0	1
			MC7824B#	—		480	480		1, 221A
			MC7824C	MC7924C	27/40				
	±1.0		MC7824A*	—	27.3/40	36	50		1
			MC7824AC	—			100		1, 221A
	±1.2		LM140-24*	—		240	240		1
			LM340-24	—					
		3000	MC78T24*		26.7/40	36	25	0.48	1
			MC78T24C						1, 221A

#T_J = −40 to +125°C
*T_J = −55 to +150°C
†Output Voltage Tolerance for Worst Case

C. SPECIALTY REGULATORS AND SWITCHING REGULATOR CONTROL CIRCUITS

In addition to the regulators of Tables 17-1 and 17-2, Motorola offers two specialty regulators: the MC1568/MC1468 ±15 V Tracking regulator and the MC1466 Precision Floating regulator. General specifications for these regulators are shown in Table 17-3. More complete data on these devices can be found in the data sheets of Section 18. An explanation of the column headings shown in Table 17-3 follows:

Device

Motorola part number for the IC regulator. (No symbol indicates 0°C to +70°C operating ambient temperature range. * indicates −55°C to +125°C operating ambient temperature range.)

Output Voltage (V_O)

For the tracking regulators, the value of the preset output voltage. (Methods for obtaining adjustable output voltages are shown in Section 3.)

For the floating regulators, the range of output voltages that can be obtained with the regulator.

* Indicates that the maximum obtainable output voltage is dependent only on the characteristics of the external pass element.

Maximum Output Current ($I_{O\ max}$)

Absolute maximum output current that can be obtained without damaging regulator. (Methods for obtaining increased output current are shown in Section 3.)

* Indicates that the maximum obtainable output current is dependent only on the characteristics of the external pass element.)

Input Voltage (V_{in})

The range of allowable DC input voltage. This is an instantaneous value. Exceeding maximum V_{IN} could result in regulator damage, while dropping below minimum value will cause loss of regulation.

Auxiliary Supply Voltage (V_{aux})

The floating regulators require an additional dedicated voltage source which is floating with respect to the output ground. The values given are the limits for this auxiliary supply voltage.

Line Regulation (Reg_{line})

Percent change in output voltage for a given change in input voltage. Test specifications are given in the data sheets of Section 18.

Load Regulation (Reg_{load})

Percent change in output voltage for a given change in output current. Test specifications are given in the data sheets of Section 18.

Load Current Regulation

Percent change in output current for a given change in load voltage while in the current regulation mode. Test specifications are given in the data sheets of Section 18.

Typical Temperature Coefficient of Output Voltage (TC of Vo)

Typical percent change in output voltage per degree Celsius change in junction temperature.

Maximum Power Dissipation ($P_{D}max$)

Maximum power which device can safely dissipate when case temperature is held at +25°C; and junction temperature is at its maximum value of +125°C. For complete thermal information, consult data sheets in Section 18. For heat sinking information, see Section 15.

Package

Case 603C: 10-pin "TO-5" type metal can
Case 614: 9-pin "TO-66" type can
Case 632: 14-pin ceramic dual-in-line package

For detailed outline drawings of these case styles, consult Section 18.

TABLE 17-3

SPECIALTY REGULATORS

FLOATING REGULATORS

DEVICE	OUTPUT VOLTAGE (V_O)		MAX OUTPUT CURRENT I_OMAX	AUXILIARY VOLTAGE		LINE REGULATION	LOAD REGULATION	CURRENT REGULATION	TYPICAL TC OF V_O	P_DMAX	PACKAGE
	MIN	MAX		MIN	MAX						
MC1566L*	0	*	*	20V	35V	.01% + 1mV	.01% + 1mV	.1% + 1mA	±.006%/°C	.75W	632
MC1466L	0	*	*	21V	30V	.03% + 3mV	.03% + 3mV	.1% + 1mA	±.01%/°C	.75W	632

TRACKING REGULATORS

DEVICE	OUTPUT VOLTAGE (V_O)		MAX OUTPUT CURRENT I_OMAX	INPUT VOLTAGE (V_{in})		LINE REGULATION %V_O	LOAD REGULATION %V_O	TYPICAL TC of V_O	P_DMAX	PACKAGE
	MIN	MAX		MIN	MAX					
MC1568G*	±14.8V	±15.2V	±100mA	±17V	±30V	.13%	.2%	±.006%/°C	.8W	603C
MC1568L*	±14.8V	±15.2V	±100mA	±17V	±30V	.13%	.2%	±.006%/°C	1.0W	632
MC1568R*	±14.8V	±15.2V	±100mA	±17V	±30V	.13%	.2%	±.006%/°C	2.4W	614
MC1468G	±14.5V	±15.5V	±100mA	±17V	±30V	.13%	.2%	±.013%/°C	.8W	603C
MC1468L	±14.5V	±15.5V	±100mA	±17V	±30V	.13%	.2%	±.013%/°C	1.0W	632
MC1468R	±14.5V	±15.5V	±100mA	±17V	±30V	.13%	.2%	±.013%/°C	2.4W	614

Switching Regulator Control Circuits

Motorola offers a complete line of switching regulator I.C.s to meet the various demands of the market. Table 17-4 lists devices offered along with key parameters. For detailed specifications, refer to Section 18.

An explanation of the column headings shown in Table 17-4 follows:

Maximum Output Current ($I_{O\ max}$)

This is the maximum output current capability of the switching control circuit outputs. Most of the devices have dual push-pull outputs, except for the MC34060/35060 and μA78S40 devices which are single ended.

Supply Voltage (V_{CC}) min/max

Minimum applied voltage to V_{CC} in which normal operation occurs. Maximum applied voltage to V_{CC}, beyond which damage to the I.C. can occur. The TL495 has an internal 39 volt zener and therefore can be operated from supplies greater than 40 volts with a series current limiting resistor. For detail specifications, refer to Section 18.

Oscillator Frequency (f_o)

The range in which the oscillator will operate to effectively drive the internal logic and outputs.

Package

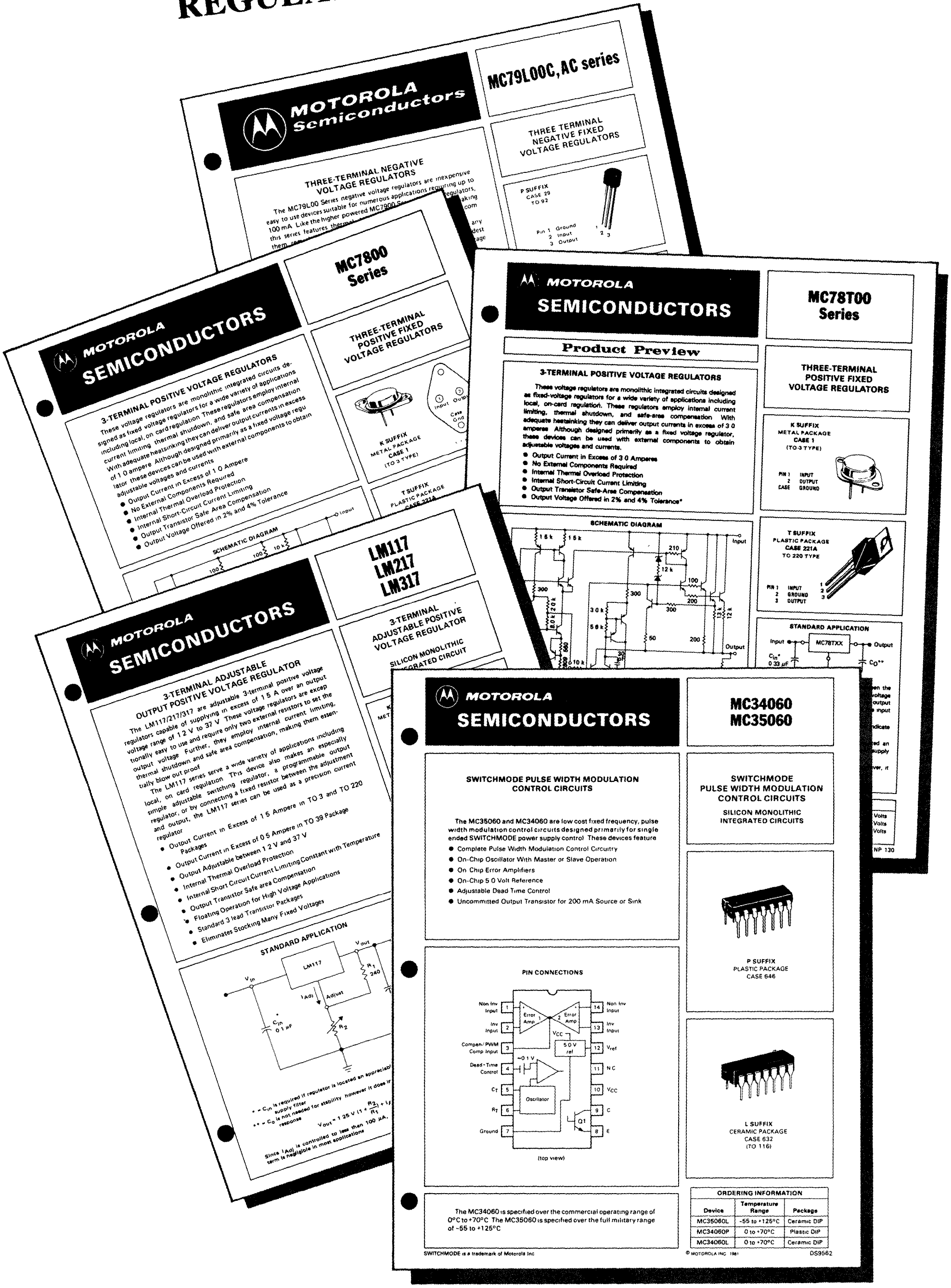
Case 620: 16-pin ceramic dual-in-line package
Case 632: 14-pin ceramic dual-in-line package
Case 646: 14-pin plastic dual-in-line package
Case 648: 16-pin plastic dual-in-line package
Case 701: 18-pin plastic dual-in-line package
Case 726: 18-pin ceramic dual-in-line package

TABLE 17-4
SWITCHING REGULATOR CONTROL CIRCUITS

I_O mA Max	V_{CC} Volts		f_o kHz		Device Number	Suffix	T_A °C	Case
	Min	Max	Min	Max				
40	10	30	2.0	100	MC3420	P	0 to +70	648
						L		620
					MC3520	L	−55 to +125	620
250*	7.0	40	1.0	300	MC34060	P	0 to + 70	646
						L		632
					MC35060	L	55 to +125	632
250	7.0	40	1.0	300	TL494	CN	0 to +70	648
						CJ		620
						IN	−25 to +.85	648
						IJ		620
						MJ	−55 to +125	620
250		>40	1.0	300	TL495	CN	0 to +70	707
						CJ		726
						IN	−25 to +85	707
						IJ	−25 to +85	726
±400	8	40	0.1	400	SG3525A	N	0° to +70	648
					SG3525A	J	0 to +70	620
					SG2525A	N	−40 to +85	648
					SG2525A	J		620
					SG1525A	J	−55 to +125	620
±400	8	40	0.1	400	SG3527A	N	0 to +70	648
					SG3527A	J		620
					SG2527A	N	−40 to +85	648
					SG2527A	J		620
					SG1527A	J	−55 to +125	620
±200	8	40	0.001	400	SG3526	N	0 to +70	707
					SG3526	J		726
					SG2526	N	−40 to +85	707
					SG2526	J		726
					SG1526	J	−55 to +125	726
1500	5	40	1	40	µA78S40	PC	0 to +70	648
					µA78S40	DC		620
					µA78S40	DM	−55 to +125	

*Single output device
**Internal 39 V zener for <40 volt operation

SECTION 18
REGULATOR DATA SHEETS



MOTOROLA SEMICONDUCTORS

MC78T00 Series

Product Preview

3-TERMINAL POSITIVE VOLTAGE REGULATORS

These voltage regulators are monolithic integrated circuits designed as fixed-voltage regulators for a wide variety of applications including local, on-card regulation. These regulators employ internal current limiting, thermal shutdown, and safe-area compensation With adequate heatsinking they can deliver output currents in excess of 3 0 amperes Although designed primarily as a fixed voltage regulator, these devices can be used with external components to obtain adjustable voltages and currents.

- Output Current in Excess of 3 0 Amperes
- No External Components Required
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Output Voltage Offered in 2% and 4% Tolerance*

THREE-TERMINAL POSITIVE FIXED VOLTAGE REGULATORS

K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 TYPE)

PIN 1 INPUT
2 OUTPUT
CASE GROUND

T SUFFIX
PLASTIC PACKAGE
CASE 221A
TO 220 TYPE

PIN 1 INPUT
2 GROUND
3 OUTPUT

SCHEMATIC DIAGRAM

STANDARD APPLICATION

MOTOROLA SEMICONDUCTORS

MC34060
MC35060

SWITCHMODE PULSE WIDTH MODULATION CONTROL CIRCUITS

The MC35060 and MC34060 are low cost fixed frequency, pulse width modulation control circuits designed primarily for single ended SWITCHMODE power supply control These devices feature

- Complete Pulse Width Modulation Control Circuitry
- On-Chip Oscillator With Master or Slave Operation
- On Chip Error Amplifiers
- On-Chip 5 0 Volt Reference
- Adjustable Dead Time Control
- Uncommitted Output Transistor for 200 mA Source or Sink

SWITCHMODE PULSE WIDTH MODULATION CONTROL CIRCUITS

SILICON MONOLITHIC INTEGRATED CIRCUITS

PIN CONNECTIONS

(top view)

P SUFFIX
PLASTIC PACKAGE
CASE 646

L SUFFIX
CERAMIC PACKAGE
CASE 632
(TO 116)

The MC34060 is specified over the commercial operating range of 0°C to +70°C The MC35060 is specified over the full military range of -55 to +125°C

ORDERING INFORMATION		
Device	Temperature Range	Package
MC35060L	-55 to +125°C	Ceramic DIP
MC34060P	0 to +70°C	Plastic DIP
MC34060L	0 to +70°C	Ceramic DIP

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DS9562

MOTOROLA

LM109
LM209
LM309

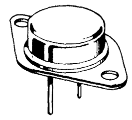
POSITIVE VOLTAGE REGULATOR

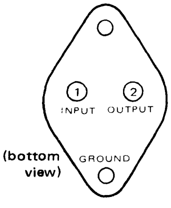
MONOLITHIC POSITIVE THREE - TERMINAL FIXED VOLTAGE REGULATOR

A versatile positive fixed +5.0-volt regulator designed for easy application as on on-card, local voltage regulator for digital logic systems. Current limiting and thermal shutdown are provided to make the units extremely rugged.

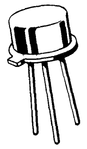
In most applications only one external component, a capacitor, is required in conjunction with the LM109 Series devices. Even this component may be omitted if the power-supply filter is not located an appreciable distance from the regulator.

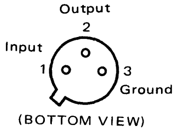
- High Maximum Output Current – Over 1.0 Ampere in TO-3 type Package – Over 200 mA in TO-39 type Package.
- Minimum External Components Required
- Internal Short-Circuit Protection
- Internal Thermal Overload Protection
- Excellent Line and Load Transient Rejection
- Designed for Use with Popular MDTL and MTTL Logic





K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 Type)

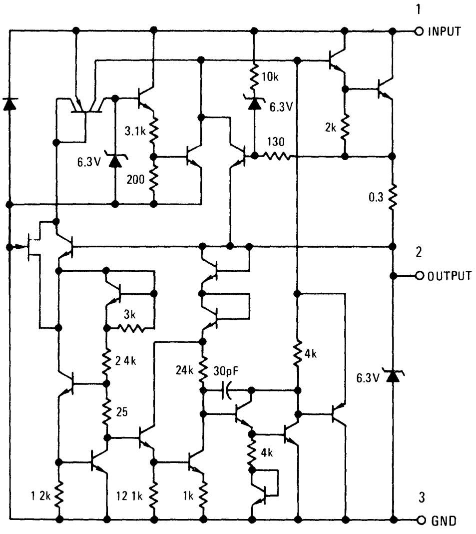




H SUFFIX
METAL PACKAGE
CASE 79
(TO-39)

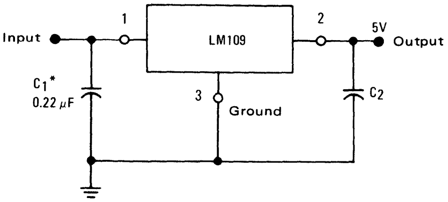
ORDERING INFORMATION		
Device	**Temperature Range**	**Package**
LM109H	$T_J = -55^{\circ}C$ to $+150^{\circ}C$	Metal Can
LM109K	$T_J = -55^{\circ}C$ to $+150^{\circ}C$	Metal Power
LM209H	$T_J = -55^{\circ}C$ to $+150^{\circ}C$	Metal Can
LM209K	$T_J = -55^{\circ}C$ to $+150^{\circ}C$	Metal Power
LM309H	$T_J = 0^{\circ}C$ to $+125^{\circ}C$	Metal Can
LM309K	$T_J = 0^{\circ}C$ to $+125^{\circ}C$	Metal Power

CIRCUIT SCHEMATIC



TYPICAL APPLICATION

FIXED 5.0 V REGULATOR



*Required if regulator is located an appreciable distance from power supply filter.
Although no output capacitor is needed for stability, it does improve transient response.

LM109, LM209, LM309

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Input Voltage	V_{in}	35	Vdc
Power Dissipation	P_D	Internally Limited	
Junction Temperature Range LM109 LM209 LM309	T_J	 -55 to +150 -55 to +150 0 to +125	°C
Storage Temperature Range	T_{stg}	-65 to +150	°C
Lead Temperature (soldering, t = 60 s)	T_S	300	°C

ELECTRICAL CHARACTERISTICS

Characteristic	Symbol	LM109/LM209 ①			LM309 ②			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	4.7	5.05	5.3	4.8	5.05	5.2	Vdc
Input Regulation (T_J = +25°C) 7 0 $\leqslant V_{in} \leqslant$ 25 V	Reg_{in}	–	4 0	50	–	4.0	50	mV
Load Regulation (T_J = +25°C) Case 11-01 **(type TO-3)** 5 0 mA $\leqslant I_O \leqslant$ 1 5A Case 79-02 **(TO-39)** 5 0 mA $\leqslant I_O \leqslant$ 0 5A	Reg_{load}	 – –	 50 20	 100 50	 – –	 50 20	 100 50	mV
Output Voltage Range 7 0 V $\leqslant V_{in} \leqslant$ 25 V 5 0 mA $\leqslant I_O \leqslant I_{max}$, P $\leqslant P_{max}$	V_O	4 6	–	5.4	4 75	–	5.25	Vdc
Quiescent Current (7 0 V $\leqslant V_{in} \leqslant$ 25 V) Quiescent Current Change (7 0 V $\leqslant V_{in} \leqslant$ 25 V) 5.0 mA $\leqslant I_O \leqslant I_{max}$	I_B ΔI_B	– – –	5.2 – –	10 0 5 0 8	– – –	5 2 – –	10 0.5 0.8	mAdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_N	–	40	–	–	40	–	µV
Long Term Stability	S	–	–	10	–	–	20	mV
Thermal Resistance, Junction to Case ③ Case 1 (type TO-3) Case 79-02 (TO-39)	θJC	 – –	 3.0 15	 – –	 – –	 3.0 15	 – –	°C/W

NOTES

① Unless otherwise specified, these specifications apply for -55°C $\leqslant T_J \leqslant$ +150° (-25°C $\leqslant T_J \leqslant$ +150°C for the LM209) For Case 79 02 (TO-39) V_{in} = 10 V, I_O = 0 1 A, I_{max} = 0 2 A and P_{max} = 2 0 W For Case 1 **(type TO-3)** V_{in} = 10 V, I_O = 0 5 A, I_{max} = 1 0 A and P_{max} = 20 W

② Unless otherwise specified, these specifications apply for 0°C $\leqslant T_J \leqslant$ +125°C, V_{in} = 10V For Case 79 02 (TO-39) I_O = 0 1A, I_{max} = 0 2A and P_{max} = 2 0 W For Case 1 (type TO-3) I_O = 0 5 A, I_{max} = 1 0 A and P_{max} = 20 W

③ Without a heat sink, the thermal resistance of the Case 79 02 **(TO-39)** package is about 150°C/W, while that of the Case 1 (type TO 3) package is aoproximately 35°C/W With a heat sink, the effective thermal resistance can only approach the values specified, depending on the efficiency of the heat sink

TYPICAL CHARACTERISTICS

(V_{in} = 10 V, T_A = +25°C unless otherwise noted.)

FIGURE 1 – MAXIMUM AVERAGE POWER DISSIPATION (LM109K, LM209K)

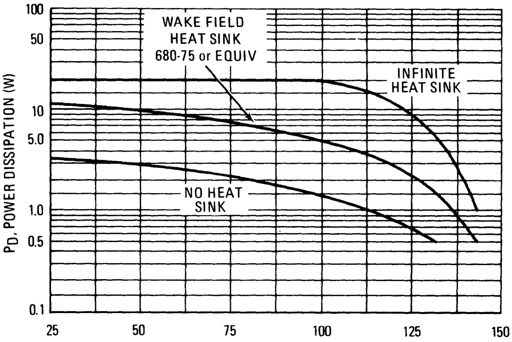
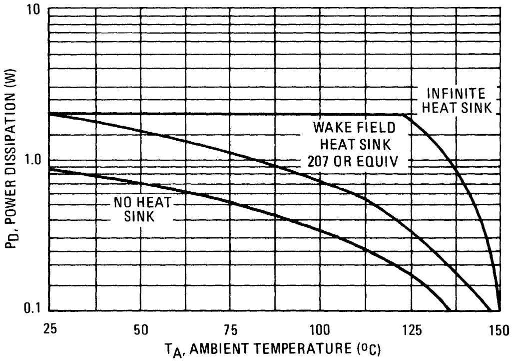


FIGURE 2 – MAXIMUM AVERAGE POWER DISSIPATION (LM109H, LM209H)



TYPICAL CHARACTERISTICS (continued)

(V_{in} = 10 V, T_A = +25°C unless otherwise noted.)

FIGURE 3 – MAXIMUM AVERAGE POWER DISSIPATION (LM309K)

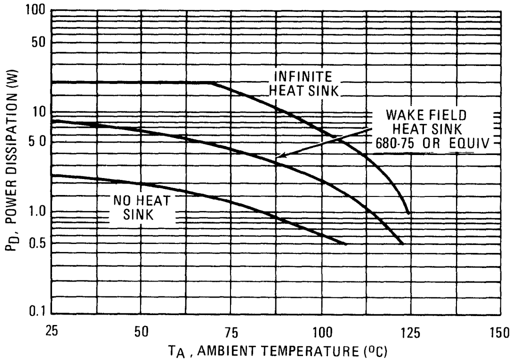


FIGURE 4 – MAXIMUM AVERAGE POWER DISSIPATION (LM309H)

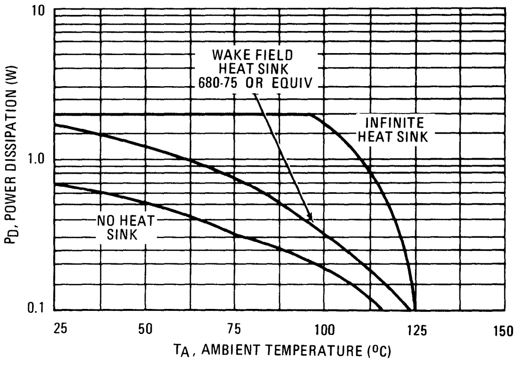


FIGURE 5 - OUTPUT IMPEDANCE versus FREQUENCY

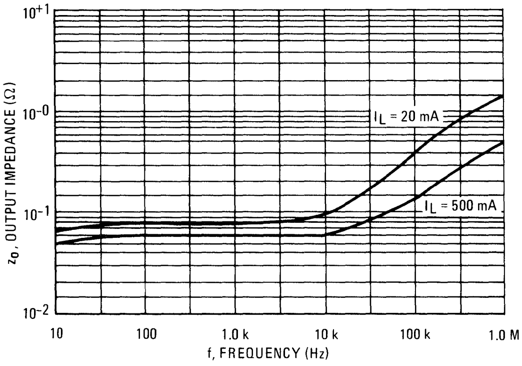


FIGURE 6 – PEAK OUTPUT CURRENT (K PACKAGE)

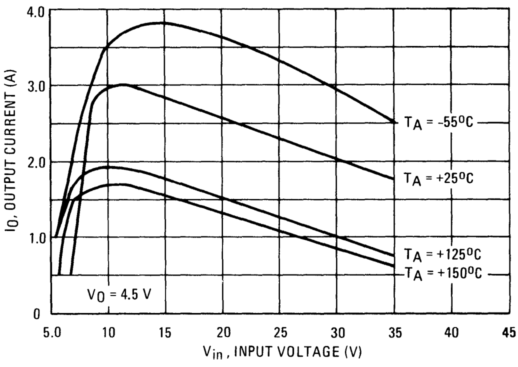


FIGURE 7 – PEAK OUTPUT CURRENT (H PACKAGE)

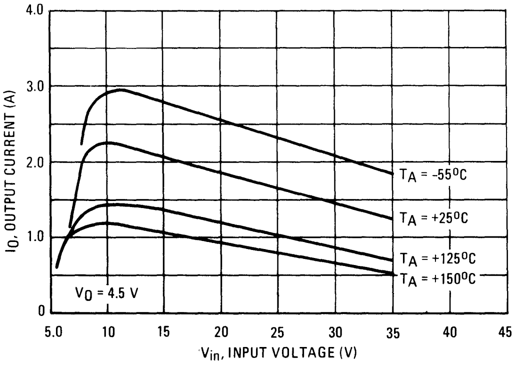
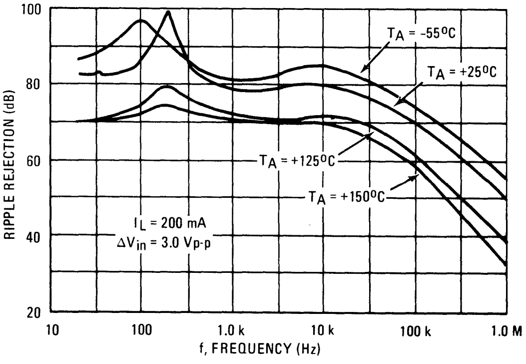


FIGURE 8 – RIPPLE REJECTION



TYPICAL CHARACTERISTICS (continued)

FIGURE 9 – DROPOUT VOLTAGE

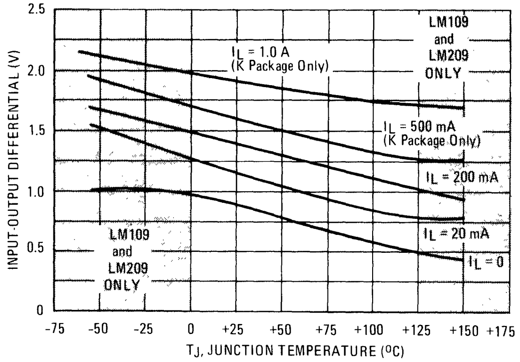


FIGURE 10 – DROPOUT CHARACTERISTIC (K PACKAGE)

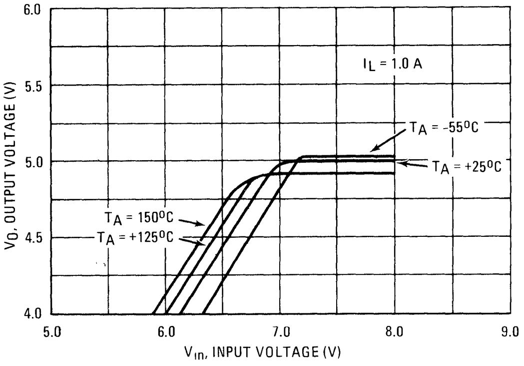


FIGURE 11 – OUTPUT VOLTAGE

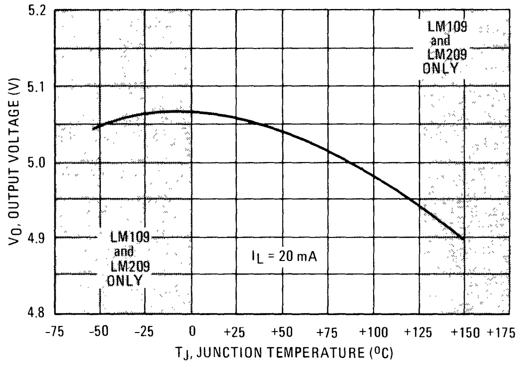


FIGURE 12 – OUTPUT NOISE VOLTAGE

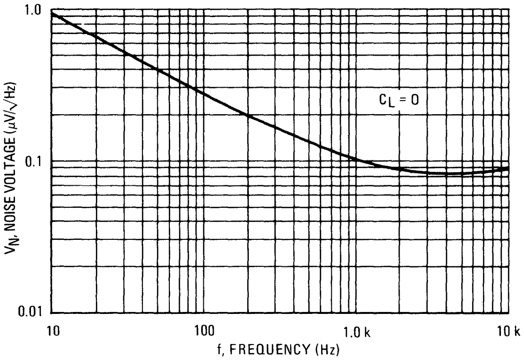


FIGURE 13 – QUIESCENT CURRENT

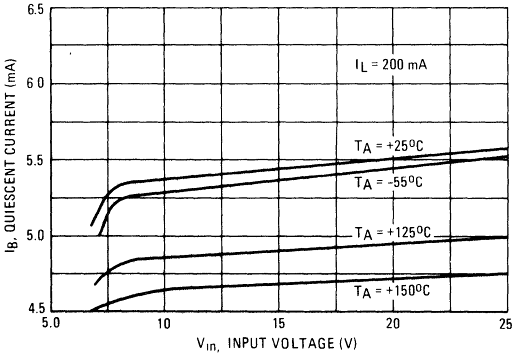
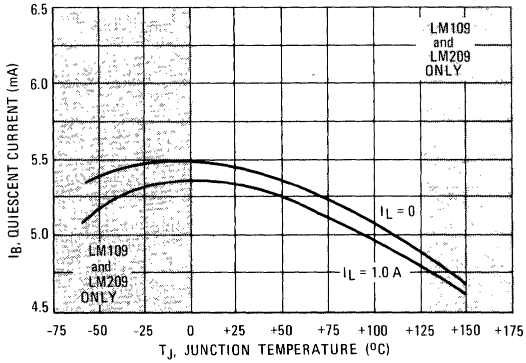


FIGURE 14 – QUIESCENT CURRENT



TYPICAL APPLICATIONS

FIGURE 15 – ADJUSTABLE OUTPUT REGULATOR

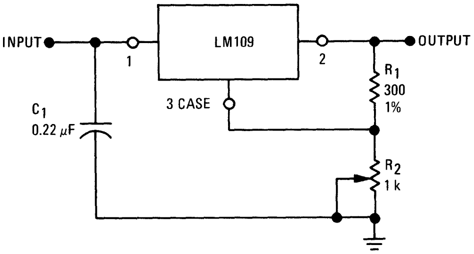
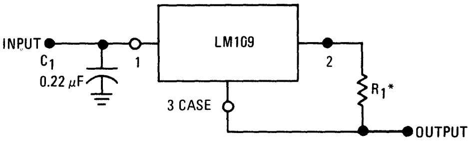


FIGURE 16 – CURRENT REGULATOR



*DETERMINES OUTPUT CURRENT.

FIGURE 17 – 5.0-VOLT, 3.0-AMPERE REGULATOR
(with plastic boost transistor)

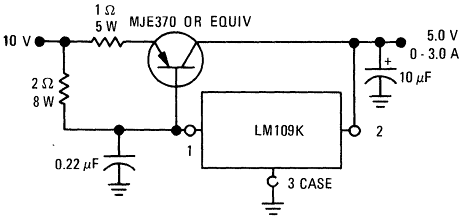


FIGURE 18 – 5.0 VOLT, 4.0-AMPERE TRANSISTOR
(with plastic Darlington boost transistor)

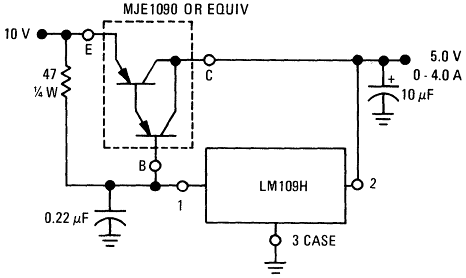


FIGURE 19 – 5.0-VOLT, 10-AMPERE REGULATOR

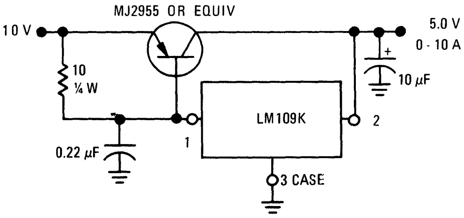
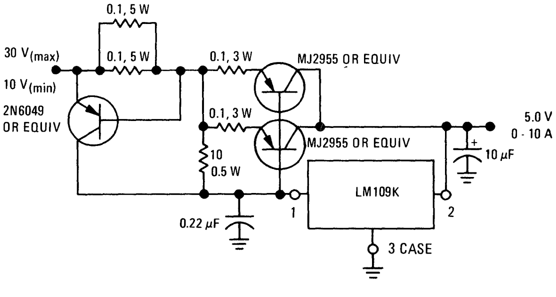


FIGURE 20 – 5.0-VOLT, 10-AMPERE REGULATOR
(with Short-Circuit Current Limiting for Safe-Area Protection of pass transistors)



MOTOROLA

LM117
LM217
LM317

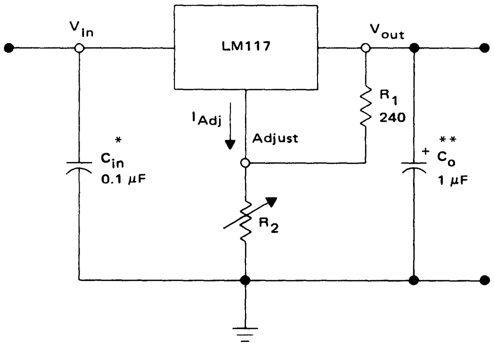
3-TERMINAL ADJUSTABLE OUTPUT POSITIVE VOLTAGE REGULATOR

The LM117/217/317 are adjustable 3-terminal positive voltage regulators capable of supplying in excess of 1.5 A over an output voltage range of 1.2 V to 37 V. These voltage regulators are exceptionally easy to use and require only two external resistors to set the output voltage. Further, they employ internal current limiting, thermal shutdown and safe area compensation, making them essentially blow-out proof.

The LM117 series serve a wide variety of applications including local, on card regulation. This device also makes an especially simple adjustable switching regulator, a programmable output regulator, or by connecting a fixed resistor between the adjustment and output, the LM117 series can be used as a precision current regulator.

- Output Current in Excess of 1.5 Ampere in TO-3 and TO-220 Packages
- Output Current in Excess of 0.5 Ampere in TO-39 Package
- Output Adjustable between 1.2 V and 37 V
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting Constant with Temperature
- Output Transistor Safe-area Compensation
- Floating Operation for High Voltage Applications
- Standard 3-lead Transistor Packages
- Eliminates Stocking Many Fixed Voltages

STANDARD APPLICATION



* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_o is not needed for stability, however it does improve transient response.

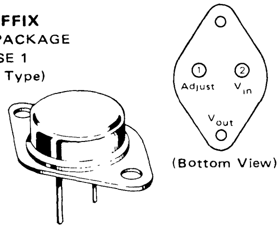
$$V_{out} = 1.25\ V\ (1 + \frac{R_2}{R_1}) + I_{Adj}\ R_2$$

Since I_{Adj} is controlled to less than 100 μA, the error associated with this term is negligible in most applications

3-TERMINAL ADJUSTABLE POSITIVE VOLTAGE REGULATOR

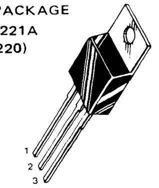
SILICON MONOLITHIC INTEGRATED CIRCUIT

K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 Type)



Pins 1 and 2 electrically isolated from case. Case is third electrical connection.

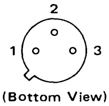
T SUFFIX
PLASTIC PACKAGE
CASE 221A
(TO-220)



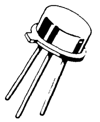
Pin 1 Adjust
Pin 2 V_{out}
Pin 3 V_{in}

Heatsink surface connected to Pin 2

H SUFFIX
METAL PACKAGE
CASE 79
(TO-39)



(Case is output)



Pin 1 V_{in}
Pin 2 Adjust
Pin 3 V_{out}

ORDERING INFORMATION

Device	Temperature Range	Package
LM117H	T_J = -55°C to +150°C	Metal Can
LM117K	T_J = -55°C to +150°C	Metal Power
LM217H	T_J = -25°C to +150°C	Metal Can
LM217K	T_J = -25°C to +150°C	Metal Power
LM317H	T_J = 0°C to +125°C	Metal Can
LM317K	T_J = 0°C to +125°C	Metal Power
LM317T	T_J = 0°C to +125°C	Plastic Power

LM117, LM217, LM317

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Input-Output Voltage Differential	V_I-V_O	40	Vdc
Power Dissipation	P_D	Internally Limited	
Operating Junction Temperature Range	T_J		°C
LM117		-55 to +150	
LM217		-25 to +150	
LM317		0 to +125	
Storage Temperature Range	T_{stg}	-65 to +150	°C

ELECTRICAL CHARACTERISTICS ($V_I - V_O$ = 5 V; I_O = 0.5 A for K and T packages; I_O = 0.1 A for H package; $T_J = T_{low}$ to T_{high} [see Note 1] ; I_{max} and P_{max} per Note 2; unless otherwise specified.)

Characteristic	Figure	Symbol	LM117/217			LM317			Unit
			Min	Typ	Max	Min	Typ	Max	
Line Regulation (Note 3)	1	Reg_{line}							%/V
T_A = 25°C, 3 V $\leq V_I - V_O \leq$ 40 V			–	0.01	0.02	–	0.01	0.04	
Load Regulation (Note 3)	2	Reg_{load}							
T_A = 25°C, 10 mA $\leq I_O \leq I_{max}$									
$V_O \leq$ 5 V			–	5	15	–	5	25	mV
$V_O \geq$ 5 V			–	0.1	0.3	–	0.1	0.5	% V_O
Adjustment Pin Current	3	I_{Adj}	–	50	100	–	50	100	µA
Adjustment Pin Current Change	1, 2	ΔI_{Adj}							µA
2.5 V $\leq V_I - V_O \leq$ 40 V									
10 mA $\leq I_L \leq I_{max}$, $P_D \leq P_{max}$			–	0.2	5	–	0.2	5	
Reference Voltage (Note 4)	3	V_{ref}							V
3 V $\leq V_I - V_O \leq$ 40 V									
10 mA $\leq I_O \leq I_{max}$, $P_D \leq P_{max}$			1.20	1.25	1.30	1.20	1.25	1.30	
Line Regulation (Note 3)	1	Reg_{line}							%/V
3 V $\leq V_I - V_O \leq$ 40 V			–	0.02	0.05	–	0.02	0.07	
Load Regulation (Note 3)	2	Reg_{load}							
10 mA $\leq I_O \leq I_{max}$									
$V_O \leq$ 5 V			–	20	50	–	20	70	mV
$V_O \geq$ 5 V			–	0.3	1	–	0.3	1.5	%V_O
Temperature Stability ($T_{low} \leq T_J \leq T_{high}$)	3	T_S	–	0.7	–	–	0.7	–	%V_O
Minimum Load Current to	3	I_{Lmin}							mA
Maintain Regulation ($V_I - V_O$ = 40 V)			–	3.5	5	–	3.5	10	
Maximum Output Current	3	I_{max}							A
$V_I - V_O \leq$ 15 V, $P_D \leq P_{max}$									
K and T Packages			1.5	2.2	–	1.5	2.2	–	
H Package			0.5	0.8	–	0.5	0.8	–	
$V_I - V_O$ = 40 V, $P_D \leq P_{max}$, T_A = 25°C									
K and T Packages			0.25	0.4	–	0.15	0.4	–	
H Package			–	0.07	–	–	0.07	–	
RMS Noise, % of V_O	–	N							%V_O
T_A = 25°C, 10 Hz $\leq$ f $\leq$ 10 KHz			–	0.003	–	–	0.003	–	
Ripple Rejection, V_O = 10 V, f = 120 Hz (Note 5)	4	RR							dB
Without C_{ADJ}			–	65	–	–	65	–	
C_{ADJ} = 10 µF			66	80	–	66	80	–	
Long Term Stability, $T_J = T_{high}$ (Note 6)	3	S							%/1.0k Hrs
T_A = 25°C for Endpoint Measurements			–	0.3	1	–	0.3	1	
Thermal Resistance Junction to Case	–	$R_{\theta JC}$							°C/W
H Package (TO-39)			–	12	15	–	12	15	
K Package (TO-3)			–	2.3	3	–	2.3	3	
T Package (TO-220)			–	–	–	–	5	–	

NOTES:

(1) T_{low} = -55°C for LM117 T_{high} = +150°C for LM117
= -25°C for LM217 = +150°C for LM217
= 0°C for LM317 = +125°C for LM317

(2) I_{max} = 1.5 A for K (TO-3) and T (TO-220) Packages
= 0.5 A for H (TO-39) Package
P_{max} = 20 W for K (TO-3) and T (TO-220) Packages
= 2 W for H (TO-39) Package

(3) Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

(4) Selected devices with tightened tolerance reference voltage available.

(5) C_{ADJ}, when used, is connected between the adjustment pin and ground.

(6) Since Long Term Stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot.

SCHEMATIC DIAGRAM

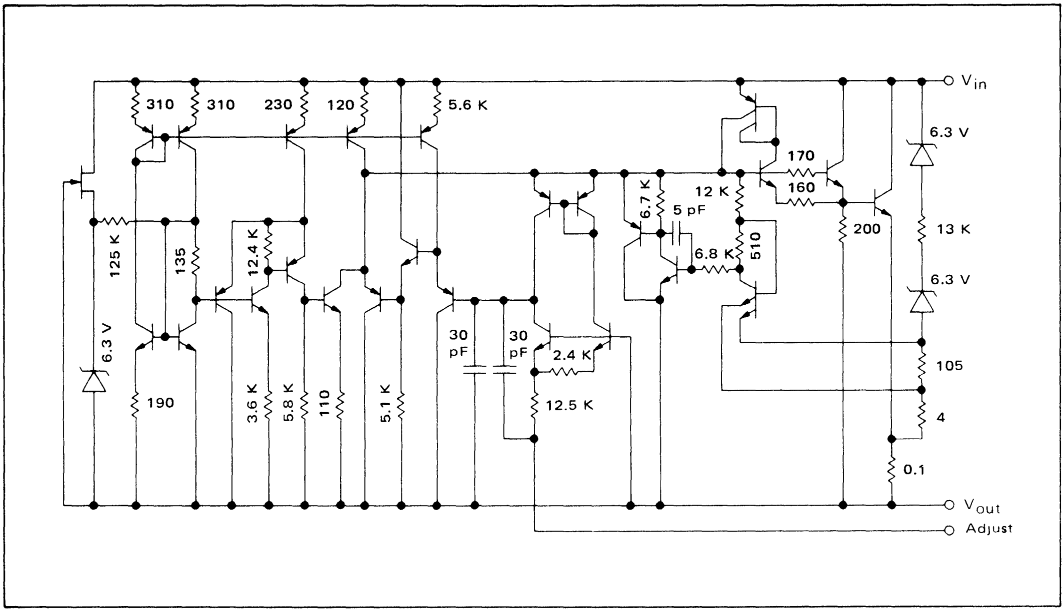
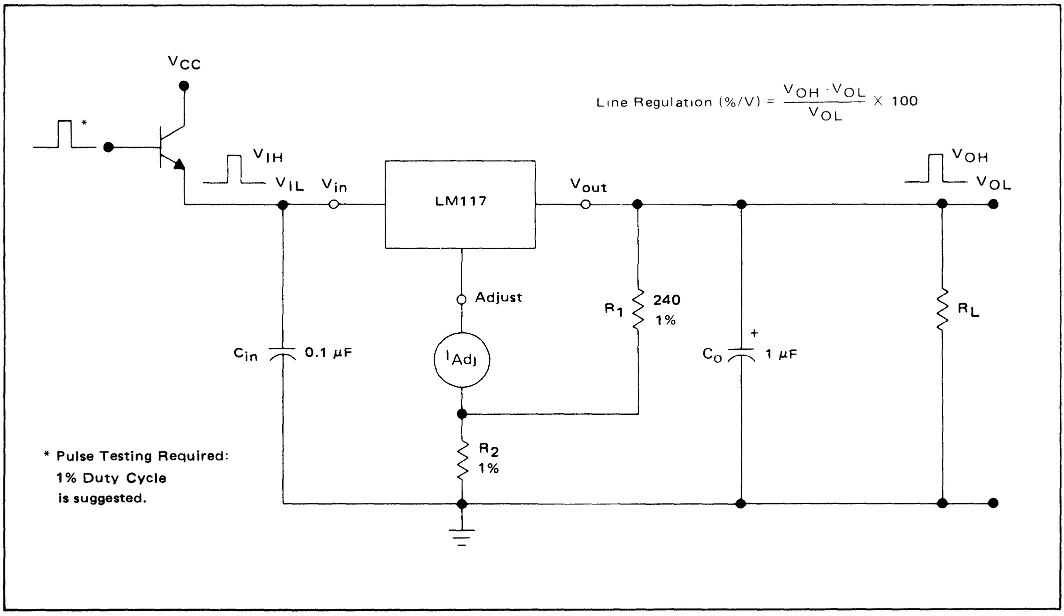
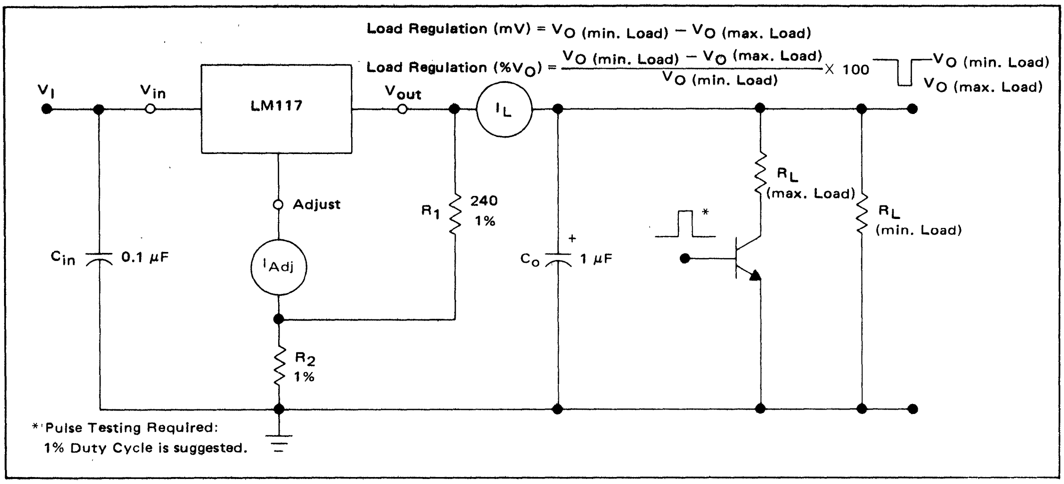


FIGURE 1 – LINE REGULATION AND ΔI_{Adj}/LINE TEST CIRCUIT



* Pulse Testing Required: 1% Duty Cycle is suggested.

FIGURE 2 – LOAD REGULATION AND ΔI_{Adj}/LOAD TEST CIRCUIT

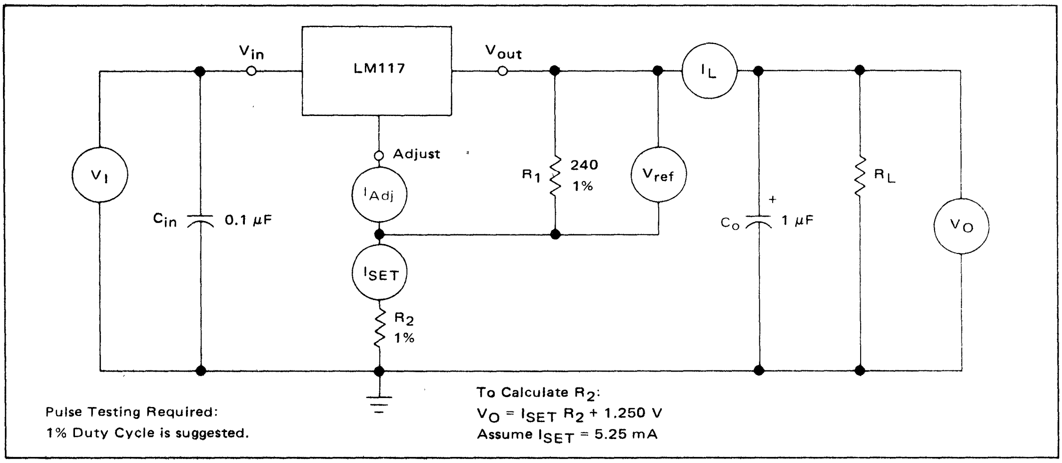


Load Regulation (mV) = V_O (min. Load) − V_O (max. Load)

$$\text{Load Regulation } (\%V_O) = \frac{V_O \text{ (min. Load)} - V_O \text{ (max. Load)}}{V_O \text{ (min. Load)}} \times 100$$

*Pulse Testing Required: 1% Duty Cycle is suggested.

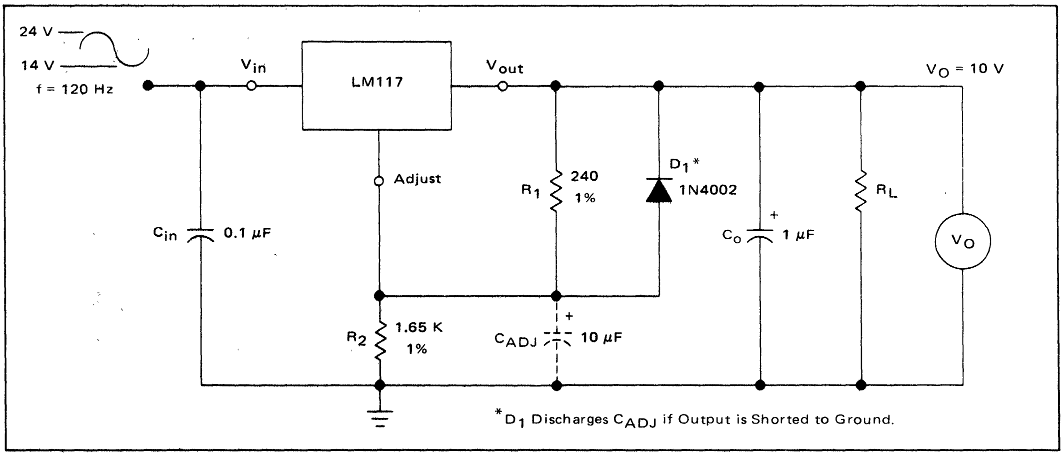
FIGURE 3 – STANDARD TEST CIRCUIT



Pulse Testing Required: 1% Duty Cycle is suggested.

To Calculate R_2:
$V_O = I_{SET} R_2 + 1.250$ V
Assume I_{SET} = 5.25 mA

FIGURE 4 – RIPPLE REJECTION TEST CIRCUIT



*D_1 Discharges C_{ADJ} if Output is Shorted to Ground.

FIGURE 5 – LOAD REGULATION

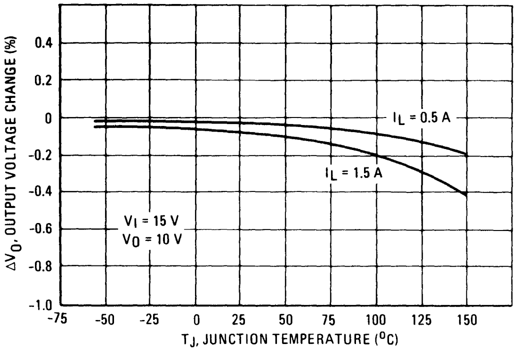


FIGURE 6 – CURRENT LIMIT

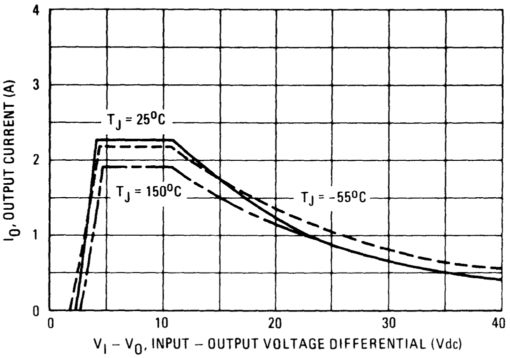


FIGURE 7 – ADJUSTMENT PIN CURRENT

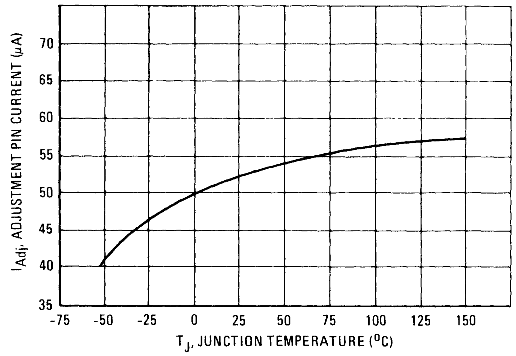


FIGURE 8 – DROPOUT VOLTAGE

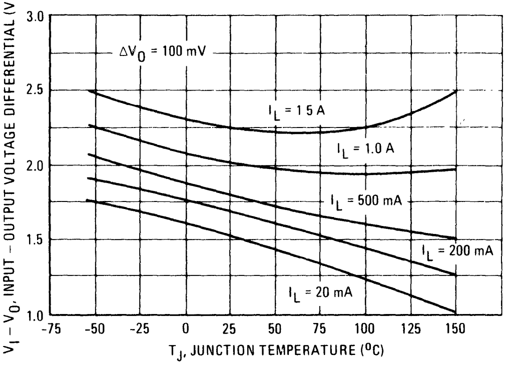


FIGURE 9 – TEMPERATURE STABILITY

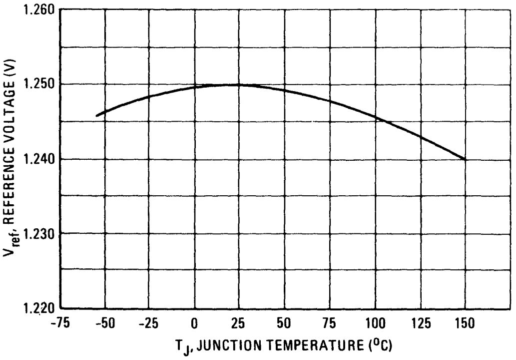


FIGURE 10 – MINIMUM OPERATING CURRENT

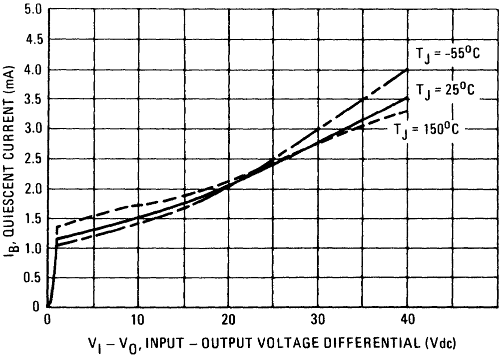


FIGURE 11 – RIPPLE REJECTION VS OUTPUT VOLTAGE

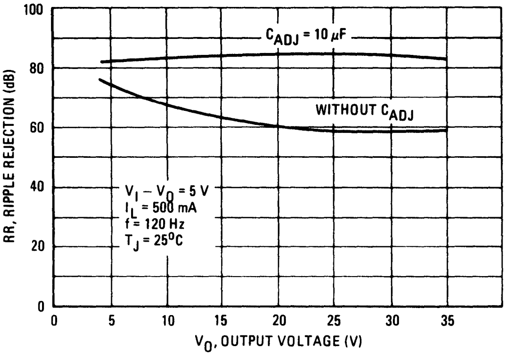


FIGURE 12 – RIPPLE REJECTION VS. OUTPUT CURRENT

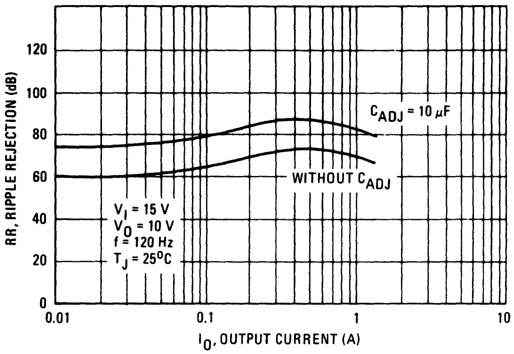


FIGURE 13 – RIPPLE REJECTION VS. FREQUENCY

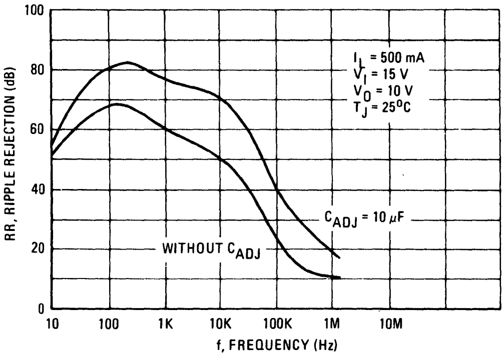


FIGURE 14 – OUTPUT IMPEDANCE

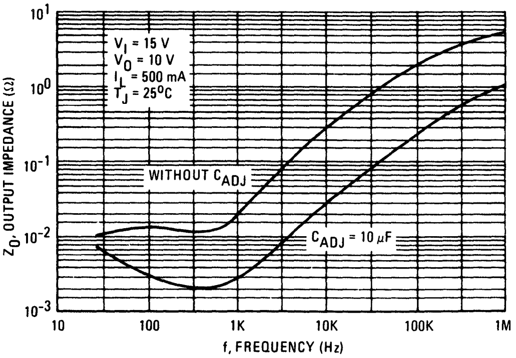


FIGURE 15 – LINE TRANSIENT RESPONSE

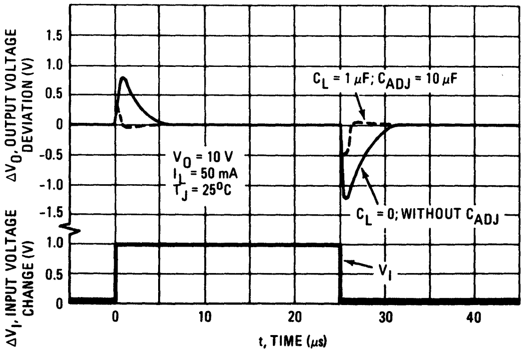
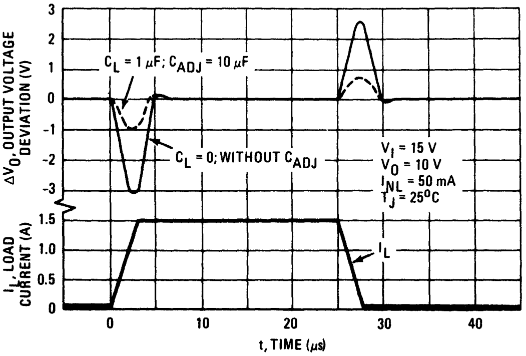


FIGURE 16 – LOAD TRANSIENT RESPONSE



APPLICATIONS INFORMATION

BASIC CIRCUIT OPERATION

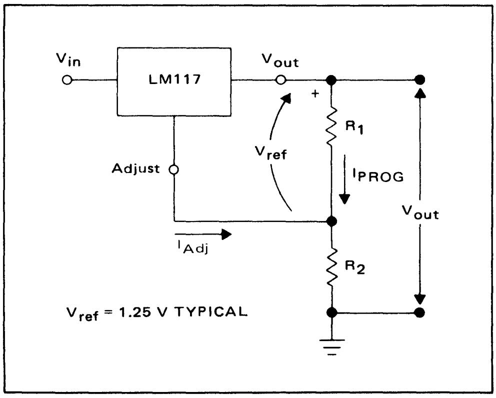
The LM117 is a 3-terminal floating regulator. In operation, the LM117 develops and maintains a nominal 1.25 volt reference (V_{ref}) between its output and adjustment terminals. This reference voltage is converted to a programming current (I_{PROG}) by R1 (see Figure 17), and this constant current flows through R2 to ground. The regulated output voltage is given by:

$$V_{out} = V_{ref} \left(1 + \frac{R2}{R1}\right) + I_{Adj}\, R2$$

Since the current from the adjustment terminal (I_{Adj}) represents an error term in the equation, the LM117 was designed to control I_{Adj} to less than 100 μA and keep it constant. To do this, all quiescent operating current is returned to the output terminal. This imposes the requirement for a minimum load current. If the load current is less than this minimum, the output voltage will rise.

Since the LM117 is a floating regulator, it is only the voltage differential across the circuit which is important to performance, and operation at high voltages with respect to ground is possible.

FIGURE 17 – BASIC CIRCUIT CONFIGURATION



LOAD REGULATION

The LM117 is capable of providing extremely good load regulation, but a few precautions are needed to obtain maximum performance. For best performance, the programming resistor (R1) should be connected as close to the regulator as possible to minimize line drops which effectively appear in series with the reference, thereby degrading regulation. The ground end of R2 can be returned near the load ground to provide remote ground sensing and improve load regulation.

EXTERNAL CAPACITORS

A 0.1 μF disc or 1 μF tantalum input bypass capacitor (C_{in}) is recommended to reduce the sensitivity to input line impedance.

The adjustment terminal may be bypassed to ground to improve ripple rejection. This capacitor (C_{ADJ}) prevents ripple from being amplified as the output voltage is increased. A 10 μF capacitor should improve ripple rejection about 15dB at 120 Hz in a 10 volt application.

Although the LM117 is stable with no output capacitance, like any feedback circuit, certain values of external capacitance can cause excessive ringing. An output capacitance (C_o) in the form of a 1 μF tantalum or 25 μF aluminum electrolytic capacitor on the output swamps this effect and insures stability.

PROTECTION DIODES

When external capacitors are used with any I.C. regulator it is sometimes necessary to add protection diodes to prevent the capacitors from discharging through low current points into the regulator.

Figure 18 shows the LM117 with the recommended protection diodes for output voltages in excess of 25 V or high capacitance values ($C_o > 25$ μF, $C_{ADJ} > 10$ μF). Diode D_1 prevents C_o from discharging thru the I.C. during an input short circuit. Diode D_2 protects against capacitor C_{ADJ} discharging through the I.C. during an output short circuit. The combination of diodes D1 and D2 prevents C_{ADJ} from discharging through the I.C. during an input short circuit.

FIGURE 18 – VOLTAGE REGULATOR WITH PROTECTION DIODES

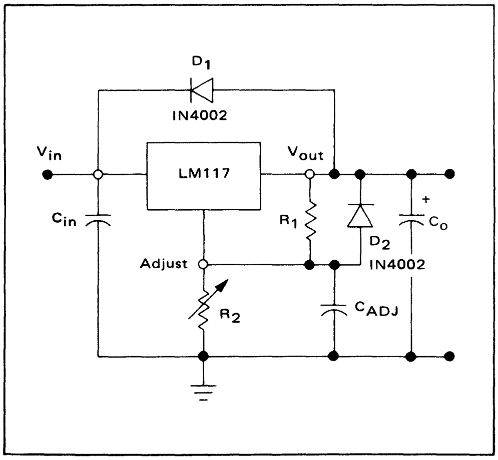
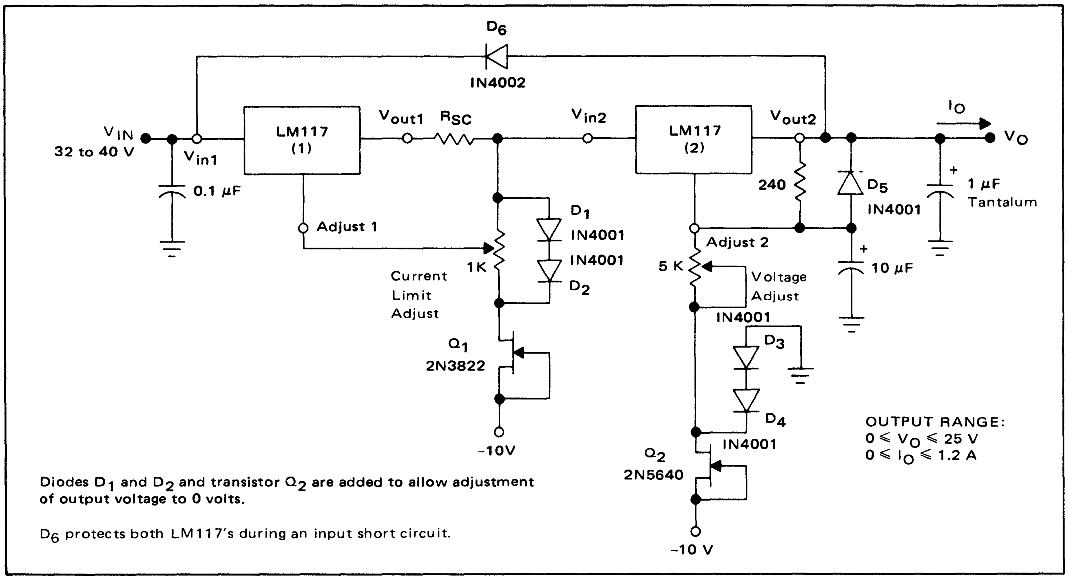


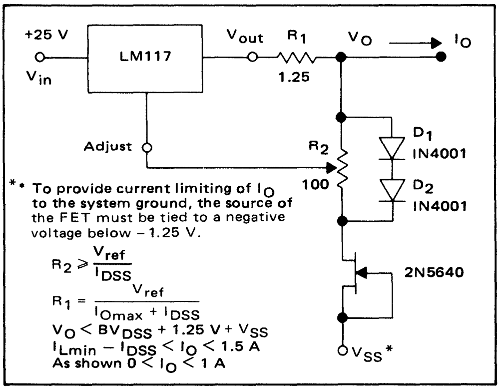
FIGURE 19 – "LABORATORY" POWER SUPPLY WITH ADJUSTABLE CURRENT LIMIT AND OUTPUT VOLTAGE



Diodes D_1 and D_2 and transistor Q_2 are added to allow adjustment of output voltage to 0 volts.

D_6 protects both LM117's during an input short circuit.

FIGURE 20 – ADJUSTABLE CURRENT LIMITER



* To provide current limiting of I_O to the system ground, the source of the FET must be tied to a negative voltage below − 1.25 V.

$$R_2 \geq \frac{V_{ref}}{I_{DSS}}$$

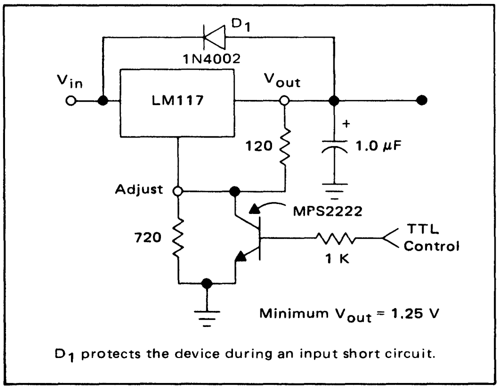
$$R_1 = \frac{V_{ref}}{I_{Omax} + I_{DSS}}$$

$$V_O < BV_{DSS} + 1.25\ V + V_{SS}$$

$$I_{Lmin} - I_{DSS} < I_O < 1.5\ A$$

As shown $0 < I_O < 1$ A

FIGURE 21 – 5 V ELECTRONIC SHUT DOWN REGULATOR



Minimum $V_{out} = 1.25$ V

D_1 protects the device during an input short circuit.

FIGURE 22 – SLOW TURN-ON REGULATOR

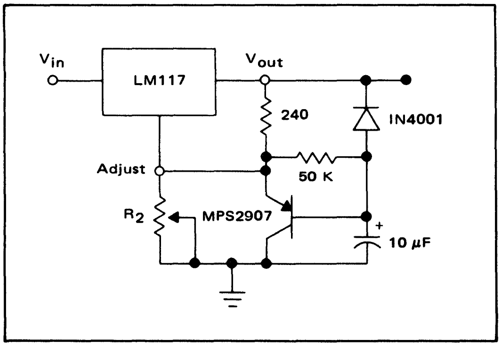
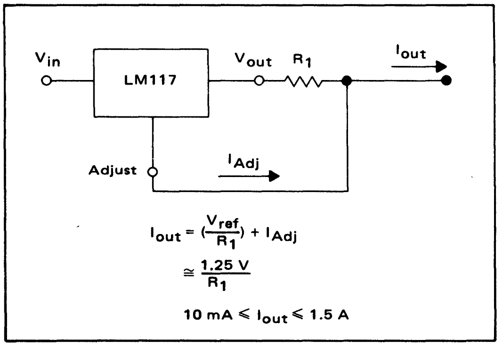


FIGURE 23 – CURRENT REGULATOR



$$I_{out} = \left(\frac{V_{ref}}{R_1}\right) + I_{Adj}$$

$$\cong \frac{1.25\ V}{R_1}$$

$$10\ mA \leq I_{out} \leq 1.5\ A$$

MOTOROLA

LM117L
LM217L
LM317L

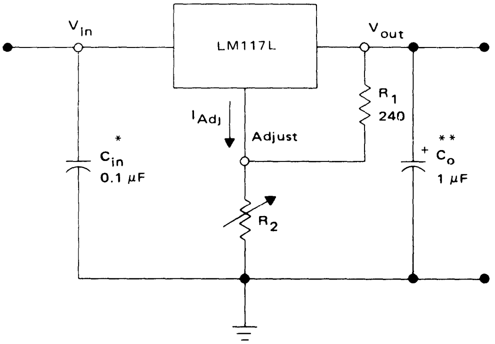
3-TERMINAL ADJUSTABLE OUTPUT POSITIVE VOLTAGE REGULATOR

The LM117L/217L/317L are adjustable 3-terminal positive voltage regulators capable of supplying in excess of 100 mA over an output voltage range of 1.2 V to 37 V. These voltage regulators are exceptionally easy to use and require only two external resistors to set the output voltage. Further, they employ internal current limiting, thermal shutdown and safe area compensation, making them essentially blow-out proof.

The LM117L series serves a wide variety of applications including local, on card regulation. This device also makes an especially simple adjustable switching regulator, a programmable output regulator, or by connecting a fixed resistor between the adjustment and output, the LM117L series can be used as a precision current regulator.

- Output Current in Excess of 100 mA
- Output Adjustable Between 1.2 V and 37 V
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Floating Operation for High Voltage Applications
- Standard 3-Lead Transistor Packages
- Eliminates Stocking Many Fixed Voltages

STANDARD APPLICATION



* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_o is not needed for stability, however it does improve transient response.

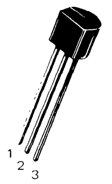
$$V_{out} = 1.25\ V\ (1 + \frac{R_2}{R_1}) + I_{Adj}\ R_2$$

Since I_{Adj} is controlled to less than 100 µA, the error associated with this term is negligible in most applications

LOW-CURRENT 3-TERMINAL ADJUSTABLE POSITIVE VOLTAGE REGULATOR

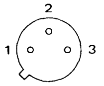
SILICON MONOLITHIC INTEGRATED CIRCUIT

Z SUFFIX
CASE 29
TO 92
PLASTIC PACKAGE
(LM317L only)



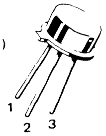
Pin 1 Adjust
Pin 2 V_{out}
Pin 3 V_{in}

H SUFFIX
METAL PACKAGE
CASE 79
(TO-39)



(Bottom View)

(Case is output)



Pin 1 V_{in}
Pin 2 Adjust
Pin 3 V_{out}

ORDERING INFORMATION

Device	Temperature Range	Package
LM117LH	T_J = −55°C to +150°C	Metal Can
LM217LH	T_J = −25°C to +150°C	Metal Can
LM317LH	T_J = 0°C to +125°C	Metal Can
LM317LZ	T_J = 0°C to +125°C	Plastic

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Input-Output Voltage Differential		V_I-V_O	40	Vdc
Power Dissipation		P_D	Internally Limited	
Operating Junction Temperature Range	LM117L	T_J	-55 to +150	°C
	LM217L		-25 to +150	
	LM317L		0 to +125	
Storage Temperature Range		T_{stg}	-65 to +150	°C

ELECTRICAL CHARACTERISTICS

($V_I - V_O = 5$ V, $I_O = 40$ mA; $T_J = T_{low}$ to T_{high} [see Note 1]; I_{max} and P_{max} per Note 2; unless otherwise specified.)

Characteristic	Figure	Symbol	LM117L/217L			LM317L			Unit
			Min	Typ	Max	Min	Typ	Max	
Line Regulation (Note 3)	1	Reg_{line}							%/V
$T_A = 25°C$, $3\ V \leq V_I$-$V_O \leq 40$ V			—	0.01	0 02	—	0.01	0.04	
Load Regulation (Note 3), $T_A = 25°C$	2	Reg_{load}							
5 mA $\leq I_O \leq I_{max}$ — LM117L/217L									
10 mA $\leq I_O \leq I_{max}$ — LM317L									
$V_O \leq 5V$			—	5	15	—	5	25	mV
$V_O \geq 5$ V			—	0.1	0.3	—	0 1	0.5	% V_O
Adjustment Pin Current	3	I_{Adj}	—	50	100	—	50	100	μA
Adjustment Pin Current Change	1,2	ΔI_{Adj}							μA
2.5 V $\leq V_I$-$V_O \leq 40$ V, $P_D \leq P_{max}$									
5 mA $\leq I_O \leq I_{max}$ — LM117L/217L			—	0.2	5	—	0 2	5	
10 mA $\leq I_O \leq I_{max}$ — LM317L									
Reference Voltage (Note 4)	3	V_{ref}							V
3 V $\leq V_I$-$V_O \leq 40$ V, $P_D \leq P_{max}$									
5 mA $\leq I_O \leq I_{max}$ — LM117L/217L			1.20	1.25	1.30	1 20	1 25	1 30	
10 mA $\leq I_O \leq I_{max}$ — LM317L									
Line Regulation (Note 3)	1	Reg_{line}							%/V
3 V $\leq V_I$-$V_O \leq 40$ V			—	0.02	0.05	—	0 02	0 07	
Load Regulation (Note 3)	2	Reg_{load}							
5 mA $\leq I_O \leq I_{max}$ — LM117L/217L									
10 mA $\leq I_O \leq I_{max}$ — LM317L									
$V_O \leq 5$ V			—	20	50	—	20	70	mV
$V_O \geq 5$ V			—	0.3	1	—	0 3	1 5	%V_O
Temperature Stability ($T_{low} \leq T_J \leq T_{high}$)	3	T_S	—	0.7	—	—	0.7	—	%V_O
Minimum Load Current to	3	I_{Lmin}							mA
Maintain Regulation (V_I-V_O = 40 V)			—	3.5	5	—	3.5	10	
Maximum Output Current	3	I_{max}							A
V_I-$V_O \leq 20$ V, $P_D \leq P_{max}$ H Package			100	200	—	100	200	—	
V_I-$V_O \leq 6.25$ V, $P_D \leq P_{max}$, Z Package			100	200	—	1005	200	—	
V_I-V_O = 40 V, $P_D \leq P_{max}$, $T_A = 25°C$									
H Package			—	50	—	—	50	—	
Z Package			—	20	—	—	20	—	
RMS Noise, % of V_O	—	N							%V_O
$T_A = 25°C$, 10 Hz $\leq f \leq$ 10 kHz			—	0 003	—	—	0.003	—	
Ripple Rejection (Note 5)	4	RR							dB
V_O = 1.25 V, f = 120 Hz			66	80	—	60	80	—	
C_{ADJ} = 10 μF V_O = 10.0 V			—	80	—	—	80	—	
Long Term Stability, $T_J = T_{high}$ (Note 6)	3	S							%/1.0 k
$T_A = 25°C$ for Endpoint Measurements			—	0.3	1	—	0.3	1	Hrs.
Thermal Resistance Junction to Case	—	$R_{\theta JC}$							°C/W
H Package (TO-39)			—	40	—	—	40	—	
Z Package (TO-92)			—	—	—	—	160	—	

NOTES:

(1) T_{low} = -55°C for LM117L　　T_{high} = +150°C for LM117L
-25°C for LM217L　　= +150°C for LM217L
0°C for LM317L　　= +125°C for LM317L

(2) I_{max} = 100 mA
P_{max} = 2 W for H (TO-39) Package
= 625 mW for Z (TO-92) Package

(3) Load and line regulation are specified at constant junction temperature Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

(4) Selected devices with tightened tolerance reference voltage available

(5) C_{ADJ}, when used, is connected between the adjustment pin and ground.

(6) Since Long Term Stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot.

SCHEMATIC DIAGRAM

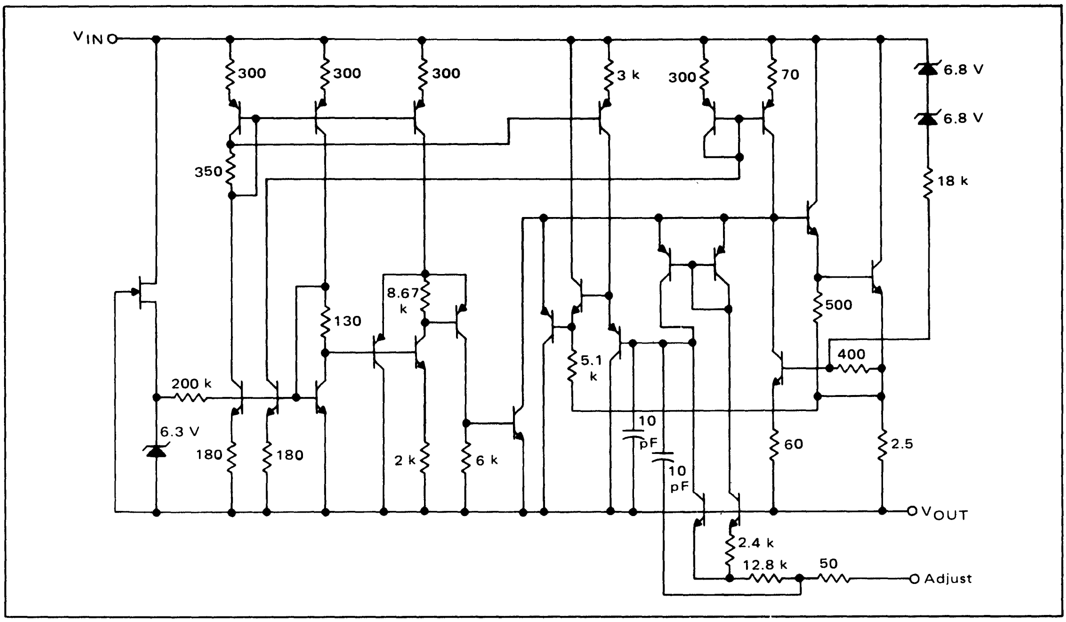


FIGURE 1 – LINE REGULATION AND ΔI_{Adj}/LINE TEST CIRCUIT

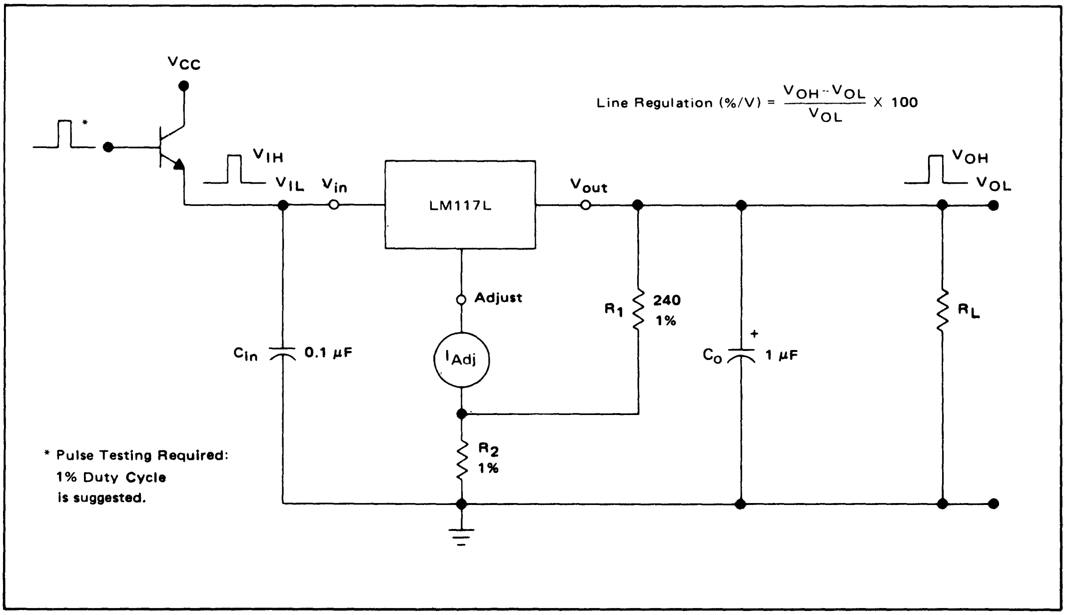
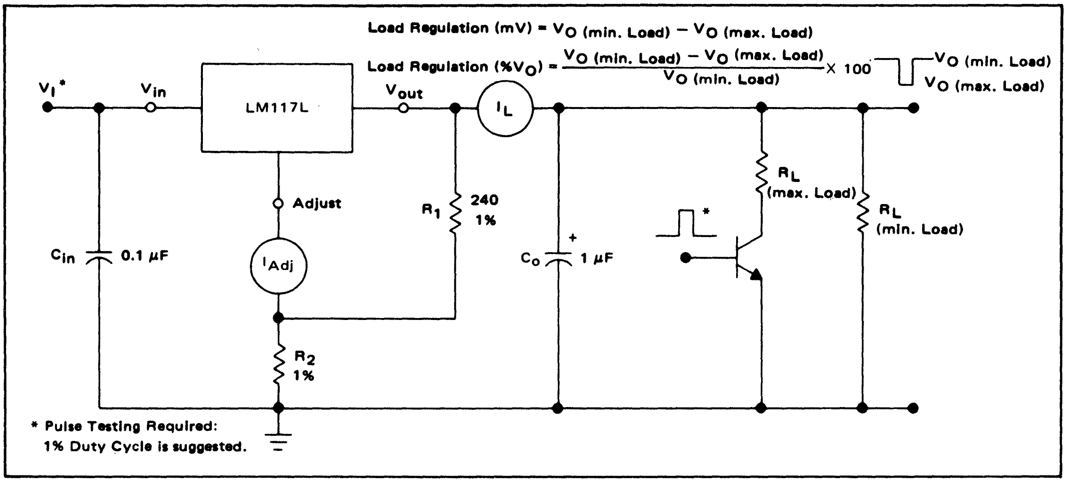


FIGURE 2 – LOAD REGULATION AND ΔI_{Adj}/LOAD TEST CIRCUIT

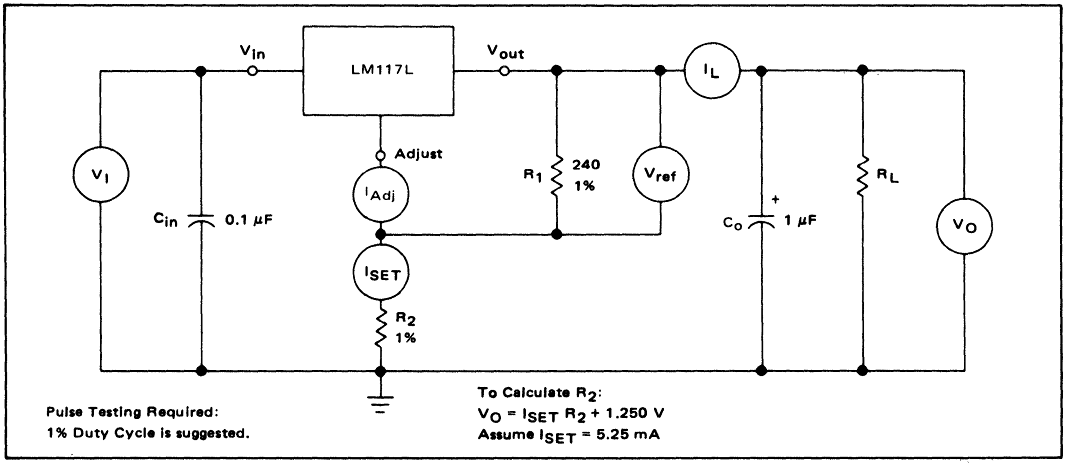


Load Regulation (mV) = V_O (min. Load) − V_O (max. Load)

Load Regulation (%V_O) = $\frac{V_O \text{ (min. Load)} - V_O \text{ (max. Load)}}{V_O \text{ (min. Load)}}$ X 100

* Pulse Testing Required:
1% Duty Cycle is suggested.

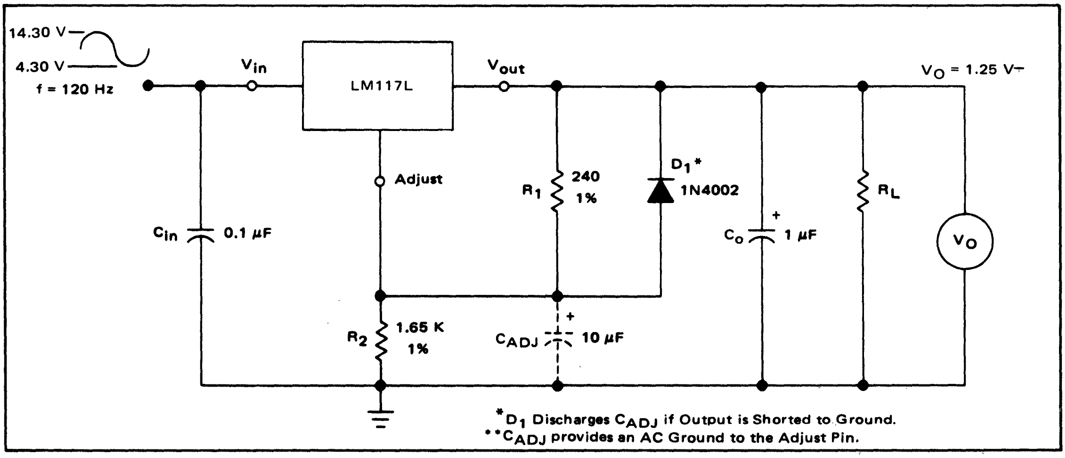
FIGURE 3 – STANDARD TEST CIRCUIT



Pulse Testing Required:
1% Duty Cycle is suggested.

To Calculate R_2:
$V_O = I_{SET} R_2 + 1.250$ V
Assume I_{SET} = 5.25 mA

FIGURE 4 – RIPPLE REJECTION TEST CIRCUIT



*D_1 Discharges C_{ADJ} if Output is Shorted to Ground.
**C_{ADJ} provides an AC Ground to the Adjust Pin.

FIGURE 5 – LOAD REGULATION

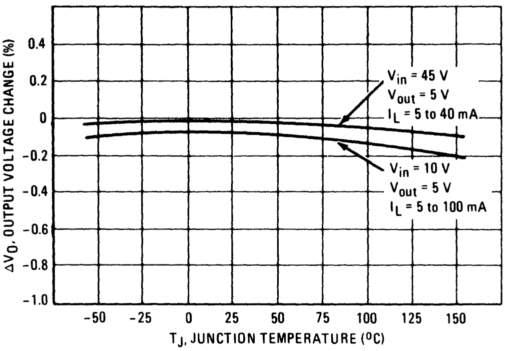


FIGURE 6 – RIPPLE REJECTION

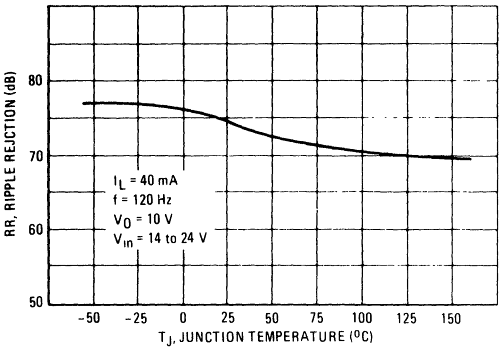


FIGURE 7 – CURRENT LIMIT

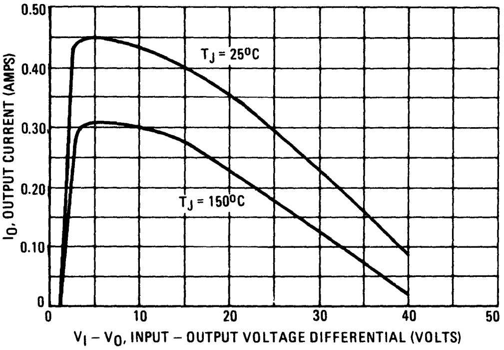


FIGURE 8 – DROPOUT VOLTAGE

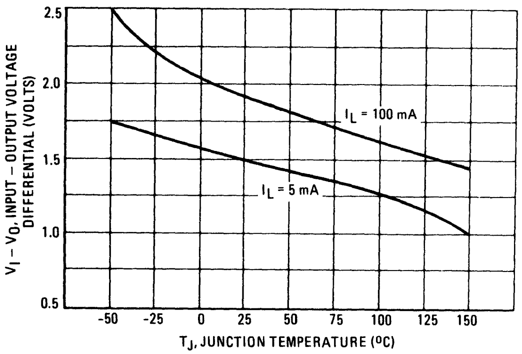


FIGURE 9 – MINIMUM OPERATING CURRENT

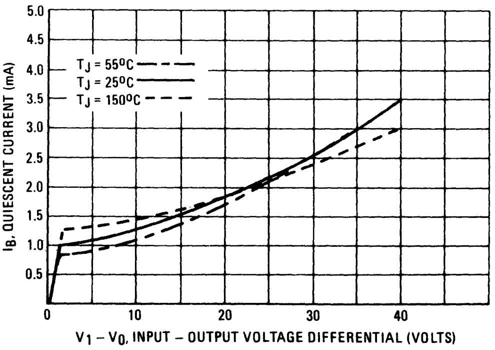


FIGURE 10 – RIPPLE REJECTION versus FREQUENCY

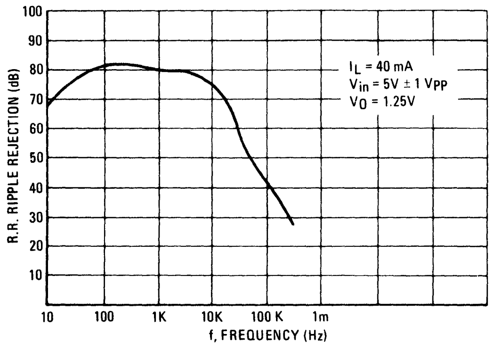


FIGURE 11 — TEMPERATURE STABILITY

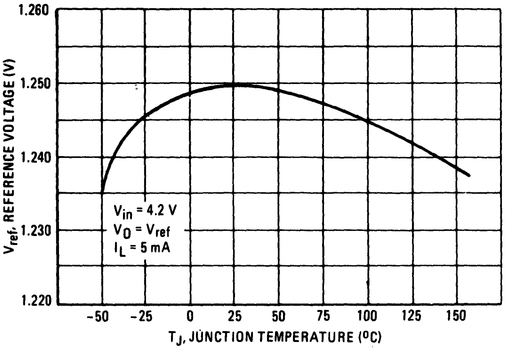


FIGURE 12 — ADJUSTMENT PIN CURRENT

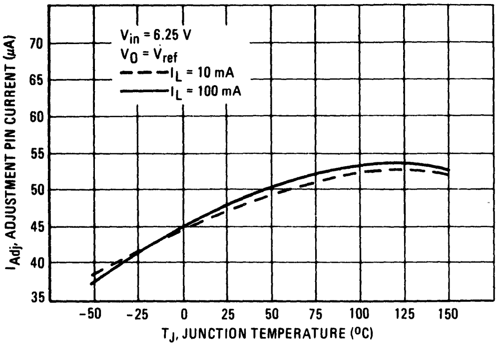


FIGURE 13 — LINE REGULATION

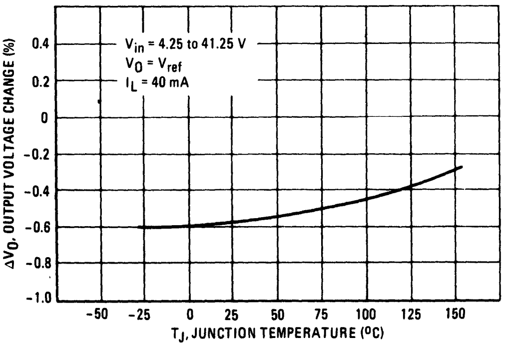


FIGURE 14 — OUTPUT NOISE

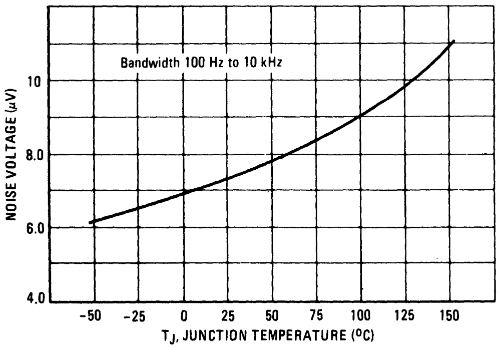


FIGURE 15 — LINE TRANSIENT RESPONSE

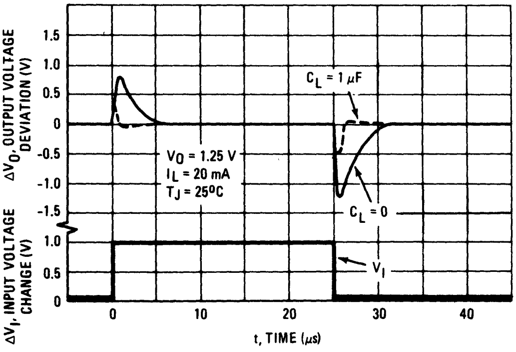
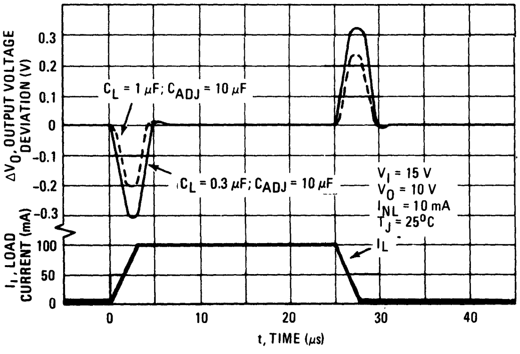


FIGURE 16 — LOAD TRANSIENT RESPONSE



APPLICATIONS INFORMATION

BASIC CIRCUIT OPERATION

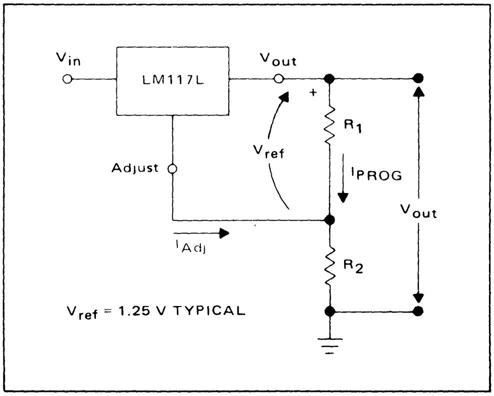
The LM117L is a 3-terminal floating regulator. In operation, the LM117L develops and maintains a nominal 1.25 volt reference (V_{ref}) between its output and adjustment terminals. This reference voltage is converted to a programming current (I_{PROG}) by R1 (see Figure 13), and this constant current flows through R2 to ground. The regulated output voltage is given by:

$$V_{out} = V_{ref} (1 + \frac{R2}{R1}) + I_{Adj}\ R2$$

Since the current from the adjustment terminal (I_{Adj}) represents an error term in the equation, the LM117L was designed to control I_{Adj} to less than 100 µA and keep it constant. To do this, all quiescent operating current is returned to the output terminal. This imposes the requirement for a minimum load current. If the load current is less than this minimum, the output voltage will rise.

Since the LM117L is a floating regulator, it is only the voltage differential across the circuit which is important to performance, and operation at high voltages with respect to ground is possible.

FIGURE 17 – BASIC CIRCUIT CONFIGURATION



LOAD REGULATION

The LM117L is capable of providing extremely good load regulation, but a few precautions are needed to obtain maximum performance. For best performance, the programming resistor (R1) should be connected as close to the regulator as possible to minimize line drops which effectively appear in series with the reference, thereby degrading regulation. The ground end of R2 can be returned near the load ground to provide remote ground sensing and improve load regulation.

EXTERNAL CAPACITORS

A 0.1 µF disc or 1 µF tantalum input bypass capacitor (C_{in}) is recommended to reduce the sensitivity to input line impedance.

The adjustment terminal may be bypassed to ground to improve ripple rejection. This capacitor (C_{ADJ}) prevents ripple from being amplified as the output voltage is increased. A 10 µF capacitor should improve ripple rejection about 15dB at 120 Hz in a 10 volt application.

Although the LM117L is stable with no output capacitance, like any feedback circuit, certain values of external capacitance can cause excessive ringing. An output capacitance (C_O) in the form of a 1 µF tantalum or 25 µF aluminum electrolytic capacitor on the output swamps this effect and insures stability.

PROTECTION DIODES

When external capacitors are used with any I.C. regulator it is sometimes necessary to add protection diodes to prevent the capacitors from discharging through low current points into the regulator.

Figure 14 shows the LM117L with the recommended protection diodes for output voltages in excess of 25 V or high capacitance values ($C_O > 10$ µF, $C_{ADJ} > 5$ µF). Diode D_1 prevents C_O from discharging thru the I.C. during an input short circuit. Diode D_2 protects against capacitor C_{ADJ} discharging through the I.C. during an output short circuit. The combination of diodes D1 and D2 prevents C_{ADJ} from discharging through the I.C. during an input short circuit.

FIGURE 18 – VOLTAGE REGULATOR WITH PROTECTION DIODES

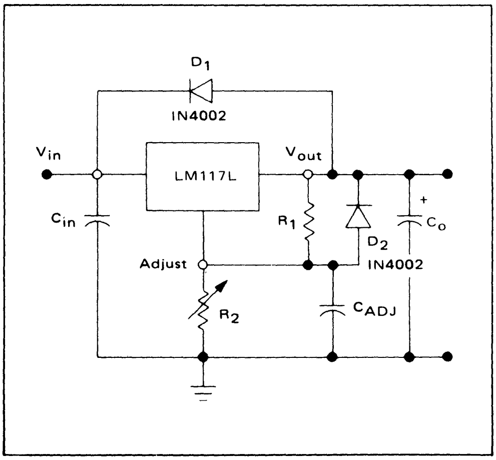
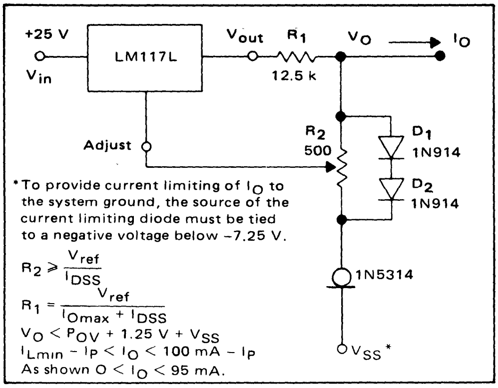


FIGURE 19 — ADJUSTABLE CURRENT LIMITER



*To provide current limiting of I_O to the system ground, the source of the current limiting diode must be tied to a negative voltage below −7.25 V.

$R_2 \geqslant \frac{V_{ref}}{I_{DSS}}$

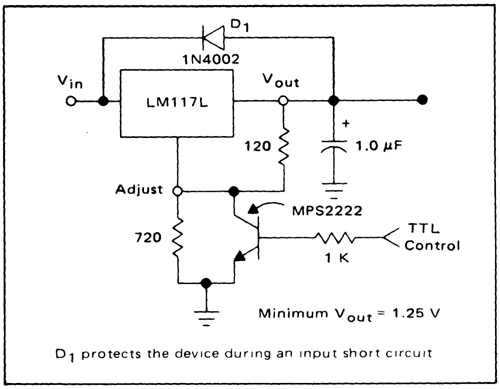
$R_1 = \frac{V_{ref}}{I_{Omax} + I_{DSS}}$

$V_O < P_{OV} + 1.25\ V + V_{SS}$

$I_{Lmin} - I_P < I_O < 100\ mA - I_P$

As shown $0 < I_O < 95$ mA.

FIGURE 20 — 5 V ELECTRONIC SHUTDOWN REGULATOR



Minimum $V_{out} = 1.25$ V

D_1 protects the device during an input short circuit

FIGURE 21 — SLOW TURN-ON REGULATOR

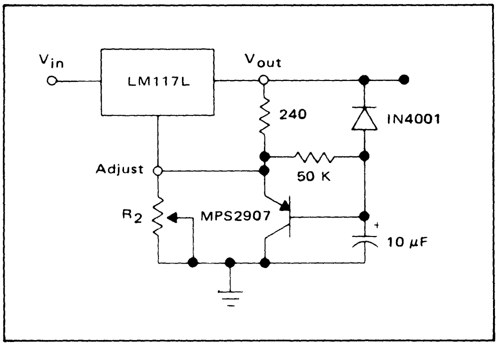
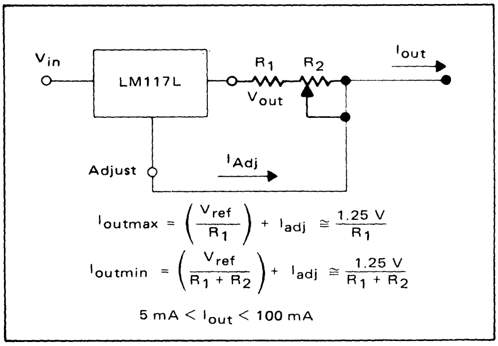


FIGURE 22 — CURRENT REGULATOR



$$I_{outmax} = \left(\frac{V_{ref}}{R_1}\right) + I_{adj} \cong \frac{1.25\ V}{R_1}$$

$$I_{outmin} = \left(\frac{V_{ref}}{R_1 + R_2}\right) + I_{adj} \cong \frac{1.25\ V}{R_1 + R_2}$$

$5\ mA < I_{out} < 100\ mA$

MOTOROLA

LM117M
LM217M
LM317M

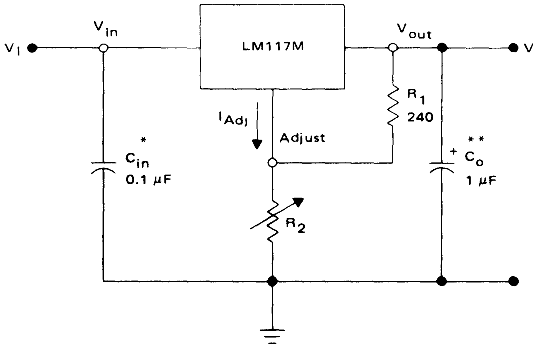
3-TERMINAL ADJUSTABLE OUTPUT POSITIVE VOLTAGE REGULATOR

The LM117M/217M/317M are adjustable 3-terminal positive voltage regulators capable of supplying in excess of 500 mA over an output voltage range of 1.2 V to 37 V. These voltage regulators are exceptionally easy to use and require only two external resistors to set the output voltage. Further, they employ internal current limiting, thermal shutdown and safe area compensation, making them essentially blow-out proof.

The LM117M series serve a wide variety of applications including local, on card regulation. This device also makes an especially simple adjustable switching regulator, a programmable output regulator, or by connecting a fixed resistor between the adjustment and output, the LM117M series can be used as a precision current regulator.

- Output Current in Excess of 500 mA
- Output Adjustable Between 1.2 V and 37 V
- Internal Thermal Overload Protection
- Internal Short-Circuit-Current Limiting
- Output Transistor Safe-Area Compensation
- Floating Operation for High Voltage Applications
- Standard 3-Lead Transistor Packages
- Eliminates Stocking Many Fixed Voltages

STANDARD APPLICATION



* C_{in} is required if regulator is located an appreciable distance from power supply filter.

** C_o is not needed for stability, however it does improve transient response.

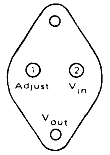
$$V_O = 1.25\ V\left(1 + \frac{R_2}{R_1}\right) + I_{adj}\,R_2$$

Since I_{adj} is controlled to less than 100 µA, the error associated with this term is negligible in most applications

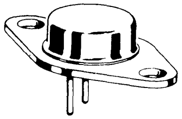
MEDIUM-CURRENT 3-TERMINAL ADJUSTABLE POSITIVE VOLTAGE REGULATOR

SILICON MONOLITHIC INTEGRATED CIRCUIT

R SUFFIX
METAL PACKAGE
CASE 80-02
(TO-66 Type)

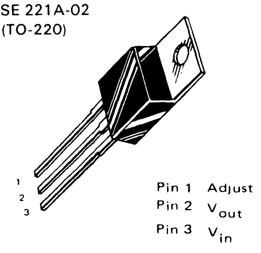


(Bottom View)



Pins 1 and 2 electrically isolated from case. Case is third electrical connection.

T SUFFIX
PLASTIC PACKAGE
CASE 221A-02
(TO-220)



Pin 1 Adjust
Pin 2 V_{out}
Pin 3 V_{in}

Heatsink surface connected to Pin 2

ORDERING INFORMATION

Device	Temperature Range	Package
LM117MR	T_J = -55°C to +150°C	Metal Power
LM217MR	T_J = -25°C to +150°C	Metal Power
LM317MR	T_J = 0°C to +125°C	Metal Power
LM317MT	T_J = 0°C to +125°C	Plastic Power

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Input-Output Voltage Differential		V_I-V_O	40	Vdc
Power Dissipation		P_D	Internally Limited	
Operating Junction Temperature Range	LM117M LM217M LM317M	T_J	-55 to +150 -25 to +150 0 to +125	°C
Storage Temperature Range		T_{stg}	-65 to +150	°C

ELECTRICAL CHARACTERISTICS

(V_I - V_O = 5 V, I_O = 0.1 A, T_J = T_{low} to T_{high} [see Note 1], P_{max} per Note 2, unless otherwise specified)

Characteristic	Figure	Symbol	LM117M/217M			LM317M			Unit
			Min	Typ	Max	Min	Typ	Max	
Line Regulation (Note 3) T_A = 25°C, 3 V ≤ V_I-V_O ≤ 40 V	1	Reg_{line}	—	0.01	0.02	—	0.01	0.04	%/V
Load Regulation (Note 3), T_A = 25°C, 10 mA ≤ I_O ≤ 0.5 A	2	Reg_{load}							
V_O ≤ 5 V			—	5	15	—	5	25	mV
V_O ≥ 5 V			—	0.1	0.3	—	0.1	0.5	% V_O
Adjustment Pin Current	3	I_{adj}	—	50	100	—	50	100	μA
Adjustment Pin Current Change 2.5 V ≤ V_I-V_O ≤ 40 V, 10 mA ≤ I_L ≤ 0.5 A, P_D ≤ P_{max}	1,2	ΔI_{adj}	—	0.2	5	—	0.2	5	μA
Reference Voltage (Note 4) 3 V ≤ V_I-V_O ≤ 40 V 10 mA ≤ I_O ≤ 0.5 A, P_D ≤ P_{max}	3	V_{ref}	1.20	1.25	1.30	1.20	1.25	1.30	V
Line Regulation (Note 3) 3 V ≤ V_I-V_O ≤ 40 V	1	Reg_{line}	—	0.02	0.05	—	0.02	0.07	%/V
Load Regulation (Note 3) 10 mA ≤ I_O ≤ 0.5 A	2	Reg_{load}							
V_O ≤ 5 V			—	20	50	—	20	70	mV
V_O ≥ 5 V			—	0.3	1	—	0.3	1.5	%V_O
Temperature Stability (T_{low} ≤ T_J ≤ T_{high})	3	T_S	—	0.7	—	—	0.7	—	%V_O
Minimum Load Current to Maintain Regulation (V_I-V_O = 40 V)	3	I_{Lmin}	—	3.5	5	—	3.5	10	mA
Maximum Output Current	3	I_{max}							A
V_I-V_O ≤ 15 V, P_D ≤ P_{max}			0.5	0.9	—	0.5	0.9	—	
V_I-V_O = 40 V, P_D ≤ P_{max}, T_A = 25°C			0.15	0.25	—	0.15	0.25	—	
RMS Noise, % of V_O T_A = 25°C, 10 Hz ≤ f ≤ 10 kHz	—	N	—	0.003	—	—	0.003	—	%V_O
Ripple Rejection, V_O = 10 V, f = 120 Hz (Note 5)	4	RR							dB
Without C_{adj}			—	65	—	—	65	—	
C_{adj} = 10 μF			66	80	—	66	80	—	
Long Term Stability, T_J = T_{high} (Note 6) T_A = 25°C for Endpoint Measurements	3	S	—	0.3	1	—	0.3	1	%/1.0 k Hrs.
Thermal Resistance Junction to Case	—	$R_{\theta JC}$							°C/W
R Package (TO-66)			—	7	7	—	7	—	
T Package (TO-220)			—	—	—	—	7	—	

NOTES

(1) T_{low} = -55°C for LM117M
  = -25°C for LM217M
  = 0°C for LM317M

T_{high} = +150°C for LM117M
  = +150°C for LM217M
  = +125°C for LM317M

(2) P_{max} = 7.5 W

(3) Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

(4) Selected devices with tightened tolerance reference voltage available.

(5) C_{adj}, when used, is connected between the adjustment pin and ground

(6) Since Long Term Stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot.

SCHEMATIC DIAGRAM

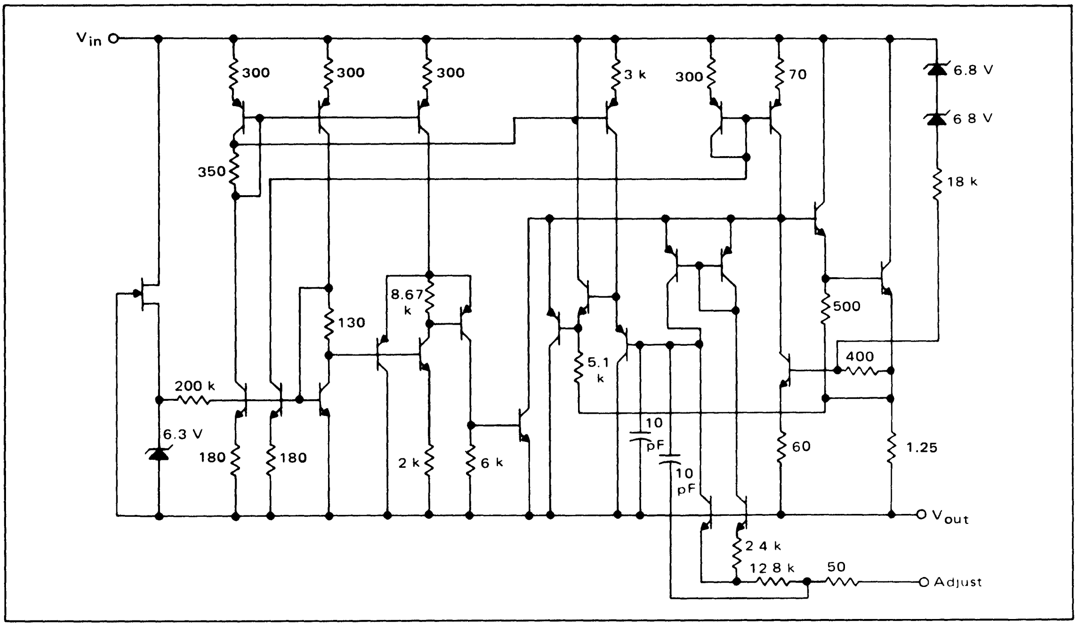


FIGURE 1 – LINE REGULATION AND ΔI_{Adj}/LINE TEST CIRCUIT

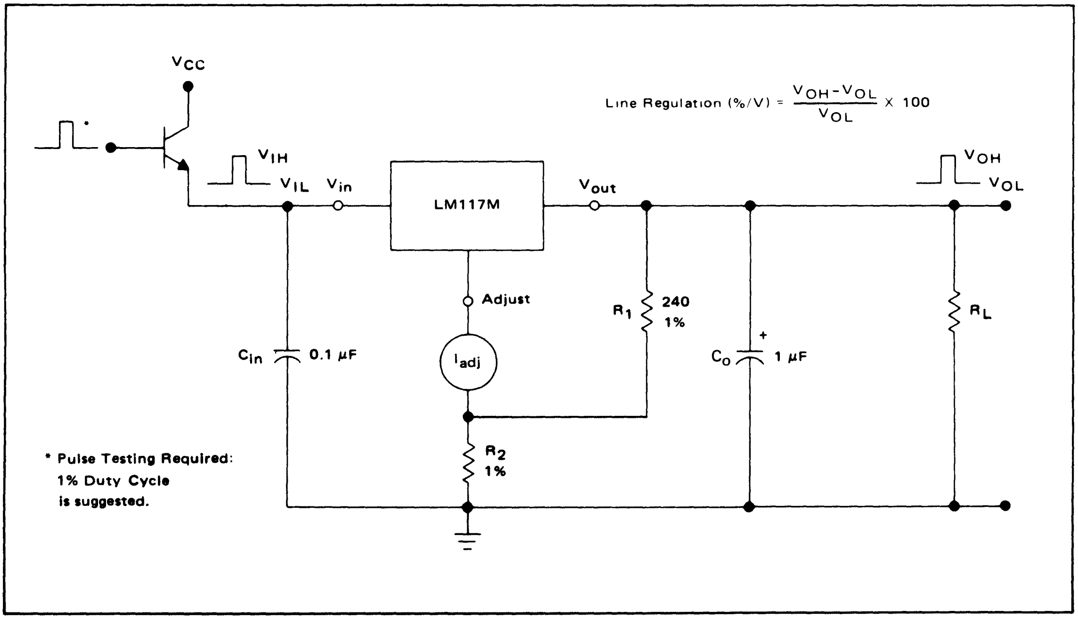
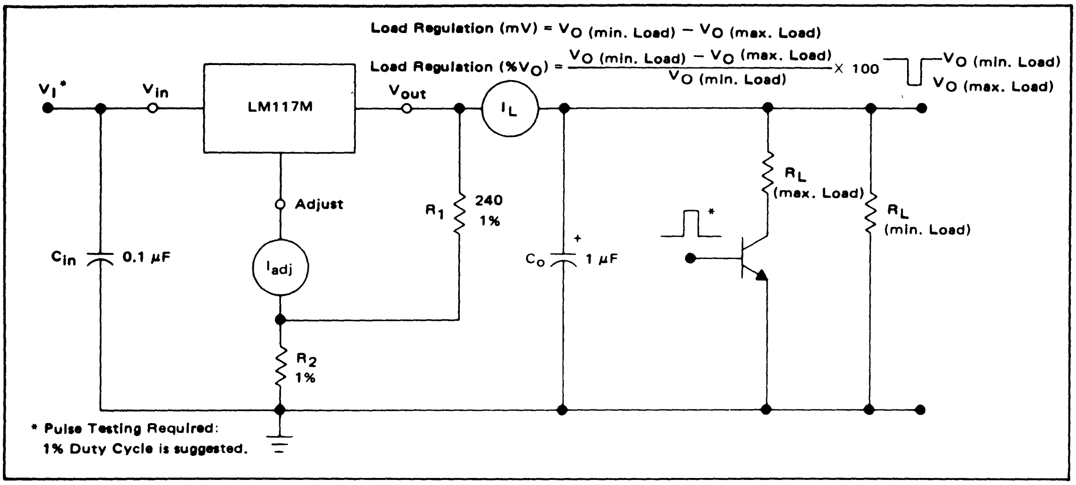
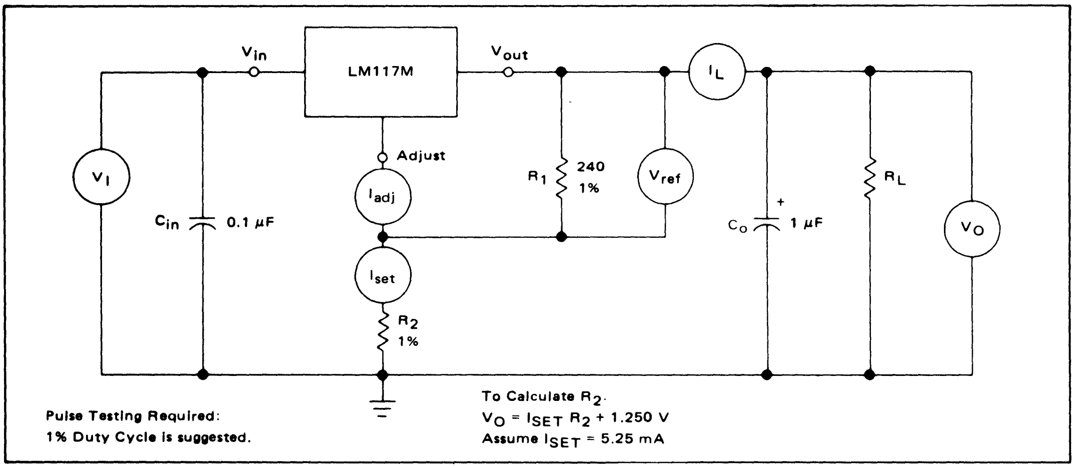


FIGURE 2 – LOAD REGULATION AND ΔI_{Adj}/LOAD TEST CIRCUIT



* Pulse Testing Required:
1% Duty Cycle is suggested.

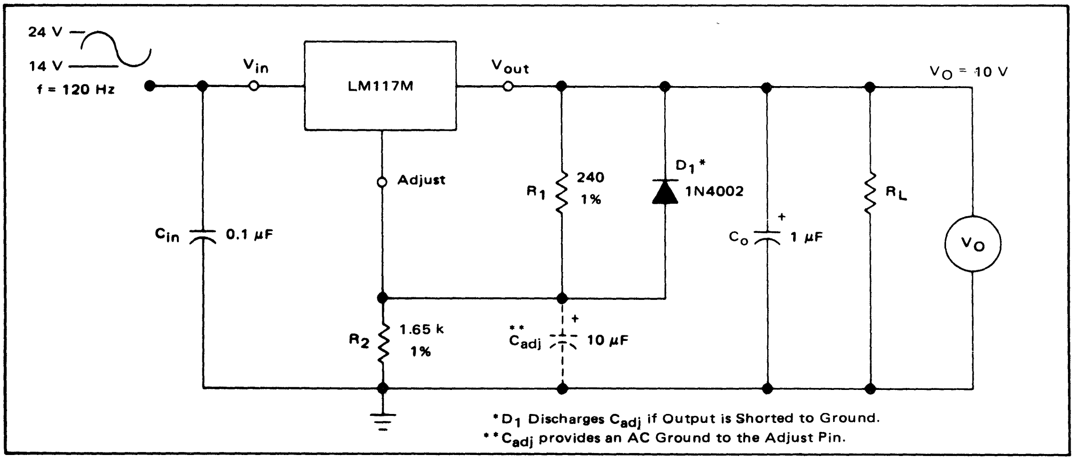
FIGURE 3 – STANDARD TEST CIRCUIT



Pulse Testing Required:
1% Duty Cycle is suggested.

To Calculate R_2.
$V_O = I_{SET} R_2 + 1.250$ V
Assume I_{SET} = 5.25 mA

FIGURE 4 – RIPPLE REJECTION TEST CIRCUIT



*D_1 Discharges C_{adj} if Output is Shorted to Ground.
**C_{adj} provides an AC Ground to the Adjust Pin.

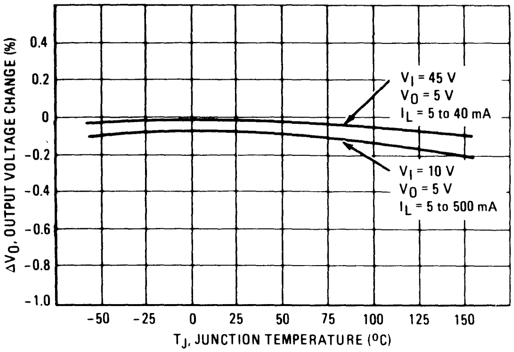


FIGURE 5 — LOAD REGULATION

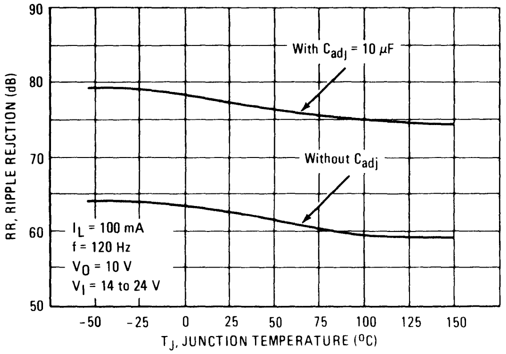


FIGURE 6 — RIPPLE REJECTION

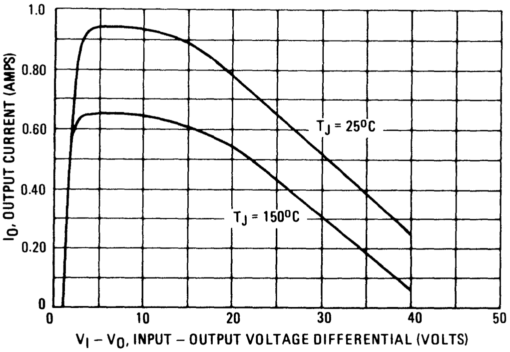


FIGURE 7 — CURRENT LIMIT

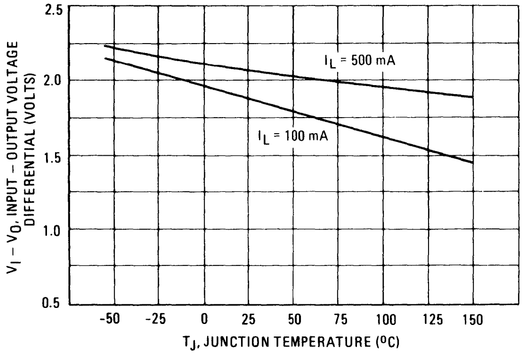


FIGURE 8 — DROPOUT VOLTAGE

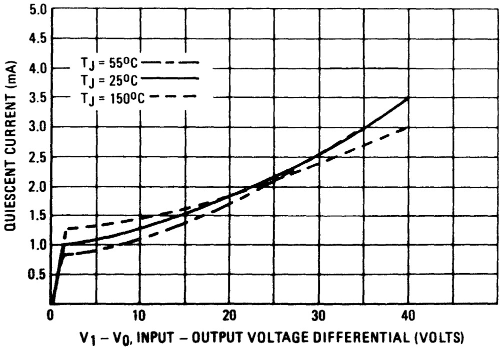


FIGURE 9 — MINIMUM OPERATING CURRENT

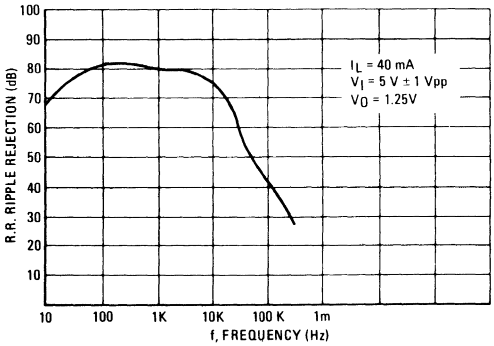


FIGURE 10 — RIPPLE REJECTION versus FREQUENCY

FIGURE 11 – TEMPERATURE STABILITY

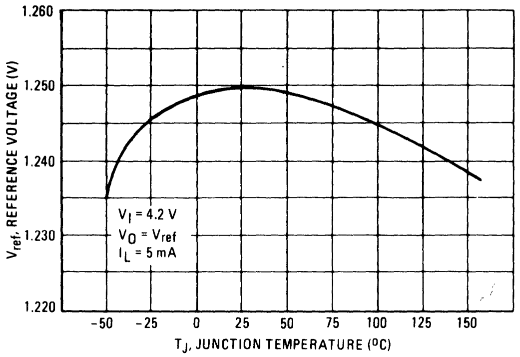


FIGURE 12 – ADJUSTMENT PIN CURRENT

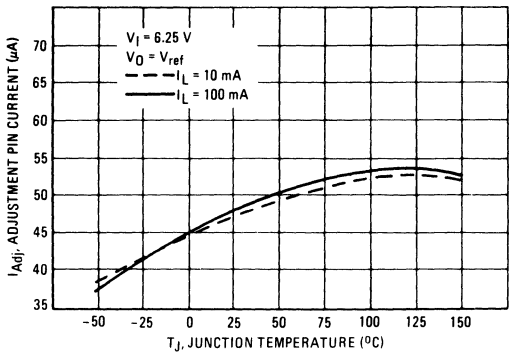


FIGURE 13 – LINE REGULATION

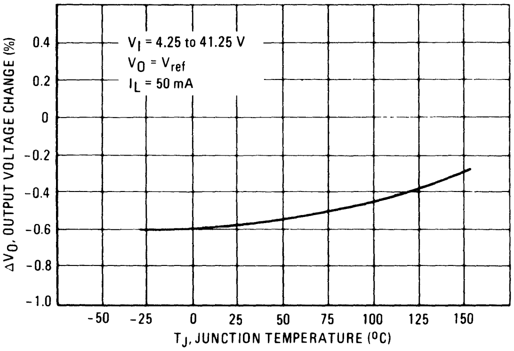


FIGURE 14 – OUTPUT NOISE

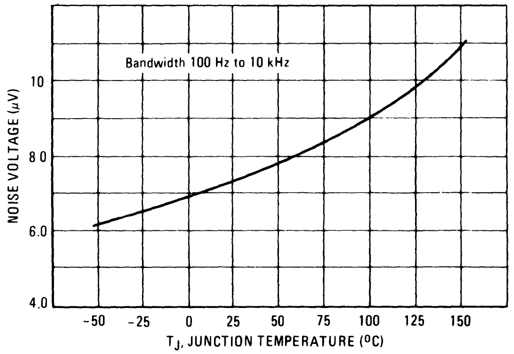


FIGURE 15 – LINE TRANSIENT RESPONSE

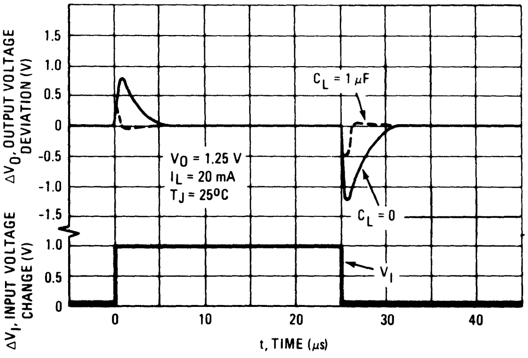
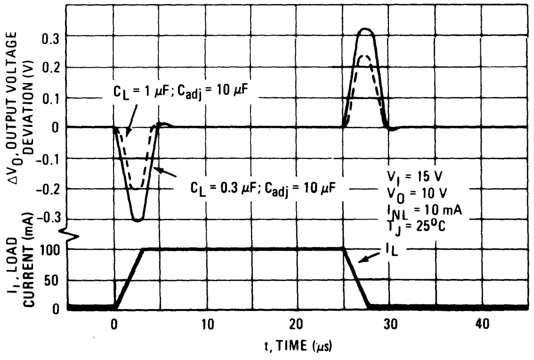


FIGURE 16 – LOAD TRANSIENT RESPONSE



APPLICATIONS INFORMATION

BASIC CIRCUIT OPERATION

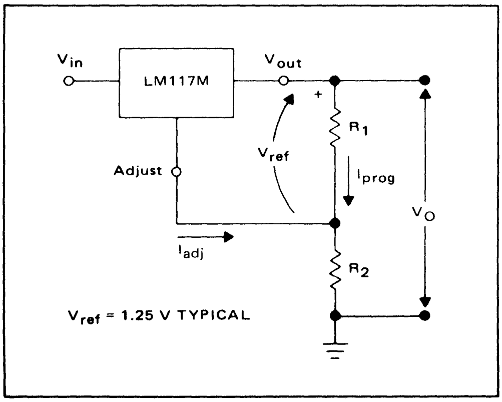
The LM117M is a 3-terminal floating regulator. In operation, the LM117M develops and maintains a nominal 1.25 volt reference (V_{ref}) between its output and adjustment terminals. This reference voltage is converted to a programming current (I_{prog}) by R1 (see Figure 17), and this constant current flows through R2 to ground. The regulated output voltage is given by:

$$V_O = V_{ref} (1 + \frac{R2}{R1}) + I_{adj}R2$$

Since the current from the adjustment terminal (I_{adj}) represents an error term in the equation, the LM117M was designed to control I_{adj} to less than 100 μA and keep it constant. To do this, all quiescent operating current is returned to the output terminal. This imposes the requirement for a minimum load current. If the load current is less than this minimum, the output voltage will rise.

Since the LM117M is a floating regulator, it is only the voltage differential across the circuit that is important to performance, and operation at high voltages with respect to ground is possible.

FIGURE 17 – BASIC CIRCUIT CONFIGURATION



LOAD REGULATION

The LM117M is capable of providing extremely good load regulation, but a few precautions are needed to obtain maximum performance. For best performance, the programming resistor (R1) should be connected as close to the regulator as possible to minimize line drops which effectively appear in series with the reference, thereby degrading regulation. The ground end of R2 can be returned near the load ground to provide remote ground sensing and improve load regulation.

EXTERNAL CAPACITORS

A 0.1 μF disc or1 μF tantalum input bypass capacitor (C_{in}) is recommended to reduce the sensitivity to input line impedance.

The adjustment terminal may be bypassed to ground to improve ripple rejection. This capacitor (C_{adj}) prevents ripple from being amplified as the output voltage is increased. A 10 μF capacitor should improve ripple rejection about 15 dB at 120 Hz in a 10 volt application.

Although the LM117M is stable with no output capacitance, like any feedback circuit, certain values of external capacitance can cause excessive ringing. An output capacitance (C_O) in the form of a 1 μF tantalum or 25 μF aluminum electrolytic capacitor on the output swamps this effect and insures stability.

PROTECTION DIODES

When external capacitors are used with any I.C. regulator it is sometimes necessary to add protection diodes to prevent the capacitors from discharging through low current points into the regulator.

Figure 18 shows the LM117M with the recommended protection diodes for output voltages in excess of 25 V or high capacitance values ($C_O > 10$ μF, $C_{adj} > 5$ μF). Diode D1 prevents C_O from discharging thru the I.C. during an input short circuit. Diode D2 protects against capacitor C_{adj} discharging through the I.C. during an output short circuit. The combination of diodes D1 and D2 prevents C_{adj} from discharging through the I.C. during an input short circuit.

FIGURE 18 – VOLTAGE REGULATOR WITH PROTECTION DIODES

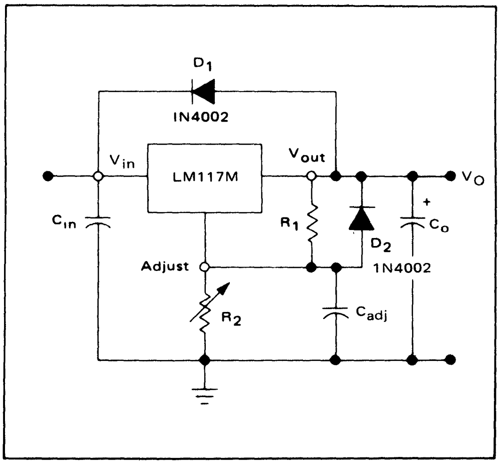
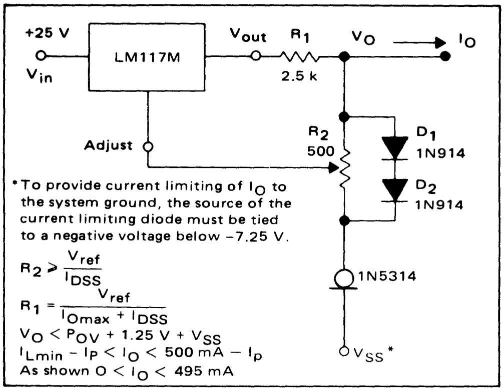


FIGURE 19 – ADJUSTABLE CURRENT LIMITER



*To provide current limiting of I_O to the system ground, the source of the current limiting diode must be tied to a negative voltage below −7.25 V.

$R_2 \geqslant \frac{V_{ref}}{I_{DSS}}$

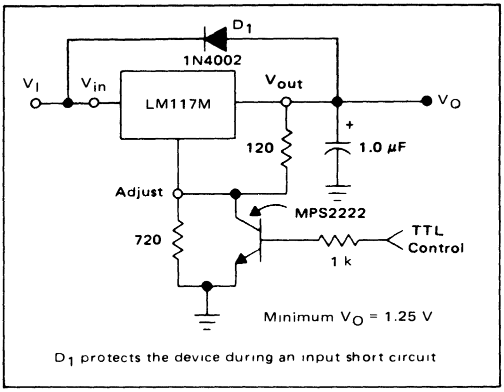
$R_1 = \frac{V_{ref}}{I_{Omax} + I_{DSS}}$

$V_O < P_{OV} + 1.25\ V + V_{SS}$

$I_{Lmin} - I_P < I_O < 500\ mA - I_P$

As shown $0 < I_O < 495$ mA

FIGURE 20 – 5 V ELECTRONIC SHUTDOWN REGULATOR



Minimum $V_O = 1.25$ V

D_1 protects the device during an input short circuit

FIGURE 21 – SLOW TURN-ON REGULATOR

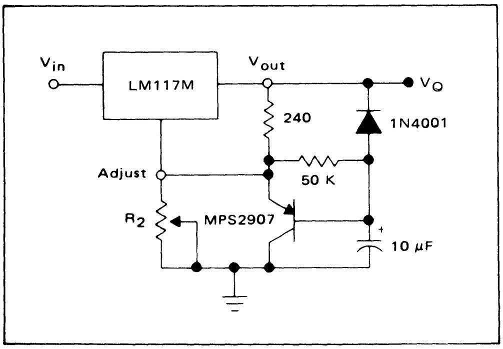
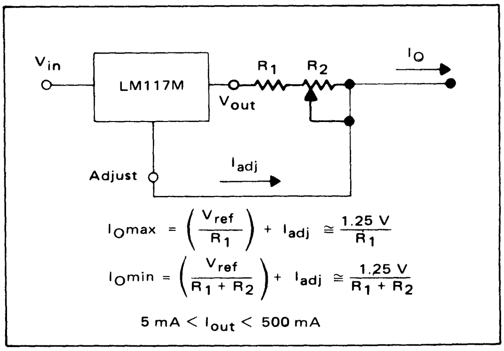


FIGURE 22 – CURRENT REGULATOR



$$I_{Omax} = \left(\frac{V_{ref}}{R_1}\right) + I_{adj} \cong \frac{1.25\ V}{R_1}$$

$$I_{Omin} = \left(\frac{V_{ref}}{R_1 + R_2}\right) + I_{adj} \cong \frac{1.25\ V}{R_1 + R_2}$$

$5\ mA < I_{out} < 500\ mA$

MOTOROLA

LM123, LM123A
LM223, LM223A
LM323, LM323A

Specifications and Applications Information

3 AMPERE, 5 VOLT POSITIVE VOLTAGE REGULATOR

The LM123, A/LM223, A/LM323, A are a family of monolithic integrated circuits which supply a fixed positive 5.0 volt output with a load driving capability in excess of 3.0 amperes. These three-terminal regulators employ internal current limiting, thermal shutdown, and safe-area compensation. An improved series with superior electrical characteristics and a 2% output voltage tolerance is available as A-suffix (LM123A/LM223A/LM323A) device types.

These regulators are offered in a hermetic TO-3 metal power package in three operating temperature ranges. A 0°C to +125°C temperature range version is also available in a low cost TO-220 plastic power package.

Although designed primarily as a fixed voltage regulator, these devices can be used with external components to obtain adjustable voltages and currents. This series of devices can be used with a series pass transistor to supply up to 15 amperes at 5.0 volts.

- Output Current in Excess of 3.0 Amperes
- Available with 2% Output Voltage Tolerance
- No external Components Required
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Thermal Regulation and Ripple Rejection Have Specified Limits

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Input Voltage		V_{in}	20	Vdc
Power Dissipation		P_D	Internally Limited	
Operating Junction Temperature Range	LM123, A LM223, A LM323, A	T_J	-55 to +150 -25 to +150 0 to +125	°C
Storage Temperature Range		T_{stg}	-65 to +150	°C
Lead Temperature (Soldering, 10 s)		T_{solder}	300	°C

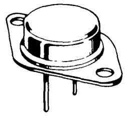
ORDERING INFORMATION

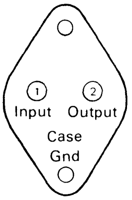
Device	Output Voltage Tolerance	Junction Temperature Range	Package
LM123K LM123AK	6% 2%	-55 to +150°C	Metal Power
LM223K LM223AK	6% 2%	-25 to +150°C	
LM323K LM323AK	4% 2%	0 to +125°C	
LM323T LM323AT	4% 2%		Plastic Power

3-AMPERE, 5 VOLT POSITIVE VOLTAGE REGULATOR

SILICON MONOLITHIC INTEGRATED CIRCUIT

K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 Type)



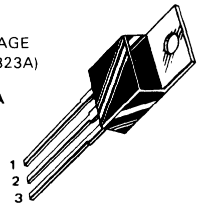


(Bottom View)

Pin 1 INPUT
2 OUTPUT
CASE GROUND

T SUFFIX
PLASTIC PACKAGE
(LM323 and LM323A)

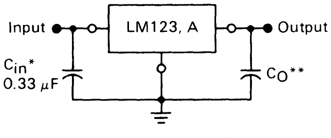
CASE 221A
(TO-220)



Pin 1 INPUT
2. GROUND
3 OUTPUT

(Heatsink surface connected to Pin 2)

STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.5 V above the output voltage even during the low point on the input ripple voltage.

* = C_{in} is required if regulator is located an appreciable distance from power supply filter. (See Applications Information for details.)

** = C_O is not needed for stability; however, it does improve transient response.

ELECTRICAL CHARACTERISTICS ($T_J = T_{low}$ to T_{high} [see Note 1] unless otherwise specified.)

Characteristic	Symbol	LM123A/LM223A/LM323A			LM123/LM223			LM323			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (V_{in} = 7.5 V, 0 ≤ I_{out} ≤ 3.0 A, T_J = 25°C)	V_O	4.9	5.0	5.1	4.7	5.0	5.3	4.8	5.0	5.2	V
Output Voltage (7.5 V ≤ V_{in} ≤ 15 V, 0 ≤ I_{out} ≤ 3.0 A, P ≤ P_{max} [Note 2])	V_O	4.8	5.0	5.2	4.6	5.0	5.4	4.75	5.0	5.25	V
Line Regulation (7.5 V ≤ V_{in} ≤ 15 V, T_J = 25°C) (Note 3)	Reg_{line}	—	1.0	15	—	1.0	25	—	1.0	25	mV
Load Regulation (V_{in} = 7.5 V, 0 ≤ I_{out} ≤ 3.0 A, T_J = 25°C) (Note 3)	Reg_{load}	—	10	50	—	10	100	—	10	100	mV
Thermal Regulation (Pulse = 10 ms, P = 20 W, T_A = 25°C)	Reg_{therm}	—	0.001	0.01	—	0.002	0.03	—	0.002	0.03	%V_O/W
Quiescent Current (7.5 V ≤ V_{in} ≤ 15 V, 0 ≤ I_{out} ≤ 3.0 A)	I_B	—	3.5	10	—	3.5	20	—	3.5	20	mA
Output Noise Voltage (10 Hz ≤ f ≤ 100 kHz, T_J = 25°C)	V_N	—	40	—	—	40	—	—	40	—	μV_{rms}
Ripple Rejection (8.0 V ≤ V_{in} ≤ 18 V, I_{out} = 2.0 A, f = 120 Hz, T_J = 25°C)	RR	66	75	—	62	75	—	62	75	—	dB
Short Circuit Current Limit (V_{in} = 15 V, T_J = 25°C) (V_{in} = 7.5 V, T_J = 25°C)	I_{SC}	 — —	 4.5 5.5	 — —	 — —	 4.5 5.5	 — —	 — —	 4.5 5.5	 — —	A
Long Term Stability	S	—	—	35	—	—	35	—	—	35	mV
Thermal Resistance Junction to Case (Note 4)	$R_{\theta JC}$	—	2.0	—	—	2.0	—	—	2.0	—	°C/W

Note 1. T_{low} = −55°C for LM123, A T_{high} = +150°C for LM123, A
= −25°C for LM223, A = +150°C for LM223, A
= 0°C for LM323, A = +125°C for LM323, A

Note 2. Although power dissipation is internally limited, specifications apply only for P ≤ P_{max}
P_{max} = 30 W for K (TO-3) package
P_{max} = 25 W for T (TO-220) package

Note 3. Load and line regulation are specified at constant junction temperature. Pulse testing is required with a pulse width ≤ 1.0 ms and a duty cycle ≤ 5%.

Note 4. Without a heat sink, the thermal resistance ($R_{\theta JA}$) is 35°C/W for the TO-3, and 65°C/W for the TO-220 packages. With a heat sink, the effective thermal resistance can approach the specified values of 2.0 °C/W, depending on the efficiency of the heat sink.

VOLTAGE REGULATOR PERFORMANCE

The performance of a voltage regulator is specified by its immunity to changes in load, input voltage, power dissipation, and temperature. Line and load regulation are tested with a pulse of short duration (< 100 µs) and are strictly a function of electrical gain. However, pulse widths of longer duration (> 1.0 ms) are sufficient to affect temperature gradients across the die. These temperature gradients can cause a change in the output voltage, in addition to changes caused by line and load regulation. Longer pulse widths and thermal gradients make it desirable to specify thermal regulation.

Thermal regulation is defined as the change in output voltage caused by a change in dissipated power for a specified time, and is expressed as a percentage output voltage change per watt. The change in dissipated power can be caused by a change in either the input voltage or the load current. Thermal regulation is a function of I.C. layout and die attach techniques, and usually occurs within 10 ms of a change in power dissipation. After 10 ms, additional changes in the output voltage are due to the temperature coefficient of the device.

Figure 1 shows the line and thermal regulation response of a typical LM123A to a 20 watt input pulse. The variation of the output voltage due to line regulation is labeled ① and the thermal regulation component is labeled ②. Figure 2 shows the load and thermal regulation response of a typical LM123A to a 20 watt load pulse. The output voltage variation due to load regulation is labeled ① and the thermal regulation component is labeled ②.

SCHEMATIC DIAGRAM

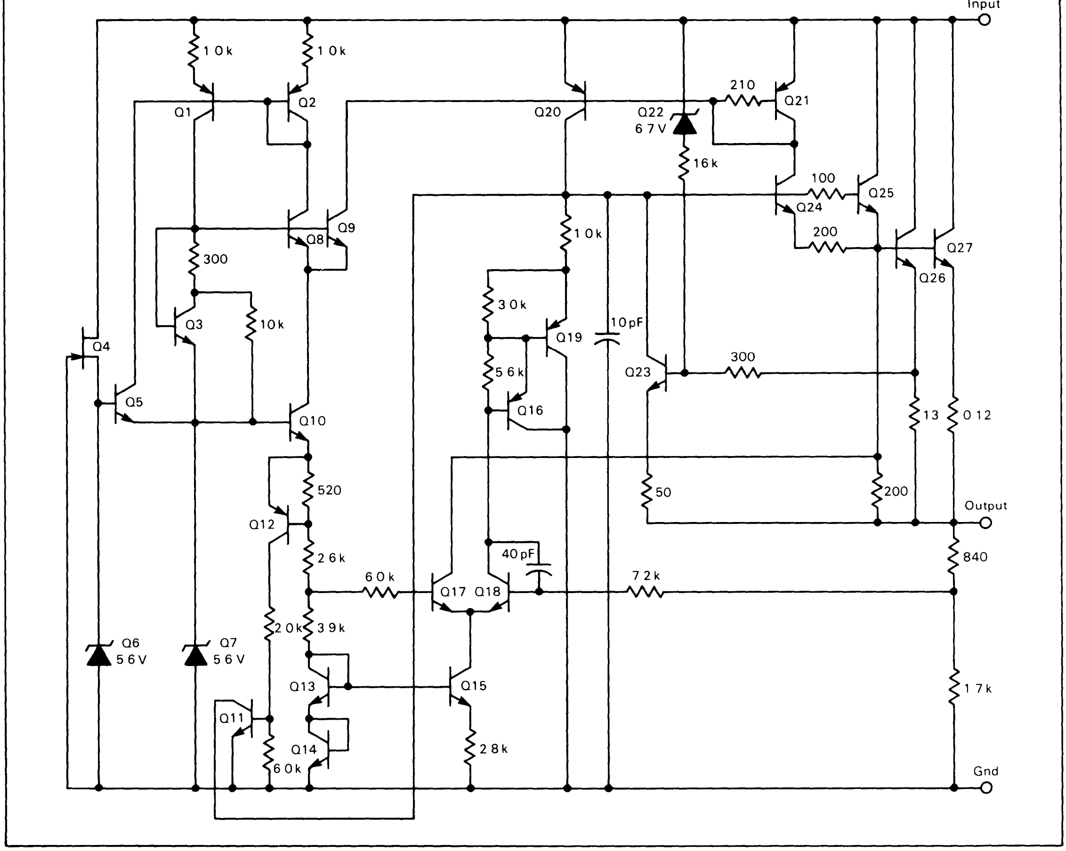
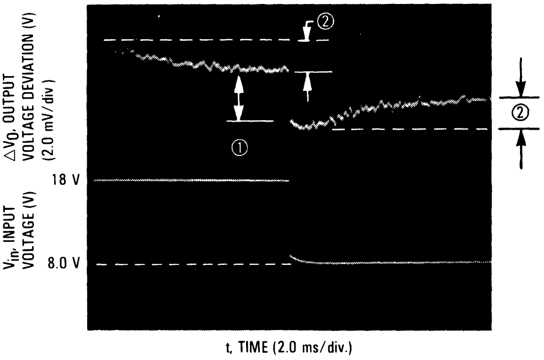
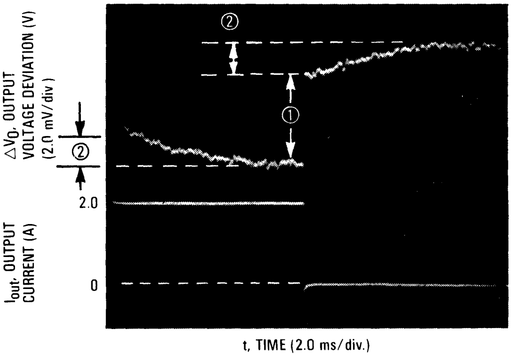


FIGURE 1 — LINE AND THERMAL REGULATION



LM123A
$V_O = 5.0$ V
$V_{in} = 8.0$ V → 18 V → 8.0 V ① = $\text{Reg}_{\text{line}} = 2.4$ mV
$I_{out} = 2.0$ A ② = $\text{Reg}_{\text{therm}} = 0.0015\% V_O$/W

FIGURE 2 — LOAD AND THERMAL REGULATION



LM123A
$V_O = 5.0$ V
$V_{in} = 15$ ① = $\text{Reg}_{\text{load}} = 4.4$ mV
$I_{out} = 0$ A → 2.0 A → 0 A ② = $\text{Reg}_{\text{therm}} = 0.0015\% V_O$/W

FIGURE 3 — TEMPERATURE STABILITY

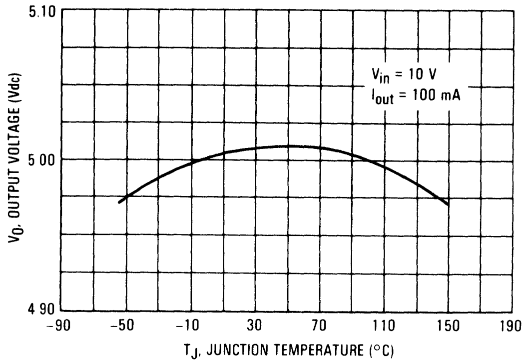


FIGURE 4 — OUTPUT IMPEDANCE

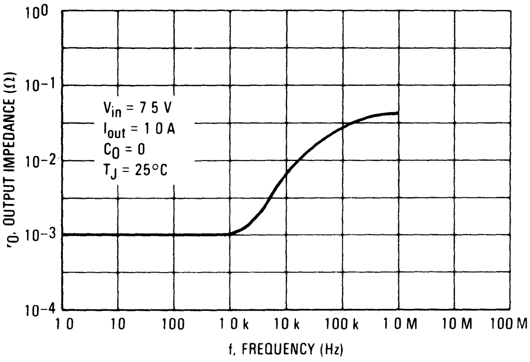


FIGURE 5 — RIPPLE REJECTION versus FREQUENCY

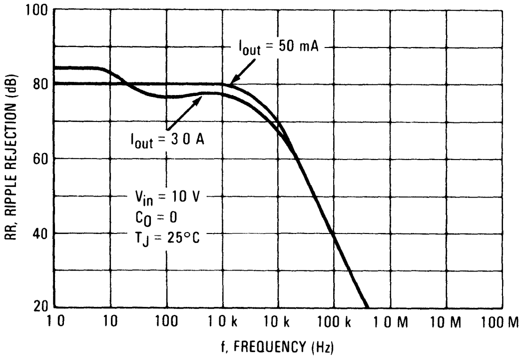


FIGURE 6 — RIPPLE REJECTION versus OUTPUT CURRENT

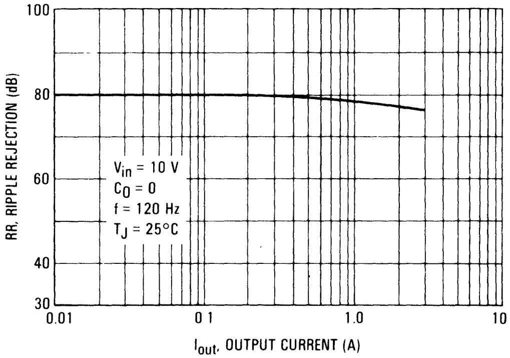


FIGURE 7 — QUIESCENT CURRENT versus INPUT VOLTAGE

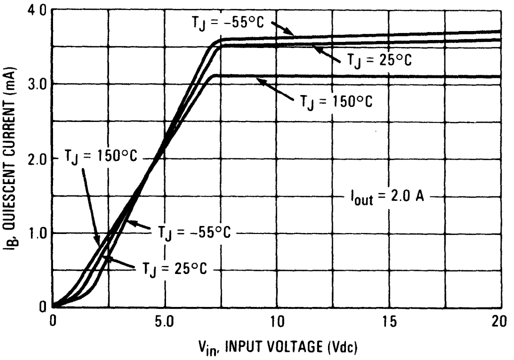


FIGURE 8 — QUIESCENT CURRENT versus OUTPUT CURRENT

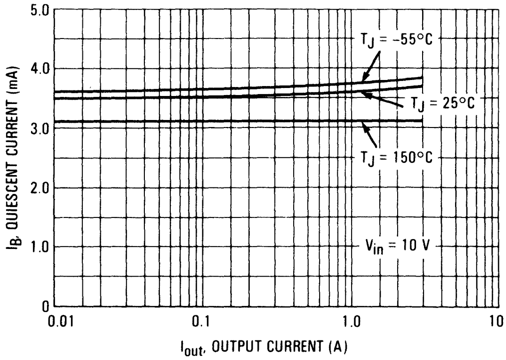


FIGURE 9 — DROPOUT VOLTAGE

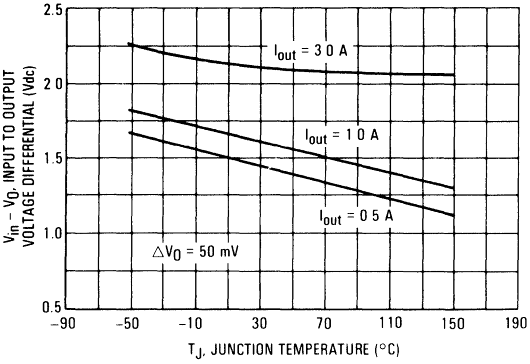


FIGURE 10 — SHORT CIRCUIT CURRENT

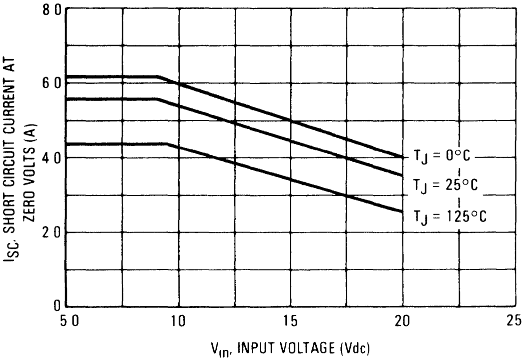


FIGURE 11 — LINE TRANSIENT RESPONSE

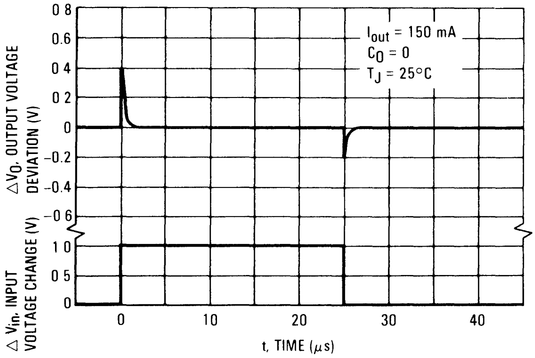


FIGURE 12 — LOAD TRANSIENT RESPONSE

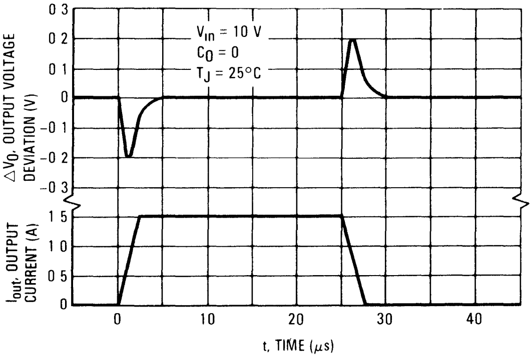


FIGURE 13 — MAXIMUM AVERAGE POWER DISSIPATION FOR LM123K and LM223K

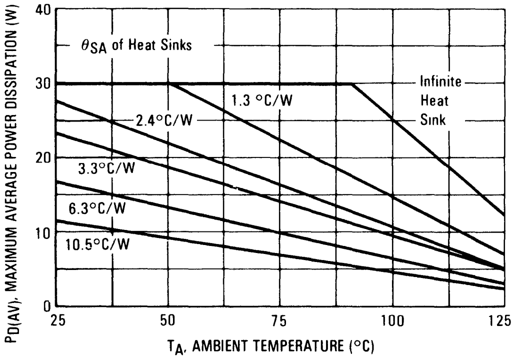
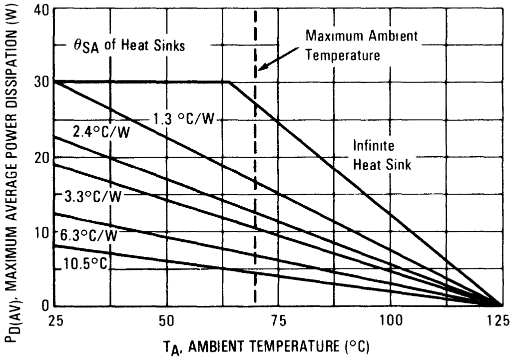


FIGURE 14 — MAXIMUM AVERAGE POWER DISSIPATION FOR LM323K



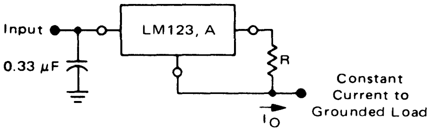
APPLICATIONS INFORMATION

Design Considerations

The LM123,A Series of fixed voltage regulators are designed with Thermal Overload Protection that shuts down the circuit when subjected to an excessive power overload condition, Internal Short-Circuit Protection that limits the maximum current the circuit will pass, and Output Transistor Safe-Area Compensation that reduces the output short-circuit current as the voltage across the pass transistor is increased

In many low current applications, compensation capacitors are not required However, it is recommended that the regulator input be bypassed with a capacitor if the regulator is connected to the power supply filter with long wire lengths, or if the output load capacitance is large An input bypass capacitor should be selected to provide good high-frequency characteristics to insure stable operation under all load conditions A 0 33 μF or larger tantalum, mylar, or other capacitor having low internal impedance at high frequencies should be chosen. The bypass capacitor should be mounted with the shortest possible leads directly across the regulator's input terminals Normally good construction techniques should be used to minimize ground loops and lead resistance drops since the regulator has no external sense lead

FIGURE 15 — CURRENT REGULATOR



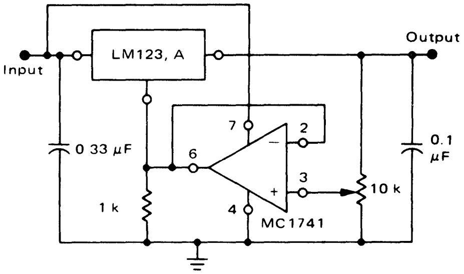
The LM123, A regulator can also be used as a current source when connected as above Resistor R determines the current as follows

$$I_O = \frac{5\ V}{R} + I_Q$$

$\Delta I_Q \cong 0.7$ mA over line, load and temperature changes
$I_Q \cong 3.5$ mA

For example, a 2-ampere current source would require R to be a 2.5 ohm, 15 W resistor and the output voltage compliance would be the input voltage less 7 5 volts

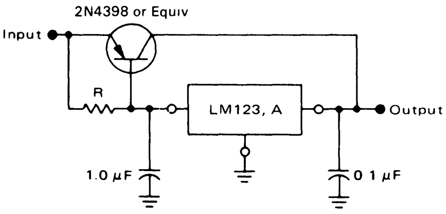
FIGURE 16 — ADJUSTABLE OUTPUT REGULATOR



V_O, 8.0 V to 20 V
$V_{in} - V_O \geqslant 2\ 5$ V

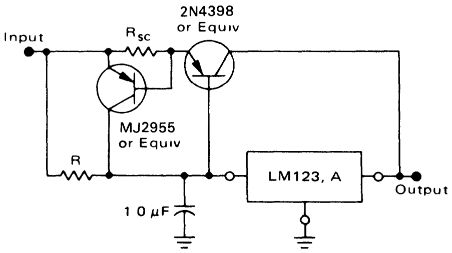
The addition of an operational amplifier allows adjustment to higher or intermediate values while retaining regulation characteristics The minimum voltage obtainable with this arrangement is 3 0 volts greater than the regulator voltage

FIGURE 17 — CURRENT BOOST REGULATOR



The LM123, A series can be current boosted with a PNP transistor. The 2N4398 provides current to 15 amperes. Resistor R in conjunction with the V_{BE} of the PNP determines when the pass transistor begins conducting; this circuit is not short-circuit proof. Input-output differential voltage minimum is increased by the V_{BE} of the pass transistor.

FIGURE 18 — CURRENT BOOST WITH SHORT-CIRCUIT PROTECTION



The circuit of Figure 17 can be modified to provide supply protection against short circuits by adding a short-circuit sense resistor, R_{SC}, and an additional PNP transistor. The current sensing PNP must be able to handle the short-circuit current of the three-terminal regulator. Therefore, an eight-ampere plastic power transistor is specified.

MOTOROLA

LM137
LM237
LM337

Specifications and Applications Information

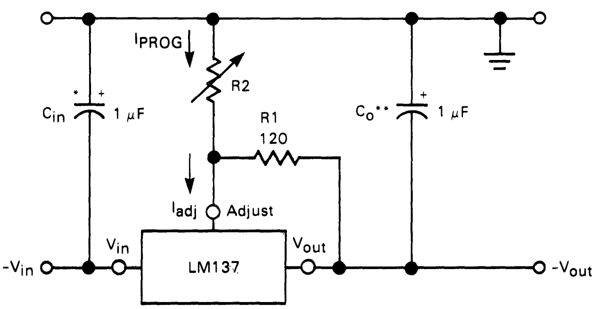
3-TERMINAL ADJUSTABLE OUTPUT NEGATIVE VOLTAGE REGULATOR

The LM137/237/337 are adjustable 3-terminal negative voltage regulators capable of supplying in excess of 1.5 A over an output voltage range of -1.2 V to -37 V These voltage regulators are exceptionally easy to use and require only two external resistors to set the output voltage. Further, they employ internal current limiting, thermal shutdown and safe area compensation, making them essentially blow-out proof.

The LM137 series serve a wide variety of applications including local, on-card regulation. This device can also be used to make a programmable output regulator; or, by connecting a fixed resistor between the adjustment and output, the LM137 series can be used as a precision current regulator.

- Output Current in Excess of 1.5 Ampere in TO-3 and TO-220 Packages
- Output Current in Excess of 0.5 Ampere in TO-39 Package
- Output Adjustable Between -1.2 V and -37 V
- Internal Thermal Overload Protection
- Internal Short-Circuit-Current Limiting, Constant with Temperature
- Output Transistor Safe-Area Compensation
- Floating Operation for High Voltage Applications
- Standard 3-Lead Transistor Packages
- Eliminates Stocking Many Fixed Voltages

STANDARD APPLICATION



*C_{in} is required if regulator is located more than 4 inches from power supply filter. A 1 µF solid tantalum or 10 µF aluminum electrolytic is recommended.

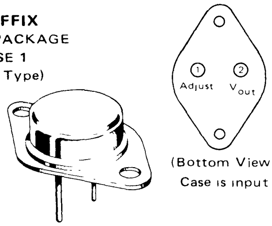
**C_O is necessary for stability. A 1 µF solid tantalum or 10 µF aluminum electrolytic is recommended.

$$V_{out} = -1.25\ V\ (1 + \frac{R2}{R1})$$

3-TERMINAL ADJUSTABLE NEGATIVE VOLTAGE REGULATOR

SILICON MONOLITHIC INTEGRATED CIRCUIT

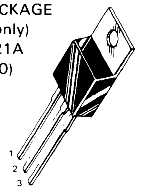
K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 Type)



(Bottom View)
Case is input

Pins 1 and 2 electrically isolated from case. Case is third electrical connection.

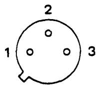
T SUFFIX
PLASTIC PACKAGE
(LM337 only)
CASE 221A
(TO-220)



Pin 1 Adjust
Pin 2 V_{in}
Pin 3 V_{out}

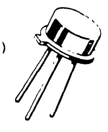
Heatsink surface connected to Pin 2

H SUFFIX
METAL PACKAGE
CASE 79
(TO 39)



(Bottom View)

(Case is output)



Pin 1 Adjust
Pin 2 Output
Pin 3 Input

ORDERING INFORMATION

Device	Temperature Range	Package
LM137H	T_J = -55°C to +150°C	Metal Can
LM137K	T_J = -55°C to +150°C	Metal Power
LM237H	T_J = -25°C to +150°C	Metal Can
LM237K	T_J = -25°C to +150°C	Metal Power
LM337H	T_J = 0°C to +125°C	Metal Can
LM337K	T_J = 0°C to +125°C	Metal Power
LM337T	T_J = 0°C to +125°C	Plastic Power

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Input-Output Voltage Differential		V_I-V_O	40	Vdc
Power Dissipation		P_D	Internally Limited	
Operating Junction Temperature Range	LM137 LM237 LM337	T_J	-55 to +150 -25 to +150 0 to +125	°C
Storage Temperature Range		T_{stg}	-65 to +150	°C

ELECTRICAL CHARACTERISTICS

($\lvert V_I - V_O \rvert$ = 5 V, I_O = 0.5 A for K and T packages, I_O = 0.1 A for H package, T_J = T_{low} to T_{high} [see Note 1], I_{max} and P_{max} per Note 2, unless otherwise specified)

Characteristic	Figure	Symbol	LM137/237			LM337			Unit
			Min	Typ	Max	Min	Typ	Max	
Line Regulation (Note 3) T_A = 25°C, 3 V $\leq \lvert V_I$-$V_O \rvert \leq$ 40 V	1	Reg_{line}	—	0.01	0.02	—	0.01	0.04	%/V
Load Regulation (Note 3), T_A = 25°C, 10 mA $\leq I_O \leq I_{max}$ $\lvert V_O \rvert \leq$ 5 V $\lvert V_O \rvert \geq$ 5 V	2	Reg_{load}	— —	15 0.3	25 0.5	— —	15 0.3	50 1.0	mV % V_O
Thermal Regulation 10 mS Pulse, T_A = 25°C	—	Reg_{therm}	—	0.002	0.02	—	0.003	0.04	% V_O/W
Adjustment Pin Current	3	I_{adj}	—	65	100	—	65	100	μA
Adjustment Pin Current Change 2.5 V $\leq \lvert V_I$-$V_O \rvert \leq$ 40 V, 10 mA $\leq I_L \leq I_{max}$, $P_D \leq P_{max}$, T_A = 25°C	1,2	ΔI_{adj}	—	2.0	5.0	—	2.0	5.0	μA
Reference Voltage (Note 4) 3 V $\leq \lvert V_I$-$V_O \rvert \leq$ 40 V, 10 mA $\leq I_O \leq I_{max}$, $P_D \leq P_{max}$, T_A = 25°C T_{low} to T_{high}	3	V_{ref}	-1.225 -1.20	-1.250 -1.25	-1.275 -1.30	-1.213 -1.20	-1.250 -1.25	-1 287 -1.30	V
Line Regulation (Note 3) 3 V $\leq \lvert V_I$-$V_O \rvert \leq$ 40 V	1	Reg_{line}	—	0.02	0.05	—	0.02	0.07	%/V
Load Regulation (Note 3) 10 mA $\leq I_O \leq I_{max}$ $\lvert V_O \rvert \leq$ 5 V $\lvert V_O \rvert \geq$ 5 V	2	Reg_{load}	— —	20 0.3	50 1.0	— —	20 0.3	70 1.5	mV %V_O
Temperature Stability ($T_{low} \leq T_J \leq T_{high}$)	3	T_S	—	0.6	—	—	0.6	—	%V_O
Minimum Load Current to Maintain Regulation ($\lvert V_I$-$V_O \rvert \leq$ 10 V) ($\lvert V_I$-$V_O \rvert \leq$ 40 V)	3	I_{Lmin}	— —	1.2 2.5	3.0 5.0	— —	1.5 2.5	6.0 10	mA
Maximum Output Current $\lvert V_I$-$V_O \rvert \leq$ 15 V, $P_D \leq P_{max}$ K and T Packages H Package $\lvert V_I$-$V_O \rvert \leq$ 40 V, $P_D \leq P_{max}$, T_J = 25°C K and T Packages H Package	3	I_{max}	1.5 0.5 0.24 0.15	2.2 0.8 0.4 0.20	— — — —	1.5 0.5 0.15 0.10	2.2 0.8 0.4 0.20	— — — —	A
RMS Noise, % of V_O T_A = 25°C, 10 Hz $\leq$ f $\leq$ 10 kHz	—	N	—	0.003	—	—	0.003	—	%V_O
Ripple Rejection, V_O = -10 V, f = 120 Hz (Note 5) Without C_{adj} C_{adj} = 10 μF	4	RR	— 66	60 77	— —	— 66	60 77	— —	dB
Long Term Stability, T_J = T_{high} (Note 6) T_A = 25°C for Endpoint Measurements	3	S	—	0.3	1.0	—	0.3	1.0	%/1.0 k Hrs.
Thermal Resistance Junction to Case H Package (TO-39) K Package (TO-3) T Package (TO-220)	—	$R_{\theta JC}$	— — —	12 2.3 —	15 3.0 —	— — —	12 2.3 4.0	15 3.0 —	°C/W

NOTES.

(1) T_{low} = -55°C for LM137 T_{high} = +150°C for LM137
= -25°C for LM237 = +150°C for LM237
= 0°C for LM337 = +125°C for LM337

(2) I_{max} = 1 5 A for K (TO-3) and T (TO-220 Packages
= 0.5 A for H (TO-39) Package
P_{max} = 20 W for K (TO-3) and T (TO-220) Packages
= 2 W for H (TO-39) Package

(3) Load and line regulation are specified at a constant junction temperature. Pulse testing with a low duty cycle is used Change in V_O because of heating effects is covered under the Thermal Regulation specification.

(4) Selected devices with tightened tolerance reference voltage available.

(5) C_{adj}, when used, is connected between the adjustment pin and ground

(6) Since Long Term Stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot.

(7) Power dissipation within an I C. voltage regulator produces a temperature gradient on the die, affecting individual I C components on the die. These effects can be minimized by proper integrated circuit design and layout techniques. Thermal Regulation is the effect of these temperature gradients on the output voltage and is expressed in percentage of output change per watt of power change in a specified time

SCHEMATIC DIAGRAM

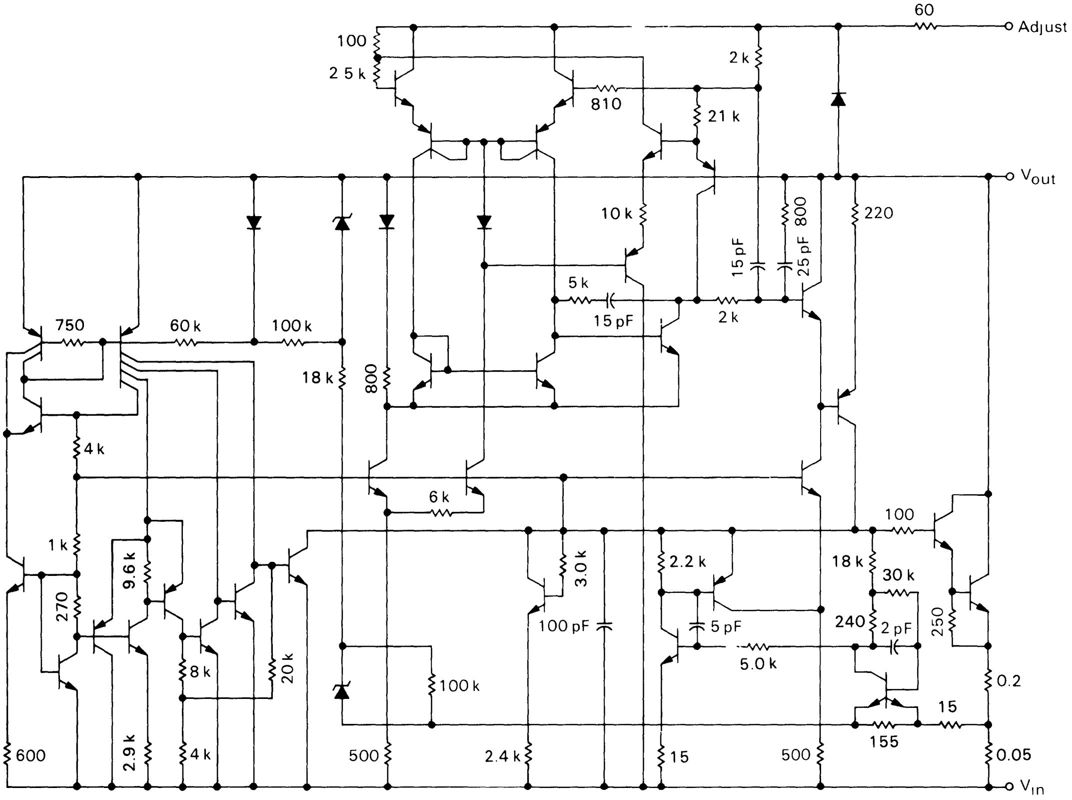


FIGURE 1 – LINE REGULATION AND ΔI_{adj}/LINE TEST CIRCUIT

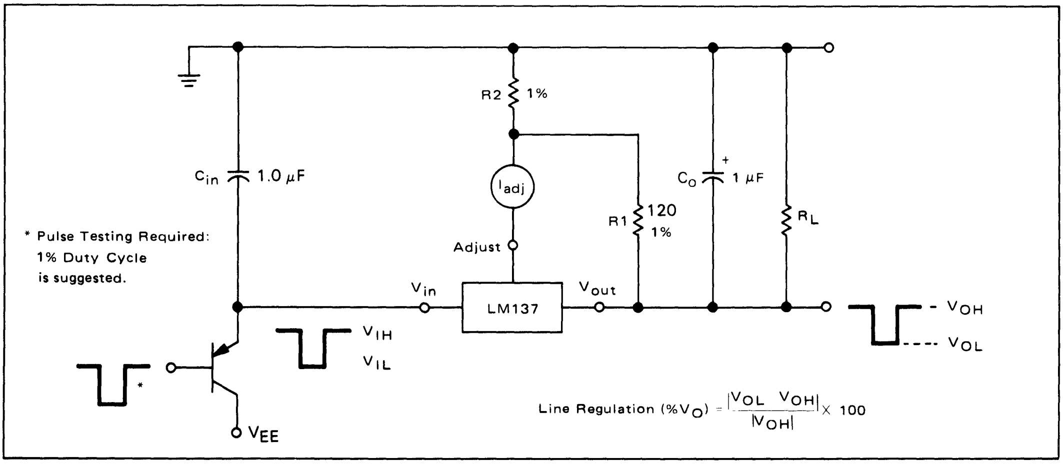
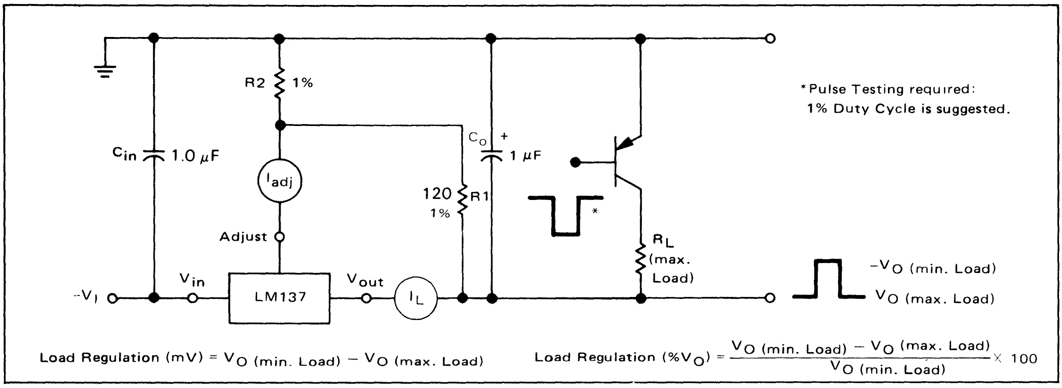


FIGURE 2 – LOAD REGULATION AND ΔI_{adj}/LOAD TEST CIRCUIT

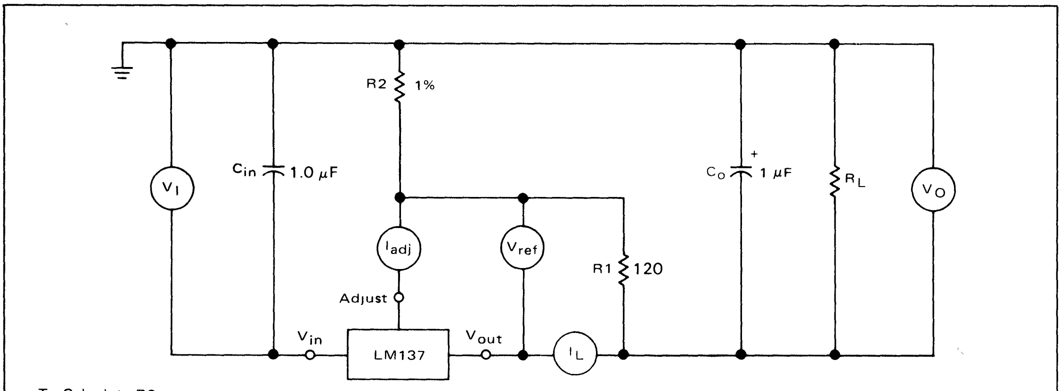


*Pulse Testing required:
1% Duty Cycle is suggested.

Load Regulation (mV) = V_O (min. Load) − V_O (max. Load)

$$\text{Load Regulation } (\%V_O) = \frac{V_O \text{ (min. Load)} - V_O \text{ (max. Load)}}{V_O \text{ (min. Load)}} \times 100$$

FIGURE 3 – STANDARD TEST CIRCUIT

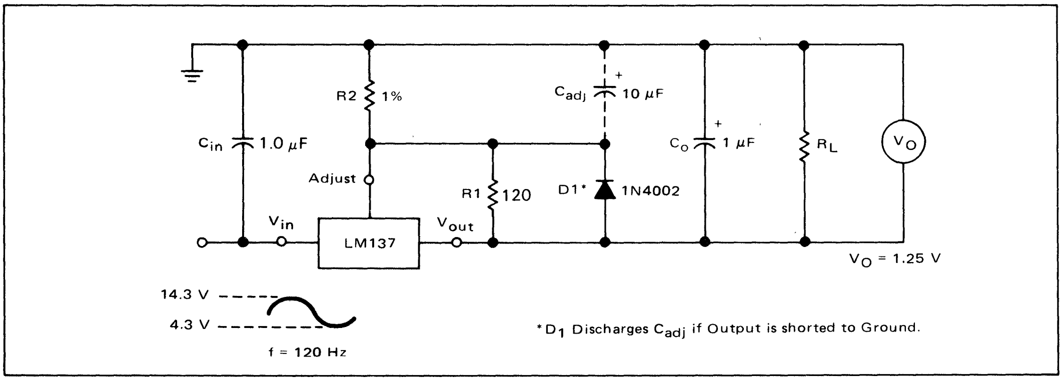


To Calculate R2:

$R_2 = \left(\frac{V_O}{V_{ref}} - 1\right) R1$ This assumes I_{adj} is negligible.

Pulse Testing Required.
1% Duty Cycle is suggested.

FIGURE 4 – RIPPLE REJECTION TEST CIRCUIT



*D_1 Discharges C_{adj} if Output is shorted to Ground.

FIGURE 5 — LOAD REGULATION

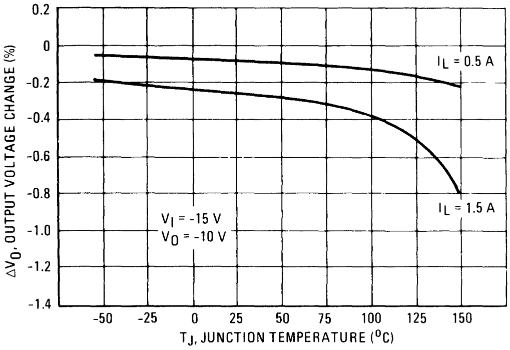


FIGURE 6 — CURRENT LIMIT

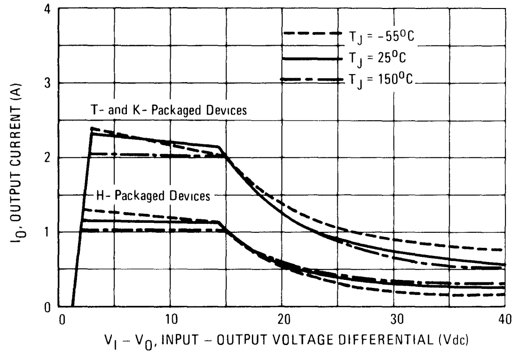


FIGURE 7 — ADJUSTMENT PIN CURRENT

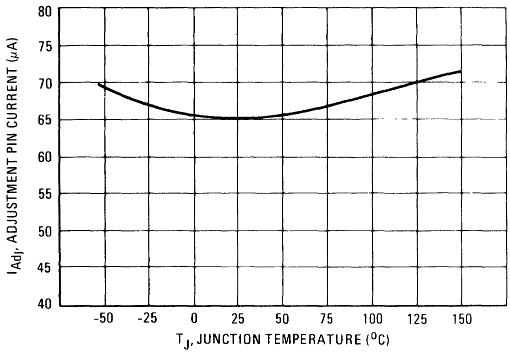


FIGURE 8 — DROPOUT VOLTAGE

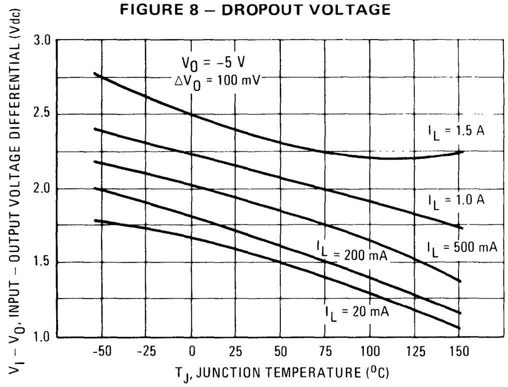


FIGURE 9 — TEMPERATURE STABILITY

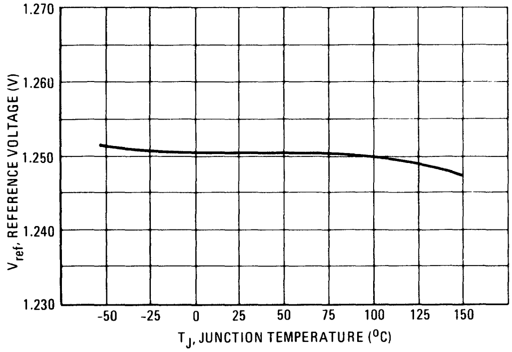


FIGURE 10 — MINIMUM OPERATING CURRENT

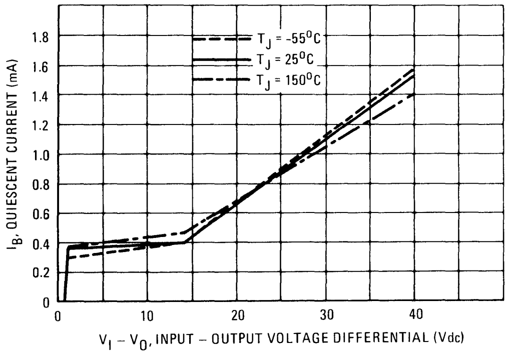


FIGURE 11 – RIPPLE REJECTION VS OUTPUT VOLTAGE

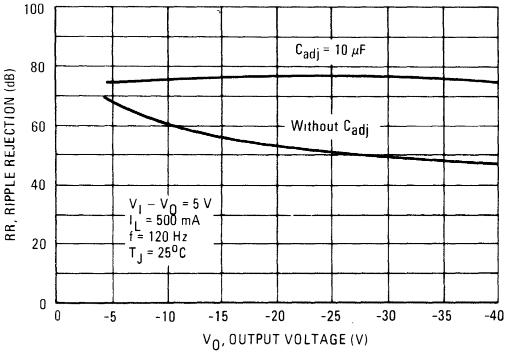


FIGURE 12 – RIPPLE REJECTION VS. OUTPUT CURRENT

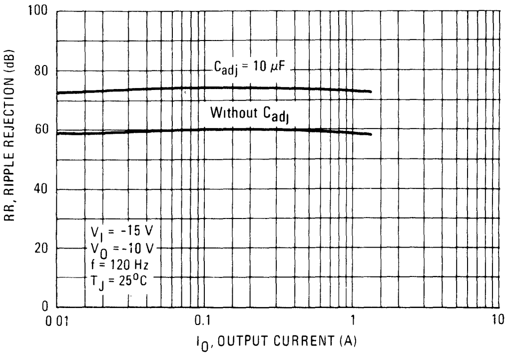


FIGURE 13 – RIPPLE REJECTION VS. FREQUENCY

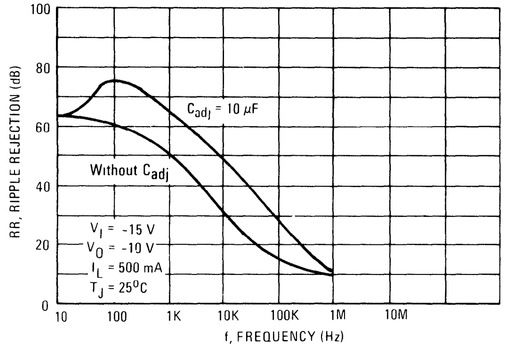


FIGURE 14 – OUTPUT IMPEDANCE

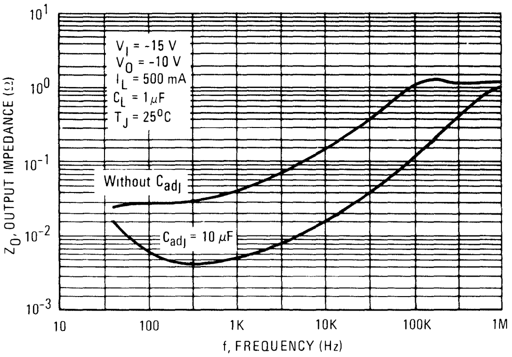


FIGURE 15 – LINE TRANSIENT RESPONSE

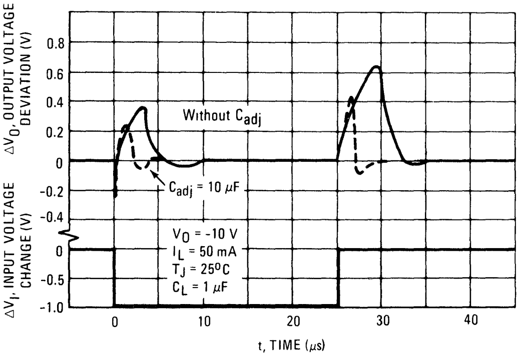
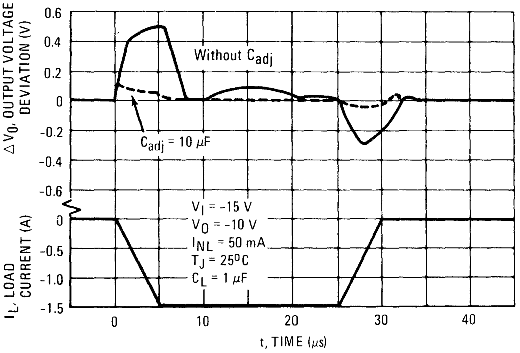


FIGURE 16 – LOAD TRANSIENT RESPONSE



APPLICATIONS INFORMATION

BASIC CIRCUIT OPERATION

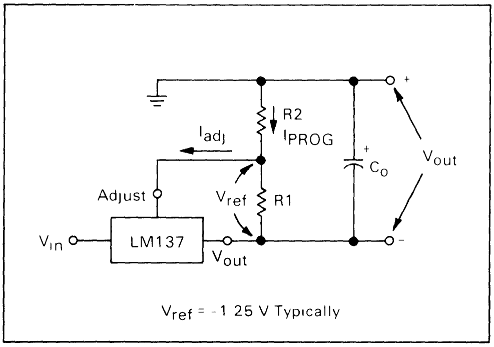
The LM137 is a 3-terminal floating regulator. In operation, the LM137 develops and maintains a nominal -1 25 volt reference (V_{ref}) between its output and adjustment terminals. This reference voltage is converted to a programming current (I_{PROG}) by R1 (see Figure 17), and this constant current flows through R2 from ground. The regulated output voltage is given by:

$$V_{out} = V_{ref} (1 + \frac{R2}{R1}) + I_{adj} R2$$

Since the current into the adjustment terminal (I_{adj}) represents an error term in the equation, the LM137 was designed to control I_{adj} to less than 100 μA and keep it constant. To do this, all quiescent operating current is returned to the output terminal. This imposes the requirement for a minimum load current. If the load current is less than this minimum, the output voltage will increase.

Since the LM137 is a floating regulator, it is only the voltage differential across the circuit that is important to performance, and operation at high voltages with respect to ground is possible

FIGURE 17 – BASIC CIRCUIT CONFIGURATION



LOAD REGULATION

The LM137 is capable of providing extremely good load regulation, but a few precautions are needed to obtain maximum performance. For best performance, the programming resistor (R1) should be connected as close to the regulator as possible to minimize line drops which effectively appear in series with the reference, thereby degrading regulation. The ground end of R2 can be returned near the load ground to provide remote ground sensing and improve load regulation

EXTERNAL CAPACITORS

A 1 μF tantalum input bypass capacitor (C_{in}) is recommended to reduce the sensitivity to input line impedance

The adjustment terminal may be bypassed to ground to improve ripple rejection. This capacitor (C_{adj}) prevents ripple from being amplified as the output voltage is increased A 10 μF capacitor should improve ripple rejection about 15 dB at 120 Hz in a 10 volt application

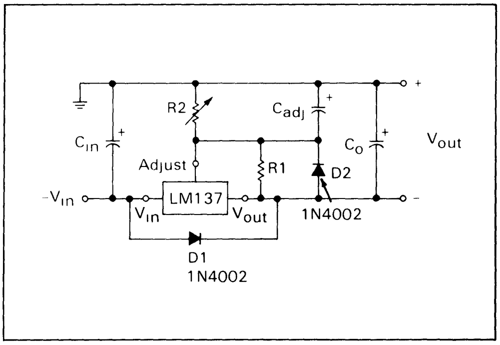
An output capacitor (C_O) in the form of a 1 μF tantalum or 10 μF aluminum electrolytic capacitor is required for stability

PROTECTION DIODES

When external capacitors are used with any I C regulator it is sometimes necessary to add protection diodes to prevent the capacitors from discharging through low current points into the regulator

Figure 18 shows the LM137 with the recommended protection diodes for output voltages in excess of -25 V or high capacitance values ($C_O > 25$ μF, $C_{adj} > 10$ μF) Diode D_1 prevents C_O from discharging thru the I C during an input short circuit Diode D_2 protects against capacitor C_{adj} discharging through the I C during an output short circuit. The combination of diodes D1 and D2 prevents C_{adj} from discharging through the I C. during an input short circuit

FIGURE 18 – VOLTAGE REGULATOR WITH PROTECTION DIODES



MOTOROLA

LM137M
LM237M
LM337M

Specifications and Applications Information

MEDIUM-CURRENT 3-TERMINAL ADJUSTABLE NEGATIVE VOLTAGE REGULATOR

SILICON MONOLITHIC INTEGRATED CIRCUIT

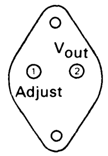
3-TERMINAL ADJUSTABLE OUTPUT NEGATIVE VOLTAGE REGULATOR

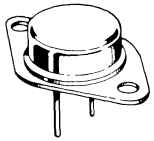
The LM137M/237M/337M are adjustable 3-terminal negative voltage regulators capable of supplying in excess of 500 mA over an output voltage range of -1.2 V to -37 V. These voltage regulators are exceptionally easy to use and require only two external resistors to set the output voltage. Further, they employ internal current limiting, thermal shutdown and safe area compensation, making them essentially blow-out proof.

The LM137M series serve a wide variety of applications including local, on-card regulation. This device can also be used to make a programmable output regulator; or, by connecting a fixed resistor between the adjustment and output, the LM137M series can be used as a precision current regulator.

- Output Current in Excess of 500 mA
- Output Adjustable Between -1.2 V and -37 V
- Internal Thermal Overload Protection
- Internal Short-Circuit-Current Limiting
- Output Transistor Safe-Area Compensation
- Floating Operation for High Voltage Applications
- Standard 3-Lead Transistor Packages
- Eliminates Stocking Many Fixed Voltages

R SUFFIX
METAL PACKAGE
CASE 80
(TO-66 Type)

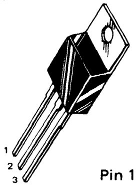




(Bottom View)
Case is input

Pins 1 and 2 electrically isolated from case.
Case is third electrical connection.

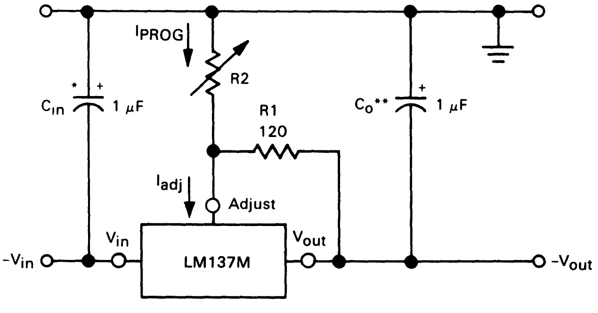
T SUFFIX
PLASTIC PACKAGE
(LM337M only)
CASE 221A
(TO-220)



Pin 1 Adjust
Pin 2 V_{in}
Pin 3 V_{out}

Heatsink surface connected to Pin 2

STANDARD APPLICATION



*C_{in} is required if regulator is located more than 4 inches from power supply filter. A 1 μF solid tantalum or 10 μF aluminum electrolytic is recommended.

**C_O is necessary for stability. A 1 μF solid tantalum or 10 μF aluminum electrolytic is recommended.

$$V_{out} = -1.25\ V\,(1 + \frac{R2}{R1})$$

ORDERING INFORMATION

Device	Temperature Range	Package
LM137MR	T_J = -55°C to +150°C	Metal Power
LM237MR	T_J = -25°C to +150°C	Metal Power
LM337MR	T_J = 0°C to +125°C	Metal Power
LM337MT	T_J = 0°C to +125°C	Plastic Power

MAXIMUM RATINGS

Rating		Symbol	Value	Unit
Input-Output Voltage Differential		V_I-V_O	40	Vdc
Power Dissipation		P_D	Internally Limited	
Operating Junction Temperature Range	LM137M LM237M LM337M	T_J	-55 to +150 -25 to +150 0 to +125	°C
Storage Temperature Range		T_{stg}	-65 to +150	°C

ELECTRICAL CHARACTERISTICS

($|V_I - V_O| = 5.0$ V, $I_O = 0.1$; $T_J = T_{low}$ to T_{high} [see Note 1], P_{max} per Note 2, unless otherwise specified.)

Characteristic	Figure	Symbol	LM137M/237M			LM337M			Unit
			Min	Typ	Max	Min	Typ	Max	
Line Regulation (Note 3) $T_A = 25°C$, $3.0\ V \leqslant \lvert V_I - V_O \rvert \leqslant 40\ V$	1	Reg_{line}	—	0.01	0.02	—	0.01	0.04	%/V
Load Regulation (Note 3), $T_A = 25°C$, $10\ mA \leqslant I_O \leqslant 0.5\ A$ $\lvert V_O \rvert \leqslant 5.0\ V$ $\lvert V_O \rvert \geqslant 5.0\ V$	2	Reg_{load}	— —	15 0.3	25 0.5	— —	15 0.3	50 1.0	mV % V_O
Thermal Regulation 10 mS Pulse, $T_A = 25°C$	—	Reg_{therm}	—	0.002	0.02	—	0.003	0.04	%V_O/W
Adjustment Pin Current	3	I_{adj}	—	65	100	—	65	100	μA
Adjustment Pin Current Change $2.5\ V \leqslant \lvert V_I - V_O \rvert \leqslant 40\ V$, $10\ mA \leqslant I_L \leqslant 0.5\ A$, $P_D \leqslant P_{max}$, $T_A = 25°C$	1,2	ΔI_{adj}	—	2.0	5.0	—	2.0	5.0	μA
Reference Voltage (Note 4) $3.0\ V \leqslant \lvert V_I - V_O \rvert \leqslant 40\ V$, $10\ mA \leqslant I_O \leqslant 0.5\ A$, $P_D \leqslant P_{max}$, $T_A = 25°C$ T_{low} to T_{high}	3	V_{ref}	-1.225 -1.20	-1.250 -1.25	-1.275 -1.30	-1.213 -1.20	-1.250 -1.25	-1.287 -1.30	V
Line Regulation (Note 3) $3.0\ V \leqslant \lvert V_I - V_O \rvert \leqslant 40\ V$	1	Reg_{line}	—	0.02	0.05	—	0.02	0.07	%/V
Load Regulation (Note 3) $10\ mA \leqslant I_O \leqslant 0.5\ A$ $\lvert V_O \rvert \leqslant 5.0\ V$ $\lvert V_O \rvert \geqslant 5.0\ V$	2	Reg_{load}	— —	20 0.3	50 1.0	— —	20 0.3	70 1.5	mV %V_O
Temperature Stability ($T_{low} \leqslant T_J \leqslant T_{high}$)	3	T_S	—	0.6	—	—	0.6	—	%V_O
Minimum Load Current to Maintain Regulation ($\lvert V_I - V_O \rvert \leqslant 10\ V$) ($\lvert V_I - V_O \rvert \leqslant 40\ V$)	3	I_{Lmin}	— —	1.2 2.5	3.0 5.0	— —	1.5 2.5	6.0 10	mA
Maximum Output Current $\lvert V_I - V_O \rvert \leqslant 15\ V$, $P_D \leqslant P_{max}$ $\lvert V_I - V_O \rvert = 40\ V$, $P_D \leqslant P_{max}$, $T_A = 25°C$	3	I_{max}	0.5 0.15	0.9 0.25	— —	0.5 0.1	0.9 0.25	— —	A
RMS Noise, % of V_O $T_A = 25°C$, $10\ Hz \leqslant f \leqslant 10\ kHz$	—	N	—	0.003	—	—	0.003	—	%V_O
Ripple Rejection, $V_O = -10\ V$, $f = 120\ Hz$ (Note 5) Without C_{adj} $C_{adj} = 10\ \mu F$	4	RR	— 66	60 77	— —	— 66	60 77	— —	dB
Long Term Stability, $T_J = T_{high}$ (Note 6) $T_A = 25°C$ for Endpoint Measurements	3	S	—	0.3	1.0	—	0.3	1.0	%/1.0 k Hrs.
Thermal Resistance Junction to Case R Package (TO-66) T Package (TO-220)	—	$R_{\theta JC}$	— —	7.0 —	— —	— —	7.0 7.0	— —	°C/W

NOTES:

(1) T_{low} = -55°C for LM137M T_{high} = +150°C for LM137M
= -25°C for LM237M = +150°C for LM237M
= 0°C for LM337M = +125°C for LM337M

(2) P_{max} = 7.5 W

(3) Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

(4) Selected devices with tightened tolerance reference voltage available.

(5) C_{adj}, when used, is connected between the adjustment pin and ground.

(6) Since Long Term Stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot.

SCHEMATIC DIAGRAM

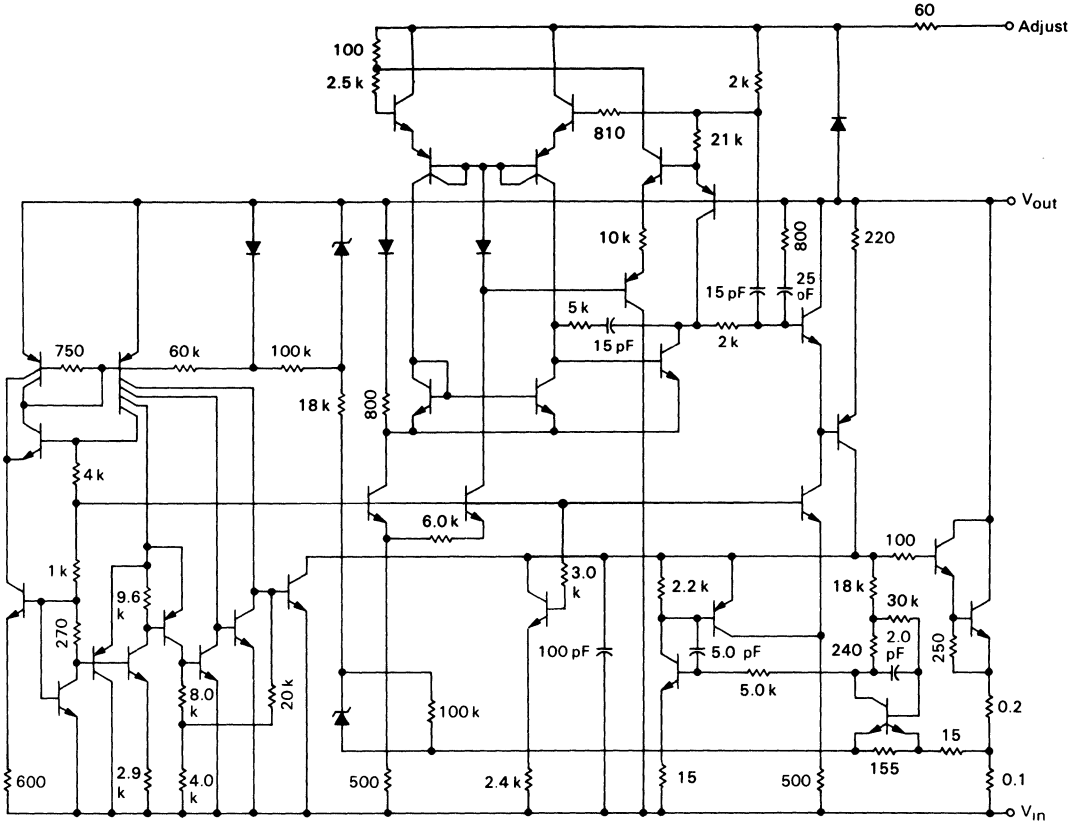
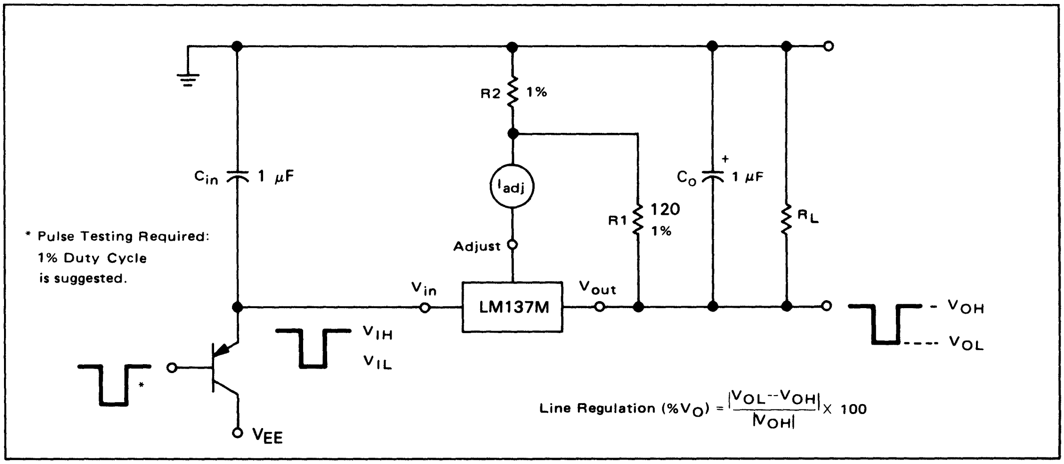


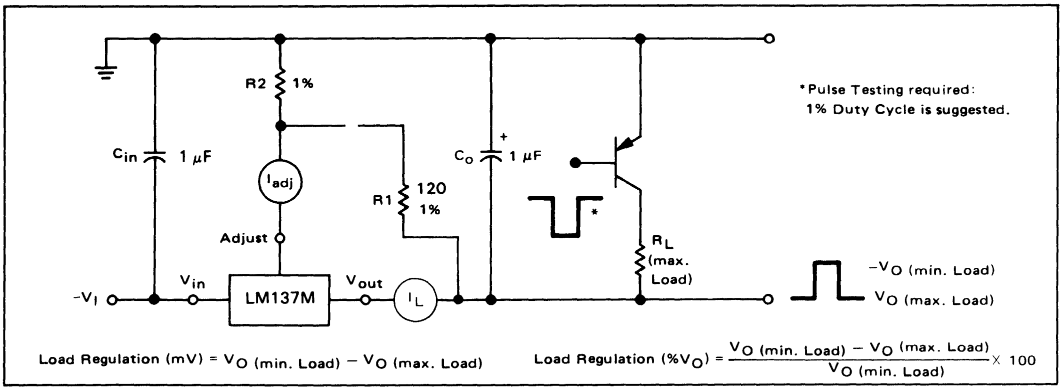
FIGURE 1 – LINE REGULATION AND ΔI_{adj}/LINE TEST CIRCUIT



* Pulse Testing Required:
1% Duty Cycle
is suggested.

Line Regulation $(\%V_O) = \frac{|V_{OL} - V_{OH}|}{|V_{OH}|} \times 100$

FIGURE 2 – LOAD REGULATION AND ΔI_{adj}/LOAD TEST CIRCUIT

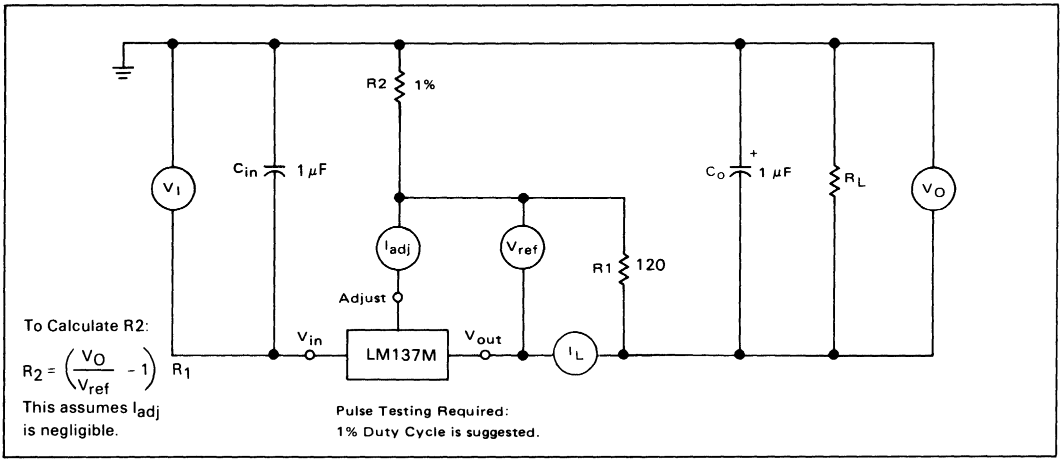


*Pulse Testing required:
1% Duty Cycle is suggested.

Load Regulation (mV) = V_O (min. Load) – V_O (max. Load)

$$\text{Load Regulation }(\%V_O) = \frac{V_O\text{ (min. Load)} - V_O\text{ (max. Load)}}{V_O\text{ (min. Load)}} \times 100$$

FIGURE 3 – STANDARD TEST CIRCUIT



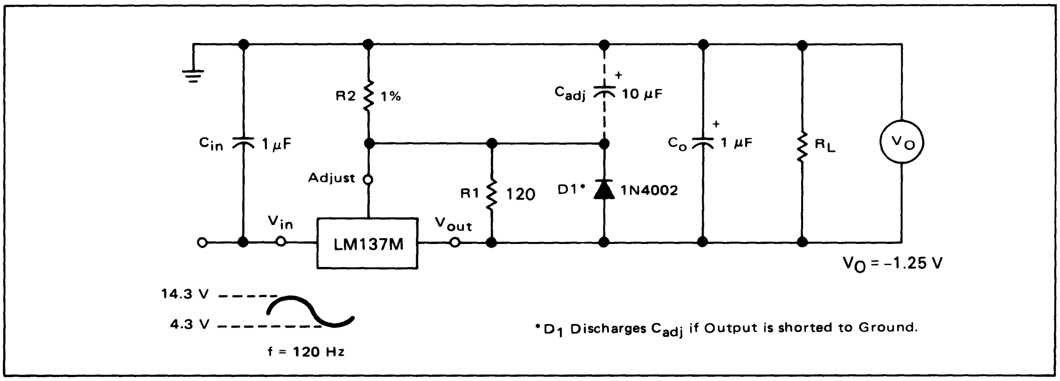
To Calculate R2:

$$R_2 = \left(\frac{V_O}{V_{ref}} - 1\right) R_1$$

This assumes I_{adj} is negligible.

Pulse Testing Required:
1% Duty Cycle is suggested.

FIGURE 4 – RIPPLE REJECTION TEST CIRCUIT



*D_1 Discharges C_{adj} if Output is shorted to Ground.

FIGURE 5 – LOAD REGULATION

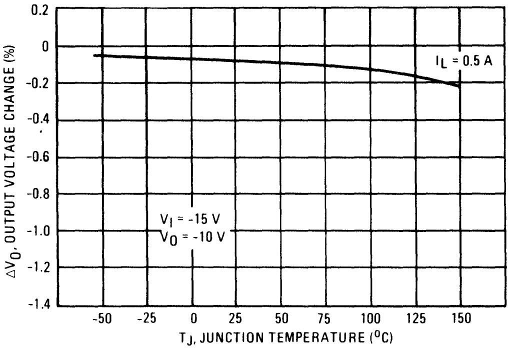


FIGURE 6 – CURRENT LIMIT

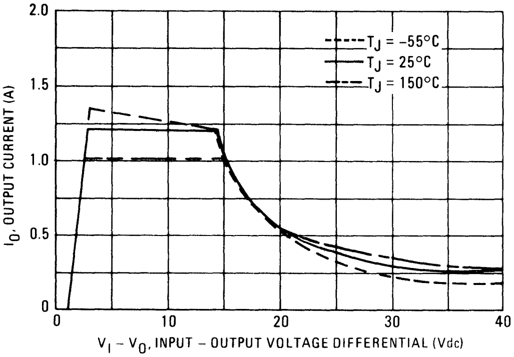


FIGURE 7 – ADJUSTMENT PIN CURRENT

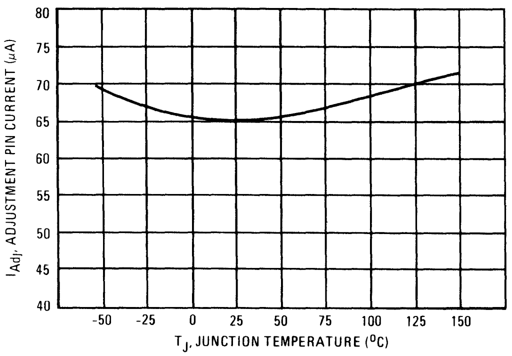


FIGURE 8 – DROPOUT VOLTAGE

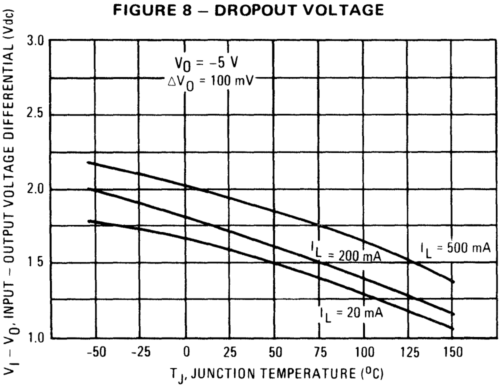


FIGURE 9 – TEMPERATURE STABILITY

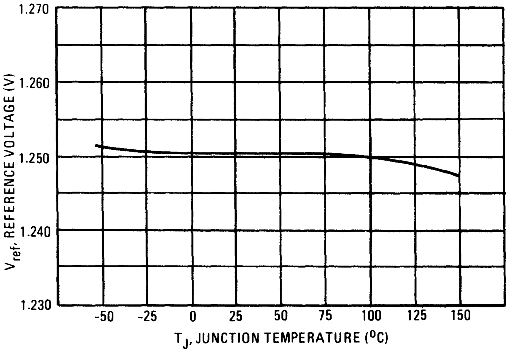


FIGURE 10 – MINIMUM OPERATING CURRENT

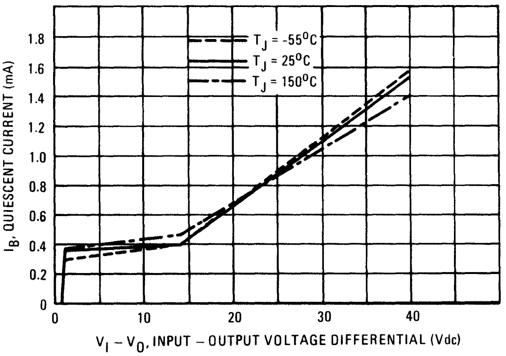


FIGURE 11 – RIPPLE REJECTION VS OUTPUT VOLTAGE

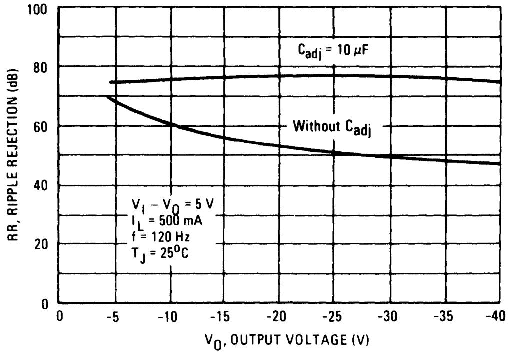


FIGURE 12 – RIPPLE REJECTION VS. OUTPUT CURRENT

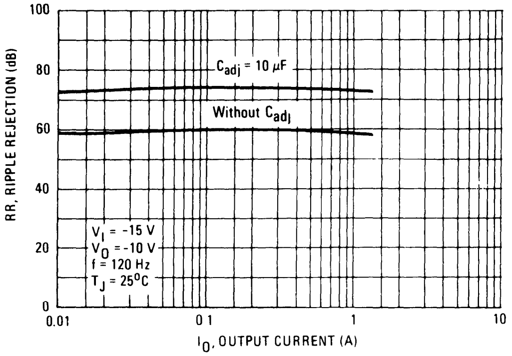


FIGURE 13 – RIPPLE REJECTION VS. FREQUENCY

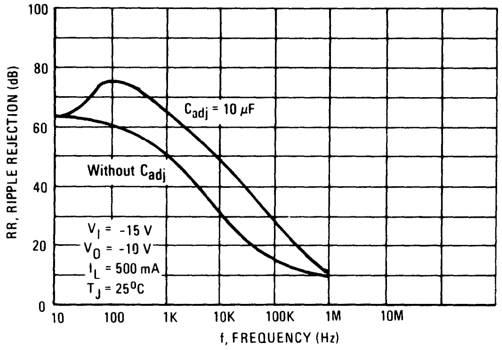


FIGURE 14 – OUTPUT IMPEDANCE

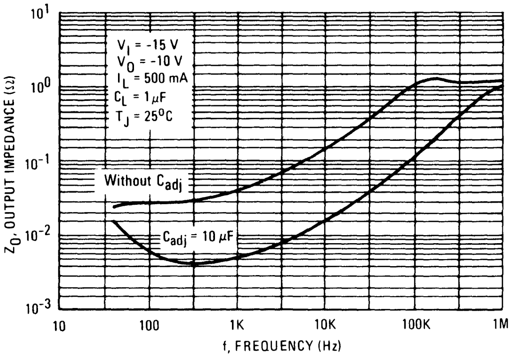


FIGURE 15 – LINE TRANSIENT RESPONSE

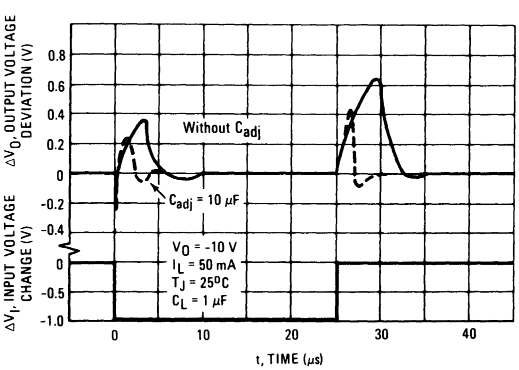
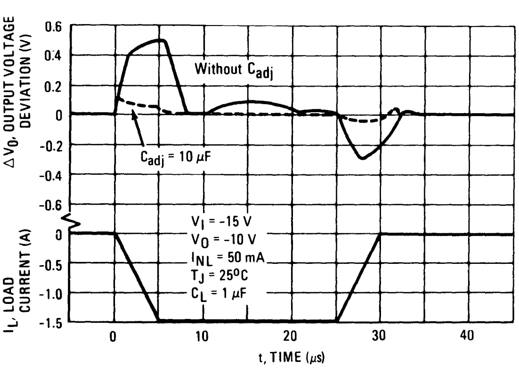


FIGURE 16 – LOAD TRANSIENT RESPONSE



APPLICATIONS INFORMATION

BASIC CIRCUIT OPERATION

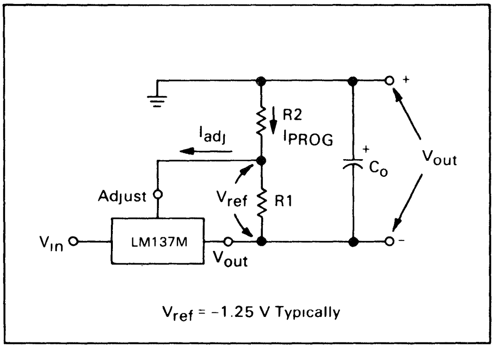
The LM137M is a 3-terminal floating regulator. In operation, the LM137M develops and maintains a nominal -1.25 volt reference (V_{ref}) between its output and adjustment terminals. This reference voltage is converted to a programming current (I_{PROG}) by R1 (see Figure 17), and this constant current flows through R2 from ground. The regulated output voltage is given by:

$$V_{out} = V_{ref}\,(1 + \frac{R2}{R1}) + I_{adj}R2$$

Since the current into the adjustment terminal (I_{adj}) represents an error term in the equation, the LM137M was designed to control I_{adj} to less than 100 μA and keep it constant. To do this, all quiescent operating current is returned to the output terminal. This imposes the requirement for a minimum load current. If the load current is less than this minimum, the output voltage will increase.

Since the LM137M is a floating regulator, it is only the voltage differential across the circuit that is important to performance, and operation at high voltages with respect to ground is possible.

FIGURE 17 – BASIC CIRCUIT CONFIGURATION



V_{ref} = -1.25 V Typically

LOAD REGULATION

The LM137M is capable of providing extremely good load regulation, but a few precautions are needed to obtain maximum performance. For best performance, the programming resistor (R1) should be connected as close to the regulator as possible to minimize line drops which effectively appear in series with the reference, thereby degrading regulation. The ground end of R2 can be returned near the load ground to provide remote ground sensing and improve load regulation.

EXTERNAL CAPACITORS

A 1 μF tantalum input bypass capacitor (C_{in}) is recommended to reduce the sensitivity to input line impedance.

The adjustment terminal may be bypassed to ground to improve ripple rejection. This capacitor (C_{adj}) prevents ripple from being amplified as the output voltage is increased. A 10 μF capacitor should improve ripple rejection about 15 dB at 120 Hz in a 10 volt application.

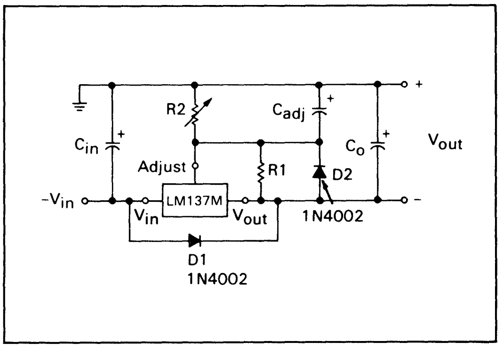
An output capacitor (C_o) in the form of a 1 μF tantalum or 10 μF aluminum electrolytic capacitor is required for stability.

PROTECTION DIODES

When external capacitors are used with any I.C. regulator it is sometimes necessary to add protection diodes to prevent the capacitors from discharging through low current points into the regulator.

Figure 18 shows the LM137M with the recommended protection diodes for output voltages in excess of -25 V or high capacitance values ($C_o > 25$ μF, $C_{adj} > 10$ μF). Diode D_1 prevents C_o from discharging thru the I.C. during an input short circuit. Diode D_2 protects against capacitor C_{adj} discharging through the I.C. during an output short circuit. The combination of diodes D1 and D2 prevents C_{adj} from discharging through the I.C. during an input short circuit.

FIGURE 18 – VOLTAGE REGULATOR WITH PROTECTION DIODES



MOTOROLA

LM140 series
LM340 series

3-TERMINAL POSITIVE VOLTAGE REGULATORS

The LM140/340 series of three-terminal positive voltage regulators are monolithic integrated circuits designed for a wide variety of applications including local on-board regulation. Available in seven fixed output voltage options from 5.0 to 24 volts, these regulators employ internal current limiting, thermal shutdown, and safe area compensation — making them virtually blowout proof. The LM140/340 series is guaranteed to have line and load regulation that is a factor of two better than the 7800 series. Although the LM140/340 series was designed primarily as a fixed regulator, it can be used with external components to obtain adjustable voltages.

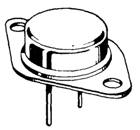
- Output Currents in Excess of 1.0 A
- Internal Thermal Overload Protection
- Internal Short Circuit Limiting
- Output Transistor Safe-Area Compensation
- No External Components Required
- Available in Both Commercial and Military Temperature Ranges

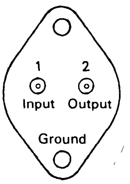
ORDERING INFORMATION

Device	Voltage	Temperature Range (T_A)
LM140K-5.0	5.0 Volts	-55 to +125°C
LM140K-6.0	6.0 Volts	-55 to +125°C
LM140K-8.0	8.0 Volts	-55 to +125°C
LM140K-12	12 Volts	-55 to +125°C
LM140K-15	15 Volts	-55 to +125°C
LM140K-18	18 Volts	-55 to +125°C
LM140K-24	24 Volts	-55 to +125°C
LM340K-5.0	5.0 Volts	0 to +70°C
LM340K-6.0	6.0 Volts	0 to +70°C
LM340K-8.0	8.0 Volts	0 to +70°C
LM340K-12	12 Volts	0 to +70°C
LM340K-15	15 Volts	0 to +70°C
LM340K-18	18 Volts	0 to +70°C
LM340K-24	24 Volts	0 to +70°C

THREE-TERMINAL POSITIVE FIXED VOLTAGE REGULATORS

K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 TYPE)

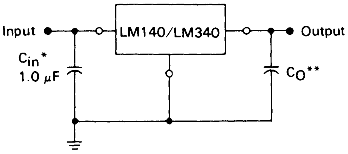




(bottom view)

Pins 1 and 2 electrically isolated from case. Case is third electrical connection.

STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.0 V above the output voltage even during the low point on the input ripple voltage.

* = C_{in} (solid tantalum) is required, if regulator is located an appreciable distance from power supply filter.

** = C_O is not needed for stability; however, it does improve transient response. If needed, its value should be greater than 0.1 μF.

LM140 series/LM340 series **MAXIMUM RATINGS** (T_A = +25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Input Voltage	V_{in}		Vdc
(5.0 V - 18 V)		35	
(24 V)		40	
Power Dissipation and Thermal Characteristics			
(Metal Package)			
T_A = +25°C	P_D	Internally Limited	Watts
Derate above T_A = +25°C	$1/R_{\theta JA}$	22.5	mW/°C
Thermal Resistance, Junction to Air	$R_{\theta JA}$	45	°C/W
T_C = +25°C	P_D	Internally Limited	Watts
Derate above T_C = +65°C (See Figure 2)	$1/R_{\theta JC}$	182	mW/°C
Thermal Resistance, Junction to Case	$R_{\theta JC}$	5.5	°C/W
Storage Junction Temperature Range	T_{stg}	-65 to +150	°C
Operating Junction Temperature Range	T_J		°C
LM140		-55 to +150	
LM340		0 to +125	

NOTES:

1. T_{low} = -55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 5.0 ELECTRICAL CHARACTERISTICS

(V_{in} = 10 V, I_O = 500 mA, T_J = T_{low} to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	4.8	5.0	5.2	Vdc
Input Regulation (Note 2) 8.0 to 20 Vdc 7.0 to 25 Vdc (T_J = +25°C) 8.0 to 12 Vdc, I_O = 1.0 A 7.3 to 20 Vdc, I_O = 1.0 A (T_J = +25°C)	Reg_{in}	 — — — —	 — — — —	 50 50 25 50	mV
Load Regulation (Note 2) 5.0 mA $\leqslant I_O \leqslant$ 1.0 A 5.0 mA $\leqslant I_O \leqslant$ 1.5 A (T_J = +25°C) 250 mA $\leqslant I_O \leqslant$ 750 mA (T_J = +25°C)	Reg_{load}	 — — —	 — — —	 50 50 25	mV
Output Voltage LM140 8.0 $\leqslant V_{in} \leqslant$ 20 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W LM340 7.0 $\leqslant V_{in} \leqslant$ 20 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W	V_O	 4.75 4.75	 5.0 5.0	 5.25 5.25	Vdc
Quiescent Current I_O = 1.0 A LM140 LM340 LM140 (T_J = +25°C) LM340 (T_J = +25°C)	I_b	 — — — —	 4.0 4.0 4.0 4.0	 7.0 8.5 6.0 8.0	mA
Quiescent Current Change 8.0 $\leqslant V_{in} \leqslant$ 25 Vdc LM140 7.0 $\leqslant V_{in} \leqslant$ 25 Vdc LM340 5.0 mA $\leqslant I_O \leqslant$ 1.0 A LM140, LM340 8.0 $\leqslant V_{in} \leqslant$ 20 Vdc, I_O = 1.0 A LM140 7.5 $\leqslant V_{in} \leqslant$ 20 Vdc, I_O = 1.0 A LM340	ΔI_b	 — — — — —	 — — — — —	 0.8 1.0 0.5 0.8 1.0	mA
Ripple Rejection LM140 LM340 I_O = 1.0 A (T_J = +25°C) LM140 LM340	RR	 68 62 68 62	 80 80 — —	 — — — —	dB
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	30	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	2.0	—	A
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	40	—	µV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±0.6	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		7.3	—	—	Vdc

NOTES:

1. T_{low} = -55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 6.0 ELECTRICAL CHARACTERISTICS

(V_{in} = 11 V, I_O = 500 mA, T_J = T_{low} to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	5.75	6.0	6.25	Vdc
Input Regulation (Note 2)	Reg_{in}				mV
9.0 to 21 Vdc		—	—	60	
8.0 to 25 Vdc (T_J = +25°C)		—	—	60	
9.0 to 13 Vdc, I_O = 1.0 A		—	—	30	
8.3 to 21 Vdc, I_O = 1.0 A (T_J = +25°C)		—	—	60	
Load Regulation (Note 2)	Reg_{load}				mV
5.0 mA $\leq I_O \leq$ 1.0 A		—	—	60	
5.0 mA $\leq I_O \leq$ 1.5 A (T_J = +25°C)		—	—	60	
250 mA $\leq I_O \leq$ 750 mA (T_J = +25°C)		—	—	30	
Output Voltage	V_O				Vdc
LM140 9.0 $\leq V_{in} \leq$ 21 Vdc, 5.0 mA $\leq I_O \leq$ 1.0 A, $P_O \leq$ 15 W		5.7	6.0	6.3	
LM340 8.0 $\leq V_{in} \leq$ 21 Vdc, 6.0 mA $\leq I_O \leq$ 1.0 A, $P_O \leq$ 15 W		5.7	5.0	6.3	
Quiescent Current I_O = 1.0 A	I_b				mA
LM140		—	4.0	7.0	
LM340		—	4.0	8.5	
LM140 (T_J = +25°C)		—	4.0	6.0	
LM340 (T_J = +25°C)		—	4.0	8.0	
Quiescent Current Change	ΔI_b				mA
9.0 $\leq V_{in} \leq$ 25 Vdc LM140		—	—	0.8	
8.0 $\leq V_{in} \leq$ 25 Vdc LM340		—	—	1.0	
5.0 mA $\leq I_O \leq$ 1.0 A LM140, LM340		—	—	0.5	
9.0 $\leq V_{in} \leq$ 21 Vdc, I_O = 1.0 A LM140		—	—	0.8	
8.6 $\leq V_{in} \leq$ 21 Vdc, I_O = 1.0 A LM340		—	—	1.0	
Ripple Rejection	RR				dB
LM140		65	78	—	
LM340		59	78	—	
I_O = 1.0 A (T_J = +25°C) LM140		65	—	—	
LM340		59	—	—	
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	35	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	1.9	—	A
Output Noise Voltage (T_A = +25°C) 10 Hz $\leq f \leq$ 100 kHz	V_n	—	45	—	µV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±0.7	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		8.3	—	—	Vdc

NOTES:

1. T_{low} = −55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 8.0 ELECTRICAL CHARACTERISTICS

(V_{in} = 14 V, I_O = 500 mA, T_J = T_{low} to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	7.7	8.0	8.3	Vdc
Input Regulation (Note 2) 11 to 23 Vdc 10.5 to 25 Vdc (T_J = +25°C) 11 to 17 Vdc, I_O= 1.0 A 10.5 to 23 Vdc, I_O = 1.0 A (T_J = +25°C)	Reg_{in}	 — — — —	 — — — —	 80 80 40 80	mV
Load Regulation (Note 2) 5.0 mA $\leqslant I_O \leqslant$ 1.0 A 5.0 mA $\leqslant I_O \leqslant$ 1.5 A (T_J = +25°C) 250 mA $\leqslant I_O \leqslant$ 750 mA (T_J = +25°C)	Reg_{load}	 — — —	 — — —	 80 80 40	mV
Output Voltage LM140 11.5 $\leqslant V_{in} \leqslant$ 23 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W LM340 10.5 $\leqslant V_{in} \leqslant$ 23 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W	V_O	 7.6 7.6	 8.0 8.0	 8.4 8.4	Vdc
Quiescent Current I_O = 1.0 A LM140 LM340 LM140 (T_J = +25°C) LM340 (T_J = +25°C)	I_b	 — — — —	 4.0 4.0 4.0 4.0	 7.0 8.5 6.0 8.0	mA
Quiescent Current Change 11.5 $\leqslant V_{in} \leqslant$ 25 Vdc LM140 10.5 $\leqslant V_{in} \leqslant$ 25 Vdc LM340 5.0 mA $\leqslant I_O \leqslant$ 1.0 A LM140, LM340 11.5 $\leqslant V_{in} \leqslant$ 23 Vdc, I_O = 1.0 A LM140 10.6 $\leqslant V_{in} \leqslant$ 23 Vdc, I_O = 1.0 A LM340	ΔI_b	 — — — — —	 — — — — —	 0.8 1.0 0.5 0.8 1.0	mA
Ripple Rejection LM140 LM340 I_O = 1.0 A (T_J = +25°C) LM140 LM340	RR	 62 56 62 56	 76 76 — —	 — — — —	dB
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	40	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	1.5	—	A
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant f \leqslant$ 100 kHz	V_n	—	52	—	µV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±1.0	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		10.5	—	—	Vdc

NOTES:

1. T_{low} = -55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 12 ELECTRICAL CHARACTERISTICS

(V_{in} = 19 V, I_O = 500 mA, T_J = T_{low} to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	11.5	12	12.5	Vdc
Input Regulation (Note 2) 15 to 27 Vdc 14.6 to 30 Vdc (T_J = +25°C) 16 to 22 Vdc, I_O= 1.0 A 14.6 to 27 Vdc, I_O = 1.0 A (T_J = +25°C)	Reg_{in}	 — — — —	 — — — —	 120 120 60 120	mV
Load Regulation (Note 2) 5.0 mA $\leqslant I_O \leqslant$ 1.0 A 5.0 mA $\leqslant I_O \leqslant$ 1.5 A (T_J = +25°C) 250 mA $\leqslant I_O \leqslant$ 750 mA (T_J = +25°C)	Reg_{load}	 — — —	 — — —	 120 120 60	mV
Output Voltage LM140 15.5 $\leqslant V_{in} \leqslant$ 27 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W LM340 14.5 $\leqslant V_{in} \leqslant$ 27 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W	V_O	 11.4 11.4	 12 12	 12.6 12.6	Vdc
Quiescent Current I_O = 1.0 A LM140 LM340 LM140 (T_J = +25°C) LM340 (T_J = +25°C)	I_b	 — — — —	 4.0 4.0 4.0 4.0	 7.0 8.5 6.0 8.0	mA
Quiescent Current Change 15 $\leqslant V_{in} \leqslant$ 30 Vdc LM140 14.5 $\leqslant V_{in} \leqslant$ 30 Vdc LM340 5.0 mA $\leqslant I_O \leqslant$ 1.0 A LM140, LM340 15 $\leqslant V_{in} \leqslant$ 27 Vdc, I_O = 1.0 A LM140 14.8 $\leqslant V_{in} \leqslant$ 27 Vdc, I_O = 1.0 A LM340	ΔI_b	 — — — — —	 — — — — —	 0.8 1.0 0.5 0.8 1.0	mA
Ripple Rejection LM140 LM340 I_O = 1.0 A (T_J = +25°C) LM140 LM340	RR	 61 55 61 55	 72 72 — —	 — — — —	dB
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	75	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	1.1	—	A
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant f \leqslant$ 100 kHz	V_n	—	75	—	µV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±1.5	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		14.6	—	—	Vdc

NOTES:

1. T_{low} = -55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 15 ELECTRICAL CHARACTERISTICS

(V_{in} = 23 V, I_O = 500 mA, $T_J = T_{low}$ to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	14.4	15	15.6	Vdc
Input Regulation (Note 2)	Reg_{in}				mV
18.5 to 30 Vdc		—	—	150	
17.5 to 30 Vdc (T_J = +25°C)		—	—	150	
20 to 26 Vdc, I_O= 1.0 A		—	—	75	
17.7 to 30 Vdc, I_O = 1.0 A (T_J = +25°C)		—	—	150	
Load Regulation (Note 2)	Reg_{load}				mV
5.0 mA $\leqslant I_O \leqslant$ 1.0 A		—	—	150	
5.0 mA $\leqslant I_O \leqslant$ 1.5 A (T_J = +25°C)		—	—	150	
250 mA $\leqslant I_O \leqslant$ 750 mA (T_J = +25°C)		—	—	75	
Output Voltage	V_O				Vdc
LM140 18.5 $\leqslant V_{in} \leqslant$ 30 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W		14.25	15	15.75	
LM340 17.5 $\leqslant V_{in} \leqslant$ 30 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W		14.25	15	15.75	
Quiescent Current I_O = 1.0 A	I_b				mA
LM140		—	4.0	7.0	
LM340		—	4.0	8.5	
LM140 (T_J = +25°C)		—	4.0	6.0	
LM340 (T_J = +25°C)		—	4.0	8.0	
Quiescent Current Change	ΔI_b				mA
18.5 $\leqslant V_{in} \leqslant$ 30 Vdc LM140		—	—	0.8	
17.5 $\leqslant V_{in} \leqslant$ 30 Vdc LM340		—	—	1.0	
5.0 mA $\leqslant I_O \leqslant$ 1.0 A LM140, LM340		—	—	0.5	
18.5 $\leqslant V_{in} \leqslant$ 30 Vdc, I_O = 1.0 A LM140		—	—	0.8	
17.9 $\leqslant V_{in} \leqslant$ 30 Vdc, I_O = 1.0 A LM340		—	—	1.0	
Ripple Rejection	RR				dB
LM140		60	70	—	
LM340		54	70	—	
I_O = 1.0 A (T_J = +25°C) LM140		60	—	—	
LM340		54	—	—	
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	95	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	800	—	mA
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	90	—	μV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±1.8	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		17.7	—	—	Vdc

NOTES:

1. T_{low} = −55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 18 ELECTRICAL CHARACTERISTICS

(V_{in} = 27 V, I_O = 500 mA, T_J = T_{low} to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	17.3	18	18.7	Vdc
Input Regulation (Note 2)	Reg_{in}				mV
21.5 to 33 Vdc		—	—	180	
21 to 33 Vdc (T_J = +25°C)		—	—	180	
24 to 30 Vdc, I_O = 1.0 A		—	—	90	
21 to 33 Vdc, I_O = 1.0 A (T_J = +25°C)		—	—	180	
Load Regulation (Note 2)	Reg_{load}				mV
5.0 mA $\leqslant I_O \leqslant$ 1.0 A		—	—	180	
5.0 mA $\leqslant I_O \leqslant$ 1.5 A (T_J = +25°C)		—	—	180	
250 mA $\leqslant I_O \leqslant$ 750 mA (T_J = +25°C)		—	—	90	
Output Voltage	V_O				Vdc
LM140 22 $\leqslant V_{in} \leqslant$ 33 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W		17.1	18	18.9	
LM340 21 $\leqslant V_{in} \leqslant$ 33 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W		17.1	18	18.9	
Quiescent Current I_O = 1.0 A	I_b				mA
LM140		—	4.0	7.0	
LM340		—	4.0	8.5	
LM140 (T_J = +25°C)		—	4.0	6.0	
LM340 (T_J = +25°C)		—	4.0	8.0	
Quiescent Current Change	ΔI_b				mA
22 $\leqslant V_{in} \leqslant$ 33 Vdc LM140		—	—	0.8	
21 $\leqslant V_{in} \leqslant$ 33 Vdc LM340		—	—	1.0	
5.0 mA $\leqslant I_O \leqslant$ 1.0 A LM140, LM340		—	—	0.5	
22 $\leqslant V_{in} \leqslant$ 33 Vdc, I_O = 1.0 A LM140		—	—	0.8	
21 $\leqslant V_{in} \leqslant$ 33 Vdc, I_O = 1.0 A LM340		—	—	1.0	
Ripple Rejection	RR				dB
LM140		59	69	—	
LM340		53	69	—	
I_O = 1.0 A (T_J = +25°C)					
LM140		59	—	—	
LM340		53	—	—	
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	110	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	500	—	mA
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	110	—	μV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±2.3	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		21	—	—	Vdc

NOTES:

1. T_{low} = −55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

LM140/340 — 24 ELECTRICAL CHARACTERISTICS

(V_{in} = 33 V, I_O = 500 mA, $T_J = T_{low}$ to T_{high} (Note 1), unless otherwise noted).

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C) I_O = 5.0 mA to 1.0 A	V_O	23	24	25	Vdc
Input Regulation (Note 2)	Reg_{in}				mV
28 to 38 Vdc		—	—	240	
27 to 38 Vdc (T_J = +25°C)		—	—	240	
30 to 36 Vdc, I_O = 1.0 A		—	—	120	
27.1 to 38 Vdc, I_O = 1.0 A (T_J = +25°C)		—	—	240	
Load Regulation (Note 2)	Reg_{load}				mV
5.0 mA $\leq I_O \leq$ 1.0 A		—	—	240	
5.0 mA $\leq I_O \leq$ 1.5 A (T_J = +25°C)		—	—	240	
250 mA $\leq I_O \leq$ 750 mA (T_J = +25°C)		—	—	120	
Output Voltage	V_O				Vdc
LM140 28 $\leq V_{in} \leq$ 38 Vdc 5.0 mA $\leq I_O \leq$ 1.0 A, $P_O \leq$ 15 W		22.8	24	25.2	
LM340 27 $\leq V_{in} \leq$ 38 Vdc, 5.0 mA $\leq I_O \leq$ 1.0 A, $P_O \leq$ 15 W		22.8	24	25.2	
Quiescent Current I_O = 1.0 A	I_b				mA
LM140		—	4.0	7.0	
LM340		—	4.0	8.5	
LM140 (T_J = +25°C)		—	4.0	6.0	
LM340 (T_J = +25°C)		—	4.0	8.0	
Quiescent Current Change	ΔI_b				mA
28 $\leq V_{in} \leq$ 38 Vdc LM140		—	—	0.8	
27 $\leq V_{in} \leq$ 38 Vdc LM340		—	—	1.0	
5.0 mA $\leq I_O \leq$ 1.0 A LM140, LM340		—	—	0.5	
28 $\leq V_{in} \leq$ 38 Vdc, I_O = 1.0 A LM140		—	—	0.8	
27.3 $\leq V_{in} \leq$ 38 Vdc, I_O = 1.0 A LM340		—	—	1.0	
Ripple Rejection	RR				dB
LM140		56	66	—	
LM340		50	66	—	
I_O = 1.0 A (T_J = +25°C)					
LM140		56	—	—	
LM340		50	—	—	
Dropout Voltage	$V_{in} - V_O$	—	2.0	—	Vdc
Output Resistance	R_O	—	150	—	mΩ
Short-Circuit Current Limit	I_{sc}	—	200	—	mA
Output Noise Voltage (T_A = +25°C) 10 Hz $\leq f \leq$ 100 kHz	V_n	—	170	—	μV
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA	TCV_O	—	±3.0	—	mV/°C
Peak Output Current (T_J = +25°C)	I_O	—	2.4	—	A
Input Voltage to Maintain Line Regulation (T_J = +25°C) I_O = 1.0 A		27.1	—	—	Vdc

NOTES:

1. T_{low} = −55°C for LM140 T_{high} = +150°C for LM140
   = 0°C for LM340 = +125°C for LM340
2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

FIGURE 1 — WORST CASE POWER DISSIPATION versus AMBIENT TEMPERATURE

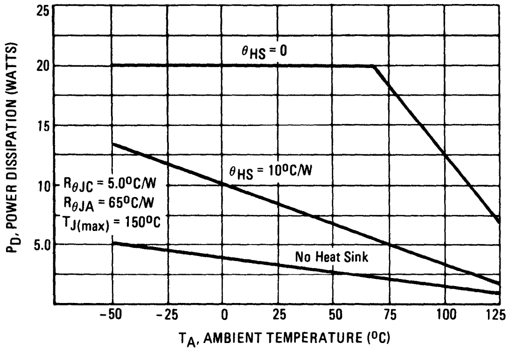


FIGURE 2 — DROPOUT CHARACTERISTICS

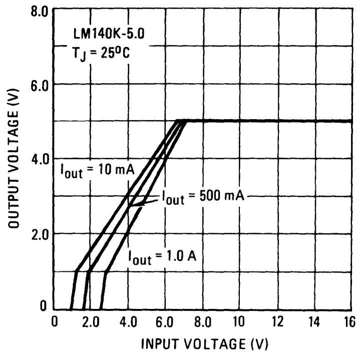


FIGURE 3 — INPUT-OUTPUT DIFFERENTIAL AS A FUNCTION OF JUNCTION TEMPERATURE

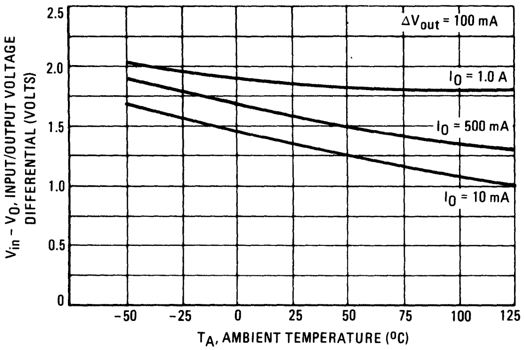


FIGURE 4 — PEAK OUTPUT CURRENT AS A FUNCTION OF INPUT-OUTPUT DIFFERENTIAL VOLTAGE

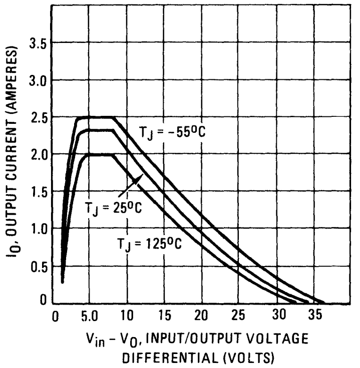


FIGURE 5 — RIPPLE REJECTION AS A FUNCTION OF FREQUENCY

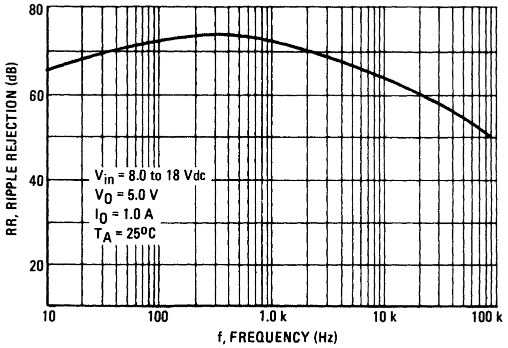
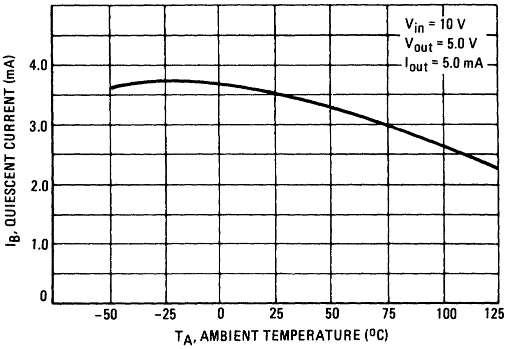


FIGURE 6 — QUIESCENT CURRENT AS A FUNCTION OF TEMPERATURE



MOTOROLA

LM150
LM250
LM350

Advance Information

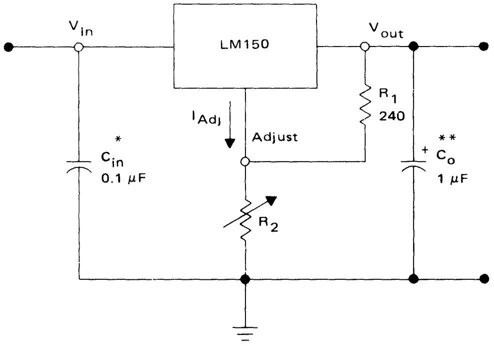
3-TERMINAL ADJUSTABLE OUTPUT POSITIVE VOLTAGE REGULATOR

The LM150/250/350 are adjustable 3-terminal positive voltage regulators capable of supplying in excess of 3.0 A over an output voltage range of 1.2 V to 33 V. These voltage regulators are exceptionally easy to use and require only two external resistors to set the output voltage. Further, they employ internal current limiting, thermal shutdown and safe area compensation, making them essentially blow-out proof.

The LM150 series serve a wide variety of applications including local, on card regulation. This device also makes an especially simple adjustable switching regulator, a programmable output regulator, or by connecting a fixed resistor between the adjustment and output, the LM150 series can be used as a precision current regulator

- Guaranteed 3.0 Amps Output Current
- Output Adjustable between 1.2 V and 33 V
- Load Regulation Typically 0.1%
- Line Regulation Typically 0 005%/V
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting Constant with Temperature
- Output Transistor Safe-area Compensation
- Floating Operation for High Voltage Applications
- Standard 3-lead Transistor Packages
- Eliminates Stocking Many Fixed Voltages

STANDARD APPLICATION



* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_o is not needed for stability, however it does improve transient response.

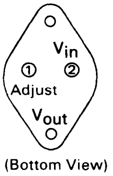
$$V_{out} = 1.25\ V\ (1 + \frac{R_2}{R_1}) + I_{Adj}\ R_2$$

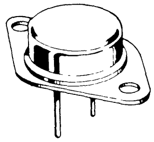
Since I_{Adj} is controlled to less than 100 µA, the error associated with this term is negligible in most applications

3-TERMINAL ADJUSTABLE POSITIVE VOLTAGE REGULATOR

SILICON MONOLITHIC
INTEGRATED CIRCUIT

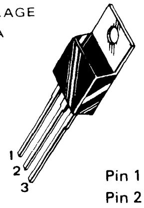
K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 Type)





Pins 1 and 2 electrically isolated from case. Case is third electrical connection.

T SUFFIX
PLASTIC PACKAGE
CASE 221A
(TO-220)



Pin 1 Adjust
Pin 2 V_{out}
Pin 3 V_{in}

Heatsink surface connected to Pin 2

ORDERING INFORMATION

Device	Temperature Range	Package
LM150K	T_J = -55°C to +150°C	Metal Power
LM250K	T_J = -25°C to +150°C	Metal Power
LM350K	T_J = 0°C to +125°C	Metal Power
LM350T	T_J = 0°C to +125°C	Plastic Power

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Input-Output Voltage Differential	V_I-V_O	35	Vdc
Power Dissipation	P_D	Internally Limited	
Operating Junction Temperature Range LM150 LM250 LM350	T_J	-55 to +150 -25 to +150 0 to +125	°C
Storage Temperature Range	T_{stg}	-65 to +150	°C
Soldering Lead Temperature (10 seconds)		300	°C

ELECTRICAL CHARACTERISTICS (Unless otherwise specified, V_I-V_O = 5 V; I_L = 1.5 A; T_J = T_{low} to T_{high} [see Note 1]; P_{max} = 30 W)

			LM150/250			LM350			
Characteristic	Figure	Symbol	Min	Typ	Max	Min	Typ	Max	Unit
Line Regulation (Note 2) T_A = 25°C, 3 V ≤ V_I - V_O ≤ 35 V	1	Reg_{line}	—	0.005	0.01	—	0.005	0.03	%/V
Load Regulation (Note 2) T_A = 25°C, 10 mA ≤ I_L ≤ 3A	2	Reg_{load}							
V_O ≤ 5V			—	5	15	—	5	25	mV
V_O ≥ 5 V			—	0.1	0.3	—	0.1	0.5	% V_O
Thermal Regulation Pulse = 20 ms	—	Reg_{therm}	—	0.002	—	—	0.002	—	%/W
Adjustment Pin Current	3	I_{Adj}	—	50	100	—	50	100	μA
Adjustment Pin Current Change 3 V ≤ V_I-V_O ≤ 35 V 10 mA ≤ I_L ≤ 3 A, P_D ≤ P_{max}	1,2	ΔI_{Adj}	—	0.2	5	—	0.2	5	μA
Reference Voltage (Note 3) 3 V ≤ V_I-V_O ≤ 35 V 10 mA ≤ I_L ≤ 3 A, P_D ≤ P_{max}	3	V_{ref}	1.20	1.25	1.30	1.20	1.25	1.30	V
Line Regulation (Note 2) 3 V ≤ V_I-V_O ≤ 35 V	1	Reg_{line}	—	0.02	0.05	—	0.02	0.07	%/V
Load Regulation (Note 2) 10 mA ≤ I_L ≤ 3 A	2	Reg_{load}							
V_O ≤ 5 V			—	20	50	—	20	70	mV
V_O ≥ 5 V			—	0.3	1	—	0.3	1.5	%V_O
Temperature Stability (T_{low} ≤ T_J ≤ T_{high})	3	T_S	—	1	—	—	1	—	%V_O
Minimum Load Current to Maintain Regulation (V_I-V_O = 35 V)	3	I_{Lmin}	—	3.5	5	—	3.5	10	mA
Maximum Output Current	3	I_{max}							A
V_I-V_O ≤ 10 V, P_D ≤ P_{max}			3.0	4.5	—	3.0	4.5	—	
V_I-V_O = 30 V, P_D ≤ P_{max}, T_A = 25°C			0.3	1	—	0.25	1	—	
RMS Noise, % of V_O T_A = 25°C, 10 Hz ≤ f ≤ 10 kHz	—	N	—	0.003	—	—	0.003	—	%V_O
Ripple Rejection, V_O = 10 V, f = 120 Hz (Note 4)	4	RR							dB
Without C_{ADJ}			—	65	—	—	65	—	
C_{ADJ} = 10 μF			66	80	—	66	80	—	
Long Term Stability, T_J = T_{high} (Note 5) T_A = 25°C for Endpoint Measurements	3	S	—	0.3	1	—	0.3	1	%/1.0 k Hrs.
Thermal Resistance Junction to Case	—	$R_{\theta JC}$							°C/W
Peak (Note 6) K Package (TO-3)			—	2.3	—	—	2.3	—	
T Package (TO-220)			—	—	—	—	2.3	—	
Average (Note 7) K Package (TO-3)			—	—	1.5	—	—	1.5	
T Package (TO-220)			—	—	—	—	—	1.5	

NOTES:

(1) T_{low} = -55°C for LM150 T_{high} = +150°C for LM150
-25°C for LM250 = +150°C for LM250
0°C for LM350 = +125°C for LM350

(2) Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

(3) Selected devices with tightened tolerance reference voltage available.

(4) C_{ADJ}, when used, is connected between the adjustment pin and ground.

(5) Since Long Term Stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot.

(6) Thermal Resistance evaluated measuring the hottest temperature on the die using an infrared scanner. This method of evaluation yields very accurate thermal resistance values which are conservative when compared to other measurement techniques.

(7) The average die temperature is used to derive the value of thermal resistance junction to case (average).

SCHEMATIC DIAGRAM

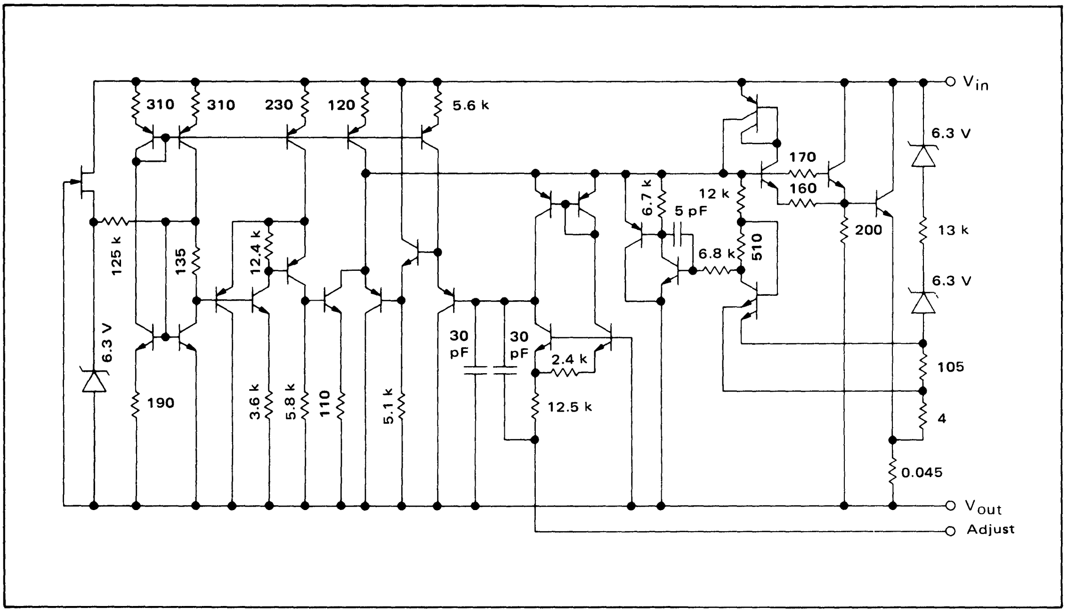
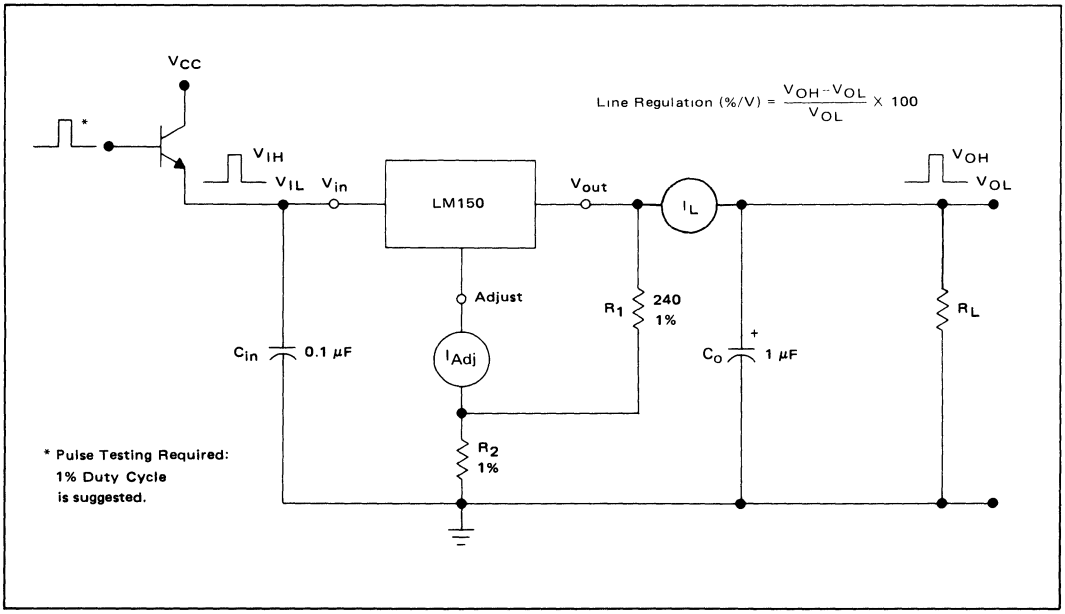


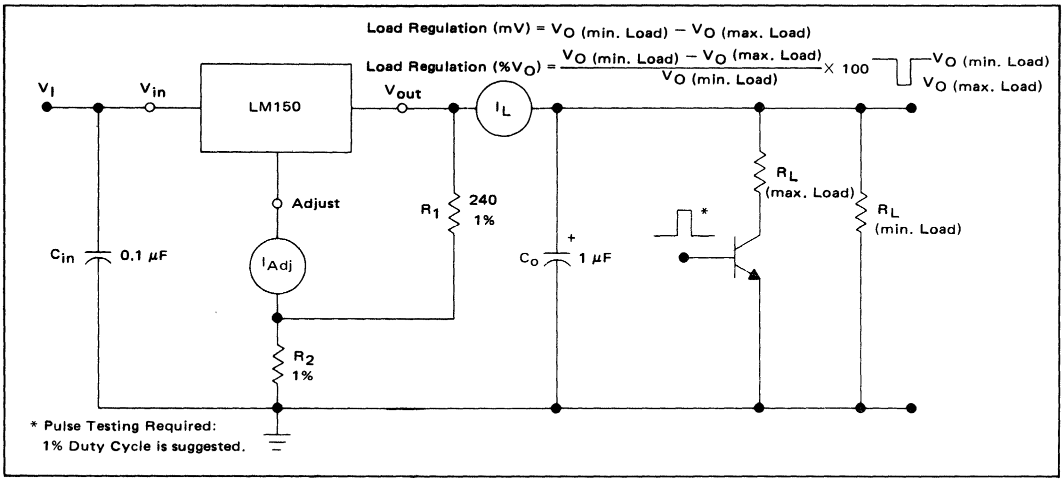
FIGURE 1 – LINE REGULATION AND ΔI_{Adj}/LINE TEST CIRCUIT



$$\text{Line Regulation (\%/V)} = \frac{V_{OH} - V_{OL}}{V_{OL}} \times 100$$

* Pulse Testing Required:
1% Duty Cycle
is suggested.

FIGURE 2 – LOAD REGULATION AND ΔI_{Adj}/LOAD TEST CIRCUIT

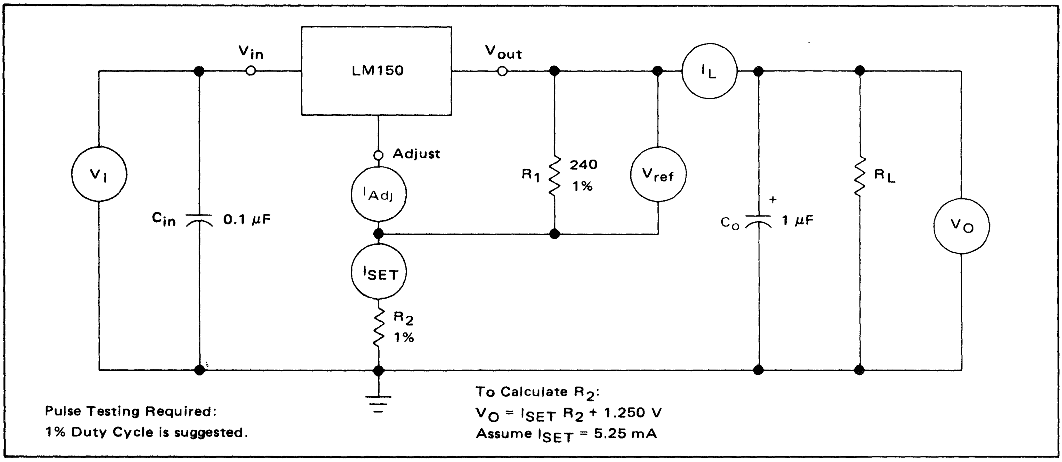


Load Regulation (mV) = V_O (min. Load) – V_O (max. Load)

Load Regulation ($\%V_O$) = $\frac{V_O \text{ (min. Load)} - V_O \text{ (max. Load)}}{V_O \text{ (min. Load)}} \times 100$

* Pulse Testing Required:
1% Duty Cycle is suggested.

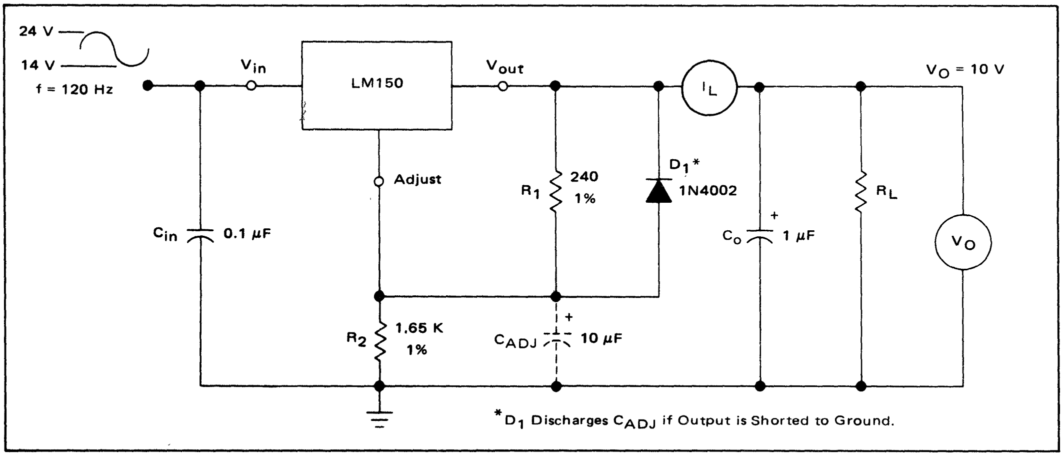
FIGURE 3 – STANDARD TEST CIRCUIT



Pulse Testing Required:
1% Duty Cycle is suggested.

To Calculate R_2:
$V_O = I_{SET}\,R_2 + 1.250$ V
Assume $I_{SET} = 5.25$ mA

FIGURE 4 – RIPPLE REJECTION TEST CIRCUIT



*D_1 Discharges C_{ADJ} if Output is Shorted to Ground.

FIGURE 5 – LOAD REGULATION

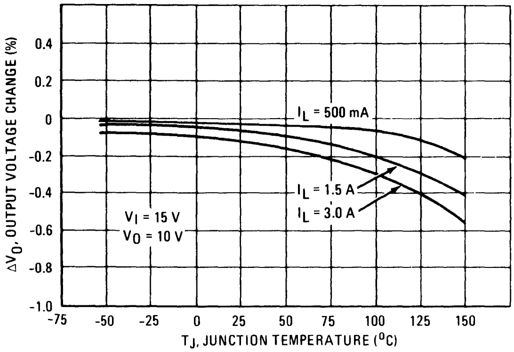


FIGURE 6 – CURRENT LIMIT

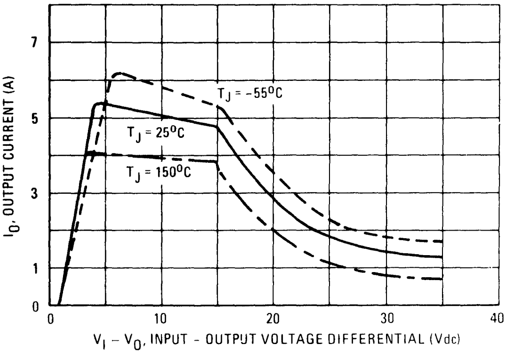


FIGURE 7 – ADJUSTMENT PIN CURRENT

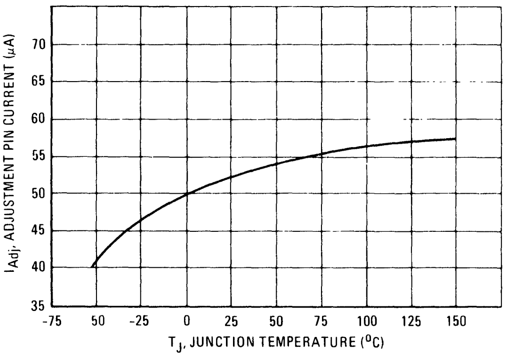


FIGURE 8 – DROPOUT VOLTAGE

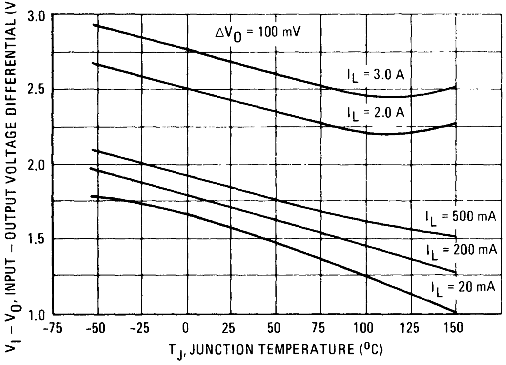


FIGURE 9 – TEMPERATURE STABILITY

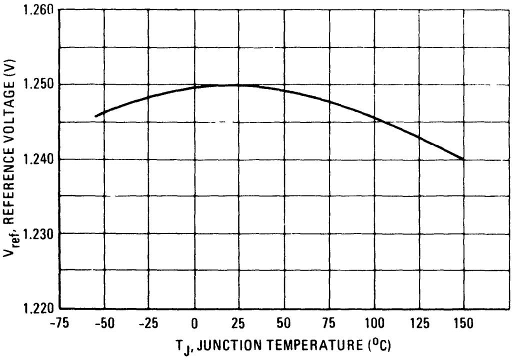


FIGURE 10 – MINIMUM OPERATING CURRENT

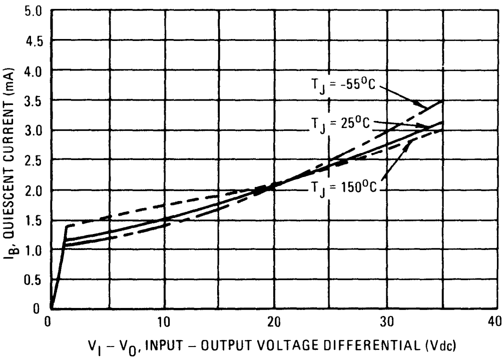


FIGURE 11 – RIPPLE REJECTION VS OUTPUT VOLTAGE

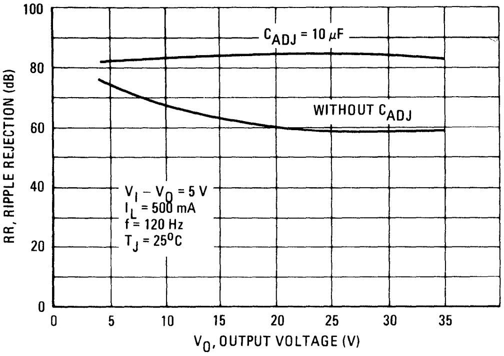


FIGURE 12 – RIPPLE REJECTION VS. OUTPUT CURRENT

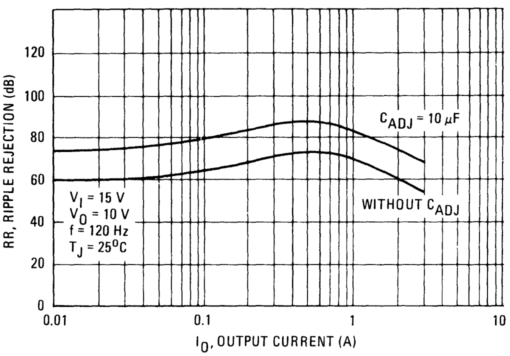


FIGURE 13 – RIPPLE REJECTION VS. FREQUENCY

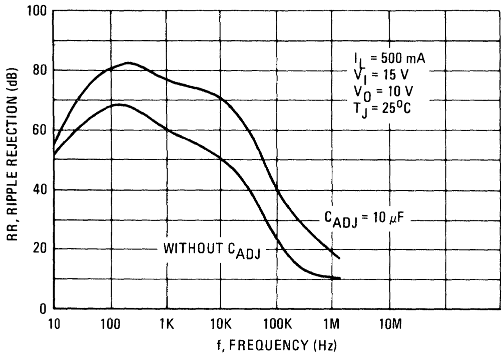


FIGURE 14 – OUTPUT IMPEDANCE

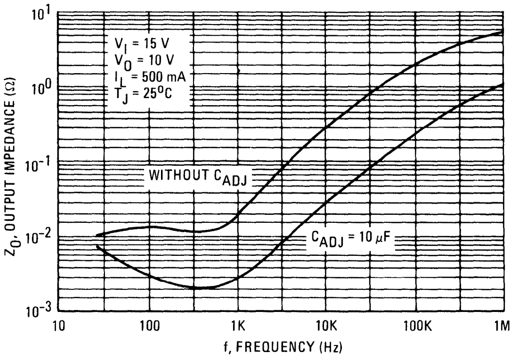


FIGURE 15 – LINE TRANSIENT RESPONSE

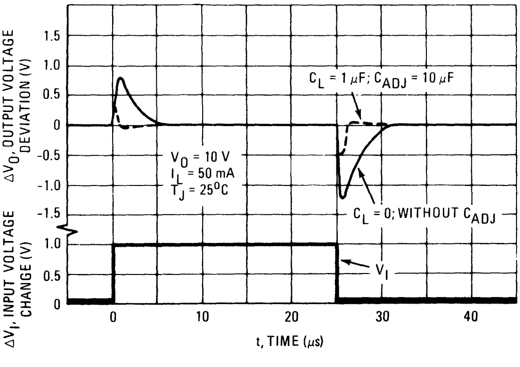
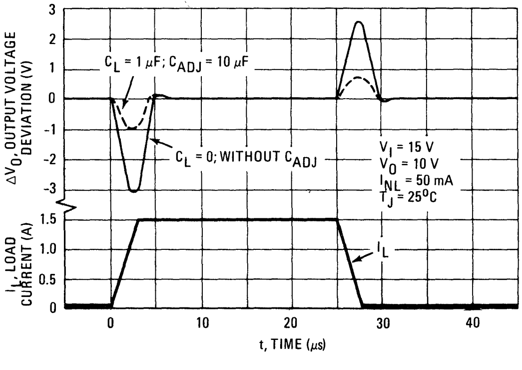


FIGURE 16 – LOAD TRANSIENT RESPONSE



APPLICATIONS INFORMATION

BASIC CIRCUIT OPERATION

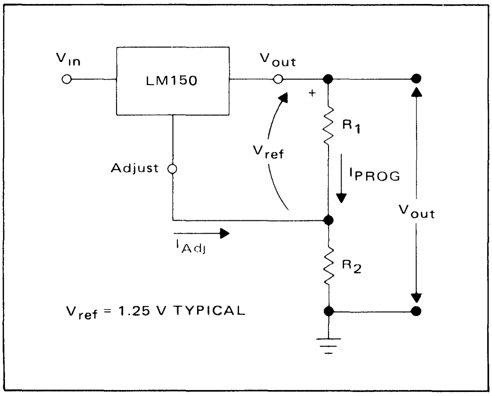
The LM150 is a 3-terminal floating regulator. In operation, the LM150 develops and maintains a nominal 1.25 volt reference (V_{ref}) between its output and adjustment terminals. This reference voltage is converted to a programming current (I_{PROG}) by R1 (see Figure 17), and this constant current flows through R2 to ground. The regulated output voltage is given by:

$$V_{out} = V_{ref} (1 + \frac{R2}{R1}) + I_{Adj} R2$$

Since the current from the adjustment terminal (I_{Adj}) represents an error term in the equation, the LM150 was designed to control I_{Adj} to less than 100 μA and keep it constant. To do this, all quiescent operating current is returned to the output terminal. This imposes the requirement for a minimum load current. If the load current is less than this minimum, the output voltage will rise.

Since the LM150 is a floating regulator, it is only the voltage differential across the circuit which is important to performance, and operation at high voltages with respect to ground is possible.

FIGURE 17 – BASIC CIRCUIT CONFIGURATION



LOAD REGULATION

The LM150 is capable of providing extremely good load regulation, but a few precautions are needed to obtain maximum performance. For best performance, the programming resistor (R1) should be connected as close to the regulator as possible to minimize line drops which effectively appear in series with the reference, thereby degrading regulation. The ground end of R2 can be returned near the load ground to provide remote ground sensing and improve load regulation.

EXTERNAL CAPACITORS

A 0.1 μF disc or 1 μF tantalum input bypass capacitor (C_{in}) is recommended to reduce the sensitivity to input line impedance.

The adjustment terminal may be bypassed to ground to improve ripple rejection. This capacitor (C_{ADJ}) prevents ripple from being amplified as the output voltage is increased. A 10 μF capacitor should improve ripple rejection about 15dB at 120 Hz in a 10 volt application.

Although the LM150 is stable with no output capacitance, like any feedback circuit, certain values of external capacitance can cause excessive ringing. An output capacitance (C_o) in the form of a 1 μF tantalum or 25 μF aluminum electrolytic capacitor on the output swamps this effect and insures stability.

PROTECTION DIODES

When external capacitors are used with any I.C. regulator it is sometimes necessary to add protection diodes to prevent the capacitors from discharging through low current points into the regulator.

Figure 18 shows the LM150 with the recommended protection diodes for output voltages in excess of 25 V or high capacitance values ($C_o > 25$ μF, $C_{ADJ} > 10$ μF). Diode D_1 prevents C_o from discharging thru the I.C. during an input short circuit. Diode D_2 protects against capacitor C_{ADJ} discharging through the I.C. during an output short circuit. The combination of diodes D1 and D2 prevents C_{ADJ} from discharging through the I.C. during an input short circuit.

FIGURE 18 – VOLTAGE REGULATOR WITH PROTECTION DIODES

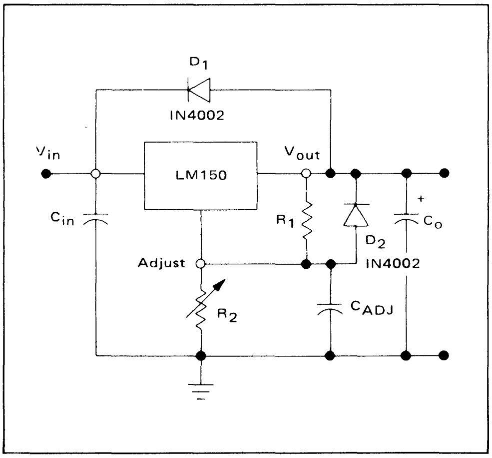
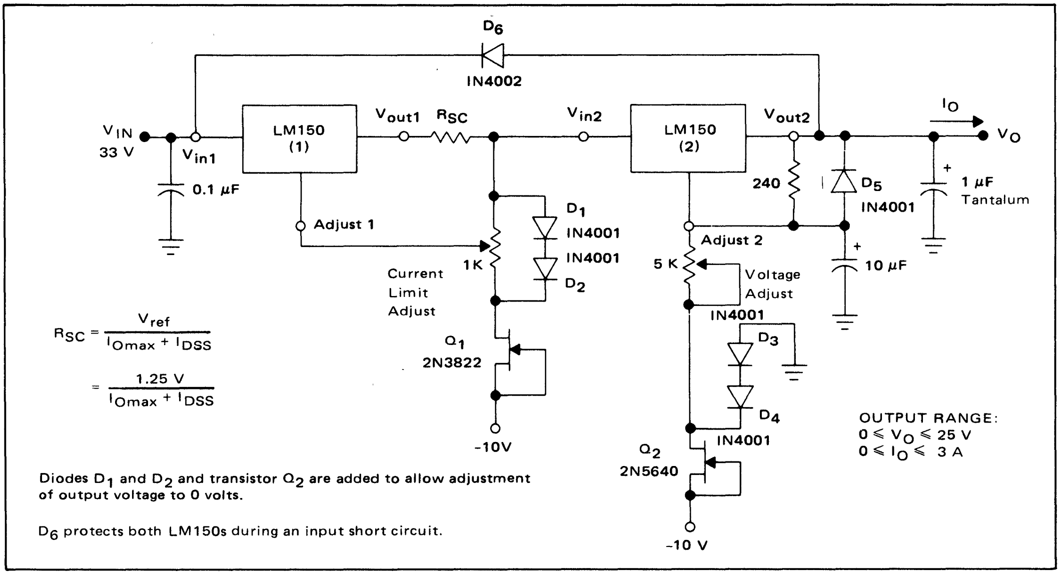


FIGURE 19 – "LABORATORY" POWER SUPPLY WITH ADJUSTABLE CURRENT LIMIT AND OUTPUT VOLTAGE



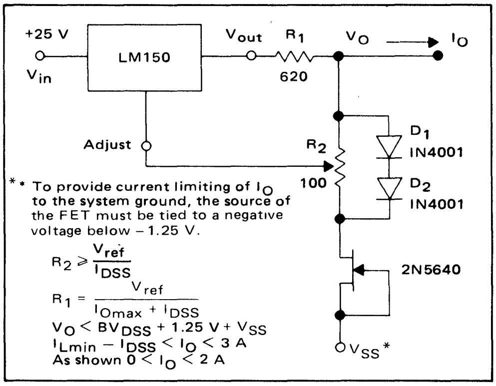
$$R_{SC} = \frac{V_{ref}}{I_{Omax} + I_{DSS}}$$

$$= \frac{1.25\ V}{I_{Omax} + I_{DSS}}$$

Diodes D_1 and D_2 and transistor Q_2 are added to allow adjustment of output voltage to 0 volts.

D_6 protects both LM150s during an input short circuit.

FIGURE 20 – ADJUSTABLE CURRENT LIMITER



* To provide current limiting of I_O to the system ground, the source of the FET must be tied to a negative voltage below – 1.25 V.

$$R_2 \geqslant \frac{V_{ref}}{I_{DSS}}$$

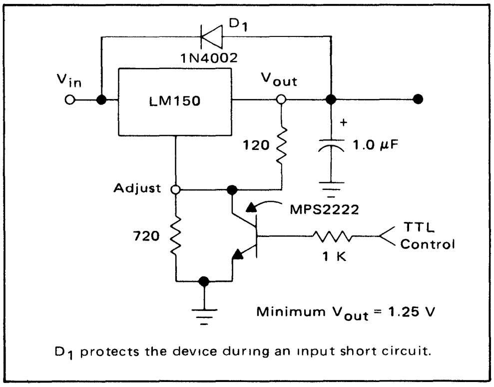
$$R_1 = \frac{V_{ref}}{I_{Omax} + I_{DSS}}$$

$V_O < BV_{DSS} + 1.25\ V + V_{SS}$

$I_{Lmin} - I_{DSS} < I_O < 3\ A$

As shown $0 < I_O < 2$ A

FIGURE 21 – 5 V ELECTRONIC SHUT DOWN REGULATOR



Minimum V_{out} = 1.25 V

D_1 protects the device during an input short circuit.

FIGURE 22 – SLOW TURN-ON REGULATOR

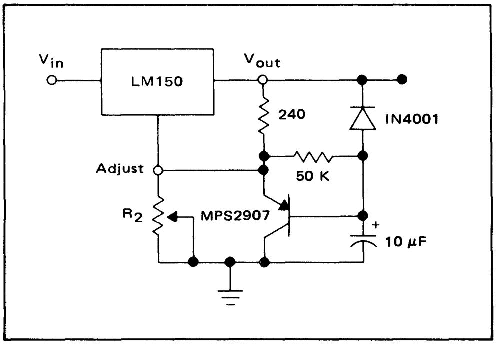
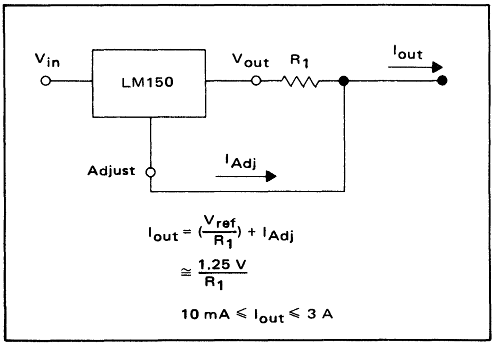


FIGURE 23 – CURRENT REGULATOR



$$I_{out} = \left(\frac{V_{ref}}{R_1}\right) + I_{Adj}$$

$$\cong \frac{1.25\ V}{R_1}$$

$10\ mA \leqslant I_{out} \leqslant 3\ A$

MOTOROLA

MC1463
MC1563

Specifications and Applications Information

NEGATIVE VOLTAGE REGULATOR

The MC1563/MC1463 is a "three terminal" negative regulator designed to deliver continuous load current up to 500 mAdc and provide a maximum negative input voltage of −40 Vdc. Output current capability can be increased to greater than 10 Adc through use of one or more external transistors.

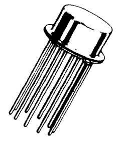
Specifications and performance of the MC1563/MC1463 Negative Voltage Regulator are nearly identical to the MC1569/MC1469 Positive Voltage Regulator. For systems requiring both a positive and negative power supply, these devices are excellent for use as complementary regulators and offer the advantage of operating with a common input ground.

The MC1563R/MC1463R case can be mounted directly to a grounded heat sink which eliminates the need for an insulator.

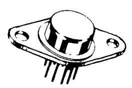
- Case is at Ground Potential (R package)
- Electronic "Shutdown" and Short-Circuit Protection
- Low Output Impedance – 20 Milliohms typical
- High Power Capability – 9.0 Watts
- Excellent Temperature Stability – $\Delta V_O/\Delta T = \pm 0.002\%/^{\circ}C$ typical
- High Ripple Rejection – 0.002% typical
- 500 mA Current Capability

NEGATIVE-POWER-SUPPLY VOLTAGE REGULATOR

SILICON MONOLITHIC INTEGRATED CIRCUIT



G SUFFIX
METAL PACKAGE
CASE 603



R SUFFIX
METAL PACKAGE
CASE 614

FIGURE 1 – TYPICAL CIRCUIT CONNECTION
$(|-3.5| \leqslant V_O \leqslant |-37|$ Vdc, $1 \leqslant I_L \leqslant 500$ mA$)$

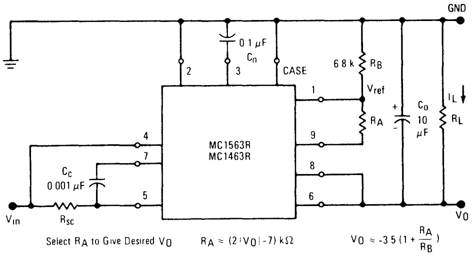


FIGURE 2 – TYPICAL NPN CURRENT BOOST CONNECTION
$(V_O = 5.2$ Vdc, $I_L = 10$ Adc [max]$)$

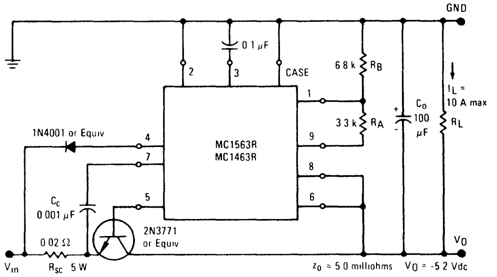
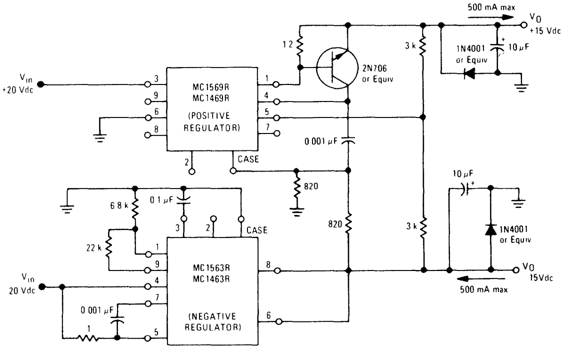


FIGURE 3 – ±15 V, ±400 mA COMPLEMENTARY TRACKING VOLTAGE REGULATOR



ORDERING INFORMATION		
DEVICE	TEMPERATURE RANGE	PACKAGE
MC1463G	0° C to +70° C	Metal Can
MC1463R	0° C to +70° C	Metal Power
MC1563G	−55° C to +125° C	Metal Can
MC1563R	−55° C to +125° C	Metal Power

MAXIMUM RATINGS (T_C = +25°C unless otherwise noted.)

Rating	Symbol	Value		Unit
Input Voltage MC1463 MC1563	V_I	 −35 −40		Vdc
		G Package	R Package	
Load Current — Peak	I_L	250	600	mA
Current, Pin 2	I_2	10	10	mA
Power Dissipation and Thermal Characteristics T_A = 25°C Derate above T_A = 25°C Thermal Resistance, Junction to Air T_C = 25°C Derate above T_C = 25°C Thermal Resistance, Junction to Case	 P_D $1/R_{\theta JA}$ $R_{\theta JA}$ P_D $1/R_{\theta JC}$ $R_{\theta JC}$	 0.68 5.44 184 1.8 14.4 69.4	 2.4 16 62 9.0 61 17	 Watts mW/°C °C/W Watts mW/°C °C/W
Operating and Storage Junction Temperature Range	T_J, T_{stg}	-65 to +150		°C

OPERATING TEMPERATURE RANGE

Operating Ambient Temperature Range MC1463 MC1563	T_A	 0 to +70 -55 to +125	°C

ELECTRICAL CHARACTERISTICS (I_L = 100 mAdc, T_C = +25°C, V_{in} = 15 V, V_O = 10 V unless otherwise noted.)

Characteristic	Fig.	Note	Symbol	MC1563			MC1463			Unit
				Min	Typ	Max	Min	Typ	Max	
Input Voltage (T_A = T_{low} ① to T_{high} ② I_L = 1.0 mA)	4	1,6	V_I	−8.5	–	−40	−9.0	–	−35	Vdc
Output Voltage Range (I_L = 1.0 mA)	4	–	V_O	−3.6	–	−37	−3.8	–	−32	Vdc
Reference Voltage (Pin 1 to Ground)	4	–	V_{ref}	−3.4	−3.5	−3.6	−3.2	−3.5	−3.8	Vdc
Minimum Input-Output Voltage Differential (R_{sc} = 0)	4	2	$\lvert V_{in} - V_O \rvert$	–	1.5	2.7	–	1.5	3.0	Vdc
Bias Current (Standby Current) (I_L = 1.0 mAdc, $I_{IB} = I_I - I_L$)	4	–	I_{IB}	–	7.0	11	–	7.0	14	mAdc
Output Noise (C_n = 0.1 μF, f = 10 Hz to 5.0 MHz)	4	–	v_N	–	120	–	–	120	–	μV(rms)
Temperature Coefficient of Output Voltage	4	3	$\Delta V_O/\Delta T$	–	±0.002	–	–	±0.002	–	%/°C
Operating Load Current Range (R_{sc} = 0.3 ohm) R Package (R_{sc} = 2.0 ohms) G Package	4	–	I_{LR}	 1.0 1.0	 – –	 500 200	 1.0 1.0	 – –	 500 200	mAdc
Input Regulation (V_{in} = 1.0 V rms, f = 1.0 kHz)	4	4	Reg_{line}	–	0.002	0.015	–	0.003	0.030	%/V_O
Load Regulation (T_J = Constant [1.0 mA $\leqslant I_L \leqslant$ 20 mA]) (T_C = +25°C [1.0 mA $\leqslant I_L \leqslant$ 50 mA]) R Package G Package	6	5	Reg_{load}	 – – –	 0.4 0.005 0.01	 1.6 0.05 0.13	 – – –	 0.7 0.005 0.01	 2.4 0.05 0.13	 mV %
Output Impedance (f = 1.0 kHz)	7	–	z_o	–	20	–	–	35	–	milliohms
Shutdown Current (V_I = -35 Vdc)	8	–	I_{sd}	–	7.0	15	–	14	50	μAdc

① T_{low} = 0°C for MC1463
= -55°C for MC1563

② T_{high} = +70°C for MC1463
= +125°C for MC1563

Heat sink required for T_{high} testing of "G" package.

Note 1 "Minimum Input Voltage" is the minimum "total instantaneous input voltage" required to properly bias the internal zener reference diode.

Note 2. This parameter states that the MC1563/MC1463 will regulate properly with the input-output voltage differential $|V_I - V_O|$ as low as 2.7 Vdc and 3.0 Vdc respectively. Typical units will regulate properly with $|V_I - V_O|$ as low as 1.5 Vdc as shown in the typical column.

Note 3. "Temperature Coefficient of Output Voltage" is defined as

$$\Delta V_O/\Delta T = \frac{\pm\,(V_O\ \text{max} - V_O\ \text{min})\ (100)}{\triangle T_A\ (V_O\ @\ T_A = +25^\circ C)}$$

where $\triangle T_A$ = +180°C for the MC1563
+75°C for the MC1463

The output-voltage adjusting resistors (R_A and R_B) must have matched temperature characteristics in order to maintain a constant ratio independent of temperature

Note 4. Input regulation is the percentage change in output voltage per volt change in the input voltage and is expressed as

$$\text{Input Regulation} = \frac{v_O}{V_O\ (v_I)}\ 100\ \ (\%/V_O).$$

where v_O is the change in the output voltage V_O for the input change v_{in}.

The following example illustrates how to compute maximum output voltage change for the conditions given:

$$Reg_{in} = 0.015\%/V_O$$
$$V_O = 10\ \text{Vdc}$$
$$v_{in} = 1.0\ \text{V(rms)}$$
$$v_O = \frac{(Reg_{line})\ (V_I)\ (V_O)}{100} = \frac{(0.015)(1.0)(10)}{100} = 0.0015\ \text{V(rms)}$$

Note 5. Temperature drift effect must be taken into account separately for conditions of high junction temperature changes due to the thermal feedback that exists on the monolithic chip.

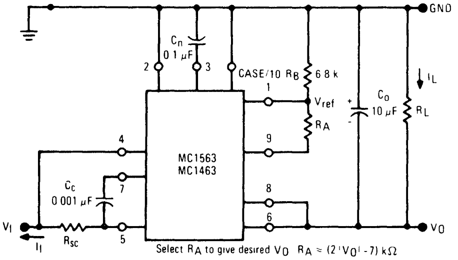
$$\text{Load Regulation} = \frac{V_O|_{I_L = 1.0\ \text{mA}}| - V_O|_{I_L = 50\ \text{mA}}|}{V_O|_{I_L = 1.0\ \text{mA}}|} \times 100$$

Note 6. Not to exceed maximum package power dissipation

TEST CIRCUITS

(I_L = 100 mAdc, T_C = +25°C unless otherwise noted.)

FIGURE 4 – GENERAL TEST CIRCUIT



Select R_A to give desired V_O $R_A \approx (2|V_O| - 7)$ kΩ

FIGURE 5 – LOAD TRANSIENT RESPONSE

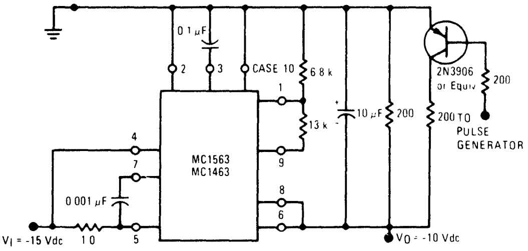


FIGURE 6 – LOAD REGULATION

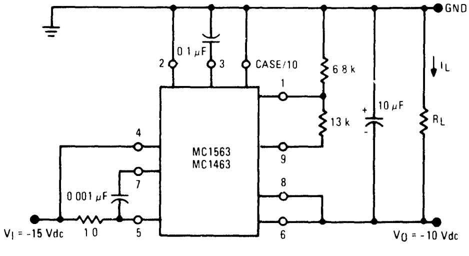


FIGURE 7 – OUTPUT IMPEDANCE

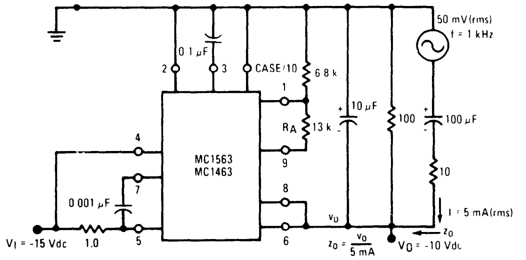
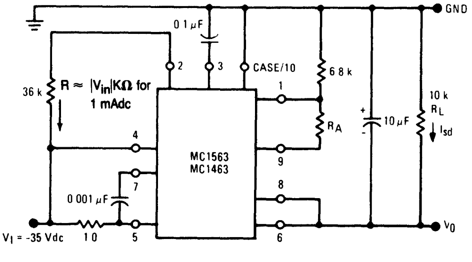


FIGURE 8 – SHUTDOWN CURRENT



GENERAL DESIGN INFORMATION

1. Output Voltage, V_O
   a) Output Voltage is set by resistors R_A and R_B (see Figure 9). Set R_B = 6.8 k ohms and determine R_A from the graph of Figure 11 or from the equation:

   $$R_A \approx (2\,|V_O|-7)\ \text{k}\Omega$$

   b) Output voltage can be varied by making R_A adjustable as shown in Figures 9 and 10.

   c) Output voltage, V_O, is determined by the ratio of R_A and R_B therefore optimum temperature performance can be achieved if R_A and R_B have the same temperature coefficient.

   d) $V_O = V_{ref}\,(1 + \frac{R_A}{R_B})$; therefore the tolerance on output voltage is determined by the tolerance of V_{ref} and R_A and R_B.

2. Short-Circuit Current, I_{SC}
   Short-Circuit Current, I_{SC} is determined by R_{sc}. R_{sc} may be chosen with the aid of Figure 11 when using the typical circuit connection of Figure 9.

3. Compensation, C_c
   A 0.001 µF capacitor (C_c, see Figure 9), will provide adequate compensation in most applications, with or without current boost. Smaller values of C_c will reduce stability and larger values of C_c will degrade pulse response and output impedance versus frequency. The physical location of C_c should be close to the MC1563/MC1463 with short lead lengths.

4. Noise Filter Capacitor, C_n
   A 0.1 µF capacitor, C_n, from Pin 3 to ground will typically reduce the output noise voltage to 120 µV(rms). The value of C_n can be increased or decreased, depending on the noise voltage requirements of a particular application. A minimum value of 0.001 µF is recommended.

5. Output Capacitor, C_O
   The value of C_O should be at least 10 µF in order to provide good stability.

6. Shutdown Control
   One method of turning "OFF" the regulator is to draw 1 mA from Pin 2 (See Figure 8). This control can be used to eliminate power consumption by circuit loads which can be put in "standby" mode. Examples include, an ac or dc "squelch" control for communications circuits, and a dissipation control to protect the regulator under sustained output short-circuiting. As the magnitude of the input-threshold voltage at Pin 2 depends directly upon the junction temperature of the integrated circuit chip, a fixed dc voltage at Pin 2 will cause automatic shutdown for high junction temperatures. This will protect the chip, independent of the heat sinking used, the ambient temperature, or the input or output voltage levels. Standard Logic levels of MRTL, MDTL* or MTTL* can also be used to turn the regulator "ON" or "OFF".

7. Remote Sensing
   The connection to Pin 8 can be made with a separate lead direct to the load. Thus, "remote sensing" can be achieved and the effect of undesired impedances (including that of the milliammeter used to measure I_L) on z_O can be greatly reduced.

FIGURE 9 – TYPICAL CIRCUIT CONNECTION

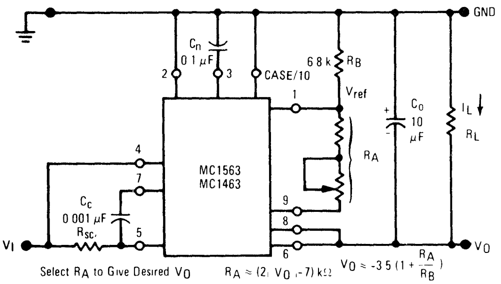


FIGURE 10 – R_A versus V_O

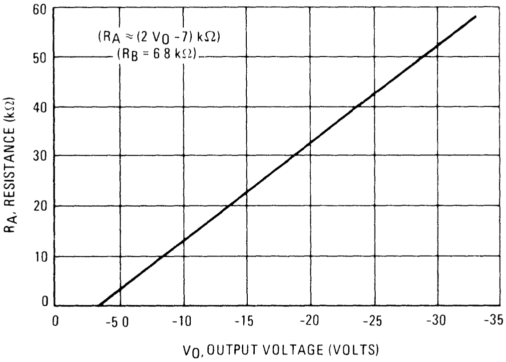
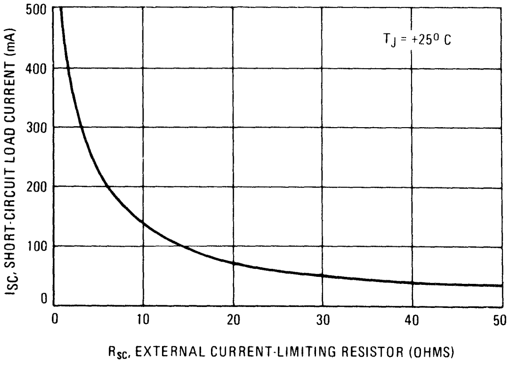


FIGURE 11 – I_{sc} versus R_{sc}



TYPICAL CHARACTERISTICS

Unless otherwise noted: $C_n = 0.1\,\mu F, C_c = 0.001\,\mu F, C_o = 10\,\mu F, T_C = +25^oC,$
$V_{I(nom)} = -15$ Vdc, $V_{O(nom)} = -10$ Vdc, $I_L = 100$ mAdc.

FIGURE 12 – TEMPERATURE DEPENDENCE OF SHORT-CIRCUIT LOAD CURRENT

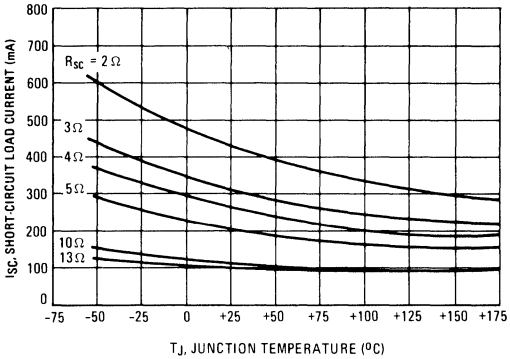


FIGURE 13 – FREQUENCY DEPENDENCE OF OUTPUT IMPEDANCE

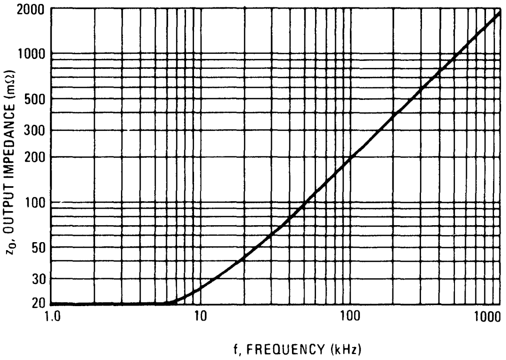


FIGURE 14 – DEPENDENCE OF OUTPUT IMPEDANCE ON OUTPUT VOLTAGE

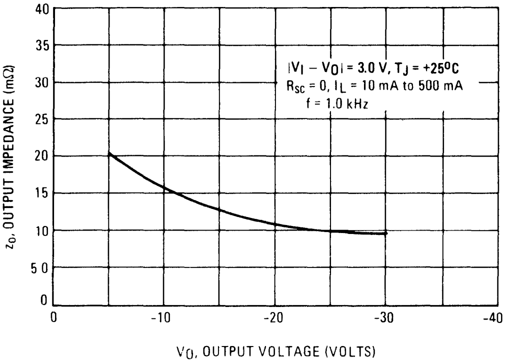


FIGURE 15 – OUTPUT IMPEDANCE versus R_{sc}

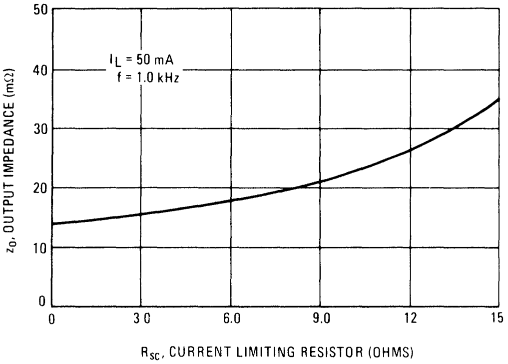
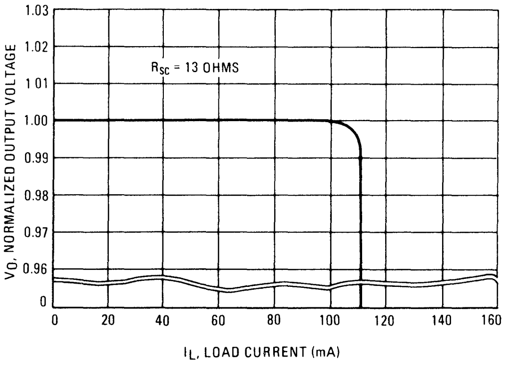


FIGURE 16 – CURRENT LIMITING CHARACTERISTICS



TYPICAL CHARACTERISTICS (continued)

FIGURE 17 – BIAS CURRENT versus INPUT VOLTAGE

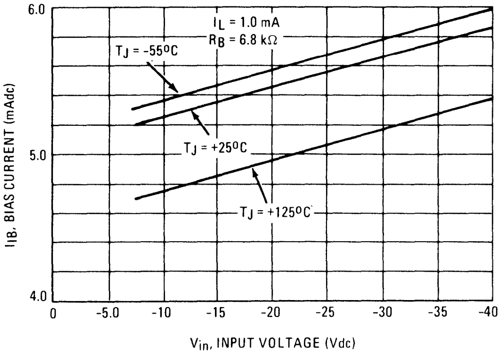


FIGURE 18 – EFFECTS OF LOAD CURRENT ON INPUT-OUTPUT VOLTAGE DIFFERENTIAL

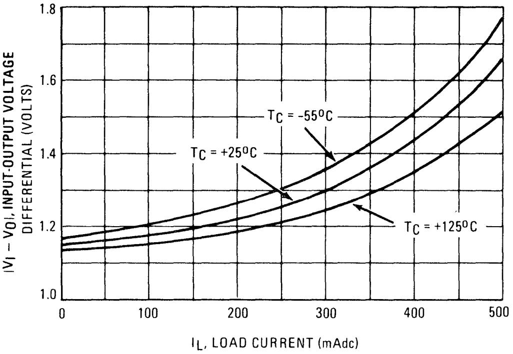


FIGURE 19 – EFFECT OF INPUT-OUTPUT VOLTAGE DIFFERENTIAL ON INPUT REGULATION

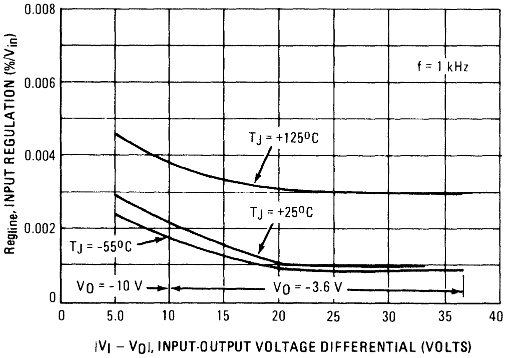


FIGURE 20 – INPUT TRANSIENT RESPONSE

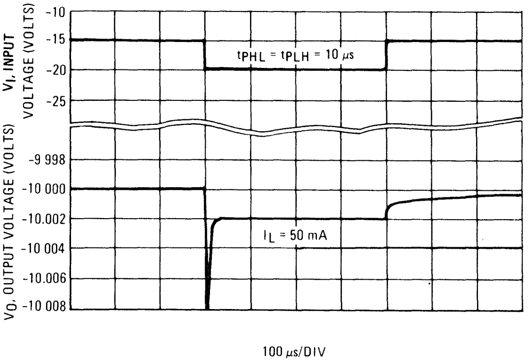


FIGURE 21 – LOAD TRANSIENT RESPONSE

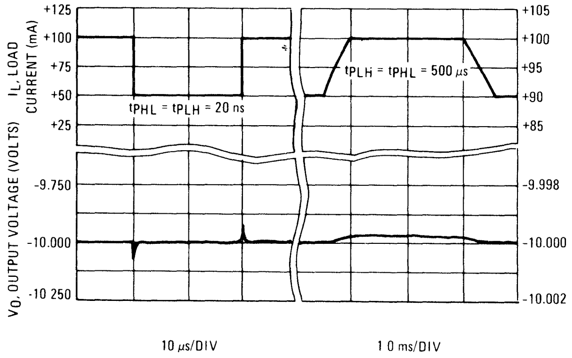
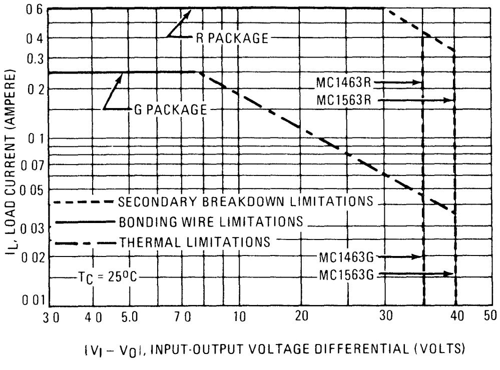


FIGURE 22 – DC OPERATING AREA



MOTOROLA

MC1466L
MC1566L

Specifications and Applications Information

MONOLITHIC VOLTAGE AND CURRENT REGULATOR

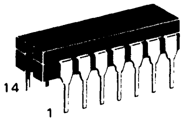
This unique "floating" regulator can deliver hundreds of volts – limited only by the breakdown voltage of the external series pass transistor. Output voltage and output current are adjustable. The MC1466/MC1566 integrated circuit voltage and current regulator is designed to give "laboratory" power-supply performance.

- Voltage/Current Regulation with Automatic Crossover
- Excellent Line Voltage Regulation, 0.01% +1.0 mV
- Excellent Load Voltage Regulation, 0.01% +1.0 mV
- Excellent Current Regulation, 0.1% +1.0 mA
- Short-Circuit Protection
- Output Voltage Adjustable to Zero Volts
- Internal Reference Voltage
- Adjustable Internal Current Source

PRECISION WIDE-RANGE VOLTAGE and CURRENT REGULATOR

EPITAXIAL PASSIVATED
INTEGRATED CIRCUIT

CERAMIC PACKAGE
CASE 632
TO-116

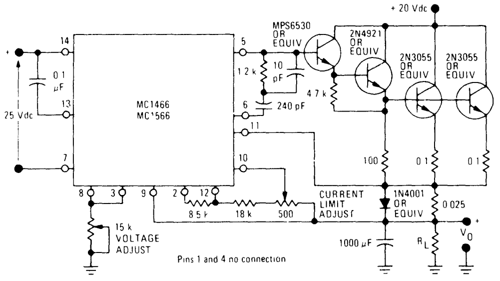


ORDERING INFORMATION

Device	Temperature Range	Package
MC1466L	0°C to +70° C	Ceramic DIP
MC1566L	−55°C to +125°C	Ceramic DIP

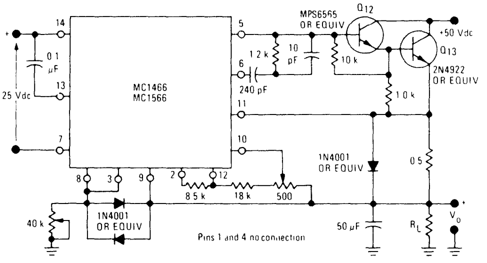
TYPICAL APPLICATIONS

FIGURE 1 – 0-TO-15 VDC, 10-AMPERES REGULATOR



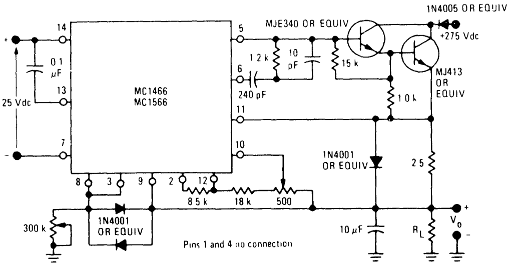
Pins 1 and 4 no connection

FIGURE 2 – 0-TO-40 VDC, 0.5-AMPERE REGULATOR



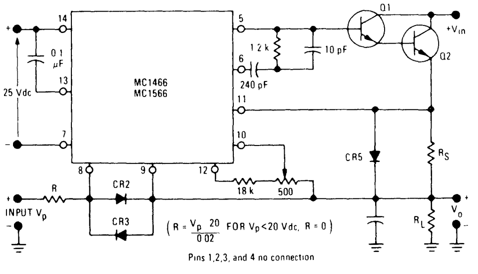
Pins 1 and 4 no connection

FIGURE 3 – 0-TO-250 VDC, 0.1·AMPERE REGULATOR



Pins 1 and 4 no connection

FIGURE 4 – REMOTE PROGRAMMING



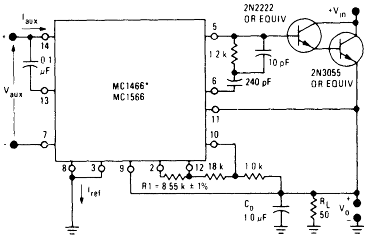
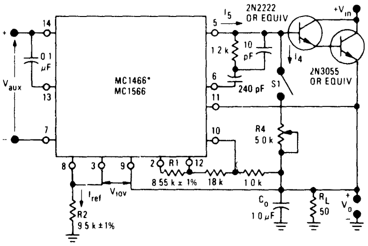
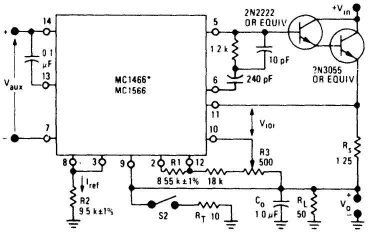
$\left(R = \frac{V_p \;\; 20}{0\,02} \text{ FOR } V_p < 20 \text{ Vdc, } R = 0 \right)$

Pins 1,2,3, and 4 no connection

MAXIMUM RATINGS (T_A = +25° unless otherwise noted)

Rating		Symbol	Value	Unit
Auxiliary Voltage		V_{aux}		Vdc
	MC1466		30	
	MC1566		35	
Power Dissipation (Package Limitation)		P_D	750	mW
Derate above T_A = +50°C		$1/\theta_{JA}$	6.0	mW/°C
Operating Temperature Range		T_A		°C
	MC1466		0 to +70	
	MC1566		−55 to +125	
Storage Temperature Range		T_{stg}	−65 to +150	°C

ELECTRICAL CHARACTERISTICS (T_A = +25°C, V_{aux} = +25 Vdc unless otherwise noted)

Characteristic Definition	Characteristic		Symbol	Min	Typ	Max	Units
	Auxiliary Voltage (See Notes 1 & 2)		V_{aux}				Vdc
	(Voltage from pin 14 to pin 7)	MC1466		21	–	30	
		MC1566		20	–	35	
	Auxiliary Current		I_{aux}				mAdc
		MC1466		–	9.0	12	
		MC1566		–	7.0	8.5	
	Internal Reference Voltage		V_{IR}				Vdc
	(Voltage from pin 12 to pin 7)	MC1466		17.3	18.2	19.7	
		MC1566		17.5	18.2	19	
	Reference Current (See Note 3)		I_{ref}				mAdc
		MC1466		0.8	1.0	1.2	
		MC1566		0.9	1.0	1.1	
	Input Current-Pin 8		I_8				µAdc
		MC1466		–	6.0	12	
		MC1566		–	3.0	6.0	
	Power Dissipation		P_D				mW
		MC1466		–	–	360	
		MC1566		–	–	300	
	Input Offset Voltage, Voltage Control		V_{iov}				mVdc
	Amplifier (See Note 4)	MC1466		0	15	40	
		MC1566		3.0	15	25	
	Load Voltage Regulation		ΔV_{iov}				
	(See Note 5)	MC1466		–	1.0	3.0	mV
		MC1566		–	0.7	1.0	
		MC1466	$\Delta V_{ref}/V_{ref}$	–	0.015	0.03	%
		MC1566		–	0.004	0.01	
	Line Voltage Regulation		ΔV_{iov}				
	(See Note 6)	MC1466		–	1.0	3.0	mV
		MC1566		–	0.7	1.0	
		MC1466	$\Delta V_{ref}/V_{ref}$	–	0.015	0.03	%
		MC1566		–	0.004	0.01	
	Temperature Coefficient of Output Voltage		TCV_o				%/°C
	(T_A = 0 to +75°C)	MC1466		–	0.01	–	
	(T_A = −55 to +25°C)	MC1566		–	0.006	–	
	(T_A = +25 to +125°C)	MC1566		–	0.004	–	
	Input Offset Voltage, Current Control		V_{ioi}				mVdc
	Amplifier (See Note 4)	MC1466		0	15	40	
	(Voltage from pin 10 to pin 11)	MC1566		3.0	15	25	
	Load Current Regulation		$\Delta I_L/I_L$				
	(See Note 7)	MC1466		–	–	0.2	%
		MC1566		–	–	0.1	
		MC1466	ΔI_{ref}	–	–	1.0	mAdc
		MC1566		–	–	1.0	

*Pins 1 and 4 no connection.

NOTE 1:
The instantaneous input voltage, V_{aux}, must not exceed the maximum value of 30 volts for the MC1466 or 35 volts for the MC1566. The instantaneous value of V_{aux} must be greater than 20 volts for the MC1566 or 21 volts for the MC1466 for proper internal regulation.

NOTE 2:
The auxiliary supply voltage V_{aux}, must "float" and be electrically isolated from the unregulated high voltage supply, V_{in}.

NOTE 3:
Reference current may be set to any value of current less than 1.2 mAdc by applying the relationship:

$$I_{ref}\ (mA) = \frac{8.55}{R_1\ (k\Omega)}$$

NOTE 4:
A built-in offset voltage (15 mVdc nominal) is provided so that the power supply output voltage or current may be adjusted to zero

NOTE 5:
Load Voltage Regulation is a function of two additive components, ΔV_{iov} and ΔV_{ref}, where ΔV_{iov} is the change in input offset voltage (measured between pins 8 and 9) and ΔV_{ref} is the change in voltage across R2 (measured between pin 8 and ground). Each component may be measured separately or the sum may be measured across the load. The measurement procedure for the test circuit shown is:

a. With S1 open (I_4 = 0) measure the value of $V_{iov}\ (1)$ and $V_{ref}\ (1)$

b. Close S1, adjust R4 so that I_4 = 500 μA and note $V_{iov}\ (2)$ and $V_{ref}\ (2)$.

Then $\Delta V_{iov} = V_{iov}\ (1) - V_{iov}\ (2)$

% Reference Regulation =

$$\frac{[V_{ref}\ (1) - V_{ref}\ (2)]}{V_{ref}\ (1)}\ (100\%) = \frac{\Delta V_{ref}}{V_{ref}}\ (100\%)$$

Load Voltage Regulation =

$$\frac{\Delta V_{ref}}{V_{ref}}\ (100\%) + \Delta V_{iov}\ .$$

NOTE 6:
Line Voltage Regulation is a function of the same two additive components as Load Voltage Regulation, ΔV_{iov} and ΔV_{ref} (see note 5). The measurement procedure is

a. Set the auxiliary voltage, V_{aux}, to 22 volts for the MC1566 or the MC1466. Read the value of $V_{iov}\ (1)$ and $V_{ref}\ (1)$.

b. Change the V_{aux} to 28 volts for the MC1566 or the MC1466 and note the value of $V_{iov}\ (2)$ and $V_{ref}(2)$. Then compute Line Voltage Regulation:

$\Delta V_{iov} = \Delta V_{iov}\ (1) - V_{iov}\ (2)$

% Reference Regulation =

$$\frac{[V_{ref}\ (1) - V_{ref}\ (2)]}{V_{ref}\ (1)}\ (100\%) = \frac{\Delta V_{ref}}{V_{ref}}\ (100\%)$$

Line Voltage Regulation =

$$\frac{\Delta V_{ref}}{V_{ref}}\ (100\%) + \Delta V_{iov}\ .$$

NOTE 7:
Load Current Regulation is measured by the following procedure:

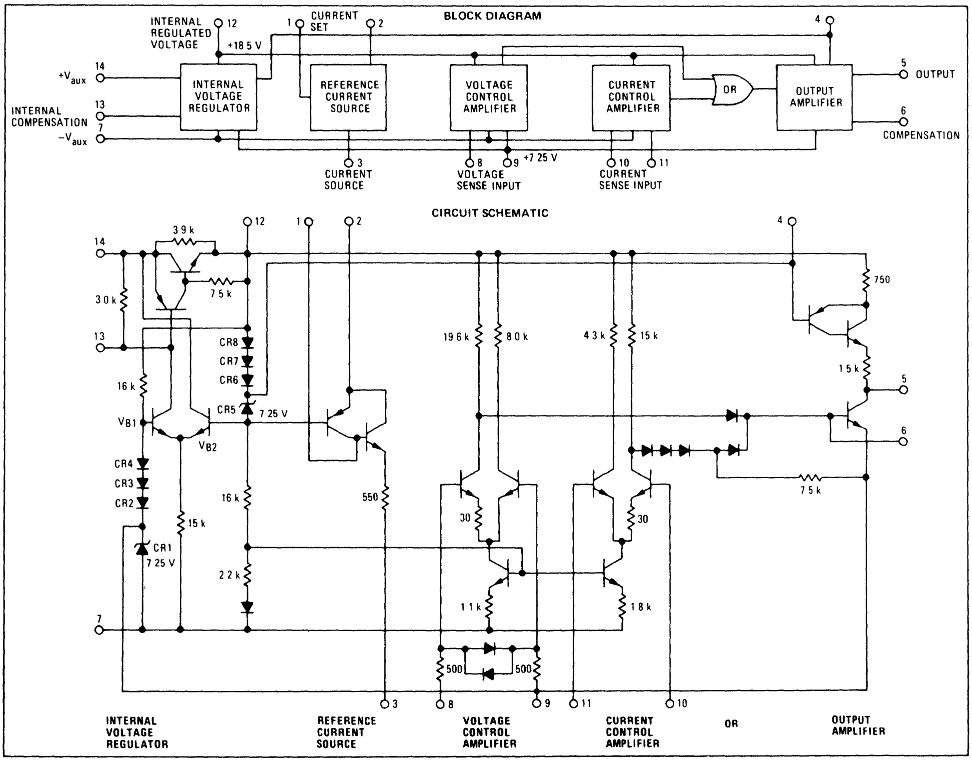
a. With S2 open, adjust R3 for an initial load current, $I_{L(1)}$, such that V_O is 8.0 Vdc.

b. With S2 closed, adjust R_T for V_O = 1.0 Vdc and read $I_{L(2)}$. Then Load Current Regulation =

$$\frac{[I_{L(2)} - I_{L(1)}]}{I_{L(1)}}\ (100\%) + I_{ref}$$

where I_{ref} is 1.0 mAdc, Load Current Regulation is specified in this manner because I_{ref} passes through the load in a direction opposite that of load current and does not pass through the current sense resistor, R_S.

FIGURE 5



MOTOROLA

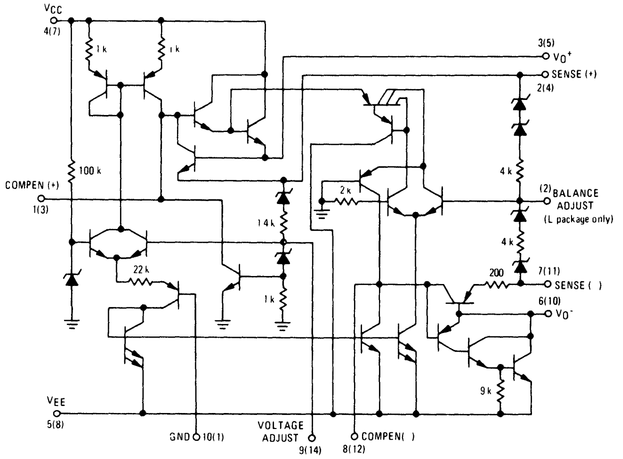
MC1468
MC1568

DUAL ±15-VOLT REGULATOR

The MC1568/MC1468 is a dual polarity tracking regulator designed to provide balanced positive and negative output voltages at currents to 100 mA. Internally, the device is set for ± 15-volt outputs but an external adjustment can be used to change both outputs simultaneously from 8.0 to 20 volts. Input voltages up to ± 30 volts can be used and there is provision for adjustable current limiting. The device is available in three package types to accomodate various power requirements.

- Internally set to ± 15 V Tracking Outputs
- Output Currents to 100 mA
- Outputs Balanced to within 1% (MC1568)
- Line and Load Regulation of 0.06%
- 1% Maximum Output Variation due to Temperature Changes
- Standby Current Drain of 3.0 mA
- Externally Adjustable Current Limit
- Remote Sensing Provisions
- Case is at Ground Potential (R suffix package)

CIRCUIT SCHEMATIC

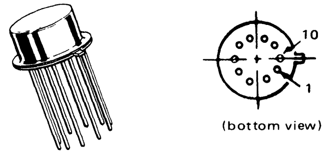


Pin numbers adjacent to terminals are for the G and R suffix packages only Pin numbers in parentheses are for the L suffix package only

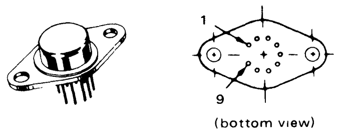
Pin 10 is ground for the G suffix package only For the R package, the case is ground

DUAL ±15-VOLT TRACKING REGULATOR

SILICON MONOLITHIC
INTEGRATED CIRCUIT

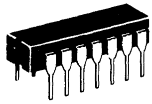


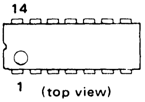
CASE 603C
METAL PACKAGE
TO-100
G SUFFIX



CASE 614
METAL PACKAGE
R SUFFIX

CASE 632
CERAMIC PACKAGE
TO-116
L SUFFIX





ORDERING INFORMATION		
DEVICE	TEMPERATURE RANGE	PACKAGE
MC1468G	0° C to +70° C	Metal Can
MC1468L	0° C to +70° C	Ceramic DIP
MC1468R	0° C to +70° C	Metal Power
MC1568G	−55° C to +125° C	Metal Can
MC1568L	−55° C to +125° C	Ceramic DIP
MC1568R	−55° C to +125° C	Metal Power

MAXIMUM RATINGS (T_C = +25°C unless otherwise noted.)

Rating	Symbol	Value			Unit
Input Voltage	$V_{CC}, \lvert V_{EE} \rvert$	30			Vdc
Peak Load Current	I_{PK}	100			mA
Power Dissipation and Thermal Characteristics		G Package	R Package	L Package	
T_A = +25°C	P_D	0.8	2.4	1.0	Watts
Derate above T_A = +25°C	$1/\theta_{JA}$	6.6	28.5	10	mW/°C
Thermal Resistance, Junction to Air	θ_{JA}	150	35	100	°C/W
T_C = +25°C	P_D	2.1	9.0	2.5	Watts
Derate above T_C = +25°C	$1/\theta_{JC}$	14	61	20	mW/°C
Thermal Resistance, Junction to Case	θ_{JC}	70	17	50	°C/W
Storage Junction Temperature Range	T_J, T_{stg}	-65 to +175			°C
Minimum Short-Circuit Resistance	R_{SC}(min)	4.0			Ohms

OPERATING TEMPERATURE RANGE

Ambient Temperature	T_A		°C
MC1468		0 to +70	
MC1568		-55 to +125	

ELECTRICAL CHARACTERISTICS (V_{CC} = +20 V, V_{EE} = -20 V, C1 = C2 = 1500 pF, C3 = C4 = 1.0 μF, $R_{SC}^+ = R_{SC}^-$ = 4.0 Ω, $I_L^+ = I_L^-$ = 0, T_C = +25°C unless otherwise noted.) (See Figure 1.)

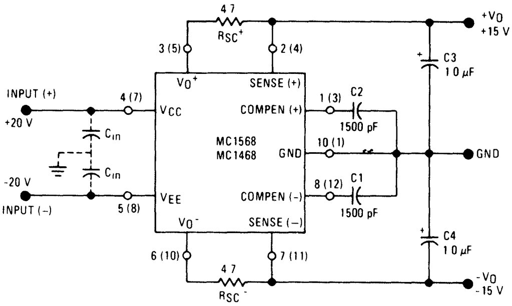
		MC1568			MC1468			
Characteristic	Symbol*	Min	Typ	Max	Min	Typ	Max	Unit
Output Voltage	V_O	±14.8	±15	±15.2	±14.5	±15	±15.5	Vdc
Input Voltage	V_{in}	–	–	±30	–	–	±30	Vdc
Input-Output Voltage Differential	$\lvert V_{in} - V_O \rvert$	2.0	–	–	2.0	–	–	Vdc
Output Voltage Balance	V_{Bal}	–	±50	±150	–	±50	±300	mV
Line Regulation Voltage	Reg_{in}							mV
(V_{in} = 18 V to 30 V)		–	–	10	–	–	10	
(T_{low}① to T_{high}②)		–	–	20	–	–	20	
Load Regulation Voltage	Reg_L							mV
(I_L = 0 to 50 mA, T_J = constant)		–	–	10	–	–	10	
(T_A = T_{low} to T_{high})		–	–	30	–	–	30	
Output Voltage Range	V_{OR}							Vdc
L Package (See Figure 4.)		±8.0	–	±20	±8.0	–	±20	
R and G Packages (See Figures 2 and 13.)		±14.5	–	±20	±14.5	–	±20	
Ripple Rejection (f = 120 Hz)	RR	–	75	–	–	75	–	dB
Output Voltage Temperature Stability (T_{low} to T_{high})	$\lvert TS_{V_O} \rvert$	–	0.3	1.0	–	0.3	1.0	%
Short-Circuit Current Limit (R_{SC} = 10 ohms)	I_{SC}	–	60	–	–	60	–	mA
Output Noise Voltage (BW = 100 Hz - 10 kHz)	V_N	–	100	–	–	100	–	μV(RMS)
Positive Standby Current (V_{in} = +30 V)	I_B^+	–	2.4	4.0	–	2.4	4.0	mA
Negative Standby Current (V_{in} = -30 V)	I_B^-	–	1.0	3.0	–	1.0	3.0	mA
Long-Term Stability	$\Delta V_O/\Delta t$	–	0.2	–	–	0.2	–	%/k Hr

① T_{low} = 0°C for MC1468
= -55°C for MC1568

② T_{high} = +70°C for MC1468
= +125°C for MC1568

TYPICAL APPLICATIONS

FIGURE 1 – BASIC 50-mA REGULATOR

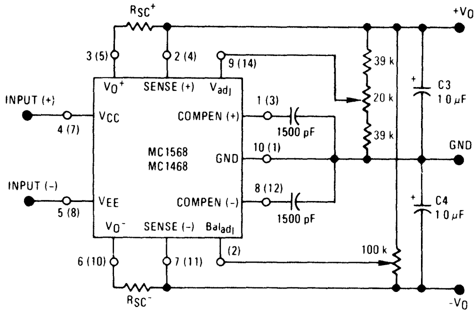


C1 and C2 should be located as close to the device as possible A 0 1 μF ceramic capacitor (C_{in}) may be required on the input lines if the device is located an appreciable distance from the rectifier filter capacitors

C3 and C4 may be increased to improve load transient response and to reduce the output noise voltage At low temperature operation, it may be necessary to bypass C4 with a 0 1 μF ceramic disc capacitor

FIGURE 2 – VOLTAGE ADJUST AND BALANCE ADJUST CIRCUIT

(14.5 V $\leqslant V_{out} \leqslant$ 20 V)



Balance adjust available in MC1568L, MC1468L ceramic dual in line package only

FIGURE 3 – ±1.5-AMPERE REGULATOR

(Short-Circuit Protected, with Proper Heatsinking)
(Metal-Packaged Devices Only, R Suffix)

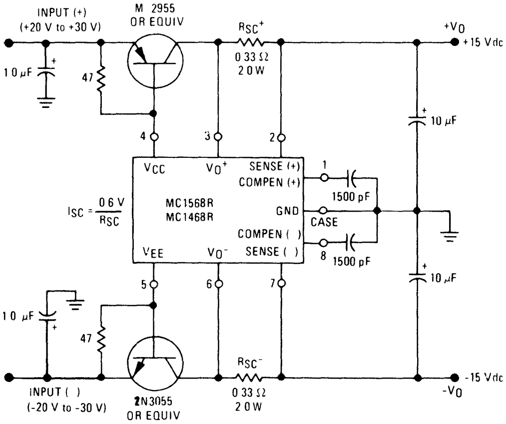
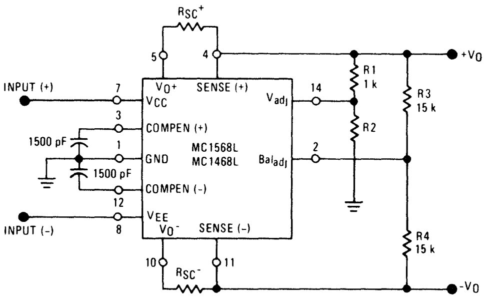


FIGURE 4 – OUTPUT VOLTAGE ADJUSTMENT FOR 8.0 V $\leqslant |\pm V_O| \leqslant$ 14.5 V

(Ceramic-Packaged Devices Only, L Suffix.)



The presence of the Bal_{adj}, pin 2, on devices housed in the dual in line package (L suffix) allows the user to adjust the output voltages down to +8 0 V The required value of resistor R2 can calculated from

$$R2 = \frac{R1\ R_{int}\ (\phi + V_z)}{R_{int}\ (V_O - \phi - V_z) - \phi\ R1}$$

Where R_{int} = An Internal Resistor = R1 = 1 kΩ
ϕ = 0 68 V
V_z = 6 6 V

Some common design values are listed below

+VO, V)	R2	TC VO (%/°C)	IB + (mA)
14	1 2 k	0 003	10
12	1 8 k	0 022	7 2
10	3 5 k	0 025	5 0
8 0	∞	0 028	2 6

TYPICAL CHARACTERISTICS

(V_{CC} = +20 V, V_{EE} = -20 V, V_O = ±15 V, T_A = +25°C unless otherwise noted.)

FIGURE 5 – LOAD REGULATION

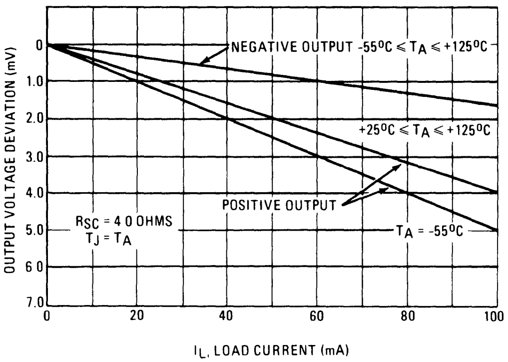


FIGURE 6 – REGULATOR DROPOUT VOLTAGE

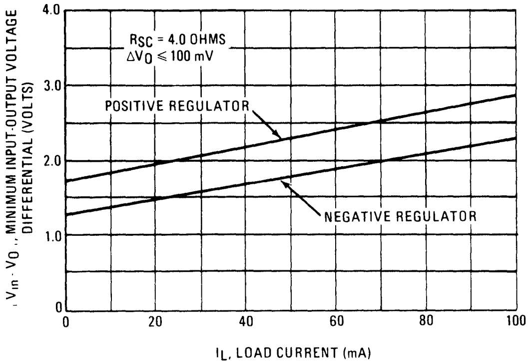


FIGURE 7 – MAXIMUM CURRENT CAPABILITY

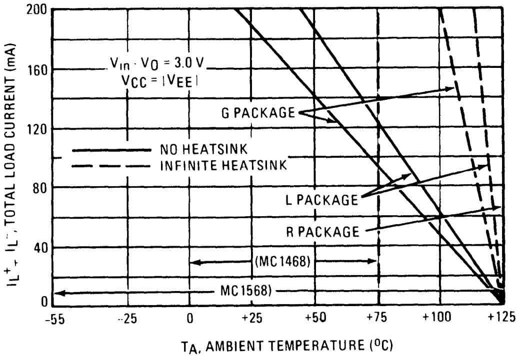


FIGURE 8 – MAXIMUM CURRENT CAPABILITY

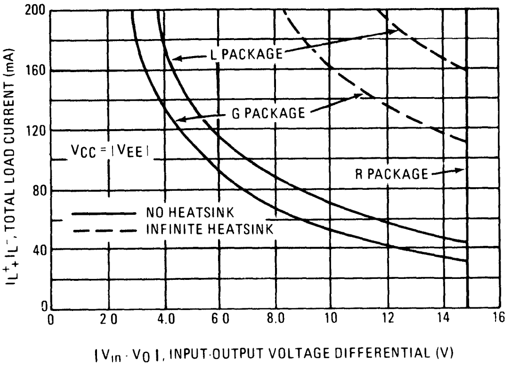


FIGURE 9 – I_{SC} versus R_{SC}

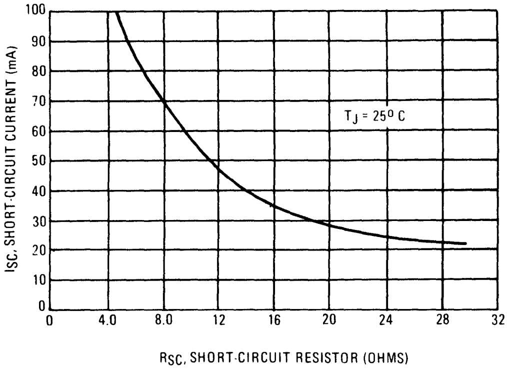
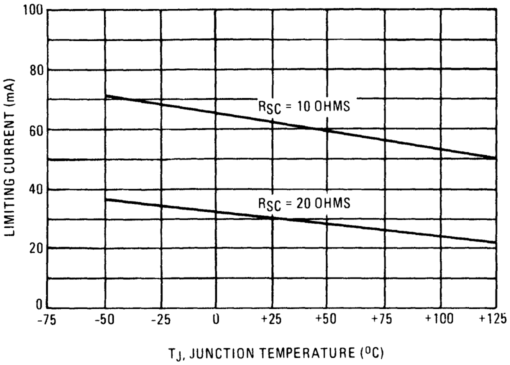


FIGURE 10 – CURRENT-LIMITING CHARACTERISTICS



TYPICAL CHARACTERISTICS (continued)

(V_{CC} = +20 V, V_{EE} = -20 V, V_O = ±15 V, T_A = +25°C unless otherwise noted.)

FIGURE 11 – STANDBY CURRENT DRAIN

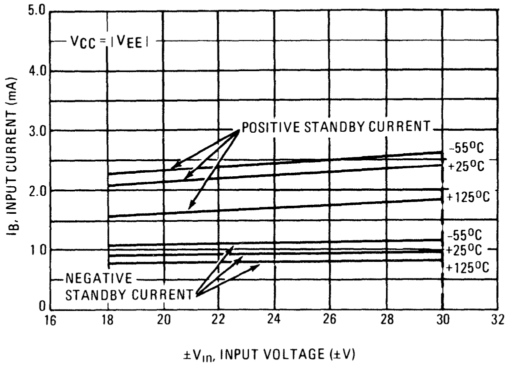


FIGURE 12 – STANDBY CURRENT DRAIN

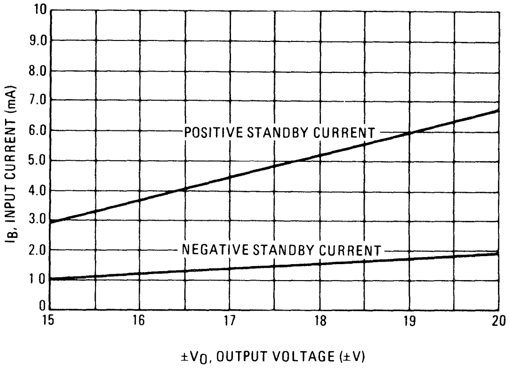


FIGURE 13 – TEMPERATURE COEFFICIENT OF OUTPUT VOLTAGE

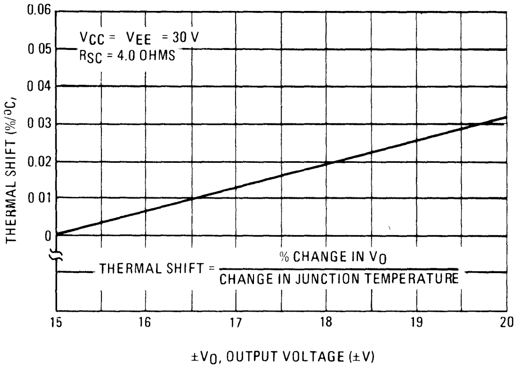


FIGURE 14 – LOAD TRANSIENT RESPONSE

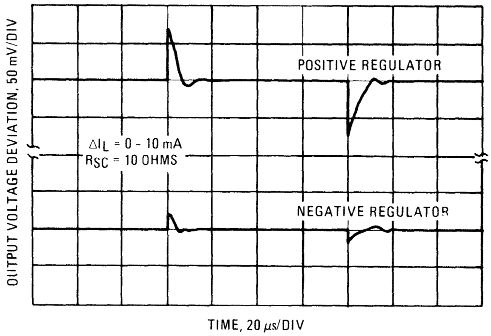


FIGURE 15 – LINE TRANSIENT RESPONSE

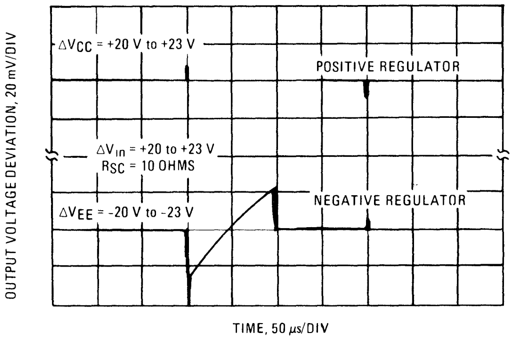
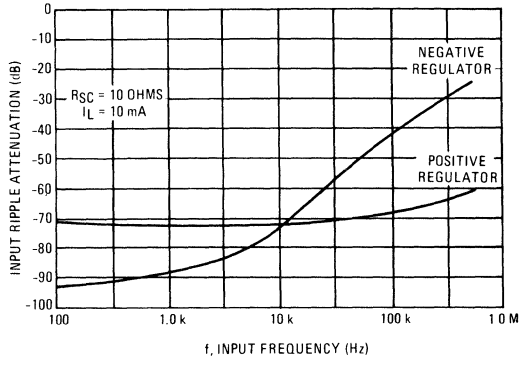


FIGURE 16 – RIPPLE REJECTION



MOTOROLA

MC1469
MC1569

Specifications and Applications Information

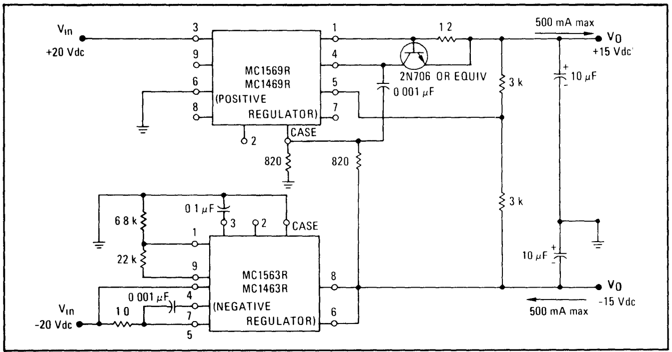
MONOLITHIC VOLTAGE REGULATOR

The MC1569/MC1469 is a positive voltage regulator designed to deliver continuous load current up to 500 mAdc. Output voltage is adjustable from 2.5 Vdc to 37 Vdc. The MC1569 is specified for use within the military temperature range (−55 to +125°C) and the MC1469 within the 0 to +70°C temperature range.

For systems requiring a positive regulated voltage, the MC1569 can be used with performance nearly identical to the MC1563 negative voltage regulator. Systems requiring both a positive and negative regulated voltage can use the MC1569 and MC1563 as complementary regulators with a common input ground.

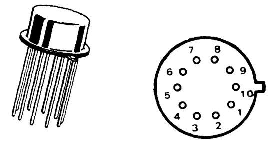
- Electronic "Shut-Down" Control
- Excellent Load Regulation (Low Output Impedance – 20 milli-ohms typ)
- High Power Capability: up to 17.5 Watts
- Excellent Temperature Stability: ±0.002 %/°C typ
- High Ripple Rejection: 0.002 %/V typ

FIGURE 1 – ±15 V, ±400 mA COMPLEMENTARY TRACKING VOLTAGE REGULATOR



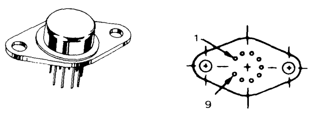
POSITIVE VOLTAGE REGULATOR INTEGRATED CIRCUIT

SILICON NONOLITHIC
EPITAXIAL PASSIVATED



CASE 603
METAL PACKAGE
G SUFFIX

(Bottom View)



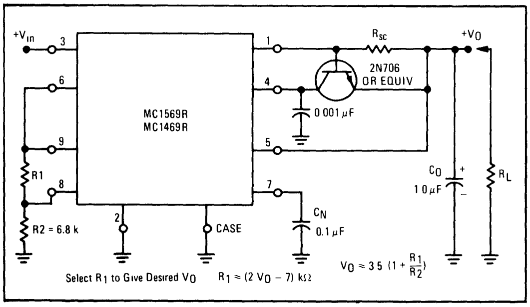
CASE 614
METAL PACKAGE
R SUFFIX

(bottom view)

ORDERING INFORMATION

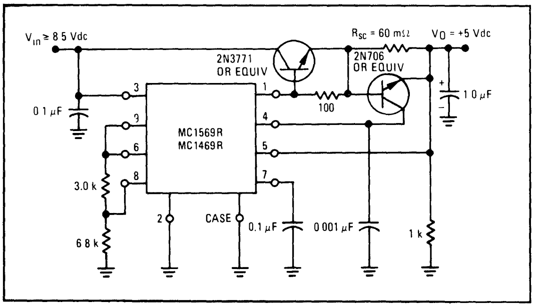
DEVICE	TEMPERATURE RANGE	PACKAGE
MC1469G	0° C to +70°C	Metal Can
MC1469R	0° C to +70° C	Metal Power
MC1569G	−55° C to +125° C	Metal Can
MC1569R	−55° C to +125° C	Metal Power

FIGURE 2 – TYPICAL CIRCUIT CONNECTION ($3.5 \leqslant V_O \leqslant 37$ Vdc, $1 \leqslant I_L \leqslant 500$ mA)



Select R_1 to Give Desired V_O $R_1 \approx (2\,V_O - 7)$ kΩ $V_O \approx 3.5\,(1 + \frac{R_1}{R_2})$

FIGURE 3 – TYPICAL NPN CURRENT BOOST CONNECTION (V_O = 5.0 Vdc, I_L = 10 Adc [max])



MAXIMUM RATINGS ($T_C = +25^oC$ unless otherwise noted)

Rating	Symbol	Value		Unit
		G Package	R Package	
Input Voltage	V_{in}			Vdc
MC1469		35		
MC1569		40		
Peak Load Current	I_{PK}	250	600	mA
Current, Pin 2	$I_{pin\ 2}$	10	10	mA
Current, Pin 9	$I_{pin\ 9}$	5.0	5.0	
Power Dissipation and Thermal Characteristics				
$T_A = +25^oC$	P_D	0.68	3.0	Watts
Derate above $T_A = +25^oC$	$1/\theta_{JA}$	5.44	24	mW/^oC
Thermal Resistance, Junction to Air	θ_{JA}	184	41.6	$^oC/W$
$T_C = +25^oC$	P_D	1.8	14	Watts
Derate above $T_C = +25^oC$	$1/\theta_{JC}$	14.4	140	mW/^oC
Thermal Resistance, Junction to Case	θ_{JC}	69.4	7.15	$^oC/W$
Operating and Storage Junction Temperature	T_J, T_{stg}	-65 to +150		^oC

OPERATING TEMPERATURE RANGE

Ambient Temperature	T_A		^oC
MC1469		0 to +70	
MC1569		-55 to +125	

ELECTRICAL CHARACTERISTICS

($T_C = +25^oC$ unless otherwise noted) (Load Current = 100 mA for "R" Package device, = 10 mA for "G" Package device, unless otherwise noted)

Characteristic	Fig.	Note	Symbol	MC1569			MC1469			Unit
				Min	Typ	Max	Min	Typ	Max	
Input Voltage ($T_A = T_{low}$ ① to T_{high} ②)	4	1	V_{in}	8.5	–	40	9.0	–	35	Vdc
Output Voltage Range	4,5		V_O	2.5	–	37	2.5	–	32	Vdc
Reference Voltage (Pin 8 to Ground, V_{in} = 15 V	4		V_{ref}	3.4	3.5	3.6	3.2	3.5	3.8	Vdc
Minimum Input-Output Voltage Differential ($R_{sc} = 0$)	4	2	$V_{in} - V_O$	–	2.1	2.7	–	2.1	3.0	Vdc
Bias Current (V_{in} = 15 V) (I_L = 1.0 mAdc, R_2 = 6.8 k ohms, $I_{IB} = I_{in} - I_L$)	4		I_{IB}	–	4.0	9.0	–	5.0	12	mAdc
Output Noise (C_N = 0.1 μF, f = 10 Hz to 5.0 MHz)	4		v_N	–	0.150	–	–	0.150	–	mV(rms)
Temperature Coefficient of Output Voltage	4	3	TCV_O	–	±0.002	–	–	±0.002	–	%/^oC
Operating Load Current Range										mAdc
($R_{sc} \leqslant 0.3$ ohms) R Package	4		I_L	1.0	–	500	1.0	–	500	
($R_{sc} \leqslant 2.0$ ohms) G Package				1.0	–	200	1.0	–	200	
Input Regulation	6	4	Reg_{in}	–	0.002	0.015	–	0.003	0.030	%/V_O
Load Regulation	7	5	Reg_{load}							
(T_J = Constant [1.0 mA$\leq I_L \leq$20 mA])				–	0.4	1.6	–	0.7	2.4	mV
($T_C = +25^oC$ [1.0 mA$\leq I_L \leq$50 mA]) R Package				–	0.005	0.05	–	0.005	0.05	%
G Package				–	0.01	0.13	–	0.01	0.13	
Output Impedance (C_c = 0.001 μF, R_{sc} = 1.0 ohm, f = 1.0 kHz, V_{in} = +14 Vdc, V_O = +10 Vdc)	8	6	z_O	–	20	–	–	35	–	milliohms
Shutdown Current (V_{in} = +35 Vdc)	9		I_{sd}	–	70	150	–	140	500	μAdc

① T_{low} = 0^oC for MC1469
= -55^oC for MC1569

② T_{high} = $+70^o$ C for MC1469
= $+125^oC$ for MC1569

Note 1. "Minimum Input Voltage" is the minimum "total instantaneous input voltage" required to properly bias the internal zener reference diode. For output voltages greater than approximately 5.5 Vdc the minimum "total instantaneous input voltage" must increase to the extent that it will always exceed the output voltage by at least the "input-output voltage differential".

Note 2. This parameter states that the MC1569/MC1469 will regulate properly with the input-output voltage differential $(V_{in} - V_O)$ as low as 2.7 Vdc and 3.0 Vdc respectively. Typical units will regulate properly with $(V_{in} - V_O)$ as low as 2.1 Vdc as shown in the typical column. (See Figure 21.)

Note 3. "Temperature Coefficient of Output Voltage" is defined as:

$$\text{MC1569, } TCV_O = \frac{\pm (V_O \text{ max} - V_O \text{ min})\ (100)}{(180^\circ C)\ (V_O\ @\ 25^\circ C)} = \%/^\circ C$$

$$\text{MC1469, } TCV_O = \frac{\pm (V_O \text{ max} - V_O \text{ min})\ (100)}{(75^\circ C)\ (V_O\ @\ 25^\circ C)} = \%/^\circ C$$

The output-voltage adjusting resistors (R1 and R2) must have matched temperature characteristics in order to maintain a constant ratio independent of temperature.

Note 4. Input regulation is the percentage change in output voltage per volt change in the input voltage and is expressed as

$$\text{Input Regulation} = \frac{v_O}{V_O\ (v_{in})}\ 100\ (\%/V_O),$$

where v_O is the change in the output voltage V_O for the input change v_{in}.

The following example illustrates how to compute maximum output voltage change for the conditions given:

$$Reg_{in} = 0.015\ \%/V_O$$
$$V_O = 10\text{ Vdc}$$
$$v_{in} = 1.0\text{ V(rms)}$$
$$v_o = \frac{(Reg_{in})\ (v_{in})\ (V_O)}{100}$$
$$= \frac{(0.015)\ (1.0)\ (10)}{100}$$
$$= 0.0015\text{ V(rms)}$$

Note 5. Load regulation is specified for small ($\leqslant$+17°C) changes in junction temperature. Temperature drift effect must be taken into account separately for conditions of high junction temperature changes due to the thermal feedback that exists on the monolithic chip.

$$\text{Load Regulation} = \frac{\left[V_O|I_L = 1.0\text{ mA}\right] - \left[V_O|I_L = 50\text{ mA}\right]}{V_O\ |I_L = 1.0\text{ mA}}\ X\ 100$$

Note 6. The resulting low level output signal (v_O) will require the use of a tuned voltmeter to obtain a reading. Special care should be used to insure that the measurement technique does not include connection resistance, wire resistance, and wire lead inductance (i.e., measure close to the case). Note that No. 22 AWG hook-up wire has approximately 4.0 milliohms/in. dc resistance and an inductive reactance of approximately 10 milliohms/in. at 100 kHz. Avoid use of alligator clips or banana plug-jack combination.

TEST CIRCUITS

FIGURE 4 – CONNECTION FOR $V_O \geqslant$3.5 Vdc

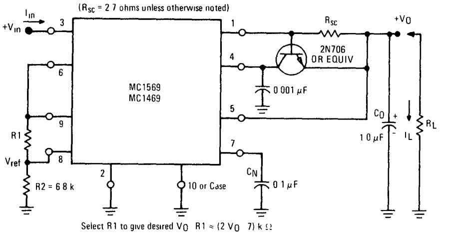


FIGURE 5 – CONNECTION FOR 2.5 Vdc $\geqslant V_O \leqslant$3.5 Vdc

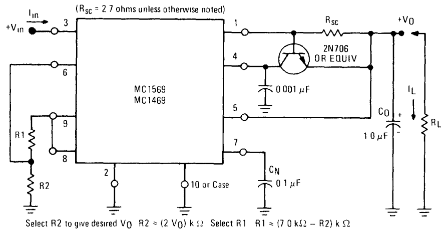


FIGURE 6 – INPUT REGULATION

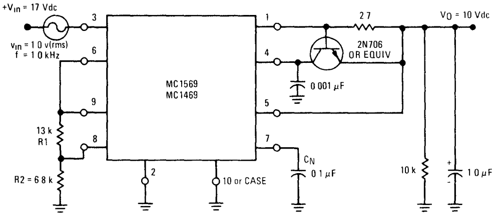


FIGURE 7 – LOAD REGULATION

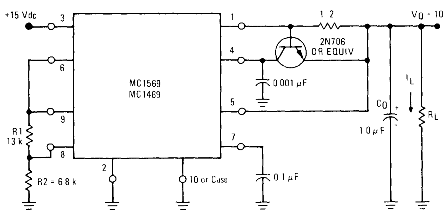


FIGURE 8 – OUTPUT IMPEDANCE

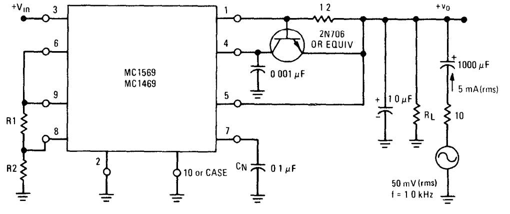
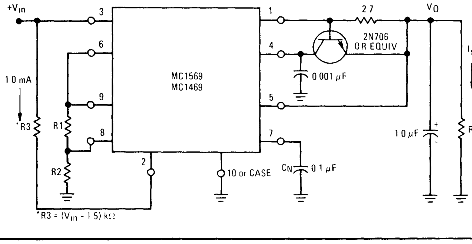


FIGURE 9 – SHUTDOWN CURRENT



GENERAL DESIGN INFORMATION

1. Output Voltage, V_O
   a) For $V_O \geqslant 3.5$ Vdc – Output voltage is set by resistors R1 and R2 (see Figure 4). Set R2 = 6.8 k ohms and determine R1 from the graph of Figure 10 or from the equation:

   $$R1 \approx (2\ V_O - 7)\ k\Omega$$

   b) For $2.5 \leqslant V_O \leqslant 3.5$ Vdc – Output voltage is set by resistors R1 and R2 (see Figure 5). Resistors R1 and R2 can be determined from the graph of Figure 11 or from the equations:

   $$R2 \approx 2\ (V_O)\ k\Omega$$
   $$R1 \approx (7\ k\Omega - R2)\ k\Omega$$

   c) Output voltage, V_O, is determined by the ratio of R1 and R2, therefore optimum temperature performance can be achieved if R1 and R2 have the same temperature coefficient.
   d) Output voltage can be varied by making R1 adjustable as shown in Figure 43.
   e) If V_O = 3.5 Vdc (to supply MRTL*for example), tie pins 6, 8 and 9 together. R1 and R2 are not needed in this case.
2. Short Circuit Current, I_{sc}
   Short Circuit Current, I_{sc}, is determined by R_{sc}. R_{sc} may be chosen with the aid of Figure 12 or the expression:

   $$R_{sc} \approx \frac{0.6}{I_{sc}} \text{ ohm}$$

   where I_{sc} is measured in amperes. This expression is also valid when current is boosted as shown in Figure 2.
3. Compensation, C_c
   A 0.001 μF capacitor, C_c, from pin 4 to ground will provide adequate compensation in most applications, with or without current boost. Smaller values of C_c will reduce stability and larger values of C_c will degrade pulse response and output impedance versus frequency. The physical location of C_c should be close to the MC1569/MC1469 with short lead lengths.
4. Noise Filter Capacitor, C_N
   A 0.1 μF capacitor, C_N, from pin 7 to ground will typically reduce the output noise voltage to 150 μV(rms). The value of C_N can be increased or decreased, depending on the noise voltage requirements of a particular application. A minimum value of 0.001 μF is recommended.
5. Output Capacitor, C_O
   The value of C_O should be at least 1.0 μF in order to provide good stability. The maximum value recommended is a function of current limit resistor R_{sc}:

   $$C_O \text{ max} \approx \frac{250\ \mu F}{R_{sc}}$$

   where R_{sc} is measured in ohms. Values of C_O greater than this will degrade the pulse response characteristics and increase the settling time.
6. Shut-Down Control
   One method of turning "OFF" the regulator is to apply a dc voltage at pin 2. This control can be used to eliminate power consumption by circuit loads which can be put in "standby" mode. Examples include, an ac or dc "squelch" control for communications circuits, and a dissipation control to protect the regulator under sustained output short-circuiting. As the magnitude of the input-threshold voltage at Pin 2 depends directly upon the junction temperature of the integrated circuit chip, a fixed dc voltage at Pin 2 will cause automatic shut-down for high junction temperatures. This will protect the chip, independent of the heat sinking used, the ambient temperature, or the input or output voltage levels. Standard Logic levels of MRTL, MDTL* or MTTL* can also be used to turn the regulator "ON" or "OFF".
7. Remote Sensina
   The connection to pin 5 can be made with a separate lead direct to the load. Thus, "remote sensing" can be achieved and the effect of undesired impedances (including that of the milliammeter used to measure I_L) on z_O can be greatly reduced.

FIGURE 10 – R1 versus V_O
($V_O \geqslant 3.5$ Vdc, See Figure 4)

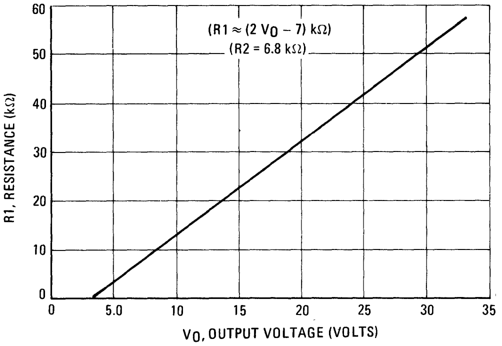


FIGURE 11 – R1 and R2 versus V_O
($2.5 \leqslant V_O \leqslant 3.5$ Vdc, See Figure 5)

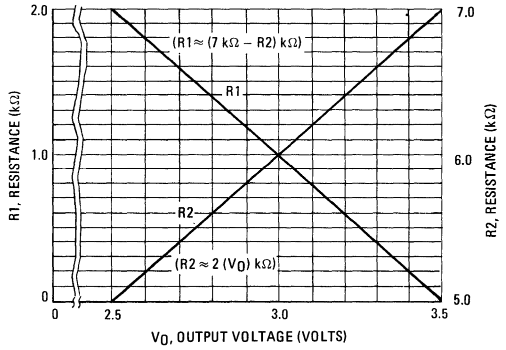
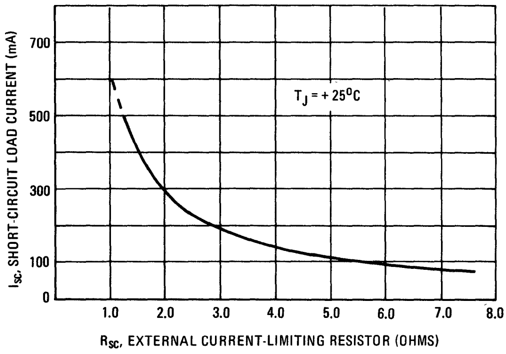


FIGURE 12 – I_{sc} versus R_{sc}



TYPICAL CHARACTERISTICS

Unless otherwise noted: C_N = 0.1 µF, C_C = 0.001 µF, C_O = 1.0 µF, T_C = +25°C,
V_{in} nom = +9.0 Vdc, V_O nom = +5.0 Vdc,
I_L > 200 mA for R package only.

FIGURE 13 – DEPENDENCE OF OUTPUT IMPEDANCE ON OUTPUT VOLTAGE

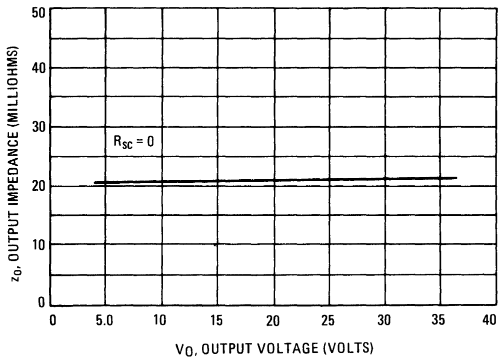


FIGURE 14 – OUTPUT IMPEDANCE versus R_{sc}

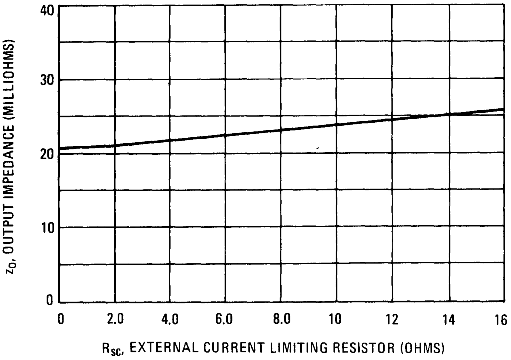


FIGURE 15 – FREQUENCY DEPENDENCE OF INPUT REGULATION, C_O = 10 µF

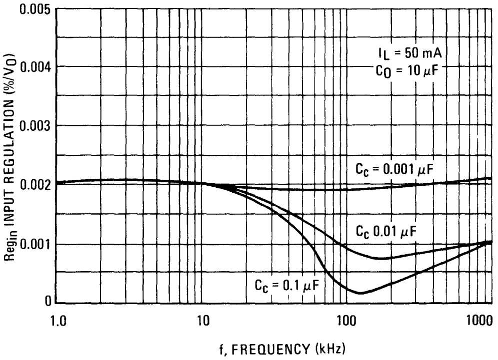


FIGURE 16 – FREQUENCY DEPENDENCE OF INPUT REGULATION, C_O = 2.0 µF

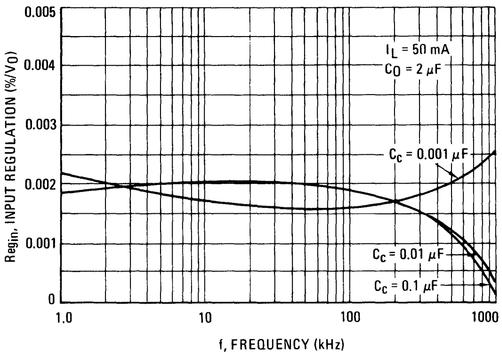


FIGURE 17 – CURRENT-LIMITING CHARACTERISTICS

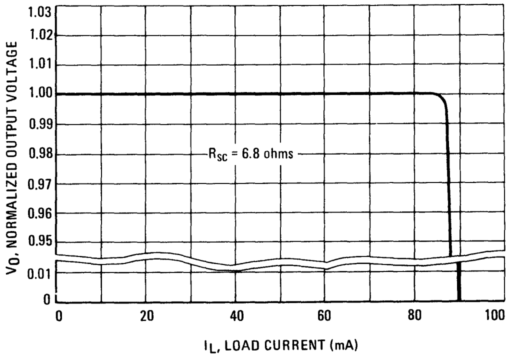
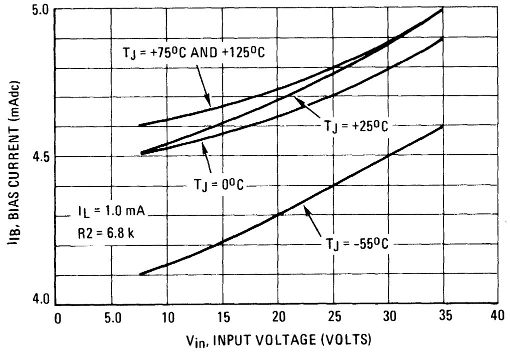


FIGURE 18 – BIAS CURRENT versus INPUT VOLTAGE



TYPICAL CHARACTERISTICS (continued)

Unless otherwise noted: $C_N = 0.1\ \mu F$, $C_c = 0.001\ \mu F$, $C_O = 1.0\ \mu F$, $T_C = +25°C$,
V_{in} nom = +9.0 Vdc, V_O nom = +5.0 Vdc,
$I_L > 200$ mA for R package only.

FIGURE 19 – EFFECT OF LOAD CURRENT ON INPUT-OUTPUT VOLTAGE DIFFERENTIAL

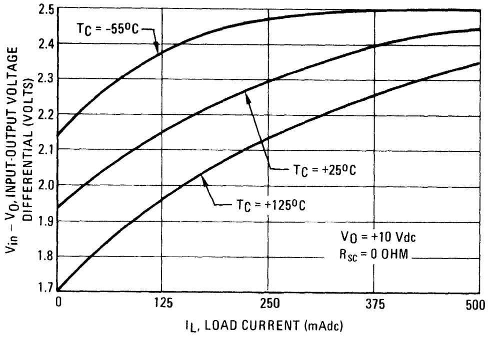


FIGURE 20 – EFFECT OF INPUT-OUTPUT VOLTAGE DIFFERENTIAL ON INPUT REGULATION

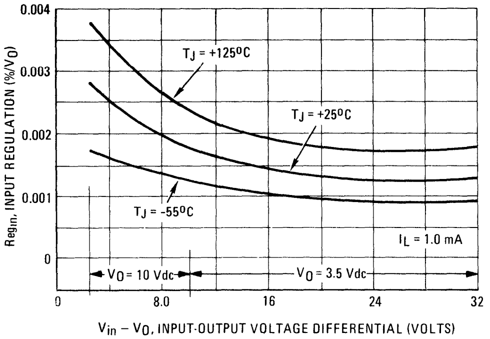


FIGURE 21 – INPUT TRANSIENT RESPONSE

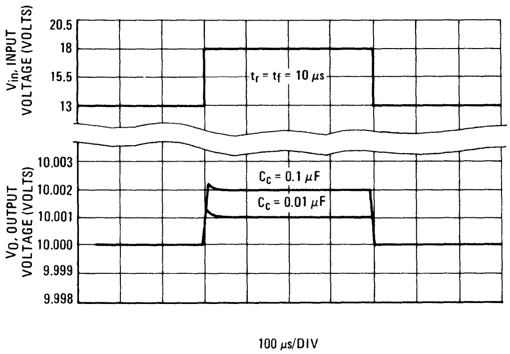


FIGURE 22 – TEMPERATURE DEPENDENCE OF SHORT-CIRCUIT LOAD CURRENT

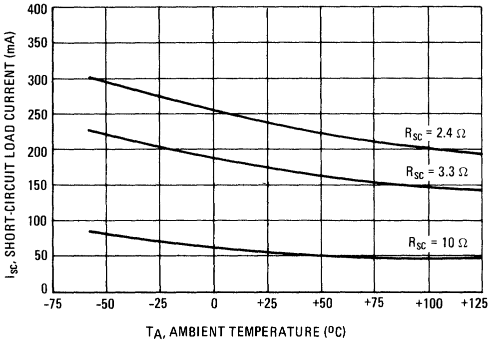


FIGURE 23 – FREQUENCY DEPENDENCE OF OUTPUT IMPEDANCE, $C_O = 10\ \mu F$

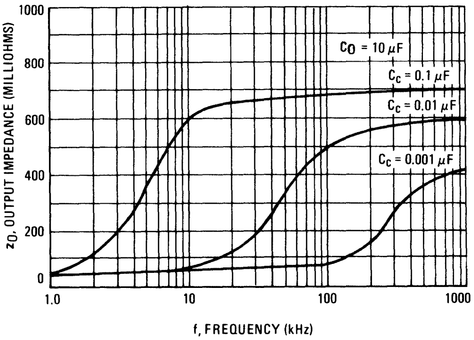
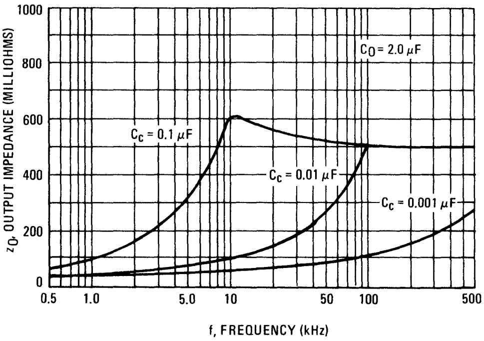


FIGURE 24 – FREQUENCY DEPENDENCE OF OUTPUT IMPEDANCE, $C_O = 2.0\ \mu F$



MOTOROLA

MC1723
MC1723C

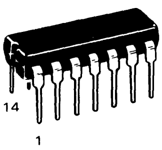
MONOLITHIC VOLTAGE REGULATOR

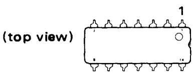
The MC1723 is a positive or negative voltage regulator designed to deliver load current to 150 mAdc. Output current capability can be increased to several amperes through use of one or more external pass transistors. MC1723 is specified for operation over the military temperature range (-55°C to +125°C) and the MC1723C over the commercial temperature range (0 to +70°C)

- Output Voltage Adjustable from 2 Vdc to 37 Vdc
- Output Current to 150 mAdc Without External Pass Transistors
- 0.01% Line and 0.03% Load Regulation
- Adjustable Short-Circuit Protection

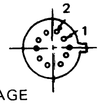
VOLTAGE REGULATOR

SILICON
MONOLITHIC
INTEGRATED CIRCUIT



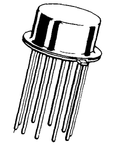


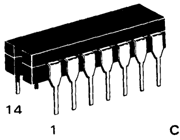
P SUFFIX
PLASTIC PACKAGE
CASE 646

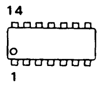


(top view)

G SUFFIX
METAL PACKAGE
CASE 603
(TO-100 Type)





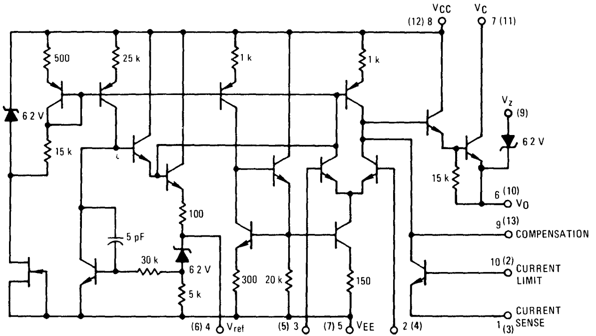


L SUFFIX
CERAMIC PACKAGE
CASE 632
(TO-116)

ORDERING INFORMATION

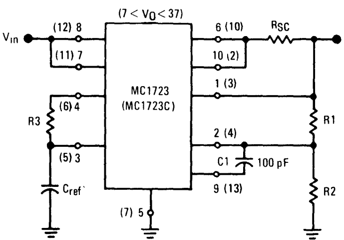
Device	Alternate	Temperature Range	Package
MC1723CG	LM723CH, µA723HC	0°C to 70°C	Metal Can
MC1723CL	LM723CJ, µA723DC	0°C to +70°C	Ceramic DIP
MC1723CP	LM723CN, µA723PC	0°C to +70°C	Plastic DIP
MC1723G	–	-55°C to +125°C	Metal Can
MC1723L	–	-55°C to +125°C	Ceramic DIP

FIGURE 1 – CIRCUIT SCHEMATIC



PIN NUMBERS ADJACENT TO TERMINALS ARE FOR THE METAL, PACKAGE
PIN NUMBERS IN PARENTHESIS ARE FOR DUAL IN LINE PACKAGES

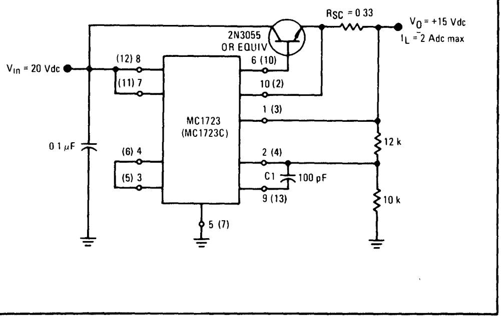
FIGURE 2 – TYPICAL CIRCUIT CONNECTION



$V_O \cong 7\left(\frac{R1 + R2}{R2}\right)$ $I_{SC} = \frac{V_{sense}}{R_{SC}} = \frac{0.66}{R_{SC}}$ at $T_J = +25°C$

For best results 10 k < R2 < 100 k
For minimum drift R3 = R1||R2

FIGURE 3 – TYPICAL NPN CURRENT BOOST CONNECTION



MAXIMUM RATINGS (T_A = +25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Pulse Voltage from V_{CC} to V_{EE} (50 ms)	$V_{in(p)}$	50	V_{peak}
Continuous Voltage from V_{CC} to V_{EE}	V_{in}	40	Vdc
Input-Output Voltage Differential	$V_{in} - V_O$	40	Vdc
Maximum Output Current	I_L	150	mAdc
Current from V_{ref}	I_{ref}	15	mAdc
Current from V_z	I_z	25	mA
Voltage Between Non-Inverting Input and V_{EE}	V_{ie}	8.0	Vdc
Differential Input Voltage	V_{id}	±5.0	Vdc
Power Dissipation and Thermal Characteristics			
Plastic Package			
T_A = +25°C	P_D	1.25	W
Derate above T_A = +25°C	$1/\theta_{JA}$	10	mW/°C
Thermal Resistance, Junction to Air	θ_{JA}	100	°C/W
Metal Package			
T_A = +25°C	P_D	1.0	Watt
Derate above T_A = +25°C	$1/\theta_{JA}$	6.6	mW/°C
Thermal Resistance, Junction to Air	θ_{JA}	150	°C/W
T_C = +25°C	P_D	2.1	Watts
Derate above T_A = +25°C	$1/\theta_{JA}$	14	mW/°C
Thermal Resistance, Junction to Case	θ_{JC}	35	°C/W
Dual In-Line Ceramic Package	P_D	1.5	Watt
Derate above T_A = +25°C	$1/\theta_{JA}$	10	mW/°C
Thermal Resistance, Junction to Air	θ_{JA}	100	°C/W
Operating and Storage Junction Temperature Range	T_J, T_{stg}		°C
Metal Package		-65 to +150	
Dual In-Line Ceramic and Ceramic Flat Packages		-65 to +175	
Operating Ambient Temperature Range	T_A		°C
MC1723C		0 to +70	
MC1723		-55 to +125	

ELECTRICAL CHARACTERISTICS (Unless otherwise noted: T_A = +25°C, V_{in} 12 Vdc, V_O = 5.0 Vdc, I_L = 1.0 mAdc, R_{SC} = 0, C1 = 100 pF, C_{ref} = 0 and divider impedance as seen by the error amplifier ⩽ 10 kΩ connected as shown in Figure 2)

Characteristic	Symbol	MC1723			MC1723C			Unit
		Min	Typ	Max	Min	Typ	Max	
Input Voltage Range	V_{in}	9.5	–	40	9.5	–	40	Vdc
Output Voltage Range	V_O	2.0	–	37	2.0	–	37	Vdc
Input-Output Voltage Differential	$V_{in}-V_O$	3.0	–	38	3.0	–	38	Vdc
Reference Voltage	V_{ref}	6.95	7.15	7.35	6.80	7.15	7.50	Vdc
Standby Current Drain (I_L = 0, V_{in} = 30 V)	I_{IB}	–	2.3	3.5	–	2.3	4.0	mAdc
Output Noise Voltage (f = 100 Hz to 10 kHz)	V_N							µV(RMS)
C_{ref} = 0		–	20	–	–	20	–	
C_{ref} = 5.0 µF		–	2.5	–	–	2.5	–	
Average Temperature Coefficient of Output Voltage (T_{low} ① $< T_A < T_{high}$ ②)	TCV_O	–	0.002	0.015	–	0.003	0.015	%/°C
Line Regulation	Reg_{in}							$\%V_O$
(T_A = +25°C) $12\text{ V} < V_{in} < 15\text{ V}$		–	0.01	0.1	–	0.01	0.1	
(T_A = +25°C) $12\text{ V} \leq V_{in} < 40\text{ V}$		–	0.02	0.2	–	0.1	0.5	
(T_{low} ① $< T_A < T_{high}$ ②) $12\text{ V} < V_{in} < 15\text{ V}$		–	–	0.3	–	–	0.3	
Load Regulation ($1.0\text{ mA} < I_L < 50\text{ mA}$)	Reg_{load}							$\%V_O$
T_A = +25°C		–	0.03	0.15	–	0.03	0.2	
T_{low} ① $< T_A < T_{high}$ ②		–	–	0.6	–	–	0.6	
Ripple Rejection (f = 50 Hz to 10 kHz)	Rej_R							dB
C_{ref} = 0		–	74	–	–	74	–	
C_{ref} = 5.0 µF		–	86	–	–	86	–	
Short Circuit Current Limit (R_{SC} = 10 Ω, V_O = 0)	I_{SC}	–	65	–	–	65	–	mAdc
Long Term Stability	$\Delta V_O/\Delta t$	–	0.1	–	–	0.1	–	%/1000 Hr

① T_{low} = 0°C for MC1723C
= -55°C for MC1723

② T_{high} = +70° C for MC1723C
= +125°C for MC1723

TYPICAL CHARACTERISTICS

(V_{in} = 12 Vdc, V_O = 5.0 Vdc, I_L = 1.0 mAdc, R_{SC} = 0, T_A = +25°C unless otherwise noted.)

FIGURE 4 – MAXIMUM LOAD CURRENT AS A FUNCTION OF INPUT-OUTPUT VOLTAGE DIFFERENTIAL

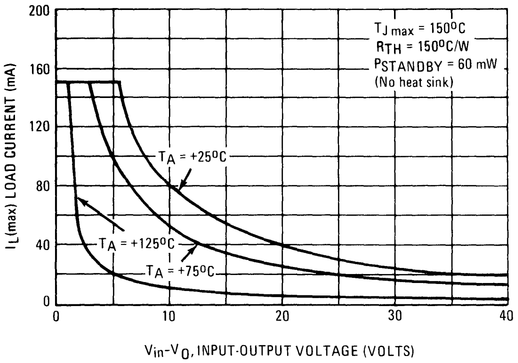


FIGURE 5 – LOAD REGULATION CHARACTERISTICS WITHOUT CURRENT LIMITING

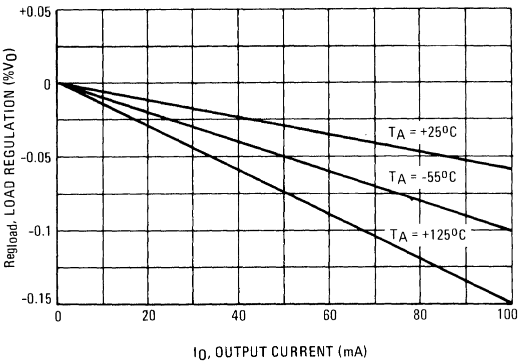


FIGURE 6 – LOAD REGULATION CHARACTERISTICS WITH CURRENT LIMITING

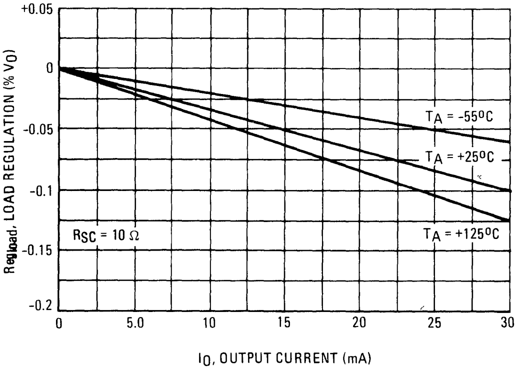


FIGURE 7 – LOAD REGULATION CHARACTERISTICS WITH CURRENT LIMITING

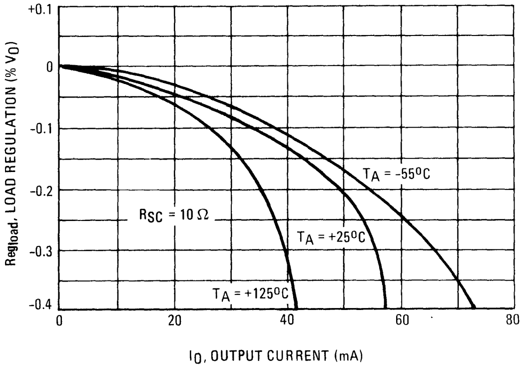


FIGURE 8 – CURRENT LIMITING CHARACTERISTICS

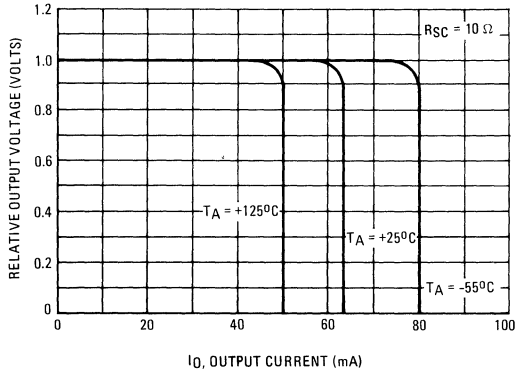
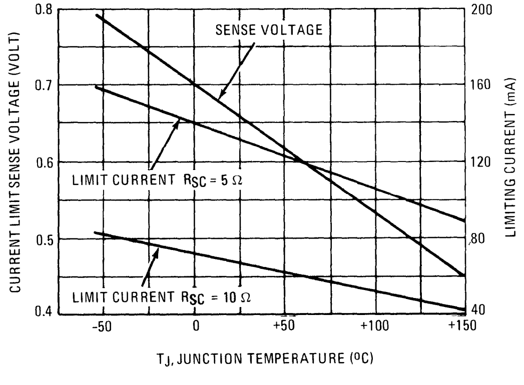


FIGURE 9 – CURRENT LIMITING CHARACTERISTICS AS A FUNCTION OF JUNCTION TEMPERATURE



TYPICAL CHARACTERISTICS (continued)

FIGURE 10 – LINE REGULATION AS A FUNCTION OF INPUT-OUTPUT VOLTAGE DIFFERENTIAL

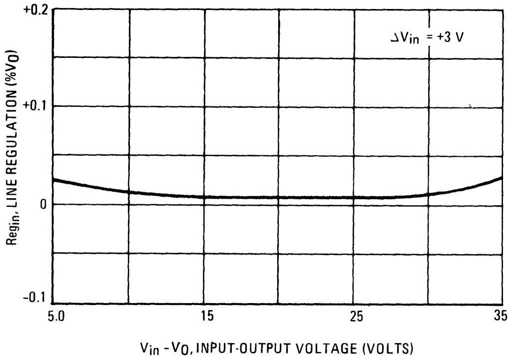


FIGURE 11 – LOAD REGULATION AS A FUNCTION OF INPUT-OUTPUT VOLTAGE DIFFERENTIAL

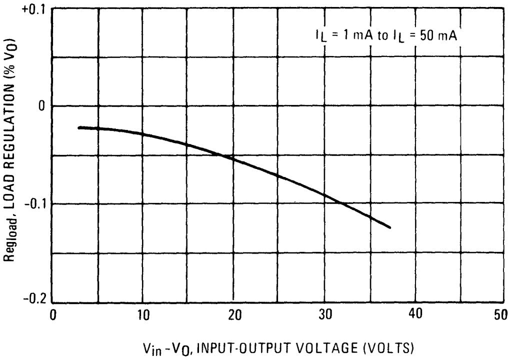


FIGURE 12 – STANDBY CURRENT DRAIN AS A FUNCTION OF INPUT VOLTAGE

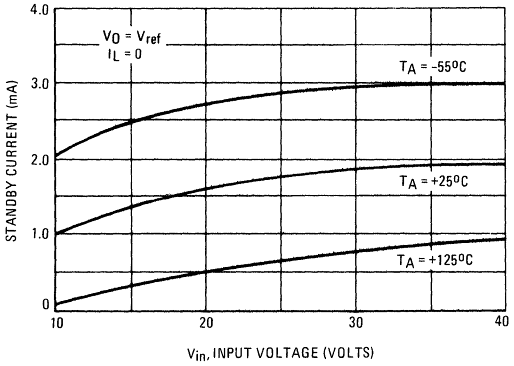


FIGURE 13 – LINE TRANSIENT RESPONSE

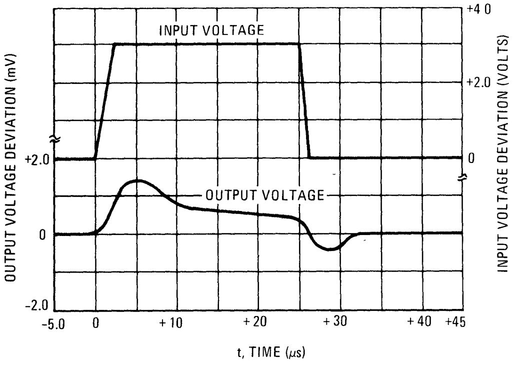


FIGURE 14 – LOAD TRANSIENT RESPONSE

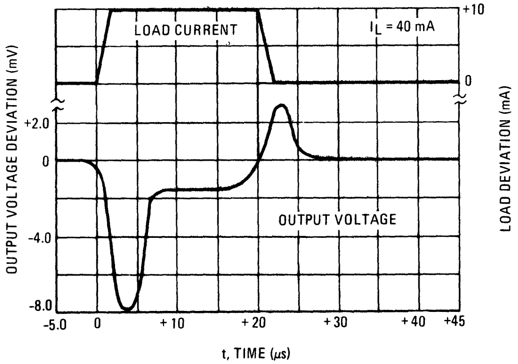
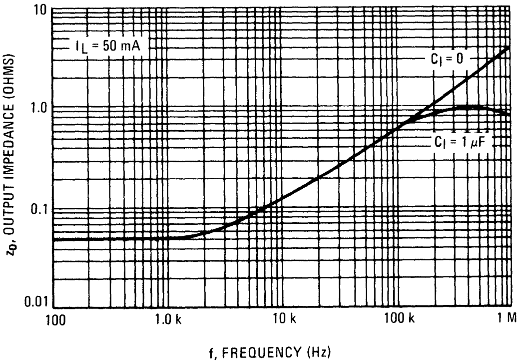


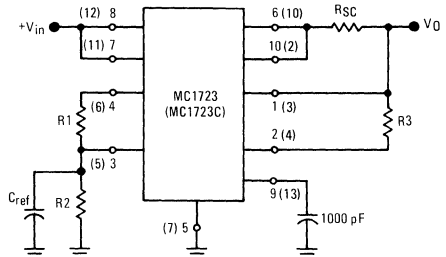
FIGURE 15 – OUTPUT IMPEDANCE AS FUNCTION OF FREQUENCY



TYPICAL APPLICATIONS

Pin numbers adjacent to terminals are for the metal package;
pin numbers in parenthesis are for the dual in-line packages.

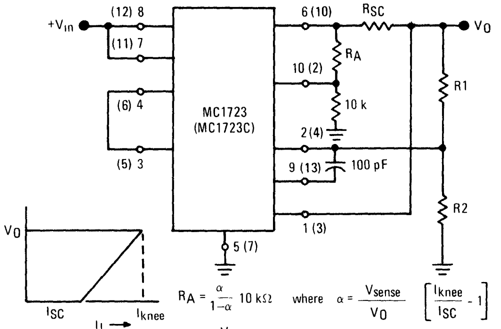
FIGURE 16 – TYPICAL CONNECTION FOR $2 < V_O < 7$



$V_O \cong 7\left[\frac{R2}{R1+R2}\right]$ $\quad I_{SC} = \frac{V_{sense}}{R_{SC}} \cong \frac{0.66}{R_{SC}}$ at $T_J = +25^{\circ}C$

For best results 10 k < R1 + R2 < 100 k.
For minimum drift R3 = R1||R2.

FIGURE 17 – MC1723,C FOLDBACK CONNECTION



$R_A = \frac{\alpha}{1-\alpha}\ 10\ k\Omega$ $\quad$ where $\alpha = \frac{V_{sense}}{V_O}\left[\frac{I_{knee}}{I_{SC}} - 1\right]$

$R_{SC} = \frac{V_{sense}}{(1-\alpha)\ I_{SC}}$

FIGURE 18 – +5 V, 1-AMPERE SWITCHING REGULATOR

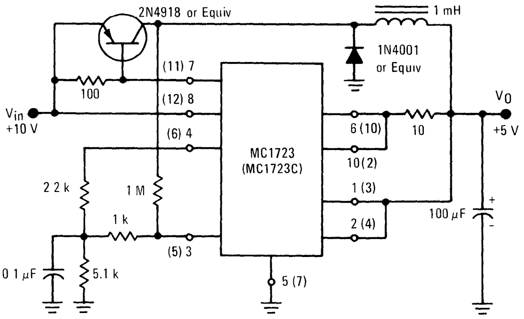


FIGURE 19 – +5 V, 1-AMPERE HIGH EFFICIENCY REGULATOR

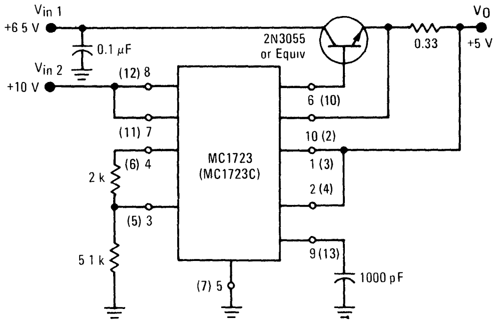


FIGURE 20 – +15 V, 1-AMPERE REGULATOR WITH REMOTE SENSE

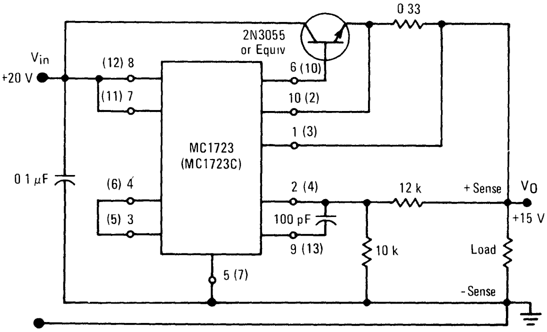
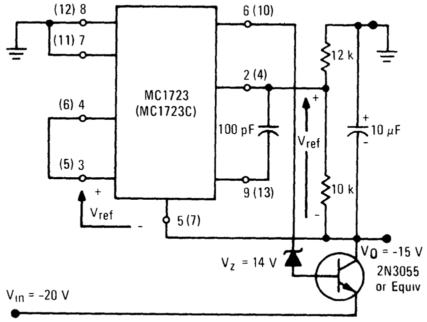


FIGURE 21 – -15 V NEGATIVE REGULATOR



MOTOROLA

MC3420
MC3520

SWITCHMODE REGULATOR CONTROL CIRCUIT

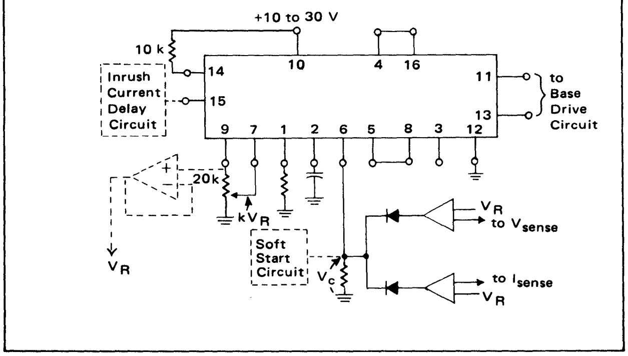
The MC3520/3420 is an inverter control unit which provides all the control circuitry for PWM push-pull, bridge and series type switchmode power supplies.

These devices are designed to supply the pulse width modulated drive to the base of two external power transistors. Other applications where these devices can be used are in transformerless voltage doublers, transformer coupled dc-to-dc converters and other power control functions.

The MC3520 is specified over the military operating range of -55°C to +125°C. The MC3420 is specified from 0°C to +70°C.

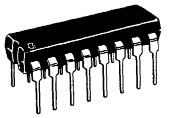
- Includes Symmetrical Oscillator
- On Chip Pulse Width Modulator, Voltage Reference, Dead Time Comparator, and Phase Splitter
- Output Frequency Adjustable (2 kHz to 100 kHz)
- Inhibit and Symmetry Correction Inputs Available
- Controlled Start-Up
- Frequency and Dead Time are Independently Adjustable (0% to 100%)
- Can be Slaved to Other MC3420s
- Open Collector Outputs
- Output Capability 50 mA (Max.)
- On Chip Protection Against Double Pulsing of Same Output During Load Transient Condition

FIGURE 1—TYPICAL APPLICATION

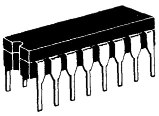


SWITCHMODE REGULATOR CONTROL CIRCUIT

SILICON MONOLITHIC INTEGRATED CIRCUITS

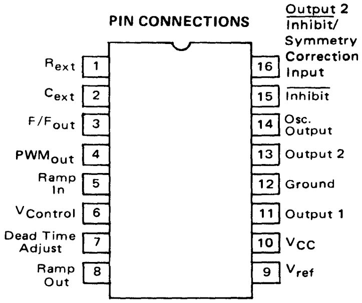


P SUFFIX
PLASTIC PACKAGE
CASE 648



L SUFFIX
CERAMIC PACKAGE
CASE 620

PIN CONNECTIONS



ORDERING INFORMATION

DEVICE	TEMPERATURE RANGE	PACKAGE
MC3420P	0 to +70°C	Plastic DIP
MC3420L	0 to +70°C	Ceramic DIP
MC3520L	-55 to +125°C	Ceramic DIP

MAXIMUM RATINGS

Rating	Symbol	MC3520	MC3420	Unit
Power Supply Voltage	V_{CC}	30		V
Output Voltage (pins 11 and 13)	V_{out}	40		V
Oscillator Output Voltage (pin 14)	V_{14}	30		V
Voltage at pin 4	V_4	2.0		V
Voltage at pins 3 and 8	V_3, V_8	5.0		V
Voltage at pin 5	V_5	7.0		V
Power Dissipation	P_D	See Thermal Information		
Operating Junction Temperature Plastic Package Ceramic Package	T_J	 – 150	 125 150	°C
Operating Ambient Temperature Range	T_A	-55 to +125	0 to +70	°C
Storage Temperature Range	T_{stg}	-65 to +150	-65 to +150	°C

ELECTRICAL CHARACTERISTICS (V_{CC} = 10 to 30 V, T_A = 25°C unless otherwise noted.)

Characteristic	Figure	Symbol	MC3520			MC3420			Unit
			Min	Typ	Max	Min	Typ	Max	
REFERENCE SECTION									
Reference Voltage (I_{ref} = 400 μA)	5	V_{ref}	7.6	7.8	8.0	7.4	7.8	8.2	V
Temperature Coefficient of Reference Voltage (V_{CC} = 15 V, I_{ref} = 400 μA)	5	TCV_{ref}	–	0.008	0.03	–	0.008	0.03	%/°C
Input Regulation of Reference Voltage (I_{ref} = 400 μA) (I_{ref} = 1.0 mA)	5	$Reg_{(in)}$	 – –	 3.0 5.0	 7.5 –	 – –	 4.0 5.0	 7.5 –	mV/V
DC SUPPLY SECTION									
Supply Voltage	5	V_{in}	10	–	30	10	–	30	V
Supply Current (R_{ext} = 10 kΩ, excluding load and current and reference current)	5	I_D	–	–	16	–	–	22	mA
OSCILLATOR SECTION									
Line Frequency Stability (f = 20 kHz) (f = 20 kHz, V_{CC} = 15 V, T_{low} to T_{high})	5	 Δf Δf	 – –	 – 0.03	 3.0 –	 – –	 – 0.04	 5.0 –	 % %/°C
Maximum Output Frequency (V_{CC} = 15 V)	6	f_{max}	100	200	–	100	200	–	kHz
Minimum Output Frequency (V_{CC} = 15 V)	6	f_{min}	–	2.0	5.0	–	2.0	5.0	kHz
Oscillator Output Saturation Voltage ($I_{14\ sink}$ = 5.0 mA)	11	$V_{osc(sat)}$	–	0.2	0.5	–	0.2	0.5	V
OUTPUT SECTION									
Output Saturation Voltage (I_L = 40 mA, T_{high} to T_{low}) (I_L = 25 mA, T_{high} to T_{low})	7	$V_{CE(sat)}$	 – –	 0.33 0.22	 0.5 –	 – –	 0.33 0.22	 0.5 –	V
Output Leakage Current (V_{CE} = 40 V, pins 11 and 13)	8	I_{CE}	–	–	50	–	–	50	μA
COMPARATOR SECTION									
Pulse Width Adjustment Range	9	ΔPW	0	–	100	0	–	100	%
Dead Time Adjustment Range	9	ΔDT	0	–	100	0	–	100	%
Temperature Coefficient of Dead Time	–	TCDT	–	0.1	–	–	0.1	–	%/°C
Comparator Bias Currents	12,13 14	I_{IB} I_{IB}	– –	5.0 10	15 30	– –	5.0 10	15 30	μA μA

ELECTRICAL CHARACTERISTICS (continued)

Characteristic	Figure	Symbol	MC3520			MC3420			Unit
			Min	Typ	Max	Min	Typ	Max	
AUXILIARY INPUTS/OUTPUTS									
Ramp Voltage	5								V
Peak High		$V_{ramp(Hi)}$	5.5	6.0	6.5	5.5	6.0	6.5	
Peak Low		$V_{ramp(Low)}$	2.0	2.4	2.8	2.0	2.4	2.8	
Ramp Voltage Change ($V_{ramp\ Hi} - V_{ramp\ Low}$)	5	ΔV_{ramp}	3.0	3.5	4.0	3.0	3.5	4.0	V
Ramp Out Sink Current	5	I_{sink}	–	400	–	–	400	–	μA
Ramp Out Source Current	5	I_{source}	–	3.0	–	–	3.0	–	mA
$\overline{\text{Inhibit}}$ Input Current – High (V_{IH} = 2.0 V)	10	I_{IH}	–	–	40	–	–	40	μA
$\overline{\text{Inhibit}}$ Input Current – Low (V_{IL} = 0.8 V)	10	I_{IL}	–	-25	-180	–	-25	-180	μA
Symmetry Correction Input/Output 2 Inhibit Current – High (V_{SY} = 2.0 V, pin 16)	10	$I_{SY/H}$	–	–	40	–	–	40	μA
Symmetry Correction Input/Output 2 Inhibit Current – Low (V_{SY} = 0.8 V, pin 16)	10	$I_{SY/L}$	–	-10	-180	–	-10	-180	μA
F/F_{out} Source Current	–	I_{source}	–	2.0	–	–	2.0	–	mA
OUTPUT AC CHARACTERISTICS ($T_A = T_{high}$, V_{CC} = +15 V, f = 20 kHz)									
Rise Time	15	t_r	–	40	–	–	40	–	ns
Fall Time	15	t_f	–	150	–	–	150	–	ns
Overlap Time	15	t_{ov}	–	275	–	–	275	–	ns
Assymmetry (Duty Cycle = 50%)	15	$\frac{t_{on1} - t_{on2}}{t_{on1}}$	–	±1.0	–	–	±1.0	–	%

NOTE:

T_{high} = +125°C for MC3520
+70°C for MC3420

T_{low} = -55°C for MC3520
0°C for MC3420

FIGURE 2–EQUIVALENT CIRCUIT

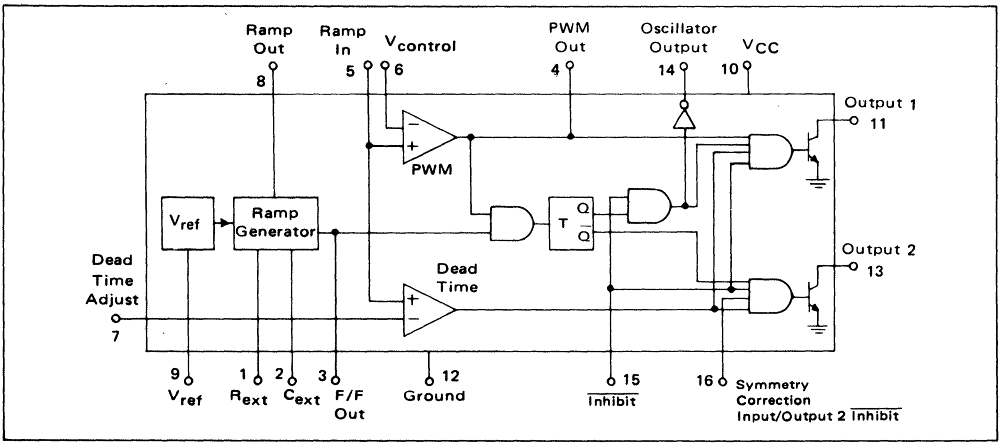
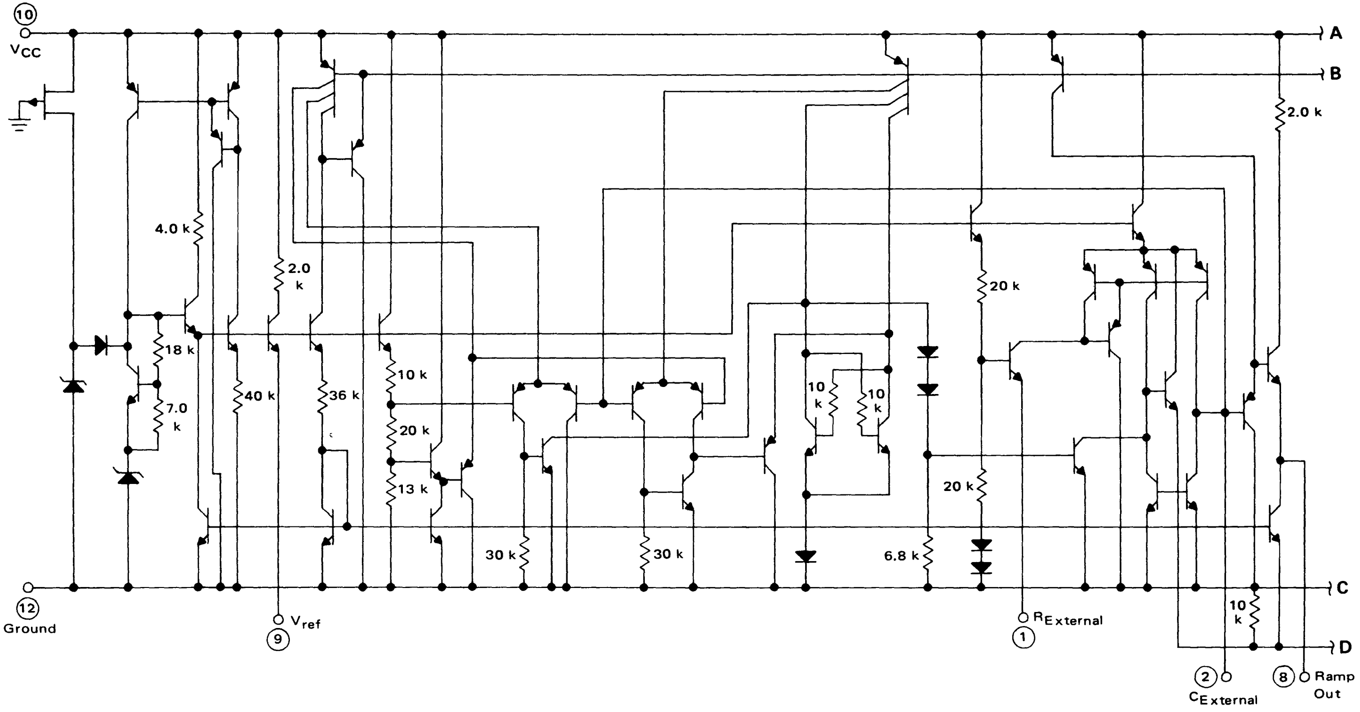
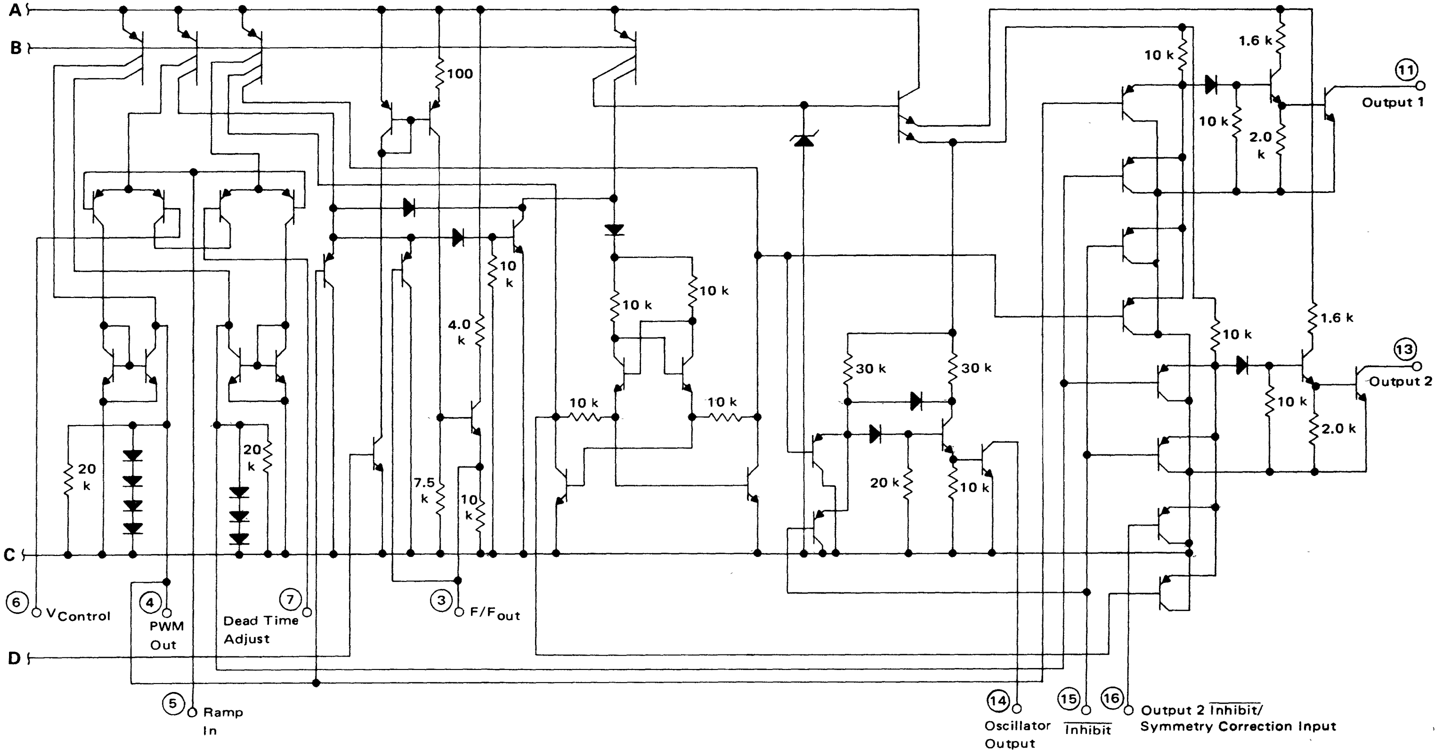


FIGURE 3 – CIRCUIT SCHEMATIC

(continued next page)



(continued) FIGURE 3 – CIRCUIT SCHEMATIC



GENERAL INFORMATION

The internal block diagram of the MC3420 is shown in Figure 2, and consists of the following sections:

Voltage Reference

A stable reference voltage is generated by the MC3420 primarily for internal use. However, it is also available externally at Pin 9 (V_{ref}) for use in setting the dead time (Pin 7) and for use as a reference for the external control loop error amplifiers.

Ramp Generator

The ramp generator section produces a symmetrical triangular waveform ramping between 2.4 V and 6.0 V, with frequency determined by an external resistor (R_{ext}) and capacitor (C_{ext}) tied from Pins 1 and 2, respectively, to ground.

PWM Comparator

The output of the ramp generator at pin 8 is normally connected to Pin 5, RAMP IN. The PWM (pulse width modulation) comparator compares the voltage at Pin 6 ($V_{control}$) to the ramp generator output. The level of $V_{control}$ determines the outputs' pulse width or duty cycle. The duty cycle of each output can vary, exclusive of dead time, from 50% (when $V_{control}$ is at approximately 2.4 V) to 0% ($V_{control}$ approximately 6.0 V).

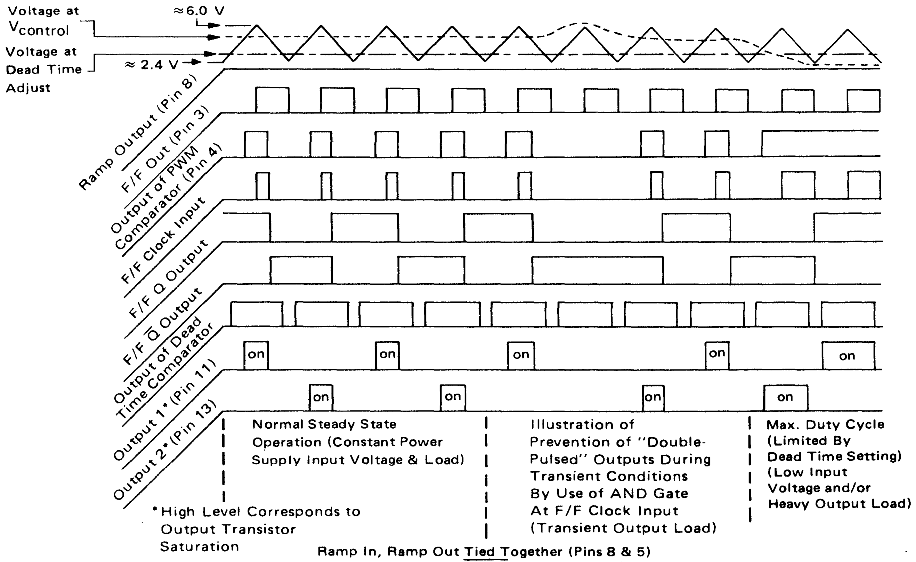
Dead Time Comparator

An additional comparator has been included in MC3420 to allow independent adjustment of system dead time or maximum duty cycle. By dividing down V_{ref} at Pin 9 with a resistive divider or potentiometer, and applying this voltage to Pin 7, a stable dead time is obtained for prevention of inverter switching transistor cross conduction at high duty cycles due to storage time delays.

Phase Splitter

A phase splitter is included to obtain two 180° out of phase outputs for use in multiple transistor inverter systems. It consists of a toggle flip-flop whose clock signal is derived by "ANDing" the output of the PWM comparator and a signal from the ramp generator section. This "AND" gate ensures that the outputs truly alternate under control loop transient conditions. Better understanding of this feature and MC3420 operation may be gained by studying the circuit waveforms, shown in Figure 4.

FIGURE 4 – INTERNAL WAVEFORMS



*High Level Corresponds to Output Transistor Saturation

Ramp In, Ramp Out Tied Together (Pins 8 & 5)
PWM Out, Output 2 Inhibit Tied Together (Pins 4 & 16)

FIGURE 5 – STANDARD AC, DC TEST CIRCUIT

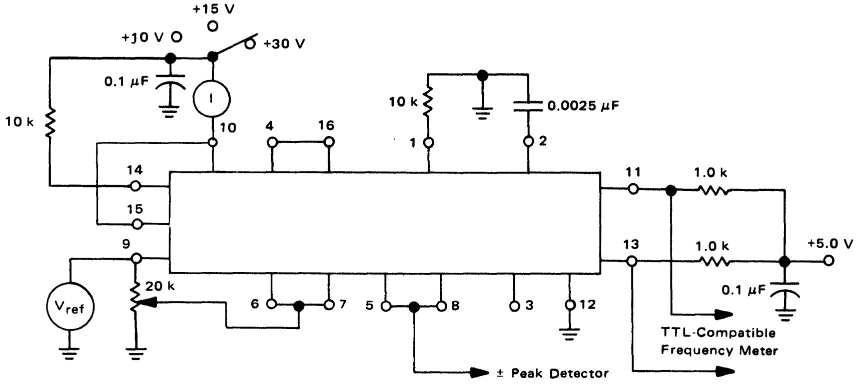


FIGURE 6 – FREQUENCY LIMIT TEST CIRCUIT

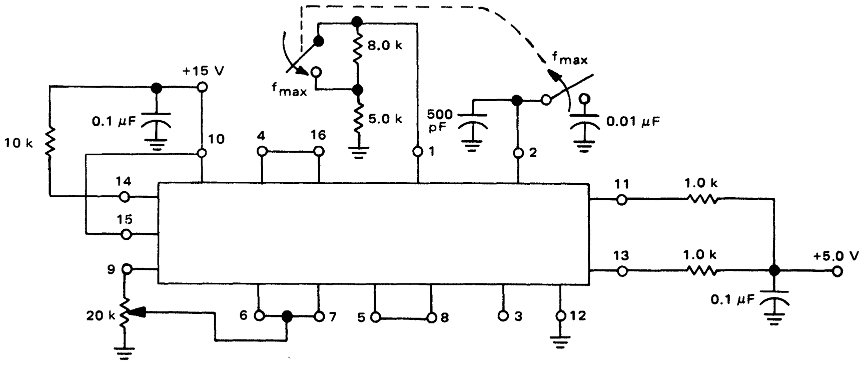
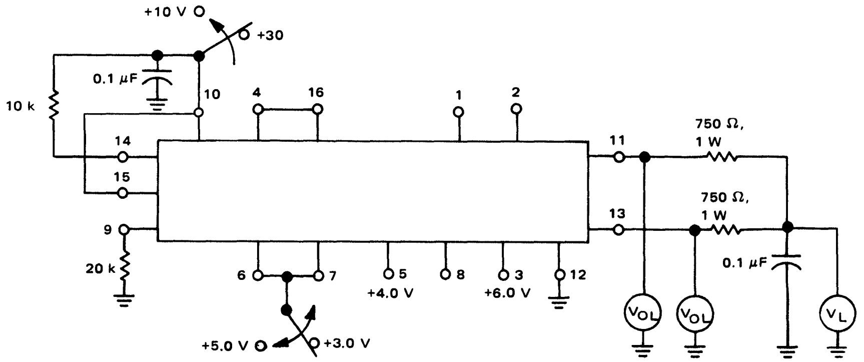
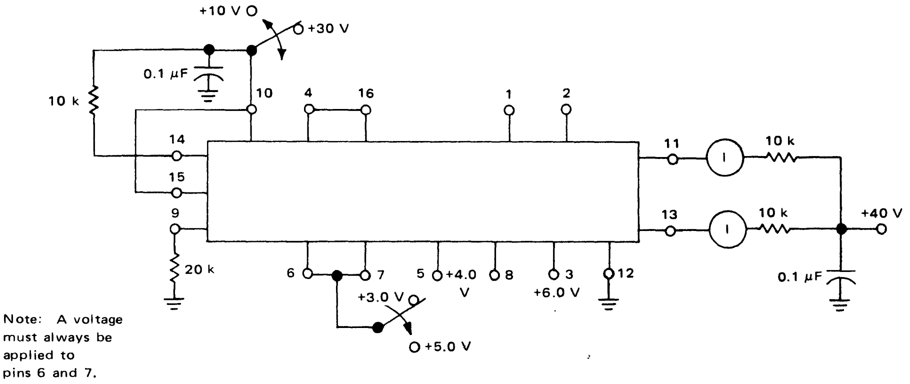


FIGURE 7 – OUTPUT SATURATION TEST CIRCUIT



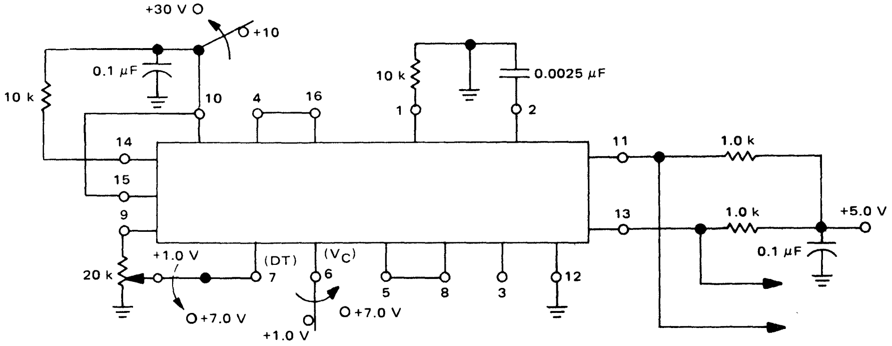
Note: Use voltage change on pins 6, 7 to change output states.
A voltage must always be present on pins 6 and 7.

FIGURE 8 – OUTPUT LEAKAGE TEST CIRCUIT



Note: A voltage must always be applied to pins 6 and 7.

FIGURE 9 – OUTPUT DUTY CYCLE TEST CIRCUIT



TYPICAL DUTY CYCLE versus DEAD TIME VOLTAGE	
PIN 7. DEAD TIME VOLTAGE (V) ($V_{control}$ = 2.0 V)	% DUTY CYCLE (FOR EACH OUTPUT)
2.0	50
2.5	46
3.0	40
3.5	33
4.0	26
4.5	18
5.0	11
5.5	4.0
6.0	0

TYPICAL DUTY CYCLE versus PWM VOLTAGE ($V_{control}$)	
PIN 6. $V_{control}$ (V) (DEAD TIME VOLTAGE = 1.0 V)	% DUTY CYCLE (FOR EACH OUTPUT)
2.0	50
2.5	46
3.0	40
3.5	33
4.0	26
4.5	18
5.0	11
5.5	4.0
6.0	0

	V_6	V_7	
	Volts		
100% Adjust			
Dead Time	1.0	1.0	(Pin 11 + Pin 13 = Logic "1")
Pulse Width	1.0	1.0	
0% Adjust			
Dead Time	7.0	1.0	(Pin 11)(Pin 13) = Logic "1"
Pulse Width	1.0	7.0	

NOTE: Logic "1" is TTL-Compatible V_{OH}.

FIGURE 10 – INHIBIT/SYMMETRY TEST CIRCUIT

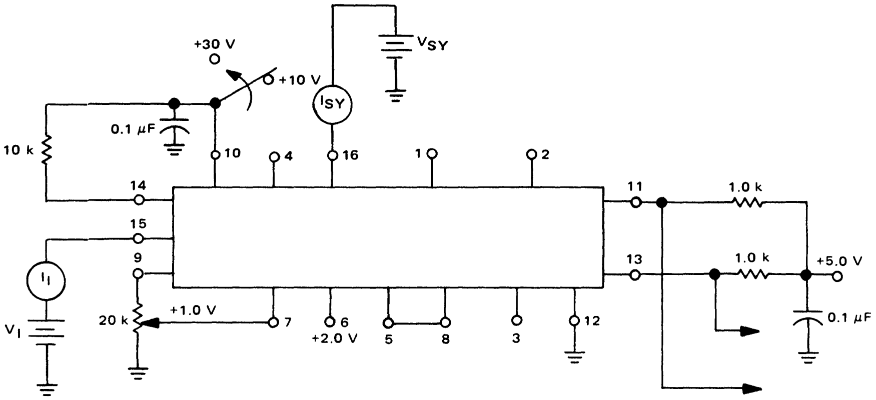


FIGURE 11 – OSCILLATOR OUTPUT (pin 14) TEST CIRCUIT

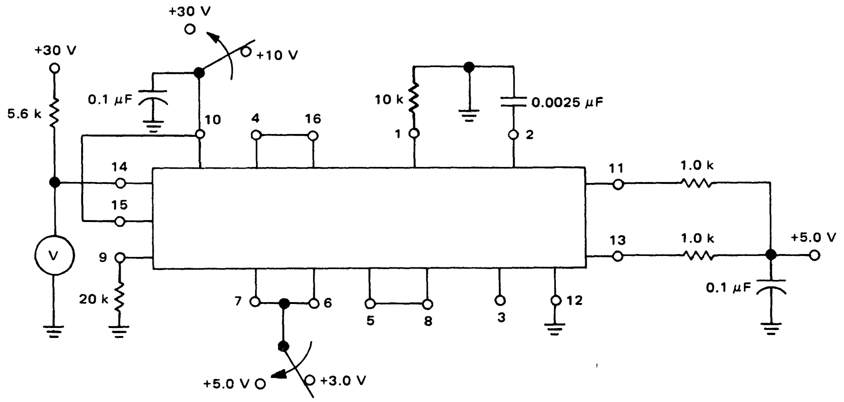


FIGURE 12 – $V_{Control}$ BIAS CURRENT TEST CIRCUIT

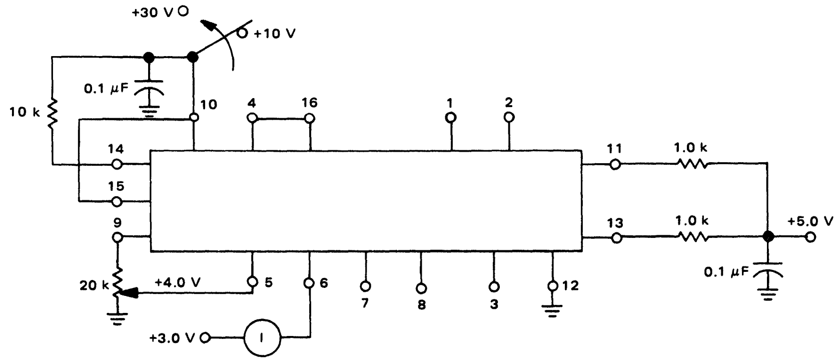


FIGURE 13 – DEAD TIME BIAS CURRENT TEST CIRCUIT

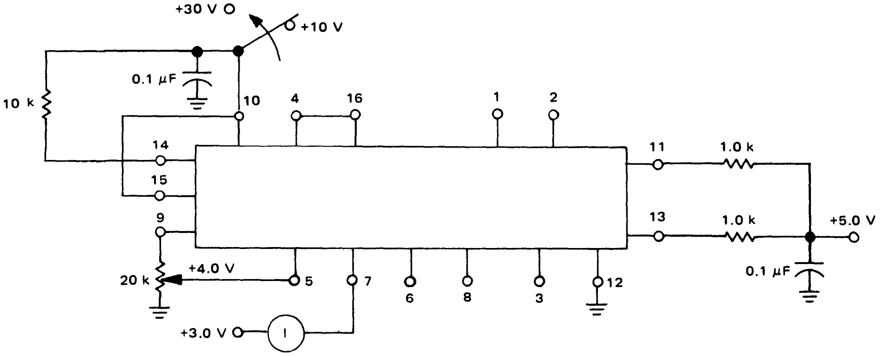


FIGURE 14 – RAMP IN BIAS CURRENT TEST CIRCUIT

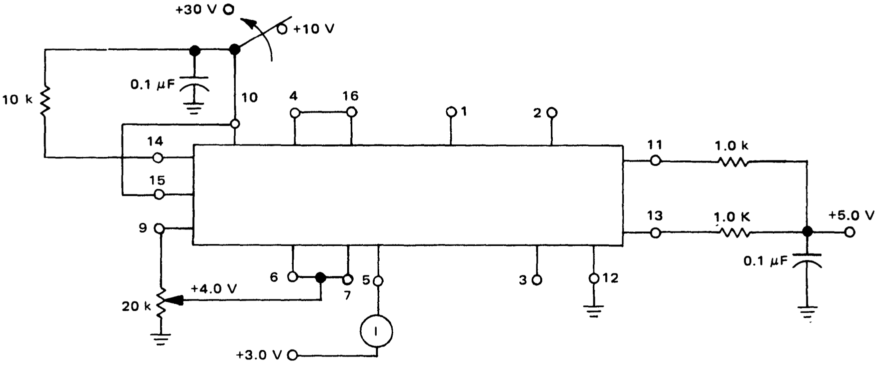
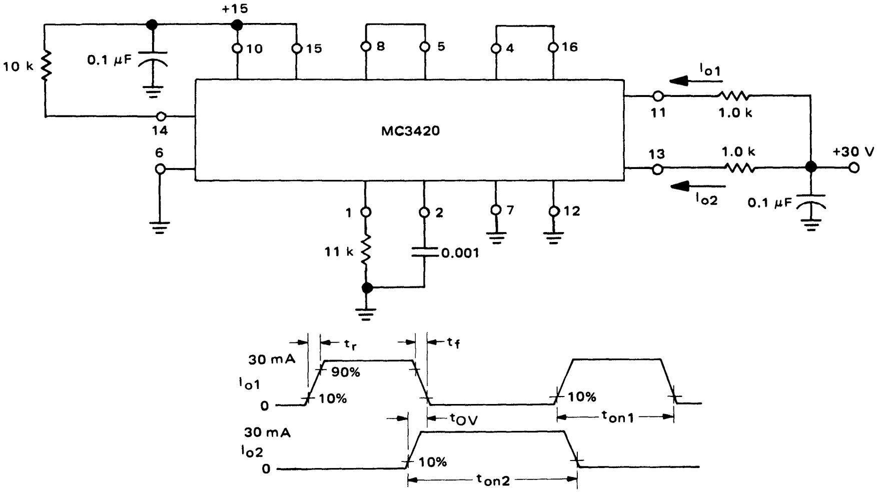


FIGURE 15 – AC TEST CIRCUIT AND WAVEFORMS



TYPICAL CHARACTERISTICS

FIGURE 16 – OUTPUT SATURATION VOLTAGE versus LOAD CURRENT

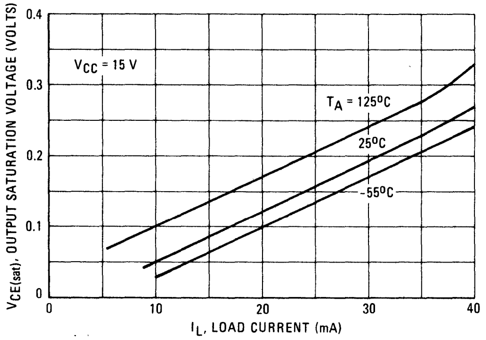


FIGURE 17 – REFERENCE VOLTAGE versus REFERENCE CURRENT

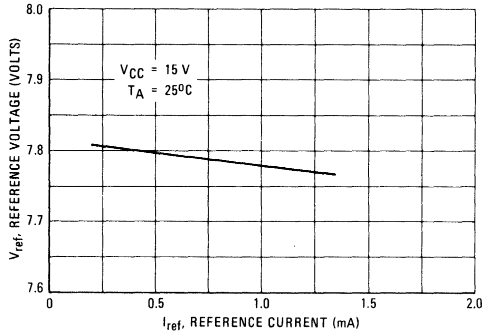


FIGURE 18 – DRAIN CURRENT versus EXTERNAL RESISTANCE

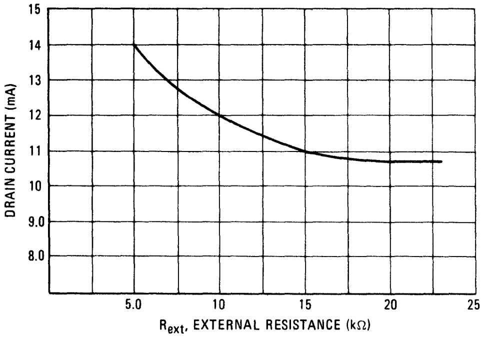


FIGURE 19 – PEAK FLIP-FLOP$_{out}$ VOLTAGE versus EXTERNAL RESISTANCE

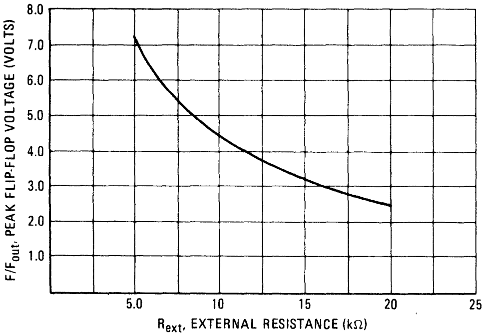


FIGURE 20 – DRAIN CURRENT versus TEMPERATURE

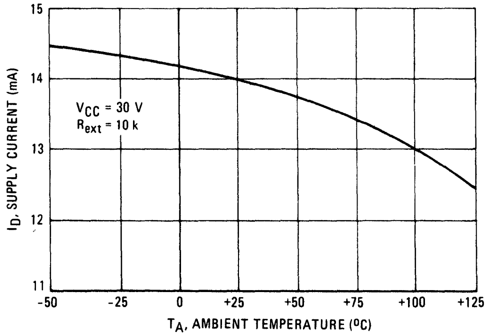
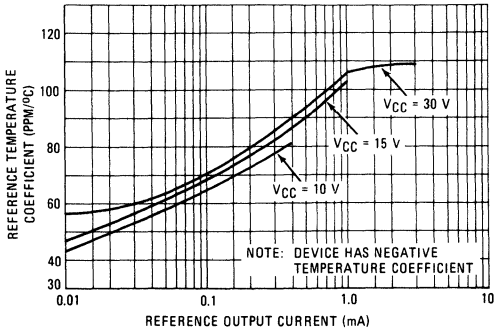


FIGURE 21 – REFERENCE VOLTAGE TEMPERATURE COEFFICIENT versus OUTPUT CURRENT

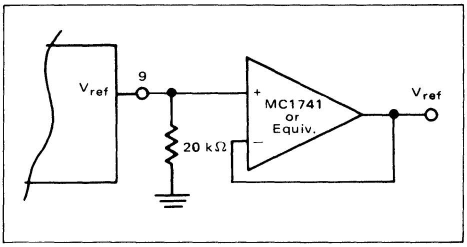


OPERATION AND APPLICATIONS INFORMATION

The Voltage Reference

The temperature coefficient of V_{ref} has been optimized for a 400 μA ($\cong$20 kΩ) load. If increased current capability is required, an op amp buffer may be used, as shown in Figure 22.

FIGURE 22



Output Frequency

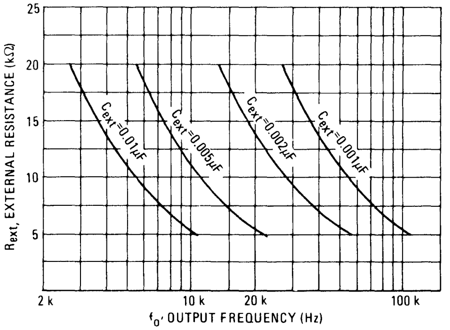
The values of R_{ext} and C_{ext} for a given output frequency, f_o, can be found from:

$$f_o \cong \frac{0.55}{R_{ext}\,C_{ext}};\ 5.0\ k\Omega \leqslant R_{ext} \leqslant 20\ k\Omega \quad \text{(Eq. 1)}$$

or from the graph shown in Figure 23.

Note that f_o refers to the frequency of Output 1 (Pin 11) or Output 2 (Pin 13). The frequency of the ramp generator output waveform at Pin 8 will be twice f_o.

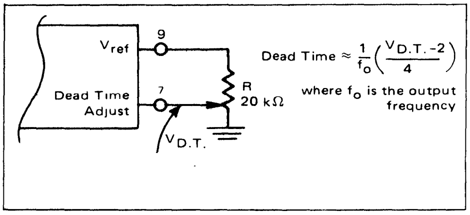
FIGURE 23



Dead Time

Figure 24 illustrates how to set or adjust the MC3420 outputs' dead time or maximum duty cycle. For minimum dead time drift with temperature or supply voltage, $V_{D.T.}$ should be derived from V_{ref} as shown.

FIGURE 24



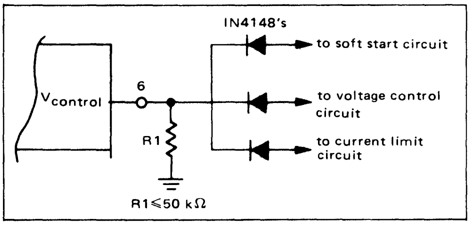
Connections to the $V_{control}$ Pin

In many systems, it is necessary to make multiple connections to the $V_{control}$ Pin in order to implement features in addition to voltage regulation such as current limiting, soft start, etc. These can be made by the use of a simple "diode-OR" connection, as shown in Figure 25. This allows whichever control element is seeking the lowest PWM duty cycle to dominate. Note that a resistor, R1, whose value is $\leqslant$ 50 kΩ is placed from the $V_{control}$ Pin to ground. This is necessary to provide a dc path for the PWM comparator input bias current under all conditions.

The system duty cycle is given by:

$$\text{D.C. (\%)} \cong \frac{V_{Control} - 2}{4} \times 100 \qquad \text{(Eq. 2)}$$

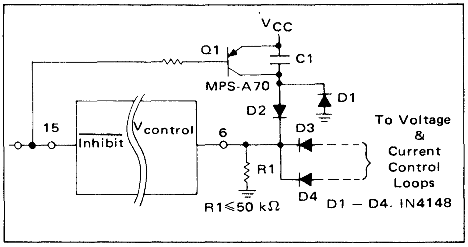
FIGURE 25



Soft Start

In most PWM switching supplies, a soft start feature is desired to prevent output voltage overshoots and magnetizing current imbalances in the power transformer primary. This feature forces the duty cycle of the switching elements to gradually increase from zero to their normal operating point during initial system power-up or after an inhibit. This feature can be easily implemented with the MC3420. One method is shown in Figure 26.

FIGURE 26



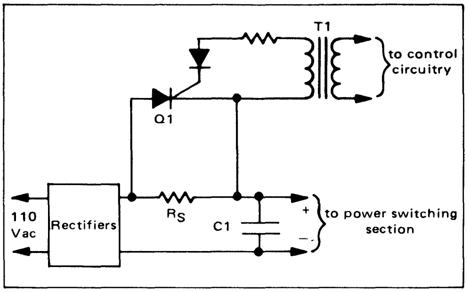
After an inhibit command or during power-up, the voltage on R1 and Pin 6 exponentially decays from V_{CC} toward ground with a time constant of R1C1, allowing a gradual increase in duty cycle. Diodes D2 – D4 provide a diode-or function at the $V_{control}$ Pin, while Q1 serves to reset the timing capacitor, C1, when an inhibit command is received thereby reinitializing the soft-start feature. D1 allows C1 to reset when power (V_{CC}) is turned off.

Inrush Current Limiting

Since many PWM switching supplies are operated directly off the rectified 110 Vac line with capacitive input filters, some means of preventing rectifier failure due to inrush surge currents is usually necessary. One method which can be used is shown in Figure 27.

In this circuit, a series resistor, R_S, is used to provide inrush surge current limiting. After the filter capacitor, C1, is charged, Q1 receives a trigger signal from the control circuitry through T1 and shorts R_S out of the circuit, eliminating its otherwise larger power dissipation. The trigger signal for Q1 may be derived from either the oscillator output (Pin 14) or one of the MC3420's outputs. If the oscillator output is used, it will be necessary to provide a time delay on the inhibit pin to keep it low until the input filter capacitor, C1, has had time to charge, whereas the initial portion of the soft start timing cycle can be used for this delay if this signal is derived from one of the output pins. However, using the Oscillator Output Pin does offer the advantage that its waveform has a constant 50% duty cycle, independent of the outputs' duty cycle which can simplify the design of a drive circuit for T1.

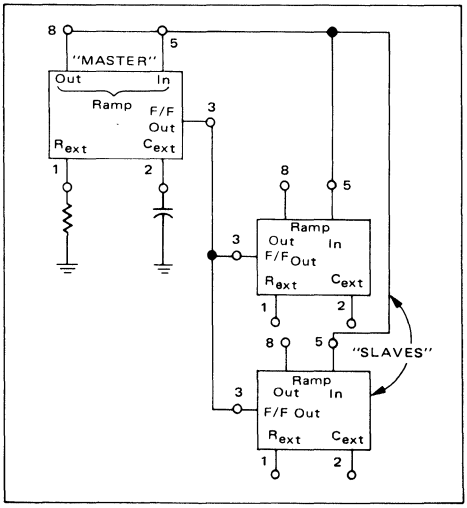
FIGURE 27



Slaving

In some applications, as when one PWM inverter/converter is used to feed another, it may be desired that their frequencies be synchronized. This can be done with multiple MC3420s as shown in Figure 28. By omitting their R_{ext} and C_{ext}, up to two MC3420s may be slaved to a master MC3420.

FIGURE 28 – SLAVING THE MC3420



MOTOROLA

MC3423
MC3523

Specifications and Applications Information

OVERVOLTAGE "CROWBAR" SENSING CIRCUIT

These overvoltage protection circuits (OVP) protect sensitive electronic circuitry from overvoltage transients or regulator failures when used in conjunction with an external "crowbar" SCR. They sense the overvoltage condition and quickly "crowbar" or short circuit the supply, forcing the supply into current limiting or opening the fuse or circuit breaker.

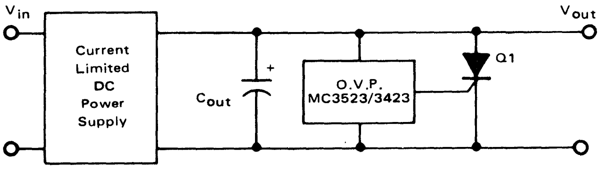
The protection voltage threshold is adjustable and the MC3423/3523 can be programmed for minimum duration of overvoltage condition before tripping, thus supplying noise immunity.

The MC3423/3523 is essentially a "two terminal" system, therefore it can be used with either positive or negative supplies.

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Differential Power Supply Voltage	V_{CC}-V_{EE}	40	Vdc
Sense Voltage (1)	$V_{Sense\ 1}$	6.5	Vdc
Sense Voltage (2)	$V_{Sense\ 2}$	6.5	Vdc
Remote Activation Input Voltage	V_{act}	7.0	Vdc
Output Current	I_O	300	mA
Operating Ambient Temperature Range MC3423 MC3523	T_A	 0 to +70 –55 to +125	°C
Operating Junction Temperature Plastic Package Ceramic Package	T_J	 125 150	°C
Storage Temperature Range	T_{stg}	–65 to +150	°C

TYPICAL APPLICATION

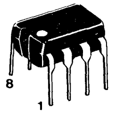


NOTE: A 2N6504 or equivalent is suggested for Q1.

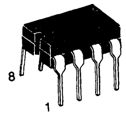
OVERVOLTAGE SENSING CIRCUIT

SILICON MONOLITHIC INTEGRATED CIRCUIT

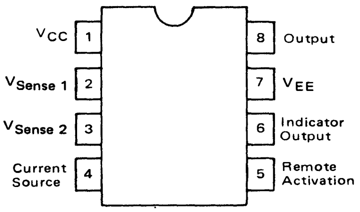
P1 SUFFIX
PLASTIC PACKAGE
CASE 626
(MC3423 only)



U SUFFIX
CERAMIC PACKAGE
CASE 693



PIN CONNECTIONS



(top view)

ORDERING INFORMATION

DEVICE	TEMPERATURE RANGE	PACKAGE
MC3423P1	0 to +70°C	Plastic DIP
MC3423U	0 to +70°C	Ceramic DIP
MC3523U	–55 to +125°C	Ceramic DIP

ELECTRICAL CHARACTERISTICS ($5\ V \leq V_{CC} - V_{EE} \leq 36\ V$, $T_{low} < T_A < T_{high}$ unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Supply Voltage Range	V_{CC}-V_{EE}	4.5	–	40	Vdc
Output Voltage (I_O = 100 mA)	V_O	V_{CC}-2.2	V_{CC}-1.8	–	Vdc
Indicator Output Voltage ($I_{O(Ind)}$ = 1.6 mA)	$V_{OL}(Ind)$	–	0.1	0.4	Vdc
Sense Voltage (T_A = 25°C)	$V_{Sense\ 1}$, $V_{Sense\ 2}$	2.45	2.6	2.75	Vdc
Temperature Coefficient of $V_{Sense\ 1}$ (Figure 2)	TCV_{S1}	–	0.06	–	%/°C
Remote Activation Input Current					μA
(V_{IH} = 2.0 V, V_{CC}-V_{EE} = 5.0 V)	I_{IH}	–	5.0	40	
(V_{IL} = 0.8 V, V_{CC}-V_{EE} = 5.0 V)	I_{IL}	–	-120	-180	
Source Current	I_{source}	0.1	0.2	0.3	mA
Output Current Risetime (T_A = 25°C)	t_r	–	400	–	mA/μs
Propagation Delay (T_A = 25°C)	t_{pd}	–	0.5	–	μs
Supply Current	I_D				mA
MC3423		–	6.0	10	
MC3523		–	5.0	7.0	

T_{low} = -55°C for MC3523
= 0°C for MC3423

T_{high} = +125°C for MC3523
= +70°C for MC3423

FIGURE 1 – BLOCK DIAGRAM

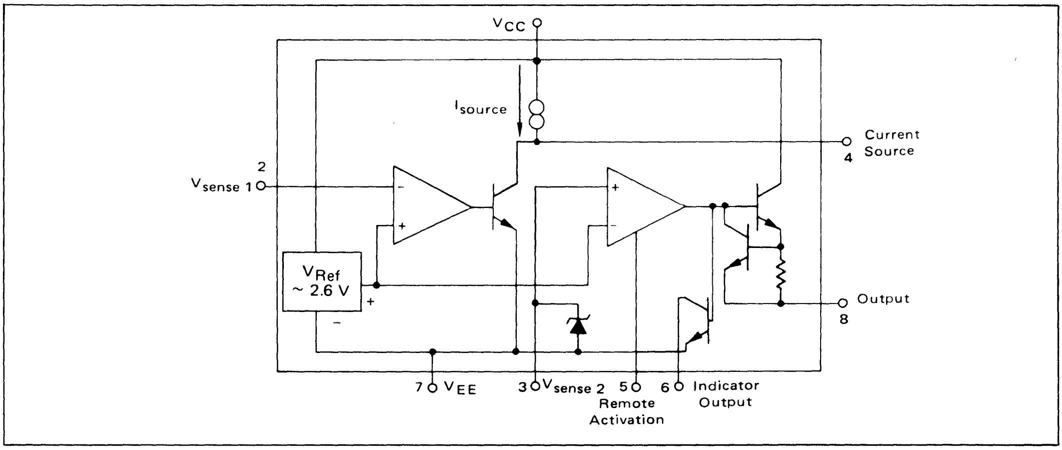
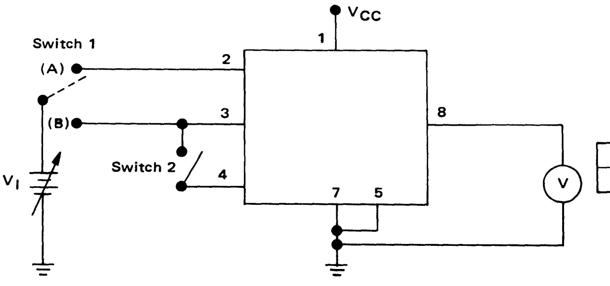


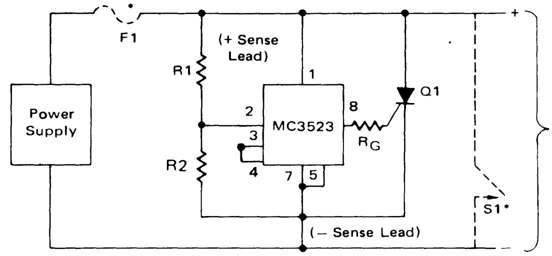
FIGURE 2 – SENSE VOLTAGE TEST CIRCUIT



	Switch 1	Switch 2
$V_{Sense\ 1}$	Position A	Closed
$V_{Sense\ 2}$	Position B	Open

Ramp V_I until output goes high; this is the V_{Sense} threshold.

FIGURE 3 – BASIC CIRCUIT CONFIGURATION



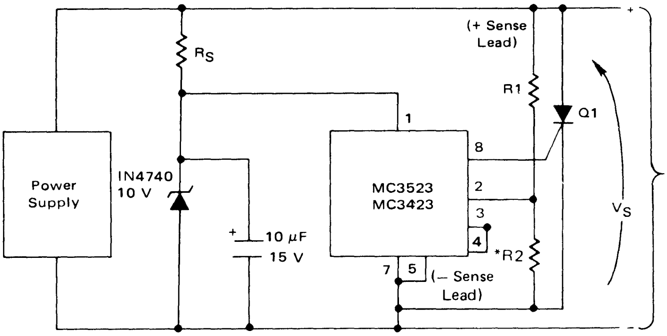
$V_{trip} = V_{ref}\ (1+\frac{R1}{R2}) \approx 2.6\ V\ (1+\frac{R1}{R2})$

R2 ⩽ 10 kΩ for minimum drift

For minimum value of R_G, see Figure 9

Q1 is 2N6504 or equivalent
*See text for explanation

FIGURE 4 – CIRCUIT CONFIGURATION FOR SUPPLY VOLTAGE ABOVE 36 V



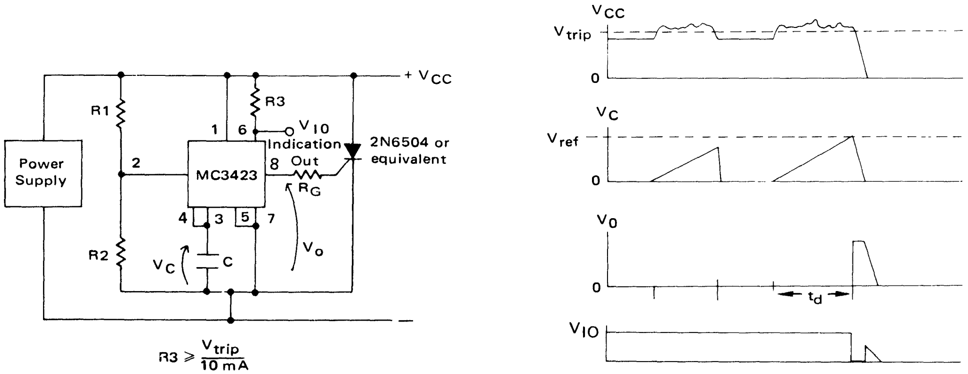
$R_S = (\frac{V_S - 10}{25})\ k\Omega$

$V_{trip} = V_{ref}\ (1+\frac{R1}{R2}) \approx 2.6\ V\ (1+\frac{R1}{R2})$

*R2 ⩽ 10 kΩ

Q1 $V_S \leqslant 50$ V; 2N6504 or equivalent
$V_S \leqslant 100$ V, 2N6505 or equivalent
$V_S \leqslant 200$ V, 2N6506 or equivalent
$V_S \leqslant 400$ V, 2N6507 or equivalent
$V_S \leqslant 600$ V, 2N6508 or equivalent
$V_S \leqslant 800$ V, 2N6509 or equivalent

FIGURE 5 – BASIC CONFIGURATION FOR PROGRAMMABLE DURATION OF OVERVOLTAGE CONDITION BEFORE TRIP



$R3 \geqslant \frac{V_{trip}}{10\ mA}$

$t_d = \frac{V_{ref}}{I_{source}} \times C \approx [12 \times 10^3]\ C$ (See Figure 10)

APPLICATIONS INFORMATION

BASIC CIRCUIT CONFIGURATION

The basic circuit configuration of the MC3423/3523 OVP is shown in Figure 3 for supply voltages from 4.5 V to 36 V, and in Figure 4 for trip voltages above 36 V. The threshold or trip voltage at which the MC3423/3523 will trigger and supply gate drive to the crowbar SCR, Q1, is determined by the selection of R1 and R2. Their values can be determined by the equation given in Figures 3 and 4, or by the graph shown in Figure 8. The minimum value of the gate current limiting resistor, R_G, is given in Figure 9. Using this value of R_G, the SCR, Q1, will receive the greatest gate current possible without damaging the MC3423/3523. If lower output currents are required, R_G can be increased in value. The switch, S1, shown in Figure 3 may be used to reset the SCR crowbar. Otherwise, the power supply, across which the SCR is connected, must be shut down to reset the crowbar. If a non current-limited supply is used, a fuse or circuit breaker, F1, should be used to protect the SCR and/or the load.

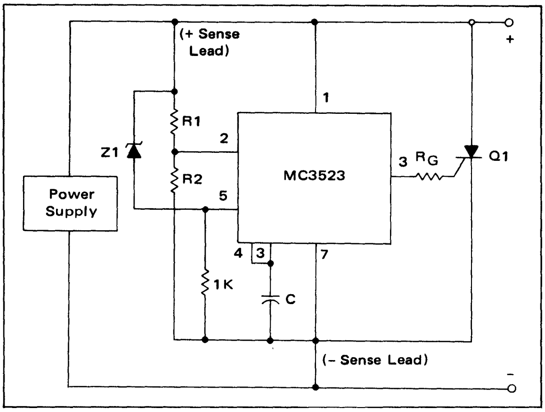
The circuit configurations shown in Figures 3 and 4 will have a typical propogation delay of 1.0 µs. If faster operation is desired, pin 3 may be connected to pin 2 with pin 4 left floating. This will result in decreasing the propogation delay to approximately 0.5 µs at the expense of a slightly increased TC for the trip voltage value.

CONFIGURATION FOR PROGRAMMABLE MINIMUM DURATION OF OVERVOLTAGE CONDITION BEFORE TRIPPING

In many instances, the MC3423/3523 OVP will be used in a noise environment. To prevent false tripping of the OVP circuit by noise which would not normally harm the load, MC3423/3523 has a programmable delay feature. To implement this feature, the circuit configuration of Figure 5 is used. In this configuration, a capacitor is connected from pin 3 to V_{EE}. The value of this capacitor determines the minimum duration of the overvoltage condition which is necessary to trip the OVP. The value of C can be found from Figure 10. The circuit operates in the following manner: When V_{CC} rises above the trip point set by R1 and R2, an internal current source (pin 4) begins charging the capacitor, C, connected to pin 3. If the overvoltage condition disappears before this occurs, the capacitor is discharged at a rate ≅ 10 times faster than the charging rate, resetting the timing feature until the next overvoltage condition occurs.

Occasionally, it is desired that immediate crowbarring of the supply occur when a high overvoltage condition occurs, while retaining the false tripping immunity of Figure 5. In this case, the circuit of Figure 6 can be used. The circuit will operate as previously described for small overvoltages, but will immediately trip if the power supply voltage exceeds V_{Z1} + 1.4 V.

FIGURE 6 – CONFIGURATION FOR PROGRAMMABLE DURATION OF OVERVOLTAGE CONDITION BEFORE TRIP/WITH IMMEDIATE TRIP AT HIGH OVERVOLTAGES



ADDITIONAL FEATURES

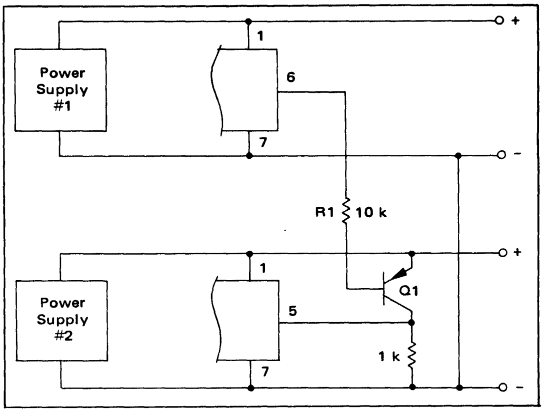
1. Activation Indication Output

An additional output for use as an indicator of OVP activation is provided by the MC3423/3523. This output is an open collector transistor which saturates when the OVP is activated. It will remain in a saturated state until the SCR crowbar pulls the supply voltage, V_{CC}, below 4.5 V as in Figure 5. This output can be used to clock an edge triggered flip-flop whose output inhibits or shuts down the power supply when the OVP trips. This reduces or eliminates the heatsinking requirements for the crowbar SCR.

2. Remote Activation Input

Another feature of the MC3423/3523 is its remote activation input, pin 5. If the volage on this CMOS/TTL compatible input is held below 0.8 V, the MC3423/3523 operates normally. However, if it is raised to a voltage above 2.0 V, the OVP output is activated independent of whether or not an overvoltage condition is present. It should be noted that pin 5 has an internal pull-up current source. This feature can be used to accomplish an orderly and sequenced shutdown of system power supplies during a system fault condition. In addition, the activation indication output of one MC3423/3523 can be used to activate another MC3423/3523 if a single transistor inverter is used to interface the former's indication output to the latter's remote activation input, as shown in Figure 7. In this circuit, the indication output (pin 6) of the MC3423 on power supply 1 is used to activate the MC3423 associated with power supply 2. Q1 is any small PNP with adequate voltage rating.

FIGURE 7 – CIRCUIT CONFIGURATION FOR ACTIVATING ONE MC3523 FROM ANOTHER



Note that both supplies have their negative output leads tied together (i.e., both are positive supplies). If their positive leads are common (two negative supplies) the emitter of Q1 would be moved to the positive lead of supply 1 and R1 would therefore have to be resized to deliver the appropriate drive to Q1.

CROWBAR SCR CONSIDERATIONS

Referring to Figure 11, it can be seen that the crowbar SCR, when activated, is subject to a large current surge from the output capacitance, C_{out}[1]. This surge current is illustrated in Figure 12, and can cause SCR failure or degradation by any one of three mechanisms: di/dt, absolute peak surge, or I^2t. The interrelationship of these failure methods and the breadth of the application make specification of the SCR by the semiconductor manufacturer difficult and expensive. Therefore, the designer must empirically determine the SCR and circuit elements which result in reliable and effective OVP operation. However, an understanding of the factors which influence the SCR's di/dt and surge capabilities simplifies this task.

1. di/dt

As the gate region of the SCR is driven on, its area of conduction takes a finite amount of time to grow, starting as a very small region and gradually spreading. Since the anode current flows through this turned-on gate region, very high current densities can occur in the gate region if high anode currents appear quickly (di/dt). This can result in immediate destruction of the SCR or gradual degradation of its forward blocking voltage capabilities – depending on the severity of the occasion.

[1] C_{out} consists of the power supply output caps, the load's decoupling caps, and in the case of Figure 11A, the supply's input filter caps.

FIGURE 8 – R1 versus TRIP VOLTAGE

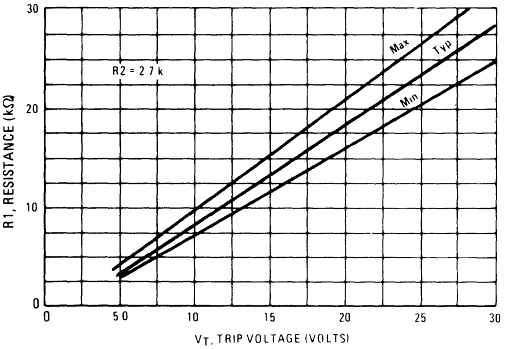


FIGURE 9 – MINIMUM R_G versus SUPPLY VOLTAGE

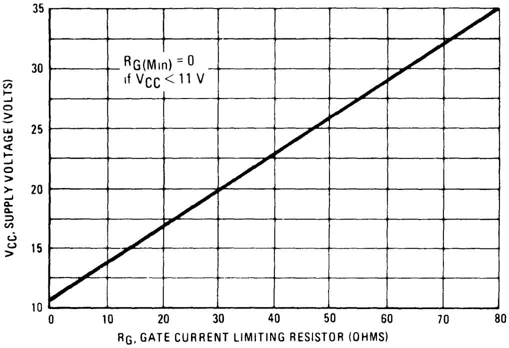


FIGURE 10 – CAPACITANCE versus MINIMUM OVERVOLTAGE DURATION

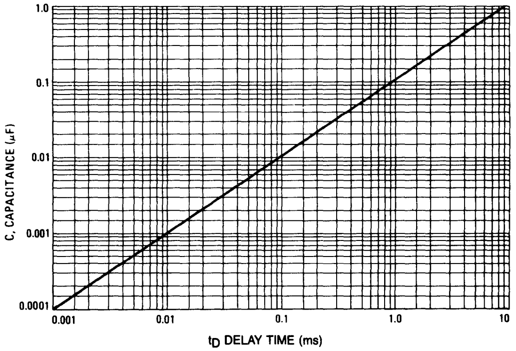


FIGURE 11 – TYPICAL CROWBAR OVP CIRCUIT CONFIGURATIONS

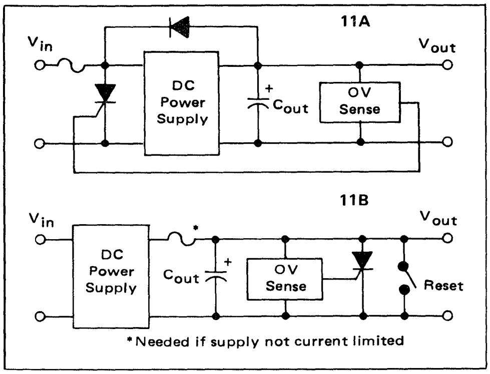


FIGURE 12 – CROWBAR SCR SURGE CURRENT WAVEFORM

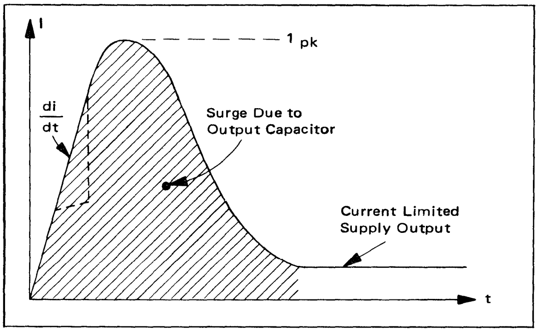
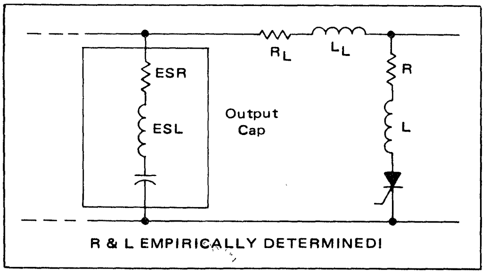


FIGURE 13 – CIRCUIT ELEMENTS AFFECTING SCR SURGE & di/dt



The usual design compromise then is to use a garden variety fuse (3AG or 3AB style) which cannot be relied on to blow before the thyristor does, and trust that if the SCR does fail, it will fail short circuit. In the majority of the designs, this will be the case, though this is difficult to guarantee. Of course, a sufficiently high surge will cause an open. These comments also apply to the fuse in Figure 11B.

The value of di/dt that an SCR can safely handle is influenced by its construction and the characteristics of the gate drive signal. A center-gate-fire SCR has more di/dt capability than a corner-gate-fire type and heavily overdriving (3 to 5 times I_{GT}) the SCR gate with a fast (< 1 μs) rise time signal will maximize its di/dt capability. A typical maximum number in phase control SCRs of less than 50 Arms rating might be 200 A/μs, assuming a gate current of five times I_{GT} and < 1 μs rise time. If having done this, a di/dt problem is seen to still exist, the designer can also decrease the di/dt of the current waveform by adding inductance in series with the SCR, as shown in Figure 13. Of course, this reduces the circuit's ability to rapidly reduce the dc bus voltage and a tradeoff must be made between speedy voltage reduction and di/dt.

2. Surge Current

If the peak current and/or the duration of the surge is excessive, immediate destruction due to device overheating will result. The surge capability of the SCR is directly proportional to its die area. If the surge current cannot be reduced (by adding series resistance – see Figure 13) to a safe level which is consistent with the system's requirements for speedy bus voltage reduction, the designer must use a higher current SCR. This may result in the average current capability of the SCR exceeding the steady state current requirements imposed by the dc power supply.

A WORD ABOUT FUSING

Before leaving the subject of the crowbar SCR, a few words about fuse protection are in order. Refering back to Figure 11A, it will be seen that a fuse is necessary if the power supply to be protected is not output current limited. This fuse is not meant to prevent SCR failure but rather to prevent a fire!

In order to protect the SCR, the fuse would have to possess an I^2t rating less than that of the SCR and yet have a high enough continuous current rating to survive normal supply output currents. In addition, it must be capable of successfully clearing the high short circuit currents from the supply. Such a fuse as this is quite expensive, and may not even be available.

CROWBAR SCR SELECTION GUIDE

As an aid in selecting an SCR for crowbar use, the following selection guide is presented.

DEVICE	I_{RMS}	I_{TSM}	PACKAGE
2N6400 Series	16A	160A	TO220 Plastic
2N6504 Series	25A	160A	TO220 Plastic
2N1842 Series	16A	125A	Metal Stud
2N2573 Series	25A	260A	Metal TO-3 Type
2N681 Series	25A	200A	Metal Stud
MCR3935-1 Series	35A	350A	Metal Stud
MCR81-5 Series	80A	1000A	Metal Stud

MOTOROLA

MC3424 • MC3424A
MC3524 • MC3524A
MC3324 • MC3324A

Product Preview

POWER SUPPLY SUPERVISORY CIRCUIT/ DUAL VOLTAGE COMPARATOR

The MC3424 series is a dual-channel supervisory circuit, consisting of two uncommitted input comparators, a reference, output comparators, and high-current drive and indicator outputs for each channel. The input comparators feature programmable hysteresis, high common-mode rejection, and wide common-mode range, capable of comparing at ground potential with single-supply operation. Separate delay-filter pins are provided to increase noise immunity by delaying activation of the outputs. A 2.5 V bandgap voltage reference is pinned-out for referencing the input comparators, or other external functions. Independent high-current drive and indicator outputs for each channel can source and sink up to 300 mA and 50 mA respectively. CMOS/TTL compatible digital inputs provide remote activation of each channel's outputs. An input-enable pin allows control of the input comparators.

Although this device is intended for power supply supervision, the pinned-out reference, uncommitted-input comparator, and many other features, enable the MC3424 series to be utilized for a wide range of applications.

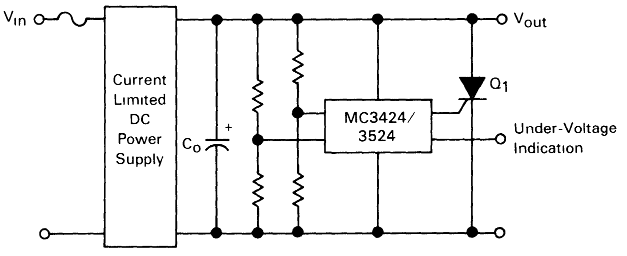
- Pinned-Out 2.5 V Reference
- Wide Common-Mode Range
- Programmable Hysteresis
- Programmable Time Delays
- Two 300 mA Drive Outputs
- Remote Activation Capability
- Wide Supply Range: 4.5 V $\leq V_{CC} \leq$ 40 V
- Low Current Drain

Applications

- Dual-Over Voltage "Crowbar" Protection
- Dual-Under Voltage Supervision
- Over/Under Voltage Protection
- Split-Supply Supervision
- Line-Loss Sensing
- Proportional Control
- Over/Under-Speed Indicator
- Sequential-Time Delay
- Battery Charging

TYPICAL APPLICATION

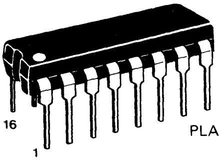
Over-Voltage Crowbar Protection, Under-Voltage Indication



This document contains information on a product under development. Motorola reserves the right to change or discontinue this product without notice

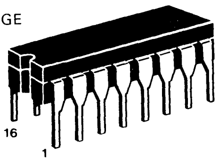
POWER SUPPLY SUPERVISORY CIRCUIT/DUAL VOLTAGE COMPARATOR

SILICON MONOLITHIC INTEGRATED CIRCUIT

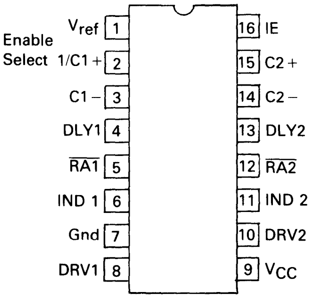


P SUFFIX
PLASTIC PACKAGE
CASE 648

L SUFFIX
CERAMIC PACKAGE
CASE 620



PIN CONNECTIONS



(Top view)

ORDERING INFORMATION

Device	Temperature Range	Package
MC3524L, AL	−55 to +125°C	Ceramic DIP
MC3324L, AL	−40 to + 85°C	Ceramic DIP
MC3324P, AP		Plastic DIP
MC3424L, AL	0 to + 70°C	Ceramic DIP
MC3424P, AP		Plastic DIP

FIGURE 1 — POWER SUPPLY OVERVOLTAGE PROTECTION (CROWBAR) AND LINE LOSS DETECTOR

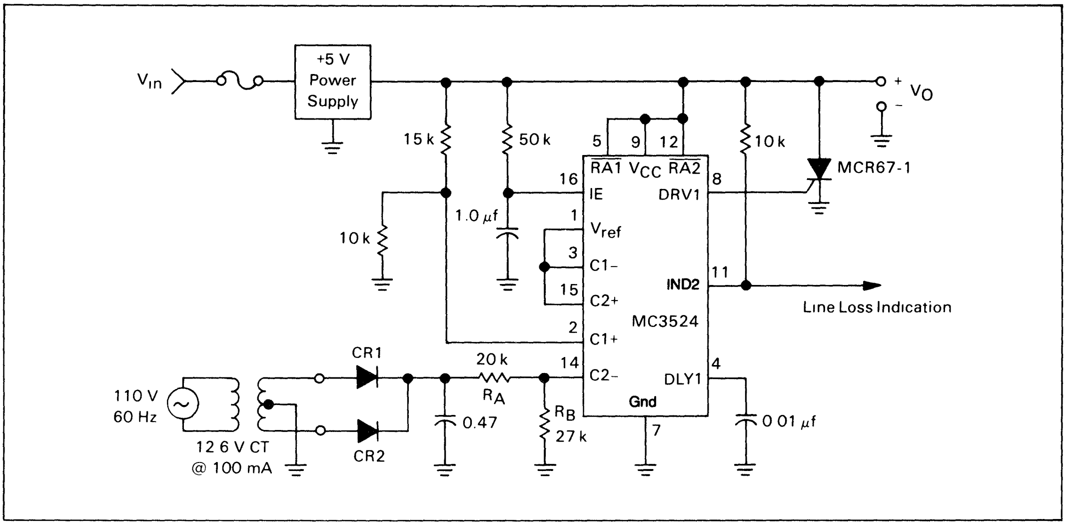
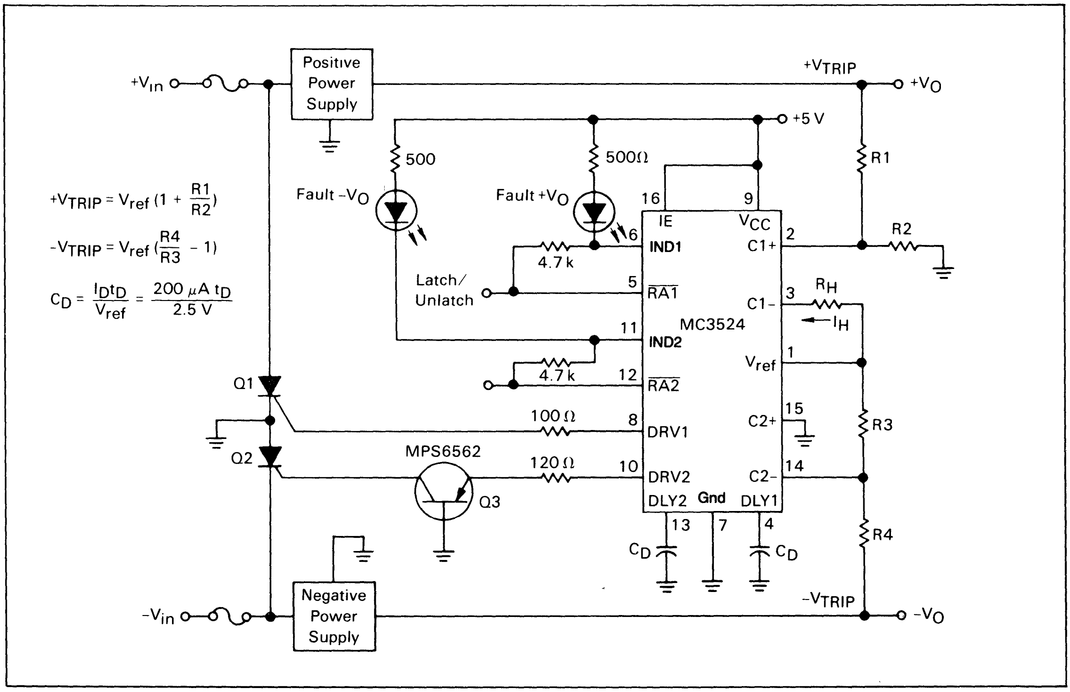
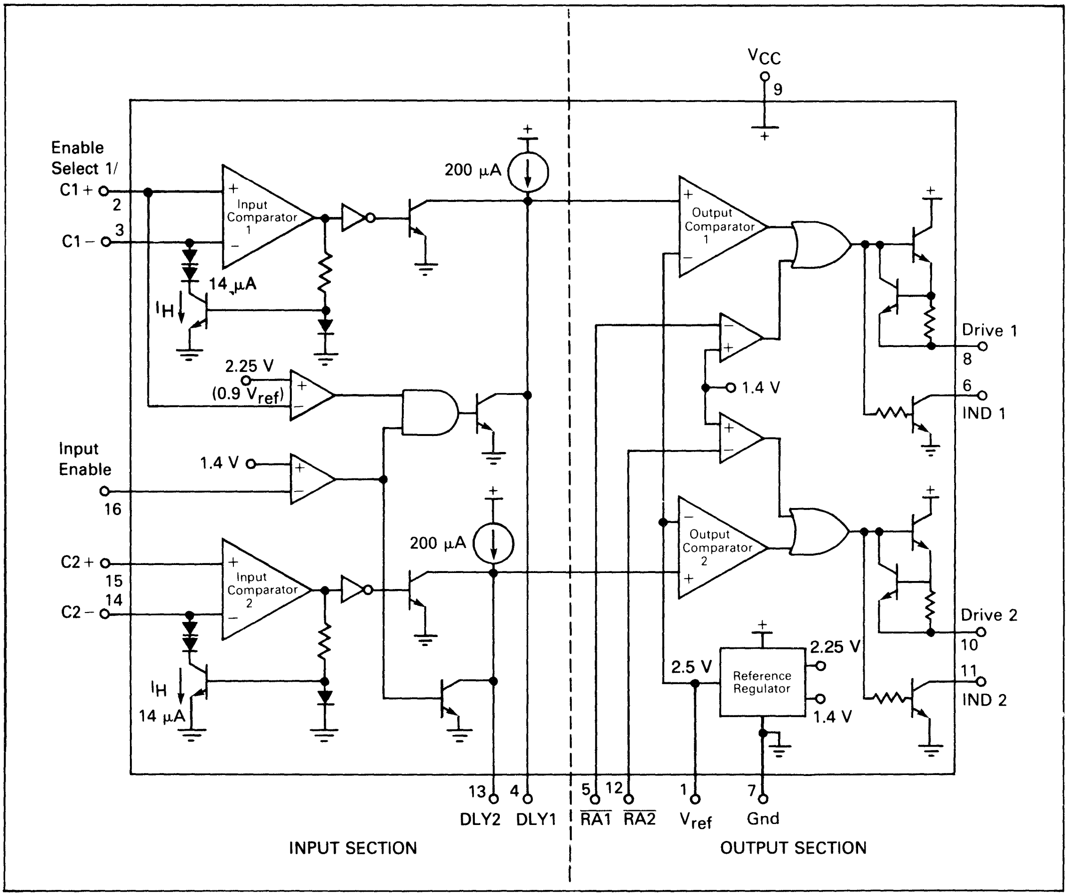


FIGURE 2 — OVERVOLTAGE PROTECTION, WITH DELAY, OF SPLIT SUPPLIES USING SCR "CROWBAR" SHUTDOWN AND LATCHED-FAULT INDICATION. (The Positive Sense is Chosen to Have $I_H R_H$ Hysteresis Voltage.)



MC3524/3424/3324 BLOCK DIAGRAM



Note: All voltages and currents are nominal.

MC34060
MC35060

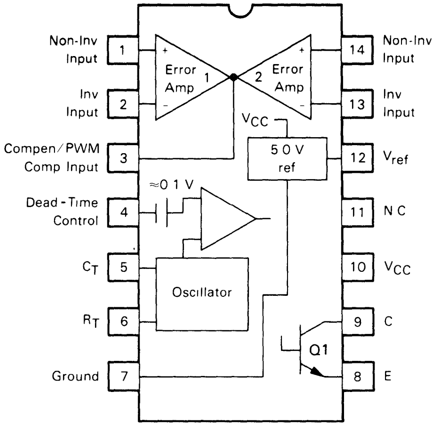
Specifications and Applications Information

SWITCHMODE PULSE WIDTH MODULATION CONTROL CIRCUITS

The MC35060 and MC34060 are low cost fixed frequency, pulse width modulation control circuits designed primarily for single ended SWITCHMODE power supply control. These devices feature:

- Complete Pulse Width Modulation Control Circuitry
- On-Chip Oscillator With Master or Slave Operation
- On-Chip Error Amplifiers
- On-Chip 5.0 Volt Reference
- Adjustable Dead Time Control
- Uncommitted Output Transistor for 200 mA Source or Sink

PIN CONNECTIONS

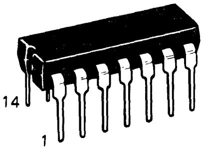


(top view)

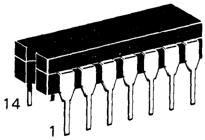
The MC34060 is specified over the commercial operating range of 0°C to +70°C. The MC35060 is specified over the full military range of -55 to +125°C.

SWITCHMODE PULSE WIDTH MODULATION CONTROL CIRCUITS

SILICON MONOLITHIC INTEGRATED CIRCUITS



P SUFFIX
PLASTIC PACKAGE
CASE 646



L SUFFIX
CERAMIC PACKAGE
CASE 632
(TO-116)

ORDERING INFORMATION

Device	Temperature Range	Package
MC35060L	-55 to +125°C	Ceramic DIP
MC34060P	0 to +70°C	Plastic DIP
MC34060L	0 to +70°C	Ceramic DIP

FIGURE 1 — BLOCK DIAGRAM

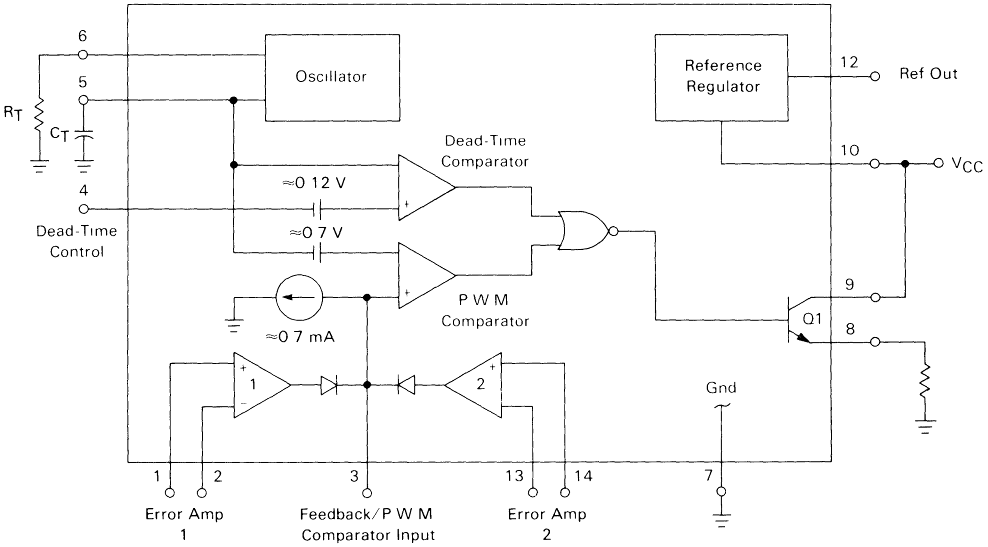
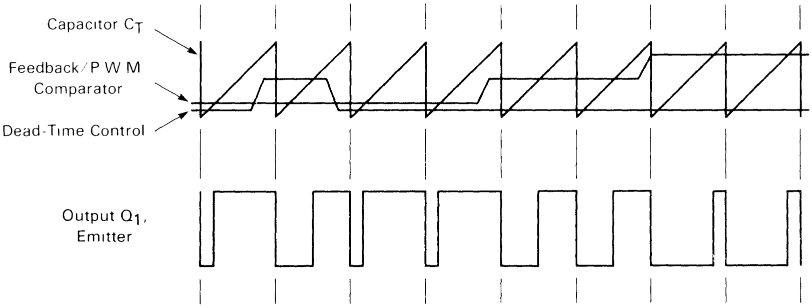


FIGURE 2 — TIMING DIAGRAM



Description

The MC35060/34060 is a fixed-frequency pulse width modulation control circuit, incorporating the primary building blocks required for the control of a switching power supply. (See Figure 1.) An internal-linear sawtooth oscillator is frequency-programmable by two external components, R_T and C_T. The oscillator frequency is determined by:

$$f_{osc} = \frac{1.1}{R_T \bullet C_T}$$

Output pulse width modulation is accomplished by comparison of the positive sawtooth waveform across capacitor C_T to either of two control signals. The output is enabled only during that portion of time when the sawtooth voltage is greater than the control signals. Therefore, an increase in control-signal amplitude causes a corresponding linear decrease of output pulse width. (Refer to the timing diagram shown in Figure 2.)

The control signals are external inputs that can be fed into the dead-time control, the error amplifier inputs, or the feedback input. The dead-time control comparator has an effective 120 mV input offset which limits the minimum output dead time to approximately the first 4% of the sawtooth-cycle time. This would result in a maximum duty cycle of 96%. Additional dead time may be imposed on the output by setting the dead time-control input to a fixed voltage, ranging between 0 to 3.3 V.

The pulse width modulator comparator provides a means for the error amplifiers to adjust the output pulse width from the maximum percent on-time, established by the dead time time control input, down to zero, as the voltage at the feedback pin varies from 0.5 to 3.5 V. Both error amplifiers have a common-mode input range from -0.3 V to (V_{CC} -2 V), and may be used to sense power supply output voltage and current. The error-amplifier outputs are active high and are ORed together at the non-inverting input of the pulse-width modulator comparator. With this configuration, the amplifier that demands minimum output on time, dominates control of the loop.

The MC35060/34060 has an internal 5.0 V reference capable of sourcing up to 10 mA of load currents for external bias circuits. The reference has an internal accuracy of ±5% with a thermal drift of less than 50 mV over an operating temperature range of 0 to +70°C.

MAXIMUM RATINGS (Full operating ambient temperature range applies unless otherwise noted)

Rating	Symbol	MC35060	MC34060	Unit
Power Supply Voltage	V_{CC}	42	42	V
Collector Output Voltage	V_C	42	42	V
Collector Output Current	I_C	250	250	mA
Amplifier Input Voltage	V_{in}	V_{CC} + 0.3	V_{CC} + 0.3	V
Power Dissipation @ $T_A \leqslant 45°C$	P_D	1000	1000	mW
Operating Junction Temperature	T_J	150	150	°C
Operating Ambient Temperature Range	T_A	-55 to 125	0 to 70	°C
Storage Temperature Range	T_{stg}	-65 to 150	-65 to 150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	L Suffix Ceramic Package	P Suffix Plastic Package	Unit
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	100	80	°C/W
Power Derating Factor	$1/R_{\theta JA}$	10	12.5	mW/°C
Derating Ambient Temperature	T_A	50	45	°C

RECOMMENDED OPERATING CONDITIONS

Condition/Value	Symbol	MC35060/MC34060			Unit
		Min	Typ	Max	
Power Supply Voltage	V_{CC}	7.0	15	40	V
Collector Output Voltage	V_C	—	30	40	V
Collector Output Current	I_C	—	—	200	mA
Amplifier Input Voltage	V_{in}	-0.3	—	V_{CC} -2.0	V
Current Into Feedback Terminal	$I_{f.b.}$	—	—	0.3	mA
Reference Output Current	I_{ref}	—	—	10	mA
Timing Resistor	R_T	1.8	47	500	kΩ
Timing Capacitor	C_T	0.00047	0.001	10	μF
Oscillator Frequency	f_{osc}	1.0	25	200	kHz

ELECTRICAL CHARACTERISTICS V_{CC} = 15 V, f_{osc} = 25 kHz unless otherwise noted. For typical values T_A = 25°C, for min/max values T_A is the operating ambient temperature range that applies unless otherwise noted.

Characteristic	Symbol	MC35060			MC34060			Unit
		Min	Typ	Max	Min	Typ	Max	
REFERENCE SECTION								
Reference Voltage (I_O = 1.0 mA)	V_{ref}	4.75	5.0	5.25	4.75	5.0	5.25	V
Reference Voltage Change with Temperature (ΔT_A = Min to Max)	$V_{ref(\Delta T)}$	—	0.2	2.0	—	1.3	2.6	%
Input Regulation (V_{CC} = 7.0 V to 40 V)	Reg_{line}	—	2.0	25	—	2.0	25	mV
Output Regulation (I_O = 1.0 mA to 10 mA)	Reg_{load}	—	3.0	15	—	3.0	15	mV
Short-Circuit Output Current (V_{ref} = 0 V, T_A = 25C)	I_{SC}	10	35	50	—	35	—	mA
OUTPUT SECTION								
Collector Off-State Current (V_{CC} = 40 V, V_{CE} = 40 V)	$I_{C(off)}$	—	2.0	100	—	2.0	100	μA
Emitter Off-State Current (V_{CC} = 40 V, V_C = 40 V, V_E = 0 V)	$I_{E(off)}$	—	—	-150	—	—	-100	μA
Collector-Emitter Saturation Voltage Common-Emitter (V_E = 0 V, I_C = 200 mA)	$V_{sat(C)}$	—	1.1	1.5	—	1.1	1.3	V
Emitter-Follower (V_C = 15 V, I_E = -200 mA)	$V_{sat(E)}$	—	1.5	2.5	—	1.5	2.5	V
Output Voltage Rise Time (T_A = 25°C) Common-Emitter (See Figure 12) Emitter-Follower (See Figure 13)	t_r	— —	100 100	200 200	— —	100 100	200 200	ns
Output Voltage Fall Time (T_A = 25°C) Common-Emitter (See Figure 12) Emitter-Follower (See Figure 13)	t_f	— —	25 40	100 100	— —	25 40	100 100	ns

Characteristic	Symbol	MC35060/MC34060			Unit
		Min	Typ	Max	
ERROR AMPLIFIER SECTIONS					
Input Offset Voltage ($V_{O[Pin\ 3]}$ = 2 5 V)	V_{IO}	—	2.0	10	mV
Input Offset Current ($V_{C[Pin\ 3]}$ = 2.5 V)	I_{IO}	—	5 0	250	nA
Input Bias Current ($V_{O[Pin\ 3]}$ = 2.5 V)	I_{IB}	—	0.1	1.0	μA
Input Common-Mode Voltage Range (V_{CC} = 7.0 V to 40 V)	V_{ICR}	-0.3	—	V_{CC}-2.0	V
Open Loop Voltage Gain (ΔV_O = 3.0 V, V_O = 0.5 to 3.5 V, R_L = 2.0 kΩ)	A_{VOL}	70	95	—	dB

ELECTRICAL CHARACTERISTICS V_{CC} = 15V, f_{osc} = 25 kHz unless otherwise noted. For typical values T_A = 25°C, for min/max values T_A is the operating ambient temperature range that applies unless otherwise noted.

Characteristic	Symbol	MC35060/MC34060			Unit
		Min.	Typ.	Max.	
ERROR AMPLIFIER SECTIONS (Continued)					
Unity-Gain Crossover Frequency (V_O = 0.5, to 3.5 V, R_L = 2.0 kΩ)	f_c	—	350	—	kHz
Phase Margin at Unity-Gain (V_O = 0.5 to 3.5 V, R_L = 2.0 kΩ)	ϕm	—	65	—	deg.
Common-Mode Rejection Ratio (V_{CC} = 40 V)	CMRR	65	90	—	dB
Power Supply Rejection Ratio (ΔV_{CC} = 33 V, V_O = 2.5 V, R_L = 2.0 kΩ)	PSRR	—	100	—	dB
Output Sink Current ($V_{O[Pin\ 3]}$ = 0.7 V)	I_{O-}	0.3	0.7	—	mA
Output Source Current ($V_{O[Pin\ 3]}$ = 3.5 V)	I_{O+}	-2.0	-4.0	—	mA
PWM COMPARATOR SECTION (Test circuit Figure 11)					
Input Threshold Voltage (Zero Duty Cycle)	V_{TH}	—	3.5	4.5	V
Input Sink Current ($V_{[Pin\ 3]}$ = 0.7 V)	I_{I-}	0.3	0.7	—	mA
DEAD-TIME CONTROL SECTION (Test Circuit Figure 11)					
Input Bias Current (Pin 4) (V_{in} = 0 to 5.25 V)	$I_{IB(DT)}$	—	-2.0	-10	μA
Maximum Output Duty Cycle (V_{in} = 0 V, C_T = 0.1 μF, R_T = 12 kΩ) (V_{in} = 0 V, C_T = 0.001 μF, R_T = 47 kΩ)	DC_{max}	 90 —	 96 92	 100 100	%
Input Threshold Voltage (Pin 4) (Zero Duty Cycle) (Maximum Duty Cycle)	V_{TH}	 — 0	 2.8 —	 3.3 —	V
OSCILLATOR SECTION					
Frequency (C_T = 0.001 μF, R_T = 47 kΩ)	f_{osc}	—	25	—	kHz
Standard Deviation of Frequency* (C_T = 0.001 μF, R_T = 47 kΩ)	σf_{osc}	—	3.0	—	%
Frequency Change with Voltage (V_{CC} = 7.0 V to 40 V, T_A = 25°C)	$\Delta f_{osc}(\Delta V)$	—	0.1	—	%
Frequency Change with Temperature (ΔT_A = 25°C to $T_{A\ low}$, 25°C to $T_{A\ high}$)	$\Delta f_{osc}(\Delta T)$	—	1.0	2.0	%
TOTAL DEVICE					
Standby Supply Current (Pin 6 at V_{ref}, all other inputs and outputs open) (V_{CC} = 15 V) (V_{CC} = 40 V)	I_{CC}	 — —	 5.5 7.0	 10 15	mA
Average Supply Current ($V_{[Pin\ 4]}$ = 2.0 V, C_T = 0.001, R_T = 47 kΩ). See Figure 11.	I_S	—	7.0	—	mA

*Standard deviation is a measure of the statistical distribution about the mean as derived from the formula, $\sigma = \sqrt{\dfrac{\sum\limits_{n=1}^{N} (X_n - \bar{X})^2}{N-1}}$

FIGURE 3 — OSCILLATOR FREQUENCY versus TIMING RESISTANCE

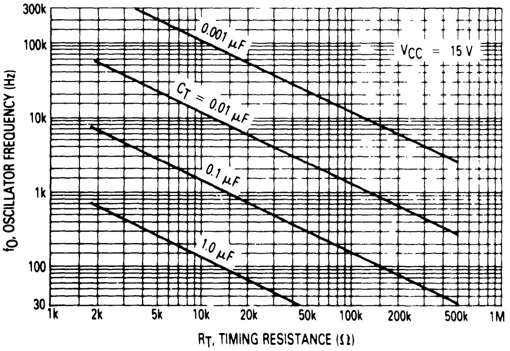


FIGURE 4 — OPEN LOOP VOLTAGE GAIN AND PHASE versus FREQUENCY

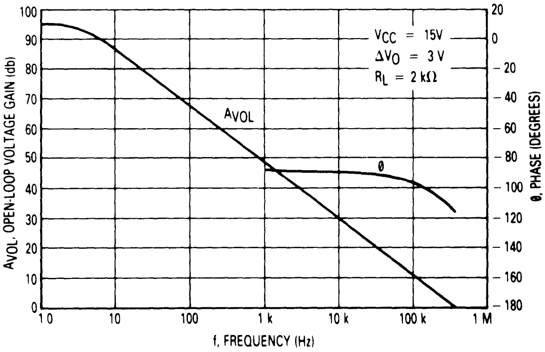


FIGURE 5 — PERCENT DEAD-TIME versus OSCILLATOR FREQUENCY

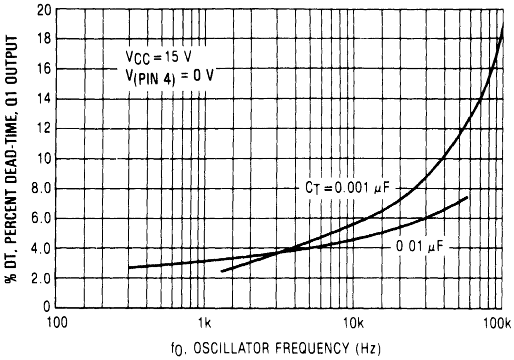


FIGURE 6 — PERCENT DUTY CYCLE versus DEAD-TIME CONTROL VOLTAGE

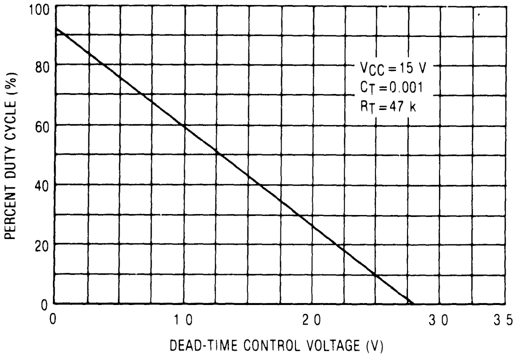


FIGURE 7 — EMITTER FOLLOWER CONFIGURATION OUTPUT-SATURATION VOLTAGE versus EMITTER CURRENT

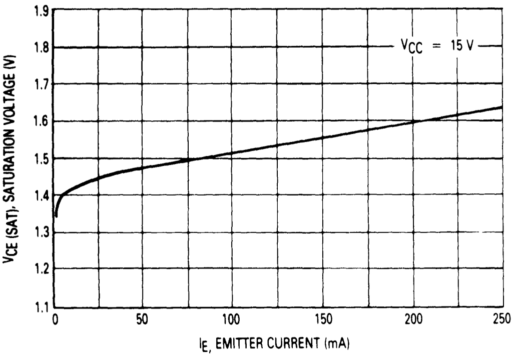


FIGURE 8 — COMMON EMITTER CONFIGURATION OUTPUT-SATURATION VOLTAGE versus EMITTER CURRENT

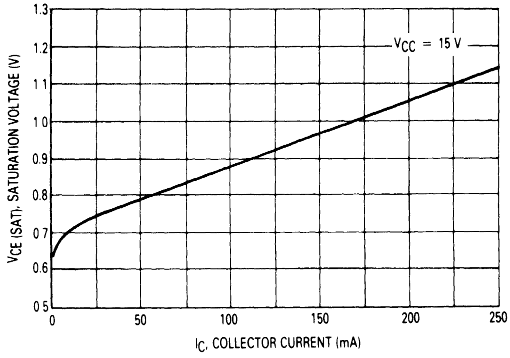


FIGURE 9 — STANDBY-SUPPLY CURRENT versus SUPPLY VOLTAGE

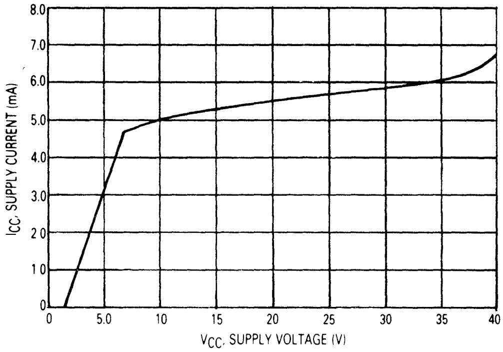


FIGURE 10 — ERROR AMPLIFIER CHARACTERISTICS

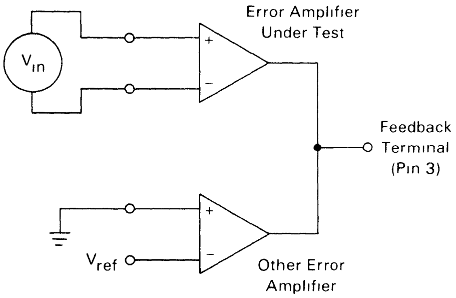


FIGURE 11 — DEAD-TIME AND FEEDBACK CONTROL TEST CIRCUIT

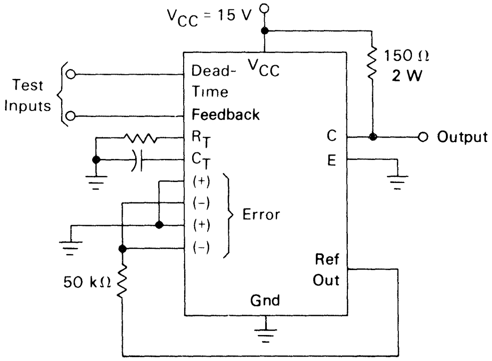


FIGURE 12 — COMMON-EMITTER CONFIGURATION TEST CIRCUIT AND WAVEFORM

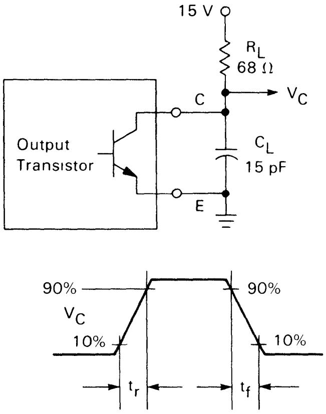


FIGURE 13 — EMITTER-FOLLOWER CONFIGURATION TEST CIRCUIT AND WAVEFORM

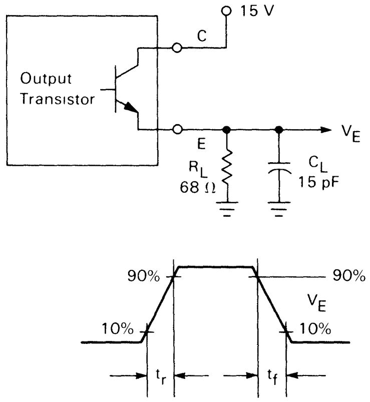
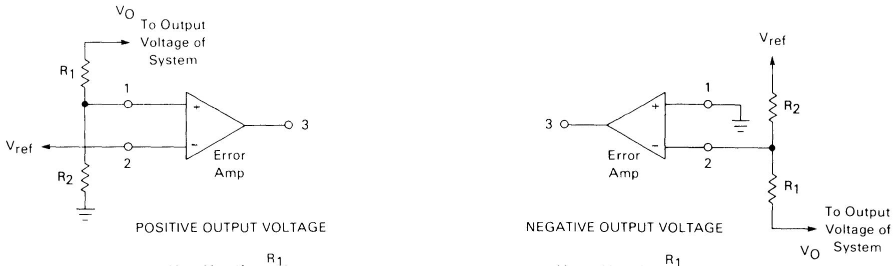


FIGURE 14 — ERROR AMPLIFIER SENSING TECHNIQUES



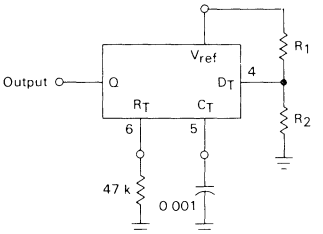
POSITIVE OUTPUT VOLTAGE

$$V_O = V_{ref}\left(1 + \frac{R_1}{R_2}\right)$$

NEGATIVE OUTPUT VOLTAGE

$$V_O = -V_{ref}\left(1 + \frac{R_1}{R_2}\right)$$

FIGURE 15 — DEAD-TIME CONTROL CIRCUIT



$$\text{Max \% On Time} \approx 92 - \left(\frac{160}{1 + \frac{R1}{R2}}\right)$$

FIGURE 16 — SOFT-START CIRCUIT

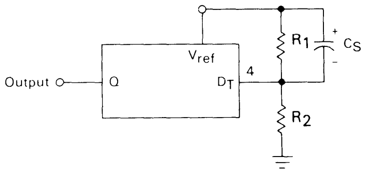


FIGURE 17 — SLAVING TWO OR MORE CONTROL CIRCUITS

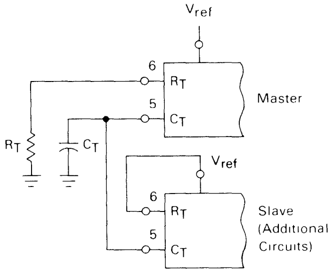
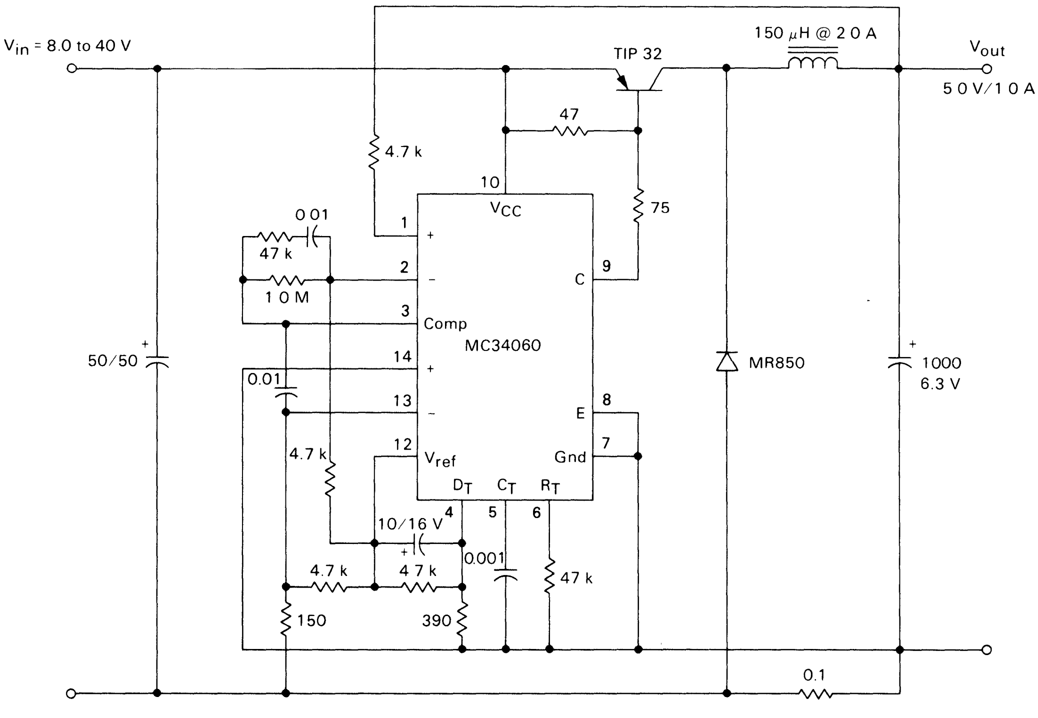
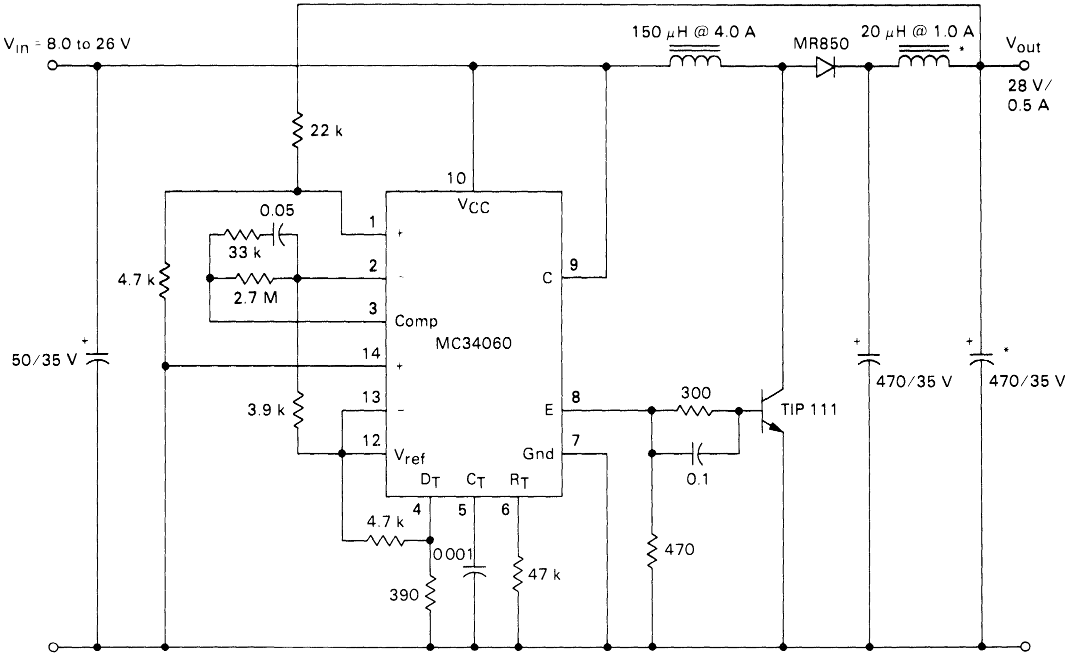


FIGURE 18 — STEP-DOWN CONVERTER WITH SOFT-START AND OUTPUT CURRENT LIMITING



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 8.0 V to 40 V, I_O = 1 0 A	25mV 0 5%
Load Regulation	V_{in} = 12 V, I_O = 1.0 mA to 1 0 A	3 0 mV 0 06%
Output Ripple	V_{in} = 12 V, I_O = 1.0 A	75 mV p-p P A R D
Short Circuit Current	V_{in} = 12 V, R_L = 0.1 Ω	1 6 A
Efficiency	V_{in} = 12 V, I_O = 1 0 A	73%

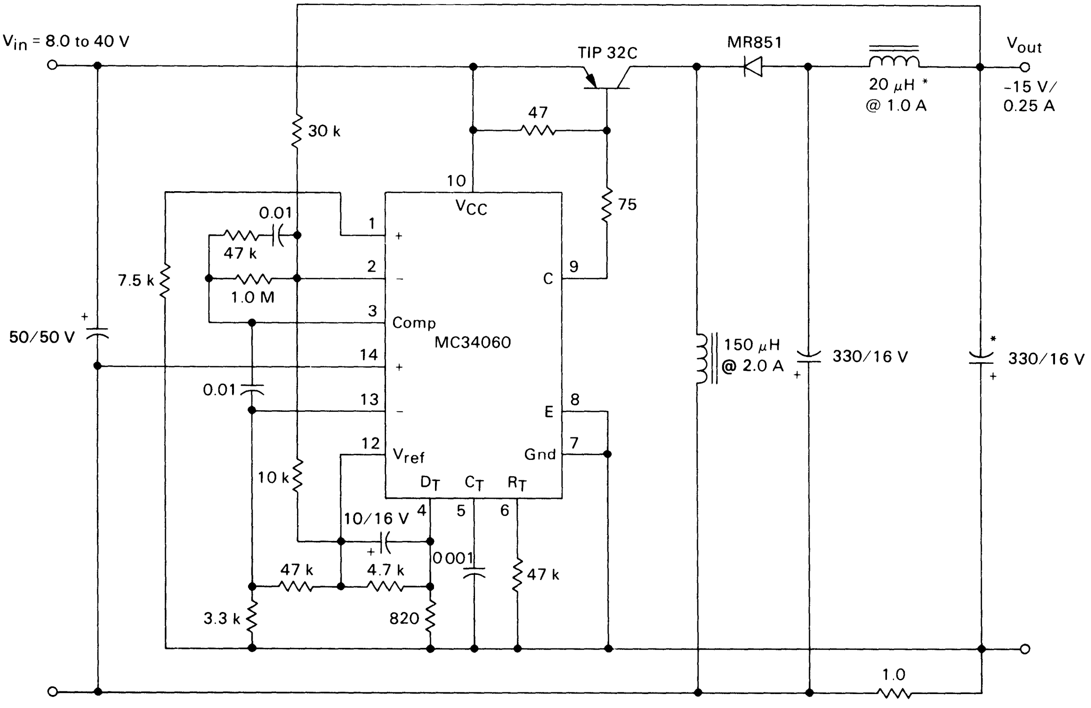
FIGURE 19 — STEP-UP CONVERTER



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 8.0 V to 26 V, I_O = 0 5 A	40mV 0 14%
Load Regulation	V_{in} = 12 V, I_O = 1 0 mA to 0 5 A	5 0 mV 0 18%
Output Ripple	V_{in} = 12 V, I_O = 0 5 A	24 mV p-p P A R D
Efficiency	V_{in} = 12 V, I_O = 0 5 A	75%

*Optional circuit to minimize output ripple

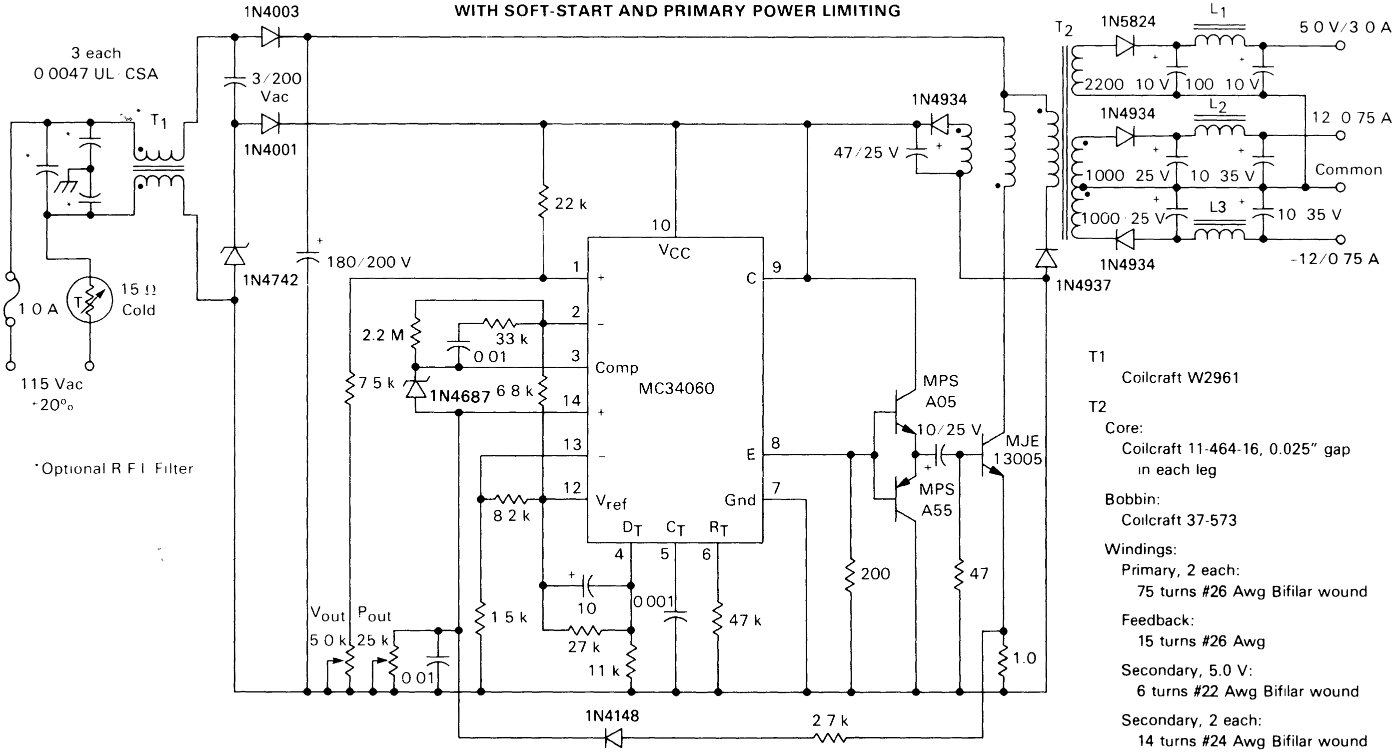
FIGURE 20 — STEP-UP/DOWN VOLTAGE INVERTING CONVERTER WITH SOFT-START AND CURRENT LIMITING



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 8.0 V to 40 V, I_O = 250 mA	52 mV 0.35%
Load Regulation	V_{in} = 12 V, I_O = 1 mA to 250 mA	47 mV 0.32%
Output Ripple	V_{in} = 12 V, I_O = 250 mA	10 mV p.p. P.A.R.D.
Short Circuit Current	V_{in} = 12 V, R_L = 0.1 Ω	330 mA
Efficiency	V_{in} = 12 V, I_O = 250 mA	86%

*Optional circuit to minimize output ripple

FIGURE 21 — 33 WATT OFF-LINE FLYBACK CONVERTER WITH SOFT-START AND PRIMARY POWER LIMITING



T1

Coilcraft W2961

T2

Core:

Coilcraft 11-464-16, 0.025" gap in each leg

Bobbin:

Coilcraft 37-573

Windings:

Primary, 2 each:

75 turns #26 Awg Bifilar wound

Feedback:

15 turns #26 Awg

Secondary, 5.0 V:

6 turns #22 Awg Bifilar wound

Secondary, 2 each:

14 turns #24 Awg Bifilar wound

L1

Coilcraft Z7156, 15 μH @ 5.0 A

L2, L3

Coilcraft Z7157, 25 μH @ 1.0 A

TEST	CONDITIONS	RESULTS
Line Regulation 5 0 V	V_{in} = 95 to 135 Vac, I_O = 3 0 A	20 mV 0 40%
Line Regulation ±12 V	V_{in} = 95 to 135 Vac, I_O = ±0 75 A	52 mV 0 26%
Load Regulation 5.0 V	V_{in} = 115 Vac, I_O = 1 0 to 4 0 A	476 mV 9 5%
Load Regulation ±12 V	V_{in} = 115 Vac, I_O = ±0.4 to ±0.9 A	300 mV 2.5%
Output Ripple 5 0 V	V_{in} = 115 Vac, I_O = 3 0 A	45 mV p-p P A R D
Output Ripple ±12 V	V_{in} = 115 Vac, I_O = ±0 75 A	75 mV p-p P A R D
Efficiency	V_{in} = 115 Vac, I_O 5 0 V = 3 0 A I_O ±12 = ±0.75 A	74%

MOTOROLA

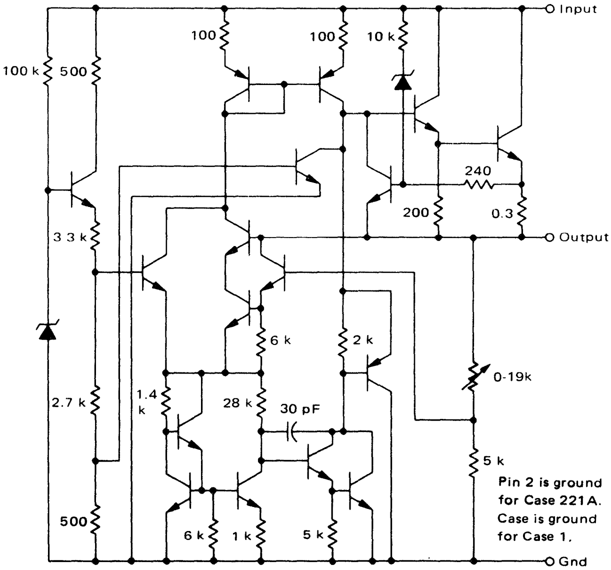
MC7800 Series

3-TERMINAL POSITIVE VOLTAGE REGULATORS

These voltage regulators are monolithic integrated circuits designed as fixed-voltage regulators for a wide variety of applications including local, on-card regulation. These regulators employ internal current limiting, thermal shutdown, and safe-area compensation. With adequate heatsinking they can deliver output currents in excess of 1.0 ampere. Although designed primarily as a fixed voltage regulator, these devices can be used with external components to obtain adjustable voltages and currents.

- Output Current in Excess of 1.0 Ampere
- No External Components Required
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Output Voltage Offered in 2% and 4% Tolerance

SCHEMATIC DIAGRAM

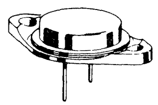


Pin 2 is ground for Case 221A. Case is ground for Case 1.

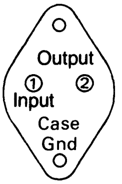
ORDERING INFORMATION

Device	Output Voltage Tolerance	Temperature Range	Package
MC78XXK MC78XXAK	4% 2%	-55 to +150°C	Metal Power
MC78XXBK	4%	-40 to +125°C	
MC78XXCK MC78XXACK	4% 2%	0 to +125°C	
MC78XXCT MC78XXACT	4% 2%		Plastic Power
MC78XXBT	4%	-40 to +125°C	

THREE-TERMINAL POSITIVE FIXED VOLTAGE REGULATORS

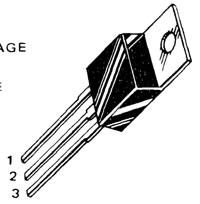


K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 TYPE)



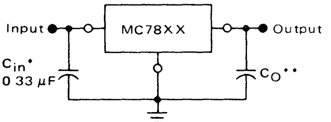
(Bottom View)

T SUFFIX
PLASTIC PACKAGE
CASE 221A
TO-220 TYPE



Pin 1. Input
2. Ground
3. Output

STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.0 V above the output voltage even during the low point on the input ripple voltage.

XX = these two digits of the type number indicate voltage.

* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_O is not needed for stability; however, it does improve transient response.

XX indicates nominal voltage

TYPE NO /VOLTAGE			
MC7805	5.0 Volts	MC7815	15 Volts
MC7806	6.0 Volts	MC7818	18 Volts
MC7808	8.0 Volts	MC7824	24 Volts
MC7812	12 Volts		

MC7800 Series MAXIMUM RATINGS (T_A = +25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Input Voltage (5.0 V - 18 V)	V_{in}	35	Vdc
(24 V)		40	
Power Dissipation and Thermal Characteristics			
Plastic Package			
T_A = +25°C	P_D	Internally Limited	Watts
Derate above T_A = +25°C	$1/\theta_{JA}$	15.4	mW/°C
Thermal Resistance, Junction to Air	θ_{JA}	65	°C/W
T_C = +25°C	P_D	Internally Limited	Watts
Derate above T_C = +95°C (See Figure 1)	$1/\theta_{JC}$	200	mW/°C
Thermal Resistance, Junction to Case	θ_{JC}	5.0	°C/W
T_C = +25°C	P_D	Internally Limited	Watts
Derate above T_A = +25°C	$1/\theta_{JA}$	22.5	mW/°C
Thermal Resistance, Junction to Air	θ_{JA}	45	°C/W
T_C = +25°C	P_D	Internally Limited	Watts
Derate above T_C = +65°C (See Figure 2)	$1/\theta_{JC}$	182	mW/°C
Thermal Resistance, Junction to Case	θ_{JC}	5.5	°C/W
Storage Junction Temperature Range	T_{stg}	-65 to +150	°C
Operating Junction Temperature Range	T_J		°C
MC7800, A		-55 to +150	
MC7800C, AC		0 to +150	
MC7800, B		-40 to +150	

DEFINITIONS

Line Regulation — The change in output voltage for a change in the input voltage The measurement is made under conditions of low dissipation or by using pulse techniques such that the average chip temperature is not significantly affected.

Load Regulation — The change in output voltage for a change in load current at constant chip temperature.

Maximum Power Dissipation — The maximum total device dissipation for which the regulator will operate within specifications.

Quiescent Current — That part of the input current that is not delivered to the load.

Output Noise Voltage — The rms ac voltage at the output, with constant load and no input ripple, measured over a specified frequency range

Long Term Stability — Output voltage stability under accelerated life test conditions with the maximum rated voltage listed in the devices' electrical characteristics and maximum power dissipation

MC7805, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 10 V, I_O = 500 mA, $T_J = T_{low}$ to T_{high} [Note 1] unless otherwise noted).

Characteristic	Symbol	MC7805			MC7805B			MC7805C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	4.8	5.0	5.2	4.8	5.0	5.2	4.8	5.0	5.2	Vdc
Output Voltage (5.0 mA $\leq I_O \leq$ 1.0 A, $P_O \leq$ 15 W)	V_O										Vdc
7.0 Vdc $\leq V_{in} \leq$ 20 Vdc		—	—	—	—	—	—	4.75	5.0	5.25	
8.0 Vdc $\leq V_{in} \leq$ 20 Vdc		4.65	5.0	5.35	4.75	5.0	5.25	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
7.0 Vdc $\leq V_{in} \leq$ 25 Vdc		—	2.0	50	—	7.0	100	—	7.0	100	
8.0 Vdc $\leq V_{in} \leq$ 12 Vdc		—	1.0	25	—	2.0	50	—	2.0	50	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA $\leq I_O \leq$ 1.5 A		—	25	100	—	40	100	—	40	100	
250 mA $\leq I_O \leq$ 750 mA		—	8.0	25	—	15	50	—	15	50	
Quiescent Current (T_J = +25°C)	I_B	—	3.2	6.0	—	4.3	8.0	—	4.3	8.0	mA
Quiescent Current Change	ΔI_B										mA
7.0 Vdc $\leq V_{in} \leq$ 25 Vdc		—	—	—	—	—	—	—	—	1.3	
8.0 Vdc $\leq V_{in} \leq$ 25 Vdc		—	0.3	0.8	—	—	1.3	—	—	—	
5 0 mA $\leq I_O \leq$ 1.0 A		—	0.04	0.5	—	—	0.5	—	—	0 5	
Ripple Rejection 8 0 Vdc $\leq V_{in} \leq$ 18 Vdc, f = 120 Hz	RR	68	75	—	—	68	—	—	68	—	dB
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2 5	—	2.0	—	—	2 0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leq$ f $\leq$ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	$\mu V/V_O$
Output Resistance f = 1 0 kHz	R_O	—	17	—	—	17	—	—	17	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0 2	1.2	—	0.2	—	—	0 2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1 3	2 5	3.3	—	2.2	—	—	2 2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±0.6	—	—	-1.1	—	—	-1 1	—	mV/°C

MC7805A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 10 V, I_O = 1 0 A, $T_J = T_{low}$ to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7805A			MC7805AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	4.9	5.0	5 1	4 9	5 0	5 1	Vdc
Output Voltage (5 0 mA $\leq I_O \leq$ 1 0 A, $P_O \leq$ 15 W) 7 5 Vdc $\leq V_{in} \leq$ 20 Vdc	V_O	4 8	5 0	5.2	4 8	5 0	5 2	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
7 5 Vdc $\leq V_{in} \leq$ 25 Vdc, I_O = 500 mA		—	2.0	10	—	7 0	50	
8 0 Vdc $\leq V_{in} \leq$ 12 Vdc		—	3.0	10	—	10	50	
8 0 Vdc $\leq V_{in} \leq$ 12 Vdc, T_J = +25°C		—	1.0	4 0	—	2 0	25	
7 3 Vdc $\leq V_{in} \leq$ 20 Vdc, T_J = +25°C		—	2.0	10	—	7 0	50	
Load Regulation (Note 2)	Reg_{load}							mV
5 0 mA $\leq I_O \leq$ 1 5 A		—	25	50	—	—	—	
5 0 mA $\leq I_O \leq$ 1.0 A		—	—	—	—	25	100	
5 0 mA $\leq I_O \leq$ 1.5 A, T_J = +25°C		—	—	—	—	25	100	
250 mA $\leq I_O \leq$ 750 mA		—	8 0	25	—	8 0	50	
Quiescent Current	I_B	—	—	5 0	—	—	6.0	mA
T_J = +25°C		—	3 2	4 0	—	4 3	6 0	
Quiescent Current Change	ΔI_B							mA
8 0 Vdc $\leq V_{in} \leq$ 25 Vdc, I_O = 500 mA		—	0.3	0.5	—	—	0 8	
7 5 Vdc $\leq V_{in} \leq$ 20 Vdc, T_J = +25°C		—	0.2	0.5	—	—	0 8	
5 0 mA $\leq I_O \leq$ 1.0 A		—	0 04	0.2	—	—	0 5	
Ripple Rejection	RR							dB
8 0 Vdc $\leq V_{in} \leq$ 18 Vdc, f = 120 Hz, T_J = +25°C		68	75	—	—	—	—	
8 0 Vdc $\leq V_{in} \leq$ 18 Vdc, f = 120 Hz, I_O = 500 mA		68	75	—	—	68	—	
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leq$ f $\leq$ 100 kHz	V_n	—	10	40	—	10	—	$\mu V/V_O$
Output Resistance (f = 1.0 kHz)	R_O	—	17	—	—	17	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±0.6	—	—	-1.1	—	mV/°C

NOTES: 1 T_{low} = -55°C for MC78XX, A $\quad T_{high}$ = +150°C for MC78XX, A
= 0° for MC78XXC, AC $\quad$ = +125°C for MC78XXC, AC, B
= -40°C for MC78XXB

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

MC7806, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 11 V, I_O = 500 mA, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted).

Characteristic	Symbol	MC7806			MC7806B			MC7806C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	5.75	6.0	6.25	5.75	6.0	6.25	5 75	6.0	6.25	Vdc
Output Voltage	V_O										Vdc
(5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W)											
8.0 Vdc $\leqslant V_{in} \leqslant$ 21 Vdc		—	—	—	—	—	—	5 7	6 0	6 3	
9.0 Vdc $\leqslant V_{in} \leqslant$ 21 Vdc		5.65	6.0	6.35	5 7	6.0	6.3	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
8.0 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc		—	3 0	60	—	9 0	120	—	9 0	120	
9 0 Vdc $\leqslant V_{in} \leqslant$ 13 Vdc		—	2.0	30	—	3.0	60	—	3 0	60	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA $\leqslant I_O \leqslant$ 1.5 A		—	27	100	—	43	120	—	43	120	
250 mA $\leqslant I_O \leqslant$ 750 mA		—	9.0	30	—	16	60	—	16	60	
Quiescent Current (T_J = +25°C)	I_B	—	3 2	6 0	—	4.3	8 0	—	4 3	8.0	mA
Quiescent Current Change	ΔI_B										mA
8 0 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc		—	—	—	—	—	—	—	—	1 3	
9 0 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc		—	0 3	0 8	—	—	1 3	—	—	—	
5.0 mA $\leqslant I_O \leqslant$ 1.0 A		—	0 04	0 5	—	—	0 5	—	—	0 5	
Ripple Rejection 9 0 Vdc $\leqslant V_{in} \leqslant$ 19 Vdc, f = 120 Hz	RR	65	73	—	—	65	—	—	65	—	dB
Dropout Voltage (I_O = 1 0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2 5	—	2 0	—	—	2 0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	µV/V_O
Output Resistance f = 1 0 kHz	R_O	—	17	—	—	17	—	—	17	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0 2	1 2	—	0.2	—	—	0 2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1 3	2.5	3 3	—	2 2	—	—	2 2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±0.7	—	—	-0 8	—	—	-0 8	—	mV/°C

MC7806A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 11 V, I_O = 1 0 A, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7806A			MC7806AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	5 88	6 0	6 12	5 88	6 0	6 12	Vdc
Output Voltage (5 0 mA $\leqslant I_O \leqslant$ 1 0 A, $P_O \leqslant$ 15 W) 8 6 Vdc $\leqslant V_{in} \leqslant$ 21 Vdc	V_O	5 76	6 0	6 24	5 76	6 0	6 24	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
8 6 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc, I_O = 500 mA		—	3 0	11	—	9 0	60	
9 0 Vdc $\leqslant V_{in} \leqslant$ 13 Vdc		—	5 0	15	—	11	60	
9 0 Vdc $\leqslant V_{in} \leqslant$ 13 Vdc, T_J = +25°C		—	2 0	5 0	—	3 0	30	
8 3 Vdc $\leqslant V_{in} \leqslant$ 21 Vdc, T_J = +25°C		—	4 0	11	—	9 0	60	
Load Regulation (Note 2)	Reg_{load}							mV
5 0 mA $\leqslant I_O \leqslant$ 1 5 A		—	27	50	—	—	—	
5 0 mA $\leqslant I_O \leqslant$ 1 0 A		—	—	—	—	43	100	
5 0 mA $\leqslant I_O \leqslant$ 1 5 A, T_J = +25°C		—	—	—	—	43	100	
250 mA $\leqslant I_O \leqslant$ 750 mA		—	9 0	25	—	16	50	
Quiescent Current	I_B	—	—	5 0	—	—	6 0	mA
T_J = +25°C		—	3 2	4.0	—	4 3	6 0	
Quiescent Current Change	ΔI_B							mA
9 0 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc, I_O = 500 mA		—	0.3	0.5	—	—	0 8	
8 6 Vdc $\leqslant V_{in} \leqslant$ 21 Vdc, T_J = +25°C		—	0 2	0 5	—	—	0 8	
5.0 mA $\leqslant I_O \leqslant$ 1 0 A		—	0 04	0 2	—	—	0.5	
Ripple Rejection	RR							dB
9 0 Vdc $\leqslant V_{in} \leqslant$ 19 Vdc, f = 120 Hz, T_J = +25°C		65	73	—	—	—	—	
9.0 Vdc $\leqslant V_{in} \leqslant$ 19 Vdc, f = 120 Hz, I_O = 500 mA		65	73	—	—	65	—	
Dropout Voltage (I_O = 1 0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2 0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	10	40	—	10	—	µV/V_O
Output Resistance (f = 1.0 kHz)	R_O	—	17	—	—	17	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1 2	—	0 2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±0.7	—	—	-0.8	—	mV/°C

NOTES: 1. T_{low} = -55°C for MC78XX, A $\quad T_{high}$ = +150°C for MC78XX, A
= 0° for MC78XXC, AC $\quad$ = +125°C for MC78XXC, AC, B
= -40°C for MC78XXB

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

MC7808, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 14 V, I_O = 500 mA, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted).

Characteristic	Symbol	MC7808			MC7808B			MC7808C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	7.7	8.0	8.3	7.7	8.0	8.3	7.7	8.0	8.3	Vdc
Output Voltage (5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W)	V_O										Vdc
10.5 Vdc $\leqslant V_{in} \leqslant$ 23 Vdc		—	—	—	—	—	—	7.6	8.0	8.4	
11.5 Vdc $\leqslant V_{in} \leqslant$ 23 Vdc		7.6	8.0	8.4	7.6	8.0	8.4	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
10.5 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc		—	3.0	80	—	12	160	—	12	160	
11 Vdc $\leqslant V_{in} \leqslant$ 17 Vdc		—	2.0	40	—	5.0	80	—	5.0	80	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA $\leqslant I_O \leqslant$ 1.5 A		—	28	100	—	45	160	—	45	160	
250 mA $\leqslant I_O \leqslant$ 750 mA		—	9.0	40	—	16	80	—	16	80	
Quiescent Current (T_J = +25°C)	I_B	—	3.2	6.0	—	4.3	8.0	—	4.3	8.0	mA
Quiescent Current Change	ΔI_B										mA
10.5 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc		—	—	—	—	—	—	—	—	1 0	
11.5 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc		—	0.3	0.8	—	—	1.0	—	—	—	
5.0 mA $\leqslant I_O \leqslant$ 1.0 A		—	0.04	0.5	—	—	0 5	—	—	0 5	
Ripple Rejection 11 5 Vdc $\leqslant V_{in} \leqslant$ 21.5 Vdc, f = 120 Hz	RR	62	70	—	—	62	—	—	62	—	dB
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	µV/ V_O
Output Resistance f = 1.0 kHz	R_O	—	18	—	—	18	—	—	18	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1 3	2 5	3.3	—	2 2	—	—	2 2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±1.0	—	—	-0.8	—	—	-0 8	—	mV/ °C

MC7808A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 14 V, I_O = 1.0 A, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7808A			MC7808AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	7.84	8.0	8 16	7 84	8 0	8 16	Vdc
Output Voltage (5 0 mA $\leqslant I_O \leqslant$ 1 0 A, $P_O \leqslant$ 15 W) 10 6 Vdc $\leqslant V_{in} \leqslant$ 23 Vdc	V_O	7 7	8 0	8 3	7 7	8 0	8 3	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
10.6 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc, I_O = 500 mA		—	4.0	13	—	12	80	
11 Vdc $\leqslant V_{in} \leqslant$ 17 Vdc		—	6.0	20	—	15	80	
11 Vdc $\leqslant V_{in} \leqslant$ 17 Vdc, T_J = +25°C		—	2 0	6 0	—	5 0	40	
10 4 Vdc $\leqslant V_{in} \leqslant$ 23 Vdc, T_J = +25°C		—	4.0	13	—	12	80	
Load Regulation (Note 2)	Reg_{load}							mV
5 0 mA $\leqslant I_O \leqslant$ 1.5 A		—	28	50	—	—	—	
5.0 mA $\leqslant I_O \leqslant$ 1.0 A		—	—	—	—	45	100	
5.0 mA $\leqslant I_O \leqslant$ 1.5 A, T_J = +25°C		—	—	—	—	45	100	
250 mA $\leqslant I_O \leqslant$ 750 mA		—	9 0	25	—	16	50	
Quiescent Current	I_B	—	—	5 0	—	—	6 0	mA
T_J = +25°C		—	3.2	4 0	—	4 3	6 0	
Quiescent Current Change	ΔI_B							mA
11 Vdc $\leqslant V_{in} \leqslant$ 25 Vdc, I_O = 500 mA		—	0.3	0 5	—	--	0 8	
10 6 Vdc $\leqslant V_{in} \leqslant$ 23 Vdc, T_J = +25°C		—	0.2	0 5	—	—	0 8	
5 0 mA $\leqslant I_O \leqslant$ 1.0 A		—	0 04	0 2	—	—	0 5	
Ripple Rejection	RR							dB
11 5 Vdc $\leqslant V_{in} \leqslant$ 21.5 Vdc, f = 120 Hz, T_J = +25°C		62	70	—	—	—	—	
11.5 Vdc $\leqslant V_{in} \leqslant$ 21.5 Vdc, f = 120 Hz, I_O = 500 mA		62	70	—	—	62	—	
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2 0	2 5	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	10	40	—	10	—	µV/V_O
Output Resistance (f = 1.0 kHz)	R_O	—	18	—	—	18	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0 2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±1.0	—	—	-0 8	—	mV/°C

NOTES: 1. T_{low} = −55°C for MC78XX, A
= 0° for MC78XXC, AC
= −40°C for MC78XXB

T_{high} = +150°C for MC78XX, A
= +125°C for MC78XXC, AC, B

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

MC7812, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 19 V, I_O = 500 mA, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted)

Characteristic	Symbol	MC7812			MC7812B			MC7812C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	11.5	12	12.5	11.5	12	12.5	11.5	12	12.5	Vdc
Output Voltage (5.0 mA ≤ I_O ≤ 1.0 A, P_O ≤ 15 W)	V_O										Vdc
14.5 Vdc ≤ V_{in} ≤ 27 Vdc		—	—	—	—	—	—	11.4	12	12.6	
15.5 Vdc ≤ V_{in} ≤ 27 Vdc		11.4	12	12.6	11.4	12	12.6	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
14.5 Vdc ≤ V_{in} ≤ 30 Vdc		—	5.0	120	—	13	240	—	13	240	
16 Vdc ≤ V_{in} ≤ 22 Vdc		—	3.0	60	—	6.0	120	—	6.0	120	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA ≤ I_O ≤ 1.5 A		—	30	120	—	46	240	—	46	240	
250 mA ≤ I_O ≤ 750 mA		—	10	60	—	17	120	—	17	120	
Quiescent Current (T_J = +25°C)	I_B	—	3.4	6.0	—	4.4	8.0	—	4.4	8.0	mA
Quiescent Current Change	ΔI_B										mA
14.5 Vdc ≤ V_{in} ≤ 30 Vdc		—	—	—	—	—	—	—	—	1.0	
15 Vdc ≤ V_{in} ≤ 30 Vdc		—	0.3	0.8	—	—	1.0	—	—	—	
5.0 mA ≤ I_O ≤ 1.0 A		—	0.04	0.5	—	—	0.5	—	—	0.5	
Ripple Rejection 15 Vdc ≤ V_{in} ≤ 25 Vdc, f = 120 Hz	RR	61	68	—	—	60	—	—	60	—	dB
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz ≤ f ≤ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	μV/ V_O
Output Resistance f = 1.0 kHz	R_O	—	18	—	—	18	—	—	18	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±1.5	—	—	-1.0	—	—	-1.0	—	mV/ °C

MC7812A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 19 V, I_O = 1.0 A, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7812A			MC7812AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	11.75	12	12.25	11.75	12	12.25	Vdc
Output Voltage (5.0 mA ≤ I_O ≤ 1.0 A, P_O ≤ 15 W) 14.8 Vdc ≤ V_{in} ≤ 27 Vdc	V_O	11.5	12	12.5	11.5	12	12.5	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
14.8 Vdc ≤ V_{in} ≤ 30 Vdc, I_O = 500 mA		—	5.0	18	—	13	120	
16 Vdc ≤ V_{in} ≤ 22 Vdc		—	8.0	30	—	16	120	
16 Vdc ≤ V_{in} ≤ 22 Vdc, T_J = +25°C		—	3.0	9.0	—	6.0	60	
14.5 Vdc ≤ V_{in} ≤ 27 Vdc, T_J = +25°C		—	5.0	18	—	13	120	
Load Regulation (Note 2)	Reg_{load}							mV
5.0 mA ≤ I_O ≤ 1.5 A		—	30	50	—	—	—	
5.0 mA ≤ I_O ≤ 1.0 A		—	—	—	—	46	100	
5.0 mA ≤ I_O ≤ 1.5 A, T_J = +25°C		—	—	—	—	46	100	
250 mA ≤ I_O ≤ 750 mA		—	10	25	—	17	50	
Quiescent Current	I_B	—	—	5.0	—	—	6.0	mA
T_J = +25°C		—	3.4	4.0	—	4.4	6.0	
Quiescent Current Change	ΔI_B							mA
15 Vdc ≤ V_{in} ≤ 30 Vdc, I_O = 500 mA		—	0.3	0.5	—	—	0.8	
14.8 Vdc ≤ V_{in} ≤ 27 Vdc, T_J = +25°C		—	0.2	0.5	—	—	0.8	
5.0 mA ≤ I_O ≤ 1.0 A		—	0.04	0.2	—	—	0.5	
Ripple Rejection	RR							dB
15 Vdc ≤ V_{in} ≤ 25 Vdc, f = 120 Hz, T_J = +25°C		61	68	—	—	—	—	
15 Vdc ≤ V_{in} ≤ 25 Vdc, f = 120 Hz, I_O = 500 mA		61	68	—	—	60	—	
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz ≤ f ≤ 100 kHz	V_n	—	10	40	—	10	—	μV/V_O
Output Resistance (f = 1.0 kHz)	R_O	—	18	—	—	18	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±1.5	—	—	-1.0	—	mV/°C

NOTES: 1. T_{low} = -55°C for MC78XX, A T_{high} = +150°C for MC78XX, A
= 0° for MC78XXC, AC = +125°C for MC78XXC, AC, B
= -40°C for MC78XXB

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

MC7815, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 23 V, I_O = 500 mA, $T_J = T_{low}$ to T_{high} [Note 1] unless otherwise noted).

Characteristic	Symbol	MC7815			MC7815B			MC7815C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	14.4	15	15.6	14.4	15	15.6	14.4	15	15.6	Vdc
Output Voltage	V_O										Vdc
(5.0 mA $\leqslant I_O \leqslant$ 1.0 A, $P_O \leqslant$ 15 W)											
17.5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc		—	—	—	—	—	—	14.25	15	15.75	
18.5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc		14.25	15	15.75	14.25	15	15.75	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
17 5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc		—	6.0	150	—	13	300	—	13	300	
20 Vdc $\leqslant V_{in} \leqslant$ 26 Vdc		—	3.0	75	—	6.0	150	—	6.0	150	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA $\leqslant I_O \leqslant$ 1.5 A		—	32	150	—	52	300	—	52	300	
250 mA $\leqslant I_O \leqslant$ 750 mA		—	10	75	—	20	150	—	20	150	
Quiescent Current (T_J = +25°C)	I_B	—	3.4	6.0	—	4 4	8.0	—	4.4	8.0	mA
Quiescent Current Change	ΔI_B										mA
17.5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc		—	—	—	—	—	—	—	—	1.0	
18 5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc		—	0.3	0 8	—	—	1 0	—	—	—	
5 0 mA $\leqslant I_O \leqslant$ 1.0 A		—	0.04	0.5	—	—	0.5	—	—	0.5	
Ripple Rejection 18.5 Vdc $\leqslant V_{in} \leqslant$ 28.5 Vdc, f = 120 Hz	RR	60	66	—	—	58	—	—	58	—	dB
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2 0	—	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	µV/ V_O
Output Resistance f = 1.0 kHz	R_O	—	19	—	—	19	—	—	19	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0 2	1 2	—	0 2	—	—	0 2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1 3	2.5	3.3	—	2.2	—	—	2 2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±1 8	—	—	-1 0	—	—	-1 0	—	mV/ °C

MC7815A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 23 V, I_O = 1 0 A, $T_J = T_{low}$ to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7815A			MC7815AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	14.7	15	15 3	14.7	15	15 3	Vdc
Output Voltage (5.0 mA $\leqslant I_O \leqslant$ 1 0 A, $P_O \leqslant$ 15 W) 17 9 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc	V_O	14.4	15	15 6	14 4	15	15 6	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
17 9 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc, I_O = 500 mA		—	6.0	22	—	13	150	
20 Vdc $\leqslant V_{in} \leqslant$ 26 Vdc		—	6 0	22	—	16	150	
20 Vdc $\leqslant V_{in} \leqslant$ 26 Vdc, T_J = +25°C		—	3.0	10	—	6 0	75	
17.5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc, T_J = +25°C		—	6.0	22	—	13	150	
Load Regulation (Note 2)	Reg_{load}							mV
5 0 mA $\leqslant I_O \leqslant$ 1.5 A		—	32	50	—	—	—	
5 0 mA $\leqslant I_O \leqslant$ 1 0 A		—	—	—	—	52	100	
5 0 mA $\leqslant I_O \leqslant$ 1 5 A, T_J = +25°C		—	—	—	—	52	100	
250 mA $\leqslant I_O \leqslant$ 750 mA		—	10	25	—	20	50	
Quiescent Current	I_B	—	—	5 5	—	—	6 0	mA
T_J = +25°C		—	3.4	4 5	—	4 4	6 0	
Quiescent Current Change	ΔI_B							mA
17 5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc, I_O = 500 mA		—	0.3	0.5	—	—	0 8	
17 5 Vdc $\leqslant V_{in} \leqslant$ 30 Vdc, T_J = +25°C		—	0.2	0 5	—	—	0 8	
5.0 mA $\leqslant I_O \leqslant$ 1.0 A		—	0 04	0 2	—	—	0.5	
Ripple Rejection	RR							dB
18.5 Vdc $\leqslant V_{in} \leqslant$ 28.5 Vdc, f = 120 Hz, T_J = +25°C		60	66	—	—	—	—	
18 5 Vdc $\leqslant V_{in} \leqslant$ 28 5 Vdc, f = 120 Hz, I_O = 500 mA		60	66	—	—	58	—	
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2 0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leqslant$ f $\leqslant$ 100 kHz	V_n	—	10	40	—	10	—	µV/V_O
Output Resistance (f = 1.0 kHz)	R_O	—	19	—	—	19	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±1.8	—	—	-1.0	—	mV/°C

NOTES: 1 T_{low} = −55°C for MC78XX, A $\quad T_{high}$ = +150°C for MC78XX, A
= 0° for MC78XXC, AC $\quad$ = +125°C for MC78XXC, AC, B
= −40°C for MC78XXB

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

MC7818, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 27 V, I_O = 500 mA, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted).

Characteristic	Symbol	MC7818			MC7818B			MC7818C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	17.3	18	18.7	17.3	18	18.7	17.3	18	18.7	Vdc
Output Voltage (5.0 mA ≤ I_O ≤ 1.0 A, P_O ≤ 15 W)	V_O										Vdc
21 Vdc ≤ V_{in} ≤ 33 Vdc		—	—	—	—	—	—	17.1	18	18.9	
22 Vdc ≤ V_{in} ≤ 33 Vdc		17.1	18	18.9	17.1	18	18.9	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
21 Vdc ≤ V_{in} ≤ 33 Vdc		—	7.0	180	—	25	360	—	25	360	
24 Vdc ≤ V_{in} ≤ 30 Vdc		—	4.0	90	—	10	180	—	10	180	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA ≤ I_O ≤ 1.5 A		—	35	180	—	55	360	—	55	360	
250 mA ≤ I_O ≤ 750 mA		—	12	90	—	22	180	—	22	180	
Quiescent Current (T_J = +25°C)	I_B	—	3.5	6.0	—	4.5	8.0	—	4.5	8.0	mA
Quiescent Current Change	ΔI_B										mA
21 Vdc ≤ V_{in} ≤ 33 Vdc		—	—	—	—	—	—	—	—	1.0	
22 Vdc ≤ V_{in} ≤ 33 Vdc		—	0.3	0.8	—	—	1.0	—	—	—	
5.0 mA ≤ I_O ≤ 1.0 A		—	0.04	0.5	—	—	0.5	—	—	0.5	
Ripple Rejection 22 Vdc ≤ V_{in} ≤ 32 Vdc, f = 120 Hz	RR	59	65	—	—	57	—	—	57	—	dB
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz ≤ f ≤ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	µV/ V_O
Output Resistance f = 1.0 kHz	R_O	—	19	—	—	19	—	—	19	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±2.3	—	—	-1.0	—	—	-1.0	—	mV/ °C

MC7818A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 27 V, I_O = 1.0 A, T_J = T_{low} to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7818A			MC7818AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	17.64	18	18.36	17.64	18	18.36	Vdc
Output Voltage (5.0 mA ≤ I_O ≤ 1.0 A, P_O ≤ 15 W) 21 Vdc ≤ V_{in} ≤ 33 Vdc	V_O	17.3	18	18.7	17.3	18	17.3	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
21 Vdc ≤ V_{in} ≤ 33 Vdc, I_O = 500 mA		—	7.0	31	—	25	180	
24 Vdc ≤ V_{in} ≤ 30 Vdc		—	12	45	—	28	180	
24 Vdc ≤ V_{in} ≤ 30 Vdc, T_J = +25°C		—	4.0	15	—	10	90	
20.6 Vdc ≤ V_{in} ≤ 33 Vdc, T_J = +25°C		—	7.0	31	—	25	180	
Load Regulation (Note 2)	Reg_{load}							mV
5.0 mA ≤ I_O ≤ 1.5 A		—	35	50	—	—	—	
5.0 mA ≤ I_O ≤ 1.0 A		—	—	—	—	55	100	
5.0 mA ≤ I_O ≤ 1.5 A, T_J = +25°C		—	—	—	—	55	100	
250 mA ≤ I_O ≤ 750 mA		—	12	25	—	22	50	
Quiescent Current	I_B	—	—	5.5	—	—	6.0	mA
T_J = +25°C		—	3.4	4.5	—	4.5	6.0	
Quiescent Current Change	ΔI_B							mA
21 Vdc ≤ V_{in} ≤ 33 Vdc, I_O = 500 mA		—	0.3	0.5	—	—	0.8	
21 Vdc ≤ V_{in} ≤ 33 Vdc, T_J = +25°C		—	0.2	0.5	—	—	0.8	
5.0 mA ≤ I_O ≤ 1.0 A		—	0.04	0.2	—	—	0.5	
Ripple Rejection	RR							dB
22 Vdc ≤ V_{in} ≤ 32 Vdc, f = 120 Hz, T_J = +25°C		59	65	—	—	—	—	
22 Vdc ≤ V_{in} ≤ 32 Vdc, f = 120 Hz, I_O = 500 mA		59	65	—	—	57	—	
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz ≤ f ≤ 100 kHz	V_n	—	10	40	—	10	—	µV/V_O
Output Resistance (f = 1.0 kHz)	R_O	—	19	—	—	19	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3.3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±2.3	—	—	-1.0	—	mV/°C

NOTES 1. T_{low} = −55°C for MC78XX, A
= 0° for MC78XXC, AC
= −40°C for MC78XXB
T_{high} = +150°C for MC78XX, A
= +125°C for MC78XXC, AC, B

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

MC7824, B, C

ELECTRICAL CHARACTERISTICS (V_{in} = 33 V, I_O = 500 mA, $T_J = T_{low}$ to T_{high} [Note 1] unless otherwise noted).

Characteristic	Symbol	MC7824			MC7824B			MC7824C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	23	24	25	23	24	25	23	24	25	Vdc
Output Voltage (5 0 mA $\leq I_O \leq$ 1.0 A, $P_O \leq$ 15 W)	V_O										Vdc
27 Vdc $\leq V_{in} \leq$ 38 Vdc		—	—	—	—	—	—	22.8	24	25 2	
28 Vdc $\leq V_{in} \leq$ 38 Vdc		22.8	24	25.2	22.8	24	25.2	—	—	—	
Line Regulation (T_J = +25°C, Note 2)	Reg_{in}										mV
27 Vdc $\leq V_{in} \leq$ 38 Vdc		—	10	240	—	31	480	—	31	480	
30 Vdc $\leq V_{in} \leq$ 36 Vdc		—	5 0	120	—	14	240	—	14	240	
Load Regulation (T_J = +25°C, Note 2)	Reg_{load}										mV
5.0 mA $\leq I_O \leq$ 1.5 A		—	40	240	—	60	480	—	60	480	
250 mA $\leq I_O \leq$ 750 mA		—	15	120	—	25	240	—	25	240	
Quiescent Current (T_J = +25°C)	I_B	—	3.6	6.0	—	4.6	8.0	—	4.6	8 0	mA
Quiescent Current Change	ΔI_B										mA
27 Vdc $\leq V_{in} \leq$ 38 Vdc		—	—	—	—	—	—	—	—	1 0	
28 Vdc $\leq V_{in} \leq$ 38 Vdc		—	0.3	0 8	—	—	1 0	—	—	—	
5.0 mA $\leq I_O \leq$ 1.0 A		—	0.04	0.5	—	—	0 5	—	—	0 5	
Ripple Rejection 28 Vdc $\leq V_{in} \leq$ 38 Vdc, f = 120 Hz	RR	56	62	—	—	54	—	—	54	—	dB
Dropout Voltage (I_O = 1 0 A, T_J = +25°C)	$V_{in} - V_O$	—	2 0	2.5	—	2.0	—	—	2 0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leq f \leq$ 100 kHz	V_n	—	10	40	—	10	—	—	10	—	$\mu V/V_O$
Output Resistance f = 1.0 kHz	R_O	—	20	—	—	20	—	—	20	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0 2	1 2	—	0.2	—	—	0 2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1 3	2 5	3 3	—	2.2	—	—	2 2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±3 0	—	—	-1 5	—	—	-1 5	—	mV/°C

MC7824A, AC

ELECTRICAL CHARACTERISTICS (V_{in} = 33 V, I_O = 1 0 A, $T_J = T_{low}$ to T_{high} [Note 1] unless otherwise noted)

Characteristics	Symbol	MC7824A			MC7824AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	23.5	24	24 5	23.5	24	24 5	Vdc
Output Voltage (5 0 mA $\leq I_O \leq$ 1 0 A, $P_O \leq$ 15 W) 27 3 Vdc $\leq V_{in} \leq$ 38 Vdc	V_O	23	24	25	23	24	25	Vdc
Line Regulation (Note 2)	Reg_{in}							mV
27 Vdc $\leq V_{in} \leq$ 38 Vdc, I_O = 500 mA		—	10	36	—	31	240	
30 Vdc $\leq V_{in} \leq$ 36 Vdc		—	15	60	—	35	240	
30 Vdc $\leq V_{in} \leq$ 36 Vdc, T_J = +25°C		—	5 0	19	—	14	120	
26 7 Vdc $\leq V_{in} \leq$ 38 Vdc, T_J = +25°C		—	10	36	—	31	240	
Load Regulation (Note 2)	Reg_{load}							mV
5 0 mA $\leq I_O \leq$ 1.5 A		—	40	50	—	—	—	
5.0 mA $\leq I_O \leq$ 1 0 A		—	—	—	—	60	100	
5 0 mA $\leq I_O \leq$ 1 5 A, T_J = +25°C		—	—	—	—	60	100	
250 mA $\leq I_O \leq$ 750 mA		—	15	25	—	25	50	
Quiescent Current	I_B	—	—	6 0	—	—	6 0	mA
T_J = +25°C		—	3 6	5.0	—	4 6	6.0	
Quiescent Current Change	ΔI_B							mA
27.3 Vdc $\leq V_{in} \leq$ 38 Vdc, I_O = 500 mA		—	0 3	0.5	—	—	0 8	
27.3 Vdc $\leq V_{in} \leq$ 38 Vdc, T_J = +25°C		—	0.2	0 5	—	—	0.8	
5.0 mA $\leq I_O \leq$ 1.0 A		—	0 04	0.2	—	—	0 5	
Ripple Rejection	RR							dB
28 Vdc $\leq V_{in} \leq$ 38 Vdc, f = 120 Hz, T_J = +25°C		56	62	—	—	—	—	
28 Vdc $\leq V_{in} \leq$ 38 Vdc, f = 120 Hz, I_O = 500 mA		56	62	—	—	54	—	
Dropout Voltage (I_O = 1.0 A, T_J = +25°C)	$V_{in} - V_O$	—	2.0	2.5	—	2.0	—	Vdc
Output Noise Voltage (T_A = +25°C) 10 Hz $\leq f \leq$ 100 kHz	V_n	—	10	40	—	10	—	$\mu V/V_O$
Output Resistance (f = 1.0 kHz)	R_O	—	20	—	—	20	—	mΩ
Short-Circuit Current Limit (T_A = +25°C) V_{in} = 35 Vdc	I_{sc}	—	0.2	1.2	—	0.2	—	A
Peak Output Current (T_J = +25°C)	I_{max}	1.3	2.5	3 3	—	2.2	—	A
Average Temperature Coefficient of Output Voltage	TCV_O	—	±3.0	—	—	-1.5	—	mV/°C

NOTES: 1. T_{low} = -55°C for MC78XX, A T_{high} = +150°C for MC78XX, A
= 0° for MC78XXC, AC = +125°C for MC78XXC, AC, B
= -40°C for MC78XXB

2. Load and line regulation are specified at constant junction temperature. Changes in V_O due to heating effects must be taken into account separately. Pulse testing with low duty cycle is used.

TYPICAL CHARACTERISTICS

(T_A = +25°C unless otherwise noted.)

FIGURE 1 — WORST CASE POWER DISSIPATION versus AMBIENT TEMPERATURE (Case 221A)

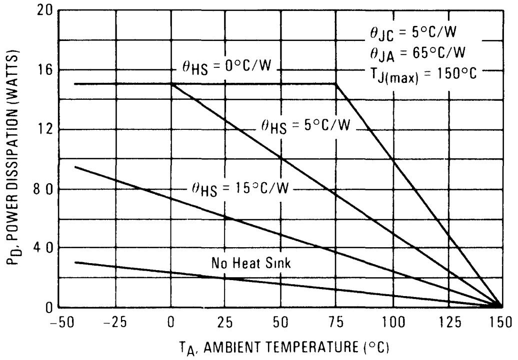


FIGURE 2 – WORST CASE POWER DISSIPATION versus AMBIENT TEMPERATURE (Case 1)

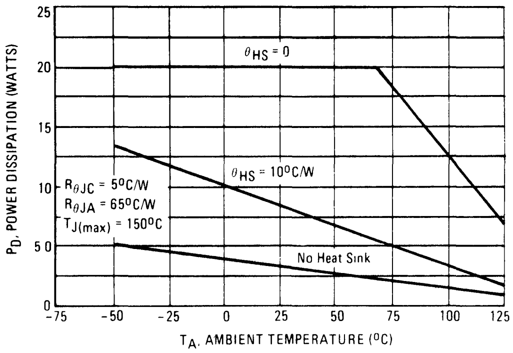


FIGURE 3 — INPUT OUTPUT DIFFERENTIAL AS A FUNCTION OF JUNCTION TEMPERATURE (MC78XXC, AC, B)

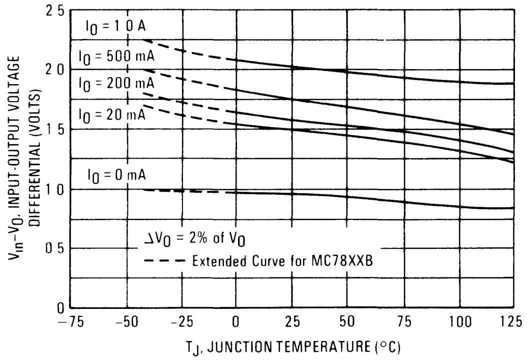


FIGURE 4 – INPUT OUTPUT DIFFERENTIAL AS A FUNCTION OF JUNCTION TEMPERATURE (MC78XX, A)

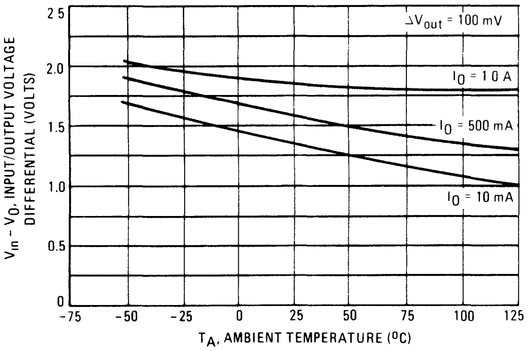


FIGURE 5 — PEAK OUTPUT CURRENT AS A FUNCTION OF INPUT-OUTPUT DIFFERENTIAL VOLTAGE (MC78XXC, AC, B)

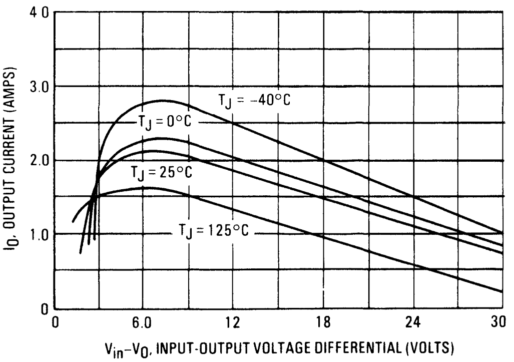
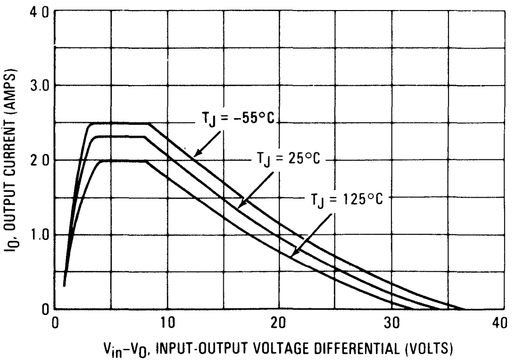


FIGURE 6 — PEAK OUTPUT CURRENT AS A FUNCTION OF INPUT-OUTPUT DIFFERENTIAL VOLTAGE (MC78XX, A)



TYPICAL CHARACTERISTICS (continued)
(T_A = 25°C unless otherwise noted.)

FIGURE 7 – RIPPLE REJECTION AS A FUNCTION OF OUTPUT VOLTAGES (MC78XXC, AC)

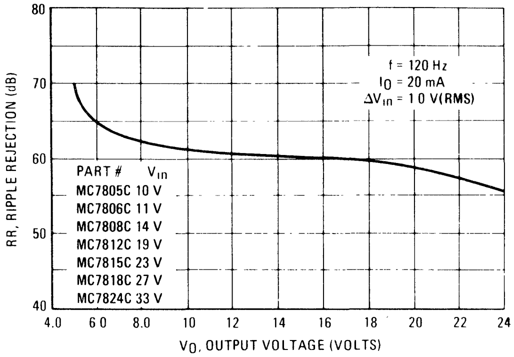


FIGURE 8 – RIPPLE REJECTION AS A FUNCTION OF FREQUENCY (MC78XXC, AC)

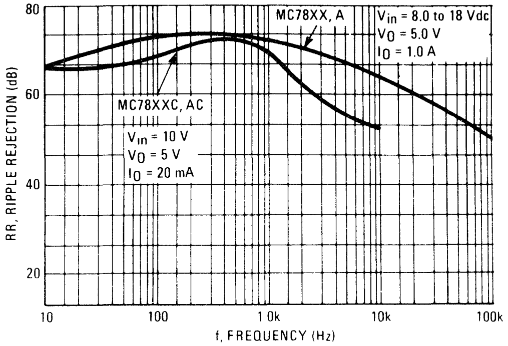


FIGURE 9 — OUTPUT VOLTAGE AS A FUNCTION OF JUNCTION TEMPERATURE (MC78XXC, AC, B)

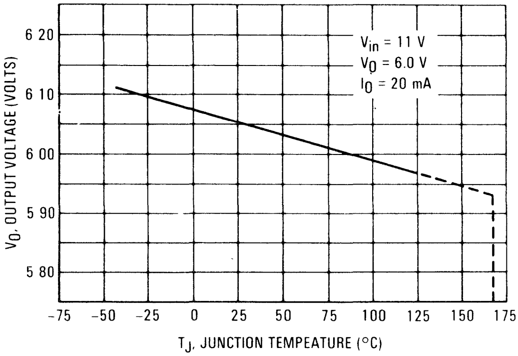


FIGURE 10 – OUTPUT IMPEDANCE AS A FUNCTION OF OUTPUT VOLTAGE (MC78XXC, AC)

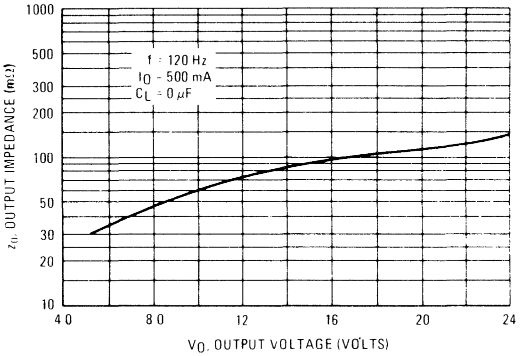


FIGURE 11 — QUIESCENT CURRENT AS A FUNCTION OF TEMPERATURE (MC78XXC, AC, B)

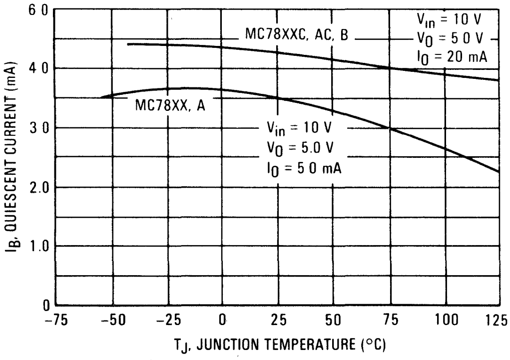
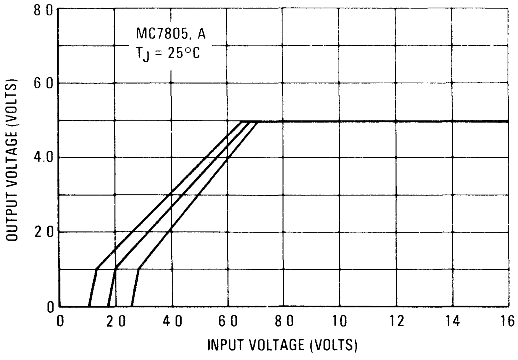


FIGURE 12 — DROPOUT CHARACTERISTICS (MC78XX, A)



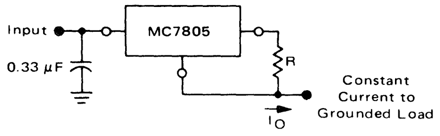
APPLICATIONS INFORMATION

Design Considerations

The MC7800 Series of fixed voltage regulators are designed with Thermal Overload Protection that shuts down the circuit when subjected to an excessive power overload condition, Internal Short-Circuit Protection that limits the maximum current the circuit will pass, and Output Transistor Safe-Area Compensation that reduces the output short-circuit current as the voltage across the pass transistor is increased.

In many low current applications, compensation capacitors are not required. However, it is recommended that the regulator input be bypassed with a capacitor if the regulator is connected to the power supply filter with long wire lengths, or if the output load capacitance is large. An input bypass capacitor should be selected to provide good high-frequency characteristics to insure stable operation under all load conditions. A 0.33 µF or larger tantalum, mylar, or other capacitor having low internal impedance at high frequencies should be chosen. The bypass capacitor should be mounted with the shortest possible leads directly across the regulators input terminals. Normally good construction techniques should be used to minimize ground loops and lead resistance drops since the regulator has no external sense lead.

FIGURE 13 – CURRENT REGULATOR



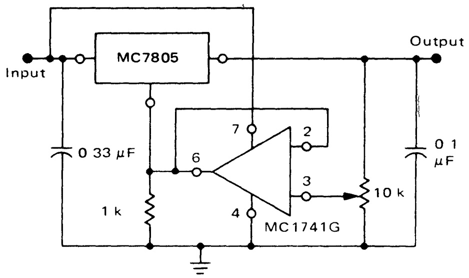
The MC7800 regulators can also be used as a current source when connected as above. In order to minimize dissipation the MC7805C is chosen in this application. Resistor R determines the current as follows:

$$I_O = \frac{5\text{ V}}{R} + I_Q$$

$I_Q \cong 1.5$ mA over line and load changes

For example, a 1-ampere current source would require R to be a 5-ohm, 10-W resistor and the output voltage compliance would be the input voltage less 7 volts

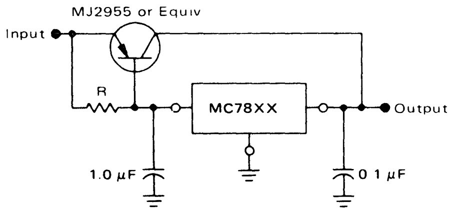
FIGURE 14 – ADJUSTABLE OUTPUT REGULATOR



V_O, 7.0 V to 20 V
$V_{IN} - V_O \geqslant 2.0$ V

The addition of an operational amplifier allows adjustment to higher or intermediate values while retaining regulation characteristics. The minimum voltage obtainable with this arrangement is 2.0 volts greater than the regulator voltage.

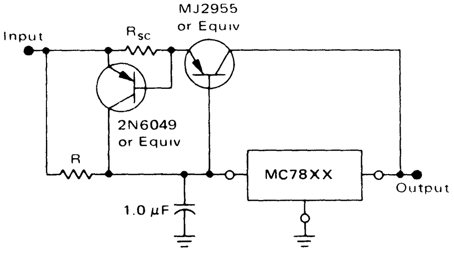
FIGURE 15 – CURRENT BOOST REGULATOR



XX = 2 digits of type number indicating voltage

The MC7800 series can be current boosted with a PNP transistor. The MJ2955 provides current to 5.0 amperes. Resistor R in conjunction with the V_{BE} of the PNP determines when the pass transistor begins conducting; this circuit is not short-circuit proof. Input-output differential voltage minimum is increased by V_{BE} of the pass transistor.

FIGURE 16 – SHORT-CIRCUIT PROTECTION



XX – 2 digits of type number indicating voltage

The circuit of Figure 15 can be modified to provide supply protection against short circuits by adding a short-circuit sense resistor, R_{SC}, and an additional PNP transistor. The current sensing PNP must be able to handle the short-circuit current of the three-terminal regulator. Therefore, a four-ampere plastic power transistor is specified.

MOTOROLA

MC78L00C,AC Series

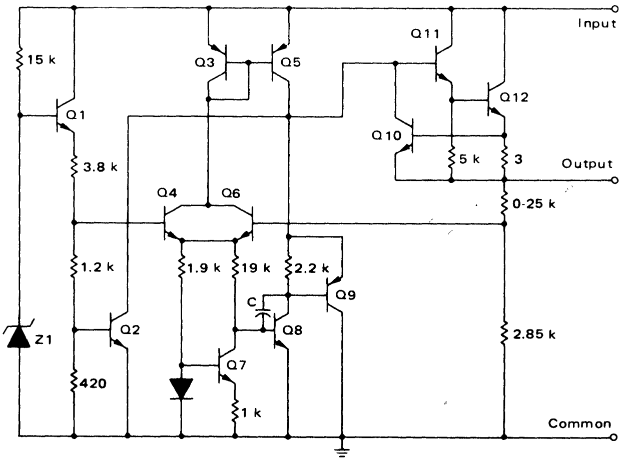
THREE-TERMINAL POSITIVE VOLTAGE REGULATORS

The MC78L00 Series of positive voltage regulators are inexpensive, easy-to-use devices suitable for a multitude of applications that require a regulated supply of up to 100 mA. Like their higher powered MC7800 and MC78M00 Series cousins, these regulators feature internal current limiting and thermal shutdown making them remarkably rugged. No external components are required with the MC78L00 devices in many applications.

These devices offer a substantial performance advantage over the traditional zener diode-resistor combination. Output impedance is greatly reduced and quiescent current is substantially reduced.

- Wide Range of Available, Fixed Output Voltages
- Low Cost
- Internal Short-Circuit Current Limiting
- Internal Thermal Overload Protection
- No External Components Required
- Complementary Negative Regulators Offered (MC79L00 Series)
- Available in Either ±5% (AC) or ±10% (C) Selections

REPRESENTATIVE CIRCUIT SCHEMATIC

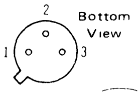


Device No. ±10%	Device No. ±5%	Nominal Voltage
MC78L05C	MC78L05AC	5.0
MC78L08C	MC78L08AC	8.0
MC78L12C	MC78L12AC	12
MC78L15C	MC78L15AC	15
MC78L18C	MC78L18AC	18
MC78L24C	MC78L24AC	24

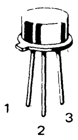
THREE-TERMINAL POSITIVE FIXED VOLTAGE REGULATORS

P SUFFIX
CASE 29
TO-92

Pin 1. Output
2. Ground
3. Input



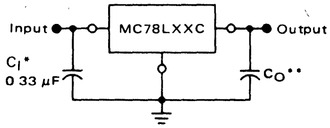
G SUFFIX
CASE 79
TO-39



Pin 1. Input
2. Output
3. Ground

(Case connected to pin 3)

STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.0 V above the output voltage even during the low point on the input ripple voltage.

* = C_I is required if regulator is located an appreciable distance from power supply filter.

** = C_O is not needed for stability; however, it does improve transient response.

ORDERING INFORMATION

Device	Temperature Range	Package
MC78LXXACG	T_J = 0°C to +150°C	Metal Can
MC78LXXACP	T_J = 0°C to +150°C	Plastic Transistor
MC78LXXCG	T_J = 0°C to +150°C	Metal Can
MC78LXXCP	T_J = 0°C to +150°C	Plastic Transistor

XX indicates nominal voltage

MC78L00 Series MAXIMUM RATINGS (T_A = +125°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Input Voltage (2.6 V – 8.0 V)	V_I	30	Vdc
(12 V – 18 V)		35	
(24 V)		40	
Storage Junction Temperature Range	T_{stg}	-65 to +150	°C
Operating Junction Temperature Range	T_J	0 to +150	°C

MC78L05C, MC78L05AC ELECTRICAL CHARACTERISTICS (V_I = 10 V, I_O = 40 mA, C_I = 0.33 µF, C_O = 0.1 µF, 0°C < T_J < +125°C unless otherwise noted.)

		MC78L05C			MC78L05AC			
Characteristic	Symbol	Min	Typ	Max	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	4.6	5.0	5.4	4.8	5.0	5.2	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C, I_O = 40 mA)								
7.0 Vdc ≤ V_I ≤ 20 Vdc		–	55	200	–	55	150	
8.0 Vdc ≤ V_I ≤ 20 Vdc		–	45	150	–	45	100	
Load Regulation	Reg_{load}							mV
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA)		–	11	60	–	11	60	
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 40 mA)		–	5.0	30	–	5.0	30	
Output Voltage	V_O							Vdc
(7.0 Vdc ≤ V_I ≤ 20 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)		4.5	–	5.5	4.75	–	5.25	
(V_I = 10 V, 1.0 mA ≤ I_O ≤ 70 mA)		4.5	–	5.5	4.75	–	5.25	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	3.8	6.0	–	3.8	6.0	
(T_J = +125°C)		–	–	5.5	–	–	5.5	
Input Bias Current Change	ΔI_{IB}							mA
(8.0 Vdc ≤ V_I ≤ 20 Vdc)		–	–	1.5	–	–	1.5	
(1.0 mA ≤ I_O ≤ 40 mA)		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	40	–	–	40	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	12	–	–	12	–	mV/1.0 k Hrs
Ripple Rejection (I_O = 40 mA, f = 120 Hz, 8.0 V ≤ V_I ≤ 18 V, T_J = +25°C)	RR	40	49	–	41	49	–	dB
Input-Output Voltage Differential (T_J = +25°C)	V_I/V_O	–	1.7	–	–	1.7	–	Vdc

MC78L08C, MC78L08AC ELECTRICAL CHARACTERISTICS (V_I = 14 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

		MC78L08C			MC78L08AC			
Characteristic	Symbol	Min	Typ	Max	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	7.36	8.0	8.64	7.7	8.0	8.3	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C, I_O = 40 mA)								
10.5 Vdc ≤ V_I ≤ 23 Vdc		–	20	200	–	20	175	
11 Vdc ≤ V_I ≤ 23 Vdc		–	12	150	–	12	125	
Load Regulation	Reg_{load}							mV
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA)		–	15	80	–	15	80	
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 40 mA)		–	6.0	40	–	8.0	40	
Output Voltage	V_O							Vdc
(10.5 Vdc ≤ V_I ≤ 23 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)		7.2	–	8.8	7.6	–	8.4	
(V_I = 14 V, 1.0 mA ≤ I_O ≤ 70 mA)		7.2	–	8.8	7.6	–	8.4	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	3.0	6.0	–	3.0	6.0	
(T_J = +125°C)		–	–	5.5	–	–	5.5	
Input Bias Current Change	ΔI_{IB}							mA
(11 Vdc ≤ V_I ≤ 23 Vdc)		–	–	1.5	–	–	1.5	
(1.0 mA ≤ I_O ≤ 40 mA)		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	52	–	–	60	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	20	–	–	20	–	mV/1.0 k Hrs.
Ripple Rejection (I_O = 40 mA, f = 120 Hz, 12 V ≤ V_I ≤ 23 V, T_J = +25°C)	RR	36	55	–	37	57	–	dB
Input-Output Voltage Differential (T_J = +25°C)	V_I/V_O	–	1.7	–	–	1.7	–	Vdc

MC78L12C, MC78L12AC ELECTRICAL CHARACTERISTICS (V_I = 19 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

		MC78L12C			MC78L12AC			
Characteristic	Symbol	Min	Typ	Max	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	11.1	12	12.9	11.5	12	12.5	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C, I_O = 40 mA)								
14.5 Vdc ≤ V_I ≤ 27 Vdc		–	120	250	–	120	250	
16 Vdc ≤ V_I ≤ 27 Vdc		–	100	200	–	100	200	
Load Regulation	Reg_{load}							mV
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA)		–	20	100	–	20	100	
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 40 mA)		–	10	50	–	10	50	
Output Voltage	V_O							Vdc
(14.5 Vdc ≤ V_I ≤ 27 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)		10.8	–	13.2	11.4	–	12.6	
(V_I = 19 V, 1.0 mA ≤ I_O ≤ 70 mA)		10.8	–	13.2	11.4	–	12.6	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	4.2	6.5	–	4.2	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
(16 Vdc ≤ V_I ≤ 27 Vdc)		–	–	1.5	–	–	1.5	
(1.0 mA ≤ I_O ≤ 40 mA)		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	80	–	–	80	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	24	–	–	24	–	mV/1.0 k Hrs.
Ripple Rejection (I_O = 40 mA, f = 120 Hz, 15 V ≤ V_I ≤ 25 V, T_J = +25°C)	RR	36	42	–	37	42	–	dB
Input-Output Voltage Differential (T_J = +25°C)	V_I/V_O	–	1.7	–	–	1.7	–	Vdc

MC78L15C, MC78L15AC ELECTRICAL CHARACTERISTICS (V_I = 23 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC78L15C			MC78L15AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	13.8	15	16.2	14.4	15	15.6	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C, I_O = 40 mA)								
17.5 Vdc ≤ V_I ≤ 30 Vdc		–	130	300	–	130	300	
20 Vdc ≤ V_I ≤ 30 Vdc		–	110	250	–	110	250	
Load Regulation	Reg_{load}							mV
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA)		–	25	150	–	25	150	
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 40 mA)		–	12	75	–	12	75	
Output Voltage	V_O							Vdc
(17.5 Vdc ≤ V_I ≤ 30 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)		13.5	–	16.5	14.25	–	15.75	
(V_I = 23 V, 1.0 mA ≤ I_O ≤ 70 mA)		13.5	–	16.5	14.25	–	15.75	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	4.4	6.5	–	4.4	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
(20 Vdc ≤ V_I ≤ 30 Vdc)		–	–	1.5	–	–	1.5	
(1.0 mA ≤ I_O ≤ 40 mA)		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	90	–	–	90	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	30	–	–	30	–	mV/1.0 k Hrs.
Ripple Rejection (I_O = 40 mA, f = 120 Hz, 18.5 V ≤ V_I ≤ 28.5 V, T_J = +25°C)	RR	33	39	–	34	39	–	dB
Input-Output Voltage Differential (T_J = +25°C)	V_I/V_O	–	1.7	–	–	1.7	–	Vdc

MC78L18C, MC78L18AC ELECTRICAL CHARACTERISTICS (V_I = 27 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC78L18C			MC78L18AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	16.6	18	19.4	17.3	18	18.7	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C, I_O = 40 mA)								
21.4 Vdc ≤ V_I ≤ 33 Vdc		–	32	325				
20.7 Vdc ≤ V_I ≤ 33 Vdc					–	45	325	
22 Vdc ≤ V_I ≤ 33 Vdc		–	27	275				
21 Vdc ≤ V_I ≤ 33 Vdc					–	35	275	
Load Regulation	Reg_{load}							mV
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA)		–	30	170	–	30	170	
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 40 mA)		–	15	85	–	15	85	
Output Voltage	V_O							Vdc
(21.4 Vdc ≤ V_I ≤ 33 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)		16.2	–	17.8				
(20.7 Vdc ≤ V_I ≤ 33 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)					17.1	–	18.9	
(V_I = 27 V, 1.0 mA ≤ I_O ≤ 70 mA)		16.2	–	17.8				
(V_I = 27 V, 1.0 mA ≤ I_O ≤ 70 mA)					17.1	–	18.9	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	3.1	6.5	–	3.1	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
(22 Vdc ≤ V_I ≤ 33 Vdc)		–	–	1.5				
(21 Vdc ≤ V_I = 33 Vdc)					–	–	1.5	
(1.0 mA ≤ I_O ≤ 40 mA)		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	150	–	–	150	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	45	–	–	45	–	mV/1.0 k Hrs.
Ripple Rejection (I_O = 40 mA, f = 120 Hz, 23 V ≤ V_I ≤ 33 V, T_J = +25°C)	RR	32	46	–	33	48	–	dB
Input-Output Voltage Differential (T_J = +25°C)	V_I/V_O	–	1.7	–	–	1.7	–	Vdc

MC78L24C, MC78L24AC ELECTRICAL CHARACTERISTICS (V_I = 33 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC78L24C			MC78L24AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	22.1	24	25.9	23	24	25	Vdc
Input Regulation (T_J = +25°C, I_O = 40 mA)	Reg_{line}							mV
27.5 Vdc ≤ V_I ≤ 38 Vdc		–	35	350	–	–	–	
28 Vdc ≤ V_I ≤ 38 Vdc		–	30	300	–	50	300	
27 Vdc ≤ V_I ≤ 38 Vdc		–	–	–	–	60	350	
Load Regulation	Reg_{load}							mV
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA)		–	40	200	–	40	200	
(T_J = +25°C, 1.0 mA ≤ I_O ≤ 40 mA)		–	20	100	–	20	100	
Output Voltage	V_O							Vdc
(28 Vdc ≤ V_I ≤ 38 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)		21.6	–	26.4				
(27 Vdc ≤ V_I ≤ 38 Vdc, 1.0 mA ≤ I_O ≤ 40 mA)					22.8	–	25.2	
(28 Vdc ≤ V_I ≤ 33 V, 1.0 mA ≤ I_O ≤ 70 mA)		21.6	–	26.4				
(27 Vdc ≤ V_I ≤ 33 V, 1.0 mA ≤ I_O ≤ 70 mA)					22.8	–	25.2	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	3.1	6.5	–	3.1	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
(28 Vdc ≤ V_I ≤ 38 Vdc)		–	–	1.5	–	–	1.5	
(1.0 mA ≤ I_O ≤ 40 mA)		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	200	–	–	200	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	56	–	–	56	–	mV/1.0 k Hrs.
Ripple Rejection (I_O = 40 mA, f = 120 Hz, 29 V ≤ V_I ≤ 35 V, T_J = +25°C)	RR	30	43	–	31	45	–	dB
Input-Output Voltage Differential (T_J = +25°C)	V_I/V_O	–	1.7	–	–	1.7	–	Vdc

TYPICAL CHARACTERISTICS

(T_A = +25°C unless otherwise noted.)

FIGURE 1 — DROPOUT CHARACTERISTICS

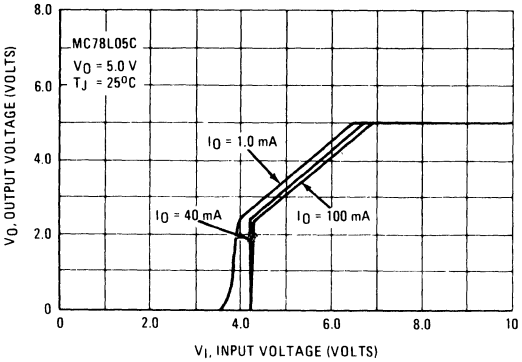


FIGURE 2 — DROPOUT VOLTAGE versus JUNCTION TEMPERATURE

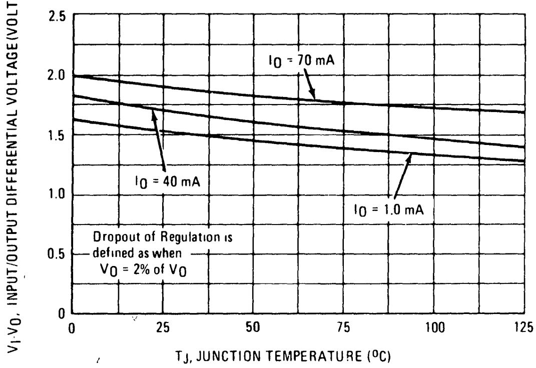


FIGURE 3 — INPUT BIAS CURRENT versus AMBIENT TEMPERATURE

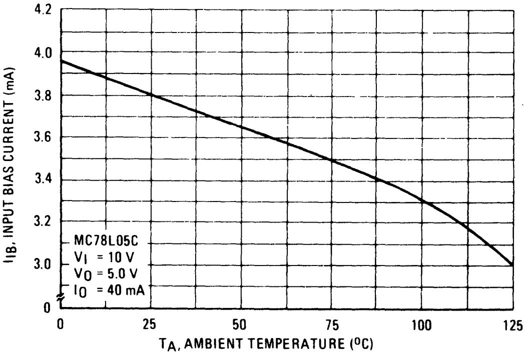


FIGURE 4 — INPUT BIAS CURRENT versus INPUT VOLTAGE

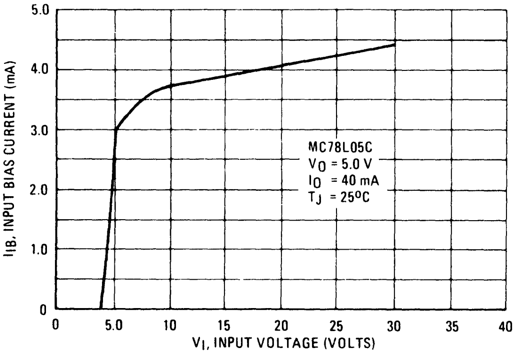


FIGURE 5 — MAXIMUM AVERAGE POWER DISSIPATION versus AMBIENT TEMPERATURE — TO-92 Type Package

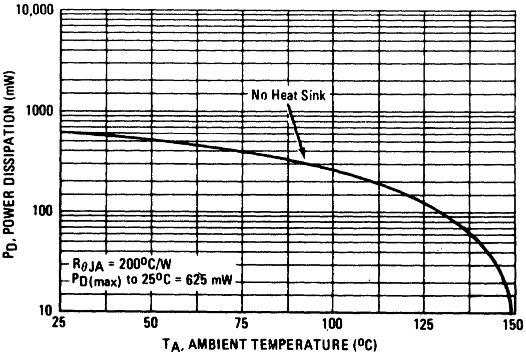
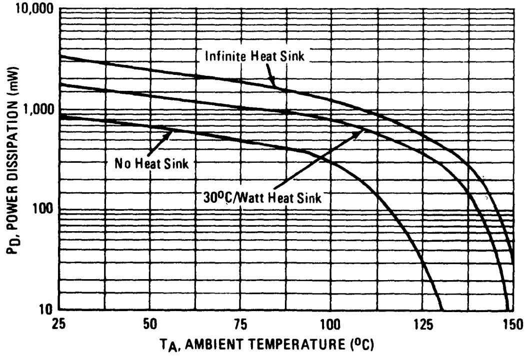


FIGURE 6 — MAXIMUM AVERAGE POWER DISSIPATION versus AMBIENT TEMPERATURE — TO-39 Type Package



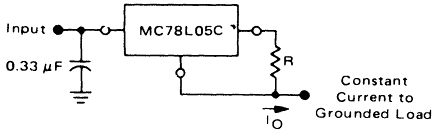
APPLICATIONS INFORMATION

Design Considerations

The MC78L00C Series of fixed voltage regulators are designed with Thermal Overload Protection that shuts down the circuit when subjected to an excessive power overload condition, Internal Short-Circuit Protection that limits the maximum current the circuit will pass.

In many low current applications, compensation capacitors are not required. However, it is recommended that the regulator input be bypassed with a capacitor if the regulator is connected to the power supply filter with long wire lengths, or if the output load capacitance is large. An input bypass capacitor should be selected to provide good high-frequency characteristics to insure stable operation under all load conditions. A 0.33 μF or larger tantalum, mylar, or other capacitor having low internal impedance at high frequencies should be chosen. The bypass capacitor should be mounted with the shortest possible leads directly across the regulators input terminals. Normally good construction techniques should be used to minimize ground loops and lead resistance drops since the regulator has no external sense lead. Bypassing the output is also recommended.

FIGURE 7 – CURRENT REGULATOR



The MC78L00C regulators can also be used as a current source when connected as above. In order to minimize dissipation the MC78L05C is chosen in this application. Resistor R determines the current as follows

$$I_O = \frac{5\ V}{R} + I_{IB}$$

I_{IB} = 3.8 mA over line and load changes

For example, a 100 mA current source would require R to be a 50-ohm, 1/2-W resistor and the output voltage compliance would be the input voltage less 7 volts.

FIGURE 8 – ±15 V TRACKING VOLTAGE REGULATOR

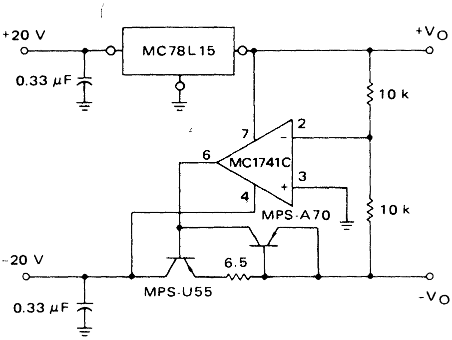
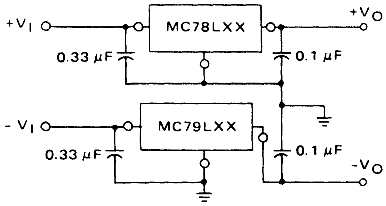


FIGURE 9 – POSITIVE AND NEGATIVE REGULATOR



MOTOROLA

MC78M00C
series

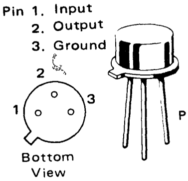
MC78M00C SERIES THREE-TERMINAL POSITIVE VOLTAGE REGULATORS

The MC78M00 Series positive voltage regulators are identical to the popular MC7800C Series devices, except that they are specified for only one-third the output current. Like the MC7800C devices, the MC78M00C three-terminal regulators are intended for local, on-card voltage regulation.

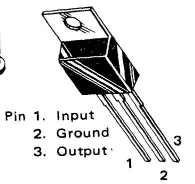
Internal current limiting, thermal shutdown circuitry and safe-area compensation for the internal pass transistor combine to make these devices remarkably rugged under most operating conditions. Maximum output current, with adequate heatsinking is 500 mA.

- No External Components Required
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Packaged in the Plastic Case 221A and Case 79 (TO-220 and Hermetic TO-39)

THREE-TERMINAL POSITIVE FIXED VOLTAGE REGULATORS

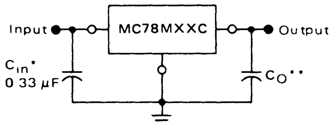


G SUFFIX
METAL PACKAGE
CASE 79
TO-39
(Case connected to Pin 3)



T SUFFIX
PLASTIC PACKAGE
CASE 221A
(TO-220 Type)
(Heatsink surface connected to Pin 2)

STANDARD APPLICATION

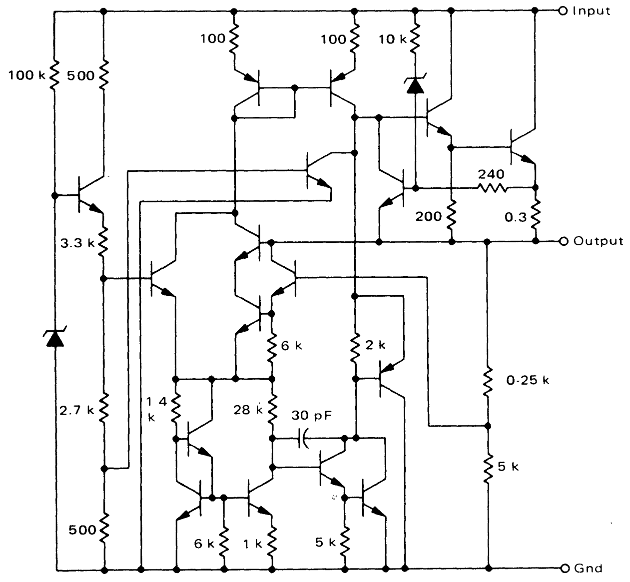


A common ground is required between the input and the output voltages. The input voltage must remain typically 2 0 V above the output voltage even during the low point on the input ripple voltage.

* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_O improves stability and transient response.

REPRESENTATIVE SCHEMATIC DIAGRAM



ORDERING INFORMATION		
DEVICE	TEMPERATURE RANGE	PACKAGE
MC78MXXCG	T_J = 0° C to +150° C	Metal Can
MC78MXXCT	T_J = 0° C to +150° C	Plastic Power
XX indicates nominal voltage		

TYPE NO./VOLTAGE

MC78M05C	5.0 Volts
MC78M06C	6.0 Volts
MC78M08C	8.0 Volts
MC78M12C	12 Volts
MC78M15C	15 Volts
MC78M18C	18 Volts
MC78M20C	20 Volts
MC78M24C	24 Volts

MC78M00C Series MAXIMUM RATINGS (T_A = +25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Input Voltage (5.0 V - 18 V) (20 V - 24 V)	V_I	35 40	Vdc
Power Dissipation (Package Limitation)			
Plastic Package			
T_A = 25°C	P_D	Internally Limited	
Derate above T_A = 25°C	θ_{JA}	70	°C/W
T_C = 25°C	P_D	Internally Limited	
Derate above T_C = 110°C	θ_{JC}	5.0	°C/W
Metal Package			
T_A = 25°C	P_D	Internally Limited	
Derate above T_A = 25°C	θ_{JA}	185	°C/W
T_C = 25°C	P_D	Internally Limited	
Derate above T_C = 85°C	θ_{JC}	25	°C/W
Operating Junction Temperature Range	T_J	0 to +150	°C
Operating Ambient Temperature Range	T_A	0 to +85	°C
Storage Temperature Range	T_{stg}		
Plastic Package		-65 to +150	°C
Metal Package		-65 to +150	°C

MC78M05C ELECTRICAL CHARACTERISTICS (V_I = 10 V, I_O = 200 mA, 0°C < T_J < +125°C, $P_D \leqslant$ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	4.8	5.0	5.2	Vdc
Line Regulation (T_J = +25°C)	Reg_{line}				mV
(7.0 Vdc $\leqslant V_I \leqslant$ 25 Vdc)		–	3.0	100	
(8.0 Vdc $\leqslant V_I \leqslant$ 25 Vdc)		–	1.0	50	
Load Regulation	Reg_{load}				mV
(T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 500 mA)		–	20	100	
(T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)		–	10	50	
Output Voltage (7.0 Vdc $\leqslant V_I \leqslant$ 25 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	V_O	4.75	–	5.25	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.5	6.0	mA
Quiescent Current Change	ΔI_{IB}				mA
(8.0 Vdc $\leqslant V_I \leqslant$ 25 Vdc)		–	–	0.8	
(5.0 mA $\leqslant I_O \leqslant$ 200 mA)		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz $\leqslant$ f $\leqslant$ 100 kHz)	e_{on}	–	40	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	20	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 8.0 V $\leqslant V_I \leqslant$ 18 V)	RR	–	80	–	dB
(I_O = 300 mA, f = 120 Hz, 8.0 $\leqslant V_I \leqslant$ 18 V, T_J = 25°C)		–	80	–	
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	300	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA)	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

MC78M06C ELECTRICAL CHARACTERISTICS

(V_I = 11 V, I_O = 200 mA, 0°C < T_J < +125°C, $P_D \leq$ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	5.75	6.0	6.25	Vdc
Line Regulation (T_J = +25°C) (8.0 Vdc ≤ V_I ≤ 25 Vdc) (9.0 Vdc ≤ V_I ≤ 25 Vdc)	Reg_{line}	 – –	 5.0 1.5	 100 50	mV
Load Regulation (T_J = +25°C, 5.0 mA ≤ I_O ≤ 500 mA) (T_J = +25°C, 5.0 mA ≤ I_O ≤ 200 mA)	Reg_{load}	 – –	 20 10	 120 60	mV
Output Voltage (8.0 Vdc ≤ V_I ≤ 25 Vdc, 5.0 mA ≤ I_O ≤ 200 mA)	V_O	5.7	–	6.3	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.5	6.0	mA
Quiescent Current Change (9.0 Vdc ≤ V_I ≤ 25 Vdc) (5.0 mA ≤ I_O ≤ 200 mA)	ΔI_{IB}	 – –	 – –	 0.8 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	45	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	24	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 9.0 V ≤ V_I ≤ 19 V) (I_O = 300 mA, f = 120 Hz, 9.0 V ≤ V_I ≤ 19 V, T_J = 25°C)	RR	– –	80 80	– –	dB
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	270	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA)	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C
Peak Output Current (T_J = 25°C) (T_J = 25°C)	I_O	–	700	–	mA

MC78M08C ELECTRICAL CHARACTERISTICS

(V_I = 14 V, I_O = 200 mA, 0°C < T_J < +125°C, $P_D \leq$ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	7.7	8.0	8.3	Vdc
Line Regulation (T_J = +25°C) (10.5 Vdc ≤ V_I ≤ 25 Vdc) (11 Vdc ≤ V_I ≤ 25 Vdc)	Reg_{line}	 – –	 6.0 2.0	 100 50	mV
Load Regulation (T_J = +25°C, 5.0 mA ≤ I_O ≤ 500 mA) (T_J = +25°C, 5.0 mA ≤ I_O ≤ 200 mA)	Reg_{load}	 – –	 25 10	 160 80	mV
Output Voltage (10.5 Vdc ≤ V_I ≤ 25 Vdc, 5.0 mA ≤ I_O ≤ 200 mA)	V_O	7.6	–	8.4	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.6	6.0	mA
Quiescent Current Change (10.5 Vdc ≤ V_I ≤ 25 Vdc) (5.0 mA ≤ I_O ≤ 200 mA)	ΔI_{IB}	 – –	 – –	 0.8 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	52	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	32	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 11.5 V ≤ V_I ≤ 21.5 V) (I_O = 300 mA, f = 120 Hz, 11.5 V ≤ V_I ≤ 21.5 V, T_J = 25°C)	RR	– –	80 80	– –	dB
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	250	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA)	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

MC78M12C ELECTRICAL CHARACTERISTICS (V_I = 19 V, I_O = 200 mA, 0°C < T_J < +125°C, P_D ≤ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	11.5	12	12.5	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C)					
(14.5 Vdc ≤ V_I ≤ 30 Vdc)		–	8.0	100	
(16 Vdc ≤ V_I ≤ 22 Vdc)		–	2.0	50	
Load Regulation	Reg_{load}				mV
(T_J = +25°C, 5.0 mA ≤ I_O ≤ 500 mA)		–	25	240	
(T_J = +25°C, 5.0 mA ≤ I_O ≤ 200 mA)		–	10	120	
Output Voltage (14.5 Vdc ≤ V_I ≤ 27 Vdc, 5.0 mA ≤ I_O ≤ 200 mA)	V_O	11.4	–	12.6	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.8	6.0	mA
Quiescent Current Change	ΔI_{IB}				mA
(14.5 Vdc ≤ V_I ≤ 30 Vdc)		–	–	0.8	
(5.0 mA ≤ I_O ≤ 200 mA)		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	75	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	48	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 15 V ≤ V_I ≤ 25 V)	RR	–	80	–	dB
(I_O = 300 mA, f = 120 Hz, 15 V ≤ V_I ≤ 25 V, T_J = 25°C)		–	80	–	
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	240	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA, 0°C ≤ T_A ≤ +125°C)	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

MC78M15C ELECTRICAL CHARACTERISTICS (V_I = 23 V, I_O = 200 mA, 0°C < T_J < +125°C, P_D ≤ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	14.4	15	15.6	Vdc
Input Regulation	Reg_{line}				mV
(T_J = +25°C)					
(17.5 Vdc ≤ V_I ≤ 30 Vdc)		–	10	100	
(20 Vdc ≤ V_I ≤ 30 Vdc)		–	3.0	50	
Load Regulation	Reg_{load}				mV
(T_J = +25°C, 5.0 mA ≤ I_O ≤ 500 mA)		–	25	300	
(T_J = +25°C, 5.0 mA ≤ I_O ≤ 200 mA)		–	10	150	
Output Voltage 17.5 Vdc ≤ V_I ≤ 30 Vdc, 5.0 mA ≤ I_O ≤ 200 mA)	V_O	14.25	–	15.75	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.8	6.0	mA
Quiescent Current Change	ΔI_{IB}				
(18.5 Vdc ≤ V_I ≤ 30 Vdc)		–	–	0.8	
(5.0 mA ≤ I_O ≤ 200 mA)		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	90	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	60	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 18.5 V ≤ V_I ≤ 28.5 V)	RR	–	70	–	dB
(I_O = 300 mA, f = 120 Hz, 18.5 V ≤ V_I ≤ 28.5 V, T_J = 25°C)		–	70	–	
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	240	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA, 0°C ≤ T_A ≤ +125°C)	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

MC78M18C ELECTRICAL CHARACTERISTICS (V_I = 27 V, I_O = 200 mA, 0°C < T_J < +125°C, $P_D \leqslant$ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	17.3	18	18.7	Vdc
Line Regulation (T_J = +25°C) (21 Vdc $\leqslant V_I \leqslant$ 33 Vdc) (24 Vdc $\leqslant V_I \leqslant$ 33 Vdc)	Reg_{line}	 – –	 10 40	 100 50	mV
Load Regulation (T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 500 mA) (T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	Reg_{load}	 – –	 30 10	 360 180	mV
Output Voltage (21 Vdc $\leqslant V_I \leqslant$ 33 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	V_O	17.1	–	18.9	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.8	6.5	mA
Quiescent Current Change (21 Vdc $\leqslant V_I \leqslant$ 33 Vdc) (5.0 mA $\leqslant I_O \leqslant$ 200 mA)	ΔI_{IB}	 – –	 – –	 0.8 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz $\leqslant$ f $\leqslant$ 100 kHz)	e_{on}	–	100	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	72	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 22 V $\leqslant V_I \leqslant$ 32 V) (I_O = 300 mA, f = 120 Hz, 22 V $\leqslant V_I \leqslant$ 32 V, T_J = 25°C)	RR	– –	70 70	– –	dB
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	240	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA, 0°C $\leqslant T_A \leqslant$ +125°C)	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

MC78M20C ELECTRICAL CHARACTERISTICS (V_I = 29 V, I_O = 200 mA, 0°C < T_J < +125°C, $P_D \leqslant$ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	19.2	20	20.8	Vdc
Line Regulation (T_J = +25°C) (23 Vdc $\leqslant V_I \leqslant$ 35 Vdc) (24 Vdc $\leqslant V_I \leqslant$ 35 Vdc)	Reg_{line}	 – –	 10 5.0	 100 50	mV
Load Regulation (T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 500 mA) (T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	Reg_{load}	 – –	 30 10	 400 200	mV
Output Voltage (23 Vdc $\leqslant V_I \leqslant$ 35 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	V_O	19	–	21	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.9	6.5	mA
Quiescent Current Change (23 Vdc $\leqslant V_I \leqslant$ 35 Vdc) (5.0 mA $\leqslant I_O \leqslant$ 200 mA)	ΔI_{IB}	 – –	 – –	 0.8 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz $\leqslant$ f $\leqslant$ 100 kHz)	e_{on}	–	110	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	80	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 24 V $\leqslant V_I \leqslant$ 34 V) (I_O = 300 mA, f = 120 Hz, 24 V $\leqslant V_I \leqslant$ 34 V, T_J = 25°C)	RR	– –	70 70	– –	dB
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C, V_I = 35 V)	I_{OS}	–	240	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA, 0°C $\leqslant T_A \leqslant$ +125°C)	$\Delta V_O/\Delta T$	–	-1.1	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

MC78M24C ELECTRICAL CHARACTERISTICS (V_I = 33 V, I_O = 200 mA, 0°C < T_J < +125°C, $P_D \leqslant$ 5.0 W unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	23	24	25	Vdc
Line Regulation (T_J = +25°C) (27 Vdc $\leqslant V_I \leqslant$ 38 Vdc) (28 Vdc $\leqslant V_I \leqslant$ 38 Vdc)	Reg_{line}	 – –	 10 5.0	 100 50	mV
Load Regulation (T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 500 mA) (T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	Reg_{load}	 – –	 30 10	 480 240	mV
Output Voltage (27 Vdc $\leqslant V_I \leqslant$ 38 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 200 mA)	V_O	22.8	–	25.2	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	5.0	7.0	mA
Quiescent Current Change (27 Vdc $\leqslant V_I \leqslant$ 38 Vdc) (5.0 mA $\leqslant I_O \leqslant$ 200 mA)	ΔI_{IB}	 – –	 – –	 0.8 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz $\leqslant$ f $\leqslant$ 100 kHz)	e_{on}	–	170	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	96	mV/1.0 k Hrs
Ripple Rejection (I_O = 100 mA, f = 120 Hz, 28 V $\leqslant V_I \leqslant$ 38 V) (I_O = 300 mA, f = 120 Hz, 28 V $\leqslant V_I \leqslant$ 38 V, T_J = 25°C)	RR	– –	70 70	– –	dB
Input-Output Voltage Differential (T_A = +25°C)	V_I-V_O	–	2.0	–	Vdc
Short-Circuit Current Limit (T_J = +25°C)	I_{OS}	–	240	–	mA
Average Temperature Coefficient of Output Voltage (I_O = 5.0 mA, 0°C $\leqslant T_A \leqslant$ +125°C)	$\Delta V_O/\Delta T$	–	-1.2	–	mV/°C
Peak Output Current (T_J = 25°C)	I_O	–	700	–	mA

DEFINITIONS

Line Regulation – The change in output voltage for a change in the input voltage. The measurement is made under conditions of low dissipation or by using pulse techniques such that the average chip temperature is not significantly affected.

Load Regulation – The change in output voltage for a change in load current at constant chip temperature.

Maximum Power Dissipation – The maximum total device dissipation for which the regulator will operate within specifications.

Input Bias Current – That part of the input current that is not delivered to the load.

Output Noise Voltage – The rms ac voltage at the output, with constant load and no input ripple, measured over a specified frequency range.

Long Term Stability – Output voltage stability under accelerated life test conditions with the maximum rated voltage listed in the devices' electrical characteristics and maximum power dissipation.

TYPICAL PERFORMANCE CURVES

FIGURE 1 – WORST CASE POWER DISSIPATION versus AMBIENT TEMPERATURE TO-220 (CASE 313)

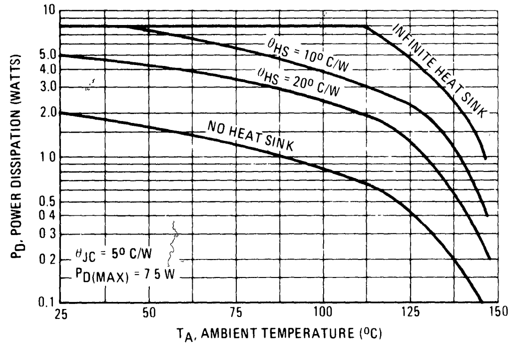


FIGURE 2 – WORST CASE POWER DISSIPATION versus AMBIENT TEMPERATURE TO-39 (CASE 79)

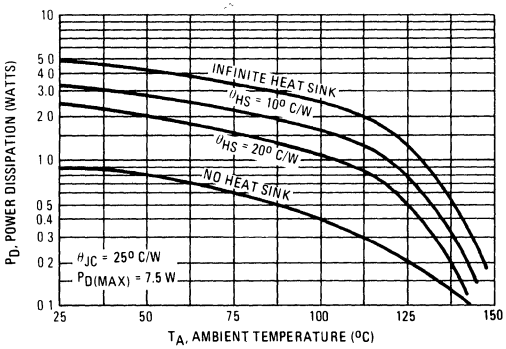


FIGURE 3 – PEAK OUTPUT CURRENT AS A FUNCTION OF INPUT-OUTPUT DIFFERENTIAL VOLTAGE

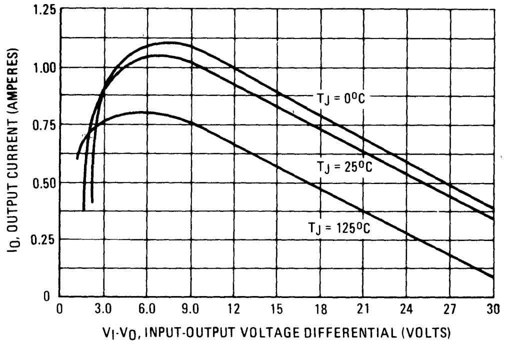
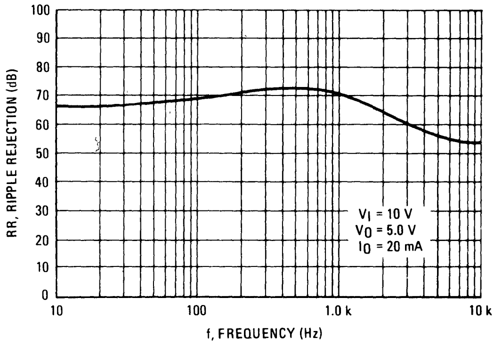


FIGURE 4 – RIPPLE REJECTION AS A FUNCTION OF FREQUENCY



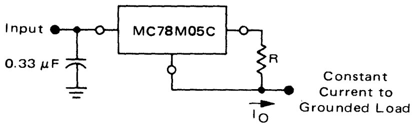
APPLICATIONS INFORMATION

Design Considerations

The MC78M00C Series of fixed voltage regulators are designed with Thermal Overload Protection that shuts down the circuit when subjected to an excessive power overload condition, Internal Short-Circuit Protection that limits the maximum current the circuit will pass, and Output Transistor Safe-Area Compensation that reduces the output short-circuit current as the voltage across the pass transistor is increased.

In many low current applications, compensation capacitors are not required. However, it is recommended that the regulator input be bypassed with a capacitor if the regulator is connected to the power supply filter with long wire lengths, or if the output load capacitance is large. An input bypass capacitor should be selected to provide good high-frequency characteristics to insure stable operation under all load conditions. A 0.33 µF or larger tantalum, mylar, or other capacitor having low internal impedance at high frequencies should be chosen. The bypass capacitor should be mounted with the shortest possible leads directly across the regulators input terminals. Normally good construction techniques should be used to minimize ground loops and lead resistance drops since the regulator has no external sense lead.

FIGURE 5 – CURRENT REGULATOR



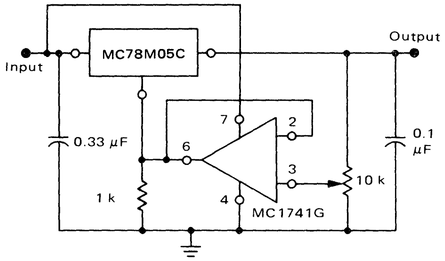
The MC7800C regulators can also be used as a current source when connected as above. In order to minimize dissipation the MC7805C is chosen in this application. Resistor R determines the current as follows:

$$I_O = \frac{5\ V}{R} + I_Q$$

I_Q = 1.5 mA over line and load changes

For example, a 500 mA current source would require R to be a 10-ohm, 10-W resistor and the output voltage compliance would be the input voltage less 7 volts.

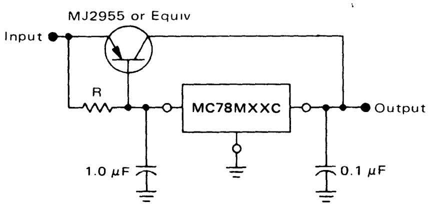
FIGURE 6 – ADJUSTABLE OUTPUT REGULATOR



V_O, 7.0 V to 20 V
$V_{IN} - V_O \geqslant 2.0$ V

The addition of an operational amplifier allows adjustment to higher or intermediate values while retaining regulation characteristics. The minimum voltage obtainable with this arrangement is 2.0 volts greater than the regulator voltage.

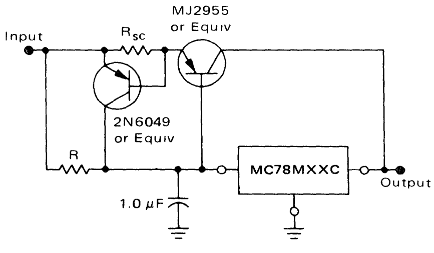
FIGURE 7 – CURRENT BOOST REGULATOR



XX = 2 digits of type number indicating voltage.

The MC78M00C series can be current boosted with a PNP transistor. The MJ2955 provides current to 5.0 amperes. Resistor R in conjunction with the V_{BE} of the PNP determines when the pass transistor begins conducting; this circuit is not short-circuit proof. Input-output differential voltage minimum is increased by V_{BE} of the pass transistor.

FIGURE 8 – SHORT-CIRCUIT PROTECTION



XX = 2 digits of type number indicating voltage.

The circuit of Figure 7 can be modified to provide supply protection against short circuits by adding a short-circuit sense resistor, R_{SC}, and an additional PNP transistor. The current sensing PNP must be able to handle the short-circuit current of the three-terminal regulator. Therefore, a two-ampere plastic power transistor is specified.

MOTOROLA

MC78T00 Series

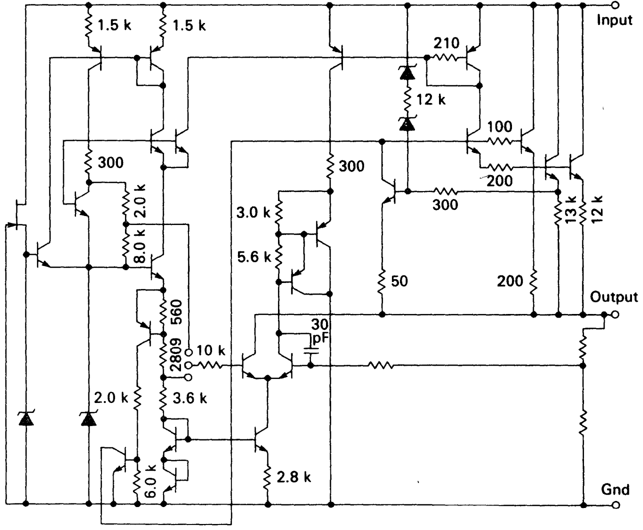
Product Preview

3-TERMINAL POSITIVE VOLTAGE REGULATORS

These voltage regulators are monolithic integrated circuits designed as fixed-voltage regulators for a wide variety of applications including local, on-card regulation. These regulators employ internal current limiting, thermal shutdown, and safe-area compensation. With adequate heatsinking they can deliver output currents in excess of 3.0 amperes. Although designed primarily as a fixed voltage regulator, these devices can be used with external components to obtain adjustable voltages and currents.

- Output Current in Excess of 3.0 Amperes
- No External Components Required
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Output Voltage Offered in 2% and 4% Tolerance*

SCHEMATIC DIAGRAM



ORDERING INFORMATION

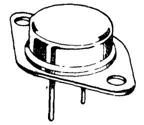
Device	Output Voltage Tolerance	Temperature Range	Package
MC78TXXK MC78TXXAK	4% 2%*	− 55 to + 150°C	Metal Power
MC78TXXCK MC78TXXACK	4% 2%*	0 to + 125°C	
MC78TXXCT MC78TXXACT	4% 2%*		Plastic Power

XX Indicates nominal voltage *2% regulators are available in 5, 12 and 15 volt devices

This document contains information on a product under development Motorola reserves the right to change or discontinue this product without notice

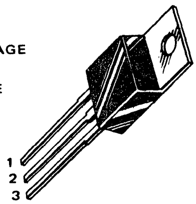
THREE-TERMINAL POSITIVE FIXED VOLTAGE REGULATORS

K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 TYPE)



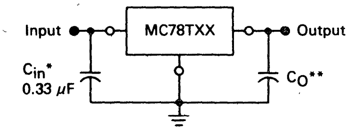
PIN 1. INPUT
2. OUTPUT
CASE GROUND

T SUFFIX
PLASTIC PACKAGE
CASE 221A
TO-220 TYPE



PIN 1. INPUT
2. GROUND
3. OUTPUT

STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.0 V above the output voltage even during the low point on the input ripple voltage.

XX = these two digits of the type number indicate voltage.

* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_O is not needed for stability; however, it does improve transient response.

XX indicates nominal voltage

TYPE NO./VOLTAGE			
MC78T05	5.0 Volts	MC78T15	15 Volts
MC78T06	6.0 Volts	MC78T18	18 Volts
MC78T08	8.0 Volts	MC78T24	24 Volts
MC78T12	12 Volts		

MOTOROLA

MC7900C
Series

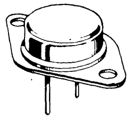
THREE-TERMINAL NEGATIVE FIXED VOLTAGE REGULATORS

MC7900C SERIES THREE-TERMINAL NEGATIVE VOLTAGE REGULATORS

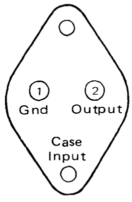
The MC7900C Series of fixed output negative voltage regulators are intended as complements to the popular MC7800C Series devices. These negative regulators are available in the same seven-voltage options as the MC7800C devices. In addition, two extra voltage options commonly employed in MECL systems are also available in the negative MC7900C Series.

Available in fixed output voltage options from -2.0 to -24 volts, these regulators employ current limiting, thermal shutdown, and safe-area compensation – making them remarkably rugged under most operating conditions. With adequate heat-sinking they can deliver output currents in excess of 1.0 ampere.

- No External Components Required
- Internal Thermal Overload Protection
- Internal Short-Circuit Current Limiting
- Output Transistor Safe-Area Compensation
- Packaged in the Plastic Case 221A and Case 1 (TO-220 and Hermetic TO-3)



K SUFFIX
METAL PACKAGE
CASE 1
(TO-3 TYPE)

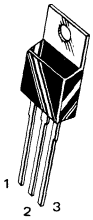


(bottom view)

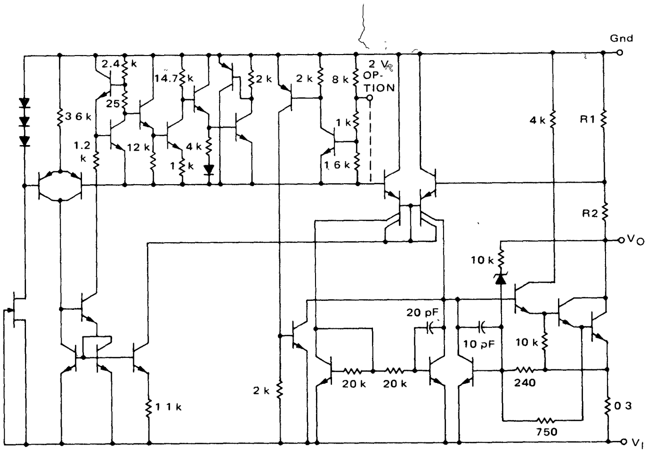
T SUFFIX
PLASTIC PACKAGE
CASE 221A

Pin 1. Ground
2. Input
3. Output

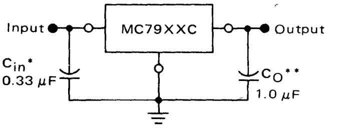
(Heatsink surface connected to Pin2)



SCHEMATIC DIAGRAM



STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.0 V more negative even during the high point on the input ripple voltage.

XX = these two digits of the type number indicate voltage.

* = C_{in} is required if regulator is located an appreciable distance from power supply filter.

** = C_O improves stability and transient response.

DEVICE TYPE/NOMINAL OUTPUT VOLTAGE

MC7902C - 2.0 Volts
MC7905C - 5.0 Volts
MC7905.2C – 5.2 Volts
MC7906C - 6.0 Volts
MC7908C - 8.0 Volts
MC7912C - 12 Volts
MC7915C - 15 Volts
MC7918C - 18 Volts
MC7924C - 24 Volts

ORDERING INFORMATION

DEVICE	TEMPERATURE RANGE	PACKAGE
MC79XXCK	T_J = 0° C to +150° C	Metal Power
MC79XXCT	T_J = 0° C to +150° C	Plastic Power

XX indicates nominal voltage

MC7900C Series MAXIMUM RATINGS (T_A = +25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Input Voltage (2.0 V – 18 V) (24 V)	V_I	-35 -40	Vdc
Power Dissipation			
Plastic Package			
T_A = +25°C	P_D	Internally Limited	Watts
Derate above T_A = +25°C	$1/R_{\theta JA}$	15.4	mW/°C
T_C = +25°C	P_D	Internally Limited	Watts
Derate above T_C = +95°C (See Figure 1)	$1/R_{\theta JC}$	200	mW/°C
Metal Package			
T_A = +25°C	P_D	Internally Limited	Watts
Derate above T_A = +25°C	$1/R_{\theta JA}$	22.2	mW/°C
T_C = +25°C	P_D	Internally Limited	Watts
Derate above T_C = +65°C	$1/R_{\theta JC}$	182	mW/°C
Storage Temperature Range	T_{stg}	-65 to +150	°C
Junction Temperature Range	T_J	0 to +150	°C

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Ambient – Plastic Package – Metal Package	$R_{\theta JA}$	65 45	°C/W
Thermal Resistance, Junction to Case – Plastic Package – Metal Package	$R_{\theta JC}$	5.0 5.5	°C/W

MC7902C ELECTRICAL CHARACTERISTICS (V_I = -10 V, I_O = 500 mA, 0°C $< T_J <$ +125°C unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-1.92	-2.00	-2.08	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-7.0 Vdc $\geqslant V_I \geqslant$ -25 Vdc		–	8.0	20	
-8.0 Vdc $\geqslant V_I \geqslant$ -12 Vdc		–	4.0	10	
(T_J = +25°C, I_O = 500 mA)					
-7.0 Vdc $\geqslant V_I \geqslant$ -25 Vdc		–	18	40	
-8.0 Vdc $\geqslant V_I \geqslant$ -12 Vdc		–	8.0	20	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA $\leqslant I_O \leqslant$ 1.5 A		–	70	120	
250 mA $\leqslant I_O \leqslant$ 750 mA		–	20	60	
Output Voltage -7.0 Vdc $\geqslant V_I \geqslant$ -20 Vdc, 5.0 mA $\leqslant I_O \leqslant$ 1.0 A, P $\leqslant$ 15 W	V_O	-1.90	–	-2.10	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.3	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-7.0 Vdc $\geqslant V_I \geqslant$ -25 Vdc		–	–	1.3	
5.0 mA $\leqslant I_O \leqslant$ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz $\leqslant$ f $\leqslant$ 100 kHz)	e_{on}	–	40	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	20	mV/1.0k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	65	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\|V_I - V_O\|$	–	3.5	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C $\leqslant T_A \leqslant$ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7905C ELECTRICAL CHARACTERISTICS (V_I = -10 V, I_O = 500 mA, 0°C < T_J < +125°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-4.8	-5.0	-5.2	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-7.0 Vdc ≥ V_I ≥ -25 Vdc		–	7.0	50	
-8.0 Vdc ≥ V_I ≥ -12 Vdc		–	2.0	25	
(T_J = +25°C, I_O = 500 mA)					
-7.0 Vdc ≥ V_I ≥ -25 Vdc		–	35	100	
-8.0 Vdc ≥ V_I ≥ -12 Vdc		–	8.0	50	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA ≤ I_O ≤ 1.5 A		–	11	100	
250 mA ≤ I_O ≤ 750 mA		–	4.0	50	
Output Voltage -7.0 Vdc ≥ V_I ≥ -20 Vdc, 5.0 mA ≤ I_O ≤ 1.0 A, P ≤ 15 W	V_O	-4.75	–	-5.25	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.3	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-7.0 Vdc ≥ V_{in} ≥ -25 Vdc		–	–	1.3	
5.0 mA ≤ I_O ≤ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	40	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	20	mV/1.0 k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	70	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\lvert V_I - V_O \rvert$	–	2.0	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C ≤ T_A ≤ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7905.2C ELECTRICAL CHARACTERISTICS (V_I = -10 V, I_O = 500 mA, 0°C < T_J < +125°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-5.0	-5.2	-5.4	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-7.2 Vdc ≥ V_I ≥ -25 Vdc		–	8.0	52	
-8.0 Vdc ≥ V_I ≥ -12 Vdc		–	2.2	27	
(T_J = +25°C, I_O = 500 mA)					
-7.2 Vdc ≥ V_I ≥ -25 Vdc		–	37	105	
-8.0 Vdc ≥ V_I ≥ -12 Vdc		–	8.5	52	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA ≤ I_O ≤ 1.5 A		–	12	105	
250 mA ≤ I_O ≤ 750 mA		–	4.5	52	
Output Voltage -7.2 Vdc ≥ V_I ≥ -20 Vdc, 5.0 mA ≤ I_O ≤ 1.0 A, P ≤ 15 W	V_O	-4.94	–	-5.46	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.3	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-7.2 Vdc ≥ V_I ≥ -25 Vdc		–	–	1.3	
5.0 mA ≤ I_O ≤ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	42	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	20	mV/1.0 k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	68	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\lvert V_I - V_O \rvert$	–	2.0	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C ≤ T_A ≤ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7906C ELECTRICAL CHARACTERISTICS (V_I = -11 V, I_O = 500 mA, $0^\circ C < T_J < +125^\circ C$ unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-5.75	-6.0	-6.25	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-8.0 Vdc $\geq V_I \geq$ -25 Vdc		–	9.0	60	
-9.0 Vdc $\geq V_I \geq$ -13 Vdc		–	3.0	30	
(T_J = +25°C, I_O = 500 mA)					
-8.0 Vdc $\geq V_I \geq$ -25 Vdc		–	43	120	
-9.0 Vdc $\geq V_I \geq$ -13 Vdc		–	10	60	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA $\leq I_O \leq$ 1.5 A		–	13	120	
250 mA $\leq I_O \leq$ 750 mA		–	5.0	60	
Output Voltage -8.0 Vdc $\geq V_I \geq$ -21 Vdc, 5.0 mA $\leq I_O \leq$ 1.0 A, P $\leq$ 15 W)	V_O	-5.7	–	-6.3	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.3	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-8.0 Vdc $\geq V_I \geq$ -25 Vdc		–	–	1.3	
5.0 mA $\leq I_O \leq$ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz $\leq f \leq$ 100 kHz)	e_{on}	–	45	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	24	mV/1.0k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	65	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\lvert V_I - V_O \rvert$	–	2.0	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C $\leq T_A \leq$ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7908C ELECTRICAL CHARACTERISTICS (V_I = -14 V, I_O = 500 mA, $0^\circ C < T_J < +125^\circ C$ unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-7.7	-8.0	-8.3	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-10.5 Vdc $\geq V_I \geq$ -25 Vdc		–	12	80	
-11 Vdc $\geq V_I \geq$ -17 Vdc		–	5.0	40	
(T_J = +25°C, I_O = 500 mA)					
-10.5 Vdc $\geq V_I \geq$ -25 Vdc		–	50	160	
-11 Vdc $\geq V_I \geq$ -17 Vdc		–	22	80	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA $\leq I_O \leq$ 1.5 A		–	26	160	
250 mA $\leq I_O \leq$ 750 mA		–	9.0	80	
Output Voltage -10.5 Vdc $\geq V_I \geq$ -23 Vdc, 5.0 mA $\leq I_O \leq$ 1.0 A, P $\leq$ 15 W	V_O	-7.6	–	-8.4	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.3	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-10.5 Vdc $\geq V_I \geq$ -25 Vdc		–	–	1.0	
5.0 mA $\leq I_O \leq$ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz $\leq f \leq$ 100 kHz)	e_{on}	–	52	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	32	mV/1.0k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	62	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\lvert V_I - V_O \rvert$	–	2.0	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C $\leq T_A \leq$ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7912C ELECTRICAL CHARACTERISTICS (V_I = -19 V, I_O = 500 mA, 0°C < T_J < +125°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-11.5	-12	-12.5	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-14.5 Vdc ≥ V_I ≥ -30 Vdc		–	13	120	
-16 Vdc ≥ V_I ≥ -22 Vdc		–	6.0	60	
(T_J = +25°C, I_O = 500 mA)					
-14.5 Vdc ≥ V_I ≥ -30 Vdc		–	55	240	
-16 Vdc ≥ V_I ≥ -22 Vdc		–	24	120	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA ≤ I_O ≤ 1.5 A		–	46	240	
250 mA ≤ I_O ≤ 750 mA		–	17	120	
Output Voltage -14.5 Vdc ≥ V_I ≥ -27 Vdc, 5.0 mA ≤ I_O ≤ 1.0 A, P ≤ 15 W	V_O	-11.4	–	-12.6	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.4	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-14.5 Vdc ≥ V_I ≥ -30 Vdc		–	–	1.0	
5.0 mA ≤ I_O ≤ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	75	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	48	mV/1.0 k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	61	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\lvert V_I - V_O \rvert$	–	2.0	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C ≤ T_A ≤ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7915C ELECTRICAL CHARACTERISTICS (V_I = -23 V, I_O = 500 mA, 0°C < T_J < +125°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-14.4	-15	-15.6	Vdc
Line Regulation	Reg_{line}				mV
(T_J = +25°C, I_O = 100 mA)					
-17.5 Vdc ≥ V_I ≥ -30 Vdc		–	14	150	
-20 Vdc ≥ V_I ≥ -26 Vdc		–	6.0	75	
(T_J = +25°C, I_O = 500 mA)					
-17.5 Vdc ≥ V_I ≥ -30 Vdc		–	57	300	
-20 Vdc ≥ V_I ≥ -26 Vdc		–	27	150	
Load Regulation	Reg_{load}				mV
T_J = +25°C, 5.0 mA ≤ I_O ≤ 1.5 A		–	68	300	
250 mA ≤ I_O ≤ 750 mA		–	25	150	
Output Voltage -17.5 Vdc ≥ V_I ≥ -30 Vdc, 5.0 mA ≤ I_O ≤ 1.0 A, P ≤ 15 W	V_O	-14.25	–	-15.75	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	–	4.4	8.0	mA
Input Bias Current Change	ΔI_{IB}				mA
-17.5 Vdc ≥ V_I ≥ -30 Vdc		–	–	1.0	
5.0 mA ≤ I_O ≤ 1.5 A		–	–	0.5	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	e_{on}	–	90	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	–	60	mV/1.0 k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	–	60	–	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\lvert V_I - V_O \rvert$	–	2.0	–	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C ≤ T_A ≤ +125°C	$\Delta V_O/\Delta T$	–	-1.0	–	mV/°C

MC7918C ELECTRICAL CHARACTERISTICS (V_I = -27 V, I_O = 500 mA, 0°C $< T_J <$ +125°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-17.3	-18	-18.7	Vdc
Line Regulation (T_J = +25°C, I_O = 100 mA) -21 Vdc $\geq V_I \geq$ -33 Vdc -24 Vdc $\geq V_I \geq$ -30 Vdc (T_J = +25°C, I_O = 500 mA) -21 Vdc $\geq V_I \geq$ -33 Vdc -24 Vdc $\geq V_I \geq$ -30 Vdc	Reg_{line}	 — — — —	 25 10 90 50	 180 90 360 180	mV
Load Regulation T_J = +25°C, 5.0 mA $\leq I_O \leq$ 1.0 A 250 mA $\leq I_O \leq$ 750 mA	Reg_{load}	 — —	 110 55	 360 180	mV
Output Voltage -21 Vdc $\geq V_I \geq$ -33 Vdc, 5.0 mA $\leq I_O \leq$ 1.0 A, P $\leq$ 15 W	V_O	-17.1	—	-18.9	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	—	4.5	8.0	mA
Input Bias Current Change -21 Vdc $\geq V_I \geq$ -33 Vdc 5.0 mA $\leq I_O \leq$ 1.0 A	ΔI_{IB}	 — —	 — —	 1.0 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz $\leq$ f $\leq$ 100 kHz)	e_{on}	—	110	—	µV
Long-Term Stability	$\Delta V_O/\Delta t$	—	—	72	mV/1.0 k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	—	59	—	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\|V_I - V_O\|$	—	2.0	—	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C $\leq T_A \leq$ +125°C	$\Delta V_O/\Delta T$	—	-1.0	—	mV/°C

MC7924C ELECTRICAL CHARACTERISTICS (V_I = -33 V, I_O = 500 mA, 0°C $< T_J <$ +125°C, unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
Output Voltage (T_J = +25°C)	V_O	-23	-24	-25	Vdc
Line Regulation (T_J = +25°C, I_O = 100 mA) -27 Vdc $\geq V_I \geq$ -38 Vdc -30 Vdc $\geq V_I \geq$ -36 Vdc (T_J = +25°C, I_O = 500 mA) -27 Vdc $\geq V_I \geq$ -38 Vdc -30 Vdc $\geq V_I \geq$ -36 Vdc	Reg_{line}	 — — — —	 31 14 118 70	 240 120 480 240	mV
Load Regulation T_J = +25°C, 5.0 mA $\leq I_O \leq$ 1.0 A 250 mA $\leq I_O \leq$ 750 mA	Reg_{load}	 — —	 150 85	 480 240	mV
Output Voltage -27 Vdc $\geq V_I \geq$ -38 Vdc, 5.0 mA $\leq I_O \leq$ 1.0 A, P $\leq$ 15 W	V_O	-22.8	—	-25.2	Vdc
Input Bias Current (T_J = +25°C)	I_{IB}	—	4.6	8.0	mA
Input Bias Current Change -27 Vdc $\geq V_I \geq$ -38 Vdc 5.0 mA $\leq I_O \leq$ 1.0 A	ΔI_{IB}	 — —	 — —	 1.0 0.5	mA
Output Noise Voltage (T_A = +25°C, 10 Hz $\leq$ f $\leq$ 100 kHz)	e_{on}	—	170	—	µV
Long-Term Stability	$\Delta V_O/\Delta t$	—	—	96	mV/1.0 k Hrs
Ripple Rejection (I_O = 20 mA, f = 120 Hz)	RR	—	56	—	dB
Input-Output Voltage Differential I_O = 1.0 A, T_J = +25°C	$\|V_I - V_O\|$	—	2.0	—	Vdc
Average Temperature Coefficient of Output Voltage I_O = 5.0 mA, 0°C $\leq T_A \leq$ +125°C	$\Delta V_O/\Delta T$	—	-1.0	—	mV/°C

TYPICAL CHARACTERISTICS

(T_A = +25°C unless otherwise noted.)

FIGURE 1 – WORST CASE POWER DISSIPATION AS A FUNCTION OF AMBIENT TEMPERATURE (TO-220)

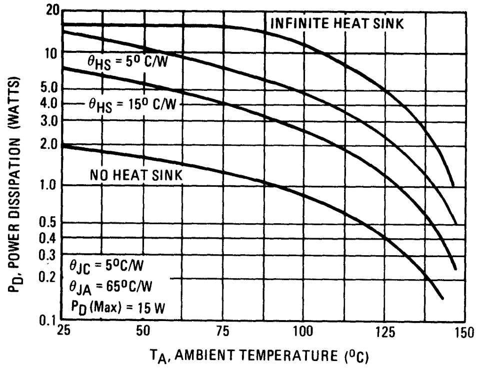


FIGURE 2 – WORST CASE POWER DISSIPATION AS A FUNCTION OF AMBIENT TEMPERATURE (TO-3)

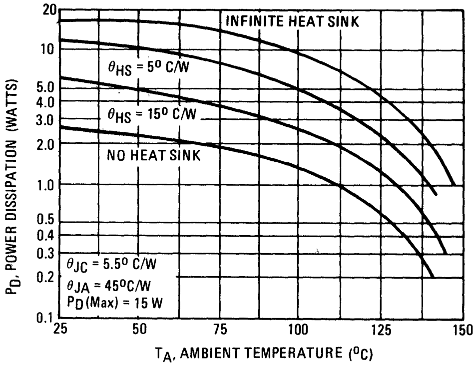


FIGURE 3 – PEAK OUTPUT CURRENT AS A FUNCTION OF INPUT-OUTPUT DIFFERENTIAL VOLTAGE

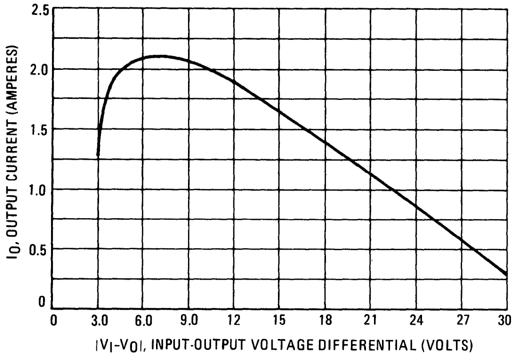


FIGURE 4 – RIPPLE REJECTION AS A FUNCTION OF FREQUENCY

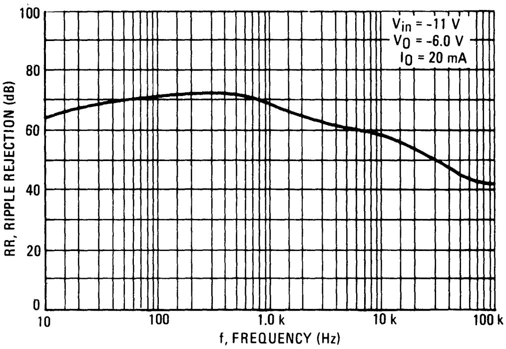


FIGURE 5 – RIPPLE REJECTION AS A FUNCTION OF OUTPUT VOLTAGES

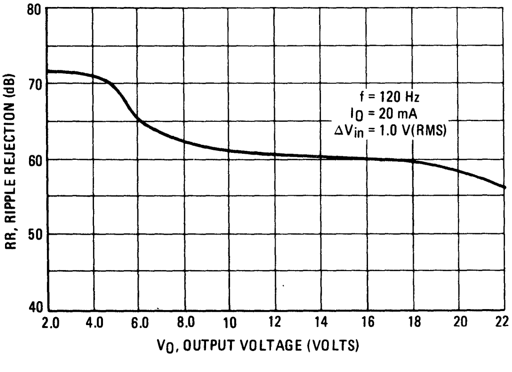
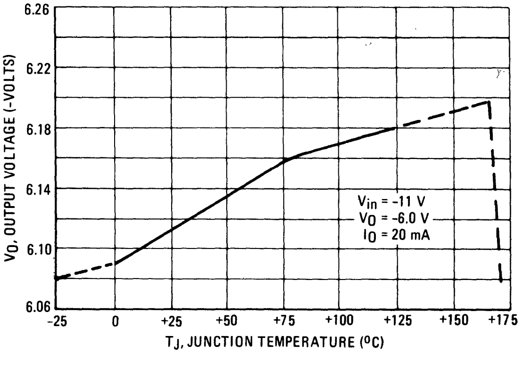
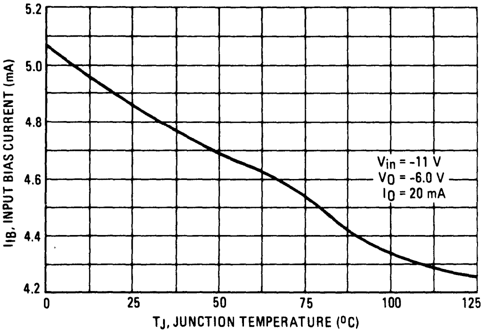


FIGURE 6 – OUTPUT VOLTAGE AS A FUNCTION OF JUNCTION TEMPERATURE



TYPICAL CHARACTERISTICS (continued)

FIGURE 7 – QUIESCENT CURRENT AS A FUNCTION OF TEMPERATURE



DEFINITIONS

Line Regulation – The change in output voltage for a change in the input voltage. The measurement is made under conditions of low dissipation or by using pulse techniques such that the average chip temperature is not significantly affected.

Load Regulation -- The change in output voltage for a change in load current at constant chip temperature.

Maximum Power Dissipation – The maximum total device dissipation for which the regulator will operate within specifications.

Input Bias Current – That part of the input current that is not delivered to the load.

Output Noise Voltage – The rms ac voltage at the output, with constant load and no input ripple, measured over a specified frequency range.

Long Term Stability - Output voltage stability under accelerated life test conditions with the maximum rated voltage listed in the devices' electrical characteristics and maximum power dissipation.

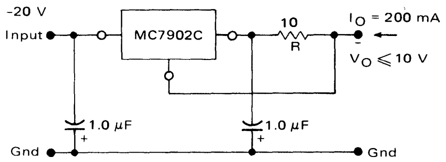
APPLICATIONS INFORMATION

Design Considerations

The MC7900C Series of fixed voltage regulators are designed with Thermal Overload Protection that shuts down the circuit when subjected to an excessive power overload condition, Internal Short-Circuit Protection that limits the maximum current the circuit will pass, and Output Transistor Safe-Area Compensation that reduces the output short-circuit current as the voltage across the pass transistor is increased.

In many low current applications, compensation capacitors are not required However, it is recommended that the regulator input be bypassed with a capacitor if the regulator is connected to the power supply filter with long wire lengths, or if the output load capacitance is large. An input bypass capacitor should be selected to provide good high-frequency characteristics to insure stable operation under all load conditions. A 0.33 μF or larger tantalum, mylar, or other capacitor having low internal impedance at high frequencies should be chosen. The bypass capacitor should be mounted with the shortest possible leads directly across the regulators input terminals. Normally good construction techniques should be used to minimize ground loops and lead resistance drops since the regulator has no external sense lead. Bypassing the output is also recommended.

FIGURE 8 – CURRENT REGULATOR



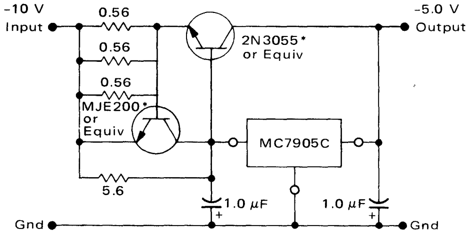
The MC7902, -2.0 V regulator can be used as a constant current source when connected as above. The output current is the sum of resistor R current and quiescent bias current as follows:

$$I_O = \frac{2\text{ V}}{R} + I_B$$

The quiescent current for this regulator is typically 4.3 mA. The 2.0 volt regulator was chosen to minimize dissipation and to allow the output voltage to operate to within 6.0 V below the input voltage.

FIGURE 9 – CURRENT BOOST REGULATOR

(-5.0 V @ 4.0 A, with 5.0 A current limiting)

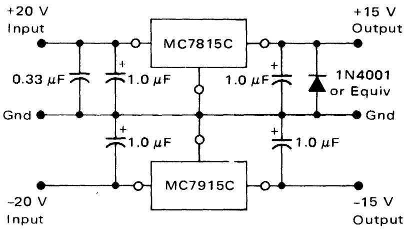


*Mounted on common heat sink, Motorola MS-10 or equivalent.

When a boost transistor is used, short-circuit currents are equal to the sum of the series pass and regulator limits, which are measured at 3.2 A and 1.8 A respectively in this case. Series pass limiting is approximately equal to 0.6 V/R_{SC}. Operation beyond this point to the peak current capability of the MC7905C is possible if the regulator is mounted on a heat sink; otherwise thermal shutdown will occur when the additional load current is picked up by the regulator.

FIGURE 10 – OPERATIONAL AMPLIFIER SUPPLY

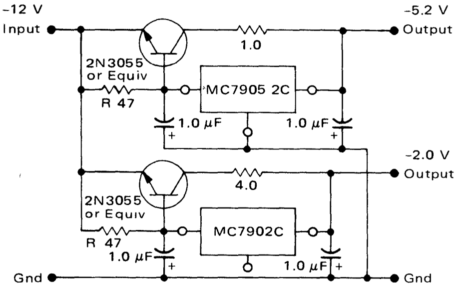
(±15 V @ 1.0 A)



The MC7815 and MC7915 positive and negative regulators may be connected as shown to obtain a dual power supply for operational amplifiers. A clamp diode should be used at the output of the MC7815 to prevent potential latch-up problems.

FIGURE 11 – TYPICAL MECL SYSTEM POWER SUPPLY

(-5.2 V @ 4.0 A and -2.0 V @ 2.0 A; for PC Board)



When current-boost power transistors are used, 47-ohm base-to-emitter resistors (R) must be used to bypass the quiescent current at no load. These resistors, in conjunction with the V_{BE} of the NPN transistors, determine when the pass transistors begin conducting. The 1-ohm and 4-ohm dropping resistors were chosen to reduce the power dissipated in the boost transistors but still leave at least 2.0 V across these devices for good regulation.

MOTOROLA

MC79L00C,AC series

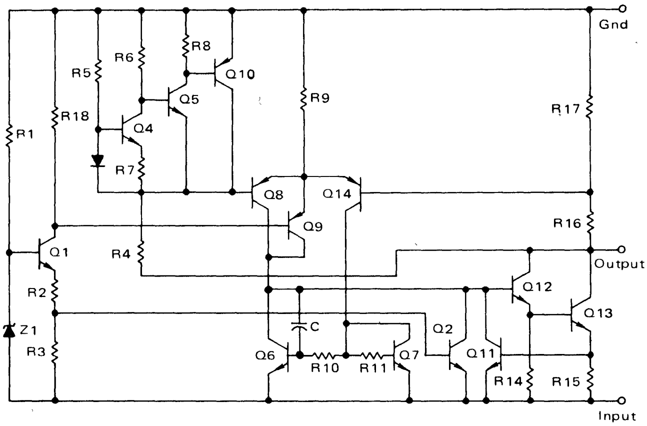
THREE-TERMINAL NEGATIVE VOLTAGE REGULATORS

The MC79L00 Series negative voltage regulators are inexpensive, easy-to-use devices suitable for numerous applications requiring up to 100 mA. Like the higher powered MC7900 Series negative regulators, this series features thermal shutdown and current limiting, making them remarkably rugged. In most applications, no external components are required for operation.

The MC79L00 devices are useful for on-card regulation or any other application where a regulated negative voltage at a modest current level is needed. These regulators offer substantial advantage over the common resistor/zener diode approach.

- No External Components Required
- Internal Short-Circuit Current Limiting
- Internal Thermal Overload Protection
- Low Cost
- Complementary Positive Regulators Offered (MC78L00 Series)
- Available in Either ±5% (AC) or ±10% (C) Selections

REPRESENTATIVE CIRCUIT SCHEMATIC

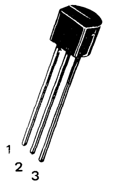


Device No. ±10%	Device No. ±5%	Nominal Voltage
MC79L03C	MC79L03AC	−3.0
MC79L05C	MC79L05AC	−5.0
MC79L12C	MC79L12AC	−12
MC79L15C	MC79L15AC	−15
MC79L18C	MC79L18AC	−18
MC79L24C	MC79L24AC	−24

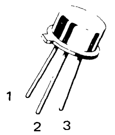
THREE-TERMINAL NEGATIVE FIXED VOLTAGE REGULATORS

P SUFFIX
CASE 29
TO-92

Pin 1. Ground
2. Input
3. Output

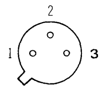


G SUFFIX
CASE 79
TO-39



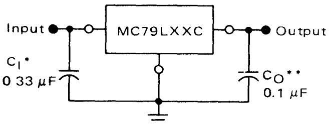
Pin 1. Ground
2. Output
3. Input

(Case connected to pin 3)



Bottom View

STANDARD APPLICATION



A common ground is required between the input and the output voltages. The input voltage must remain typically 2.0 V above the output voltage even during the low point on the input ripple voltage.

* = C_I is required if regulator is located an appreciable distance from power supply filter.

** = C_O improves stability and transient response.

ORDERING INFORMATION

Device	Temperature Range	Package
MC79LXXACG	T_J = 0°C to +150°C	Metal Can
MC79LXXACP	T_J = 0°C to +150°C	Plastic Power
MC79LXXCG	T_J = 0°C to +150°C	Metal Can
MC79LXXCP	T_J = 0°C to +150°C	Plastic Power

XX indicates nominal voltage

MC79L00C Series MAXIMUM RATINGS (T_A = +25°C unless otherwise noted.)

Rating	Symbol	Value	Unit
Input Voltage (-3,-5 V) (-12,-15,-18 V) (-24 V)	V_I	-30 -35 -40	Vdc
Storage Temperature Range	T_{stg}	-65 to +150	°C
Junction Temperature Range	T_J	0 to +150	°C

MC79L03C, AC ELECTRICAL CHARACTERISTICS (V_I = -10 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC79L03C			MC79L03AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	-2.76	-3.00	-3.24	-2.88	-3.0	-3.12	Vdc
Input Regulation (T_J = +25°C)	Reg_{line}							mV
-7.0 Vdc ⩾ V_I ⩾ -20 Vdc		–	–	80	–	–	60	
-8.0 Vdc ⩾ V_I ⩾ -20 Vdc		–	–	60	–	–	40	
Load Regulation	Reg_{load}							mV
T_J = +25°C, 1.0 mA ⩽ I_O ⩽ 100 mA		–	–	72	–	–	72	
1.0 mA ⩽ I_O ⩽ 40 mA		–	–	36	–	–	36	
Output Voltage	V_O							Vdc
-7.0 Vdc ⩾ V_I ⩾ -20 Vdc, 1.0 mA ⩽ I_O ⩽ 40 mA		-2.7	–	-3.3	-2.85	–	-3.15	
V_I = -10 Vdc, 1.0 mA ⩽ I_O ⩽ 70 mA		-2.7	–	-3.3	-2.85	–	-3.15	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	–	6.0	–	–	6.0	
(T_J = +125°C)		–	–	5.5	–	–	5.5	
Input Bias Current Change	ΔI_{IB}							mA
-8.0 Vdc ⩾ V_I ⩾ -20 Vdc		–	–	-1.5	–	–	-1.5	
1.0 mA ⩽ I_O ⩽ 40 mA		–	–	-0.2	–	–	-0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ⩽ f ⩽ 100 kHz)	V_N	–	30	–	–	30	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	10	–	–	10	–	mV/1.0 k Hrs.
Ripple Rejection (-8.0 ⩾ V_I ⩾ -18 Vdc, f = 120 Hz, T_J = 25°C)	RR	44	51	–	45	51	–	dB
Input-Output Voltage Differential I_O = 40 mA, T_J = +25°C	$/V_I-V_O/$	–	1.7	–	–	1.7	–	Vdc

MC79L05C, AC Series ELECTRICAL CHARACTERISTICS

(V_I = -10 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC79L05C			MC79L05AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	-4.6	-5.0	-5.4	-4.8	-5.0	-5.2	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C)								
-7.0 Vdc ≥ V_I ≥ -20 Vdc		–	–	200	–	–	150	
-8.0 Vdc ≥ V_I ≥ -20 Vdc		–	–	150	–	–	100	
Load Regulation	Reg_{load}							mV
T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA		–	–	60	–	–	60	
1.0 mA ≤ I_O ≤ 40 mA		–	–	30	–	–	30	
Output Voltage	V_O							Vdc
-7.0 Vdc ≥ V_I ≥ -20 Vdc, 1.0 mA ≤ I_O ≤ 40 mA		-4.5	–	-5.5	-4.75	–	-5.25	
V_I = -10 Vdc, 1.0 mA ≤ I_O ≤ 70 mA		-4.5	–	-5.5	-4.75	–	-5.25	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	–	6.0	–	–	6.0	
(T_J = +125°C)		–	–	5.5	–	–	5.5	
Input Bias Current Change	ΔI_{IB}							mA
-8.0 Vdc ≥ V_I ≥ -20 Vdc		–	–	1.5	–	–	1.5	
1.0 mA ≤ I_O ≤ 40 mA		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	40	–	–	40	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	12	–	–	12	–	mV/1.0 k Hrs.
Ripple Rejection (-8.0 ≥ V_I ≥ 18 Vdc, f = 120 kHz, T_J = 25°C)	RR	40	49	–	41	49	–	dB
Input-Output Voltage Differential I_O = 40 mA, T_J = +25°C	$/V_I-V_O/$	–	1.7	–	–	1.7	–	Vdc

MC79L12C, AC ELECTRICAL CHARACTERISTICS

(V_I = -19 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC79L12C			MC79L12AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	-11.1	-12	-12.9	-11.5	-12	-12.5	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C)								
-14.5 Vdc ≥ V_I ≥ -27 Vdc		–	–	250	–	–	250	
-16 Vdc ≥ V_I ≥ -27 Vdc		–	–	200	–	–	200	
Load Regulation	Reg_{load}							mV
T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA		–	–	100	–	–	100	
1.0 mA ≤ I_O ≤ 40 mA		–	–	50	–	–	50	
Output Voltage	V_O							Vdc
-14.5 Vdc ≥ V_I ≥ -27 Vdc, 1.0 mA ≤ I_O ≤ 40 mA		-10.8	–	-13.2	-11.4	–	-12.6	
V_I = -19 Vdc, 1.0 mA ≤ I_O ≤ 70 mA		-10.8	–	-13.2	-11.4	–	-12.6	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	–	6.5	–	–	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
-16 Vdc ≥ V_I ≥ -27 Vdc		–	–	1.5	–	–	1.5	
1.0 mA ≤ I_O ≤ 40 mA		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	80	–	–	80	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	24	–	–	24	–	mV/1.0 k Hrs.
Ripple Rejection (-15 ≤ V_I ≤ -25 Vdc, f = 120Hz, T_J = +25°C)	RR	36	42	–	37	42	–	dB
Input-Output Voltage Differential I_O = 40 mA, T_J = +25°C	$/V_I-V_O/$	–	1.7	–	–	1.7	–	Vdc

MC79L15C, AC ELECTRICAL CHARACTERISTICS (V_I = -23 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC79L15C			MC79L15AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	-13.8	-15	-16.2	-14.4	-15	-15.6	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C)								
-17.5 Vdc ≥ V_I ≥ -30 Vdc		–	–	300	–	–	300	
-20 Vdc ≥ V_I ≥ -30 Vdc		–	–	250	–	–	250	
Load Regulation	Reg_{load}							mV
T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA		–	–	150	–	–	150	
1.0 mA ≤ I_O ≤ 40 mA		–	–	75	–	–	75	
Output Voltage	V_O							Vdc
-17.5 Vdc ≥ V_I ≥ -30 Vdc, 1.0 mA ≤ I_O ≤ 40 mA		-13.5	–	-16.5	-14.25	–	-15.75	
V_I = -23 Vdc, 1.0 mA ≤ I_O ≤ 70 mA		-13.5	–	-16.5	-14.25	–	-15.75	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	–	6.5	–	–	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
-20 Vdc ≥ V_I ≥ -30 Vdc		–	–	1.5	–	–	1.5	
1.0 mA ≤ I_O ≤ 40 mA		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	90	–	–	90	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	30	–	–	30	–	mV/1.0 k Hrs.
Ripple Rejection (-18.5 ≤ V_I ≤ -28.5 Vdc, f = 120 Hz)	RR	33	39	–	34	39	–	dB
Input-Output Voltage Differential I_O = 40 mA, T_J = +25°C	$/V_I-V_O/$	–	1.7	–	–	1.7	–	Vdc

MC79L18C, AC ELECTRICAL CHARACTERISTICS (V_I = -27 V, I_O = 40 mA, C_I = 0.33 μF, C_O = 0.1 μF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC79L18C			MC79L18AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	-16.6	-18	-19.4	-17.3	-18	-18.7	Vdc
Input Regulation	Reg_{line}							mV
(T_J = +25°C)								
-20.7 Vdc ≥ V_I ≥ -33 Vdc		–	–	–	–	–	325	
-21.4 Vdc ≥ V_I ≥ -33 Vdc		–	–	325	–	–	–	
-22 Vdc ≥ V_I ≥ -33 Vdc		–	–	275	–	–	–	
-21 Vdc ≥ V_I ≥ -33 Vdc		–	–	–	–	–	275	
Load Regulation	Reg_{load}							mV
T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA		–	–	170	–	–	170	
1.0 mA ≤ I_O ≤ 40 mA		–	–	85	–	–	85	
Output Voltage	V_O							Vdc
-20.7 Vdc ≥ V_I ≥ -33 Vdc, 1.0 mA ≤ I_O ≤ 40 mA		–	–	–	-17.1	–	-18.9	
-21.4 Vdc ≥ V_I ≥ -33 Vdc, 1.0 mA ≤ I_O ≤ 40 mA		-16.2	–	-19.8	–	–	–	
V_I = -27 Vdc, 1.0 mA ≤ I_O ≤ 70 mA		-16.2	–	-19.8	-17.1	–	-18.9	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	–	6.5	–	–	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
-21 Vdc ≥ V_I ≥ -33 Vdc		–	–	–	–	–	1.5	
-27 Vdc ≥ V_I ≥ -33 Vdc		–	–	1.5	–	–	–	
1.0 mA ≤ I_O ≤ 40 mA		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A= +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	150	–	–	150	–	μV
Long-Term Stability	$\Delta V_O/\Delta t$	–	45	–	–	45	–	mV/1.0 k Hrs.
Ripple Rejection (-23 ≤ V_I ≤ -33 Vdc, f = 120 Hz, T_J = +25°C)	RR	32	46	–	33	48	–	dB
Input-Output Voltage Differential I_O = 40 mA, T_J = +25°C	$/V_I-V_O/$	–	1.7	–	–	1.7	–	Vdc

MC79L24C, AC ELECTRICAL CHARACTERISTICS (V_I = -33 V, I_O = 40 mA, C_I = 0.33 µF, C_O = 0.1 µF, 0°C < T_J < +125°C unless otherwise noted.)

Characteristic	Symbol	MC79L24C			MC79L24AC			Unit
		Min	Typ	Max	Min	Typ	Max	
Output Voltage (T_J = +25°C)	V_O	-22.1	-24	-25.9	-23	-24	-25	Vdc
Input Regulation (T_J = +25°C)	Reg_{line}							mV
-27 Vdc ≥ V_I ≥ -38 V		–	–	–	–	–	350	
-27.5 Vdc ≥ V_I ≥ -38 Vdc		–	–	350	–	–	–	
-28 Vdc ≥ V_I ≥ -38 Vdc		–	–	300	–	–	300	
Load Regulation	Reg_{load}							mV
T_J = +25°C, 1.0 mA ≤ I_O ≤ 100 mA		–	–	200	–	–	200	
1.0 mA ≤ I_O ≤ 40 mA		–	–	100	–	–	100	
Output Voltage	V_O							Vdc
-27 Vdc ≥ V_I ≥ -38 V, 1.0 mA ≤ I_O ≤ 40 mA		–	–	–	-22.8	–	-25.2	
-28 Vdc ≥ V_I ≥ -38 Vdc, 1.0 mA ≤ I_O ≤ 40 mA		-21.4	–	-26.4	–	–	–	
V_I = -33 Vdc, 1.0 mA ≤ I_O ≤ 70 mA		-21.4	–	-26.4	-22.8	–	-25.2	
Input Bias Current	I_{IB}							mA
(T_J = +25°C)		–	–	6.5	–	–	6.5	
(T_J = +125°C)		–	–	6.0	–	–	6.0	
Input Bias Current Change	ΔI_{IB}							mA
-28 Vdc ≥ V_I ≥ -38 Vdc		–	–	1.5	–	–	1.5	
1.0 mA ≤ I_O ≤ 40 mA		–	–	0.2	–	–	0.1	
Output Noise Voltage (T_A = +25°C, 10 Hz ≤ f ≤ 100 kHz)	V_N	–	200	–	–	200	–	µV
Long-Term Stability	$\Delta V_O/\Delta t$	–	56	–	–	56	–	mV/1.0 k Hrs.
Ripple Rejection (-29 ≤ V_I ≤ -35 Vdc, f = 120 Hz, T_J = 25°C)	RR	30	43	–	31	47	–	dB
Input-Output Voltage Differential I_O = 40 mA, T_J = +25°C	$/V_I-V_O/$	–	1.7	–	–	1.7	–	Vdc

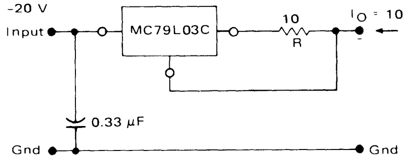
APPLICATIONS INFORMATION

Design Considerations

The MC79L00C Series of fixed voltage regulators are designed with Thermal Overload Protection that shuts down the circuit when subjected to an excessive power overload condition, Internal Short-Circuit Protection that limits the maximum current the circuit will pass.

In many low current applications, compensation capacitors are not required However, it is recommended that the regulator input be bypassed with a capacitor if the regulator is connected to the power supply filter with long wire lengths, or if the output load capacitance is large An input bypass capacitor should be selected to provide good high-frequency characteristics to insure stable operation under all load conditions. A 0.33 µF or larger tantalum, mylar, or other capacitor having low internal impedance at high frequencies should be chosen. The bypass capacitor should be mounted with the shortest possible leads directly across the regulators input terminals. Normally good construction techniques should be used to minimize ground loops and lead resistance drops since the regulator has no external sense lead. Bypassing the output is also recommended.

CURRENT REGULATOR

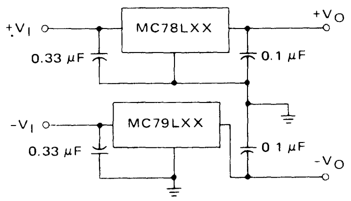


The MC79L03, -3.0 V regulator can be used as a constant current source when connected as above. The output current is the sum of resistor R current and quiescent bias current as follows

$$I_O = \frac{3\ V}{R} + I_B$$

The quiescent current for this regulator is typically 3.8 mA The -3.0 volt regulator was chosen to minimize dissipation and to allow the output voltage to operate to within 6 0 V below the input voltage

POSITIVE AND NEGATIVE REGULATOR



TYPICAL CHARACTERISTICS

(T_A = +25°C unless otherwise noted.)

FIGURE 1 – DROPOUT CHARACTERISTICS

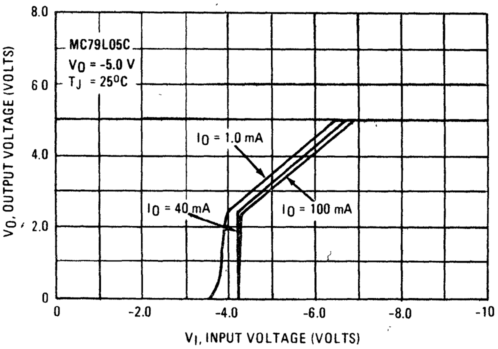


FIGURE 2 – DROPOUT VOLTAGE versus JUNCTION TEMPERATURE

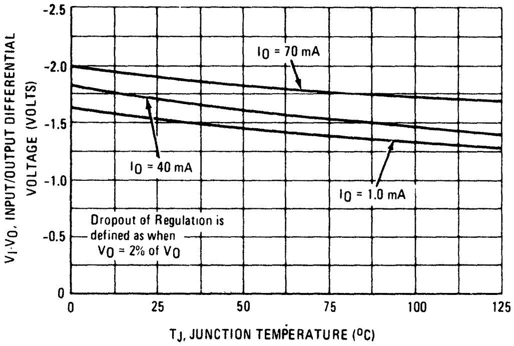


FIGURE 3 – INPUT BIAS CURRENT versus AMBIENT TEMPERATURE

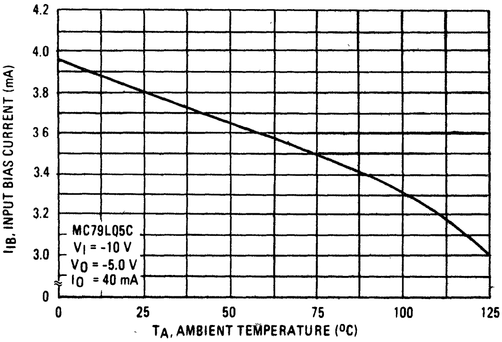


FIGURE 4 – INPUT BIAS CURRENT versus INPUT VOLTAGE

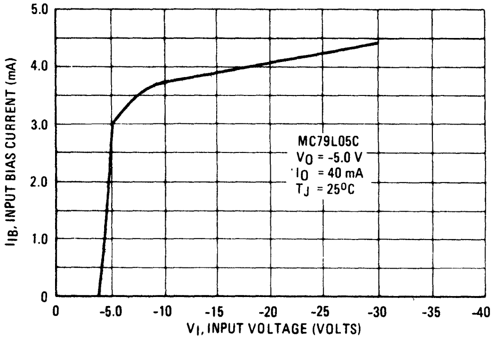


FIGURE 5 – MAXIMUM AVERAGE POWER DISSIPATION versus AMBIENT TEMPERATURE – TO-92 Type Package

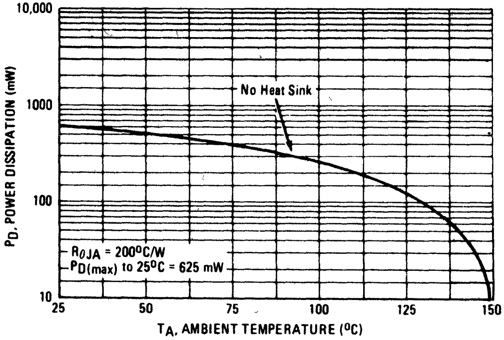
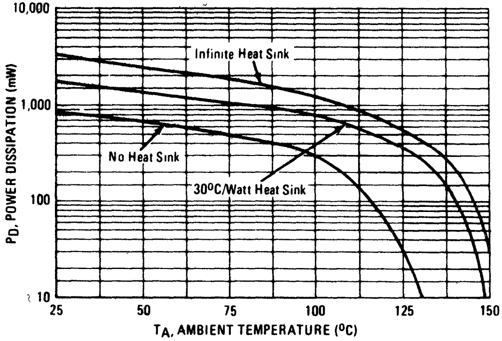


FIGURE 6 – MAXIMUM AVERAGE POWER DISSIPATION versus AMBIENT TEMPERATURE – TO-39 Type Package



MOTOROLA

SG1525A/SG1527A
SG2525A/SG2527A
SG3525A/SG3527A

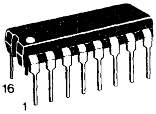
PULSE WIDTH MODULATOR CONTROL CIRCUIT

The SG1525A/1527A series of pulse width modulator control-circuits offer improved performance and lower external parts count when implemented for controlling all types of switching power supplies. The device includes a +5.1 volt ±1% reference and an error amplifier with a common-mode range including the reference voltage to eliminate external divider resistors. A sync input to the oscillator enables multiple units to be slaved together, or a single unit can be synchronized to an external system clock A wide range of dead time is programmable with a single resistor between the C_T pin and the Discharge pin. Other features included are soft-start circuitry requiring only an external timing capacitor. A shutdown pin controls both the soft-start circuitry and the output stages, allowing fast output turn-off with soft-start recycle turn-on. Undervoltage lockout keeps the outputs off when V_{CC} is less than the required level for normal operation The output stages are a totem-pole design capable of sinking and sourcing in excess of 200 mA. The SG1525A series output stage features NOR Logic, giving a low output for an off state The SG1527A utilizes OR Logic which results in a high output level when off. These devices are available in Military, Industrial and Commercial temperature ranges and feature·

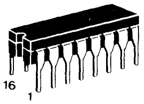
- 8 0 to 35 Volt Operation
- 5 1 Volt ±1% Trimmed Reference
- 100 Hz to 400 kHz Oscillator Range
- Separate Oscillator Sync Pin
- Adjustable Dead Time
- Input Undervoltage Lockout
- Latching PWM to Prevent Multiple Pulses
- Dual Source/Sink Output Current. ±400 mA Peak

PULSE WIDTH MODULATOR CONTROL CIRCUITS

SILICON MONOLITHIC INTEGRATED CIRCUITS

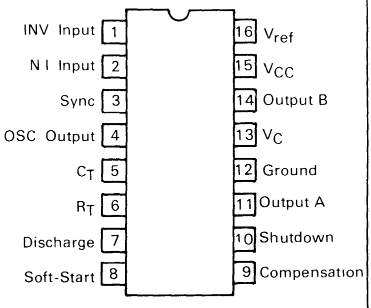


N SUFFIX
PLASTIC PACKAGE
CASE 648



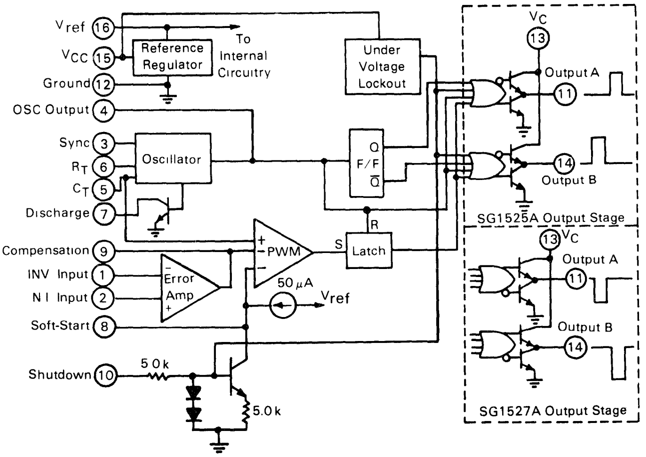
J SUFFIX
CERAMIC PACKAGE
CASE 620

PIN CONNECTIONS



(Top View)

FUNCTIONAL BLOCK DIAGRAM



ORDERING INFORMATION

Device	Temperature Range	Package
SG1525AJ	-55 to +125°C	Ceramic Dip
SG1527AJ	-55 to +125°C	Ceramic Dip
SG2525AJ	-40 to +85°C	Ceramic Dip
SG2525AN	-40 to +85°C	Plastic Dip
SG2527AJ	-40 to +85°C	Ceramic Dip
SG2527AN	-40 to +85°C	Plastic Dip
SG3525AJ	0 to +70°C	Ceramic Dip
SG3525AN	0 to +70°C	Plastic Dip
SG3527AJ	0 to +70°C	Ceramic Dip
SG3527AN	0 to +70°C	Plastic Dip

MAXIMUM RATINGS (Note 1)

Rating	Symbol	Value	Unit
Supply Voltage	V_{CC}	+40	Vdc
Collector Supply Voltage	V_C	+40	Vdc
Logic Inputs	—	-0.3 to +5.5	V
Analog Inputs	—	-0.3 to V_{CC}	V
Output Current, Source or Sink	I_O	±500	mA
Reference Output Current	I_{ref}	50	mA
Oscillator Charging Current	—	5.0	mA
Power Dissipation (Plastic & Ceramic Package) Note 2, T_A = +25°C Note 3, T_C = +25°C	P_D	 1000 2000	mW
Thermal Resistance Junction to Air Plastic and Ceramic Package	$R_{\theta JA}$	100	°C/W
Thermal Resistance Junction to Case Plastic and Ceramic Package	$R_{\theta JC}$	60	°C/W
Operating Junction Temperature	T_J	+150	°C
Storage Temperature Range Ceramic Package Plastic Package	T_{stg}	-65 to +150 -55 to +125	°C
Lead Temperature (Soldering, 10 Seconds)	T_{Solder}	+300	°C

NOTES
1 Values beyond which damage may occur
2 Derate at 10 mW/°C for ambient temperatures above +50°C
3 Derate at 16 mW/°C for case temperatures above +25°C

RECOMMENDED OPERATING CONDITIONS

Characteristic	Symbol	Min.	Max.	Unit
Supply Voltage	V_{CC}	+8.0	+35	Vdc
Collector Supply Voltage	V_C	+4.5	+35	Vdc
Output Sink/Source Current (Steady State) (Peak)	I_O	 0 0	 ±100 ±400	mA
Reference Load Current	I_{ref}	0	20	mA
Oscillator Frequency Range	f_{osc}	0.1	400	kHz
Oscillator Timing Resistor	R_T	2.0	150	kΩ
Oscillator Timing Capacitor	C_T	0.001	0.1	μF
Deadtime Resistor Range	R_D	0	500	Ω
Operating Ambient Temperature Range SG1525A, SG1527A SG2525A, SG2527A SG3525A, SG3527A	T_A	 -55 -40 0	 +125 +85 +70	°C

ELECTRICAL CHARACTERISTICS (V_{CC} = +20 Vdc, T_A = T_{low} to T_{high} [Note 4], unless otherwise specified)

		SG1525A/2525A SG1527A/2527A			SG3525A SG3527A			
Characteristic	**Symbol**	**Min**	**Typ**	**Max**	**Min**	**Typ**	**Max**	**Unit**
REFERENCE SECTION								
Reference Output Voltage (T_J = +25°C)	V_{ref}	5.05	5.10	5.15	5.00	5.10	5.20	Vdc
Line Regulation (+8.0 V ≤ V_{CC} ≤ +35 V)	Reg_{line}	—	10	20	—	10	20	mV
Load Regulation (0 mA ≤ I_L ≤ 20 mA)	Reg_{load}	—	20	50	—	20	50	mV
Temperature Stability	$\Delta V_{ref}/\Delta T$	—	20	50	—	20	50	mV
Total Output Variation Includes Line and Load Regulation over Temperature	ΔV_{ref}	5.00	—	5.20	4 95	—	5 25	Vdc
Short Circuit Current (V_{ref} = 0 V, T_J = +25°C)	I_{SC}	—	80	100	—	80	100	mA
Output Noise Voltage (10 Hz ≤ f ≤ 10 kHz, T_J = +25°C)	V_N	—	40	200	—	40	200	μV_{rms}
Long Term Stability (T_J = +125°C) (Note 5)	S	—	20	50	—	20	50	mV/khr
OSCILLATOR SECTION (Note 6, unless otherwise specified)								
Initial Accuracy (T_J = +25°C)	—	—	±2.0	±6 0	—	±2 0	±6 0	%
Frequency Stability with Voltage (+8.0 V ≤ V_{CC} ≤ +35 V)	$\frac{\Delta f_{osc}}{\Delta V_{CC}}$	—	±0 3	±1 0	—	±1 0	±2 0	%
Frequency Stability with Temperature	$\frac{\Delta f_{osc}}{\Delta T}$	—	±3 0	±6 0	—	±3 0	±6 0	%
Minimum Frequency (R_T = 150 kΩ, C_T = 0 1 μF)	f_{min}	—	—	100	—	—	100	Hz
Maximum Frequency (R_T = 2 0 kΩ, C_T = 1.0 nF)	f_{max}	400	—	—	400	—	—	kHz
Current Mirror (I_{RT} = 2 0 mA)	—	1 7	2 0	2 2	1 7	2 0	2 2	mA
Clock Amplitude	—	3.0	3 5	—	3 0	3 5	—	V
Clock Width (T_J = +25°C)	—	0 3	0 5	1.0	0 3	0 5	1 0	μs
Sync Threshold	—	1 2	2 0	2.8	1 2	2 0	2 8	V
Sync Input Current (Sync Voltage = +3 5 V)	—	—	1 0	2.5	—	1 0	2 5	mA
ERROR AMPLIFIER SECTION (V_{CM} = +5.1 V)								
Input Offset Voltage	V_{IO}	—	0 5	5 0	—	2.0	10	mV
Input Bias Current	I_{IB}	—	1 0	10	—	1 0	10	μA
Input Offset Current	I_{IO}	—	—	1 0	—	—	1 0	μA
DC Open Loop Gain (R_L ≥ 10 MΩ)	A_{VOL}	60	75	—	60	75	—	dB
Gain Bandwidth Product (A_{VOL} = 0 dB, T_J = +25°C)	GBW	1 0	2 0	—	1 0	2.0	—	MHz
Low Level Output Voltage	V_{OL}	—	0.2	0.5	—	0 2	0 5	V
High Level Output Voltage	V_{OH}	3.8	5.6	—	3 8	5 6	—	V
Common Mode Rejection Ratio (+1 5 V ≤ V_{CM} ≤ +5.2 V)	CMRR	60	75	—	60	75	—	dB
Power Supply Rejection Ratio (+8.0 V ≤ V_{CC} ≤ +35 V)	PSRR	50	60	—	50	60	—	dB
PWM COMPARATOR SECTION								
Minimum Duty Cycle	DC_{min}	—	—	0	—	—	0	%
Maximum Duty Cycle	DC_{max}	45	49	—	45	49	—	%
Input Threshold, Zero Duty Cycle (Note 6)	V_{TH}	0.6	0.9	—	0.6	0.9	—	V
Input Threshold, Maximum Duty Cycle (Note 6)	V_{TH}	—	3.3	3.6	—	3.3	3 6	V
Input Bias Current	I_{IB}	—	0.05	1.0	—	0.05	1 0	μA

ELECTRICAL CHARACTERISTICS (Continued)

		SG1525A/2525A SG1527A/2527A			SG3525A SG3527A			
Characteristic	**Symbol**	**Min**	**Typ**	**Max**	**Min**	**Typ**	**Max**	**Unit**
SOFT-START SECTION								
Soft-Start Current ($V_{shutdown}$ = 0 V)	—	25	50	80	25	50	80	μA
Soft-Start Voltage ($V_{shutdown}$ = 2.0 V)	—	—	0.4	0.6	—	0.4	0.6	V
Shutdown Input Current ($V_{shutdown}$ = 2.5 V)	—	—	0.4	1.0	—	0.4	1.0	mA
OUTPUT DRIVERS (Each Output, V_C = +20 V)								
Output Low Level	V_{OL}							V
(I_{sink} = 20 mA)		—	0.2	0.4	—	0.2	0.4	
(I_{sink} = 100 mA)		—	1.0	2.0	—	1.0	2.0	
Output High Level	V_{OH}							V
(I_{source} = 20 mA)		18	19	—	18	19	—	
(I_{source} = 100 mA)		17	18	—	17	18	—	
Under Voltage Lockout (V8 and V9 = High)	V_{UL}	6.0	7.0	8.0	6.0	7.0	8.0	V
Collector Leakage, V_C = +35 V (Note 7)	$I_{C(leak)}$	—	—	200	—	—	200	μA
Rise Time (C_L = 1.0 nF, T_J = 25°C)	t_r	—	100	600	—	100	600	ns
Fall Time (C_L = 1.0 nF, T_J = 25°C)	t_f	—	50	300	—	50	300	ns
Shutdown Delay (V_{SD} = +3.0 V, C_S = 0, T_J = +25°C)	t_{ds}	—	0.2	0.5	—	0.2	0.5	μs
Supply Current, V_{CC} = +35 V	I_{CC}	—	14	20	—	14	20	mA

NOTES.

4 T_{low} = −55°C for SG1525A/1527A
−40°C for SG2525A/2527A
0°C for SG3525A/3527A

T_{high} = +125°C for SG1525A/1527A
+85°C for SG2525A/2527A
+70°C for SG3525A/3527A

5 Since long term stability cannot be measured on each device before shipment, this specification is an engineering estimate of average stability from lot to lot

6 Tested at f_{osc} = 40 kHz (R_T = 3.6 kΩ, C_T = 0.01 μF, R_D = 0 Ω)

7 Applies to SG1525A/2525A/3525A only, due to polarity of output pulses

APPLICATION INFORMATION

Shutdown Options (see block diagram, front page)

1. An external open collector comparator or transistor can be used to pull down the Compensation pin (9). This will set the PWM latch and turn off both outputs. Pulse-by-pulse protection can be accomplished if the shutdown signal is momentary, since the PWM latch will be reset with each clock pulse.

2. Shutdown can also be accomplished by pulling down on the SOFT-START pin (8). When using this approach, shutdown will not affect the amplifier compensation network; however, if a SOFT-START capacitor is used, it must be discharged, possible slowing shutdown response.

3. Applying a positive-going signal to the Shutdown pin (10) will provide the most rapid shutdown of the outputs if a soft-start capacitor is not used at Pin 8. An external soft-start capacitor at Pin 8 will slow shutdown response due to the discharge time of the soft-start capacitor. Dishcarge current is approximately twice the charging current.

4. The Shutdown terminal can be used to set the PWM latch on a pulse-by-pulse basis if there is no external capacitance on Pin 8. Soft-start characteristics may still be accomplished by applying an external capacitor, blocking diode and charging resistor to the Compensation pin (9).

TYPICAL CHARACTERISTICS

FIGURE 1 — SG1525A OSCILLATOR SCHEMATIC

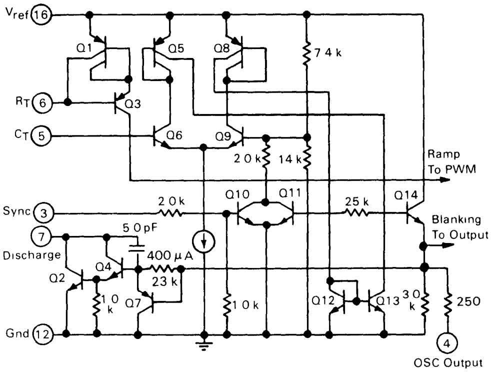


FIGURE 2 — OSCILLATOR CHARGE TIME versus R_T

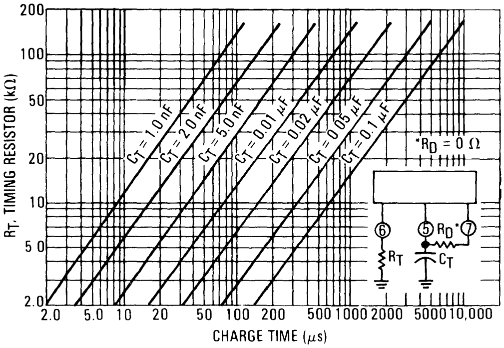


FIGURE 3 — OSCILLATOR DISCHARGE TIME versus R_D

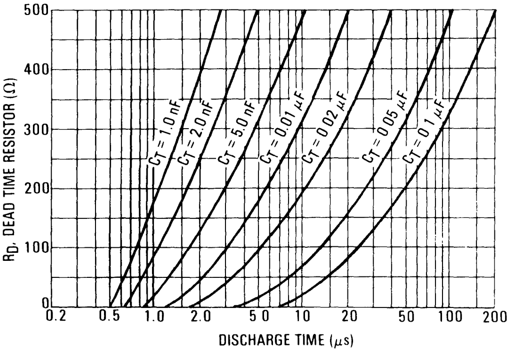


FIGURE 4 — SG1525A ERROR AMPLIFIER SCHEMATIC

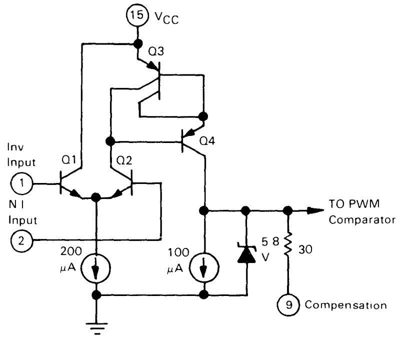


FIGURE 5 — ERROR AMPLIFIER OPEN-LOOP FREQUENCY RESPONSE

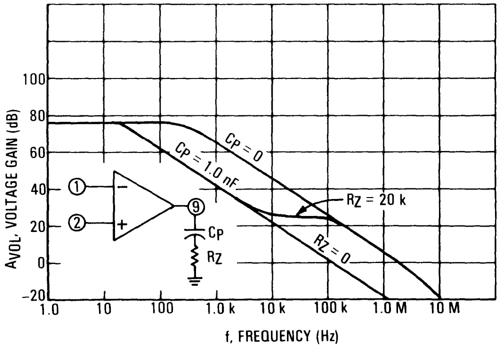


FIGURE 6 — SG1525A OUTPUT CIRCUIT
(1/2 CIRCUIT SHOWN)

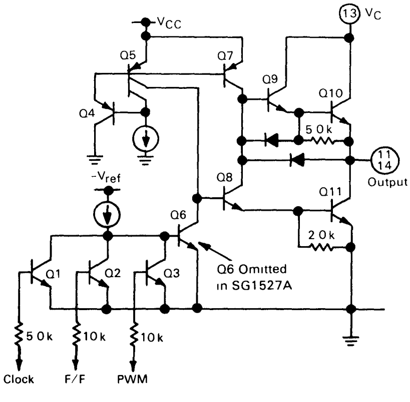


**FIGURE 7 — SG1525A/2525A/3525A
OUTPUT SATURATION CHARACTERISTICS**

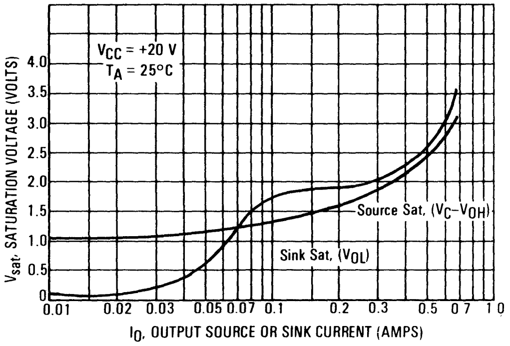
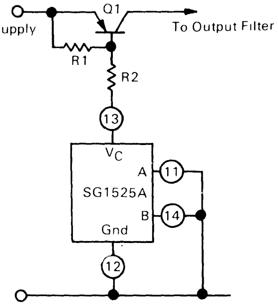
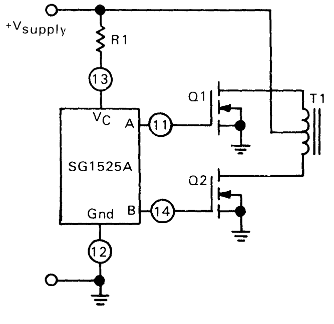


FIGURE 8 — SINGLE ENDED SUPPLY



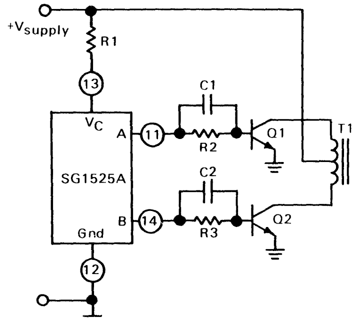
For single-ended supplies, the driver outputs are grounded. The V_C terminal is switched to ground by the totem-pole source transistors on alternate oscillator cycles.

FIGURE 10 — DRIVING POWER FETS



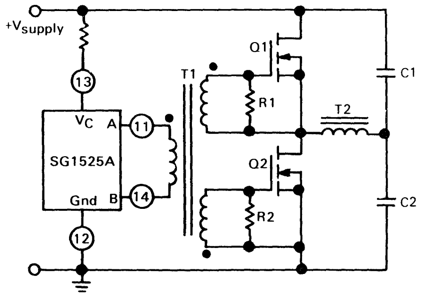
The low source impedance of the output drivers provides rapid charging of power FET input capacitance while minimizing external components.

FIGURE 9 — PUSH-PULL CONFIGURATION



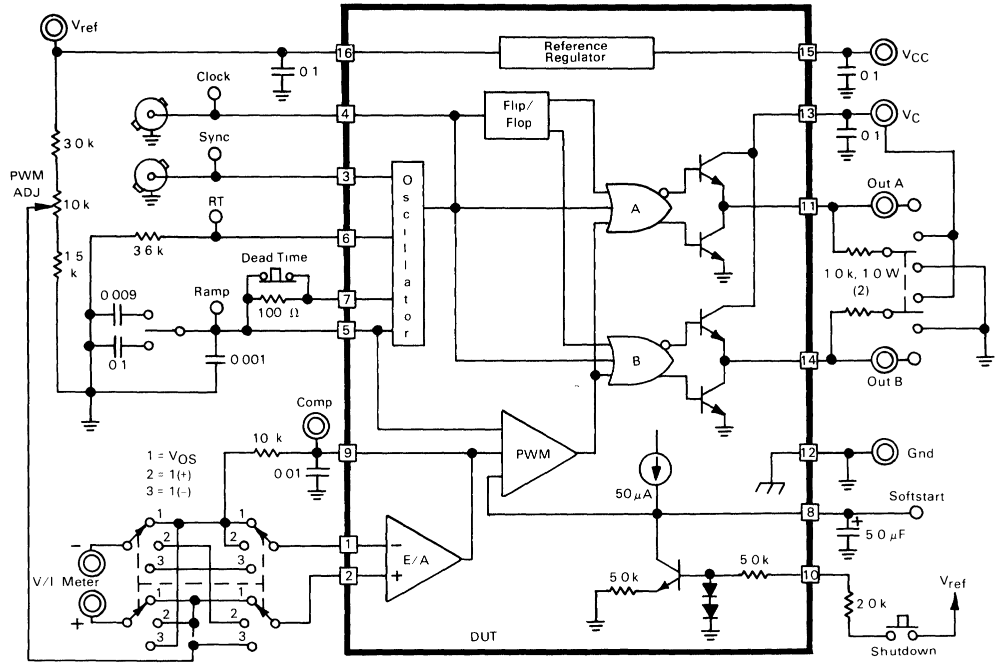
In conventional push-pull bipolar designs, forward base drive is controlled by R1–R3. Rapid turn-off times for the power devices are achieved with speed-up capacitors C1 and C2.

FIGURE 11 — DRIVING TRANSFORMERS IN A HALF-BRIDGE CONFIGURATION



Low power transformers can be driven directly by the SG1525A. Automatic reset occurs during deadtime, when both ends of the primary winding are switched to ground.

FIGURE 12 — LAB TEST FIXTURE



MOTOROLA

SG1526
SG2526
SG3526

PULSE WIDTH MODULATION CONTROL CIRCUIT

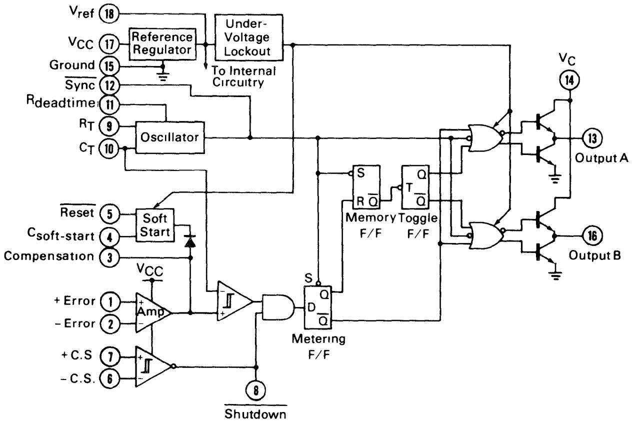
The SG1526 is a high performance pulse width modulator integrated circuit intended for fixed frequency switching regulators and other power control applications.

Functions included in this IC are a temperature compensated voltage reference, sawtooth oscillator, error amplifier, pulse width modulator, pulse metering and steering logic, and two high current totem pole outputs ideally suited for driving the capacitance of power FETs at high speeds.

Additional protective features include soft-start and undervoltage lockout, digital current limiting, double pulse inhibit, adjustable dead time and a data latch for single pulse metering. All digital control ports are TTL and B-series CMOS compatible. Active low logic design allows easy wired-OR connections for maximum flexibility. The versatility of this device enables implementation in single-ended or push-pull switching regulators that are transformerless or transformer coupled. The SG1526 is specified over the full military junction temperature range of -55°C to +150°C. The SG2526 is specified over a junction temperature range of -40°C to +150°C while the SG3526 is specified over a range of 0°C to +125°C.

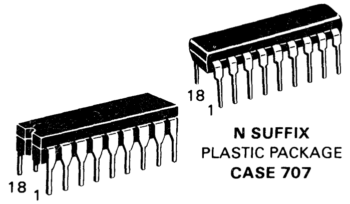
- 8.0 to 35 Volt Operation
- 5.0 Volt ±1% Trimmed Reference
- 1.0 Hz to 400 kHz Oscillator Range
- Dual Source/Sink Current Outputs: ±100 mA
- Digital Current Limiting
- Programmable Dead Time
- Undervoltage Lockout
- Single Pulse Metering
- Programmable Soft-Start
- Wide Current Limit Common Mode Range
- Guaranteed 6 Unit Synchronization

BLOCK DIAGRAM



PULSE WIDTH MODULATION CONTROL CIRCUITS

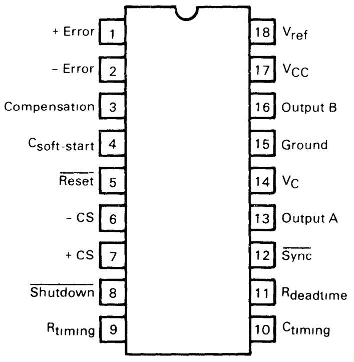
SILICON MONOLITHIC INTEGRATED CIRCUITS



N SUFFIX
PLASTIC PACKAGE
CASE 707

J SUFFIX
CERAMIC PACKAGE
CASE 726

PIN CONNECTIONS



Top View

ORDERING INFORMATION

Device	Junction Temperature Range	Package
SG1526J	-55 to +150°C	Ceramic DIP
SG2526J SG2526N	-40 to +150°C	Ceramic DIP Plastic DIP
SG3526J SG3526N	0 to +125°C	Ceramic DIP Plastic DIP

MAXIMUM RATINGS (Note 1)

Rating	Symbol	Value	Unit
Supply Voltage	V_{CC}	+40	Vdc
Collector Supply Voltage	V_C	+40	Vdc
Logic Inputs	—	-0.3 to +5.5	V
Analog Inputs	—	-0.3 to V_{CC}	V
Output Current, Source or Sink	I_O	±200	mA
Reference Output Current	I_{ref}	50	mA
Logic Sink Current	—	15	mA
Power Dissipation (Plastic & Ceramic Package) Note 2, $T_A = +25°C$ Note 3, $T_C = +25°C$	P_D	 1000 3000	mW
Thermal Resistance Junction to Air (Plastic and Ceramic Package)	$R_{\theta JA}$	100	°C/W
Thermal Resistance Junction to Case (Plastic and Ceramic Package)	$R_{\theta JC}$	42	°C/W
Operating Junction Temperature	T_J	+150	°C
Storage Temperature Range	T_{stg}	-65 to +150	°C
Lead Temperature (Soldering, 10 Seconds)	T_{Solder}	±300	°C

Notes:
1 Values beyond which damage may occur
2. Derate at 10 mW/°C for ambient temperatures above +50°C
3 Derate at 24 mW/°C for case temperatures above +25°C

RECOMMENDED OPERATING CONDITIONS

Characteristic	Symbol	Min	Max	Unit
Supply Voltage	V_{CC}	+8.0	+35	Vdc
Collector Supply Voltage	V_C	+4.5	+35	Vdc
Output Sink/Source Current (Each Output)	I_O	0	±100	mA
Reference Load Current	I_{ref}	0	20	mA
Oscillator Frequency Range	f_{osc}	0.001	400	kHz
Oscillator Timing Resistor	R_T	2.0	150	kΩ
Oscillator Timing Capacitor	C_T	0.001	20	μF
Available Deadtime Range (40 kHz)		3.0	50	%
Operating Junction Temperature Range SG1526 SG2526 SG3526	T_J	 -55 -40 0	 +150 +150 +125	°C

ELECTRICAL CHARACTERISTICS (V_{CC} = +15 Vdc, T_J = T_{low} to T_{high} [Note 4] unless otherwise specified)

Characteristic	Symbol	SG1526/2526			SG3526			Unit
		Min	Typ	Max	Min	Typ	Max	
REFERENCE SECTION (Note 5)								
Reference Output Voltage (T_J = +25°C)	V_{ref}	4.95	5.00	5.05	4.90	5.00	5.10	V
Line Regulation ($+8.0\ V \leqslant V_{CC} \leqslant +35\ V$)	Reg_{line}	—	10	20	—	10	30	mV
Load Regulation, $0\ mA \leqslant I_L \leqslant 20\ mA$	Reg_{load}	—	10	30	—	10	50	mV
Temperature Stability	$\Delta V_{ref}/\Delta T_J$	—	15	50	—	15	50	mV
Total Reference Output Voltage Variation ($+8.0\ V \leqslant V_{CC} \leqslant +35\ V$, $0\ mA \leqslant I_L \leqslant 20\ mA$)	ΔV_{ref}	4.90	5.00	5.10	4.85	5.00	5.15	V
Short Circuit Current (V_{ref} = 0 V)	I_{SC}	25	50	100	25	50	100	mA
UNDERVOLTAGE LOCKOUT								
$\overline{\text{Reset}}$ Output Voltage (V_{ref} = +3.8 V)	—	—	0.2	0.4	—	0.2	0.4	V
$\overline{\text{Reset}}$ Output Voltage (V_{ref} = +4.8 V)	—	2.4	4.8	—	2.4	4.8	—	V
OSCILLATOR SECTION (Note 6)								
Initial Accuracy (T_J = +25°C)	—	—	±3.0	±8.0	—	±3.0	±8.0	%
Frequency Stability over Power Supply Range ($+8.0\ V \leqslant V_{CC} \leqslant +35\ V$)	$\frac{\Delta f_{osc}}{\Delta V_{CC}}$	—	0.5	1.0	—	0.5	1.0	%
Frequency Stability over Temperature (ΔT_J = T_{low} to T_{high})	$\frac{\Delta f_{osc}}{\Delta T_J}$	—	7.0	10	—	3.0	5.0	%
Minimum Frequency (R_T = 150 k Ω, C_T = 20 μF)	f_{min}	—	—	1.0	—	—	1.0	Hz
Maximum Frequency (R_T = 2.0 kΩ, C_T = 0.001 μF)	f_{max}	400	—	—	400	—	—	kHz
Sawtooth Peak Voltage (V_{CC} = +35 V)	$V_{osc(P)}$	—	3.0	3.5	—	3.0	3.5	V
Sawtooth Valley Voltage (V_{CC} = +8.0 V)	$V_{osc(V)}$	0.5	1.0	—	0.5	1.0	—	V
ERROR AMPLIFIER SECTION (Note 7)								
Input Offset Voltage ($R_S \leqslant 2.0\ k\Omega$)	V_{IO}	—	2.0	5.0	—	2.0	10	mV
Input Bias Current	I_{IB}	—	−350	−1000	—	−350	−2000	nA
Input Offset Current	I_{IO}	—	35	100	—	35	200	nA
DC Open Loop Gain ($R_L \geqslant 10\ M\Omega$)	A_{Vol}	64	72	—	60	72	—	dB
High Output Voltage ($V_{Pin\ 1} - V_{Pin\ 2} \geqslant +150\ mV$, I_{source} = 100 μA)	V_{OH}	3.6	4.2	—	3.6	4.2	—	V
Low Output Voltage ($V_{Pin\ 2} - V_{Pin\ 1} \geqslant +150\ mV$, I_{sink} = 100 μA)	V_{OL}	—	0.2	0.4	—	0.2	0.4	V
Common Mode Rejection Ratio ($R_S \leqslant 2.0\ k\Omega$)	CMRR	70	94	—	70	94	—	dB
Power Supply Rejection Ratio ($+12\ V \leqslant V_{CC} \leqslant +18\ V$)	PSRR	66	80	—	66	80	—	dB
PWM COMPARATOR SECTION (Note 6)								
Minimum Duty Cycle ($V_{compensation}$ = +0.4 V)	DC_{min}	—	—	0	—	—	0	%
Maximum Duty Cycle ($V_{compensation}$ = +3.6 V)	DC_{max}	45	49	—	45	49	—	%

ELECTRICAL CHARACTERISTICS (Continued)

Characteristic	Symbol	SG1526/2526			SG3526			Unit
		Min	Typ	Max	Min	Typ	Max	
DIGITAL PORTS ($\overline{\text{SYNC}}$, $\overline{\text{SHUTDOWN}}$, $\overline{\text{RESET}}$)								
Output Voltage — High Logic Level (I_{source} = 40 μA)	V_{OH}	2.4	4.0	—	2.4	4.0	—	V
Output Voltage — Low Logic Level (I_{sink} = 3.6 mA)	V_{OL}	—	0.2	0.4	—	0.2	0.4	V
Input Current — High Logic Level (V_{IH} = +2.4 V)	I_{IH}	—	-125	-200	—	-125	-200	μA
Input Current — Low Logic Level (V_{IL} = +0.4 V)	I_{IL}	—	-225	-360	—	-225	-360	μA
CURRENT LIMIT COMPARATOR SECTION (Note 8)								
Sense Voltage ($R_S \leq 50\ \Omega$)	V_{sense}	90	100	110	80	100	120	mV
Input Bias Current	I_{IB}	—	-3.0	-10	—	-3.0	-10	μA
SOFT-START SECTION								
Error Clamp Voltage ($\overline{\text{Reset}}$ = +0.4 V)	—	—	0.1	0.4	—	0.1	0 4	V
$C_{Soft\text{-}Start}$ Charging Current ($\overline{\text{Reset}}$ = +2.4 V)	I_{CS}	50	100	150	50	100	150	μA
OUTPUT DRIVERS (Each Output, V_C = +15 Vdc unless otherwise specified)								
Output High Level	V_{OH}							V
I_{source} = 20 mA		12.5	13.5	—	12.5	13.5	—	
I_{source} = 100 mA		12	13	—	12	13	—	
Output Low Level	V_{OL}							V
I_{sink} = 20 mA		—	0.2	0.3	—	0.2	0.3	
I_{sink} = 100 mA		—	1.2	2.0	—	1.2	2.0	
Collector Leakage, V_C = +40 V	$I_{C(leak)}$	—	50	150	—	50	150	μA
Rise Time (C_L = 1000 pF)	t_r	—	0.3	0.6	—	0.3	0.6	μs
Fall Time (C_L = 1000 pF)	t_f	—	0.1	0.2	—	0.1	0.2	μs
Supply Current ($\overline{\text{Shutdown}}$ = +0.4 V, V_{CC} = +35 V, R_T = 4.12 kΩ)	I_{CC}	—	18	30	—	18	30	mA

Notes

4. T_{low} = -55°C for SG1526
   -40°C for SG2526
   0°C for SG3526
   T_{high} = +150°C for SG1526/2526
   +125°C for SG3526
5. I_L = 0 mA unless otherwise noted.
6. f_{osc} = 40 kHz (R_T = 4.12 kΩ ±1%, C_T = 0.01 μF ±1%, R_D = 0 Ω)
7. $0\ V \leq V_{CM} \leq +5.2\ V$
8. $0\ V \leq V_{CM} \leq +12\ V$

TYPICAL CHARACTERISTICS

FIGURE 1 — SG1526 REFERENCE STABILITY OVER TEMPERATURE

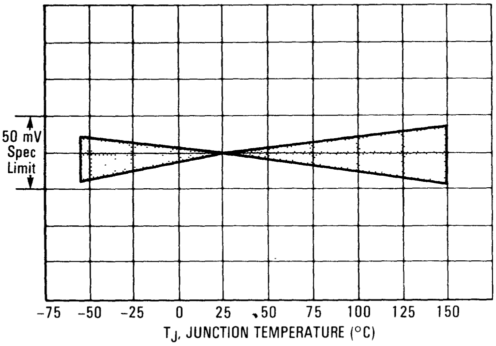


FIGURE 2 — REFERENCE VOLTAGE AS A FUNCTION SUPPLY VOLTAGE

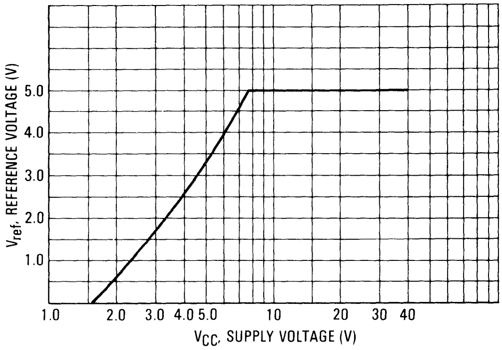


FIGURE 3 — ERROR AMPLIFIER OPEN LOOP FREQUENCY RESPONSE

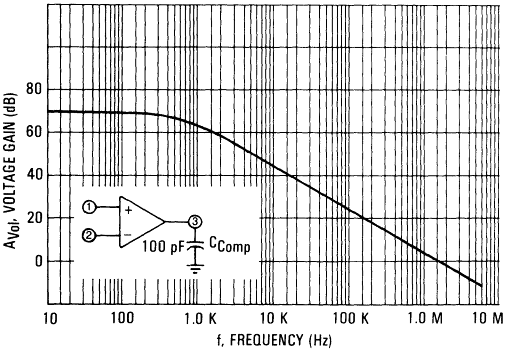


FIGURE 4 — CURRENT LIMIT COMPARATOR THRESHOLD

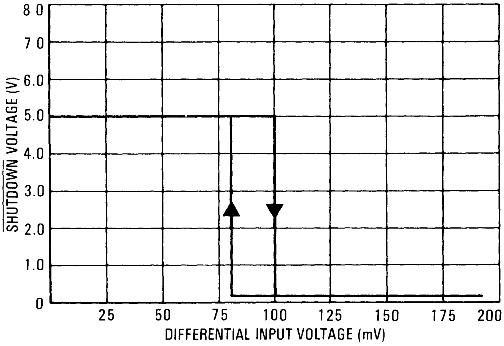


FIGURE 5 — UNDERVOLTAGE LOCKOUT CHARACTERISTIC

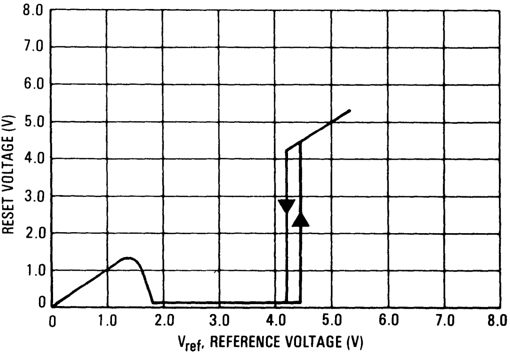


FIGURE 6 — OUTPUT DRIVER SATURATION VOLTAGE AS A FUNCTION OF SINK CURRENT

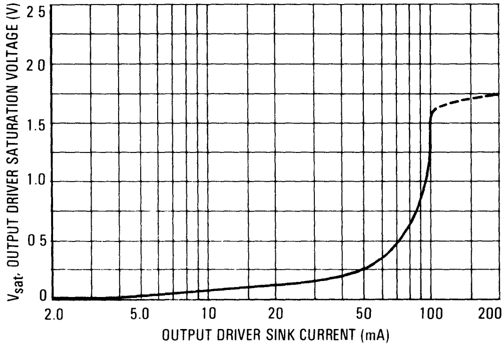


FIGURE 7 — V_C SATURATION VOLTAGE AS A FUNCTION OF SINK CURRENT

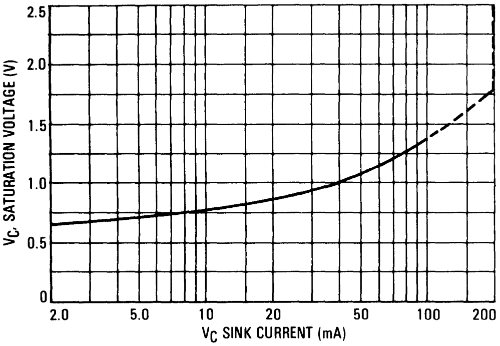


FIGURE 8 — SG1526 OSCILLATOR PERIOD

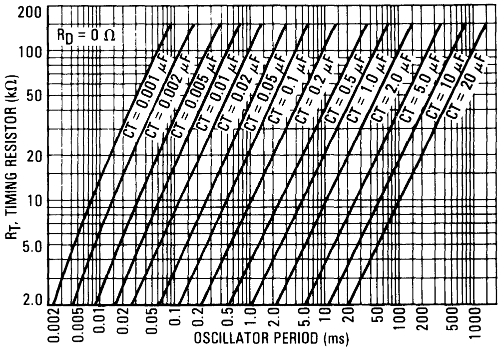


FIGURE 9 — SG1526 ERROR AMPLIFIER

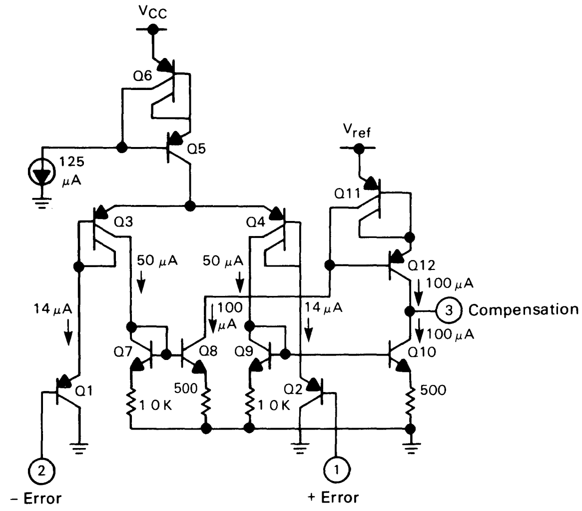


FIGURE 10 — SG1526 UNDERVOLTAGE LOCKOUT

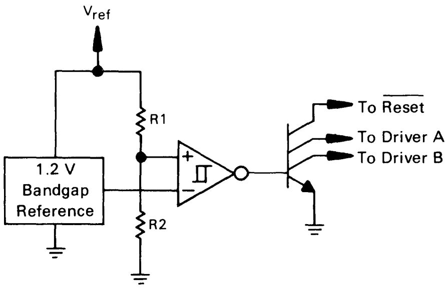
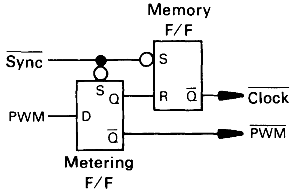


FIGURE 11 — SG1526 PULSE PROCESSING LOGIC

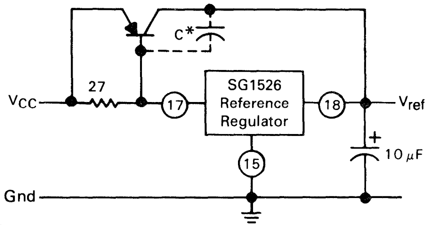


The metering FLIP-FLOP is an asynchronous data latch which suppresses high frequency oscillations by allowing only one PWM pulse per oscillator cycle.

The memory FLIP-FLOP prevents double pulsing in a push-pull configuration by remembering which output produced the last pulse.

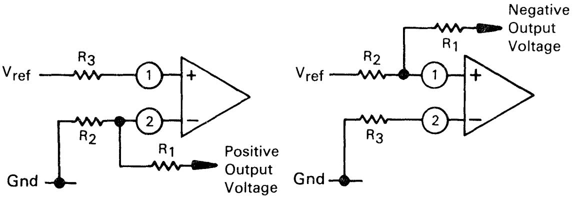
APPLICATIONS INFORMATION

FIGURE 12 — EXTENDING REFERENCE OUTPUT CURRENT CAPABILITY



*May be required with some types of transistors

FIGURE 13 — ERROR AMPLIFIER CONNECTIONS



$V_{out} = V_{ref}\left(\frac{R_1 + R_2}{R2}\right)$

$V_{out} = V_{ref}\left(\frac{R_1}{R2}\right)$

$R_3 = \left(\frac{R_1 R_2}{R_1 + R_2}\right)$

FIGURE 14 — OSCILLATOR CONNECTIONS

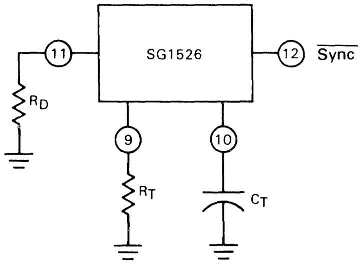
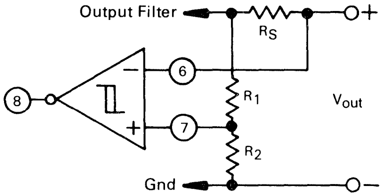


FIGURE 15 — FOLDBACK CURRENT LIMITING



$I_{max} = \frac{\left(0.1\text{ V} + \frac{V_{out} R_1}{R1 + R2}\right)}{R_S}$

$I_{SC} = \left(\frac{0.1\text{ V}}{R_S}\right)$

FIGURE 16 — SG1526 SOFT-START CIRCUITRY

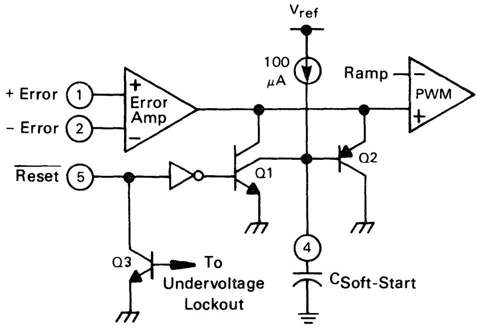
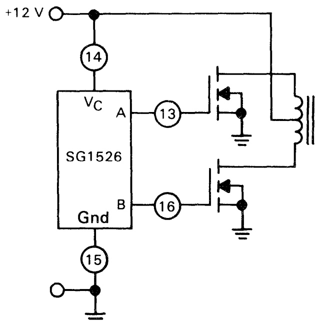


FIGURE 17 — DRIVING VMOS POWER FETS



The totem-pole output drivers of the SG1526 are ideally suited for driving the input capacitance of power FETs at high speeds.

FIGURE 18 — HALF-BRIDGE CONFIGURATION

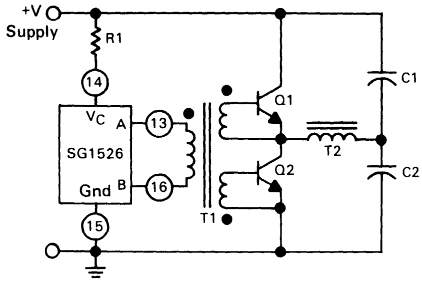
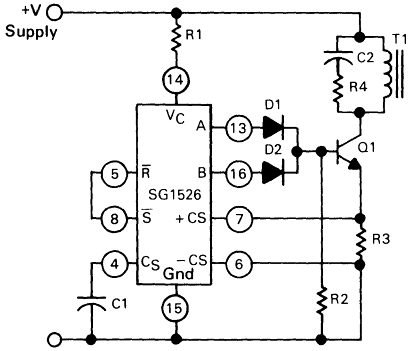


FIGURE 19 — FLYBACK CONVERTER WITH CURRENT LIMITING



In the above circuit, current limiting is accomplished by using the current limit comparator output to reset the soft-start capacitor.

FIGURE 20 — SINGLE-ENDED CONFIGURATION

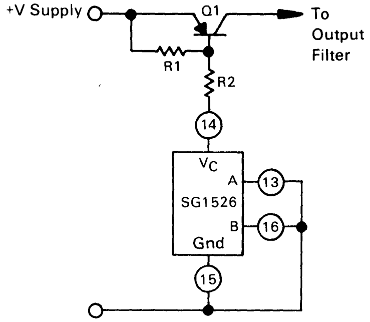
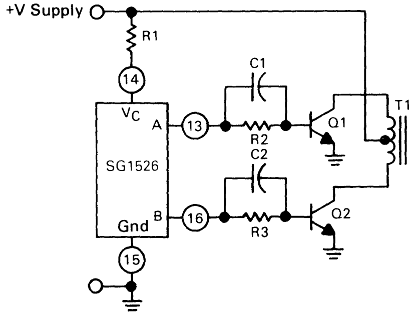


FIGURE 21 — PUSH-PULL CONFIGURATION



MOTOROLA

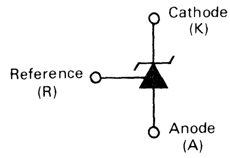
TL431
series

Specifications and Applications Information

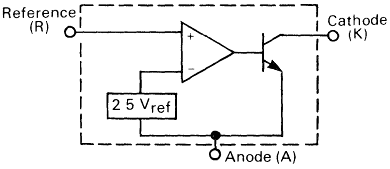
PROGRAMMABLE PRECISION REFERENCES

The TL431 integrated circuits are three-terminal programmable shunt regulator diodes. These monolithic IC voltage references operate as a low temperature coefficient zener which is programmable from V_{ref} to 36 volts with two external resistors. These devices exhibit a wide operating current range of 1.0 to 100 mA with a typical dynamic impedance of 0.22 Ω. The characteristics of these references make them excellent replacements for zener diodes in many applications such as digital voltmeters, power supplies, and op amp circuitry. The 2.5 volt reference makes it convenient to obtain a stable reference from 5.0 volt logic supplies, and since the TL431 operates as a shunt regulator, it can be used as either a positive or negative voltage reference.

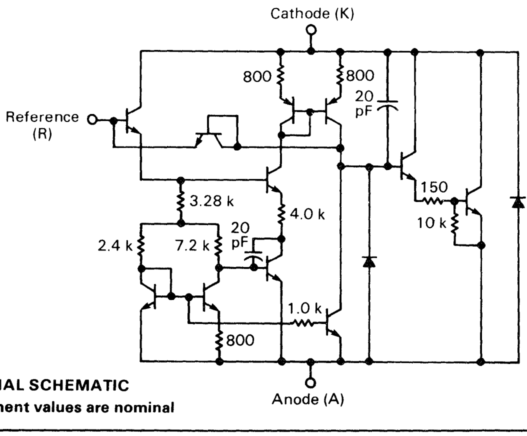
- Programmable Output Voltage to 36 Volts
- Low Dynamic Output Impedance, 0 22 Ω Typical
- Sink Current Capability of 1 0 to 100 mA.
- Equivalent Full-Range Temperature Coefficient of 50 ppm/°C Typical
- Temperature Compensated for Operation over Full Rated Operating Temperature Range
- Low Output Noise Voltage



SYMBOL



FUNCTIONAL BLOCK DIAGRAM



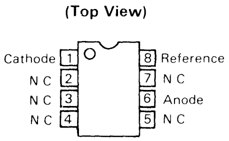
INTERNAL SCHEMATIC
Component values are nominal

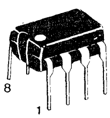
PROGRAMMABLE PRECISION REFERENCES

SILICON MONOLITHIC
INTEGRATED CIRCUITS

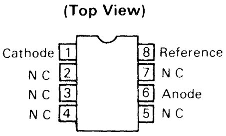
LP SUFFIX
PLASTIC PACKAGE
CASE 29
TO-92

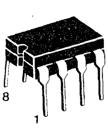
Pin 1 Reference
2 Anode
3 Cathode





P SUFFIX
PLASTIC DUAL-IN-LINE PACKAGE
CASE 626





JG SUFFIX
CERAMIC DUAL-IN-LINE PACKAGE
CASE 693

ORDERING INFORMATION

Device	Temperature Range	Package
TL431CLP	0 to +70°C	Plastic TO-92
TL431CP	0 to +70°C	Plastic DIP
TL431CJG	0 to +70°C	Ceramic DIP
TL431ILP	-40 to +85°C	Plastic TO-92
TL431IP	-40 to +85°C	Plastic DIP
TL431IJG	-40 to +85°C	Ceramic DIP
TL431MJG	-55 to +125°C	Ceramic DIP

MAXIMUM RATINGS (Full operating ambient temperature range applies unless otherwise noted.)

Rating	Symbol	Value	Unit
Cathode To Anode Voltage	V_{KA}	37	V
Cathode Current Range, Continuous	I_K	-100 to +150	mA
Reference Input Current Range, Continuous	I_{ref}	-0.05 to +10	mA
Operating Junction Temperature	T_J	150	°C
Operating Ambient Temperature Range TL431M TL431I TL431C	T_A	 -55 to +125 -40 to +85 0 to +70	°C
Storage Temperature Range	T_{stg}	-65 to +150	°C
Total Power Dissipation @ $T_A = 25°C$ Derate above 25°C Ambient Temperature LP Suffix Plastic Package P Suffix Plastic Package JG Suffix Ceramic Package	P_D	 0.775 1 10 1.25	W
Total Power Dissipation @ $T_C = 25°C$ Derate above 25°C Case Temperature LP Suffix Plastic Package P Suffix Plastic Package JG Suffix Ceramic Package	P_D	 1.5 3.0 3.3	W

THERMAL CHARACTERISTICS

Characteristics	Symbol	LP Suffix Package	P Suffix Package	JG Suffix Package	Unit
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	178	114	100	°C/W
Thermal Resistance, Junction to Case	$R_{\theta JC}$	83	41	38	°C/W

RECOMMENDED OPERATING CONDITIONS

Condition/Value	Symbol	Min	Max	Unit
Cathode To Anode Voltage	V_{KA}	V_{ref}	36	V
Cathode Current	I_K	1.0	100	mA

ELECTRICAL CHARACTERISTICS (Ambient temperature at 25°C unless otherwise noted)

Characteristic	Symbol	TL431M			TL431I			TL431C			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
Reference Input Voltage (Figure 1) $V_{KA} = V_{ref}$, $I_K = 10$ mA	V_{ref}	2.440	2.495	2 550	2.440	2.495	2.550	2.440	2.495	2.550	V
Reference Input Voltage Deviation Over Temperature Range. (Figure 1, Note 1) $V_{KA} = V_{ref}$, $I_K = 10$ mA	ΔV_{ref}	—	15	44	—	7.0	30	—	3 0	17	mV
Ratio of Change in Reference Input Voltage to Change in Cathode to Anode Voltage $I_K = 10$ mA (Figure 2), $\Delta V_{KA} = 10$ V to V_{ref} $\Delta V_{KA} = 36$ V to 10 V	$\frac{\Delta V_{ref}}{\Delta V_{KA}}$	— —	-1.4 -1.0	-2.7 -2.0	— —	-1.4 -1.0	-2.7 -2.0	— —	-1.4 -1.0	-2.7 -2.0	mV/V
Reference Input Current (Figure 2) $I_K = 10$ mA, R1 = 10 k, R2 = ∞	I_{ref}	—	1.8	4.0	—	1.8	4.0	—	1.8	4.0	μA
Reference Input Current Deviation Over Temperature Range. (Figure 2) $I_K = 10$ mA, R1 = 10 k, R2 = ∞	ΔI_{ref}	—	1.0	3.0	—	0.8	2.5	—	0.4	1.2	μA
Minimum Cathode Current For Regulation $V_{KA} = V_{ref}$ (Figure 1)	I_{min}	—	0.5	1.0	—	0.5	1.0	—	0 5	1.0	mA
Off-State Cathode Current (Figure 3) $V_{KA} = 36$ V, $V_{ref} = 0$ V	I_{off}	—	2.6	1000	—	2.6	1000	—	2.6	1000	nA
Dynamic Impedance (Figure 1, Note 2) $V_{KA} = V_{ref}$, $\Delta I_K = 1.0$ mA to 100 mA $f \leqslant 1.0$ kHz	$\lvert Z_{ka} \rvert$	—	0.22	0.5	—	0.22	0.5	—	0.22	0.5	Ω

FIGURE 1 — TEST CIRCUIT FOR $V_{KA} = V_{ref}$

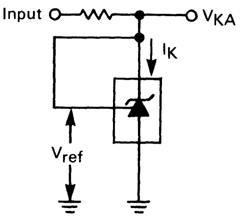


FIGURE 2 — TEST CIRCUIT FOR $V_{KA} > V_{ref}$

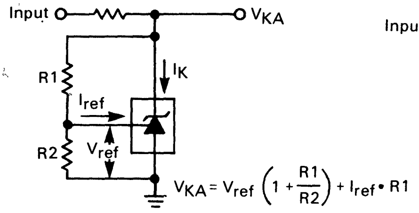
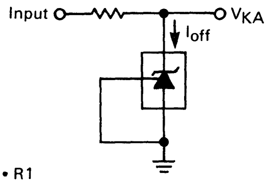
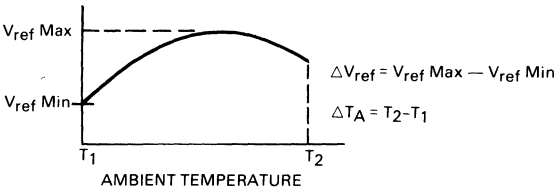


FIGURE 3 — TEST CIRCUIT FOR I_{off}



Note 1

The deviation parameter ΔV_{ref} is defined as the differences between the maximum and minimum values obtained over the full operating ambient temperature range that applies.



The average temperature coefficient of the reference input voltage, $\alpha\ V_{ref}$, is defined as:

$$\alpha\ V_{ref}\ \frac{ppm}{°C} = \frac{\left(\frac{\Delta V_{ref}}{V_{ref}\ @\ 25°C}\right)\times 10^6}{\Delta T_A} = \frac{\Delta V_{ref}\times 10^6}{\Delta T_A\ (V_{ref}\ @\ 25°C)}$$

αV_{ref} can be positive or negative depending on whether V_{ref} Min or V_{ref} Max occurs at the lower ambient temperature. (Refer to Figure 6)

Example: $\Delta V_{ref} = 8.0$ mV and slope is positive, V_{ref} @ 25°C = 2.495 V, $\Delta T_A = 70°C$

$$\alpha V_{ref} = \frac{0.008\times 10^6}{70\ (2.495)} = 45.8\ ppm/°C$$

Note 2

The dynamic impedance Z_{ka} is defined as:

$$|Z_{ka}| = \frac{\Delta V_{KA}}{\Delta I_K}$$

When the device is programmed with two external resistors, R1 and R2, (refer to Figure 2) the total dynamic impedance of the circuit is defined as:

$$|Z_{ka}'| \approx |Z_{ka}| \left(1 + \frac{R1}{R2}\right)$$

FIGURE 4 — CATHODE CURRENT versus CATHODE VOLTAGE

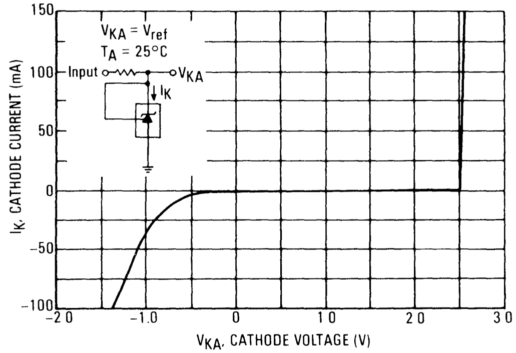


FIGURE 5 — CATHODE CURRENT versus CATHODE VOLTAGE

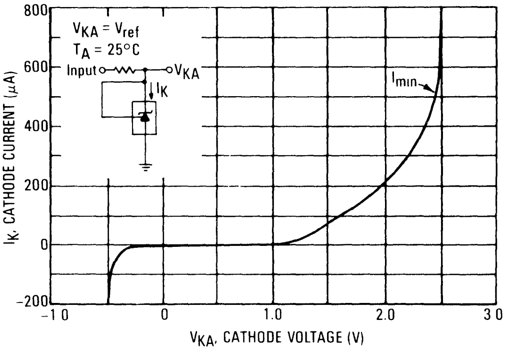


FIGURE 6 — REFERENCE INPUT VOLTAGE versus AMBIENT TEMPERATURE

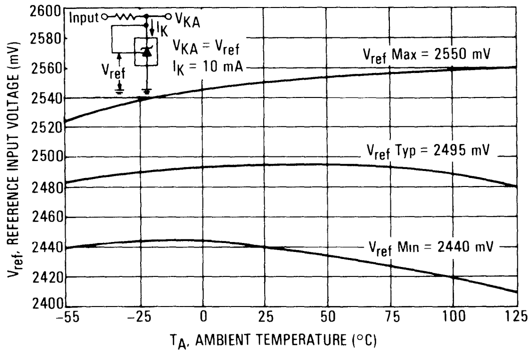


FIGURE 7 — REFERENCE INPUT CURRENT versus AMBIENT TEMPERATURE

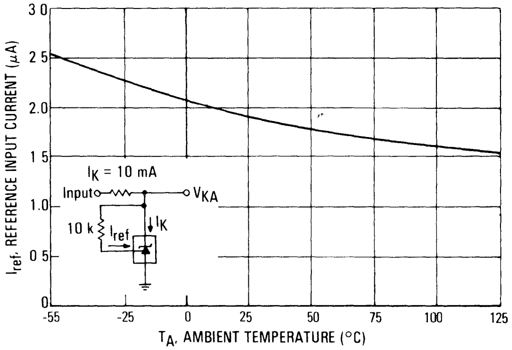


FIGURE 8 — CHANGE IN REFERENCE INPUT VOLTAGE versus CATHODE VOLTAGE

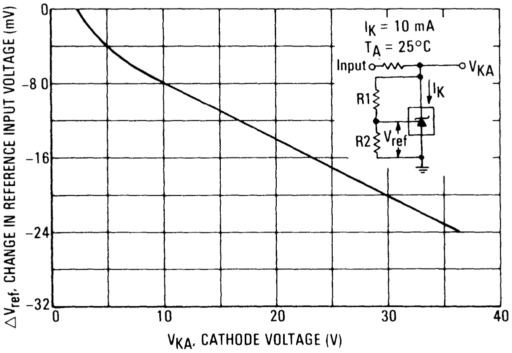


FIGURE 9 — OFF-STATE CATHODE CURRENT versus AMBIENT TEMPERATURE

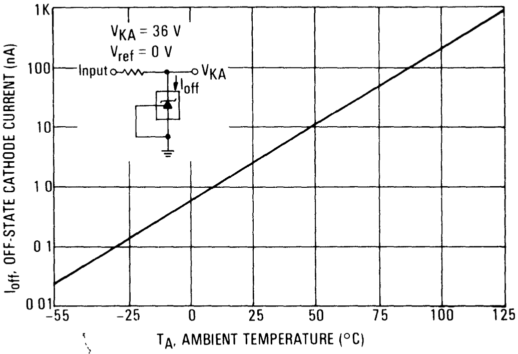


FIGURE 10 — DYNAMIC IMPEDANCE versus FREQUENCY

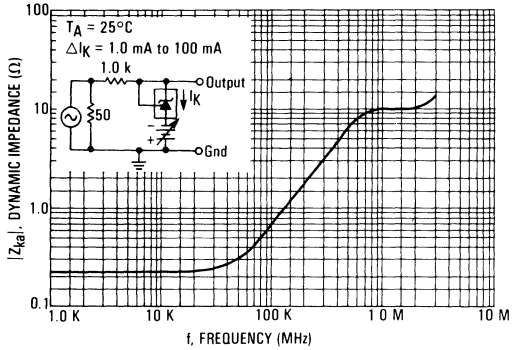


FIGURE 11 — DYNAMIC IMPEDANCE versus AMBIENT TEMPERATURE

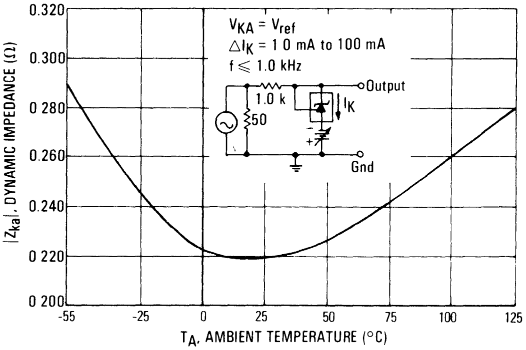


FIGURE 12 — OPEN LOOP VOLTAGE GAIN versus FREQUENCY

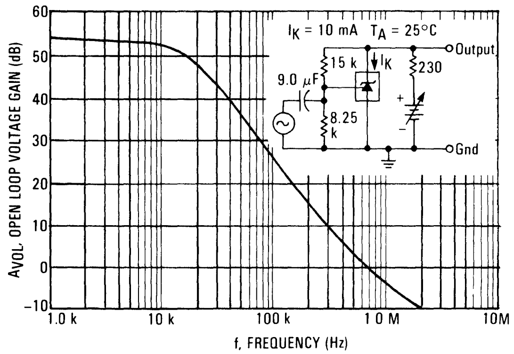


FIGURE 13 — SPECTRAL NOISE DENSITY

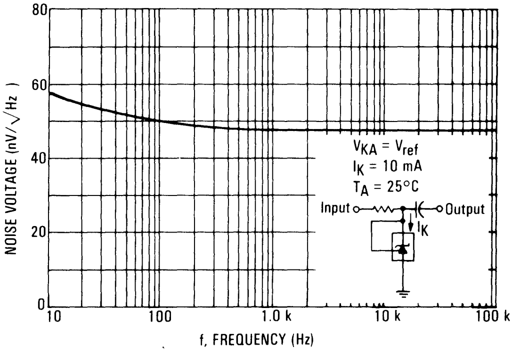


FIGURE 14 — PULSE RESPONSE

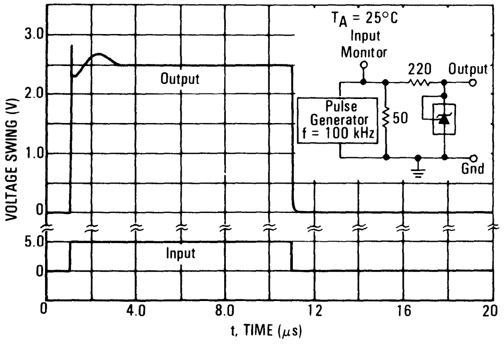


FIGURE 15 — STABILITY BOUNDARY CONDITIONS

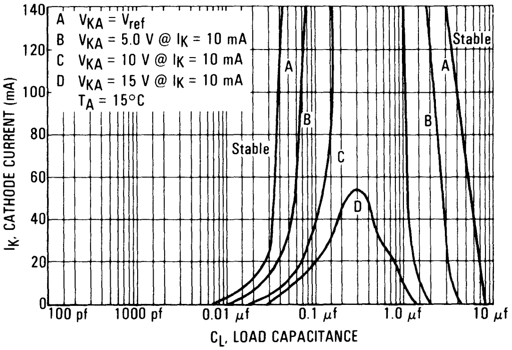


FIGURE 16 — TEST CIRCUIT FOR CURVE A OF STABILITY BOUNDARY CONDITIONS

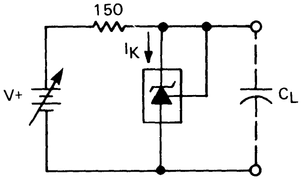
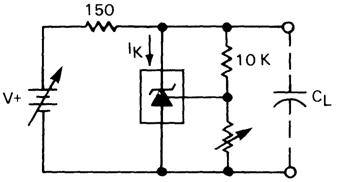
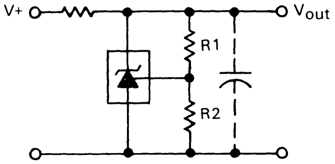


FIGURE 17 — TEST CIRCUIT FOR CURVES B, C, AND D OF STABILITY BOUNDARY CONDITIONS



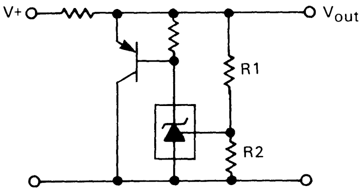
TYPICAL APPLICATIONS

FIGURE 18 — SHUNT REGULATOR



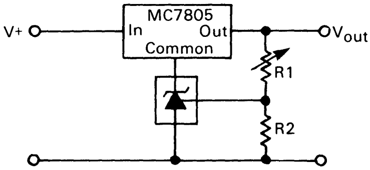
$$V_{out} = \left(1 + \frac{R1}{R2}\right) V_{ref}$$

FIGURE 19 — HIGH CURRENT SHUNT REGULATOR



$$V_{out} = \left(1 + \frac{R1}{R2}\right) V_{ref}$$

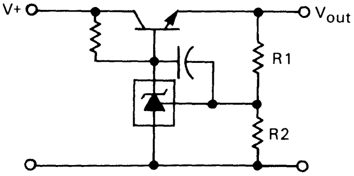
FIGURE 20 — OUTPUT CONTROL OF A THREE-TERMINAL FIXED REGULATOR



$$V_{out} = \left(1 + \frac{R1}{R2}\right) V_{ref}$$

V_{out} Min = V_{ref} + 5.0 V

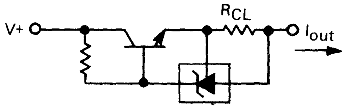
FIGURE 21 — SERIES PASS REGULATOR



$$V_{out} = \left(1 + \frac{R1}{R2}\right) V_{ref}$$

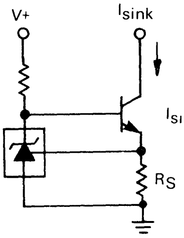
V_{out} Min = V_{ref} + 5.0 V

FIGURE 22 — CONSTANT CURRENT SOURCE



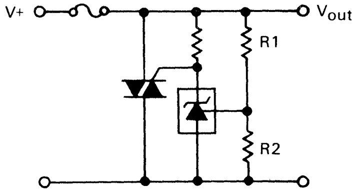
$$I_{out} = \frac{V_{ref}}{R_{CL}}$$

FIGURE 23 — CONSTANT CURRENT SINK



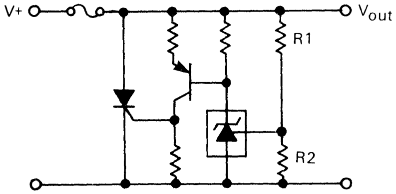
$$I_{sink} = \frac{V_{ref}}{R_S}$$

FIGURE 24 — TRIAC CROWBAR



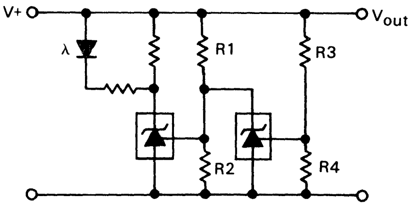
$$V_{out(trip)} = \left(1 + \frac{R1}{R2}\right)V_{ref}$$

FIGURE 25 — SCR CROWBAR



$$V_{out(trip)} = \left(1 + \frac{R1}{R2}\right)V_{ref}$$

FIGURE 26 — VOLTAGE MONITOR

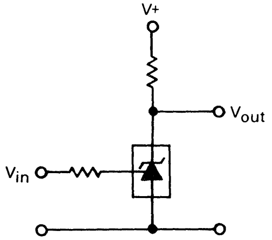


L.E.D. indicator is 'on' when V+ is between the upper and lower limits.

$$\text{Lower Limit} = \left(1 + \frac{R1}{R2}\right)V_{ref}$$

$$\text{Upper Limit} = \left(1 + \frac{R3}{R4}\right)V_{ref}$$

FIGURE 27 — SINGLE-SUPPLY COMPARATOR WITH TEMPERATURE-COMPENSATED THRESHOLD



$V_{th} = V_{ref}$

V_{in}	V_{out}
$<V_{ref}$	V+
$>V_{ref}$	≈ 2.0 V

FIGURE 28 — LINEAR OHMMETER

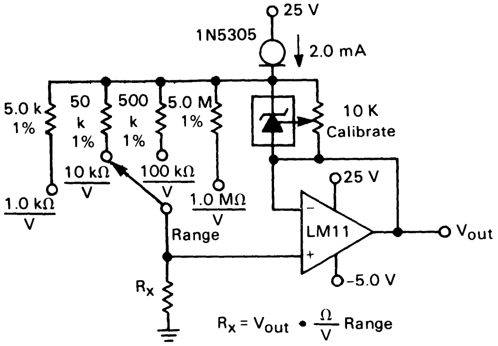


FIGURE 29 — SIMPLE 400 mW PHONO AMPLIFIER

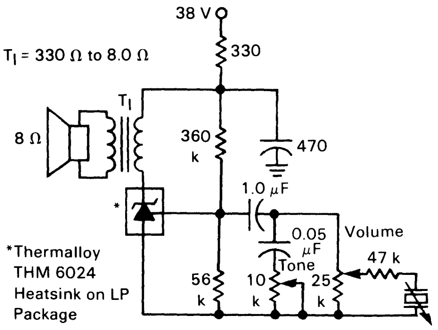
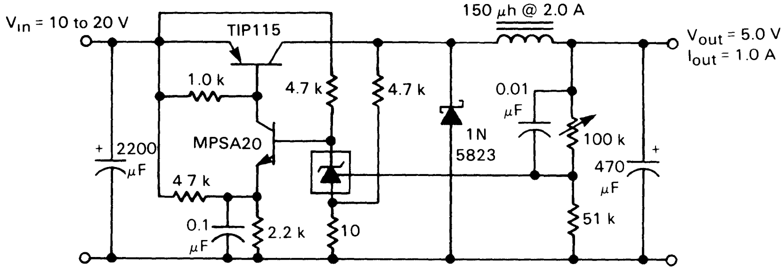


FIGURE 30 — HIGH EFFICIENCY STEP-DOWN SWITCHING CONVERTER



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 10 V to 20 V, I_o = 1.0 A	53 mV (1 1%)
Load Regulation	V_{in} = 15 V, I_o = 0A to 1.0 A	25 mV (0.5%)
Output Ripple	V_{in} = 10 V, I_o = 1.0 A	50 mV_{p-p} P.A.R.D.
Output Ripple	V_{in} = 20 V, I_o = 1.0 A	100 mV_{p-p} P.A.R.D.
Efficiency	V_{in} = 15 V, I_o = 1.0 A	82%

MOTOROLA

TL494
TL495

Specifications and Applications Information

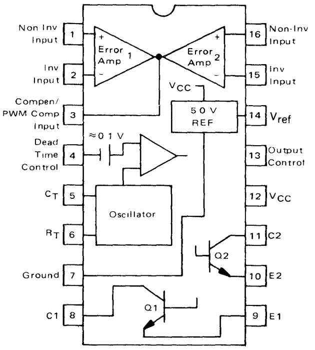
SWITCHMODE PULSE WIDTH MODULATION CONTROL CIRCUITS

The TL494 and TL495 are fixed frequency, pulse width modulation control circuits designed primarily for Switchmode power supply control. These devices feature:

- Complete Pulse Width Modulation Control Circuitry
- On-Chip Oscillator With Master Or Slave Operation
- On-Chip Error Amplifiers
- On-Chip 5 Volt Reference
- Adjustable Dead-Time Control
- Uncommitted Output Transistors For 200 mA Source Or Sink
- Output Control For Push-Pull Or Single-Ended Operation
- On-Chip 39 Volt Zener (TL495 Only)
- Output Steering Control (TL495 Only)

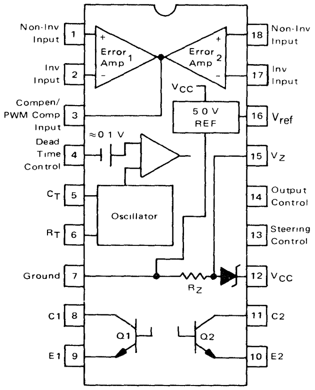
PIN CONNECTIONS

TL494



(top view)

TL495



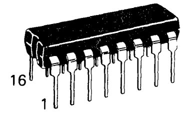
(top view)

The TL494C/495C are specified over the commercial operating range of 0°C to 70°C. The TL494I/495I are specified over the industrial range of −25°C to 85°C. The TL494M is specified over the full military range of −55°C to 125°C.

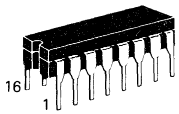
SWITCHMODE PULSE WIDTH MODULATION CONTROL CIRCUITS

SILICON MONOLITHIC INTEGRATED CIRCUITS

TL494

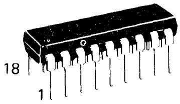


N SUFFIX
PLASTIC PACKAGE
CASE 648

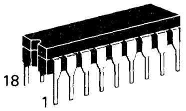


J SUFFIX
CERAMIC PACKAGE
CASE 620

TL495



N SUFFIX
PLASTIC PACKAGE
CASE 707



J SUFFIX
CERAMIC PACKAGE
CASE 726

ORDERING INFORMATION

Device	Temperature Range	Package
TL494CN	0 To 70°C	Plastic DIP
TL494CJ	0 To 70°C	Ceramic DIP
TL494IN	−25 To 85°C	Plastic DIP
TL494IJ	−25 To 85°C	Ceramic DIP
TL494MJ	−55 To 125°C	Ceramic DIP
TL495CN	0 To 70°C	Plastic DIP
TL495CJ	0 To 70°C	Ceramic DIP
TL495IN	−25 To 85°C	Plastic DIP
TL495IJ	−25 To 85°C	Ceramic DIP

FIGURE 1 — BLOCK DIAGRAM

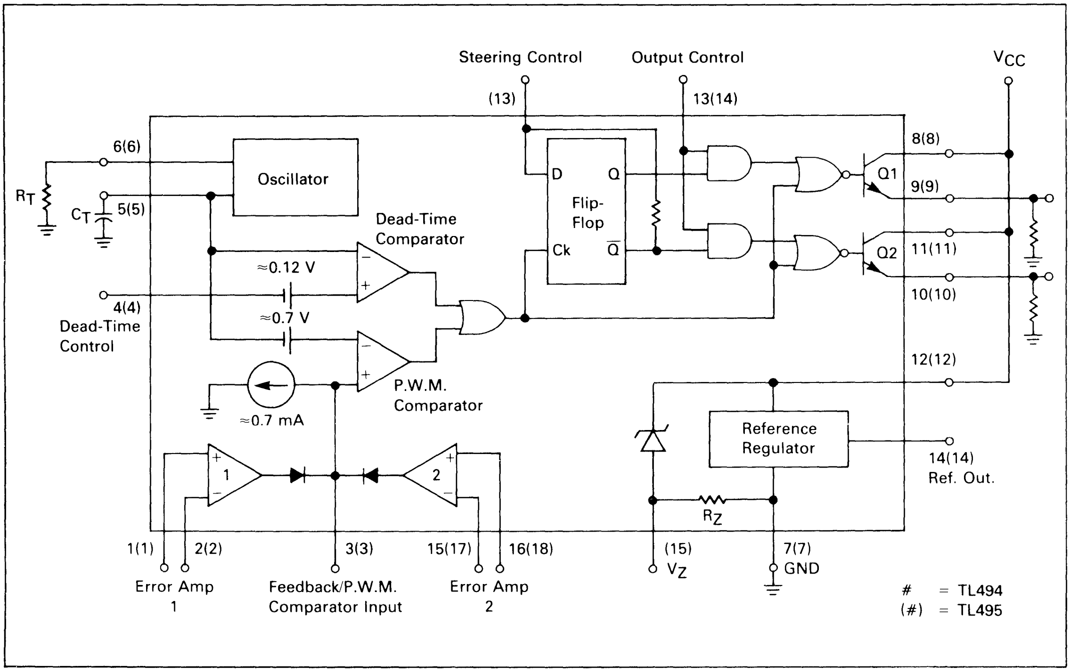
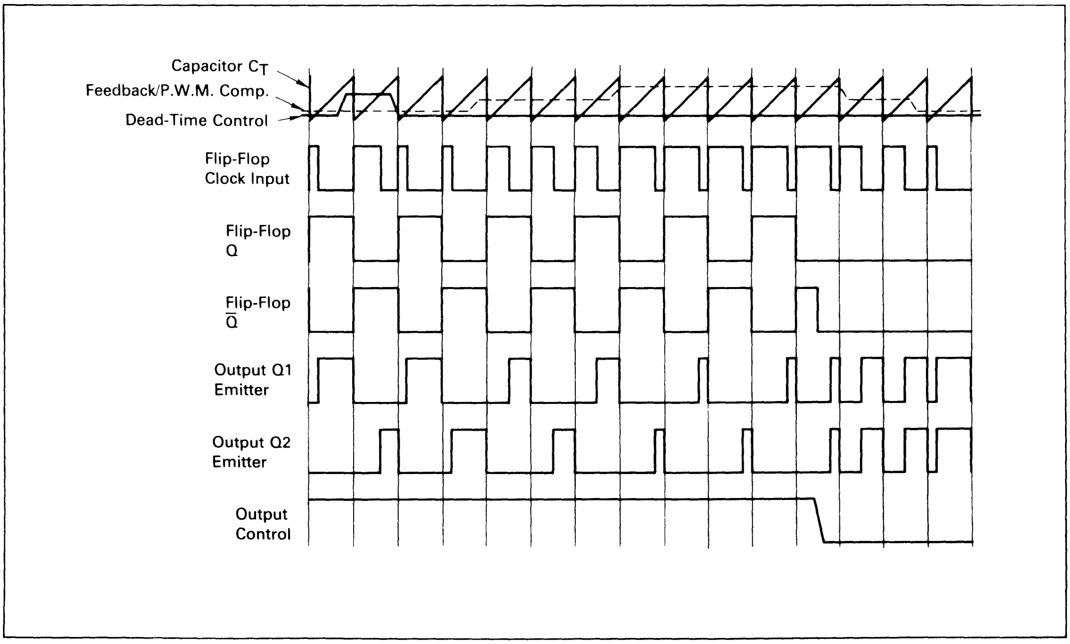


FIGURE 2 — TIMING DIAGRAM



Description

The TL494/495 are fixed-frequency pulse width modulation control circuit, incorporating the primary building blocks required for the control of a switching power supply. (See Figure 1.) An internal-linear sawtooth oscillator is frequency-programmable by two external components, R_T and C_T. The oscillator frequency is determined by:

$$f_{osc} \approx \frac{1.1}{R_T \bullet C_T}$$

Output pulse width modulation is accomplished by comparison of the positive sawtooth waveform across capacitor C_T to either of two control signals. The NOR gates, which drive output transistors Q1 and Q2, are enabled only when the flip-flop clock-input line is in its low state. This happens only during that portion of time when the sawtooth voltage is greater than the control signals. Therefore, an increase in control-signal amplitude causes a corresponding linear decrease of output pulse width. (Refer to the timing diagram shown in Figure 2.)

The control signals are external inputs that can be fed into the dead-time control, the error amplifier inputs, or the feedback input. The dead-time control comparator has an effective 120 mV input offset which limits the minimum output dead time to approximately the first 4% of the sawtooth-cycle time. This would result in a maximum duty cycle on a given output of 96% with the output control grounded, and 48% with it connected to the reference line. Additional dead time may be imposed on the output by setting the dead time-control input to a fixed voltage, ranging between 0 to 3.3 V.

The pulse width modulator comparator provides a means for the error amplifiers to adjust the output pulse width from the maximum percent on-time, established by the dead time control input, down to zero, as the voltage at the feedback pin varies from 0.5 to 3.5 V. Both error amplifiers have a common-mode input range from -0.3 V to ($V_{CC} - 2$ V), and may be used to sense power-supply output voltage and current. The error-amplifier outputs are active high and are ORed together at the non-inverting input of the pulse-width modulator comparator. With this configuration, the amplifier that demands minimum output on time, dominates control of the loop.

When capacitor C_T is discharged, a positive pulse is generated on the output of the dead-time comparator, which clocks the pulse-steering flip-flop and inhibits the output transistors, Q1 and Q2. With the output-control connected to the reference line, the pulse-steering flip-flop directs the modulated pulses to each of the two output transistors alternately for push-pull operation. The output frequency is equal to half that of the oscillator. Output drive can also be taken from Q1 or Q2, when single-ended operation with a maximum on-time of less than 50% is required. This is desirable when the output transformer has a ringback winding with a catch diode used for snubbing. When higher output-drive currents are required for single-ended operation, Q1 and Q2 may be connected in parallel, and the output-mode pin must be tied to ground to disable the flip-flop. The output frequency will now be equal to that of the oscillator.

The TL494/495 has an internal 5 V reference capable of sourcing up to 10 mA of load current for external bias circuits. The reference has an internal accuracy of ± 5% with a thermal drift of less than 50 mV over an operating temperature range of 0 to 70°C.

The TL495 contains an on-chip 39 volt zener diode for high voltage applications where V_{CC} is greater than 40 volts, and an output steering control that overrides the internal control of the pulse-steering flip-flop. (Refer to the functional table shown in figure 3.)

FIGURE 3 — FUNCTIONAL TABLE

Inputs		Output Function	$\frac{f_{out}}{f_{osc}} =$
Output Control	**Steering Control**		
Grounded	Open	Single-ended P.W.M. at Q1 and Q2	1
At V_{ref}	Open	Push-pull operation	0.5
At V_{ref}	V1 <0.4 V	Single-ended P.W.M. at Q1 only	1
At V_{ref}	V1 >2.4 V	Single-ended P.W.M. at Q2 only	1

MAXIMUM RATINGS (Full operating ambient temperature range applies unless otherwise noted)

Rating	Symbol	TL494M	TL494I/TL495I	TL494C/TL495C	Unit
Power Supply Voltage	V_{CC}	42	42	42	V
Collector Output Voltage	V_{C1}, V_{C2}	42	42	42	V
Collector Output Current (each transistor)	I_{C1}, I_{C2}	250	250	250	mA
Amplifier Input Voltage	V_{in}	V_{CC} + .03	V_{CC} + .03	V_{CC} + .03	V
Power Dissipation @ $T_A \leq 45°C$	P_D	1000	1000	1000	mW
Operating Junction Temperature	T_J	150	150	150	°C
Operating Ambient Temperature Range	T_A	−55 to 125	−25 to 85	0 to 70	°C
Storage Temperature Range	T_{stg}	−65 to 150	−65 to 150	−65 to 150	°C

THERMAL CHARACTERISTICS

Characteristics	Symbol	J Suffix Ceramic Package	N Suffix Plastic Package	Unit
Thermal Resistance, Junction to Ambient	$R_{\theta JA}$	100	80	°C/W
Power Derating Factor	$1/R_{\theta JA}$	10.0	12.5	mW/°C
Derating Ambient Temperature	T_A	50	45	°C

RECOMMENDED OPERATING CONDITIONS

Condition/Value	Symbol	TL494/TL495			Unit
		Min.	Typ.	Max.	
Power Supply Voltage	V_{CC}	7.0	15	40	V
Collector Output Voltage	V_{C1}, V_{C2}	—	30	40	V
Collector Output Current (each transistor)	I_{C1}, I_{C2}	—	—	200	mA
Amplifier Input Voltage	V_{in}	−0.3	—	V_{CC} − 2.0	V
Current Into Feedback Terminal	$I_{f.b.}$	—	—	0.3	mA
Reference Output Current	I_{ref}	—	—	10	mA
Timing Resistor	R_T	1.8	30	500	kΩ
Timing Capacitor	C_T	0.00047	0.001	10	μF
Oscillator Frequency	f_{osc}	1.0	40	200	kHz

ELECTRICAL CHARACTERISTICS (V_{CC} = 15 V, f_{osc} = 10 kHz unless otherwise noted.)
For typical values T_A = 25°C, for min/max values T_A is the operating ambient temperature range that applies unless otherwise noted.

Characteristic	Symbol	TL494M			TL494C, I/TL495C, I			Unit
		Min	Typ	Max	Min	Typ	Max	
REFERENCE SECTION								
Reference Voltage (I_O = 1.0 mA)	V_{ref}	4.75	5.0	5.25	4.75	5.0	5.25	V
Reference Voltage Change with Temperature (ΔT_A = Min to Max)	ΔV_{ref} (ΔT)	—	0.2	2.0	—	1.3	2.6	%
Input Regulation (V_{CC} = 7.0 V to 40 V)	Reg_{line}	—	2.0	25	—	2.0	25	mV
Output Regulation (I_O = 1.0 mA to 10 mA)	Reg_{load}	—	3.0	15	—	3.0	15	mV
Short-Circuit Output Current (V_{ref} = 0 V, T_A = 25°C)	I_{SC}	10	35	50	—	35	—	mA

ELECTRICAL CHARACTERISTICS (V_{CC} = 15 V, f_{osc} = 10 kHz unless otherwise noted.)
For typical values T_A = 25°C, for min/max values T_A is the operating ambient temperature range that applies unless otherwise noted.

Characteristic	Symbol	TL494M			TL494C, I/TL495C, I			Unit
		Min	Typ	Max	Min	Typ	Max	
OUTPUT SECTION								
Collector Off-State Current (V_{CC} = 40 V, V_{CE} = 40 V)	$I_{C(off)}$	—	2.0	100	—	2.0	100	μA
Emitter Off-State Current (V_{CC} = 40 V, V_C = 40 V, V_E = 0 V)	$I_{E(off)}$	—	—	−150	—	—	−100	μA
Collector-Emitter Saturation Voltage Common-Emitter (V_E = 0V, I_C = 200 mA)	$V_{sat(C)}$	—	1.1	1.5	—	1.1	1.3	V
Emitter-Follower (V_C = 15 V, I_E = −200 mA)	$V_{sat(E)}$	—	1.5	2.5	--	1.5	2.5	V
Output Control Pin Current Low State (V_{OC} ≤ 0.4 V)	I_{OCL}	—	10	—	—	10	—	μA
High State (V_{OC} = V_{ref})	I_{OCH}	—	0.2	3.5	—	0.2	3.5	mA
Output Voltage Rise Time (T_A = 25°C) Common-Emitter (See Figure 13)	t_r	—	100	200	—	100	200	ns
Emitter-Follower (See Figure 14)		—	100	200	—	100	200	ns
Output Voltage Fall Time (T_A = 25°C) Common-Emitter (See Figure 13)	t_f	—	25	100	—	25	100	ns
Emitter-Follower (See Figure 14)		—	40	100	—	40	100	ns

Characteristic	Symbol	TL494/TL495			Unit
		Min	Typ	Max	
ERROR AMPLIFIER SECTIONS					
Input Offset Voltage (V_O (Pin 3) = 2.5 V)	V_{IO}	—	2.0	10	mV
Input Offset Current (V_O (Pin 3) = 2.5 V)	I_{IO}	—	5.0	250	nA
Input Bias Current (V_O (Pin 3) = 2.5 V)	I_{IB}	—	0.1	1.0	μA
Input Common-Mode Voltage Range (V_{CC} = 7.0 V to 40 V)	V_{ICR}	−0.3	—	V_{CC} − 2.0	V
Open-Loop Voltage Gain (ΔV_O = 3.0 V, V_O = 0.5 to 3.5 V, R_L = 2.0 kΩ)	A_{VOL}	70	95	—	dB
Unity-Gain Crossover Frequency (V_O = 0.5 to 3.5 V, R_L = 2.0 kΩ)	f_C	—	350	—	kHz
Phase Margin at Unity-Gain (V_O = 0.5 to 3.5 V, R_L = 2.0 kΩ)	ϕm	—	65	—	deg.
Common-Mode Rejection Ratio (V_{CC} = 40 V)	CMRR	65	90		dB
Power Supply Rejection Ratio (ΔV_{CC} = 33 V, V_O = 2.5 V, R_L = 2.0 kΩ)	PSRR	—	100	—	dB
Output Sink Current (V_O (Pin 3) = 0.7 V)	I_O-	0.3	0.7	—	mA
Output Source Current (V_O (Pin 3) = 3.5 V)	I_O+	−2.0	−4.0	—	mA

ELECTRICAL CHARACTERISTICS (V_{CC} = 15 V, f_{osc} = 10 kHz unless otherwise noted.)
For typical values T_A = 25°C, for min/max values T_A is the operating ambient temperature range that applies unless otherwise noted.

Characteristic	Symbol	TL494/TL495			Unit
		Min	Typ	Max	
PWM COMPARATOR SECTION (Test Circuit Figure 12)					
Input Threshold Voltage (Zero duty cycle)	V_{TH}	—	3.5	4.5	V
Input Sink Current (V $_{(Pin\ 3)}$ = 0.7 V)	I_I –	0.3	0.7	—	mA
DEAD-TIME CONTROL SECTION (Test Circuit Figure 12)					
Input Bias Current (Pin 4) (V_{in} = 0 to 5.25 V)	$I_{IB\ (DT)}$	—	– 2.0	– 10	μA
Maximum Duty Cycle, Each Output, Push-Pull Mode (V_{in} = 0 V, C_T = 0.1 μF, R_T = 12 kΩ) (V_{in} = 0 V, C_T = 0.001 μF, R_T = 30 kΩ)	DC_{max}	 45 —	 48 45	 50 50	%
Input Threshold Voltage (Pin 4) (Zero Duty Cycle) (Maximum Duty Cycle)	V_{TH}	 — 0	 2.8 —	 3.3 —	V
OSCILLATOR SECTION					
Frequency (C_T = 0.001 μf, R_T = 30 kΩ)	f_{osc}	—	40	—	kHz
Standard Deviation of Frequency* (C_T = 0.001 μf, R_T = 30 kΩ)	σf_{osc}	—	3.0	—	%
Frequency Change with Voltage (V_{CC} = 7.0 V to 40 V, T_A = 25°C)	$\Delta f_{osc\ (\Delta V)}$	—	0.1	—	%
Frequency Change with Temperature (ΔT_A = 25°C to $T_{A\ low}$, 25°C to $T_{A\ high}$)	$\Delta f_{osc\ (\Delta T)}$	—	1.0	2.0	%

Characteristic	Symbol	TL495			Unit
		Min	Typ	Max	
STEERING CONTROL					
Input Current Low ($V_{(Pin\ 13)}$ = 0.4 V)	I_{STL}	—	– 25	– 200	μA
Input Current High ($V_{(Pin\ 13)}$ = 2.4 V) ($V_{(Pin\ 13)}$ = V_{ref})	I_{STH}	 — —	 25 75	 200 —	μA
ZENER CHARACTERISTICS					
Zener Breakdown Voltage (I_Z = 2mA)	V_Z	—	39	—	V
Sink Current ($V_{(Pin\ 15)}$ = 1.0 V)	I_{RZ}	—	0.3	—	mA
TOTAL DEVICE					
Standby Supply Current (Pin 6 at V_{ref}, All Other Inputs and Outputs Open) (V_{CC} = 15 V) (V_{CC} = 40 V)	I_{CC}	 — —	 5.5 7.0	 10 15	mA
Average Supply Current ($V_{(Pin\ 4)}$ = 2.0 V) (See Figure 12.) (C_T = 0.01, R_T = 12 kΩ, V_{CC} = 15V)	—	—	7.0	—	mA

* Standard deviation is a measure of the statistical distribution about the mean as derived from the formula, $\sigma = \sqrt{\frac{\sum_{n=1}^{N} (X_n - \overline{X})^2}{N - 1}}$

FIGURE 4 — OSCILLATOR FREQUENCY VERSUS TIMING RESISTANCE

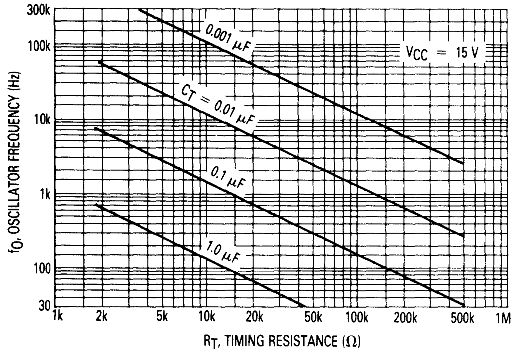


FIGURE 5 — OPEN LOOP VOLTAGE GAIN AND PHASE VERSUS FREQUENCY

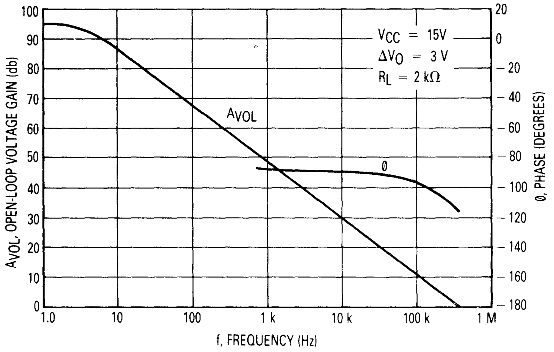


FIGURE 6 — PERCENT DEAD TIME VERSUS OSCILLATOR FREQUENCY

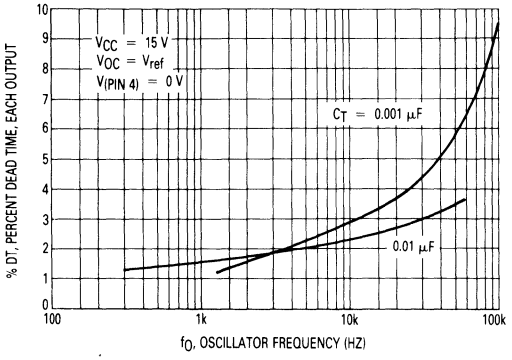


FIGURE 7 — PERCENT DUTY CYCLE VERSUS DEAD-TIME CONTROL VOLTAGE

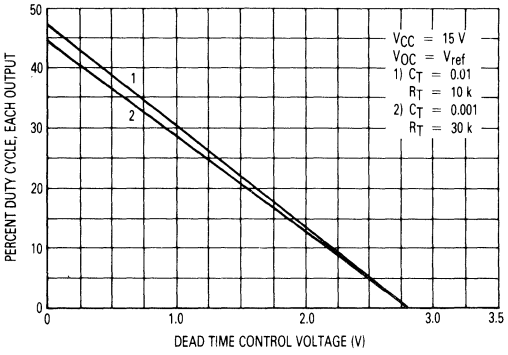


FIGURE 8 — EMITTER-FOLLOWER CONFIGURATION, OUTPUT-SATURATION VOLTAGE VERSUS EMITTER CURRENT

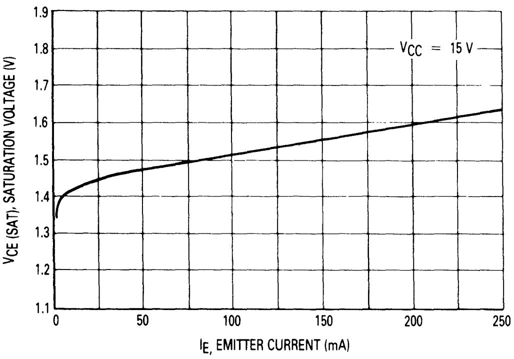


FIGURE 9 — COMMON-EMITTER CONFIGURATION OUTPUT-SATURATION VOLTAGE VERSUS COLLECTOR CURRENT

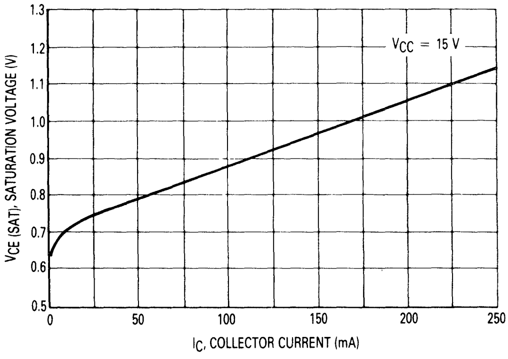


FIGURE 10 — STANDBY-SUPPLY CURRENT VERSUS SUPPLY VOLTAGE

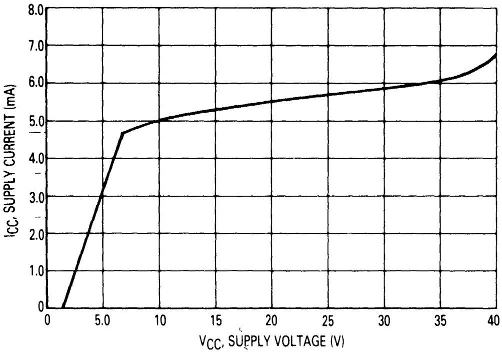


FIGURE 11 — ERROR AMPLIFIER CHARACTERISTICS

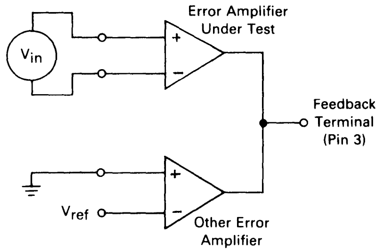


FIGURE 12 — DEAD-TIME AND FEEDBACK CONTROL TEST CIRCUIT

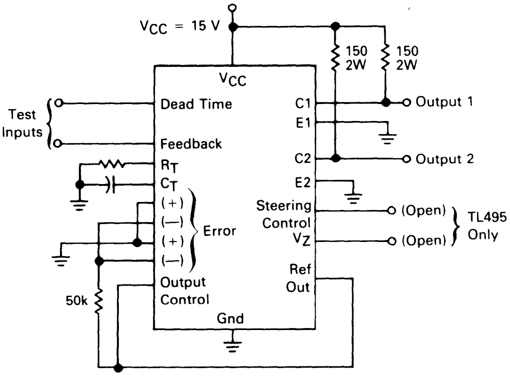


FIGURE 13 — COMMON-EMITTER CONFIGURATION TEST CIRCUIT AND WAVEFORM

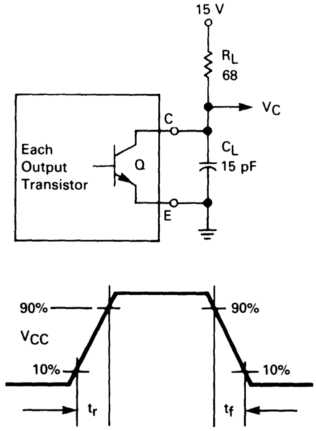


FIGURE 14 — EMITTER-FOLLOWER CONFIGURATION TEST CIRCUIT AND WAVEFORM

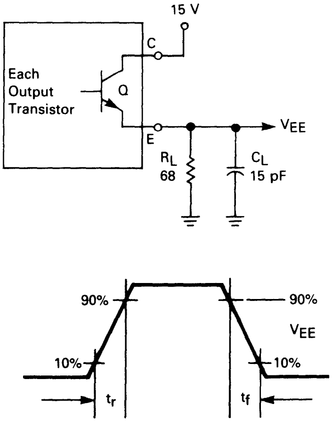


FIGURE 15 — ERROR-AMPLIFIER SENSING TECHNIQUES

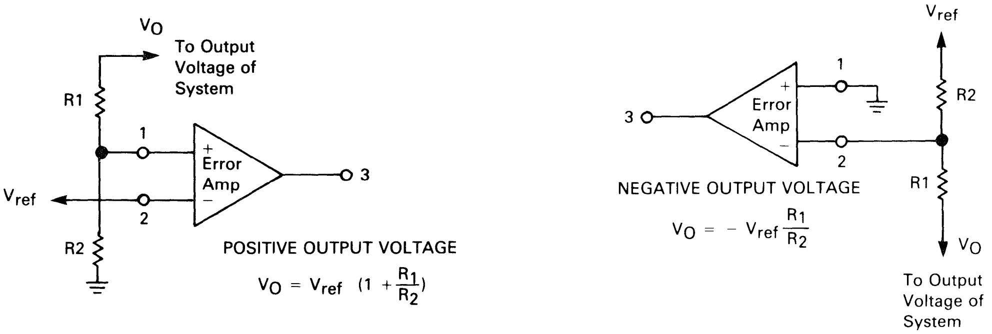


FIGURE 16 — DEAD-TIME CONTROL CIRCUIT

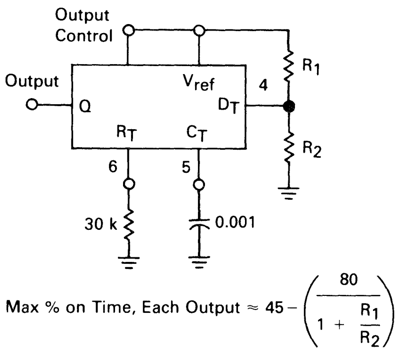


FIGURE 17 — SOFT-START CIRCUIT

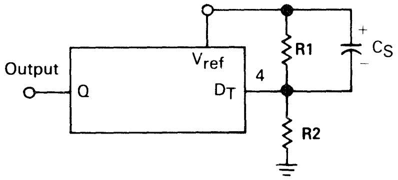
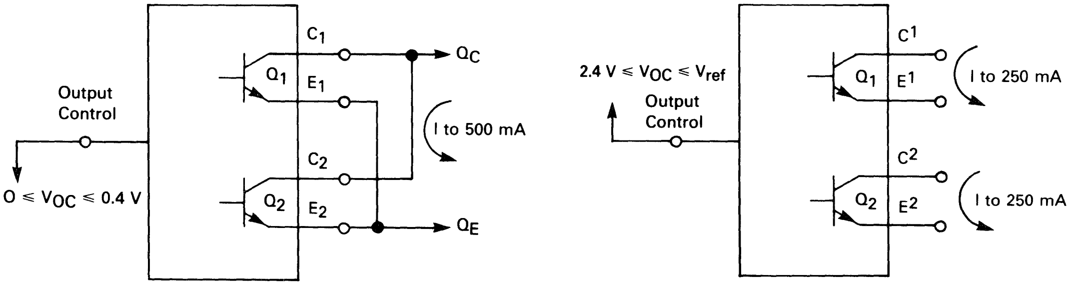


FIGURE 18 — OUTPUT CONNECTIONS FOR SINGLE-ENDED AND PUSH-PULL CONFIGURATIONS



Single Ended Configuration

Push-Pull Configuration

FIGURE 19 — SLAVING TWO OR MORE CONTROL CIRCUITS

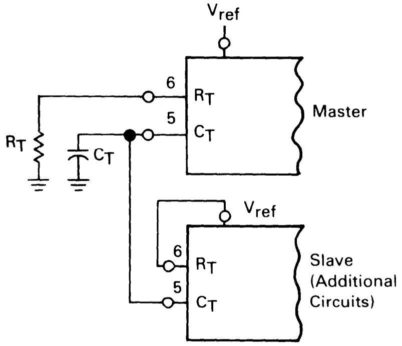


FIGURE 20 — OPERATION WITH V_{IN} > 40 V USING INTERNAL ZENER (TL495 ONLY)

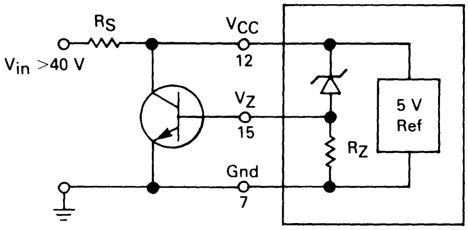
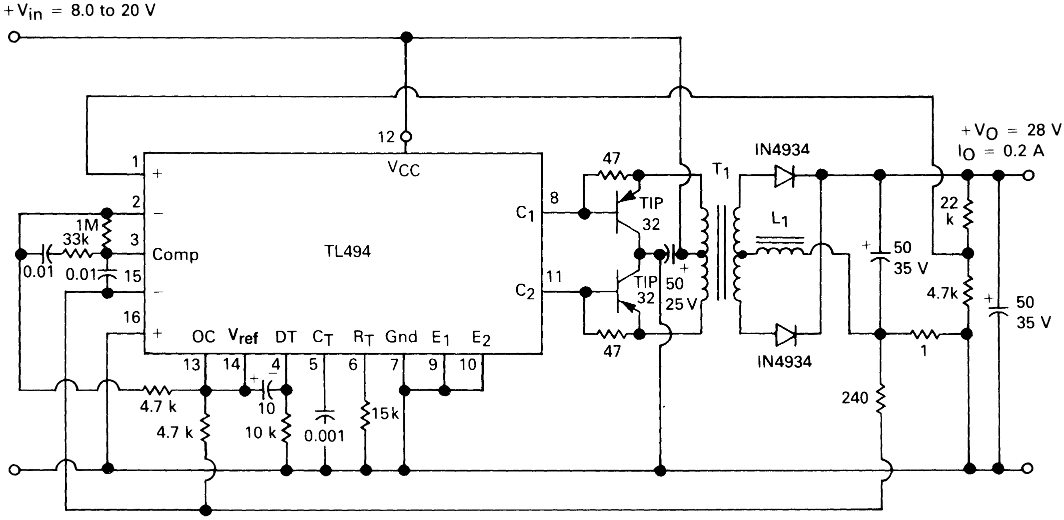


FIGURE 21 — PULSE-WIDTH MODULATED PUSH-PULL CONVERTER

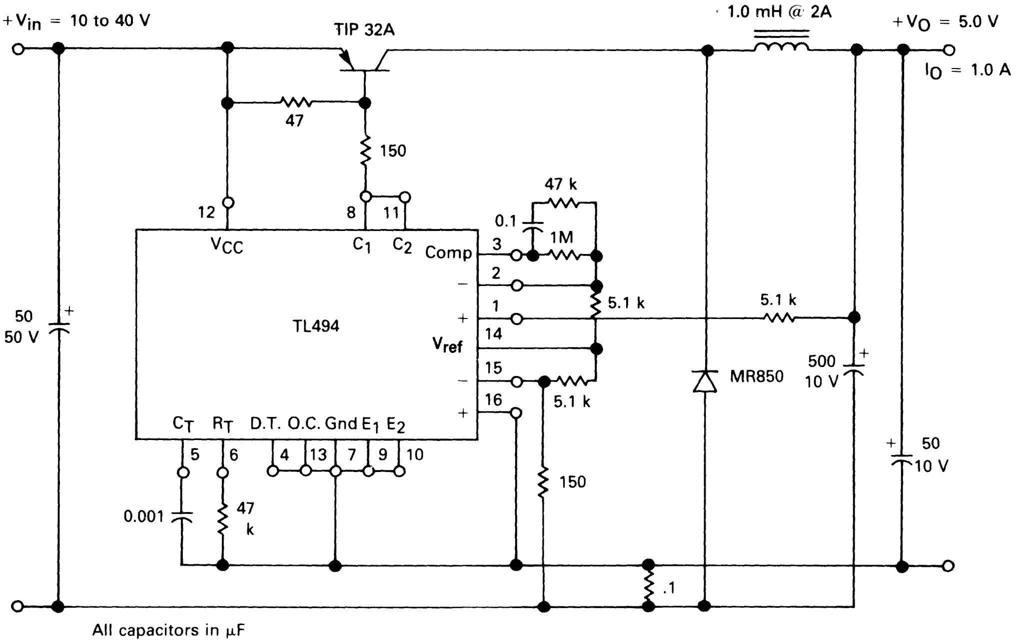


All capacitors in μF

L1 — 3.5 mh @ 0.3A
T1 — Primary: 20T C.T. #28 AWG
Secondary: 120T C.T. #36 AWG
Core: Ferroxcube 1408P-L00-3C8

TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 8.0 to 20 V	3.0 mV 0.01%
Load Regulation	V_{in} = 12.6 V, I_O = 0.2 to 200 mA	5.0 mV 0.02%
Output Ripple	V_{in} = 12.6 V, I_O = 200 mA	40 mV p-P P.A.R.D.
Short Circuit Current	V_{in} = 12.6 V, R_L = 0.1 Ω	250 mA
Efficiency	V_{in} = 12.6 V, I_O = 200 mA	72%

FIGURE 22 — PULSE-WIDTH MODULATED STEP-DOWN CONVERTER



TEST	CONDITIONS	RESULTS
Line Regulation	V_{in} = 10V to 40V	14mV 0.28%
Load Regulation	V_{in} = 28V, I_O = 1 mA to 1 A	3.0mV 0.06%
Output Ripple	V_{in} = 28V, I_O = 1.0A	65mV P-P P.A.R.D.
Short Circuit Current	V_{in} = 28V, R_L = 0.1Ω	1.6 amps
Efficiency	V_{in} = 28V, I_O = 1A	71%

MOTOROLA

μA78S40

Advance Information

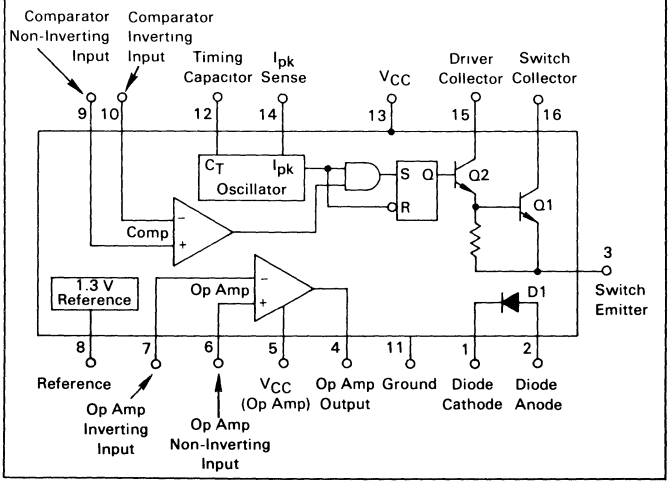
UNIVERSAL SWITCHING REGULATOR SUBSYSTEM

The μA78S40 is a monolithic-switching regulator subsystem, providing all active functions necessary for a switching regulator system. The device consists of a tight-tolerance temperature-compensated voltage reference, controlled-duty cycle oscillator with an active peak-current limit circuit, comparator, high-current and high-voltage output switch, capable of 1.5 A and 40 V, pinned-out power diode and an uncommitted operational amplifier, which can be powered up or down independent of the I.C. supply. The switching output can drive external NPN or PNP transistors when voltages greater than 40 V, or currents in excess of 1.5 A, are required. Some of the features are wide-supply voltage range, low standby current, high efficiency and low drift. The μA78S40 is available in both commercial (0°C to +70°C) and military (-55°C to +125°C) temperature ranges.

Some of the applications include use in step-up, step-down, and inverting regulators, with extremely good results obtained in battery-operated systems.

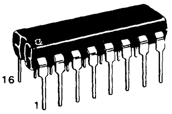
- Output Adjustable from 1.3 V to 40 V
- Peak Output Current of 1.5 A Without External Transistor
- 80 dB Line and Load Regulation
- Operation from 2.5 V to 40 V Supply
- Low Standby Current Drain
- High Gain, High Output Current, Uncommitted Op Amp.
- Uncommitted Power Diode
- Low Cost

μA78S40 EQUIVALENT CIRCUIT



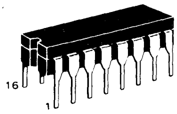
UNIVERSAL SWITCHING REGULATOR SUBSYSTEM

SILICON MONOLITHIC INTEGRATED CIRCUIT

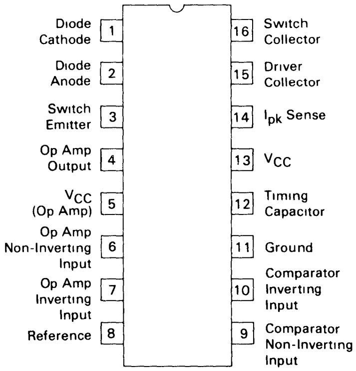


P SUFFIX
PLASTIC PACKAGE
CASE 648

D SUFFIX
CERAMIC PACKAGE
CASE 620



PIN CONNECTIONS



ORDERING INFORMATION

Device	Temperature Range	Package
μA78S40PC	0°C to +70°C	Plastic DIP
μA78S40DC	0°C to +70°C	Ceramic DIP
μA78S40DM	-55°C to +125°C	Ceramic DIP

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Power Supply Voltage	V_{CC}	40	V
Op Amp Power Supply Voltage	V_{CC} (Op Amp)	40	V
Common Mode Input Range (Comparator and Op Amp)	V_{ICR}	-0.3 to V_{CC}	V
Differential Input Voltage (Note 2)	V_{ID}	±30	V
Output Short-Circuit Duration (Op Amp)	—	Continuous	—
Reference Output Current	I_{ref}	10	mA
Voltage from Switch Collectors to Gnd	—	40	V
Voltage from Switch Emitters to Gnd	—	40	V
Voltage from Switch Collectors to Emitter	—	40	V
Voltage from Power Diode to Gnd	—	40	V
Reverse-Power Diode Voltage	V_{DR}	40	V
Current through Power Switch	I_{SW}	1.5	A
Current through Power Diode	I_D	1.5	A
Power Dissipation and Thermal Characteristics Plastic Package - T_A = +25°C Derate above +25°C (Note 1) Ceramic Package - T_A = 25°C Derate above +25°C (Note 1)	 P_D $1/R_{\theta JA}$ P_D $1/R_{\theta JA}$	 1500 14 1000 8	 mW mW/°C mW mW/°C
Storage Temperature Range	T_{stg}	-65 to +150	°C
Operating Temperature Range μA78S40M μA78S40C	T_A	 -55 to +125 0 to +70	°C

Notes:

1. T_{low} = -55°C for μA78S40DM
   = 0°C for μA78S40DC and μA78S40PC
   T_{high} = +125°C for μA78S40DM
   = +70°C for μA78S40DC and μA78S40PC
2. For supply voltages less than 30 V the maximum differential input voltage (Error Amp and Op Amp) is equal to the supply voltage.

ELECTRICAL CHARACTERISTICS (V_{CC} = 5.0 V, V_{CC} (Op Amp) = 5.0 V, T_A = T_{low} to T_{high} unless otherwise noted.)

Characteristic	Symbol	Min	Typ	Max	Unit
GENERAL					
Supply Voltage	V_{CC}	2.5	—	40	V
Supply Current (Op Amp Disconnected) (V_{CC} = 5.0 V) (V_{CC} = 40 V)	I_{CC}	 — —	 1.8 2.3	 3.5 5.0	mA
Supply Current (Op Amp Connected) (V_{CC} = 5.0 V) (V_{CC} = 40 V)	I_{CC}	 — —	 — —	 4.0 5.5	mA
REFERENCE					
Reference Voltage (I_{ref} = 1.0 mA)	V_{ref}	1.180	1.245	1.310	V
Reference Voltage Line Regulation (3.0 V ≤ V_{CC} ≤ 40 V, I_{ref} = 1.0 mA, T_A = 25°C)	Reg_{Line}	—	0.04	0.2	mV/V
Reference Voltage Load Regulation (1.0 mA ≤ I_{ref} ≤ 10 mA, T_A = 25°C)	Reg_{Load}	—	0.2	0.5	mV/mA

ELECTRICAL CHARACTERISTICS (Continued)

Characteristic	Symbol	Min	Typ	Max	Unit
OSCILLATOR					
Charging Current (T_A = 25°C) (V_{CC} = 5.0 V) (V_{CC} = 40 V)	I_{chg}	 20 20	 — —	 50 70	μA
Discharge Current (T_A = 25°C) (V_{CC} = 5.0 V) (V_{CC} = 40 V)	I_{chg}	 150 150	 — —	 250 350	μA
Oscillator Voltage Swing (T_A = 25°C) (V_{CC} = 5.0 V)	V_{osc}	—	0.5	—	V
Turn-on/Turn-off	t_{on}/t_{off}	—	6.0	—	μs/μs
CURRENT LIMIT					
Current-Limit Sense Voltage (T_A = 25°C) (V_{CC} – V_{IPK} [Sense])	—	250	—	350	mV
OUTPUT SWITCH					
Output Saturation Voltage 1 (I_{SW} = 1.0 A, Pin 15 tied to Pin 16)	V_{sat1}	—	1.1	1.3	V
Output Saturation Voltage 2 (I_{SW} = 1.0 A, I_{15} = 50 mA)	V_{sat2}	—	0.45	0.7	V
Output Transistor Current Gain (T_A = 25°C) (I_C = 1.0 A, V_{CE} = 5.0 V)	h_{FE}	—	70	—	—
Output Leakage Current (T_A = 25°C) (V_O = 40 V)	—	—	10	—	nA
POWER DIODE					
Forward Voltage Drop (I_D = 1.0 A)	V_D	—	1.25	1.5	V
Diode Leakage Current (T_A = 25°C) (V_{DR} = 40 V)	I_{DR}	—	10	—	nA
COMPARATOR					
Input Offset Voltage (V_{CM} = V_{ref})	V_{IO}	—	1.5	15	mV
Input Bias Current (V_{CM} = V_{ref})	I_{IB}	—	35	200	nA
Input Offset Current (V_{CM} = V_{ref})	I_{IO}	—	5.0	75	nA
Common-Mode Voltage Range (T_A = 25°C)	V_{ICR}	0	—	V_{CC}-2	V
Power-Supply Rejection Ratio (T_A = 25°C) ($3.0 \leqslant V_{CC} \leqslant 40$ V)	P_{SRR}	70	96	—	dB
OUTPUT OPERATIONAL AMPLIFIER					
Input Ofset Voltage (V_{CM} = 2.5 V)	V_{IO}	—	4.0	15	mV
Input Bias Current (V_{CM} = 2.5 V)	I_{IB}	—	30	200	nA
Input Offset Current (V_{CM} = 2.5 V)	I_{IO}	—	5.0	75	nA
Voltage Gain + (T_A = 25°C) (R_L = 2.0 kΩ to Gnd, $1.0\ V \leqslant V_O \leqslant 2.5$ V)	A_{vol+}	25000	250000	—	V/V
Voltage Gain – (T_A = 25°C) (R_L = 2.0 kΩ to V_{CC} (op amp), $1.0\ V \leqslant V_O \leqslant 2.5$ V)	A_{vol-}	25000	250000	—	V/V
Common-Mode Voltage Range (T_A = 25°C)	V_{ICR}	0	—	V_{CC}-2	V
Common-Mode Rejection Ratio (T_A = 25°C) (V_{CM} = 0 to 3.0 V)	C_{MRR}	76	100	—	dB
Power-Supply Rejection Ratio (T_A = 25°C) ($3.0\ V \leqslant V_{CC}$ (op amp) $\leqslant 40$ V)	P_{SRR}	76	100	—	dB
Output Source Current (T_A = 25°C)	I_{Source}	75	150	—	mA
Output Sink Current (T_A = 25°C)	I_{Sink}	10	35	—	mA
Slew Rate (T_A = 25°C)	SR	—	0.6	—	V/μs
Output Low Voltage (T_A = 25°C) (I_L = –5.0 mA)	V_{OL}	—	—	1.0	V
Output High Voltage (T_A = 25°C) (I_L = 50 mA)	V_{CC} (Op Amp) -3.0 V	—	—	—	V

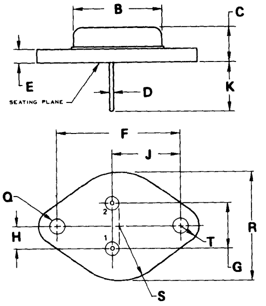
SECTION 19

PACKAGE OUTLINE DIMENSIONS

K SUFFIX

METAL PACKAGE

CASE 1-03



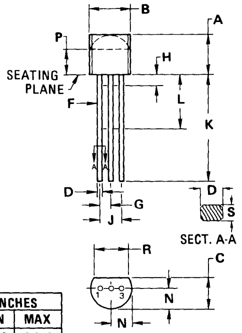
DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
B	–	22.23	–	0.875
C	6.35	11.43	0.250	0.450
D	0.97	1.09	0.038	0.043
E	–	3.43	–	0.135
F	29.90	30.40	1.177	1.197
G	10.67	11.18	0.420	0.440
H	5.21	5.72	0.205	0.225
J	16.64	17.15	0.655	0.675
K	7.92	–	0.312	–
Q	3.84	4.09	0.151	0.161
S	–	13.34	–	0.525
T	–	4.78	–	0.188

All JEDEC dimensions and notes apply

LP, P, Z SUFFIX

PLASTIC PACKAGE

CASE 29-02



DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
A	4.32	5.33	0.170	0.210
B	4.44	5.21	0.175	0.205
C	3.18	4.19	0.125	0.165
D	0.41	0.56	0.016	0.022
F	0.41	0.48	0.016	0.019
G	1.14	1.40	0.045	0.055
H	–	2.54	–	0.100
J	2.41	2.67	0.095	0.105
K	12.70	–	0.500	–
L	6.35	–	0.250	–
N	2.03	2.92	0.080	0.115
P	2.92	–	0.115	–
R	3.43	–	0.135	–
S	0.36	0.41	0.014	0.016

All JEDEC dimensions and notes apply.

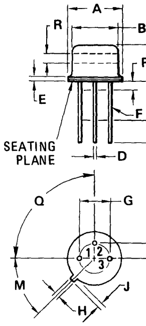
NOTES.

1. CONTOUR OF PACKAGE BEYOND ZONE "P" IS UNCONTROLLED.
2. DIM "F" APPLIES BETWEEN "H" AND "L". DIM "D" & "S" APPLIES BETWEEN "L" & 12.70 mm (0.5") FROM SEATING PLANE. LEAD DIM IS UNCONTROLLED IN "H" & BEYOND 12.70 mm (0.5") FROM SEATING PLANE.

G, H SUFFIX

METAL PACKAGE

CASE 79-02



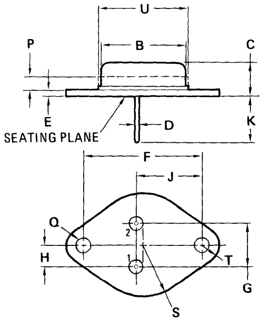
All JEDEC dimensions and notes apply.

DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
A	8.89	9.40	0.350	0.370
B	8.00	8.51	0.315	0.335
C	6.10	6.60	0.240	0.260
D	0.406	0.533	0.016	0.021
E	0.229	3.18	0.009	0.125
F	0.406	0.483	0.016	0.019
G	4.83	5.33	0.190	0.210
H	0.711	0.864	0.028	0.034
J	0.737	1.02	0.029	0.040
K	12.70	–	0.500	–
L	6.35	–	0.250	–
M	45° NOM		45° NOM	
P	–	1.27	–	0.050
Q	90° NOM		90° NOM	
R	2.54	–	0.100	–

R SUFFIX

METAL PACKAGE

CASE 80-02

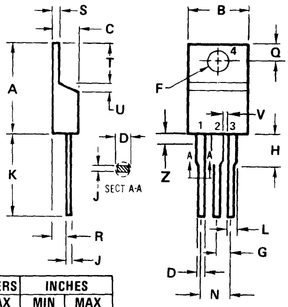


DIM	MILLIMETERS MIN	MILLIMETERS MAX	INCHES MIN	INCHES MAX
B	11.94	12.70	0.470	0.500
C	6.35	8.64	0.250	0.340
D	0.71	0.86	0.028	0.034
E	1.27	1.91	0.050	0.075
F	24.33	24.43	0.958	0.962
G	4.83	5.33	0.190	0.210
H	2.41	2.67	0.095	0.105
J	14.48	14.99	0.570	0.590
K	9.14	–	0.360	–
P	–	1.27	–	0.050
Q	3.61	3.86	0.142	0.152
S	–	8.89	–	0.350
T	–	3.68'	–	0.145
U	–	15.75	–	0.620

All JEDEC Dimensions and and Notes Apply.

PACKAGE OUTLINE DIMENSIONS (continued)

T SUFFIX
PLASTIC PACKAGE
CASE 221A-02

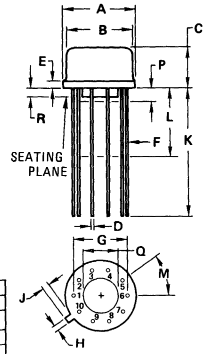


DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	15.11	15.75	0.595	0.620
B	9.65	10.29	0.380	0.405
C	4.06	4.82	0.160	0.190
D	0.64	0.89	0.025	0.035
F	3.61	3.73	0.142	0.147
G	2.41	2.67	0.095	0.105
H	2.79	3.30	0.110	0.130
J	0.36	0.56	0.014	0.022
K	12.70	14.27	0.500	0.562
L	1.14	1.27	0.045	0.050
N	4.83	5.33	0.190	0.210
Q	2.54	3.04	0.100	0.120
R	2.04	2.79	0.080	0.110
S	1.14	1.39	0.045	0.055
T	5.97	6.48	0.235	0.255
U	0.76	1.27	0.030	0.050
V	1.14	-	0.045	-
Z	–	2.03	–	0.080

NOTES:

1. DIMENSIONS L AND H APPLIES TO ALL LEADS.
2. DIMENSION Z DEFINES A ZONE WHERE ALL BODY AND LEAD IRREGULARITIES ARE ALLOWED.
3. DIMENSIONING AND TOLERANCING PER ANSI Y14.5 1973.
4. CONTROLLING DIMENSION: INCH

G SUFFIX
METAL PACKAGE
CASE 603-04

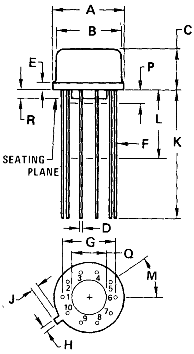


DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	8.51	9.39	0.335	0.370
B	7.75	8.51	0.305	0.335
C	4.19	4.70	0.165	0.185
D	0.407	0.533	0.016	0.021
E	–	1.02	–	0.040
F	0.406	0.483	0.016	0.019
G	5.84 BSC		0.230 BSC	
H	0.712	0.864	0.028	0.034
J	0.737	1.14	0.029	0.045
K	12.70	–	0.500	–
L	6.35	12.70	0.250	0.500
M	36^{0} BSC		36^{0} BSC	
P	–	1.27	–	0.050
Q	3.56	4.06	0.140	0.160
R	0.254	1.02	0.010	0.040

All JEDEC dimensions and notes apply

NOTE:
LEADS WITHIN 0.18 mm (0.007) RADIUS OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.

G SUFFIX
METAL PACKAGE
CASE 603C-01

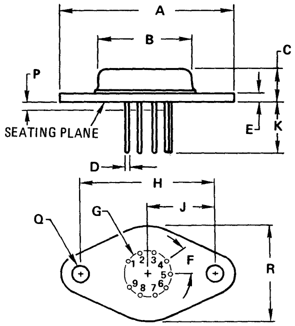


DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	8.51	9 39	0.335	0.370
B	7 75	8.51	0.305	0.335
C	4.19	6.73	0.165	0.265
D	0.407	0.533	0.016	0 021
E	–	1.02	–	0.040
F	0.406	0.483	0.016	0.019
G	5.84 BSC		0.230 BSC	
H	0.712	0.864	0.028	0.034
J	0.737	1.14	0.029	0.045
K	12 70	–	0.500	–
L	6.35	12.70	0.250	0.500
M	36^{0} BSC		36^{0} BSC	
P	–	1.27	–	0.050
Q	3.56	4.06	0.140	0 160
R	0.254	1.02	0.010	0.040

NOTES:

1. LEADS WITHIN 0.18 mm (0.007) RADIUS OF TRUE POSITION TO DIM. "A" & "H" AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.
2. LEAD DIA UNCONTROLLED BEYOND DIM "K" MIN.

R SUFFIX
METAL PACKAGE
CASE 614-02



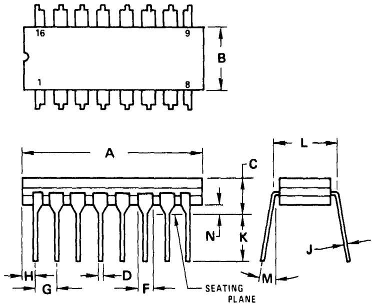
DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	–	31.80	–	1.252
B	11.94	12.70	0.470	0.500
C	6.35	8.64	0.250	0.340
D	0.71	0.81	0.028	0.032
E	1.27	1.90	0.050	0.075
F	36^{0} BSC		36^{0} BSC	
G	8.26 BSC		0.325 BSC	
H	24.33	24.43	0.958	0.962
J	12.17	12.22	0.479	0.481
K	9.14	–	0.360	–
P	1.40 BSC		0.055 BSC	
Q	3.61	3.86	0.142	0.152
R	–	17.78	–	0.700

NOTE:

1. LEADS TRUE POSITIONED WITHIN 0.36 mm (0.014) DIA. to DIM. "A" & "H" AT MAX. MATERIAL CONDITIONS AND DIM. "P"
2. LEAD DIAMETERS ARE UNCONTROLLED BEYOND 12.70 mm (0.500) FROM BASE PLANE.

PACKAGE OUTLINE DIMENSIONS (continued)

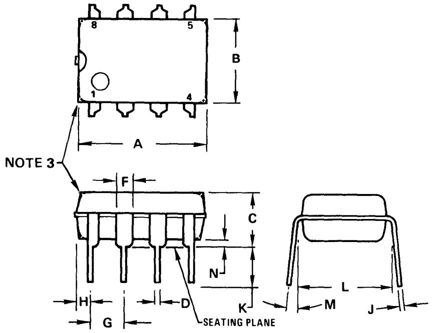
D, J, L SUFFIX
CERAMIC PACKAGE
CASE 620-02



DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	19.05	19.81	0.750	0.780
B	6.22	6.98	0.245	0.275
C	4.06	5.08	0.160	0.200
D	0.38	0.51	0.015	0.020
F	1.40	1.65	0.055	0.065
G	2.54 BSC		0.100 BSC	
H	0.51	1.14	0.020	0.045
J	0.20	0.30	0.008	0.012
K	3.18	4.06	0.125	0.160
L	7.37	7.87	0.290	0.310
M	–	15^0	–	15^0
N	0.51	1.02	0.020	0.040

NOTES:
1 LEADS WITHIN 0.13 mm (0.005) RADIUS OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION
2 PKG. INDEX: NOTCH IN LEAD NOTCH IN CERAMIC OR INK DOT
3 DIM "L" TO CENTER OF LEADS WHEN FORMED PARALLEL

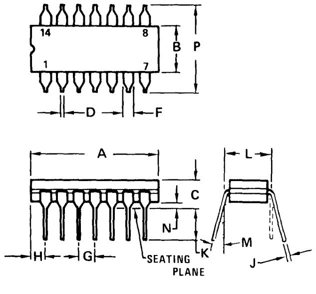
P, P1 SUFFIX
PLASTIC PACKAGE
CASE 626-04



DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	9.40	10.16	0.370	0.400
B	6.10	6.60	0.240	0.260
C	3.94	4.45	0.155	0.175
D	0.38	0.51	0.015	0.020
F	1.02	1.52	0.040	0.060
G	2.54 BSC		0.100 BSC	
H	0.76	1.27	0.030	0.050
J	0.20	0.30	0.008	0.012
K	2.92	3.43	0.115	0.135
L	7.62 BSC		0.300 BSC	
M	–	10^0	–	10^0
N	0.51	0.76	0.020	0.030

NOTES:
1. LEADS WITHIN 0.13 mm (0.005) RADIUS OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.
2. DIM "L" TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. PACKAGE CONTOUR OPTIONAL (ROUND OR SQUARE CORNERS)

L SUFFIX
CERAMIC PACKAGE
CASE 632-02

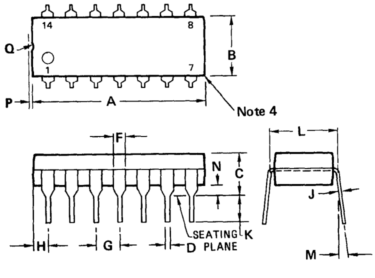


DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	16.8	19.9	0.660	0.785
B	5.59	7.11	0.220	0.280
C	–	5.08	–	0.200
D	0.381	0.584	0.015	0.023
F	0.77	1.77	0.030	0.070
G	2.54 BSC		0.100 BSC	
J	0.203	0.381	0.008	0.015
K	2.54	–	0.100	–
L	7.62 BSC		0.300 BSC	
M	–	15^0	–	15^0
N	0.51	0.76	0.020	0.030
P	–	8.25	–	0.325

All JEDEC dimensions and notes apply.

NOTES:
1. ALL RULES AND NOTES ASSOCIATED WITH MO-001 AA OUTLINE SHALL APPLY.
2. DIMENSION "L" TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. LEADS WITHIN 0.25mm (0.010) DIA OF TRUE POSITION AT SEATING PLANE AND MAXIMUM MATERIAL CONDITION.

P SUFFIX
PLASTIC PACKAGE
CASE 646-05

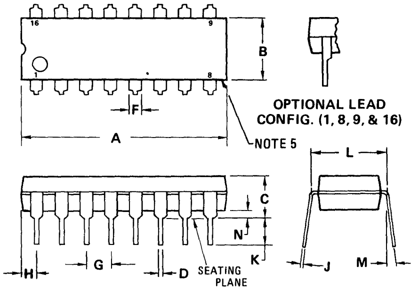


DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	18.16	19.56	0.715	0.770
B	6.10	6.60	0.240	0.260
C	4.06	5.08	0.160	0.200
D	0.38	0.53	0.015	0.021
F	1.02	1.78	0.040	0.070
G	2.54 BSC		0.100 BSC	
H	1.32	2.41	0.052	0.095
J	0.20	0.38	0.008	0.015
K	2.92	3.43	0.115	0.135
L	7.62 BSC		0.300 BSC	
M	0^0	10^0	0^0	10^0
N	0.51	1.02	0.020	0.040

NOTES:
1. LEADS WITHIN 0.13 mm (0.005) RADIUS OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.
2. DIMENSION "L" TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. DIMENSION "B" DOES NOT INCLUDE MOLD FLASH.
4. ROUNDED CORNERS OPTIONAL.

PACKAGE OUTLINE DIMENSIONS (continued)

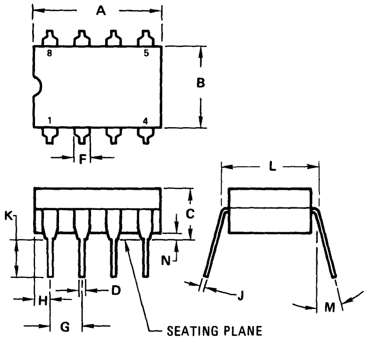
N, P SUFFIX
PLASTIC PACKAGE
CASE 648-05



DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	18.80	21.34	0.740	0.840
B	6.10	6.60	0.240	0.260
C	4.06	5.08	0.160	0.200
D	0.38	0.53	0.015	0.021
F	1.02	1.78	0.040	0.070
G	2.54 BSC		0.100 BSC	
H	0.38	2.41	0.015	0.095
J	0.20	0.38	0.008	0.015
K	2.92	3.43	0.115	0.135
L	7.62 BSC		0.300 BSC	
M	0°	10°	0°	10°
N	0.51	1.02	0.020	0.040

NOTES:
1. LEADS WITHIN 0.13 mm (0.005) RADIUS OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.
2. DIMENSION "L" TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. DIMENSION "B" DOES NOT INCLUDE MOLD FLASH.
4. "F" DIMENSION IS FOR FULL LEADS. "HALF" LEADS ARE OPTIONAL AT LEAD POSITIONS 1, 8, 9, and 16).
5. ROUNDED CORNERS OPTIONAL.

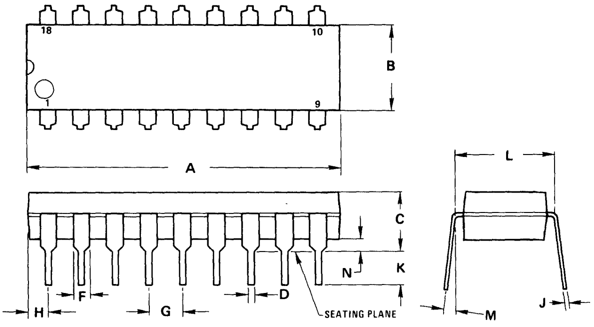
JG, U SUFFIX
CERAMIC PACKAGE
CASE 693-02



DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	9.91	10.92	0.390	0.430
B	6.22	6.99	0.245	0.275
C	4.32	5.08	0.170	0.200
D	0.41	0.51	0.016	0.020
F	1.40	1.65	0.055	0.065
G	2.54 BSC		0.100 BSC	
H	1.14	1.65	0.045	0.065
J	0.20	0.30	0.008	0.012
K	3.18	4.06	0.125	0.160
L	7.37	7.87	0.290	0.310
M	–	15°	–	15°
N	0.51	1.02	0.020	0.040

NOTES:
1. LEADS WITHIN 0.13 mm (0.005) RAD OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.
2. DIMENSION "L" TO CENTER OF LEADS WHEN FORMED PARALLEL.

N SUFFIX
PLASTIC PACKAGE
CASE 707-02

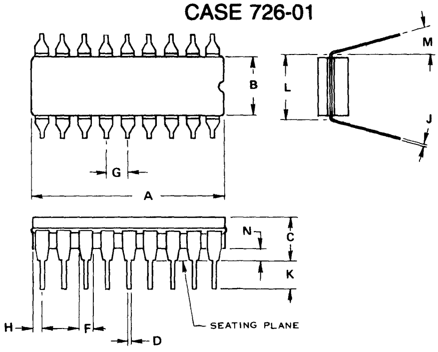


NOTES:
1. POSITIONAL TOLERANCE OF LEADS (D), SHALL BE WITHIN 0.25mm(0.010) AT MAXIMUM MATERIAL CONDITION, IN RELATION TO SEATING PLANE AND EACH OTHER.
2. DIMENSION L TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. DIMENSION B DOES NOT INCLUDE MOLD FLASH.

DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	22.22	23.24	0.875	0.915
B	6.10	6.60	0.240	0.260
C	3.56	4.57	0.140	0.180
D	0.36	0.56	0.014	0.022
F	1.27	1.78	0.050	0.070
G	2.54 BSC		0.100 BSC	
H	1.02	1.52	0.040	0.060
J	0.20	0.30	0.008	0.012
K	2.92	3.43	0.115	0.135
L	7.62 BSC		0.300 BSC	
M	0°	15°	0°	15°
N	0.51	1.02	0.020	0.040

PACKAGE OUTLINE DIMENSIONS (continued)

J SUFFIX
CERAMIC PACKAGE
CASE 726-01

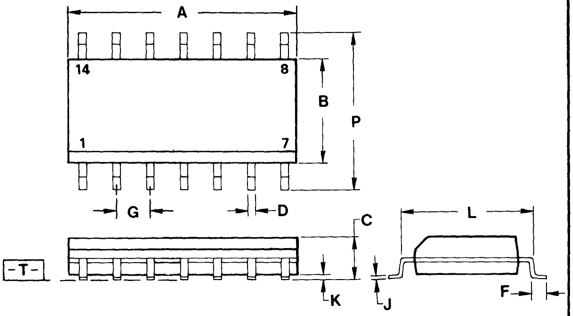


DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	22.35	23.11	0.880	0.910
B	6.63	7.24	0.261	0.285
C	–	5.08	–	0.200
D	0.41	0.51	0.016	0.020
F	1.40	1.65	0.055	0.065
G	2.54 BSC		0.100 BSC	
H	0.76	1.02	0.030	0.040
J	0.13	0.38	0.005	0.015
K	–	4.44	–	0.175
L	7.37	8.00	0.290	0.315
M	0^0	15^0	0^0	15^0
N	0.51	0.76	0.020	0.030

NOTES:

1. LEADS, TRUE POSITIONED WITHIN 0.25 mm (0.010) DIA. AT SEATING PLANE, AT MAXIMUM MATERIAL CONDITION.
2. DIM "L" TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. DIM "A" & "B" INCLUDES MENISCUS.

D SUFFIX
PLASTIC PACKAGE
CASE 751A-01



DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	8.54	8.74	0.336	0.344
B	3.81	4.01	0.150	0.158
C	1.35	1.75	0.053	0.069
D	0.35	0.46	0.014	0.018
F	0.67	0.77	0.026	0.030
G	1.27 BSC		0.050 BSC	
J	0.19	0.22	0.007	0.009
K	0.10	0.20	0.004	0.008
L	4.82	5.21	0.189	0.205
P	5.79	6.20	0.228	0.244

NOTES:

1. -T- IS SEATING PLANE.
2. DIMENSION A IS DATUM.
3. POSITIONAL TOLERANCE FOR LEADS:

⌖	0.25 (0.010) Ⓜ	A Ⓢ

SECTION 20
VOLTAGE REGULATOR CROSS REFERENCE GUIDE

This cross reference provides a complete interchangeability list linking the most common voltage regulators offered by major Linear Integrated Circuits manufacturers to the nearest equivalent Motorola device. The Motorola ''Direct Replacement'' column lists devices with identical pin connections and package and the same or better electrical characteristics and temperature range. The Motorola ''Functional Equivalent'' column provides a device which performs the same function but with possible differences in package configurations, pin connections, temperature range or electrical characteristics.

Grouped by individual manufacturers, reference numbers are listed in alphanumeric sequence, with Greek ''μ'' preface numbers appearing first.

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
FAIRCHILD		
µA109KM	LM109K	
µA117KM	LM117K	
µA209KM	LM209K	
µ217UV		LM217K
µA309KC	LM309K	
µA317KC	LM317K	
µA317UC	LM317T	
µA494DC	TL494CJ	
µA494DM	TL494MJ	
µA494PC	TL494CN	
µA723DC	MC1723CL	
µA723DM	MC1723L	
µA723HC	MC1723CG	
µA723HM	MC1723G	
µA723PC	MC1723CP	
µA7805KC	MC7805CK	
µA7805KM	MC7805K	
µA7805UC	MC7805CT	
µA7805UV	MC7805BT	
µA7806KC	MC7806CK	
µA7806KM	MC7806K	
µA7806UC	MC7806CT	
µA7806UV	MC7806BT	
µA7808KC	MC7808K	
µA7808KM	MC7808K	
µA7808UC	MC7808CT	
µA7808UV	MC7808BT	
µA7812KC	MC7812CK	
µA7812KM	MC7812K	
µA7812UC	MC7812CT	
µA7812UV	MC7812BT	
µA7815KC	MC7815CK	
µA7815KM	MC7815K	
µA7815UC	MC7815CT	
µA7815UV	MC7815BT	
µA7818KC	MC7818CK	
µA7818KM	MC7818K	
µA7818UC	MC7818CT	
µ7818UV	MC7818BT	
µ7824KC	MC7824CK	
µA7824KM	MC7824K	
µA7824UC	MC7824CT	
µA7824UV	MC7824BT	
µA78GKC		LM317K
µA78GKM		LM117K
µA78GUC		LM317T
µA78L05AHC	MC78L05ACG	
µA78L05AWC	MC78L05ACP	
µA78L08AWC	MC78L08ACP	
µA78L12AHC	MC78L12ACG	
µA78L12AWC	MC78L12ACP	
µA78L15AHC	MC78L15ACG	
µA78L15AWC	MC78L15ACP	
µA78L18AHC	MC78L18ACG	
µA78L18AWC	MC78L18ACP	
µA78L24AHC	MC78L24ACG	
µA78L24AWC	MC78L24ACP	
µA78MGHC		LM317MR
µA78MGHM		LM117MR
µA78MGUC		LM317MT
µA78M05HC	MC78M05CG	
µA78M05UC	MC78M05CT	
µA78M06HC	MC78M06CG	
µA78M06UC	MC78M06CT	
µA78M08HC	MC78M08CG	
µA78M08UC	MC78M08CT	
µA78M12HC	MC78M12CG	
µA78M12UC	MC78M12CT	
µA78M15HC		MC78M15CG
µA78M15UC	MC78M15CT	
µA78M24HC	MC78M24CG	
µA7905KM		MC7905CK
µA7905UC	MC7905CT	
µA7906KC	MC7906CK	
µA7906KM		MC7906CK
µA7906UC	MC7906CT	
µA7908KC	MC7908CT	
µA7908KM		MC7908CK
µA7908UC		MC7908CT
µA7912KC	MC7912CK	
µA7912KM		MC7912CK
µA7912UC	MC7912CT	
µA7915KC	MC7915CK	
µA7915KM		MC7915CK
µA7915UC	MC7915CT	
µA7918KC	MC7918CK	
µ7918KM		MC7918CK
µA7918UC	MC7918CT	
µA7924KC	MC7924CK	
µA7924KM		MC7924CK
µA7924UC	MC7924CT	
µA79M05AUC		MC7905CT
µA79M06AUC		MC7906CT
µA79M08AUC		MC7908CT
µA79M12AUC		MC7912CT
µA79M15AUC		MC7915CT
µA79M24AUC		MC7924CT
SH323SKC	LM323K	
NATIONAL		
LM109H	LM109H	
LM109K	LM109K	
LM117H	LM117H	
LM117K	LM117K	
LM120H-5.0		MC7905CK
LM120H-12		MC7912CK
LM120K-5.0		MC7905CK
LM120K-12		MC7912CK
LM120H-15		MC7915CK
LM120K-15		MC7915CK
LM123K	LM123K	
LM125H		MC1568G
LM126H		MC1568G
LM137K	LM137K	
LM140AK-5	MC7805AK	
LM140AK-12	MC7812AK	
LM140AK-15	MC7815AK	
LM140K-5.0	LM140K-5.0	
LM140K-12	LM140K-12	
LM140K-15	LM140K-15	
LM140LAH-5.0		MC78L05ACG
LM140LAH-12		MC78L12ACG
LM140LAH-15		MC78L15ACG
LM150K	LM150K	

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
LM209H	LM209H	
LM209K	LM209K	
LM217H	LM217H	
LM217K	LM217K	
LM223K	LM223K	
LM225H		MC1568G
LM226H		MC1568G
LM237K	LM237K	
LM250K	LM250K	
LM309H	LM309H	
LM309K	LM309K	
LM317H	LM317H	
LM317K	LM317K	
LM317MP	LM317MT	
LM317T	LM317T	
LM320H-5.0		MC7905CK
LM320H-12		MC7912CK
LM320H-15		MC7915CK
LM320K-5.0	MC7905CK	
LM320K-12	MC7912CK	
LM320K-15	MC7915CK	
LM320LZ-5.0	MC79L05ACP	
LM320LZ-12	MC79L12ACP	
LM320LZ-15	MC79L15ACP	
LM320T-5.0	MC7905CT	
LM320T-12	MC7912CT	
LM320T-15	MC7915CT	
LM323K	LM323K	
LM325AN		MC1468L
LM325AS		MC1468L
LM325G		MC1468L
LM325H		MC1468L
LM325N		MC1468L
LM326H		MC1468G
LM326N		MC1468L
LM326S		MC1468L
LM337K	LM337K	
LM337MP		LM337MT
LM337T	LM337T	
LM340AK-5.0	MC7805ACK	
LM340AK-12	MC7812ACK	
LM340AK-15	MC7815ACK	
LM340AT-5.0	MC7805ACT	
LM340AT-12	MC7812ACT	
LM340AT-15	MC7815ACT	
LM340K-5.0	LM340K-5.0	
LM340K-12	LM340K-12	
LM340K-15	LM340K-15	
LM340LAH-5.0		MC78L05ACG
LM340LAH-12		MC78L12ACG
LM340LAH-15		MC78L15ACG
LM340LAZ-5.0		MC78L05ACP
LM340LAZ-12		MC78L12ACP
LM340LAZ-15		MC78L15ACP
LM340T-5.0	MC7805CT	
LM340T-12	MC7812CT	
LM341P-5.0	MC78M05CT	
LM341P-12	MC78M12CT	
LM341P-15	MC78M15CT	
LM342P-5.0	MC78M05CT	
LM342P-12	MC78M12CT	
LM342P-15	MC78M15CT	
LM350K	LM350K	
LM723CH	MC1723CG	
LM723CJ	MC1723CL	
LM723CN	MC1723CP	
LM723H	MC1723G	
LM723J	MC1723L	
LM7805CK	MC7805CK	
LM7805CT	MC7805CT	
LM7812CK	MC7812CK	
LM7812CT	MC7812CT	
LM7815CK	MC7815CK	
LM7815CT	LM7815CT	
LM78L05ACH	MC78L05ACG	
LM78L05ACZ	MC78L05ACP	
LM78L05CH	MC78L05CG	
LM78L05CZ	MC78L05CP	
LM78L12ACH	MC78L12ACG	
LM78L12ACZ	MC78L12ACP	
LM78L12CH	MC78L12CG	
LM78L12CZ	MC78L12CP	
LM78L15ACH	MC78L15ACG	
LM78L15ACZ	MC78L15ACP	
LM78L15CH	MC78L15CG	
LM78L15CZ	MC78L15CP	
LM78M05CP		MC78M05CT
LM78M12CP		MC78M12CT
LM78M15CP		MC78M15CT
LM7905CK	MC7905CK	
LM7905CT	MC7905CT	
LM7912CK	MC7912CK	
LM7912CT	MC7912CT	
LM7915CK	MC7915CK	
LM7915CT	MC7915CT	
LM79L05ACZ	MC79L05ACP	
LM79L12ACZ	MC79L12ACP	
LM79L15ACZ	MC79L15ACP	
RAYTHEON		
LM109H	LM109H	
LM209H	LM209H	
LM309H	LM309H	
RC4194DC		MC1468L
RC4194TK		MC1468R
RC4195NB		MC1468L
RC4195T		MC1468G
RC4195TK		MC1468R
RC723DB	MC1723CP	
RC723DC	MC1723CL	
RC723T	MC1723CG	
RM4194DC		MC1568L
RM4194TK		MC1568R
RM4195T		MC1568G
RM4195TK		MC1568R
RM723DC	MC1723L	
RM723T	MC1723G	
RCA		
CA3085		MC1723G
CA3085A		MC1723G
CA3085AF		MC1723L
CA3085AS		MC1723G
CA3085B		MC1723G
CA3085BF		MC1723L

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
CA3085BS		MC1723G
CA3085F		MC1723L
CA3085S		MC1723G
CA723CE	MC1723CP	
C723CT	MC1723CG	
CA723T	MC1723G	
CA723E	MC1723L	
SIGNETICS		
µA723F	MC1723L	
µA723CF	MC1723CL	
µA723CL	MC1723CG	
µA723CN	MC1723CP	
NE550A		MC1723CP
NE550L		MC1723CG
SE550L		MC1723G
SILICON GENERAL		
SG109K	LM109K	
SG109R		MC109K
SG109T	LM109H	
SG117T	LM117H	
SG117R		LM117K
SG117K	LM117K	
SG123K	LM123K	
SG137T	LM137H	
SG137R		LM137K
SG137K	LM137K	
SG140K-05	LM140K-5.0	
SG140K-06	LM140K-6.0	
SG140K-08	LM140K-8.0	
SG140K-12	LM140K-12	
SG140K-15	LM140K-15	
SG140K-18	LM140K-18	
SG140K-24	LM140K-24	
SG1468T	MC1468G	
SG1468R	MC1468R	
SG1468J	MC1468L	
SG1468N		MC1468L
SG150K	LM150K	
SG1501AJ		MC1568L
SG1501J	MC1568L	
SG1501T	MC1568G	
SG1502J		MC1568L
SG1503Y		MC1503U
SG1503T		MC1503U
SG1511T		MC1563G
SG1511J		MC1563G
SG1568T	MC1568G	
SG1568R	MC1568R	
SG1568J	MC1568L	
SG209K	LM209K	
SG209R		MC209K
SG209T	LM209H	
SG217T	LM217H	
SG217R		LM217K
SG217K	LM217K	
SG223K	LM223K	
SG237T	LM237H	
SG237R		LM237K
SG237K	LM237K	

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
SG2501AT		
SG2501J	MC1468L	
SG2501T	MC1468G	
SG2502J		MC1468L
SG2502N		MC1468L
SG2503M		MC1403AU
SG2503Y		MC1403AU
SG2503T		MC1403AU
SG250K	LM250K	
SG309K	LM309K	
SG309P		LM309K
SG309R		MC309K
SG309T	LM309H	
SG317T	LM317H	
SG317R		LM317T
SG317K	LM317K	
SG317P	LM317T	
SG337T	LM337H	
SG337R		LM337T
SG337K	LM337K	
SG337P	LM337T	
SG340K-05	LM340K-5.0	
SG340K-06	LM340K-6.0	
SG340K-08	LM340K-8.0	
SG340K-12	LM340K-12	
SG340K-15	LM340K-15	
SG340K-18	LM340K-18	
SG340K-24	LM340K-24	
SG3501AJ	MC1468L	
SG3501AN		MC1468L
SG3501AT	MC1468G	
SG3501J	MC1468L	
SG3501T	MC1468G	
SG3502J		MC1468L
SG3503Y	MC1403U	
SG3503T		MC1403U
SG3503M		MC1403U
SG350K	LM350K	
SG3511T		MC1463G
SG3511J		MC1463G
SG3511N		MC1463G
SG4194CJ		MC1468L
SG4194J		MC1568L
SG4194CR		MC1468R
SG4194R		MC1568R
SG4501T		MC1468G
SG4501J		MC1468L
SG4501N		MC1468L
SG501AJ		MC1468G
SG723CJ	MC1723CL	
SG723CN	MC1723CP	
SG723CT	MC1723CG	
SG723J	MC1723L	
SG723T	MC1723G	
SG7805ACK	MC7805ACK	
SG7805ACP	MC7805ACT	
SG7805ACR		MC7805ACT
SG7805ACT		MC7805ACT
SG7805AK	MC7805AK	
SG7805AR		MC7805AK
SG7805AT		MC7805AK
SG7805CK	MC7805CK	

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
SG7805CP	MC7805CT	
SG7805CR		MC7805CT
SG7805CT		MC78M05CG
SG7805K	MC7805K	
SG7805R		MC7805K
SG7805T		MC7805K
SG7806ACK	MC7806ACK	
SG7806ACP	MC7806ACT	
SG7806ACR		MC7806ACT
SG7806ACT		MC7806ACT
SG7806AK	MC7806AK	
SG7806AR		MC7806AK
SG7806AT		MC7806AK
SG7806CK	MC7805CK	
SG7806CP	MC7806CT	
SG7806CR		MC7806CT
SG7806CT		MC78M06CG
SG7806K	MC7806K	
SG7806R		MC7806K
SG7806T		MC7806K
SG7808ACK	MC7808ACK	
SG7808ACP	MC7808ACT	
SG7808ACR		MC78M08ACT
SG7808ACT		MC7808ACT
SG7808AK	MC7808AK	
SG7808AR		MC7808AK
SG7808AT		MC7808AK
SG7808CK	MC7808CK	
SG7808CP	MC7808CT	
SG7808CR		MC7808CT
SG7808CT		MC7808CG
SG7808K	MC7808K	
SG7808R		MC7808K
SG7808T		MC7808K
SG7812ACK	MC7812ACK	
SG7812ACP	MC7812ACT	
SG7812ACR		MC7812ACT
SG7812ACT		MC7812ACT
SG7812AK	MC7812AK	
SG7812AR		MC7812AK
SG7812AT		MC7812AK
SG7812CK	MC7812CK	
SG7812CP	MC7812CT	
SG7812CR		MC7812CT
SG7812CT		MC78M12CG
SG7812K	MC7812K	
SG7815ACK	MC7815ACK	
SG7815ACP	MC7815ACT	
SG7815ACR		MC7815ACT
SG7815ACT		MC7815ACT
SG7815AK	MC7815AK	
SG7815AR		MC7815AK
SG7815AT		MC7815AK
SG7815CK	MC7815CK	
SG7815CP	MC7815CT	
SG7815CR		MC7815CT
SG7815CT		MC78M15CG
SG7815K	MC7815K	
SG7815R		MC7815K
SG7815T		MC7815K
SG7818ACK	MC7818ACK	

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
SG7818ACP	MC7818ACT	
SG7818ACR		MC7818ACT
SG7818ACT		MC7818ACT
SG7818AK	MC7818AK	
SG7818AR		MC7818AK
SG7818AT		MC7818AK
SG7818CK	MC7818CK	
SG7818CP	MC7818CT	
SG7818CR	MC7818CT	
SG7818CT		MC7818CG
SG7818K	MC7818K	
SG7818R		MC7818K
SG7818T		MC7818K
SG7824ACK	MC7824ACK	
SG7824ACP	MC7824ACT	
SG7824ACR		MC7824ACT
SG7824ACT		MC7824ACT
SG7824AK	MC7824AK	
SG7824AR		MC7824AK
SG7824AT		MC7824AK
SG7824CK	MC7824CK	
SG7824CP	MC7824CT	
SG7824CR		MC7824CT
SG7824CT		MC78M24CG
SG7824K	MC7824K	
SG7824R		MC7824K
SG7824T		MC7824K
SG7905ACK	MC7905ACK	
SG7905ACP	MC7905ACT	
SG7905ACR		MC7905ACT
SG7905ACT		MC7905ACT
SG7905CK		MC7905CK
SG7905CP	MC7905CT	
SG7905CR		MC7905CT
SG7905CT		MC7905CT
SG7905.2CK	MC7905.2CK	
SG7905.2CP	MC7905.2CT	
SG7905.2CR		MC7905.2CT
SG7905.2CT		MC7905.2CT
SG7908CK	MC7908CK	
SG7908CP	MC7908CT	
SG7908CR		MC7908CT
SG7908CT		MC7908CT
SG7912ACK	MC7912ACK	
SG7912ACP	MC7912ACT	
SG7912ACR		MC7912ACT
SG7912ACT		MC7912ACT
SG7912CK	MC7912CK	
SG7912CP	MC7912CT	
SG7912CR		MC7912CT
SG7912CT		MC7912CT
SG7915ACK	MC7915ACK	
SG7915ACP	MC7915ACT	
SG7915ACR		MC7915ACT
SG7915ACT		MC7915ACT
SG7915CK	MC7915CK	
SG7915CP	MC7915CT	
SG7915CR		MC7915CT
SG7915CT		MC7915CT
SG7918CK	MC7918CK	
SG7918CP	MC7918CT	

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
T.I.		
μA723CJ	MC1723CL	
μA723CL	MC1723CG	
μA723CN	MC1723CP	
μ723MJ	MC1723L	
μ723ML	MC1723G	
μA7805CKC	MC7805CT	
μA7806CKC	MC7806CT	
μA7808CKC	MC7808CT	
μA7812CKC	MC7812CT	
μA7815CKC	MC1715CT	
μA7818CKC	MC7818CT	
μA7824CKC	MC7824CT	
μA78L05ACJG		MC78L05ACG
μA78L05ACLP	MC78L05ACP	
μA78L05CJG		MC78L05CG
μA78L05CLP	MC78L05CP	
μA78L08ACJG		MC78L08ACG
μA78L08ACLP		MC78L08ACP
μA78L08CJG		MC78L08CG
μA78L08CLP	MC78L08CP	
μA78L12ACJG		MC78L12ACG
μA78L12ACLP	MC78L12ACP	
μA78L12CJG		MC78L12CG
μA78L12CLP	MC78L12CP	
μA78L15ACJG		MC78L15ACG
μA78L15ACLP	MC78L15ACP	
μA78L15CJG		MC78L15CG
μA78L15CLP	MC78L15CP	
μA78M05CKC	MC78M05CT	
μA78M05CKD		MC78M05CT
μA78M05CLA	MC78M05CG	
μA78M06CKC	MC78M06CT	
μA78M06CKD		MC78M06CT
μA78M06CLA	MC78M06CG	
μA78M08CKC	MC78M08CT	
μA78M08CKD		MC78M08CT
μA78M08CLA	MC78M08CG	
μA78M12CKC	MC78M12CT	
μA78M12CKD		MC7812CT
μA78M12CLA	MC78M12CG	
μA78M15CKC	MC78M15CT	
μA7815CKD		MC78M15CT
μA78M15CLA	MC78M15CG	

REFERENCE NUMBER	MOTOROLA DIRECT REPLACEMENT	MOTOROLA FUNCTIONAL EQUIVALENT
μA78M20CKC	MC78M20CT	
μA78M20CKD		MC78M20CT
μA78M20CLA	MC78M20CG	
μA78M24CKC	MC78M24CT	
μA78M24CKD		MC78M24CT
μA78M24CLA	MC78M24CG	
μA7905CKC	MC7905CT	
μA7905.2CKC	MC7905.2CT	
μA7906CKC	MC7906CT	
μA7908CKC	MC7908CT	
μA7912CKC	MC7912CT	
μA7915CKC	MC7915CT	
μA7918CKC	MC7918CT	
μA7924CKC	MC7924CT	
μA79M05CKC		MC7905CT
μA79M06CKC		MC7806CT
μA79M08CKC		MC7908CT
μA79M12CKC		MC7912CT
μA79M15CKC		MC7915CT
μA79M24CKC		MC7924CT
LM109LA	LM109H	
LM117LA	LM117H	
LM209LA	LM209H	
LM217KC		LM217K
LM217KD		LM217H
LM217LA	LM217H	
LM309LA	LM309H	
LM317KC	LM317T	
LM317KD		LM317T
LM317LA	LM317H	
LM340KC-5		LM340K-5.0
LM340KC-6		LM340K-6.0
LM340KC-8		LM340K-8.0
LM340KC-12		LM340K-12
LM340KC-15		LM340K-15
LM340KC-18		LM340K-18
LM340KC-24		LM340K-24
TL494CJ	TL494CJ	
TL494CN	TL494CN	
TL494MJ	TL494MJ	
TL495CJ	TL495CJ	
TL495CN	TL495CN	
TL495MJ	TL495MJ	
TL7805ACKC	MC7805ACT	

APPENDIX A

SWITCHMODE POWER TRANSISTOR APPLICATION SELECTOR GUIDE

For line-operated SWITCHMODE power supplies (20 to 50 kHz, 40 to 3200 watts), this guide offers the power supply design engineer an easy way to identify those Motorola SWITCHMODE Transistors most ideally-suited for his particular application. To use the five tables in this guide, the designer must first:

1. Determine which of five circuits he will be using (i.e., full-bridge, half-bridge, push-pull, forward or flyback).
2. Determine which of three line voltages he will be using (i.e., 120, 220, or 380 Vac).
3. Determine the output power capability needed by his design (the table covers the area of 40 to 3200 watts).

Tables 1 through 3 list devices by $V_{CEO\,(sus)}$ for use in bridge circuits at either 120, 220 or 380 volts. Tables 4 and 5 list the same devices by V_{CEV} for use in the push-pull, forward and flyback circuits at either 120 or 220 volts. Within each table, the devices are grouped by the output power capability of that circuit, and the equivalent operating current level is also noted.

TABLE 1

CIRCUIT: HALF AND FULL* BRIDGE
LINE VOLTAGE: 120 VRMS
DEVICE V_{CEO} RATING ≥200 V

Circuit Rating		Metal—TO-204**, TO-66		Plastic—TO-220AB, TO-126		Darlington—TO-204**	
Output Power* (Watts)	$I_{C(OP)}$ (Amps)	Device Type	Rated V_{CEO} (Volts)	Device Type	Rated V_{CEO} (Volts)	Device Type	Rated V_{CEO} (Volts)
40	1	2N6233	225	MJE13002	300		
		2N6421PNP	250				
		2N6078	250				
		2N3584	250				
		2N6077	275				
		2N6234	275				
		2N3585	300				
		2N6212PNP	300				
		2N6422PNP	300				
		MJ4360	300				
		2N6235	325				
		2N6213PNP	350				
		MJ4361	400				
80	2	2N5838	250	MJE13004	300		
		2N5839	275				
120	3	2N6306	250	2N6497	250		
		MJ6502PNP	250				
		2N6307	300	2N6498	300		
		2N6542	300	2N6499	350		
		MJ4380	300				
		MJ4400	300				
		2N6308	350				
		MJ4381	400				
		MJ4401	400				
200	5	2N6544	300	MJE13006	300	MJ10006	350
		MJ13014	350	MJE5850PNP	300		
		MJ6502PNP	250	MJE5851PNP	350		
320	8	MJ13014	350	MJE13008	300		
400	10	2N6249	200			MJ10004	350
		MJ13330	200				
		MJ13331	250				
		2N6250	275				
		2N6546	300				
		2N6251	350				
		MJ13332	350				
800	20					MJ10015	400
						MJ10022	350
1200	30					MJ10020	200
						MJ10021	250

*NOTE: Power output ratings are for half-bridge circuit configurations, multiply by 2 for full-bridge.

**Formerly TO-3

TABLE 2

CIRCUIT: HALF AND FULL* BRIDGE
LINE VOLTAGE: 220 VRMS
DEVICE V_{CEO} RATING ≥400V

Circuit Rating		Metal—TO-204**, TO-66		Plastic—TO-220AB, TO-126		Darlington—TO-204**	
Output Power* (Watts)	$I_{C(OP)}$ (Amps)	Device Type	Rated V_{CEO} (Volts)	Device Type	Rated V_{CEO} (Volts)	Device Type	Rated V_{CEO} (Volts)
80	1	MJ4361	400	MJE13003	400		
160	2	MJ4381	400	MJE13005	400		
240	3	2N6543 MJ4401	400 400				
400	5	2N6545 MJ6503PNP MJ13015	400 400 400	MJE13007 MJE5852PNP	400 400	MJ10007	400
640	8	MJ13333	400	MJE13009	400	MJ10013	550
800	10	2N6547 MJ13333 MJ13334 MJ13335	40 400 450 500			MJ10005 MJ10008 MJ10009 MJ10013 MJ10014	400 450 500 550 600
1600	20					MJ10023 MJ10015 MJ10016	400 400 500

*NOTE: Power output ratings are for half-bridge circuit configurations, multiply by 2 for full-bridge.

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TABLE 3

CIRCUIT: HALF AND FULL* BRIDGE
LINE VOLTAGE: 380 VRMS
DEVICE V_{CEO} RATING ≥600V

Circuit Rating		Metal—TO-204**, TO-66		Plastic—TO-220AB, TO-126		Darlington—TO–204**	
Output Power* (Watts)	$I_{C(OP)}$ (Amps)	Device Type	Rated V_{CEO} (Volts)	Device Type	Rated V_{CEO} (Volts)	Device Type	Rated V_{CEO} (Volts)
240	2	MJ8500 MJ12002 MJ8501	700 750 800	MJE12007	750		
360	3	MJ8502 MJ12003 MJ8503	700 750 800				
480	4	MJ12004	750			MJ10011	700
600	5	MJ8504 MJ12005 MJ8505	700 750 800				
1200	10					MJ10014	600

*NOTE: Power output ratings are for half-bridge circuit configurations, multiply by 2 for full-bridge.

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TABLE 4

CIRCUIT: FORWARD, PUSH-PULL* AND FLYBACK*
LINE VOLTAGE: 120 VRMS
DEVICE V_{CEV} RATING ≥450 V

Circuit Rating		Metal—TO-204**, TO-66		Plastic—TO-220AB, TO-126		Darlington—TO-204**	
Output Power* (Watts)	$I_{C(OP)}$ (Amps)	Device Type	Rated V_{CEV} (Volts)	Device Type	Rated V_{CEV} (Volts)	Device Type	Rated V_{CEV} Volts
40	1	2N3585	450	MJE13002	600		
		2N6422PNP	450	MJE13003	700		
		2N6423PNP	450				
		2N4240	450				
80	2			MJE13004	600		
				MJE13005	700		
120	3	2N6306	500	2N6499	450		
		2N6307	600				
		2N6542	650				
		2N6308	700				
		2N6543	850				
200	5	MJ6503PNP	450	MJE5852PNP	450	MJ10005	450
		2N6544	650	MJE5740	600	MJ10007	500
		2N6545	850	MJE13006	600	MJ10012	550
				MJE5741	700		
				MJE13007	700		
				MJE5742	800		
320	8			MJE13008	600		
				MJE13009	700		
400	10	MJ13332	450			MJ10004	450
		MJ13333	500			MJ10005	500
		MJ13334	550			MJ10008	650
		MJ13335	600			MJ10009	750
		2N6546	650			MJ10008	650
						MJ10014	700
800	20					MJ10009	750
						MJ10015	600
						MJ10016	750

*NOTE: Power output ratings are for forward converter configurations (one transistor). Multiply by 2 for push-pull circuits and divide by 2 for flyback configurations.

**Formerly TO-3

TABLE 5

CIRCUIT: FORWARD, PUSH-PULL* AND FLYBACK*
LINE VOLTAGE: 220 VRMS
DEVICE V_{CEV} RATING $\geqslant$850 V

Circuit Rating		Metal—TO-204**, TO-66		Plastic—TO-220AB, TO-126		Darlington—TO-204**	
Output Power* (Watts)	$I_{C(OP)}$ (Amps)	Device Type	Rated V_{CEV} (Volts)	Device Type	Rated V_{CEV} (Volts)	Device Type	Rated V_{CEV} Volts
160	2	MJ8500	1200				
		MJ8501	1400				
		MJ12002	1500	MJE12007	1500		
240	3	2N6543	850				
		MJ8502	1200				
		MJ8503	1400				
		MJ12003	1500				
320	4	MJ12004	1500				
400	5	2N6545	850				
		MJ8504	1200				
		MJ8505	1400				
		MJ12005	1500			MJ10011	1500
560	7	MJ12010	950				
800	10	2N6547	850				

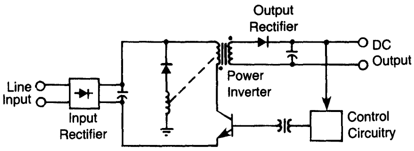
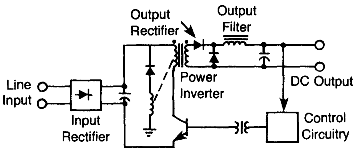
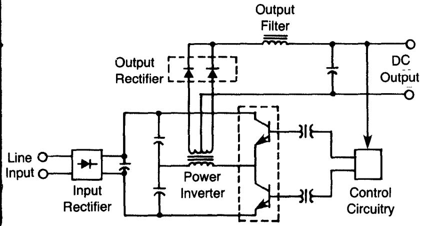
*NOTE: Power output ratings are for forward converter configurations (one transistor). Multiply by 2 for push-pull circuits and divide by 2 for flyback configurations.

**Formerly TO-3

APPENDIX B

MOTOROLA SWITCHMODE RECTIFIERS FOR SWITCHING POWER SUPPLIES

MOTOROLA SWITCHMODE INPUT RECTIFIERS

Total Supply Power	Typical Circuit	Standard Recovery for Line Voltage Operation			
		Input Current	Suggested Devices		
			Type	I_O	V_R
10 W	Flyback (Ringing-Choke)  	<1.0 A	1N4004 MDA104A MDA920A6	1.0 A 1.0 A 1.5 A	400 V
50 W		1.0 A	1N4004 MDA104A MDA920A6	1.0 A 1.0 A 1.5 A	
75 W		2.0 A	MR504 1N5404 MDA204	3.0 A 3.0 A 2.0 A	
	Basic Forward Converter  	2.0 A	MR504 1N5404 MDA204	3.0 A 3.0 A 2.0 A	
100 W		3.0 A	MR504 1N5404 MDA970A5	3.0 A 3.0 A 4.0 A	
250 W	Basic Half-Bridge Configuration  	6.0 A	MR754 1N1204,A,B,C MR1124 MDS804	6.0 A 12 A 12 A 8.0 A	
1000 W		12 A	1N1204,A,B,C MR1124 MDA1204 MR2004S	12 A 12 A 12 A 20 A	
2500 W	+ Full-Bridge and Push-Pull	25 A	MDA2504 MDA3504 1N1183,A	25 A 35 A 40 A	

MOTOROLA SWITCHMODE OUTPUT RECTIFIERS

Schottky for 5.0 V Outputs				Fast Recovery for >5.0 V Outputs			
Output Current	Suggested Devices			Output Current	Suggested Devices		
	Type	I_O	V_R		Type	I_O	V_R
1.0–2.0 A	1N5818	1.0 A	30 V	<0.5 A	1N4934	1.0 A	100 V ↓
	1N5821	3.0 A	30 V				
	MBR330M	3.0 A	30 V				
	MBR330M	3.0 A	30 V				
	1N5824	5.0 A	30 V				
5.0–10 A	1N5827	15 A	30 V	0.5–1.5 A	1N4934	1.0 A	
	MBR1530	15 A	30 V		MR851	1.0 A	
	1N5830	25 A	30 V		MR831	3.0 A	
	1N6095	25 A	30 V		MR801	3.0 A	
10–15 A	1N5830	25 A	30 V	1.5–2.5 A	MR851	3.0 A	
	MBR2535	25 A	35 V		MR821	5.0 A	
	SD41	30 A	35 V		MR831	3.0 A	
	MBR3535	35 A	35 V		MR801	3.0 A	
8.0–16 A	1N5827	15 A	30 V	2.0–2.5 A			
	MBR1530	15 A	30 V		1N4934	1.0 A	
	1N5830	25 A	30 V		MR851	3.0 A	
	1N6095	25 A	30 V		MR801	3.0 A	
	MBR3035CT	30 A	35 V				
10–20 A	1N5827	15 A	30 V	2.0–2.5 A			
	MBR1530	15 A	30 V		1N4934	1.0 A	
	1N5830	25 A	30 V		MR851	3.0 A	
	1N6095	25 A	30 V		MR801	3.0 A	
	MBR3035CT	30 A	35 V				
30–50 A	1N5830	25 A	30 V	2.0–8.0 A	1N4934	1.0 A	
	SD41	30 A	35 V		MR851	3.0 A	
	1N6095	25 A	30 V		MR821	5.0 A	
	MBR3535	35 A	35 V		1N3880,A	6.0 A	
	MBR3035CT	30 A	35 V		MDA2501FR	25 A	
200 A	SD51	60 A	35 V	40 A			
	MBR6035	60 A	35 V		1N3900	20 A	
	MBR7535	75 A	35 V		1N3910	30 A	
	1N6097	50 A	30 V		MDA3501FR	35 A	
	(IN PARALLEL)						
500 A	SD51	60 A	35 V	100 A	MR871	50 A	
	MBR6035	60 A	35 V				
	MBR7535	75 A	35 V				
	1N6097	50 A	30 V				
	(IN PARALLEL)						

FURTHER INFORMATION ON SWITCHING REGULATORS

1. ''100 kHz FET Switcher,'' R. Haver, Power Conversion International, April 1982.
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10639-4 PRINTED IN USA 6-82 IMPERIAL LITHO C06082 27,500