APPLICATIONS OF THE CA3085 SERIES MONOLITHIC IC VOLTAGE REGULATORS

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The Harris CA3085, CA3085A, and CA3085B monolithic IC's are positive-voltage regulators capable of providing output currents up to 100mA over the temperature range from -55°C to +125°C. They are supplied in 8 lead TO-5 type packages. The following tabulation shows some key characteristics and salient differences between devices in the CA3085 Series.

TYPE	V_{IN} (V_I) RANGE (V)	V_{OUT} (V_O) RANGE (V)	MAX. I_{OUT} (I_O) (mA)	MAX LOAD REGULATION (% V_O)
CA3085	7.5 - 30	1.8 - 26	12*	0.1
CA3085A	7.5 - 40	1.7 - 36	100	0.15
CA3085B	7.5 - 50	1.7 - 46	100	0.15

*This value may be extended to 100mA; however, regulation is not specified beyond 12mA.

In addition to these differences, the range of some specified performance parameters is more tightly controlled in the CA3085B than in the CA3085A, and more in the CA3085A than in the CA3085.

This note describes the basic circuit of the CA3085 series devices and some typical applications that include a high current regulator, constant current regulations, a switching regulator, a negative-voltage regulator, a dual-tracking regulator, high-voltage regulators, and various methods of providing current limiting, A circuit in which the CA3085 is used as a general-purpose amplifier is also shown.

Circuit Description

The block diagram of the CA3085 series circuits is shown in Figure 1. Fundamentally, the circuit consists of a frequency compensated error-amplifier which compares an internally generated reference voltage with a sample of the output voltage and controls a series-pass amplifier to regulate the output. The starting circuit assures stable latch-in of the voltage-reference circuitry. The current-limiting portion of the circuit is an optional feature that protects the IC in the event of overload.

Terminal 5 provides a source of stable reference voltage for auxiliary use; a current of about 250µA can be supplied to an external circuit without significantly disturbing reference-voltage stability. If necessary, filtering of the inherent noise of the reference-voltage circuit can be accomplished by connecting a suitable bypass capacitor between terminals 5 and 4.

Terminal 6 (the "inverting input" in accordance with operational-amplifier terminology) is the input through which a sample of the regulated output voltage is applied.

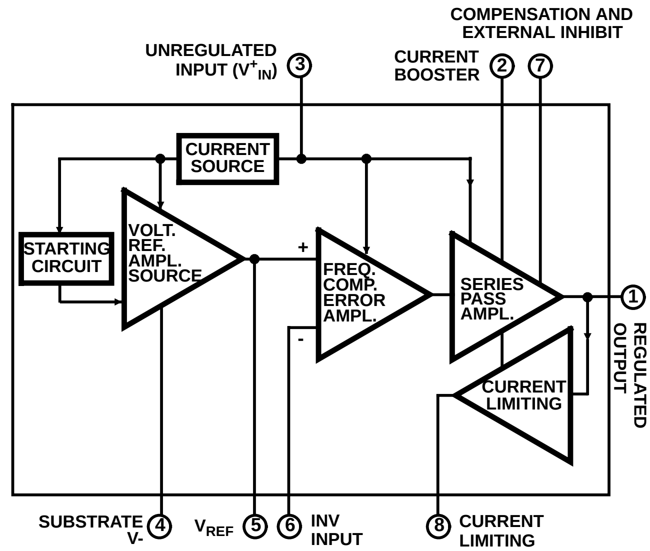


FIGURE 1. BLOCK DIAGRAM OF CA3085 SERIES

The collector of the series-pass output transistor is brought out separately at terminal 2 ("current booster") to provide base drive for an external p-n-p transistor; this approach is one method of regulating currents greater than 100mA.

Because the voltage regulator is essentially an operational amplifier having considerable feedback, frequency compensation is necessary in some circuits to prevent oscillations. Terminal 7 is provided for if external frequency compensation is necessary. Terminal 7 can also be used to "inhibit" (strobe, squelch, pulse, key) the operation of the series-pass amplifier.

Brief Description of CA3085 Schematic Diagram

The schematic diagram of the CA3085 series circuits is shown in Figure 2. The left-hand section includes the starting circuit, the voltage reference circuit, and the constant-current circuit. The center section is basically an elementary operational amplifier which serves as the voltage-error amplifier controlling the series-pass. Darlington pair (Q13, Q14) shown in the right-hand section when controlled by an appropriate external sensing network, transistor Q15, serves to provide protective current-limiting characteristics by diverting base drive from the series pass circuit. For operation at the highest current levels, terminals 2 and 3 are tied together to eliminate the voltage drop which would otherwise be developed across resistor R5.

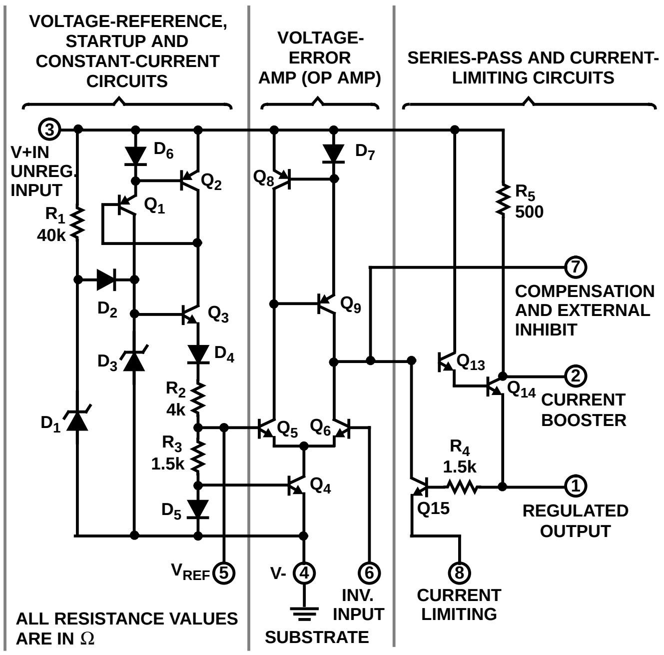


FIGURE 2. SCHEMATIC DIAGRAM OF CA3085 SERIES

Voltage Reference Circuits

The basic voltage referenced element used in the CA3085 is zener diode D3. It provides a nominal reference voltage of 5.5V and exhibits a positive temperature coefficient of approximately 2.5mV/^{o}C. If this reference voltage were used directly in conjunction with the error-amplifier (Q5, Q6, etc.), the IC would exhibit two major undesirable characteristics: (1) its performance with temperature variations would be poor, and (2) its use as a regulator would be restricted to circuits in which the minimum regulated output voltages are in excess of 5.5V. Consequently, it is necessary to provide means of compensating for the positive temperature coefficient of D3 and at the same time provide for obtaining a stable source of lower reference voltage. Both temperature compensation and the reduction of the reference voltage are accomplished by means of the series divider network consisting of the base-emitter junction of Q3, diode D4, resistors R2 and R3, and diode 5.

The voltage developed across D3 drives the divider network and a voltage of approximately 4V is developed between the cathode of D4 and the cathode of D5 (terminal 4). The current through this divider network is held nearly constant with temperature because of the combined temperature coefficients of the zener diode (D3), Q3 base-emitter junction, D4, D5, and the resistors R2 and R3. This constant current through the diode D5 and the resistor R3 produces a voltage drop between terminals 4 and 5 that results in the reference voltage (≈ 1.6V) having an effective temperature coefficient of about 0.0035%/^{o}C.

The reference diode D3 receives a currant of approximately 620µA from a constant-current circuit consisting of Q3 and the current-mirror* D6, Q1, and Q2. Current to start-up the constant-current source initially is provided by auxiliary zener diode D1 and R1. Diode D2 blocks current from the R1-D1 source after latch-in of the constant-current source establishes a stable reference potential, and thereby prevents modulation of the reference voltage by ripple voltage on the unregulated input voltage.

Voltage-Error Amplifier

Transistors Q5 and Q6 comprise the basic differential amplifier that is used as a voltage-error amplifier to compare the stable reference voltage applied at the base of Q5 with a sample of the regulator output voltage applied at terminal 6. The D5-Q4 combination is a current-mirror which maintains essentially constant-current flow to Q5 and Q6 despite variations in the unregulated input voltage. The Q8, Q9, and D7 network provides a "mirrored" active collector load for Q5 and Q6 and also provides a variable single-ended drive to the Q13 and Q14 series-pass transistors in accordance with the difference signal developed between the bases of Q5 and Q6. The open-loop gain of the error-amplifier is greater than 1000.

Series-Pass and Current-Limiting Circuits

In the normal mode of operation, or in the current-boost mode when terminals 2 and 3 are tied together, the Darlington pair Q13-Q14 performs the basic series-pass regulating function between the unregulated input voltage and the regulated output voltage at terminal 1. In the current-limiting mode transistor Q15 provides current-limiting to protect the CA3085 and/or limit the load current. To provide current-limiting protection, a resistor (e.g., 5Ω) is connected between terminals 1 and 8; terminal 8 becomes the source of regulated output voltage. As the voltage drop across this resistor increases, base drive is supplied to transistor Q15 so that it becomes increasingly conductive and diverts base drive from the Q13-Q14 pass transistor to reduce output current accordingly. Resistor R4 is provided to protect Q15 against overdrive by limiting its base current under transient and load-short conditions.

Because the CA3085 regulator is essentially an op-amp having considerable feedback, frequency compensation may be required to prevent oscillations. Stability must also be maintained despite line and load transients, even during operation into reactive loads (e.g., filter capacitors). Provisions are included in the CA3085 so that a small-value capacitor may be connected between terminals 6 and 7 to compensate the regulator, when necessary, by "rolling-off" the amplifier frequency-response. Terminal 7 is also used to externally "inhibit" operation of the CA3085 by diverting base current supplied to Q13-Q14, thereby permitting the use of keying, strobing, programming, and/or auxiliary overload-protection circuits.

Applications

A Simple Voltage Regulator

Figure 3 shows the schematic diagram of a simple regulated power supply using the CA3085. The ac supply voltage is stepped down by T1, full-wave rectified by the diode bridge circuit, and smoothed by the large electrolytic capacitor C1 to provide unregulated dc to the CA3085 regulator circuit. Frequency compensation of the error-amplifier is provided by capacitor C2. Capacitor C3 bypasses residual noise in the reference-voltage source, and thus decreases the incremental noise-voltage in the regulator circuit output.

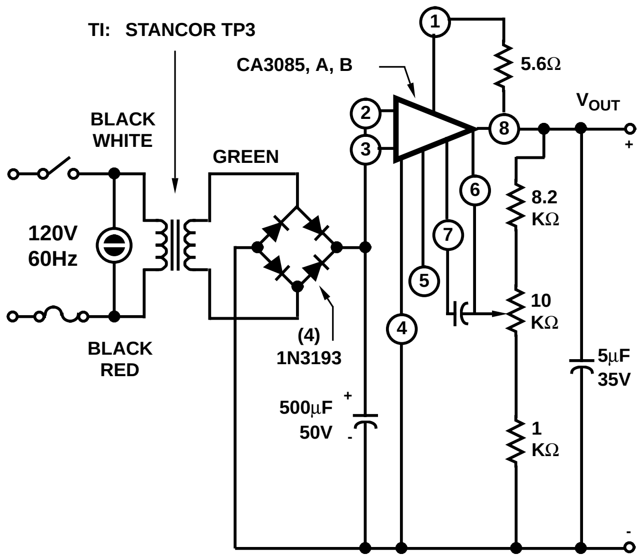


FIGURE 3. BASIC POWER SUPPLY

Because the open-loop gain of the error-amplifier is very high (greater than 1000), the output voltage may be directly calculated from the following expression:

$$V_O = \frac{(R2 + R1)}{R1} V_{REF} \quad \text{(EQ. 1)}$$

In the circuit shown in Figure 3, the output voltage can be adjusted from 1.8V to 20V by varying R2. The maximum output current is determined by R_{SC}; load-regulation characteristics for various values of R_{SC} are shown in Figure 4.

When this circuit is used to provide high output currents at low output voltages, care must be exercised to avoid excessive IC dissipation. In the circuit of Figure 3, this dissipation control can be accomplished by increasing the primary-to-secondary transformer ratio (a reduction in V_I) or by using a dropping resistor between the rectifier and the CA3085 regulator. Figure 5 gives data on dissipation limitation (V_I - V_O vs. I_O) for CA3085 series circuits. The short-circuit current is determined as follows:

$$I_{SC} = \frac{V_{BE}}{R_{SC}} \approx \frac{0.7}{R_{SC}} \text{ amperes} \quad \text{(EQ. 2)}$$

The line-and-load regulation characteristics for the circuit shown in Figure 3 are approximately 0.05 percent of the output voltage.

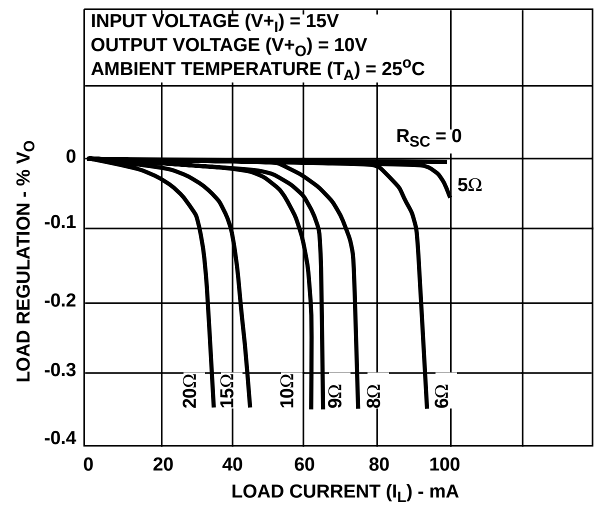


FIGURE 4. LOAD REGULATION CHARACTERISTICS FOR CIRCUIT OF FIGURE 3

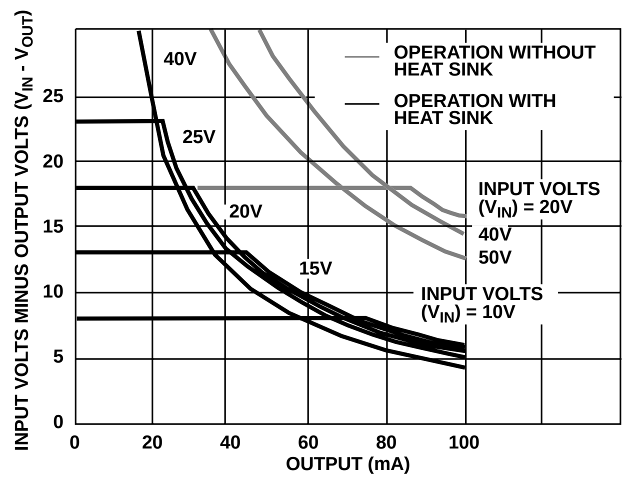


FIGURE 5. DISSIPATION LIMITATION (V_I - V_O vs I_O) FOR CA3085 SERIES CIRCUITS

High-Current Voltage Regulator

When regulated voltages at currents greater than 100mA are required, the CA3085 can be used in conjunction with an external n-p-n pass transistor as shown in the circuits of Figure 6. In these circuits the output current available from the regulator is increased in accordance with the h_{FE} of the external n-p-n pass transistor. Output currents up to 8A can be regulated with these circuits. A Darlington power transistor can be substituted for the 2N5497 transistor when currents greater than 8A are to be regulated.

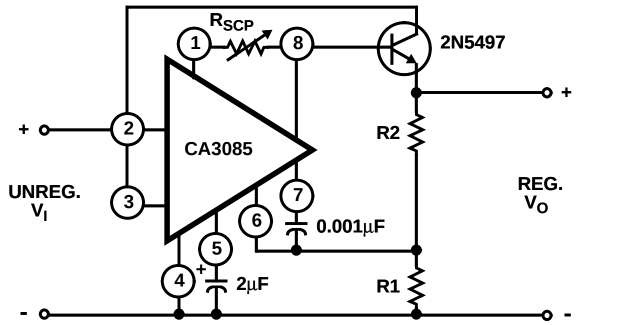


FIGURE 6A. WITH SIMPLIFIED SHORT-CIRCUIT PROTECTION

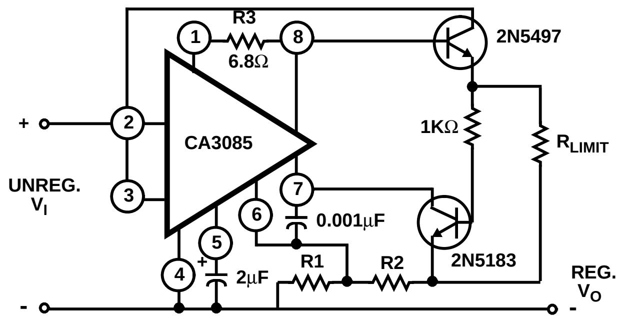


FIGURE 6B. WITH AUXILLIARY SHORT-CIRCUIT PROTECTION

FIGURE 6. HIGH-CURRENT VOLTAGE REGULATOR USING n-p-n PASS TRANSISTOR

A simplified method of short circuit protection is used in connection with the circuit of Figure 6A. The variable resistor R_{SCP} serves two purposes: 1) it can be adjusted to optimize the base drive requirements (h_{FE}) of the particular 2N5497 transistor being used, and 2) in the event of a short circuit in the regulated output voltage the base drive current in the 2N5497 will increase, thereby increasing the voltage drop across R_{SCP}. As this voltage drop increases the short circuit protection system within the CA3085 correspondingly reduces the output current available at terminal 8, as described previously. It should be noted that the degree of short circuit protection depends on the value of R_{SCP}, i.e., design compromise is required in choosing the value of R_{SCP} to provide the desired base drive for the 2N5497 while maintaining the desired short circuit protection. Figure 6B shows an alternate circuit in which an additional transistor (2N2102) and two resistors have been added as an auxiliary short circuit protection feature. Resistor R3 is used to establish the desired base drive for the 2N5497, as described above. Resistor R_{LIMIT} now controls the short circuit output current because, in the event of a short circuit, the voltage drop developed across its terminals increases sufficiently to increase the base drive to the 2N2102 transistor. This increase in base drive results in reduced output from the CA3085 because collector current flow in the 2N2102 diverts base drive from the Darlington output stage of the CA3085 (see Figure 2) through terminal 7. The load regulation of this circuit is typically 0.025 per cent with 0 to 3A load-current variation; line regulation is typically 0.025%/V change in input voltage.

Voltage Regulator with Low V_I - V_O Difference

In the voltage regulators described in the previous section, it is necessary to maintain a minimum difference of about 4V between the input and output voltages. In some applications this requirement is prohibitive. The circuit shown in Figure 7 can deliver an output current in the order of 2A with a V_I - V_O difference of only one volt.

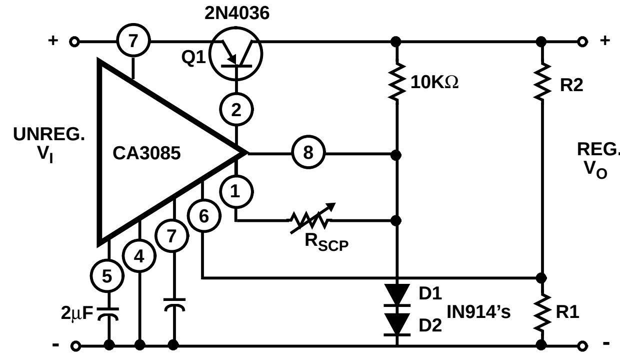


FIGURE 7. VOLTAGE REGULATOR FOR LOW V_I - V_O DIFFERENCE

It employs a single external p-n-p transistor having its base and emitter connected to terminals 2 and 3, respectively, of the CA3085. In this circuit, the emitter of the output transistor (Q14 in Figure 2) in the CA3085 is returned to the negative supply rail through an external resistor (R_{SCP}) and two series-connected diodes (D1, D2). These forward biased diodes maintain Q6 in the CA3085 within linear-mode operation. The choice of resistors R1 and R2 is made in accordance with Equation 1. Adequate frequency compensation for this circuit is provided by the 0.01µF capacitor connected between terminal 7 of the CA3085 and the negative supply rail.

Figure 8 which shows the output impedance of the circuit of Figure 7 as a function of frequency, illustrates the excellent ripple-rejection characteristics of this circuit at frequencies below 1kHz. Lower output impedances at the higher frequencies can be provided by connecting an appropriate capacitor across the output voltage terminals. The addition of a capacitor will, however, degrade the ability of the system to react to transient-load conditions.

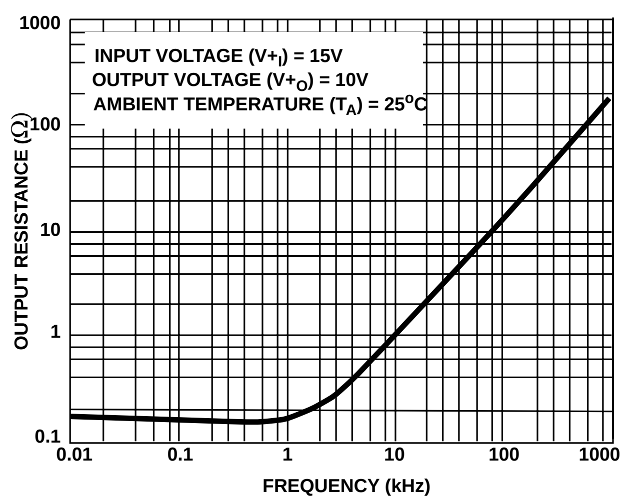


FIGURE 8. OUTPUT RESISTANCE vs FREQUENCY FOR CIRCUIT OF FIGURE 7

High Voltage Regulator

Figure 9 shows a circuit that uses the CA3085 as a voltage reference and regulator control device for high-voltage power supplies in which the voltages to be regulated are well above the input-voltage ratings of the CA3085 series circuits. The external transistors Q1 and Q2 require voltage ratings in excess of the maximum input voltage to be regulated, Series-pass transistor Q2 is controlled by the collector current of Q1, which in turn is controlled by the normally regulated current output supplied by the CA3085. The input voltage for the CA3085 regulator at terminal 3 is supplied through dropping resistor R3 and the clamping zener diode D1. The values for resistor R1 and R2 are determined in accordance with Equation 1.

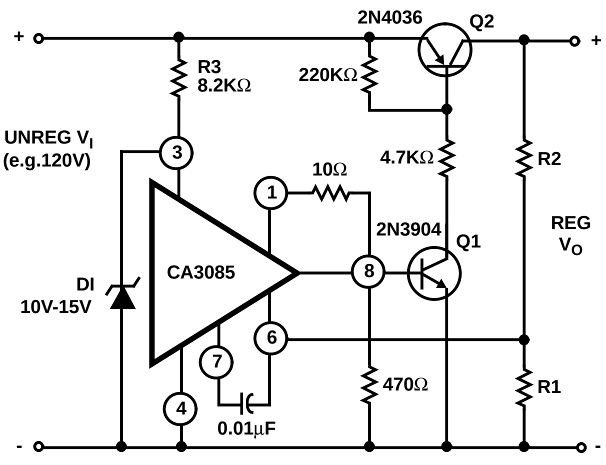


FIGURE 9. HIGH VOLTAGE REGULATOR

Negative Voltage Regulator

The CA3085 is used as a negative-supply voltage regulator in the circuit shown in Figure 10. Transistor Q3 is the series pass transistor. It should be noted that the CA3085 is effectively connected across the load side of the regulated system. Diode D1 is used initially in a "circuit-starter" function; transistor Q2 "latches" D1 out of its starter-circuit function so that the CA3085 can assume its role in controlling the pass transistor Q3 by means of Q1.

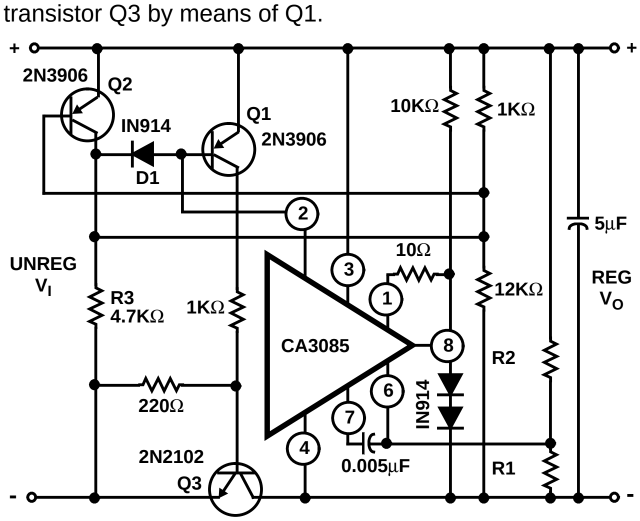


FIGURE 10. NEGATIVE VOLTAGE REGULATOR

Operation of the circuit is as follows: current through R3 and D1 provides base drive for Q1, which in turn provides base drive for the pass-transistor Q3. By this means operating potential for the CA3085 is developed between the collector of Q3 (terminal 4 of the CA3085) and the positive supply-rail (terminal 3 of the CA3085). When the output voltage has risen sufficiently to maintain operation of the CA3085 (approx. 7.5V), transistor Q2 is driven into conduction by the base drive supplied from the 1KΩ-12KΩ voltage divider. As Q2 becomes conductive, it diverts the base drive being supplied to Q1 through the R3-D1 path, and diode D1 ceases to conduct. Under these conditions, base-current drive to Q1 through terminal 2 of the CA3085 regulates the base drive to Q3. Values of R1 and R2 are determined in accordance with Equation 1.

The circuit shown in Figure 11 is similar to that of Figure 10, except for the addition of a constant-current limiting circuit consisting of transistor Q4, a 1KΩ resistor, and resistor R_{SCP}. When the load current increases above a particular design value, the corresponding increase in the voltage drop across resistor R_{SCP} provides additional base drive to transistor Q4. Thus, as transistor Q4 becomes increasingly conductive, its collector current diverts sufficient base drive from Q3 to limit the current in the pass transistor feeding the regulated load. With the types of transistors shown in Figures 10 and 11, maximum currents in the order of 5A can be regulated.

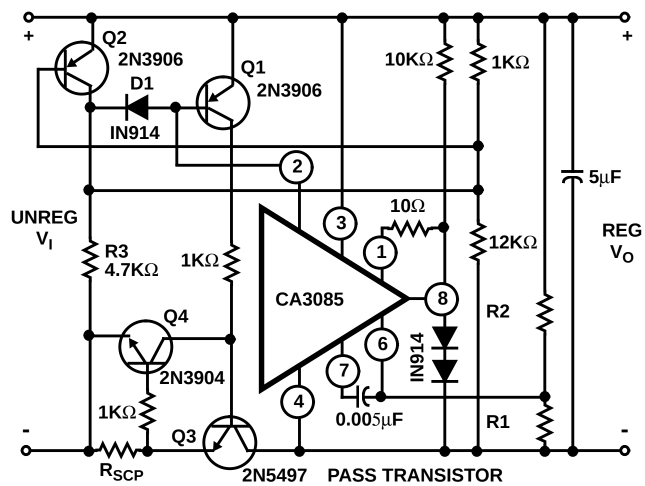


FIGURE 11. NEGATIVE VOLTAGE REGULATOR WITH CONSTANT CURRENT LIMITING CIRCUIT

High-Output-Current Voltage Regulator With "Foldback" Current-Limiting (Also known as "Switch-Back" Current-Limiting)

In high-current voltage regulators employing constant current limiting (e.g., Figures 6 and 7), it is possible to develop excessive dissipation in the series-pass transistor when a short circuit develops across the output terminals. This situation can be avoided by the use of the "foldback" current-limiting circuitry as shown in Figure 12. In this circuit, terminal 8 of the CA3085 senses the output voltage, and terminal 1 is tied to a tap on a voltage-divider network connected between the emitter of the pass-transistor (Q3) and ground. The current-foldback trip-point is established by the value of resistor R_{SC}.

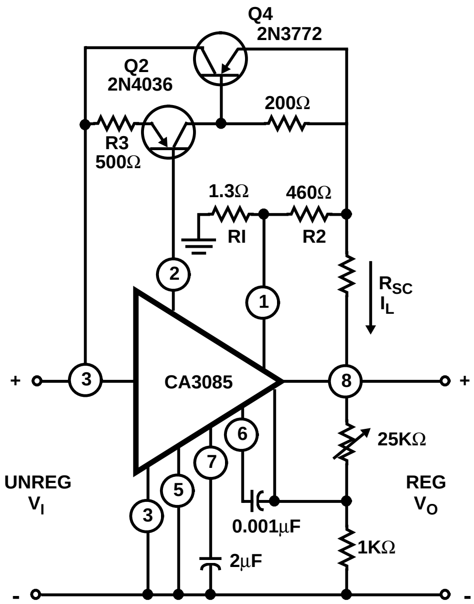


FIGURE 12. HIGH OUTPUT CURRENT VOLTAGE REGULATOR WITH "FOLDBACK" CURRENT LIMITING

The protective tripping action is accomplished by forward-biasing Q15 in the CA3085 (see Figure 2). Conditions for tripping circuit operation are defined by the following expressions:

$$V_{BE(Q15)} = \text{(voltage at terminal 1) - (output voltage)}$$

$$= \left[(V_O + I_L R_{SC}) \frac{R1}{R1 + R2} \right] - V_O \qquad \text{(EQ. 3)}$$

$$\text{If} \quad \frac{R1}{R1 + R2} = K, \text{ then}$$

$$V_{SE(Q15)} = (V_O + I_L R_{SC}) K - V_O = KV_O + KI_L R_{SC} - V_O$$

and therefore

$$R_{SC} = \frac{V_O + V_{BE(Q15)} - KV_O}{KI_L} \qquad \text{(EQ. 4)}$$

Under load short-circuit conditions, terminal 8 is forced to ground potential and current flows from the emitter of Q14 in the CA3085, establishing terminal 1 at one V_{SE}-drop [≅0.7V] above ground and Q15 in a partially conducting state. The current through Q14 necessary to establish this one V_{SE} condition is the sum of currents flowing to ground through R1 and [R2 + R_{SC}]. Normally R_{SC} is much smaller than R2 and can be ignored; therefore, the equivalent resistance R_{eq} to ground is the parallel combination of R1 and R2.

The Q14 current is then given by:

$$I_{Q14} = \frac{V_{BE(Q15)}}{R_{EQ}} = \frac{V_{BE(Q15)}}{\frac{R1R2}{R1 + R2}} = \frac{0.7\,[1.3 + 0.46]}{1.3 \times 0.46} \quad 2.06\text{mA} \qquad \text{(EQ. 5)}$$

This current provides a voltage between terminals 2 and 3 as follows: Ω

$$V_{2\text{-}3} = I_{Q14} \times 250\Omega = 2.06 \times 10^{-3} \times 250 = 0.515\text{V} \qquad \text{(EQ. 6)}$$

The effective resistance between terminals 2 and 3 is 250Ω because the external 500Ω resistor R3 is in parallel with the internal 500Ω resistor R5. It should be understood that the $V_{2\text{-}3}$ potential of 0.515V is insufficient to maintain the external p-n-p transistor Q2 in conduction, and, therefore, Q3 has no base drive. Thus the output current is reduced to zero by the protective circuitry. Figure 13 shows the foldback characteristic typical of the circuit of Figure 12.

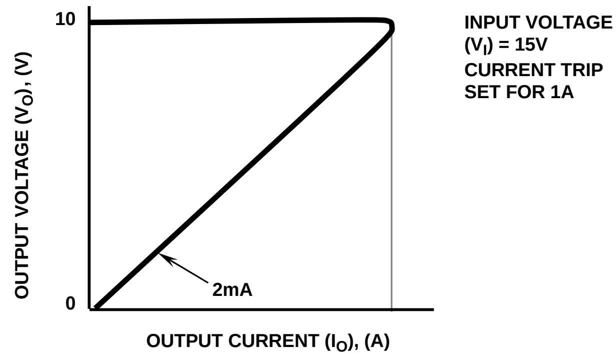


FIGURE 13. TYPICAL "FOLDBACK" CURRENT-LIMITING CHARACTERISTIC FOR CIRCUIT OF FIGURE 12

An alternative method of providing "foldback" current-limiting is shown in Figure 14. The operation of this circuit is similar to that of Figure 12 except that the foldback-control transistor Q2 is external to the CA3085 to permit added flexibility in protection-circuit design.

Under low load conditions Q2 is effectively reverse-biased by a small amount, depending upon the values of R3 and R4. As the small amount, depending upon the values of R3 and R4. As the load current increases the voltage drop across R_{trip} increases, thereby raising the voltage at the base of Q1, and Q2 starts to conduct. As Q2 becomes increasingly conductive it diverts base current from transistors Q13 and Q14 in the CA3085, and thus reduces base drive to the external pass-transistor Q1 with a consequent reduction in the output voltage. The point at which current-limiting occurs, I_{tnp}, is calculated as follows:

$V_{BE(Q1)}$ = voltage at terminal 8 - V_O (assuming a low value for R_{TRIP})

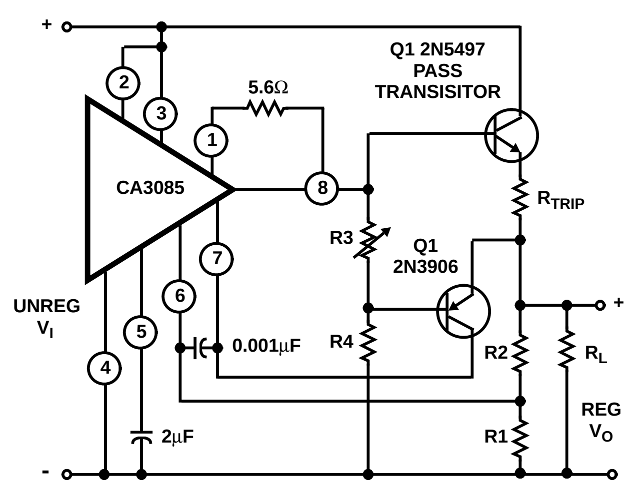


FIGURE 14. HIGH-OUTPUT-CURRENT VOLTAGE REGULATOR USING AUXILIARY TRANSISTOR TO PROVIDE "FOLDBACK" CURRENT LIMITING

$$V_{BE(Q2)} = \text{voltage at terminal 8}\left(\frac{R4}{R3+R4}\right) - V_O$$

$$= \left[V_O + I_L R_{TRIP} + V_{BE(Q1)}\right]\left[\frac{R4}{R3+R4}\right] - V_O$$

if $K = \frac{R4}{R3+R4}$, then the trip current is given by:

$$I_{trip} = \frac{V_{BE(Q2)} - K[V_O + V_{BE(Q1)}] + V_O}{KR_{TRIP}} \quad \text{(EQ. 7)}$$

In the circuit in Figure 12 the load current goes to zero when a short circuit occurs. In the circuit of Figure 14 the load current is significantly reduced but does not go to zero. The value for I_{SC} is computed as follows:

$$V_{BE(Q2)} + \left[\frac{V_{BE(Q2)}}{R2} + I_{B(Q2)}\right] R1 = V_{BE(Q1)} + I_{SC}R_{TRIP}$$

$$I_{SC} = \frac{V_{BE(Q2)} + \left[\frac{V_{BE(Q2)}}{R2} + I_{B(Q2)}\right] R1 = V_{BE(Q1)}}{R_{TRIP}} \quad \text{(EQ. 8)}$$

Figure 15 shows that the transfer characteristic of the load current is essentially linear between the “trip-point” and the “short-circuit” point.

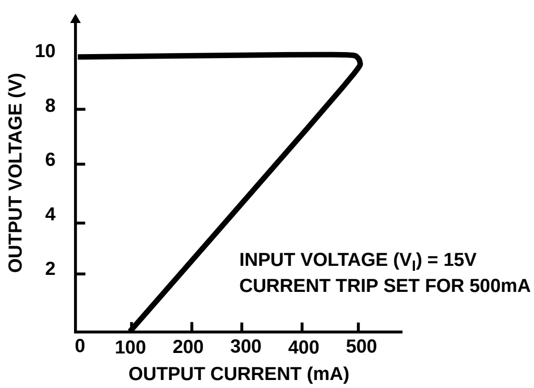


FIGURE 15. TYPICAL FOLDBACK CURRENT-LIMITING CHARACTERISTIC FOR CIRCUIT OF FIGURE 14

High-Voltage Regulator Employing Current “Snap-Back” Protection

In high-voltage regulators (e.g., see Figure 9), “foldback” current-limiting cannot be used safely because the high voltage across the pass transistor can cause second breakdown despite the reduction in current flow. To adequately protect the pass transistor in this type of high-voltage regulator, the so-called “snap-back” method of current limiting can be employed to reduce the current to zero in a few microseconds, and thus prevent second-breakdown destruction of the device.

The circuit diagram of a high-voltage regulator employing current “snap-back” protection is shown in Figure 16. The basic regulator circuit is similar to that shown in Figure 9. The additional circuitry in the circuit of Figure 16 quickly interrupts base drive to the pass transistor in event of load fault. The point of current-trip is established as follows:

$$I_{trip} = \frac{V_{BE(Q1)}}{R_{SC}} \quad \text{(EQ. 9)}$$

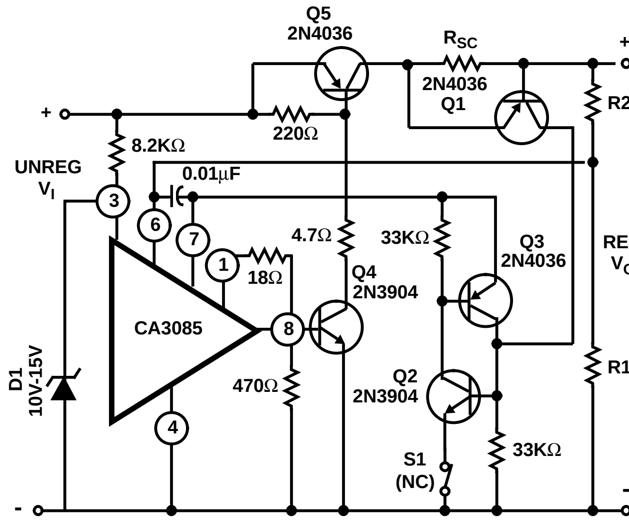


FIGURE 16. HIGH-VOLTAGE REGULATOR INCORPORATING CURRENT “SNAP-BACK” PROTECTION

Thus, when a sufficient voltage drop is developed across R_{SC}, transistor Q1 becomes conductive and current flows into the base of Q2 so that it also becomes conductive. Transistor Q3, in turn, is driven into conduction, thereby latching the Q2-Q3 combination (basic SCR action) so that it diverts (through terminal 7) base drive from the output stage (Q13, Q14) in the CA3085. By this means, base drive is diverted from Q4 and the pass transistor Q5. To restore regulator operation, normally closed switch S1 is momentarily opened and unlatches Q2-Q3.

Switching Regulator

When large input-to-output voltage differences are necessary, the regulators described above are inefficient because they dissipate significant power in the series-pass transistor. Under these conditions, high-efficiency operation can be achieved by using a switching-type regulator of the generic type shown in Figure 17A. Transistor Q1 acts as a keyed switch and operates in either a saturated or cut-off condition to minimize dissipation. When transistor Q1 is conductive, diode D1 is reversed-biased and current in the inductance L1 increases in accordance with the following relationship:

$$i_L = \frac{1}{L}\int_{t_0}^{t_1} V\,dt \quad \text{(EQ. 10)}$$

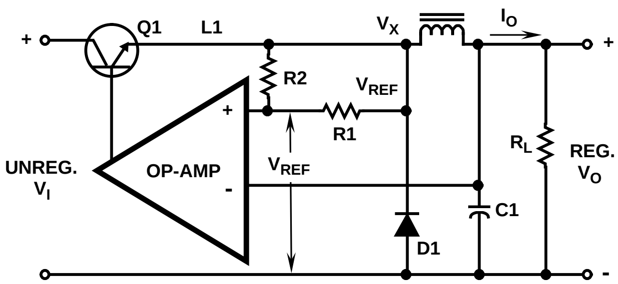


FIGURE 17A. SELF-OSCILLATING SWITCHING REGULATOR

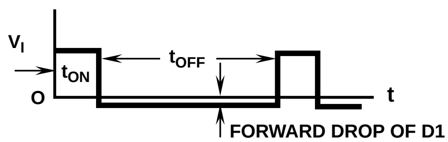


FIGURE 17B. VOLTAGE AT POINT V_X

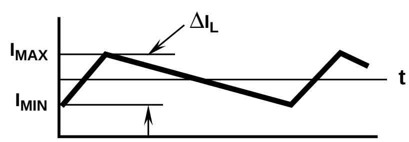


FIGURE 17C. INDUCTOR CURRENT I_{LI}

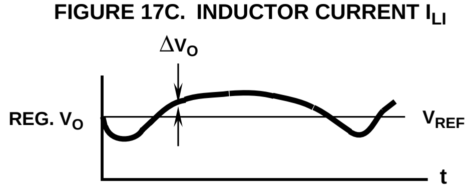


FIGURE 17D. OUTPUT VOLTAGE

FIGURE 17. SWITCHING REGULATOR AND ASSOCIATED WAVEFORMS

Where V is the voltage across the inductance L1. The current through the inductance charges the capacitor C1 and supplies current to the load. The output voltage rises until it slightly exceeds the reference voltage V_{ref}. At this point the op-amp removes base drive to Q1 and the unregulated input voltage V1 is "switched off". The energy stored in the inductor L1 now causes the voltage at V_X to swing in the negative direction and current flows through diode D1, while continuing to supply current into the load R_L. As the current in the inductor falls below the load current, the capacitor C1 begins to discharge and V_O decreases. When V_O falls slightly below the value of V_{REF}, the op-amp turns on Q1 and the cycle is repeated. It should be apparent that the output voltage oscillates about V_{REF} with an amplitude determined by R1 and R2. Actually, the value of V_{REF} varies from being slightly more positive than V_{REF} when Q1 is conducting, to being slightly more negative than V_{REF} when D1 is conducting. The voltage and current waveforms are shown in Figure 17B, C, and D.

Design Example

The following specifications are used in decomputations for a switching regulator:

V_I = 30V, V_O = 5V, I_O = 500mA,
switching frequency = 20kHz,
output ripple = 100mV

If it is assumed that transistor Q1 is in steady-state saturated operation with a low voltage-drop, the current in the inductor is given by Eq.10, as follows:

$$i_L = \frac{1}{L}\int_{t_0}^{t1} V dt = \left(\frac{V_I - V_O}{L_1}\right) t_{ON} \quad \text{(EQ. 11)}$$

When transistor Q1 is off, the current in the inductor is given by:

$$i_L \cong \frac{(V_O + V_{D1})\, t_{OFF}}{L1} \quad \text{(EQ. 12)}$$

From Equation 11,

$$L_1 = \frac{(V_I - V_O)}{i_L} \cdot \frac{1}{f} \cdot \frac{V_O}{V_I} \quad \text{(EQ. 13)}$$

If i_{max} is 1.3 I_L, then during t_{on} the current in the inductor (i_L) will be 0.5A x 1.3 = 0.65A; therefore, Δi_L = 0.15A.

Substitution in Equation 13 yields

$$L_1 = \frac{(30 - 5)}{0.15} \cdot \frac{1}{(20 \times 10^3)} \cdot \frac{5}{30} . \quad 1.4\text{mH} \quad \text{(EQ. 14)}$$

Current discharge from the capacitor C1 is given by:

$$i_C = C \frac{dv}{dt}$$

$$\text{Thus, } \Delta i_C = C \frac{\Delta v}{\Delta t} \quad \text{or } C = \frac{\Delta_{IC}\, \Delta t}{\Delta v}$$

Since $i_C = i_L$ and $\Delta t = t_{OFF}$, then

$$C = \frac{\Delta i_L\, t_{OFF}}{\Delta v} \quad \text{(EQ. 15)}$$

Substitution for the value of iL from Equation 13 yields

$$C = \frac{\left(\frac{V_I - V_O}{L1}\right) \cdot \frac{1}{f} \cdot \left(\frac{V_O}{V_I}\right) . \quad t_{OFF}}{\Delta v} \quad \text{(EQ. 16)}$$

The total period $T = t_{OFF} + t_{ON}$, and $T = \frac{1}{f}$ Therefore,

$$t_{OFF} = \frac{1}{f} - t_{ON} \quad \text{(EQ. 17)}$$

For optimum efficiency t_{on} should be

$$\cong \left(\frac{V_O}{V_I}\right) T \cong \left(\frac{V_O}{V_I}\right) \frac{1}{f} \quad \text{(EQ. 18)}$$

Substitution for t_{on} in Equation 18 yields

$$t_{OFF} = \frac{1}{f} - \left(\frac{V_O}{V_I}\right)\frac{1}{f} = \frac{1}{f}\left(1 - \frac{V_O}{V_I}\right) \quad \text{(EQ. 19)}$$

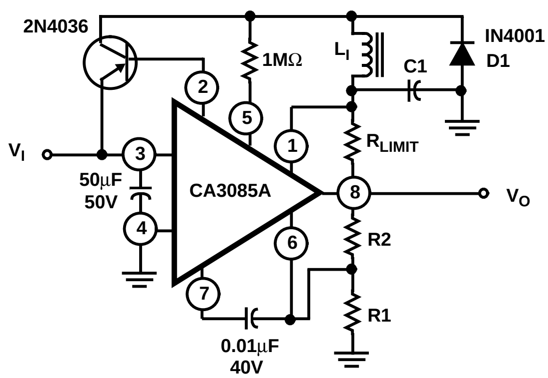
Substitution for t_{on} in Equation 16 yields

$$C = \frac{\frac{(V_I - V_O)}{L_1} \cdot \frac{1}{f} \cdot \frac{V_O}{V_I} \cdot \frac{1}{f} \cdot \left(1 - \frac{V_O}{V_I}\right)}{\Delta v} \qquad \text{(EQ. 20)}$$

Substitution of numerical values in Equation 20 produces the following value for C:

$$C = \frac{\frac{30 - 5}{1.4 \times 10^{-3}} \cdot \frac{1}{20 \times 10^{3}} \cdot \frac{5}{30} \cdot \frac{1}{20 \times 10^{3}} \cdot \left(1 - \frac{5}{30}\right)}{10^{-1}} = 63\mu F$$

A switching-regulator circuit using the CA3085 is shown in Figure 18. The values of L and C (1.5mH and 50mF, respectively) are commercially available components having values approximately equal to the computed values in the previous design example.



$$R_{LIMIT} = \frac{0.7}{1.2\ I_{L\,(MAX)}} \qquad V_O = V_{REF.}\left(\frac{R1 + R1}{R1}\right)$$

FIGURE 18. TYPICAL SWITCHING REGULATOR CIRCUIT

Current Regulators

The CA3085 series of voltage regulators can be used to provide a constant source or sink current. A regulated-current supply capable of delivering up to 100mA is shown in Figure 19A . The regulated load current is controlled by R1 because the current flowing through this resistor must establish a voltage difference between terminals 6 and 4 that is equal to the internal reference voltage developed between terminals 5 and 4.

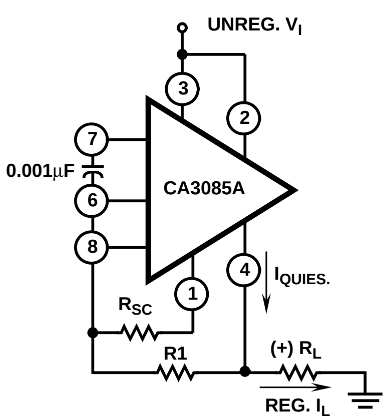
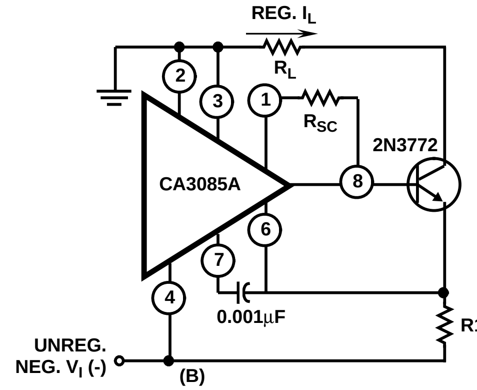


FIGURE 19A. CURRENT REGULATOR



HIGH-CURRENT REGULATOR

FIGURE 19B. HIGH-CURRENT REGULATOR

FIGURE 19. CONSTANT CURRENT REGULATORS

The actual regulated current, reg I_L is the sum of the quiescent regulator current and the current through R1, i.e.,

$$\text{reg } I_L = I_{QUIESCENT} + I_{R1}$$

Figure 19B shows a high-current regulator using the CA3085 in conjunction with an external n-p-n transistor to regulate currents up to 3A. In this circuit the quiescent regulator current does not flow through the load and the output current can be directly programmed by R1, i.e.,

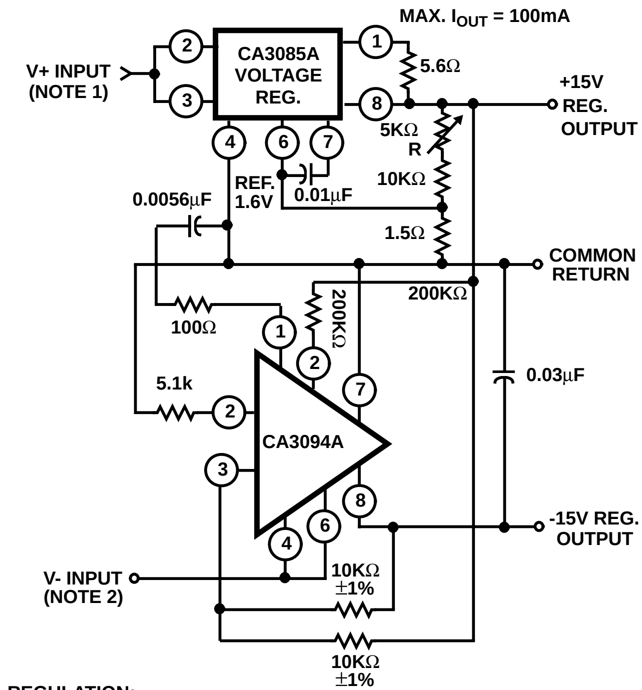
$$\text{Reg } I_L = \frac{V_{REF}}{R1}$$

With this regulator currents between 1mA and 3A can be programmed directly. At currents below 1mA inaccuracies may occur as a result of leakage in the external transistor.

A Dual-Tracking Voltage Regulator

A dual-tracking voltage regulator using a CA3085 and a CA3094A is shown in Figure 20. The CA3094A is basically an op-amp capable of supplying 100mA of output current. Specifications for the CA3094A appear in datasheet file number 598.

The positive output voltage is regulated by a CA3085 operating in a configuration essentially similar to that described in connection with Figure 3. Resistor R is used as a vernier adjustment of output voltage. The negative output voltage is regulated by the CA3094A, which is “slaved” to the regulated positive voltage supplied by the CA3085. It should be noted that the non-inverting input of the CA3094A and the negative supply terminal of the CA3085 are connected to a common ground reference. The “slaving” potential for the CA3094A is derived from an accurate 1:1 voltage-divider network comprised of two 10KΩ resistors connected between the +15V and the -15V output terminals. The junction of these two resistors is connected to the inverting input of the CA3094A. The voltage at this junction is compared with the voltage at the non-inverting input, and the CA3094A then automatically adjusts the output current at the negative terminal to maintain a negative regulated output voltage essentially equal to the regulated positive output voltage. Typical performance data for this circuit are shown in Figure 20.



REGULATION:

$$\text{MAX. LINE} = \frac{\Delta V_{OUT}}{\left[V_{OUT}(\text{INITIAL})\right]\Delta V_{IN}} \times 100 = 0.075\%/\text{V}$$

$$\text{MAX. LOAD} = \frac{\Delta V_{OUT}}{\left[V_{OUT}(\text{INITIAL})\right]\Delta V_{IN}} \times 100 = 0.075\%/V_{OUT}$$

(I_L FROM 1µA TO 50mA)

NOTE:

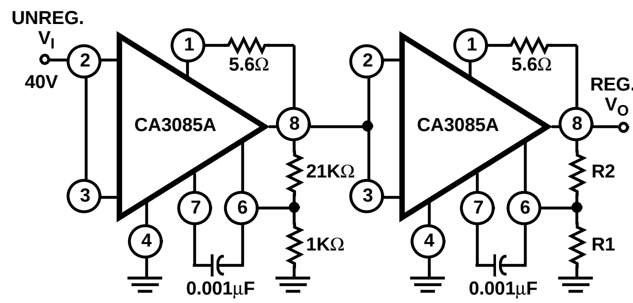
1. V+ Input Range = 19V to 30V for 15V Output
2. V- Input Range = -16V to -30V for -15V Output

FIGURE 20. DUAL-VOLTAGE TRACKING REGULATOR

The basic circuit of Figure 20 can be modified to regulate dissimilar positive and negative voltages (e.g., +15V, -5V) by appropriate selection of resistor ratios in the voltage-divider network discussed previously. As an example, to provide tracking of the -15V and -5V regulated voltages with the circuit of Figure 20, it is only necessary to replace the 10KΩ resistor connected between terminals 3 and 8 of the CA3094A with a 3.3KΩ resistor.

Regulators With High Ripple Rejection

When the reference-voltage source in the CA3085 is adequately filtered, the typical ripple rejection provided by the circuit is 56dB. It is possible to achieve higher ripple-rejection performance by cascading two stages of the CA3085, as shown in Figure 21. The voltage-regulator circuit in Figure 21A provides 90dB of ripple rejection. The output voltage is adjustable over the range from 1.8V to 30V by appropriate adjustment of resistors R1 and R2. Higher regulated output currents up to 1A can be obtained with this circuit by adding an external n-p-n transistor as shown in Figure 21B.



90dB RIPPLE REJECTION
LINE REG. <0.0001%/ V_I
LOAD REG. <0.1% V_O FOR LOAD CURRENTS UP TO 50mA
V_O RANGE FROM 1.8V TO 30V

FIGURE 21A. VOLTAGE REGULATOR WITH HIGH RIPPLE REJECTION

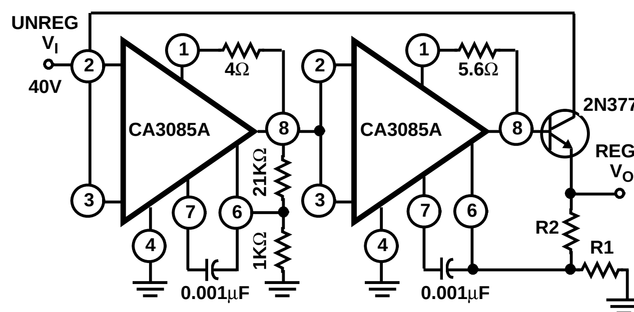


FIGURE 21B. HIGH-CURRENT VOLTAGE REGULATOR WITH HIGH RIPPLE REJECTION

The CA3085 As A Power Source For Sensors

Certain types of sensor applications require a regulated power source. Additionally, low-impedance sensors can consume significant power. An example of a circuit with these requirements, in which a CA3085 provides regulated power for a low-impedance sensor and the CA3059 zero-voltage switch, is shown in Figure 22. Terminal 12 on the CA3059 provides the ac trigger-signal which actuates the zero-voltage switch synchronously with the power line to control the load-switching triac. Specifications for the CA3059 appear in datasheet file number 490.

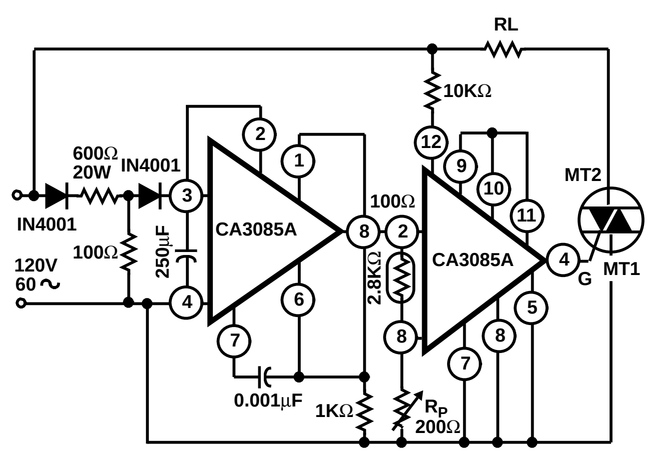


FIGURE 22. VOLTAGE REGULATOR FOR SENSOR AND ZERO-VOLTAGE SWITCH

The CA3085 As A General-Purpose Amplifier

As described above, the CA3085 series regulators contain a high-gain linear amplifier having a current-output capability up to 100mA. The premium type (CA3085B) can operate at supply voltages up to 50V. When equipped with an appropriate radiator or heat sink, the TO-5 package of these devices can dissipate up to 1.6W at 55^{o}C. A very stable internal voltage-reference source is used to bias the high-gain amplifier and/or provide an external voltage-reference despite extreme temperature or supply-voltage variations. These factors, plus economics, prompt consideration of this circuit for general-purpose uses, such as amplifiers, relay controls, signal-lamp controls, and thyristor firing.

As an example, Figure 23 shows the application of the CA3085 in a general-purpose amplifier. Under the conditions shown, the circuit has a typical gain of 70bB with a flat response to at least 100kHz without the RC network connected between terminals 6 and 7. The RC network is useful as a tone control or to “roll-off” the amplifier response for other reasons. Current limiting is not used in this circuit. The network connected between terminals 8 and 6 provides both dc and ac feedback. This circuit is also applicable for directly driving an external discrete n-p-n power transistor.

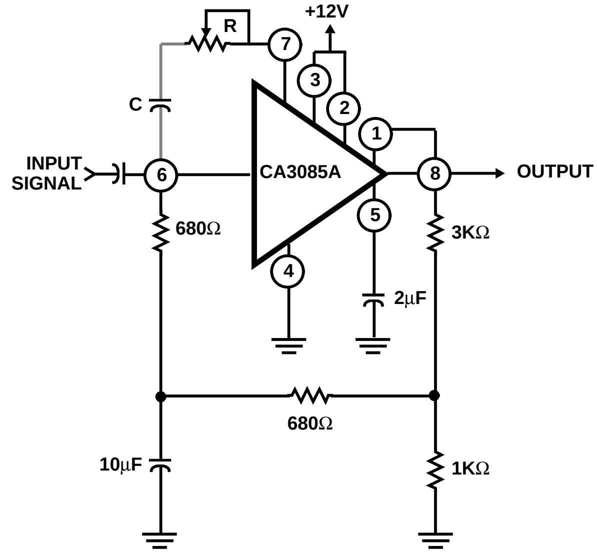


FIGURE 23. GENERAL-PURPOSE AMPLIFIER USING CA3085A