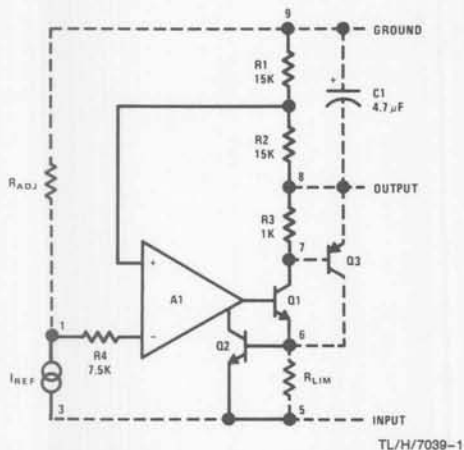# Designs for Negative Voltage Regulators

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Introduction

A number of IC voltage regulators have been introduced to date, but these have been designed primarily to regulate positive voltages. Most can be adapted as negative regulators, at some sacrifice in complexity, performance and flexibility. This note, however, describes an IC, which is designed specifically as a negative regulator. It is intended to complement the LM100 and LM105 positive regulators, providing a line of IC's for practically every regulator application.

Unique features of the circuit are that it supplies any output voltage from 0V down to −40V, while operating from a single unregulated supply. The output voltage is proportional to a single programming resistor, and remote sensing can be done at the load. It also regulates within 0.01% in circuits using a separate, floating bias supply, where the maximum output voltage is limited only by the breakdown of external pass transistors. The device is designed for either linear or switching regulator applications.

In the circuits described, emphasis is placed on practical considerations for the design of reliable regulators. Many of the pitfalls which cause unexpected failures are explained, and protection schemes for many of the hazards facing regulators are given. Most of the design hints are sufficiently general to apply equally to other IC's or even regulators designed entirely with discrete components.

increase the output-current capability. Short-circuit protection makes the circuit exhibit a constant-current characteristic when Q2 is turned on by the voltage drop across an external current-limit resistor, R_{lim}. A more complete description of the integrated circuit itself is given in the back of the text.
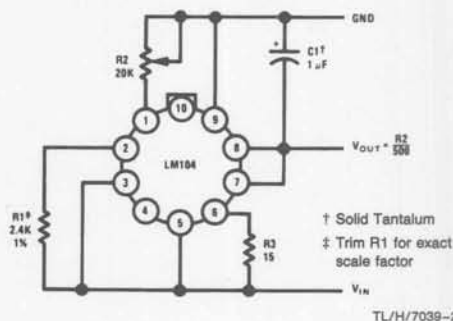
low power regulator or bias supply

This circuit can provide output voltages between 0V and −40V at currents up to 25 mA. The output voltage is linearly dependent on the value of R2, giving approximately 2V for each 1 KΩ of resistance. The exact scale factor can be set up by trimming R1. This should be done at the maximum output voltage setting in order to compensate for any mismatch in the internal divider resistors of the integrated circuit.

Short-circuit protection is provided by R3. The value of this resistor should be chosen so that the voltage drop across it is 300 mV at the maximum load current. This insures worst-case operation up to full load over a −55°C to 125°C temperature range. With a lower maximum operating temperature, the design value for this voltage can be increased linearly to 525 mV at 25°C.

For an output voltage setting of 15V, the regulation, no load to full load, is better than 0.05%; and the line regulation is better than 0.2% for a ±20% input voltage variation. Noise and ripple can be greatly reduced by bypassing R2 with a 10 μF capacitor. This will keep the ripple on the output less than 0.5 mV for a 1V, 120 Hz ripple on the unregulated input. The capacitor also improves the line-transient response by a factor of five.

An output capacitor of at least 1 μF is required to keep the regulator from oscillating. This should be a low inductance capacitor, preferably solid tantalum, installed with short leads. It is not usually necessary to bypass the input, but at least a 0.01 μF bypass is advisable when there are long leads connecting the circuit to the unregulated power source.

TL/H/7039−1

A functional diagram of the LM104 regulator and external circuitry (dash line) is shown in the figure. The internal reference is a temperature compensated current source, I_{ref}. A voltage which is proportional to an external programming resistor, R_{adj}, is fed into an error amplifier, A1. This drives an internal series pass transistor, Q1, to supply an output voltage equal to twice the voltage across the programming resistor. External pass transistors can be added, as is Q3, to

TL/H/7039−2

It is important to watch power dissipation in the integrated circuit even with load currents of 25 mA or less. The dissipation can be in excess of 1W with large input-output voltage differentials, and this is above ratings for the device.

Increased output current

When output currents above 25 mA are required or when the dissipation in the series pass transistor can be higher than about 0.2W, under worst-case conditions, it is advisable to add an external transistor to the LM104 to handle the power. The connection of an external booster transistor is shown here. The output current capability of the regulator is increased by the current gain of the added PNP transistor, but it is still necessary to watch dissipation in the external pass transistor. Excessive dissipation can burn out both the series pass transistor and the integrated circuit.

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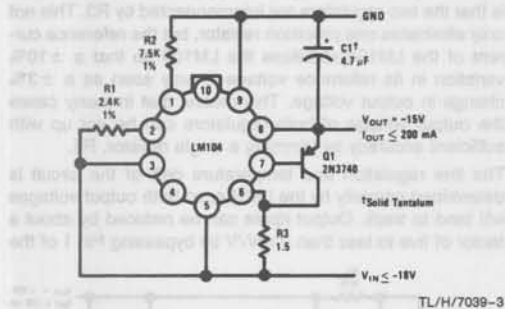
For example, with the circuit shown, the worst-case input voltage can be 25V. With a shorted output at 125°C, the current through the pass transistor will be 300 mA; and the dissipation in it will be 7.5W. This clearly establishes the need for an efficient heat sink.

For lower-power operation, a 2N2905 with a clip on heat sink can be used for the external pass transistor. However, when the worst case dissipation is above 0.5W, it is advisable to employ a power device such as the 2N3740 with a good heat sink.

The current limit resistor is chosen so that the voltage drop across it is 300 mV, with maximum load current, for operation to 125°C. With lower maximum ambients this voltage drop could be increased by 2.2 mV/°C. If possible, a fast-acting fuse rated about 25% higher than the maximum load current should be included in series with the unregulated input.

When a booster transistor is used, the minimum input-output voltage differential of the regulator will be increased by the emitter-base voltage of the added transistor. This establishes the minimum differential at 2 to 3V, depending on the base drive required by the external transistor.

high current regulator

When output currents in the ampere range are needed, it is necessary to add a second booster transistor to the LM104 circuitry. This connection is shown in the accompanying figure. The output current capability of the LM104 is increased by the product of the current gains of Q1 and Q2. However, it is still necessary to watch the dissipation in both the series pass transistor, Q2, and its driver, Q1. A clip-on heat sink is definitely required for Q1, and it is advisable to replace the 2N2905 with a 2N3740 which has a good heat sink when

output currents greater than 1A are needed. A 1000 pF capacitor should also be added between Pins 4 and 5 to compensate for the poorer frequency response of the 2N3740. The need for an efficient heat sink on Q2 should be obvious.

Experience shows that a single-diffused transistor such as a 2N3055 (or a 2N3772 for higher currents) is preferred over a double diffused, high-frequency transistor for the series pass element. The slower, single-diffused devices are less prone to secondary breakdown and oscillations in linear regulator applications.

As with the lower-current regulators, C1 is required to frequency compensate the regulator and prevent oscillations. It is also advisable to bypass the input with C2 if the regulator is located any distance from the output filter of the unregulated supply. The resistor across the emitter base junction of Q2 fixes the minimum collector current of Q1 to minimize oscillation problems with light loads. It is still possible to experience oscillations with certain physical layouts, but these can almost always be eliminated by stringing a ferrite bead, such as a Ferroxcube K5-001-00/3B, on the emitter lead of Q2.

The use of two booster transistors does not appreciably increase the minimum input-output voltage differential over that for a single transistor. The minimum differential will be 2 to 3V, depending on the drive current required from the integrated circuit.

With high current regulators, remote sensing is sometimes required to eliminate the effect of line resistance between the regulator and the load. This can be accomplished by returning R2 and Pin 9 of the LM104 to the ground end of the load and connecting Pin 8 directly to the high end of the load.

The low resistance values required for the current limit resistor, R3, are sometimes not readily available. A suitable resistor can be made using a piece of resistance wire or even a short length of kovar lead wire from a standard TO-5 transistor.

The current limit sense voltage can be reduced to about 400 mV by inserting a germanium diode (or a diode-connected germanium transistor) in series with Pin 6 of the LM104. This diode will also compensate the sense voltage and make the short circuit current essentially independent of temperature.

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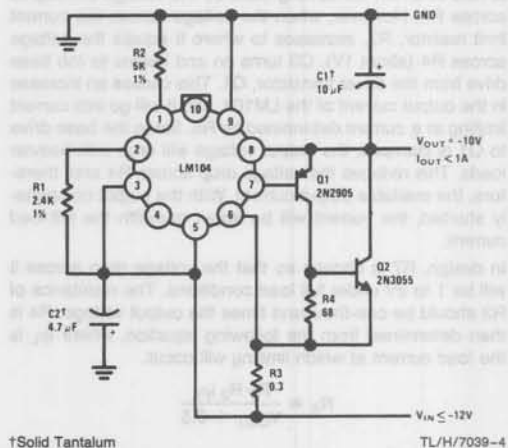
With high current regulators it is especially important to use a low-inductance capacitor on the output. The lead length on this capacitor must also be made short. Otherwise, the capacitor leads can resonate with smaller bypass capacitors (like 0.1 µF ceramic) which may be connected to the output. These resonances can lead to oscillations. With short leads on the output capacitor, the Q of the tuned circuit can be made low enough so that it cannot cause trouble.

foldback current limiting

High current regulators dissipate a considerable amount of power in the series pass transistor under full-load conditions. When the output is shorted, this dissipation can easily increase by a factor of four. Hence, with normal current limiting, the heat sink must be designed to handle much more power than the worst case full load dissipation if the circuit is to survive short-circuit conditions. This can increase the bulk of the regulator substantially.
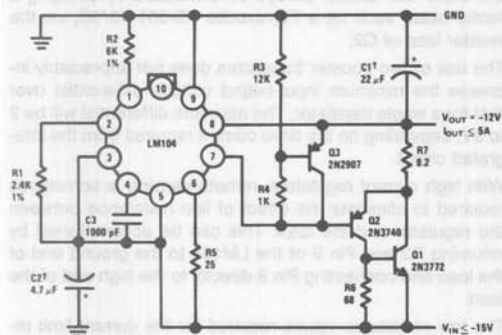
†Solid Tantalum

TL/H/7039-5

This situation can be eased considerably by using foldback current limiting. With this method of current limiting, the available output current actually decreases as the maximum load on the regulator is exceeded and the output voltage falls off. The short-circuit current can be adjusted to be a fraction of the full load current, minimizing dissipation in the pass transistor.

The circuit shown here accomplishes just this. Normally Q3 is held in a non-conducting state by the voltage developed across R4. However, when the voltage across the current limit resistor, R7, increases to where it equals the voltage across R4 (about 1V), Q3 turns on and begins to rob base drive from the driver transistor, Q1. This causes an increase in the output current of the LM104, and it will go into current limiting at a current determined by R5. Since the base drive to Q1 is clamped, the output voltage will drop with heavier loads. This reduces the voltage drop across R4 and, therefore, the available output current. With the output completely shorted, the current will be about one-fifth the full-load current.

In design, R7 is chosen so that the voltage drop across it will be 1 to 2V under full load conditions. The resistance of R3 should be one-thousand times the output voltage. R4 is then determined from the following equation, where I_{FL} is the load current at which limiting will occur.

$$R_4 \cong \frac{R_7 R_3 I_{FL}}{V_{OUT} + 0.5}$$

If it is desired to reduce the ratio of full load to short circuit current, this can be done by connecting a resistance of 2 to 10 kΩ across the emitter-base of Q3.

symmetrical power supplies

In many applications, such as powering operational amplifiers, there is a need for symmetrical positive and negative supply voltages. A circuit which is a particularly-economical solution to this design problem is shown in the adjoining figure. It uses a minimum number of components, and the voltage at both outputs can be set up within ±1.5% by a single adjustment. Further, the output voltages will tend to track with temperature and variations on the unregulated supply.

The positive voltage is regulated by an LM105, while an LM104 regulates the negative supply. The unusual feature is that the two regulators are interconnected by R3. This not only eliminates one precision resistor, but the reference current of the LM105 stabilizes the LM104 so that a ±10% variation in its reference voltage is only seen as a ±3% change in output voltage. This means that in many cases the output voltage of both regulators can be set up with sufficient accuracy by trimming a single resistor, R1.

The line regulation and temperature drift of the circuit is determined primarily by the LM104, so both output voltages will tend to track. Output ripple can be reduced by about a factor of five to less than 2 mV/V by bypassing Pin 1 of the
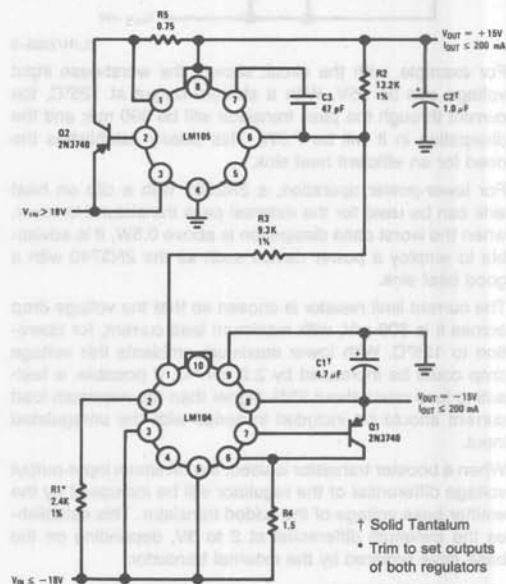

† Solid Tantalum
* Trim to set outputs of both regulators

TL/H/7039-6

LM104 to ground with a 10 µF capacitor. A center-tapped transformer with a bridge rectifier can be used for the unregulated power source.

adjustable current limiting

In laboratory power supplies, it is often necessary to adjust the limiting current of a regulator. This, of course, can be done by using a variable resistance for the current limit resistor. However, the current-limit resistor can easily have a value below that of commercially-available potentiometers. Discrete resistance values can be switched to vary the limiting current, but this does not provide continuously-variable adjustment.
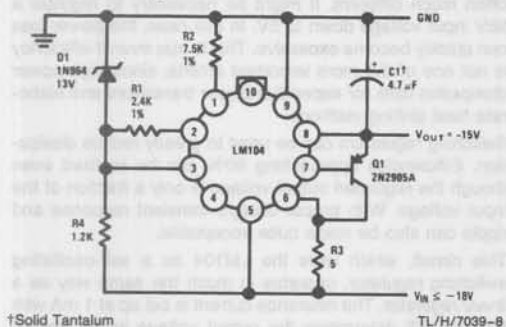
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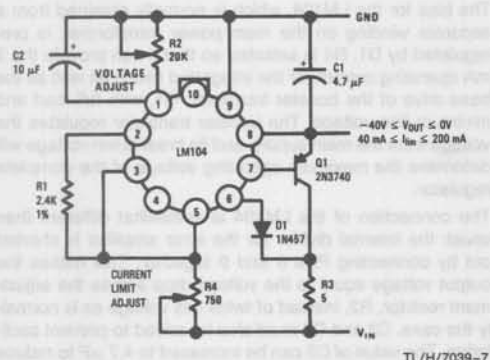
The circuit shown here solves this problem, giving a linear adjustment of limiting current over a five-to-one range. A silicon diode, D1, is included to reduce the current limit sense voltage to approximately 50 mV. Approximately 1.3 mA from the reference supply is passed through a potentiometer, R4, to buck out the diode voltage. Therefore, the effective current limit sense voltage is nearly proportional to the resistance of R4. The current through R4 is fairly insensitive to changes in ambient temperature, and D1 compensates for temperature variations in the current limit sense voltage of the LM104. Therefore, the limiting current will not be greatly affected by temperature.

It is important that a potentiometer be used for R4 and connected as shown. If a rheostat connection were used, it could open while it was being adjusted and momentarily increase the current limit sense voltage to many times its normal value. This could destroy the series pass transistors under short-circuit conditions.

The inclusion of R4 will soften the current limiting characteristics of the LM104 somewhat because it acts as an emitter-degeneration resistor for the current-limit transistor. This can be avoided by reducing the value of R4 and developing the voltage across R4 with additional bleed current to ground.

improving line regulation

The line regulation for voltage variations on the reference supply terminal of the LM104 is about five times worse than it is for changes on the unregulated input. Therefore, a zener-diode preregulator can be used on the reference supply to improve line regulation. This is shown in the figure below.

†Solid Tantalum

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The design of this circuit is fairly simple. It is only necessary that the minimum current through R4 be greater than 2 mA with low input voltage. Further, the zener voltage of D1 must be five volts greater than one-half the maximum output voltage to keep the transistors in the reference current source from saturating.

using protective diodes

It is a little known fact that most voltage regulators can be damaged by shorting out the unregulated input voltage while the circuit is operating—even though the output may have short-circuit protection. When the input voltage to the regulator falls instantaneously to zero, the output capacitor is still charged to the nominal output voltage. This applies voltage of the wrong polarity across the series pass transistor and other parts of the regulator, and they try to discharge the output capacitor into the short. The resulting current surge can damage or destroy these parts.

When the LM104 is used as the control element of the regulator, the discharge path is through internal junctions forward biased by the voltage reversal. If the charge on the output capacitor is in the order of 40 volt • μF, the circuit can be damaged during the discharge interval. However, the problem is not only seen with integrated circuit regulators. It also happens with discrete regulators where the series-pass transistor usually gets blown out.

The problem can be eliminated by connecting a diode between the output and the input such that it discharges the output capacitor when the input is shorted. The diode should be capable of handling large current surges without excessive voltage drop, but it does not have to be a power diode since it does not carry current continuously. It should also be relatively fast. Ordinary rectifier diodes will not do because they look like an open circuit in the forward direction until minority carriers are injected into the intrinsic base region of the PIN structure.

This problem is not just caused by accidental physical shorts on the input. It has shown up more than once when regulators are driven from high-frequency dc-dc converters. Tantalum capacitors are frequently used as output filters for the rectifiers. When these capacitors are operated near their maximum voltage ratings with excessive high frequency ripple across them, they have a tendency to sputter—that is, short momentarily and clear themselves. When they short, they can blow out the regulator; but they look innocent after the smoke has cleared.

The solution to this problem is to use capacitors with conservative voltage ratings, to observe the maximum ripple ratings for the capacitor and to include a protective diode between the input and output of the regulator to protect it in case sputtering does occur.

Heavy loads operating from the unregulated supply can also destroy a voltage regulator. When the input power is switched off, the input voltage can drop faster than the output voltage, causing a voltage reversal across the regulator, especially when the output of the regulator is lightly loaded. Inductive loads such as a solenoid are particularly troublesome in this respect. In addition to causing a voltage reversal between the input and the output, they can reverse the input voltage causing additional damage.

In cases like this, it is advisable to use a multiple-pole switch or relay to disconnect the regulator from the unregulated supply separate from the other loads. If this cannot be done, it is necessary to put a diode across the input of the regulator to clamp any reverse voltages, in addition to the protective diode between the input and the output.
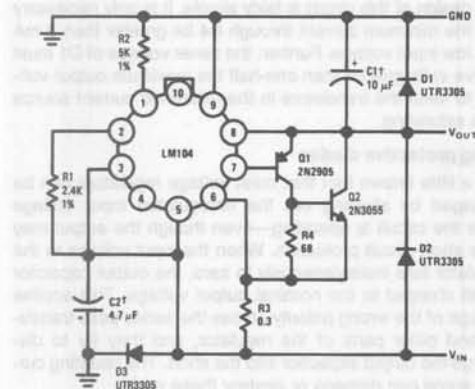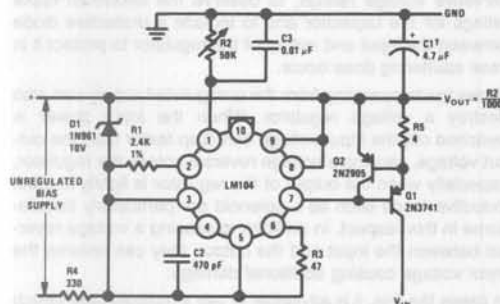
†Solid Tantalum TL/H/7039–9

Yet another failure mode can occur if the regulated supply drives inductive loads. When power is shut off, the inductive current can reverse the output voltage polarity, damaging the regulator and the output capacitor. This can be cured with a clamp diode on the output. Even without inductive loads it is usually good practice to include this clamp diode to protect the regulator if its output is accidentally shorted to a negative supply.

A regulator with all these protective diodes is shown here. D1 protects against output voltage reversal. D2 prevents a voltage reversal between the input and the output of the regulator. And D3 prevents a reversal of the input-voltage polarity. In many cases, D3 is not needed if D1 and D2 are used, since these diodes will clamp the input voltage within two diode drops of ground. This is adequate if the input voltage reversals are of short duration.

high voltage regulator

In the design of commercial power supplies, it is common practice to use a floating bias supply to power the control circuitry of the regulator. As shown here, this connection can be used with the LM104 to regulate output voltages that are higher than the ratings of the integrated circuit. Better regulation can also be obtained because it is a simple matter to preregulate the low current bias supply so that the integrated circuit does not see ripple or line voltage variations and because the reduced operating voltage minimizes power dissipation and associated thermal effects from the current delivered to the booster transistor.

†Solid Tantalum TL/H/7039–10

The bias for the LM104, which is normally obtained from a separate winding on the main power transformer, is pre-regulated by D1. R4 is selected so that it can provide the 3 mA operating current for the integrated circuit as well as the base drive of the booster transistor, Q1, with full load and minimum line voltage. The booster transistor regulates the voltage from the main supply, and its breakdown voltage will determine the maximum operating voltage of the complete regulator.

The connection of the LM104 is somewhat different than usual: the internal divider for the error amplifier is shorted out by connecting Pins 8 and 9 together. This makes the output voltage equal to the voltage drop across the adjustment resistor, R2, instead of twice this voltage as is normally the case. C2 and C3 must also be added to prevent oscillation. The value of C3 can be increased to 4.7 μF to reduce noise on the output.

It is necessary to add Q2 and R5 to provide current limiting. When the output current becomes high enough to turn on Q2, there will be an abrupt rise in the output current of the LM104 as Q2 tries to remove base drive from the booster transistor. Any further increases in load current will cause the LM104 to limit at a current determined by R3, and the output voltage will collapse. The value of R3 must be selected so that the integrated circuit can deliver the base current of Q1, at full load, without limiting.

A second, NPN booster transistor can be used in a compound connection with Q1 to increase the output current of the regulator. However, with very-high-voltage regulators, the most economical solution may be to use a high voltage PNP driving a vacuum tube for the series pass element.

Remote sensing, which eliminates the effects of voltage dropped in the leads connecting the regulator to the load, can be provided by connecting R2 to the ground end of the load and Pins 8 and 9 to the high end of the load.

switching regulator

Linear regulators have the advantages of fast response to load transients as well as low noise and ripple. However, since they must dissipate the difference between the unregulated-supply power and the output power, they sometimes have a low efficiency. This is not always a problem with ac line-operated equipment because the power loss is easily afforded, because the input voltage is already fairly-well regulated and because losses can be minimized by adjustment of transformer ratios in the power supply. In systems operating from a fixed dc input voltage, the situation is often much different. It might be necessary to regulate a 28V input voltage down to 5V. In this case, the power loss can quickly become excessive. This is true even if efficiency is not one of the more important criteria, since high power dissipation calls for expensive power transistors and elaborate heat sinking methods.

Switching regulators can be used to greatly reduce dissipation. Efficiencies approaching 90% can be realized even though the regulated output voltage is only a fraction of the input voltage. With proper design, transient response and ripple can also be made quite acceptable.

This circuit, which uses the LM104 as a self-oscillating switching regulator, operates in much the same way as a linear regulator. The reference current is set up at 1 mA with R1, and R2 determines the output voltage in the normal fashion. The circuit is made to oscillate by applying positive feedback through R5 to the non-inverting input on the error amplifier of the LM104. When the output voltage is low, the

internal pass transistor of the integrated circuit turns on and drives Q1 into saturation. The current feedback through R5 then increases the magnitude of the reference voltage developed across R2. Q1 will remain on until the output voltage comes up to twice this reference voltage. At this point, the error amplifier goes into linear operation, and the positive feedback makes the circuit switch off. When this happens, the reference voltage is lowered by feedback through R5, and the circuit will stay off until the output voltage drops to where the error amplifier again goes into linear operation. Hence, the circuit regulates with the output voltage oscillating about the nominal value with a peak-to-peak ripple of around 40 mV.

The power conversion from the input voltage to a lower output voltage is obtained by the action of the switch transistor, Q1, the catch diode, D1, and the LC filter. The inductor is made large enough so that the current through it is essentially constant throughout the switching cycle. When Q1 turns on, the voltage on its collector will be nearly equal to the unregulated input voltage. When it turns off, the magnetic field in L1 begins to collapse, driving the collector voltage of Q1 to ground where it is clamped by D1.

If, for example, the input voltage is 10V and the switch transistor is driven at a 50% duty cycle, the average voltage on the collector of Q1 will be 5V. This waveform will be filtered by L1 and C1 and appear as a 5V dc voltage on the output. Since the inductor current comes from the input while Q1 is on but from ground through D1 while Q1 is off, the average value of the input current will be half the output current. The power output will therefore equal the input power if switching losses are neglected.

In design, the value of R3 is chosen to provide sufficient base drive to Q1 at the maximum load current. R4 must be low enough so that the bias current coming out of Pin 5 of the LM104 (approximately 300 μA) does not turn on the switch transistor. The purpose of C2 is to remove transients that can appear across R2 and cause erratic switching. It should not be made so large that it severely integrates the waveform fed back to this point.


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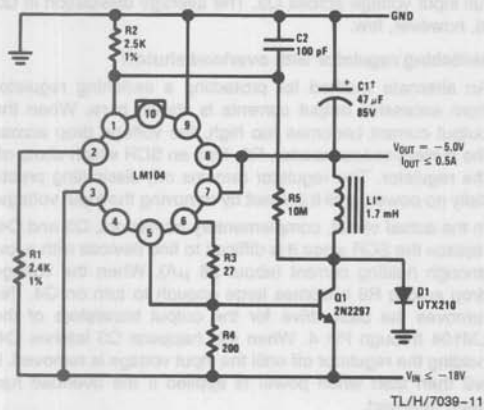
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Arnold Engineering
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Molybdenum Permalloy Core

high current switching regulator

Output currents up to 3A can be obtained using the switching regulator circuit shown here. The circuit is identical to the one described previously, except that Q2 has been added to increase the output current capability by about an order of magnitude. It should be noted that the reference supply terminal is returned to the base of Q2, rather than the unregulated input. This is done because the LM104 will not function properly if Pin 5 gets more than 2V more positive than Pin 3. The reference current, as well as the bias currents for Pins 3 and 5, is supplied from the unregulated input through R5, so its resistance must be low enough so that Q2 is not turned on with about 2 mA flowing through it.

The line regulation of this circuit is worsened somewhat by the unregulated input voltage being fed back into the reference for the regulator through R6. This effect can be eliminated by connecting a 0.01 μF capacitor in series with R6 to remove the dc component of the feedback.

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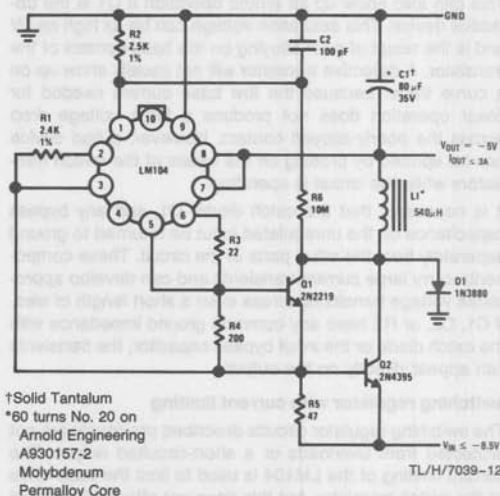
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Molybdenum
Permalloy Core

There are a number of precautions that should be observed with all switching regulators, although they are more inclined to cause problems in high-current applications:

For one, fast switching diodes and transistors must be used. If D1 is an ordinary junction rectifier, voltages in the order of 10V can be developed across it in the forward direction when the switch transistor turns off. This happens because low-frequency rectifiers are usually manufactured with a PIN structure which presents a high forward impedance until enough minority carriers are injected into the diode base region to increase its conductance. This not only causes excessive dissipation in the diode, but the diode also presents a short circuit to the switch transistor, when it first turns on, until all the charge stored in the base region of the diode is removed. Similarly, a high frequency switch transistor must be used as excessive switching losses in low frequency transistors, like the 2N3055, make them overheat.

It is important that the core material used for the inductor have a soft saturation characteristic. Cores that saturate abruptly produce excessive peak currents in the switch transistor if the output current becomes high enough to run the core close to saturation. Powdered molybdenum-permalloy cores, on the other hand, exhibit a gradual reduction in per-

meability with excessive current, so the only effect of output currents above the design value is a gradual increase in switching frequency.

One thing that is frequently overlooked in the design of switching circuits is the ripple rating of the filter capacitors. Excessive high-frequency ripple can cause these capacitors to fail. This is an especially-important consideration for capacitors used on the unregulated input as the ripple current through them can be higher than the dc load current. The situation is eased somewhat for the filter capacitor on the output of the regulator since the ripple current is only a fraction of the load current. Nonetheless, proper design usually requires that the voltage rating of this capacitor be higher than that dictated by the DC voltage across it for reliable operation.

One unusual problem that has been noted in working with switching regulators is excessive dissipation in the switch transistors caused by high emitter-base saturation voltage. This can also show up as erratic operation if Q1 is the defective device. This saturation voltage can be as high as 5V and is the result of poor alloying on the base contact of the transistor. A defective transistor will not usually show up on a curve tracer because the low base current needed for linear operation does not produce a large voltage drop across the poorly-alloyed contact. However, a bad device can be spotted by probing on the bases of the switch transistors while the circuit is operating.

It is necessary that the catch diode, D1, and any bypass capacitance on the unregulated input be returned to ground separately from the other parts of the circuit. These components carry large current transients and can develop appreciable voltage transients across even a short length of wire. If C1, C2, or R2 have any common ground impedance with the catch diode or the input bypass capacitor, the transients can appear directly on the output.

switching regulator with current limiting

The switching regulator circuits described previously are not protected from overloads or a short-circuited output. The current limiting of the LM104 is used to limit the base drive of the switch transistor, but this does not effectively protect the switch transistor from excessive current. Providing short circuit protection is no simple problem, since it is necessary to keep the regulator operating in the switching mode when the output is shorted. Otherwise, the dissipation in the switch transistor will become excessive even though the current is limited.

A circuit which provides current limiting and protects the regulator from short circuits is shown here. The current through the switch transistor produces a voltage drop across R9. When this voltage becomes large enough to turn on Q3, current limiting is initiated. This occurs because Q3 takes over as the control transistor and regulates the voltage on Pin 8 of the LM104. This point, which is the feedback terminal of the error amplifier, is separated from the actual output of the regulator by not shorting the regulated output and booster output terminals of the integrated circuit. Hence, with excessive output current, the circuit still operates as a switching regulator with Q3 regulating the voltage fed back to the error amplifier as the output voltage falls off.

A resistor, R7, is included so that excessive base current will not be driven into the base of Q3. C4 insures that Q3 does not turn on from the current spikes through the switch transistor caused by pulling the stored charge out of the catch diode (these are about twice the load current). This

capacitor also operates in conjunction with C2 to produce sufficient phase delay in the feedback loop so that the circuit will oscillate in current limiting. However, C4 should not be made so large that it appreciably integrates the rectangular waveform of the current through the switch transistor.
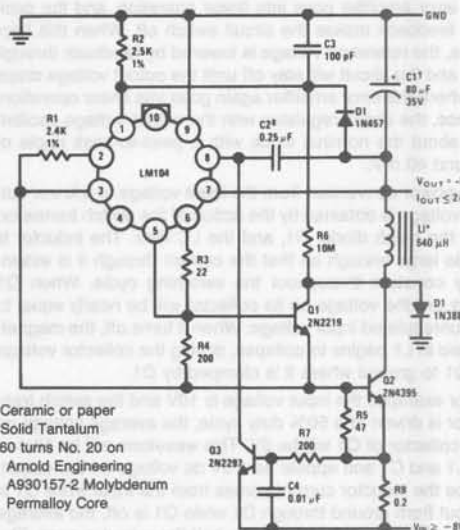
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A930157-2 Molybdenum
Permalloy Core

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As the output voltage falls below half the design value, D1 pulls down the reference voltage across R2. This permits the current limiting circuitry to keep operating when the unregulated input voltage drops below the design value of output voltage, with a short on the output of the regulator.

A transistor with good high-current capability was chosen for Q3 so that it does not suffer from secondary breakdown effects from the large peak currents (about 200 mA) through it. With a shorted output, these peakcurrents occur with the full input voltage across Q3. The average dissipation in Q3 is, however, low.

switching regulator with overload shutoff

An alternate method for protecting a switching regulator from excessive output currents is shown here. When the output current becomes too high, the voltage drop across the current-sense resistor, R8, fires an SCR which shuts off the regulator. The regulator remains off, dissipating practically no power, until it is reset by removing the input voltage.

In the actual circuit, complementary transistors, Q3 and Q4, replace the SCR since it is difficult to find devices with a low enough holding current (about 25 μA). When the voltage drop across R8 becomes large enough to turn on Q4, this removes the base drive for the output transistors of the LM104 through Pin 4. When this happens Q3 latches Q4, holding the regulator off until the input voltage is removed. It will then start when power is applied if the overload has been removed.

With this circuit, it is necessary that the shutoff current be 1.5 times the full load current. Otherwise, the circuit will shut off when it is switched on with a full load because of the excess current required to charge the output capacitor. The shutoff current can be made closer to the full load current by connecting a 10 μF capacitor across R2 which will limit the charging current for C1 by slowing the risetime of the

output voltage when the circuit is turned on. However, this capacitor will also bypass the positive feedback from R6 which makes the regulator oscillate. Therefore, it is necessary to put a 270Ω resistor in the ground end of the added capacitor and provide feedback to this resistor from the collector of Q1 through a 1 MΩ resistor.
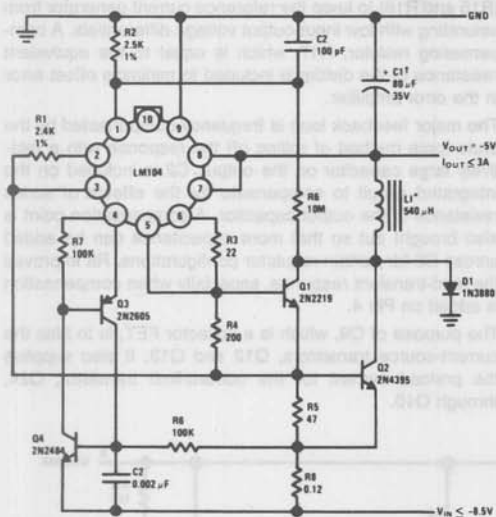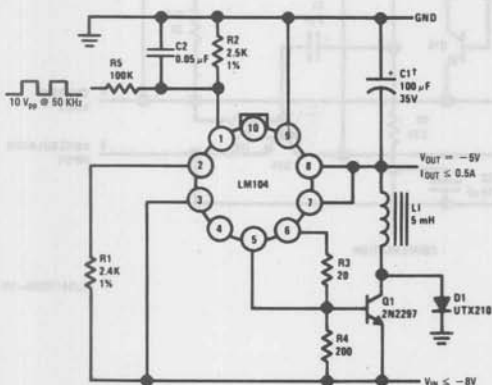

TL/H/7039–14

†Solid Tantalum
*60 turns No. 20 on
Arnold Engineering
A930157-2 Molybdenum
Permalloy Core

driven switching regulator

When a number of switching regulators are operated from a common power source, it is desirable to synchronize their operation to more uniformly distribute the switched current waveforms in the input line. Synchronous operation can also be beneficial when a switching regulator is operated in conjunction with a power converter.

TL/H/7039–15

†Solid Tantalum

A circuit which synchronizes the switching regulator with a square wave drive signal is shown here. It differs from the switching regulators described previously in that positive feedback is not used. Instead, a triangular wave with a peak-to-peak amplitude of 25 mV is applied to the noninverting input of the error amplifier. The waveform is obtained by integrating the square wave synchronizing signal. This triangular wave causes the error amplifier to switch because its gain is high enough that the waveform easily overdrives it. The switching duty cycle is controlled by the output voltage fed back to the error amplifier. If the output voltage goes up, the duty cycle will decrease since the error amplifier will pick off a smaller portion of the triangular wave. Similarly, the duty cycle will decrease if the output voltage drops. Hence, the duty cycle is controlled to produce the desired output voltage.

Without a synchronous drive signal, the circuit will self oscillate at a frequency determined by L1 and C1. This self-oscillation frequency must be lower than the synchronous drive frequency. Therefore, more filtering is required for a driven regulator than for a self-oscillating regulator operating at the same frequency. This also means that a driven regulator will have less output ripple.

The value of C2 is chosen so that its capacitive reactance at the drive frequency is less than one-tenth the resistance of R2. The amplitude of the triangular wave is set at 25 mV with R5. It is advisable to ac couple the drive signal by putting a capacitor in series with R5 so that it does not disturb the dc reference voltage developed for the error amplifier.

the LM104 regulator

The basic reference for the regulator is zener diode D1. The reference diode is supplied from a PNP current source, Q8, which has a fixed current gain of 2. This arrangement permits the circuit to operate with unregulated input voltages as low as 7V, substantially increasing the efficiency of low-voltage regulators.

The reference supply is temperature compensated by using the negative temperature coefficient of the transistor emitter-base voltages to cancel the positive coefficient of the zener diode. The design produces a nominal 2.4V between the reference and reference supply terminals of the integrated circuit. Connecting an external 2.4 KΩ resistor between those terminals gives a 1 mA reference current from the collectors of Q1 and Q2, which is independent of temperature. The reference voltage supplied to the error amplifier is developed across a second external resistor connected between the adjustment terminal and ground.

The reference supply terminal is normally connected to the unregulated supply. However, improved line regulation can be obtained by preregulating the voltage on this terminal. This improvement occurs because Q1, Q2, and Q7 do not see changes in input voltage. Normally, it is the change in the emitter-base voltage of these transistors with changes in collector-base voltage which determines the line regulation.

When the reference supply and unregulated input terminals are operated from separate voltage sources, it is important to make sure that the unregulated input terminal of the integrated circuit does not get more than 2V more positive than the reference supply terminal. If this happens, the collector-isolation junction of Q6 becomes forward biased and disrupts the reference.

The error amplifier of the regulator is quite similar to the LM101 operational amplifier. Emitter follower input transistors, Q18 and Q19, drive a dual PNP which is operated in the common-base configuration. The current gain of these PNP transistors is fixed at 4 so that the base can be driven by a current source (Q13). Active collector loads are used for the input stage so that a voltage gain of 2000 is obtained. Q21 and Q22 provide enough current gain to keep the internal, series-pass transistor from loading the input stage. R14 limits the base drive on Q23 when it saturates with low, unregulated input voltages. The collector of Q23 is brought out separately so that an external booster transistor can be added for increased output current capability. R13 established the minimum operating current in Q23 when booster transistors are used.

One feature of the error amplifier is that it operates properly with common mode voltages all the way up to ground. Because of this, the circuit will regulate with output voltages to zero volts.
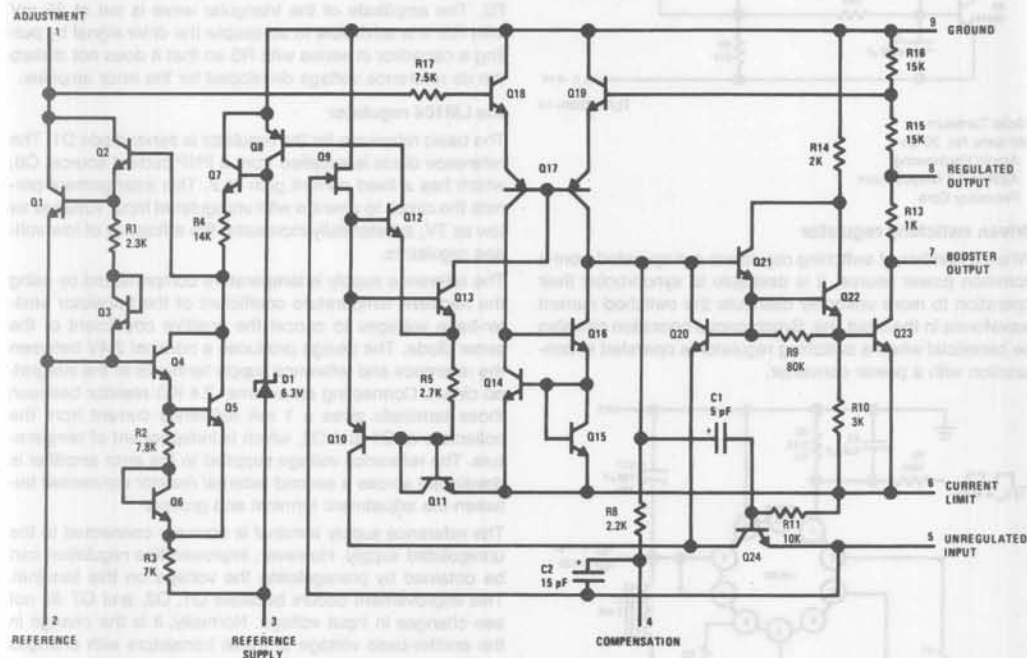
Current limiting is provided by Q24. When the voltage between the current limit and unregulated input terminals becomes large enough to turn on Q24, it will pull Q10 out of saturation and remove base drive from Q21 through Q20.

This causes the series pass transistor to exhibit a constant current characteristic. The pre-load current, provided for Q24 by Q10 before current limiting is initiated, gives a much sharper current-limit characteristic. C1 and R11 are included in the limiting circuitry to suppress oscillations.

The error amplifier is connected to a divider on the output (R15 and R16) to keep the reference current generator from saturating with low input-output voltage differentials. A compensating resistor, R17, which is equal to the equivalent resistance of the divider is included to minimize offset error in the error amplifier.

The major feedback loop is frequency compensated by the brute-force method of rolling off the response with a relatively large capacitor on the output. C2 is included on the integrated circuit to compensate for the effects of series resistance in the output capacitor. A compensation point is also brought out so that more capacitance can be added across C2 for certain regulator configurations. R8 improves the load-transient response, especially when compensation is added on Pin 4.

The purpose of Q9, which is a collector FET, is to bias the current-source transistors, Q12 and Q13. It also supplies the preload current for the current-limit transistor, Q24, through Q10.

TL/H/7039–16