March 1969

DESIGNING SWITCHING REGULATORS

INTRODUCTION

The series pass element in a conventional series regulator operates as a variable resistance which drops an unregulated input voltage down to a fixed output voltage. This element, usually a transistor, must be able to dissipate the voltage difference between the input and output at the load current. The power generated can become excessive, particularly when the input voltage is not well regulated and the difference between the input and output voltages is large.

Switching regulators, on the other hand, are capable of high efficiency operation even with large differences between the input and output voltages. The efficiency is, in fact, negligibly affected by the voltage difference since this type of regulator acts as a continuously-variable power converter.

Switching regulators are, therefore, useful in battery-powered equipment where the required output voltage is considerably lower than the battery voltage. An example of this is a missile with a 30V battery as its only power source, containing a large number of integrated logic circuits which require a 5V supply. Switching regulators are also useful in space vehicles where conservation of power is extremely important. In addition, they are frequently the most economical solution in commercial and industrial applications where the increased efficiency reduces the cost of the series-pass transistors and simplifies heat sinking.

One of the disadvantages of switching regulators is that they are more complex than linear regulators, but this is often a substitution of electrical complexity for the thermal and mechanical complexity of high power linear regulators. Another disadvantage is higher output ripple. However, this can be held to a minimum (about 10 mV) and it is at a high enough frequency so that it can be easily filtered out. Another limitation is that the response to load transients is not always as fast as with linear regulators, but this can be largely overcome by proper design. The rejection of line transients, however, is every bit as good if not better than linear regulators. Lastly, switching regulators throw current transients back into the unregulated supply which are somewhat larger than the maximum load current. These, in some cases, can be troublesome unless adequate filtering is used.

This article will demonstrate the use of a monolithic voltage regulator in a number of switching regulator applications. These include both self-oscillating and synchronously driven regulators in the 0.1A to 5A range. Circuits are shown for both positive and negative regulators with output voltages in the 2V to 30V range. Methods of isolating the integrated circuit from the input voltage are given, permitting input voltages in excess of 100V. Further, current limiting schemes which keep peak currents and dissipation well within safe limits for both over-load and short-circuit conditions are presented. Finally, component selection details peculiar to switching regulators are covered.

SWITCHING REGULATOR OPERATION

The method by which a switching regulator produces a voltage conversion with high efficiency can be explained with the aid of Figure 1. Q_1 is a switch transistor which is turned on and off by a pulse waveform with a given duty cycle, and D_1 is a catch diode which provides a continuous path for the inductor current when Q_1 turns off. The voltage waveform on the collector of Q_1 will be as shown in the figure. The output of the LC filter will be the average value of the switched waveform, V_1. If the voltage drops across the transistor and diode are neglected, the output voltage will be

$$V_{OUT} = V_{IN} \frac{t_{on}}{T}; \qquad (1)$$

and it is independent of the load current. It is obvious from the equation that changes in input voltage can be compensated for by varying the duty cycle of the switched waveform. This is what is done in a switching regulator.

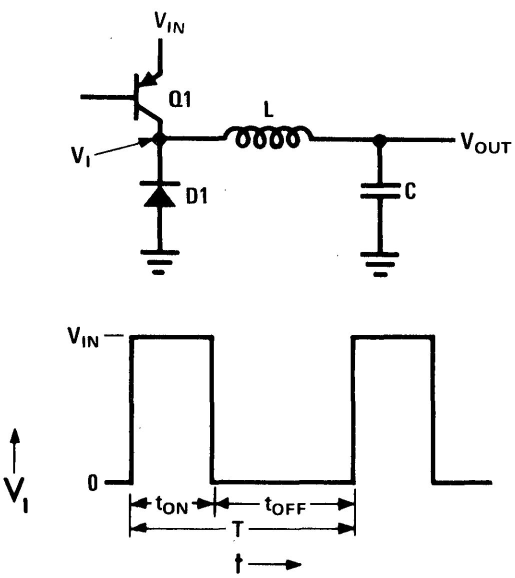


FIGURE 1
Switching Circuit for Voltage Conversion

Figure 2 shows a self-oscillating switching regulator which produces this duty-cycle control. A reference voltage, V_{ref}, equal to the desired output voltage, is supplied to one input of an operational amplifier, A_1. The operational amplifier, in turn, drives the switch transistor. The resistive divider, arranged such that $R_1 >> R_2$, provides a slight amount of positive feedback at high frequencies to make the circuit oscillate. At lower frequencies where the attenuation of the LC filter is less than the attenuation of the resistive divider, there is net negative feedback to the inverting input of the operational amplifier.

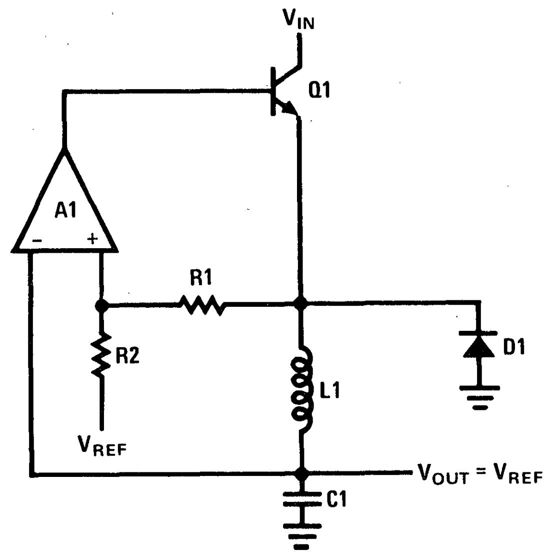


FIGURE 2
Self-oscillating Switching Regulator

In operation, when the circuit is first turned on, the output voltage is less than the reference voltage so the switch transistor is turned on. When this happens, current flow through R_1 raises the voltage on the non-inverting input of the operational amplifier slightly above the reference voltage. The circuit will remain switched on until the output rises to this voltage. The amplifier now goes into the active region, causing the switch to turn off. At this point, the reference voltage seen by the amplifier is *lowered* by feedback through R_1, and the circuit will stay off until the output voltage drops to this lower voltage. Hence, the output voltage oscillates about the reference voltage. The amplitude of this oscillation (or the output ripple) is nearly equal to the voltage fed back through R_1 to R_2 and can be made quite small.

THE LM100

The switching regulator circuits described here use the LM100 integrated voltage regulator as the control element. This device contains, on a single silicon chip, the voltage reference, the operational amplifier and the circuitry for driving a PNP switch transistor. Discrete switch transistors, catch diodes and reactive elements are employed since these components are not easily integrated.

A complete circuit description of the LM100 is given in Application Note AN-1 along with a number of its applications as a linear regulator. However, a brief description will be included here in order to facilitate understanding of the regulator circuits which follow.

Figure 3 shows a schematic diagram of the LM100. The voltage reference portion of the circuit starts with a breakdown diode, D_1, which is supplied by a current source from the unregulated input (one of the collectors of Q_2). The output of the reference diode, which has a positive temperature coefficient of 2.4 mV/°C, is buffered by an emitter follower, Q_4, which increases the temperature coefficient to +4.7 mV/°C. This is further increased to 7 mV/°C by the diode-connected transistor, Q_6. A resistor divider reduces this voltage as well as the temperature coefficient to exactly compensate for the negative temperature coefficient of Q_7, producing a temperature-compensated output of 1.8V.

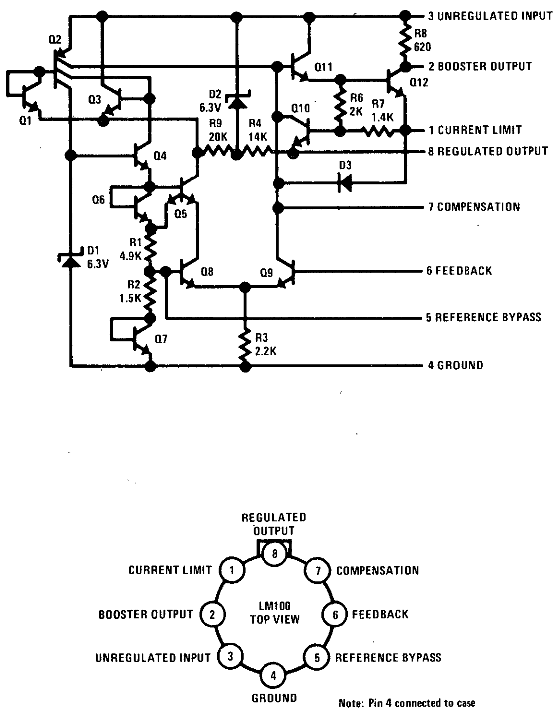


FIGURE 3

Schematic and Connection Diagrams of the LM100 Voltage Regulator

The transistor pair, Q_8 and Q_9, form the input stage of the operational amplifier. The gain of the stage is made high by the use of a current source, one of the collectors of Q_2, as a collector load. The output of this stage drives a compound emitter follower, Q_{11} and Q_{12}. The output of Q_{12} is taken across R_8 to drive the PNP switch transistor. An additional transistor, Q_{10}, is used to limit the output current of Q_{12} to the value required for driving a PNP transistor connected on the booster output. This current is determined by a resistor placed between the current limit and regulated output terminals. The value of the drive current can be determined from Figure 4 which plots the output current as a function of temperature for various current limit resistors.

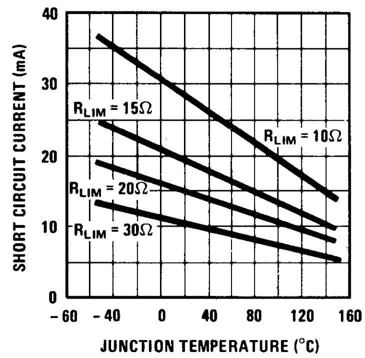


FIGURE 4

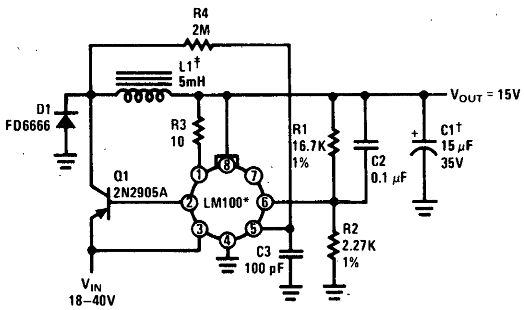
Switched Output Current as a Function of Temperature for Various Values of Current Limit Resistors

As for the remaining details of the circuit, Q_5, Q_3 and Q_1 are part of a bias stabilization circuit for Q_2 to set its collector currents at the desired value. R_9, R_4 and D_2 serve the sole function of starting the regulator. Lastly, D_3 is a clamp diode which keeps Q_9 from saturating when it is switching.

SWITCHING REGULATOR CIRCUITS

Figure 5 demonstrates the use of the LM100 as a switching regulator. Feedback to the inverting input of the operational amplifier (Pin 6 of the LM100) is obtained through a resistive divider which can be used to set the output voltage anywhere in the 2-30V range. R_3 determines the base drive for the switch transistor, Q_1, providing enough drive to saturate it with maximum load current. R_4 works into the 1 kΩ impedance at the reference terminal, producing the positive feedback. C_2 serves to minimize output ripple by causing the full ripple to appear on the feedback terminal. The remaining capacitor, C_3, removes the fast-risetime transients which would otherwise be coupled into Pin 5 through the shunt capacitance of R_4. It must be made small enough so that it does not seriously integrate the waveform at this point.

The circuit shown in Figure 5 is suitable for output currents as high as 500 mA. This limit is set by the output current available from the LM100 to saturate the switch transistor, Q_1. For lower currents, the value of R_3 should be increased so that the base of Q_1 is not driven unnecessarily hard.



*Basing diagram is Top View

†Solid tantalum

‡125 turns #22 on Arnold Engineering A262123-2 molybdenum permalloy core.

FIGURE 5

Switching Regulator Using the LM100

The optimum switching frequency for these regulators has been determined to be between 20 kHz and 100 kHz. At lower frequencies, the core becomes unnecessarily large; and at higher frequencies, switching losses in Q_1 and D_1 become excessive. It is important, in this respect, that both Q_1 and D_1 be fast-switching devices to minimize switching losses.

The output ripple of the regulator at the switching frequency is mainly determined by R_4. It should be evident from the description of circuit operation that the peak-to-peak output ripple will be nearly equal to the peak-to-peak voltage fed back to Pin 5 of the LM100. Since the resistance looking into Pin 5 is approximately 1000Ω, this voltage will be

$$\Delta V_{ref} \simeq \frac{1000\, V_{IN}}{R_4}. \tag{2}$$

In practice, the ripple will be somewhat larger than this. When the switch transistor shuts off, the current in the inductor will be greater than the load current so the output voltage will continue to rise above the value required to shut off the regulator. An important consideration in choosing the value of the inductor is that it be large enough so that the current through it does not change drastically during the switching cycle. If it does, the switch transistor and catch diode must be able to handle peak currents which are significantly larger than the load current. The change in inductor current can be written as

$$\Delta I_L \simeq \frac{V_{OUT} t_{off}}{L}. \tag{3}$$

In order for the peak current to be about 1.2 times the maximum load current, it is necessary that

$$L_1 = \frac{2.5\, V_{OUT} t_{off}}{I_{OUT\,(max)}}. \tag{4}$$

A value for t_{off} can be estimated from

$$t_{off} = \frac{1}{f}\left(1 - \frac{V_{OUT}}{V_{IN}}\right), \tag{5}$$

where f is the desired switching frequency and V_{IN} is the nominal input voltage.

The size of the output capacitor can now be determined from

$$C_1 = \left(\frac{V_{IN} - V_{OUT}}{2L_1\, \Delta V_{OUT}}\right)\left(\frac{V_{OUT}}{fV_{IN}}\right)^2, \tag{6}$$

where ΔV_{OUT} is the peak-to-peak output ripple and V_{IN} is the nominal input voltage.

It now remains to determine if the component values obtained above give satisfactory load-transient response. The overshoot of the regulator can be determined from

$$\Delta V_{OUT} = \frac{L_1\,(\Delta I_L)^2}{C_1\,(V_{IN} - V_{OUT})} \tag{7}$$

for increasing loads, and

$$\Delta V_{OUT} = \frac{L_1\,(\Delta I_L)^2}{C_1\, V_{OUT}} \tag{8}$$

for decreasing loads, where ΔI_L is the load-current transient. The recovery time is

$$t_r = \frac{2L_1 \Delta I_L}{V_{IN} - V_{OUT}} \tag{9}$$

and

$$t_r = \frac{2L_1 \Delta I_L}{V_{OUT}} \tag{10}$$

for increasing and decreasing loads respectively.

In order to improve the load transient response, it is necessary to allow larger peak to average current ratios in the switch transistor and catch diode. Reducing the value of inductance given by Equation (4) by a factor of 2 will reduce the overshoot by 4 times and halve the response time. This, of course, assumes that the output capacitance is doubled to maintain a constant switching frequency.

The above equations outline a design procedure for determining the value for R_4, L_1, and C_1, given the switching frequency and the output ripple. These equations are not exact, but they do provide a starting point for designing a regulator to fit a given application.

As an example, this design method will be applied to a regulator which must deliver 15V at a maximum current of 300 mA from a 28V supply. To start, a 40 kHz switching frequency will be selected along with an output ripple of 14 mV, peak-to-peak.

From (2), R_4 is calculated to be 2 MΩ. In determining L_1, t_{off} is found to be 11.6 μs from (5). Inserting this into (4) gives a value of 1.45 mH for L_1. The value of C_1 obtained from (6) is then 57.5 μF.

In the actual circuit of Figure 5, a standard value of 47 μF is used for C_1; and L_1 is adjusted to 1.7 mH. The switching frequency obtained experimentally on this circuit is 60 kHz and the peak-to-peak output ripple is 20 mV. The fairly-large disagreement between the calculated and experimental values is not alarming since many simplifying assumptions were made in the derivation of the equations. They do, however, provide a convenient method of handling a large number of mutually-dependent variables to arrive at a working circuit.

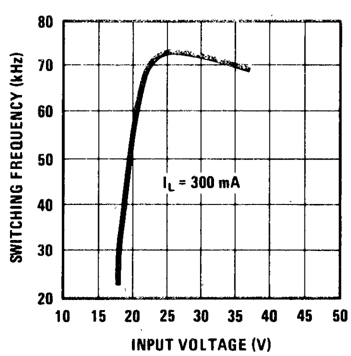


FIGURE 6

Switching Frequency as a Function of Input Voltage

More exact expressions would involve a design procedure which is too cumbersome to be of practical value.

The variation of switching frequency with input voltage and load current is shown in Figures 6 and 7. The sharp rise in frequency at low output currents happens because the output transistor of the LM100 (Q_{12}) begins to supply an appreciable portion of the load current directly.

The efficiency of the regulator over a wide range of input voltages and output currents is given in Figures 8 and 9.

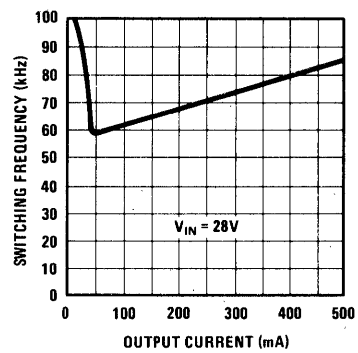


FIGURE 7

Switching Frequency as a Function of Output Current

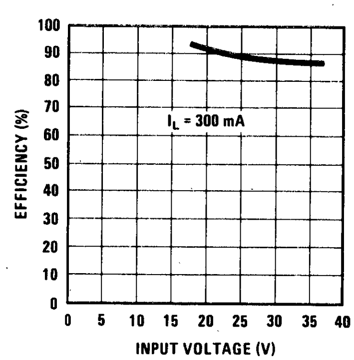


FIGURE 8

Efficiency as a Function of Input Voltage

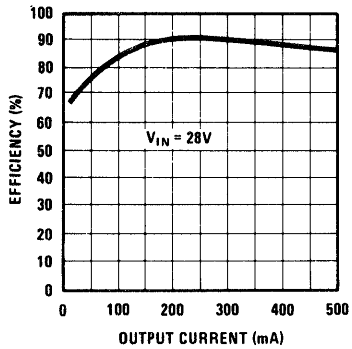


FIGURE 9
Efficiency as a Function of Output Current

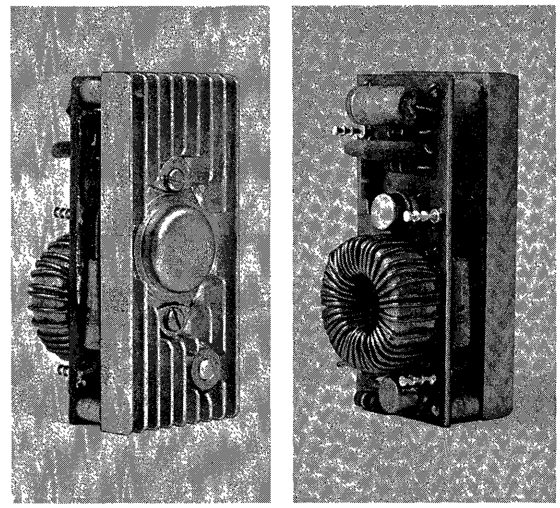


FIGURE 11
High Current Switching Regulator

HIGHER CURRENT REGULATORS

If output currents greater than about 500 mA are required, it is necessary to add another switch transistor to obtain more current gain. This is illustrated in Figure 10. With the exception of the added NPN power switch, Q_2, this circuit is the same as that described previously.

A photograph of a high-current regulator is shown in Figure 11. It is capable of delivering output currents of 3A continuously with only a small heat sink. Figure 12 shows that the efficiency is better than 80 percent at this level. Output currents to 5A can be obtained at reduced efficiency. However, the case temperature of the power switch and catch diode approach 100°C under this condition, so continuous operation is not recommended unless more heat sink is provided.

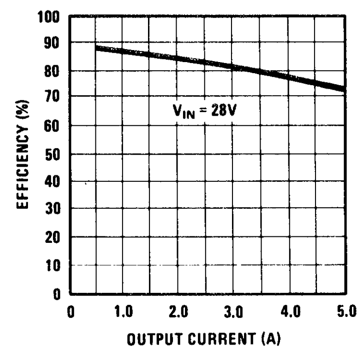
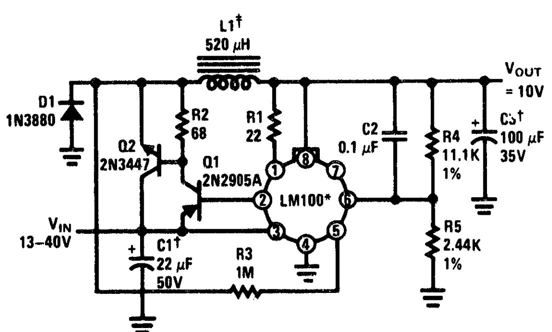


FIGURE 12
Efficiency as a Function of Output Current



*Basing diagram is Top View
†Solid tantalum
‡60 turns =20 on Arnold Engineering A930157-2 molybdenum permalloy core

FIGURE 10
Switching Regulator for Higher Output Currents

Figure 13 shows that the efficiency is not significantly affected by input voltage. In Figure 14 it can be seen that the switching frequency is fairly constant over a wide range of input voltages. Figure 15 shows that the switching frequency increases with increasing load current. The higher dc current through the inductor reduces the incremental inductance causing the frequency to go up. The last graph, Figure 16, illustrates the line regulation of the device. this can be improved by putting a small capacitor (0.01 μF) in series with the positive feedback resistor, R_3, to isolate the reference terminal from the dc input voltage changes.

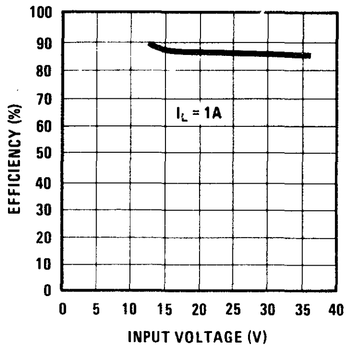


FIGURE 13
Efficiency as a Function of Input Voltage

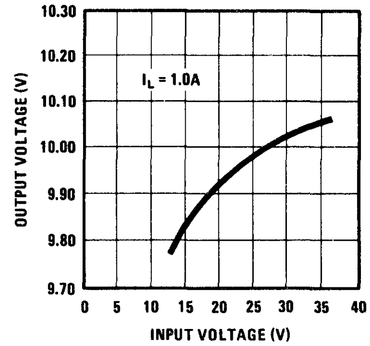


FIGURE 16
Line Regulation

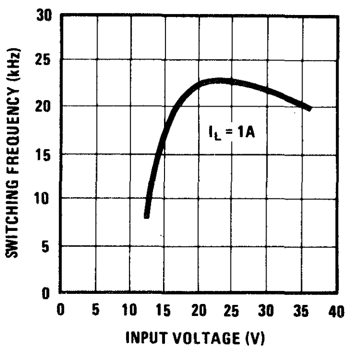


FIGURE 14
Variation of Switching Frequency with Input Voltage

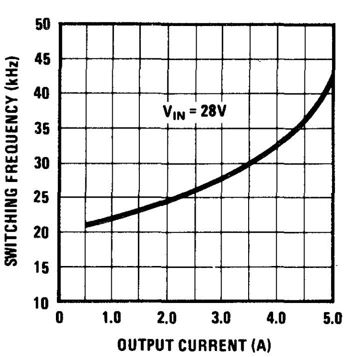


FIGURE 15
Variation of Switching Frequency with Output Current

At low output currents the inductor current can drop to zero at some time after the switch transistor turns off. When this happens, ringing occurs on the switching waveform. This is perfectly normal and causes no ill effects.

The use of solid tantalum capacitors for C_1 and C_3 is recommended when the regulator is expected to perform over the full military temperature range. The reason for using 35V capacitors on the output, even though the output voltage is only 10V, is that the 40 mV peak-to-peak ripple on the output would, for example, exceed the ratings of a 100 μF, 15V capacitor.

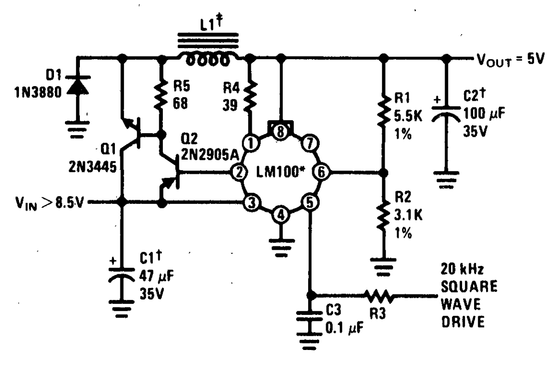
Aluminum electrolytic capacitors have been used successfully over a limited temperature range. And there is basically no reason why wet foil or wet slug tantalums could not be used as long as their equivalent series resistance is low enough so that they behave like capacitors with the high frequency switched-current waveform. It is also important that manufacturer's data be consulted to insure that they can withstand the high frequency ripple.

As was mentioned with the low current regulator, it is necessary to use fast-switching diodes and transistors in these circuits. Ordinary silicon rectifiers or low-frequency power transistors will operate at drastically-reduced efficiencies and will quickly overheat in these circuits.

DRIVEN SWITCHING REGULATOR

When a number of switching regulators are used together in a system it is sometimes desirable to synchronize their operation to more uniformly distribute the switched current waveforms on the input line. Synchronous operation is also wanted when a switching regulator is operated in conjunction with a power converter.

A circuit for synchronizing the switching regulator with a square wave drive signal is shown in Figure 17. In this circuit, positive feedback is not used. Instead, the square wave drive signal is integrated; and the resulting triangular wave (about 40 mV peak-to-peak) is applied to the reference bypass terminal of the LM100. This triangular wave will cause the regulator to switch since its gain is so high that the waveform overdrives it. The duty cycle of the switched waveform is controlled by the voltage on the feedback terminal, Pin 6. If this voltage goes up, the duty cycle will decrease since it is picking off a smaller portion of the triangular wave on Pin 5. By the same token, the duty cycle will decrease if the voltage on Pin 6 drops.



*Basing diagram is Top View

†Solid tantalum

†100 turns #22 on Arnold Engineering A930157-2 molybdenum permalloy core

FIGURE 17
Driven Switching Regulator

This action produces the desired regulation: if the output voltage starts to go up, it will raise the voltage on Pin 6 such that a smaller portion of the triangular wave is picked off. This reduces the duty cycle, counteracting the output voltage increase.

In order for this circuit to work properly, the ripple voltage on Pin 6 should be less than a quarter of the peak-to-peak amplitude of the triangular wave. If this condition is not satisfied, the regulator will try to oscillate at its own frequency. Further, since the resistance looking into Pin 5 is about 1 kΩ, the integrating capacitor, C_3, should have a capacitive reactance of less than 100Ω at the drive frequency. The value of R_3 is determined so that the amplitude of the triangular wave on Pin 5 is about 40 mV.

Driven regulators also have other advantages. For one, it is possible to design the LC filter independent of switching frequency considerations. Hence, lower output ripple and better transient response can be realized. A second advantage is the frequency stability. In a self-oscillating regulator, the switching frequency is controlled by a relatively large number of factors. As a result, it is not well determined when normal tolerances are taken into account. With low and medium power regulators, this is not usually a problem since the efficiency does not vary greatly with frequency. However, high power regulators tend to be more frequency sensitive and it is desirable to operate them at constant frequency.

CURRENT LIMITING

In the circuits described previously, the regulator is not protected from overloads or short-circuited output. Providing short-circuit protection is no simple problem, since it is necessary to keep the regulator switching when the output is shorted. Otherwise, the dissipation will become excessive even though the current is limited.

A circuit that does this is shown in Figure 18. The peak current through the switch transistor is sensed by R_6. When the voltage drop across this resistor becomes large enough to turn on Q_3, the output voltage begins to fall since current is being supplied to the feedback terminal of the regulator from the collector of Q_3 so less has to be supplied from the output through R_1. Furthermore, the circuit will continue to oscillate, even with a shorted output, because of positive feedback through R_6 and the relatively-long discharge time constant of C_2.

It is necessary to put a resistor, R_7, in series with the base of Q_3 to insure that excessive current will not be driven into the base. In addition, a capacitor, C_4, must be added across the input of Q_3 so that it does not turn on prematurely on the large current spike (about twice the load current) through the switch transistor caused by pulling the stored charge out of the catch diode. A zener diode bias supply must also be used on the output of the LM100 since the current limiting will not work if the voltage on this point drops below about 1V.

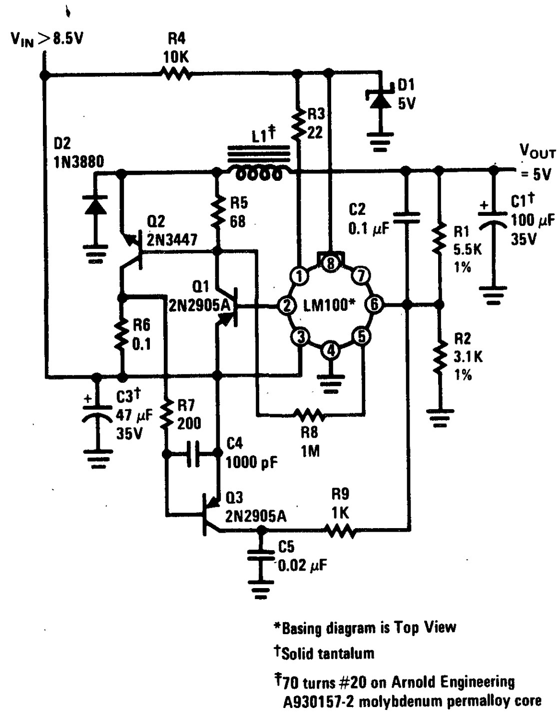


FIGURE 18
Switching Regulator with Current Limiting

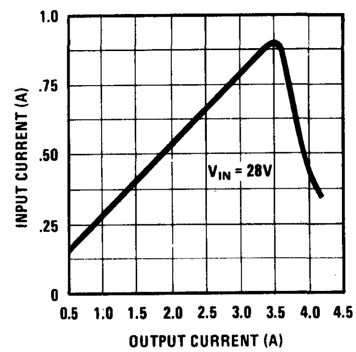


FIGURE 20
Illustrating Drop in Input Current as Regulator Goes Into Limiting

The current limiting characteristics of this circuit are shown in Figure 19. Figure 20 shows how the average input current actually drops off as the circuit goes into current limiting.

This current limiting scheme protects the switching transistors from overload or short-circuited output. However, the drop-out current and short-circuit current are not well controlled, so it is difficult to prove that the circuit will sustain a continuous short circuit under worst-case conditions. This is particularly true with high current regulators where the required amount of overdesign can become quite expensive.

Figure 21 shows a circuit which is more easily designed for continuous short-circuit protection under worst-case conditions. In this circuit, the current-sensing resistor is located in series with the inductor. Therefore, the peak-limiting current can be more precisely determined since the current spike generated by pulling the stored charge out of the catch diode does not flow through the sense resistor.

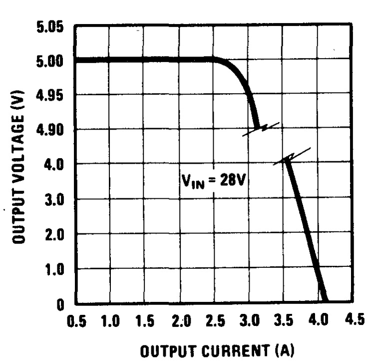


FIGURE 19
Current Limiting Characteristics

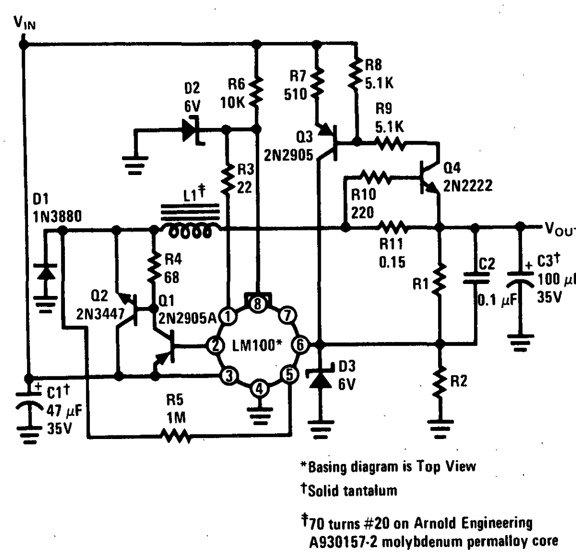


FIGURE 21
Switching Regulator with Continous Short-Circuit Protection

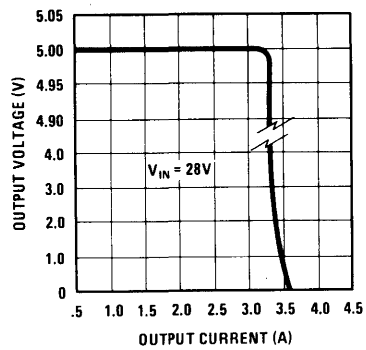


FIGURE 22
Current Limiting Characteristics

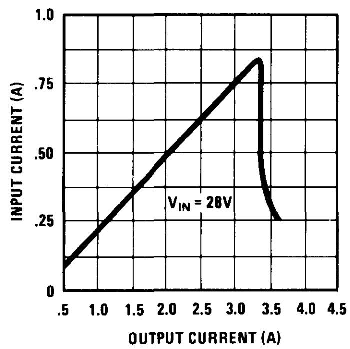


FIGURE 23
Plot of Input Current as Regulator Goes Into Limiting

Operation of this circuit is essentially the same as the previous one in that an NPN transistor, Q_4, senses the overcurrent condition and turns on Q_3 which supplies the current-limit signal to the feedback terminal. The zener diode, D_3, is required on the feedback terminal to guarantee that this terminal cannot go more than 0.5V higher than Pin 1. If this does happen, the circuit can latch up and burn out. The performance of this current-limiting scheme is illustrated in Figures 22 and 23.

With this circuit it is not only possible to more accurately determine the limiting current, but as can be seen from Figures 22 and 23, the limiting characteristic is considerably sharper. One disadvantage of this circuit is that the load current flows continuously through the current sense resistor, reducing efficiency. As an example, with a 5V regulated output the efficiency will be reduced by 10 percent at full load.

NEGATIVE REGULATORS

All circuits discussed thus far are for regulators with positive outputs. Although negative regulators can be obtained by floating the unregulated supply and grounding the output, this is not always convenient.

Figure 24 shows a circuit for a negative switching regulator where the unregulated input and regulated output have a common ground. The only limitation of the circuit is that there must be a positive voltage greater than 3V available in order to properly bias the negative regulator.

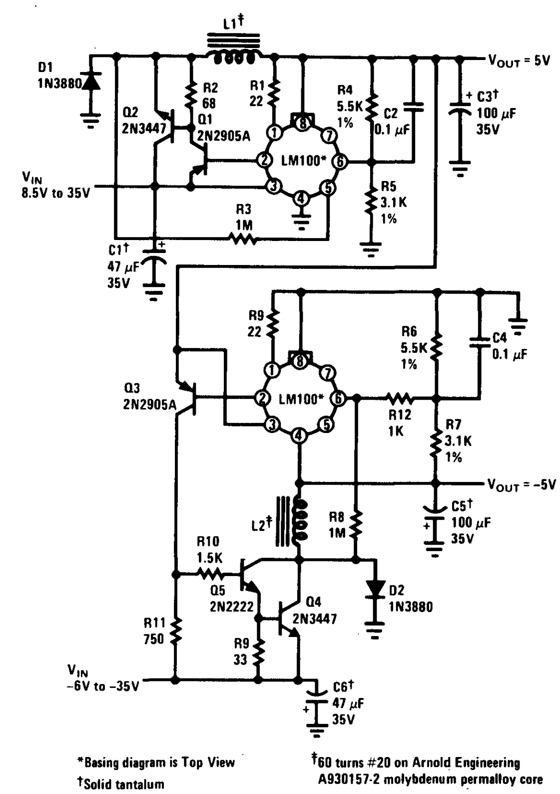


FIGURE 24
Positive and Negative Switching Regulators

In this circuit, the normal output terminal of the LM100 (Pin 8) is grounded and the ground terminal (Pin 4) is connected to the regulated negative output. Hence, as before, it regulates the voltage between the output and ground terminals. The unregulated input terminal (Pin 3) is run from a positive voltage for proper biasing. A PNP booster transistor, Q_3, is connected in the normal manner; and it drives a Darlington-connected NPN switch. Positive feedback is developed by the resistive divider, R_8 and R_{12}.

It is necessary to use a Darlington switch even though the current gain is not needed. The power switch transistor, Q_4, cannot be operated with a fixed base drive: if the base drive is made large enough to insure saturation at maximum load current, it will overstore so badly at lower currents that the output ripple will increase radically. With the extra transistor, however, it is kept out of saturation at lower output currents, eliminating the problem.

HIGH VOLTAGE REGULATORS

With switching regulators, an application can easily arise where the input voltage can be higher than the 40V maximum rating of the LM100, even though the output voltage is within the 30V maximum. As shown in Figure 25, it is possible to isolate the LM100 from the unregulated supply so that it can be used with input voltages limited only by the switch transistors and the catch diode.

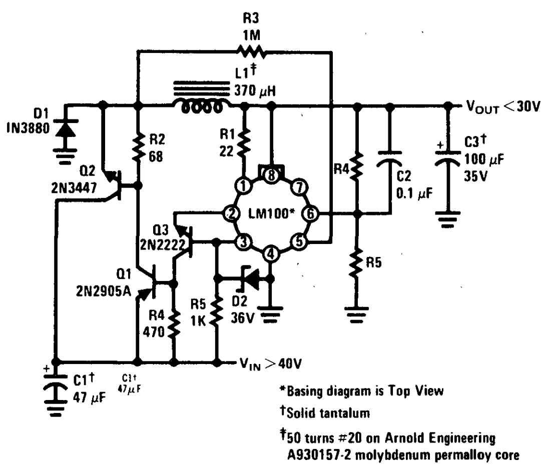


FIGURE 25
Switching Regulator for High-voltage Inputs

In this circuit, the voltage seen by the LM100 is maintained at a fixed level within ratings by the zener diode, D_2. The zener voltage must be at least 3V greater than the output voltage. The output of the LM100 is level-shifted up to the input voltage by an additional NPN transistor, Q_3, which is operated common base. This drives the PNP switch driver in the normal manner.

SWITCHING AND LINEAR REGULATOR COMBINATION

In certain applications, the output ripple and load transient response requirements rule out the use of a switching regulator, yet the input-output voltage differential is still high. In this case, a power converter might be used to reduce the input voltage and this reduced voltage would be regulated by a linear regulator. This arrangement, however, is not nearly as efficient as the switching and linear regulator combination shown in Figure 26. The switching regulator not only reduces the input voltage with high efficiency, but it also regulates it. Therefore, the linear regulator operates with a fixed input-output voltage differential which holds dissipation to a minimum.

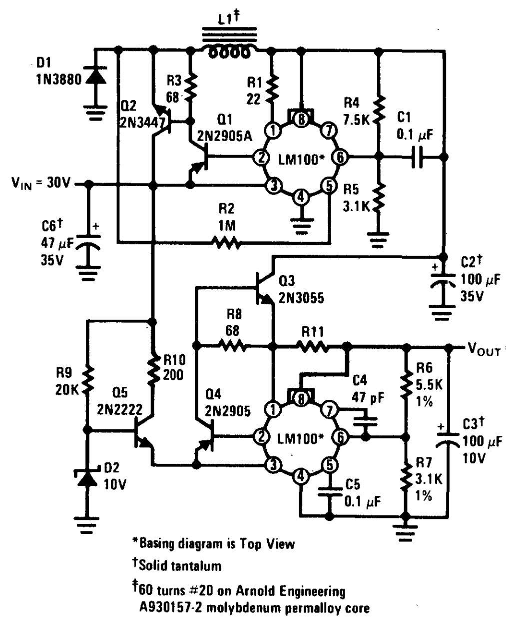


FIGURE 26
Switching and Linear Regulator Combination for Obtaining Very Low Ripple and Fast Transient Response

In this circuit, the linear regulator is biased by a zener pre-regulator (R_9, D_2 and Q_5) to isolate it from noise on the unregulated supply. This separate bias supply permits the linear pass transistor, Q_3, to operate right down into saturation. The collector of Q_3 is supplied by the output of a switching regulator which is made enough higher than the linear regulator output to allow for the maximum overshoot of the switching regulator plus the saturation of Q_3.

SUMMARY

A number of switching regulator circuits which use a readily-available monolithic voltage regulator as the voltage reference and control circuitry have been described. These regulators are useful over a 2V to 30V range for either positive or negative supplies. Although the discussion was limited to circuits providing maximum output currents from 100 mA to 5A, it is possible to obtain even higher output currents. The output current is, in fact, limited by the discrete components – not by the basic design or the integrated circuit.

The majority of the circuits shown were self-oscillating regulators; however, a method of driving the regulator in synchronism with an external clock signal was given. In addition, circuits which provide overload protection, limiting both the output current as well as the power dissipation, were presented. The performance of the regulator circuits was described in detail, and a design procedure was outlined. Suggestions were also made on the selection of components for switching regulators.

The circuits which have been described here for the LM100 work equally well with the LM200 or the LM300. These devices are identical, except that the LM200 is specified over a $-25^\circ C$ to $85^\circ C$ temperature range and the LM300 is specified from $0^\circ C$ to $70^\circ C$ instead of the $-55^\circ C$ to $125^\circ C$ temperature range for the LM100.