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APPLICATION NOTE

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A Low-Noise Low Drop-Out Regulator for Portable Equipment

by James Wong

Designed for laptop computers and portable instruments, this low-noise, low drop-out regulator achieves all objectives of small size, low power and safe operation.

The primary goal of designing a battery power supply system for laptop computers or portable instruments is to maximize operating time before recharging becomes necessary. Translating the goal into design criteria, the system must have high efficiency and extraordinarily low standby power drain, not only during normal operation, but also during potential short-circuit occurrences. In addition, the regulator circuit must be capable of recovering rapidly and maintaining a stable output even when sudden surges of current occur in the load.

A linear series-pass regulator is particularly attractive when compared to a switchmode design. The linear design offers the following features: low noise, fast response, simplicity, and comparable efficiency. Low noise is inherent because the linear supply has no fast switching circuits. This feature is especially desirable for powering the linear circuits used in sensitive portable instruments where high frequency noise can easily enter high impedance circuits and destroy accuracy. The system can be designed for as wide a bandwidth as necessary, determined by the choice of op amp and its operating current. The linear circuit is quite simple compared to a switcher; fewer component parts helps ensure a reliable design. The full load efficiency can be designed to be comparable to a switcher because the input battery voltage can be chosen to be only slightly higher than the desired output and the input does not vary over wide limits. Furthermore, the zero load quiescent current can be made much lower than a switcher by the proper choice of components.

Figure 1 shows a complete 5V, 1A low drop-out linear voltage regulator that can operate at up to 90% conversion efficiency near drop-out voltage, while drawing less than 50µA of standby current. The design owes its low operating current to several contributing factors. The heart of the regulation loop is an ultra-low supply current op amp, the OP-90, which typically draws only 15µA of supply current. Another OP-90 is used to generate a precision bandgap reference that consumes less than 20µA current. Instead of using a bipolar power transistor for the pass element, a MOSFET power transistor is used. The reasoning behind selecting the MOSFET is that the MOSFET gate draws only picoamps rather than milliamps of steady-state current. In addition, a P-channel MOSFET allows the regulator to operate with an exceptionally low drop-out voltage and still maintain regulation. The details of the design are discussed in the following sections.

The Power Transistor

Using a P-channel MOSFET permits a low drop-out design to be easily implemented. The gate drive is simple because the FET is turned on when the gate is returned to ground. Several types of FETs are available that can handle amperes of current with a gate-source potential of only 5V.

To allow sufficient headroom for maintaining regulation the following inequality must be satisfied:

$$V_B \geq V_o + I_o\, r_{DS(on)}$$

where

V_B – Battery voltage at the limit selected for % discharge
V_o – Output voltage of regulator
I_o – Output current of regulator
$r_{DS(on)}$ – drain-source on-resistance of MOSFET

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The choice of MOSFET used is, as usual, a compromise between conflicting goals. A lower $r_{DS(on)}$ transistor allows V_B to more closely approach V_0. The penalty paid is higher part cost and higher capacitance, which lowers bandwidth. For the design example of *Figure 1*, the IRF 9530 MOSFET was chosen. It has a specified maximum $r_{DS(on)}$ of 0.5Ω. At operating temperature this provides enough headroom to operate satisfactorily when V_B is 6V and I_o is 1A.

The Reference Voltage

To achieve a stable, accurate regulation loop, a fairly precise reference voltage needs to be generated. It is difficult to find a reasonably accurate reference using a zener or a bandgap device that operates at microamps of current. In particular, low power zener diodes generally have poor dynamic resistance characteristics and poor initial device-to-device tolerance. Thus, as the battery voltage varies so will reference voltage accuracy.

However, a pair of matched transistors (MAT-01), closing the loop around an amplifier, easily forms a stable bandgap circuit, as shown in *Figure 1*. The circuit typically draws 17µA of quiescent current yet exhibits low temperature drift characteristics. The circuit requires one adjustment to set the bandgap voltage of 1.230V at the amplifier output. This calibration also establishes a low temperature drift coefficient point. Its typical drift is 5.5µV/°C over a −40°C to +85°C temperature range.

To achieve this high degree of precision, the proper passive components must be used. All resistors, except the one to the FET gate, must be 1% metal film types with a temperature coefficient of 50 ppm/°C or better supplied from the same manufacturer. The trimpot must have a similarly low temperature coefficient. It should also be a multiturn type in order to permit an accurate voltage adjustment.

In applications where low supply current is not of critical importance, the reference trim can be eliminated altogether using a monolithic precision reference device such as the REF-43 in *Figure 2*. The device's quiescent current consumption is on the order of 450µA. Besides offering a temperature stable (10ppm/°C) 2.5V output, its initial tolerance is 0.2% maximum over the full temperature range, which renders trimming unnecessary. In addition, it can tolerate an input drop to as low as 4.5V and still maintain regulation. Finally, the REF-43 has a temperature output that can be used to temperature compensate the circuit, saving components over a discrete circuit scheme.

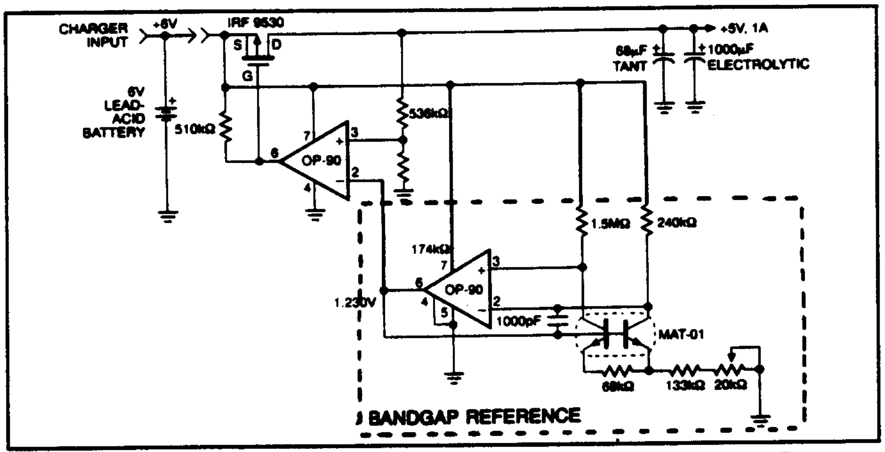

Figure 1. A Low Drop-Out Regulator That Supplied 5V, 1A Output. The Circuit Draws 50µA Standby Current.

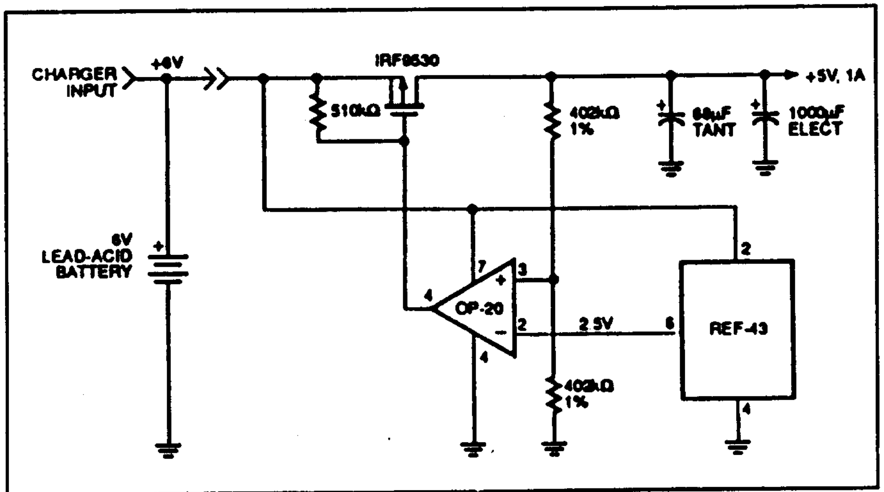

Figure 2. Less Critical Supply Current Applications Can Use an Adjustment-Free Design Using a Precision 2.500V Reference Supplies by the REF-43. Quiescent Current of the REF-43 Is 450µA.

The Op Amp

For this design an ultra-low current, single supply op amp is used to regulate the loop. Of the many op amp types available, few can match the ultra-low 15µA supply current of the OP-90. In addition, the OP-90 has excellent output swing range, which is critical when operating on only 6V of supply voltage. For example, in the case where the regulator is not delivering any load current, the op amp's output must swing to within one volt of the positive supply rail in order to cut off the FET gate drive. Conversely, when a rated one amp current flows, the amplifier output must swing sufficiently low, to within one volt of ground, to ensure adequate enhancement drive to the FET.

The regulation loop is closed by feeding back a portion of the 5V regulated output to the noninverting input of the op amp. Negative feedback is accomplished because the P-channel MOSFET provides a phase inversion.

AC characteristics of the regulation loop are important whenever the load goes through a sudden disturbance. The lower the bandwidth the slower the recovery to a loop perturbation. *Figure 3* shows how the regulator of *Figure 1* reacts to such a load change. Starting from no load, a step change of 150mA of load current was applied to the regulator. Notice in *Figure 3a*, Trace 2, the output initially dipped by some 80mV. It took about 2msec before it recovered to a stable state. The amount of glitch noise and slow recovery may not be acceptable for certain equipment such as precision instruments where glitch noise in the supply can disrupt a sensitive measurement, resulting in errors. On the other hand, if the regulator is used to power logic circuits, the 80mV glitch is usually well within their noise margin.

Faster response can be obtained by substituting an OP-20 as the error amplifier. While the OP-20 consumes 50µA, versus 15µA for the OP-90, its bandwidth is considerably higher. The middle trace of *Figure 2b* shows that both glitch amplitude and recovery time improved significantly. The glitch amplitude is reduced from 80mV to about 35mV and its recovery time shortens from 2msec to 0.5msec.

The bottom trace of the photos of

Figure 2 shows how the two regulator circuits behave when a 150mA step current is applied on top of a normal 750mA quiescent load, which more realistically simulates the real-world condition. Notice in both cases, the peak glitch amplitude is on the order of 10mV, a much more tolerable behavior. The OP-20's performance shows a dramatic improvement in recovery time.

Battery Selection and Charging

The need for laptop computers or portable instruments to run for hours is a foremost design consideration. Advancements in CMOS circuits and battery technology today allow laptop computers to operate for as long as 4 to 6 hours without recharging.

Lead-acid batteries have several desirable characteristics that have caused sealed small cell units to be developed for portable products. One of their characteristics is that they will tolerate a relatively fast charging time, as fast as 2 hours, to reach 90% of capacity. Another is relatively high capacity for the weight and volume. Because of their extremely low internal impedance, they can deliver many amperes of current in a short period of time, although some care needs to be exercised to not deep-drain the battery. In addition, unlike Ni-Cad batteries, lead-acid cells do not have a "memory" problem, which after a while reduces the batteries' capability to recharge to full capacity, resulting in markedly reduced operating time.

Because of the regulator's low drop-out characteristic, it is well suited for use with a 6V lead-acid battery. The combination minimizes power loss and therefore maximizes conversion efficiency, resulting in considerable size and weight savings, particularly in the battery.

A unique characteristic of the lead-acid battery is that it requires a constant voltage source to charge as opposed to a current source charger for Ni-Cad batteries. Consequently, the same voltage regulator design as in *Figure 2* can be used. The battery requires a higher charging voltage of 7.3V so that it can be left float-charging indefinitely to maintain a full charge. A voltage-mode charging scheme also allows easy implementation of a temperature compensation circuit to achieve maximum charging efficiency.

For lead-acid batteries, the charging efficiency decreases as the battery temperature increases. Since charging and ambient environment can increase the battery temperature, a longer than expected charging time occurs. Applying a temperature-dependent charging voltage that has a coefficient of −2.5mV/°C/cell speeds up charging.

To implement a charging circuit that has this temperature coefficient, the previously mentioned temperature output of the REF-43 is utilized as shown in *Figure 4*. Its temperature output pin (pin 3), whose output impedance is quite high, is necessarily buffered with a unity gain amplifier. This pin normally outputs a DC voltage of 630mV at room temperature, with a temperature coefficient of +1.9mV/°C. This voltage is summed into the error amplifier that otherwise regulates at an output voltage of 7.3V. Because of phase inversion, the temperature coefficient of +1.9mV/°C is converted to a −2.5mV/°C at the regulated output by ratio of the 68.1kΩ and the two voltage divider resistors. To be sure the temperature is being measured accurately, it is necessary for the REF-43 to be in direct thermal contact with the battery body.

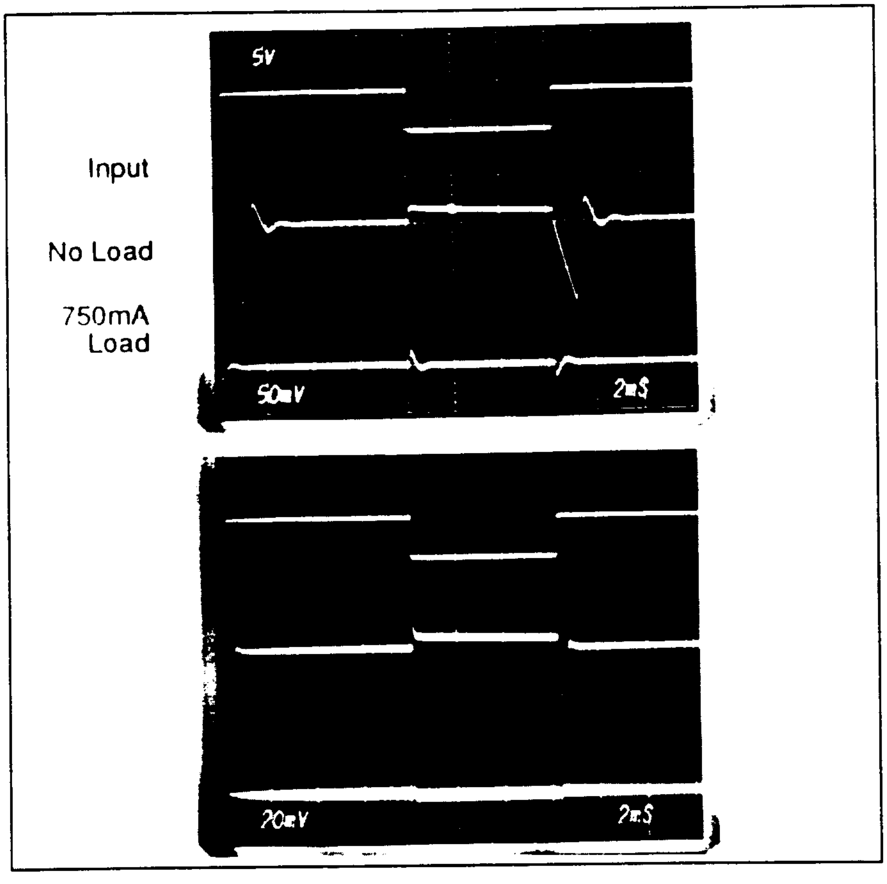

Figure 3. Comparing a Regulator Load Recovery Between a Slower OP-90(a) and a Wider Bandwidth OP-20(b). The Top Trace Is the Control Signal to the Dynamic Load, Shown for Timing Reference. The Middle Trace Shows the Transient Recovery When a Step Current From No Load to 150mA Is Applied. The Bottom Trace Shows the Output When a Step Change From a DC 750mA to 900mA Occurs.

Current-Limiting Options

Because lead-acid batteries are capable of many amperes of current output for short periods of time, a short-circuit can potentially damage the cell by overheating. In addition, it is not a good idea to allow a deep discharge, as it reduces the charge capacity during subsequent recharges. A simple current-limiting circuit can provide the necessary protection. Two alternative arrangements, easily implemented into the circuits of *Figure 1* or *Figure 2*, are shown in *Figure 5*. To simplify the schematics the reference voltage circuit is omitted.

The first, and simplest, is a constant current limiter using a transistor acting as a threshold switch, plus a current sense resistor and a pull-up resistor. During normal operation, the regulator load current develops less than the 0.6V threshold voltage across the 0.6Ω sense resistor that is necessary to turn on the PNP transistor. Therefore normal voltage regulation ensues. However, as the load current exceeds the threshold of about 1.1A, which develops more than 0.6V across the sense resistor, the PNP transistor begins to draw current and thus pull up the gate voltage which prevents the MOSFET from passing more current. In this manner the PNP will maintain the current at the limit level, to the extreme point where the regulator output voltage begins to drop. Since only a low level of PNP collector current is required, the transistor need only be a general-purpose type, such as a 2N3905.

Although simple in implementation, there is a serious drawback with this design. During normal operation, the load current drawn might be, say 0.75A. It would develop 0.45V across the sense resistor, which is in series with the pass element. In effect, this 0.45V reduces the drop-out "headroom" by the same amount, which is a serious drawback in a low drop-out design.

The second alternative is shown in *Figure 5b*. Although it costs a bit more circuitry, it is far better in terms of preserving the low drop-out feature. Instead of a 0.6Ω sense resistor, a lower 0.1Ω resistor is used. Thus at the current limit threshold of 1.1A, it only degrades the drop-out voltage by 0.11V, which is far more tolerable than the 0.66V in the other circuit.

The limit threshold voltage is set up by the voltage divider at the noninverting input of the current limit detector/amplifier. Sensing is done by the sense resistor plus the second voltage divider. If the threshold limit is not reached, the inverting input has a higher voltage than the noninverting input, causing the current limit amplifier output to go low, reverse-biasing the diode, thus opening the amplifier loop. Normal voltage regulation ensues. Once a 0.11V is developed across the sense resistor, which is the threshold of the current limit, the amplifier output rises to close the loop. Its low output resistance drives the gate of the MOSFET pass element high to maintain the current limit. This design preserves a low drop-out voltage that is critical to an efficient power system.

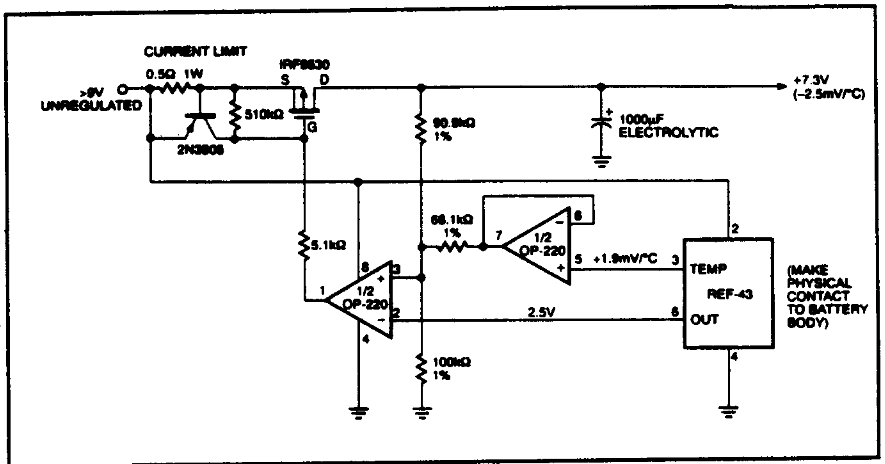

Figure 4. Lead-Acid Battery Requires Simple Regulated Voltage Charging Circuit. Optimum Charging Efficiency Can Be Achieved by Providing a Temperature Sense and Compensation. Charging Voltage Has Temperature Coefficient of −2.5mV/°C/Cell.

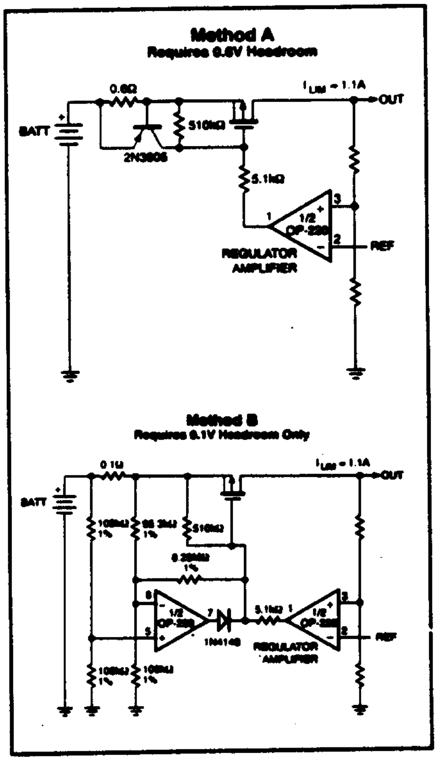

Figure 5. Two Alternate Methods of Implementing Current Limiting. Method A(a) Is Simpler But Has Poor Drop-Out Characteristic. Method B(b) Is Preferred Because It Preserves Low Drop-Out Features.

Even though constant current limiting saves the battery from self-destruction, sufficient heat sinking must be provided for the power MOSFET transistor in order to prevent it from being damaged under a sustained short-circuit condition. That is, although the output current is limited to 1.1A, a short circuit forces nearly 6V across the power MOSFET causing it to dissipate 6½W.

Foldback current limiting can be used to eliminate the need for a heat sink. Fortunately, such a feature comes free with the design. By moving the previously described current limit sensing circuit from the input side of the MOSFET to the output side as shown in *Figure 6*, foldback current limiting action results. The current limiting action initially works the same way as previously described. The foldback action kicks in as the current limit of 1.1A is reached, the output voltage begins to drop, pulling the threshold voltage on pin 6 down proportionally. As it drops, less and less voltage is needed across the sense resistor to trip the threshold. In effect, the regulated current "folds" back as the output voltage drops. For example, when the output drops to 1V, the output current drops to about 0.2A.

The significance of this design is that the power MOSFET is called on to dissipate less than 1W instead of more than 6W of power when in short circuit. The worst-case dissipation is indeed 2W when the output is three volts and the load current is 0.65A. This reduces heat sink size and improves operating reliability. Thus the design achieves all objectives of small size, low power and safe operation.

References

1. Derek Bowers, Art Kappor, Andrew Jenkins, "Low Power Op Amp Delivers Precision at Low Signal Levels," EDN January 1987.
2. Precision Monolithics, Inc. 1989 Analog Applications Seminar Handbook.

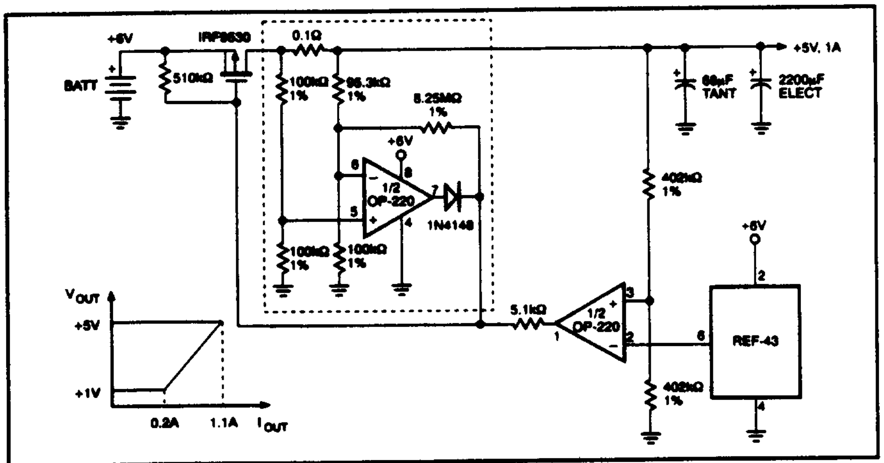

Figure 6. By Moving the Current Limiting Circuit to the Output Side of the Regulation Loop, a Foldback Effect Is Achieved. As the Output Voltage Drops, the Output Current Also Drops, Reducing Power Dissipation in the Power MOSFET Device.