A DESIGNER'S GUIDE TO THE L200 VOLTAGE REGULATOR

Delivering 2A at a voltage variable from 2.85V to 36V, the L200 voltage regulator is a versatile device that simplifies the design of linear supplies. This design guide describes the operation of the device and its applications.

The introduction of integrated regulator circuits has greatly simplified the work involved in designing supplies. Regulation and protection circuits required for the supply, previously realized using discrete components, are now integrated in a single chip. This has led to significant cost and space saving as well as increased reliability. Today the designer has a wide range of fixed and adjustable, positive and negative series regulators to choose from as well as an increasing number of switching regulators.

The L200 is a positive variable voltage regulator which includes a current limiter and supplies up to 2A at 2.85 to 36V.

The output voltage is fixed with two resistors or, if a continuously variable output voltage is required, with one fixed and one variable resistor.

The maximum output current is fixed with a low value resistor. The device has all the characteristics common to normal fixed regulators and these are described in the datasheet. The L200 is particularly suitable for applications requiring output voltage variation or when a voltage not provided by the standard regulators is required or when a special limit must be placed on the output current.

The L200 is available in two packages:

Pentawatt — Offers easy assembly and good reliability. The guaranteed thermal resistance ($R_{th\ j\text{-}case}$) is 3°C/W (typically 2°C/W) while if the device is used without heatsink we can consider a guaranteed junction-ambient thermal resistance of 50°C/W.

TO-3 — For professional and military use or where good hermeticity is required.
The guaranteed junction-case thermal resistance is 4°C/W, while the junction-ambient thermal resistance is 35°C/W.
The junction-case thermal resistance of this package, which is greater than that of the Pentawatt, is partly compensated by the lower contact resistance with the heatsink, especially when an electrical insulator is used.

CIRCUIT OPERATION

As can be seen from the block diagram (fig. 1) the voltage regulation loop is almost identical to that of fixed regulators. The only difference is that the negative feedback network is external, so it can be varied (fig. 3). The output is linked to the reference by:

$$V_{out} = V_{ref} \left(1 + \frac{R2}{R1}\right) \qquad (1)$$

Considering V_{out} as the output of an operational amplifier with gain equal to $G_V = 1 + R2/R1$ and input signal equal to V_{ref}, variability of the output voltage can be obtained by varying R1 or R2 (or both). It's best to vary R1 because in this way the current in resistors R1 and R2 remains constant (this current is in fact given by $V_{ref}/R1$).

(Equation (1) can also be found in another way which is more useful in order to understand the descriptions of the applications discussed.

$$V_{out} = R1\ i_1 + R2\ i_2$$

and since in practice $i_1 \gg i_4$ (i_4 has a typical value of 10 μA) we can say that

$$V_{out} = R1\ i_1 + R2\ i_1 \quad \text{with} \quad i_1 = \frac{V_{ref}}{R1}$$

Therefore

$$V_{out} = \frac{R2}{R1} V_{ref} + V_{ref} = V_{ref} \left(1 + \frac{R2}{R1}\right)$$

In other words R1 fixes the value of the current circulating in R2 so R2 is determined.

Fig. 1 - Block diagram

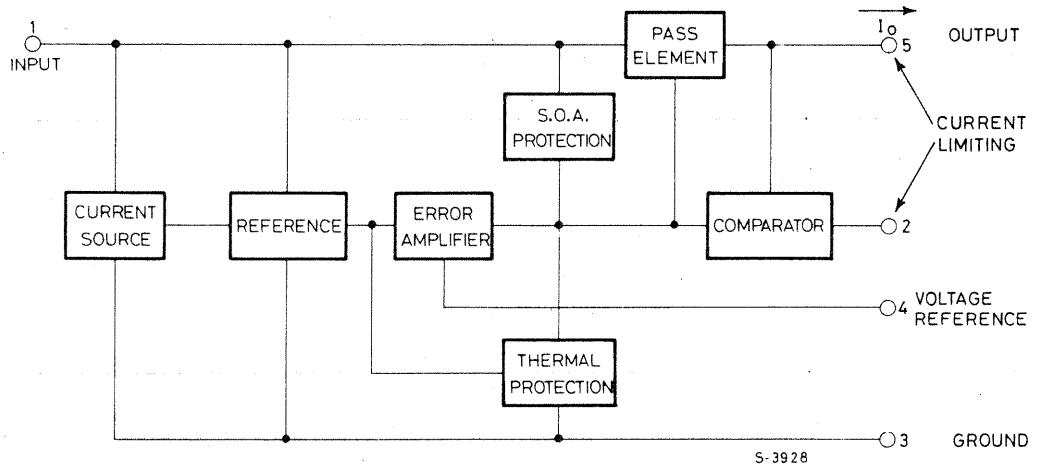

Fig. 2 - Schematic diagram

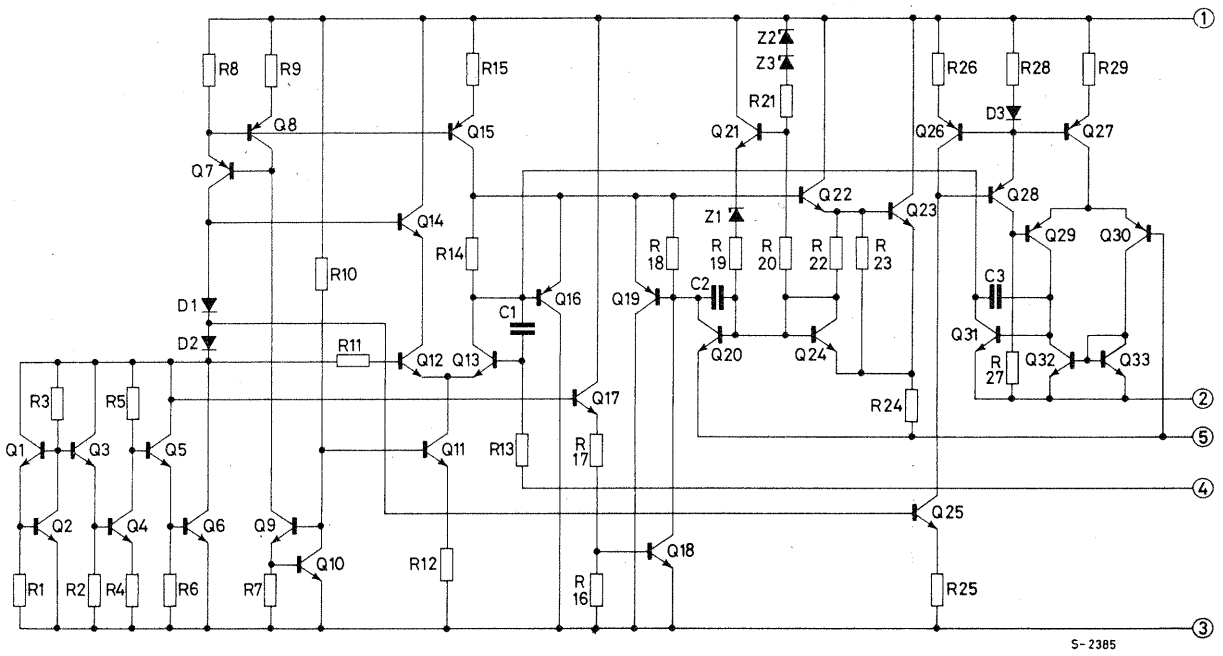

Fig. 3

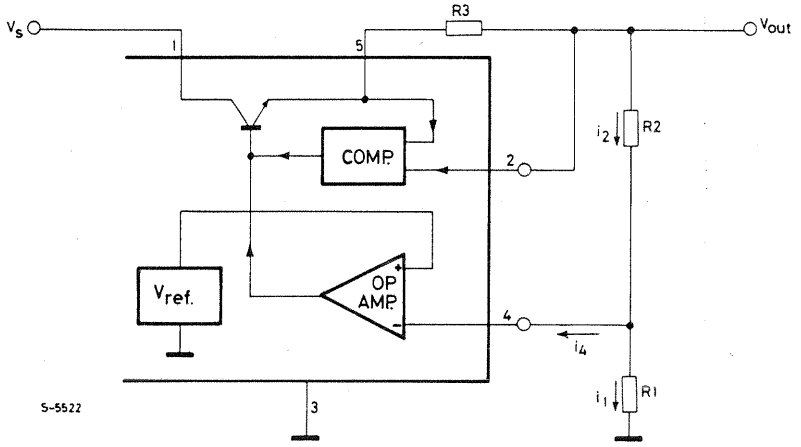

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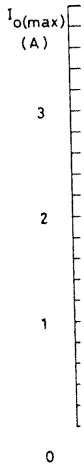
Fig. 4



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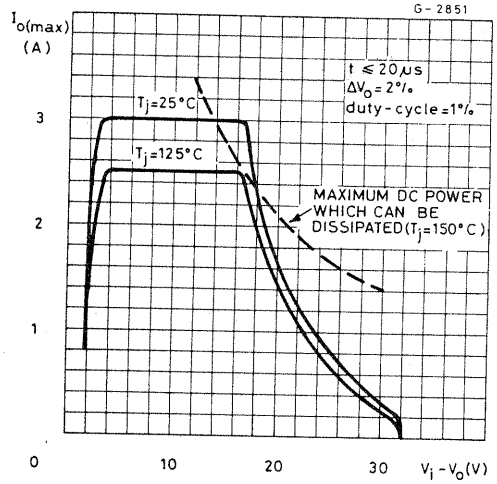
Overload protection

The device has an overload protection circuit which limits the current available.

Referring to fig. 2, R24 operates as a current sensor. When at the terminals of R24 there is a voltage drop sufficient to make Q20 conduct, Q19 begins to draw current from the base of the power transistor (darlington formed by Q22 and Q23) and the output current is limited. The limit depends on the current which Q21 injects into the base of Q20. This current depends on the drop-out and the temperature which explains the trend of the curves in fig. 4.

Fig. 4



Thermal protection

The junction temperature of the device may reach destructive levels during a short circuit at the output or due to an abnormal increase in the ambient temperature. To avoid having to use heatsinks which are costly and bulky, a thermal protection circuit has been introduced to limit the output current so that the dissipated power does not bring the junction temperature above the values allowed. The operation of this circuit can be summarized as follows.

In Q17 there is a constant current equal to:

$$\frac{V_{ref} - V_{BE17}}{R17 + R16} \qquad (V_{ref} = 2.75V \text{ typ})$$

The base of Q18 is therefore biased at:

$$V_{BE18} = \frac{V_{ref} - V_{BE17}}{R16 + R17} \cdot R16 \cong 350 \text{ mV}$$

Therefore at $T_j = 25°C$ Q18 is off (since 600 mV is needed for it to start conducting). Since the V_{BE} of a silicon transistor decreases by about 2 mV/°C, Q18 starts conducting at the junction temperature:

$$T_j = \frac{600-350}{2} + 25 = 150°C$$

Current limitation

The innovative feature of this device is the possibility of acting on the current regulation loop, i.e. of limiting the maximum current that can be supplied to the desired value by using a simple resistor (R3 in fig. 2). Obviously if R3 = 0 the maximum output current is also the maximum current that the device can supply because of its internal limitation.

The current loop consists of a comparator circuit with fixed threshold whose value is V_{sc}. This comparator intervenes when $I_o \cdot R3 = V_{sc}$, hence $I_o = \frac{V_{sc}}{R3}$ (V_{sc} is the voltage between pins 5 and 2 with typical value of 0.45V).

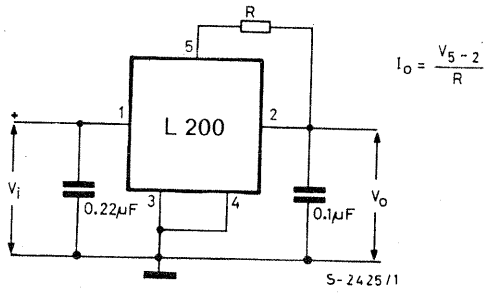
Special attention has been given to the comparator circuit in order to ensure that the device behaves as a current generator with high output impedance.

TYPICAL APPLICATIONS

Programmable current regulator

Fig. 5 shows the device used as current generator. In this case the error amplifier is disabled by short-circuiting pin 4 to ground.

Fig. 5



The output current I_o is fixed by means of R:

$$I_o = \frac{V_{5-2}}{R}$$

The output voltage can reach a maximum value $V_i - V_{drop} \cong V_i - 2V$ (V_{drop} depends on I_o).

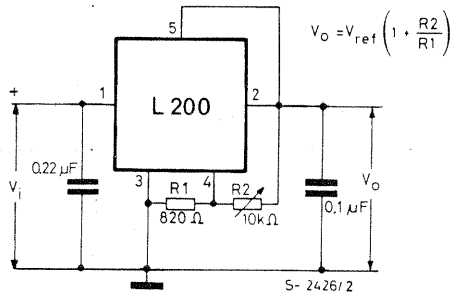
Programmable voltage regulator

Fig. 6 shows the device connected as a voltage regulator and the maximum output current is the maximum current that the device can supply. The output voltage V_o is fixed using potentiometer R2. The equation which gives the output voltage is as follows:

$$V_o = V_{ref}\left(1 + \frac{R2}{R1}\right)$$

By substituting the potentiometer with a fixed resistor and choosing suitable values for R1 and R2, it is possible to obtain a wide range of fixed output voltages.

Fig. 6



The following formulas and tables can be used to calculate some of the most common output voltages.

Having fixed a certain V_o, using the previous formula, the maximum value is:

$$V_{o\ max} = V_{ref\ max}\ (1 + \frac{R2_{max}}{R1_{min}})$$ and the minimum value is:

$$V_{o\ min} = V_{ref\ min}\ (1 + \frac{R2_{min}}{R1_{max}})$$

The table below indicates resistor values for typical output voltages:

V_o ± 4%	R1 ± 1%	R2 ± 1%
5V	1.5 kΩ	1.2 kΩ
12V	1 kΩ	3.3 kΩ
15V	750 Ω	3.3 kΩ
18V	330 Ω	1.8 kΩ
24V	510 Ω	3.9 kΩ

Programmable current and voltage regulator

The typical configuration used by the device as a voltage regulator with external current limitation is shown in fig. 7. The fixed voltage of 2.77V at the terminals of R1 makes it possible to force a constant current across variable resistor R2. If R2 is varied, the voltage at pin 2 is varied and so is the output voltage.

The output voltage is given by:

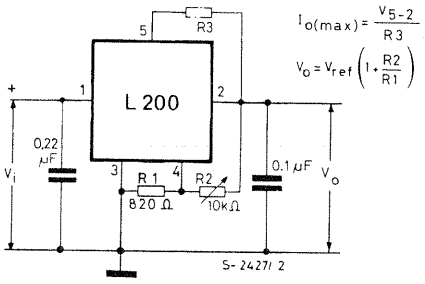
$$V_o = V_{ref} \cdot (1 + \frac{R2}{R1})$$, with V_{ref}= 2.77V typ

and the maximum output current is given by:

$$I_{o\ max} = \frac{V_{5-2}}{R3}$$ with V_{5-2} = 0.45V typ.

To maintain a sufficient current for good regulation the value of R1 should be kept low. When there is no load, the output current is $V_{ref}/R1$. Suitable values of R1 are between 500Ω and 1.5 kΩ. If the load is always present the maximum value for R1 is limited by the current value (10 µA) at the input of the error amplifier (pin 4).

Fig. 7

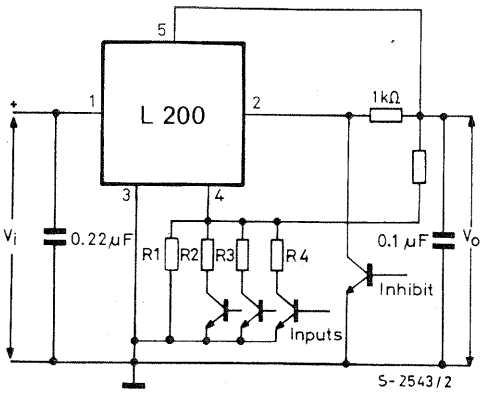


Digitally selected regulator with inhibit

The output voltage of the device can be regulated digitally as shown in fig. 8. The output voltage depends on the divider formed by R5 and a combination of R1, R2, R3 and P2. The device can be switched off with a transistor.

When the inhibit transistor is saturated, pin 2 is brought to ground potential and the output voltage does not exceed 0.45V.

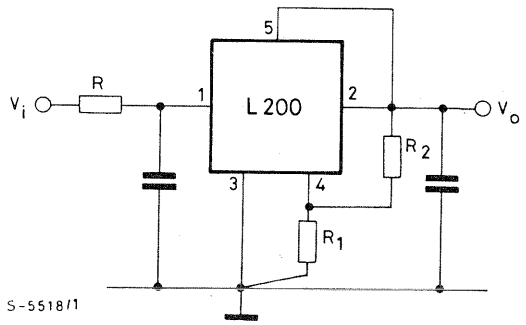
Fig. 8



Reducing power dissipation with dropping resistor

If may sometimes be advisable to reduce the power dissipated by the device. A simple and economic method of doing this is to use a resistor connected in series to the input as shown in fig. 9. The input-output differential voltage on the device is thus reduced.

Fig. 9



The formula for calculating R is as follows:

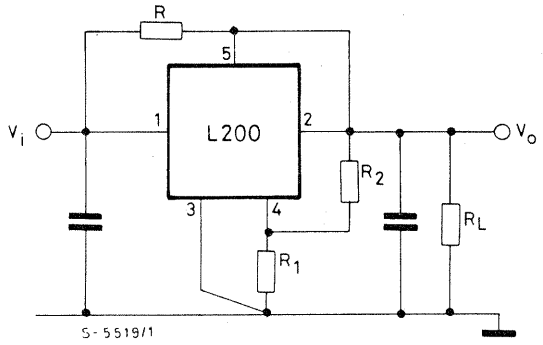
$$R = \frac{V_{i\,min} - (V_o + V_{drop})}{I_o}$$

Where V_{drop} is the minimum differential voltage between the input and the output of the device at current I_o. $V_{in\,min}$ is the minimum input voltage. V_o is the output voltage and I_o the output current.

With constant load, resistor R can be connected between pins 1 and 2 of the IC instead of in series with the input (fig. 10). In this way, part of the load current flows through the device and part through the resistor. This configuration can be used when the minimum current by the load is:

$$I_{o\,min} = \frac{V_{drop}}{R} \quad \text{(instant by instant)}$$

Fig. 10



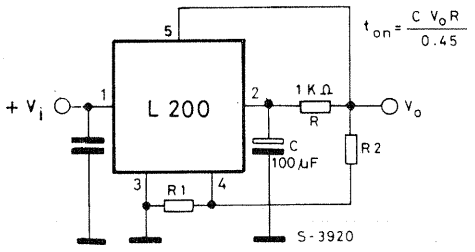
Soft start

When a slow rise time of the output voltage is required, the configuration in fig. 11 can be used. The rise time can be found using the following formula:

$$t_{on} = \frac{CVo\,R}{0.45}$$

At switch on capacitor C is discharged and it keeps the voltage at pin 2 low; or rather, since a voltage of more than 0.45V cannot be generated between pins 5 and 2, the V_o follows the voltage at pin 2 at less than 0.45V.

Fig. 11



Capacitor C is charged by the constant current i_c.

$$i_c = \frac{V_{sc}}{R}$$

Therefore the output reaches its nominal value after the time t_{on}:

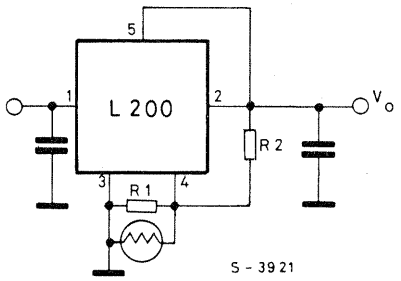
$$V_o - V_{sc} = \frac{i_c \cdot t_{on}}{C}$$

$$t_{on} = C \cdot \frac{(V_o - 0.45)}{0.45} \cdot R \cong \frac{CV_o\,R}{0.45}$$

Light controller

Fig. 12 shows a circuit in which the output voltage is controlled by the brightness of the surrounding environment. Regulation is by means of a photoresistor in parallel with R1. In this case, the output voltage increases as the brightness increases. The opposite effect, i.e. dimming the light as the ambient light increases, can be obtained by connecting the photoresistor in parallel with R2.

Fig. 12



Light dimmer for car display

Although digital displays in cars are often more aesthetically pleasing and frequently more easily read they do have a problem. Under varying ambient light conditions they are either lost in the background or alternatively appear so bright as to distract the driver. With the system proposed here, this problem is overcome by automatically adjusting the display brightness during daylight conditions and by giving the driver control over the brightness during dusk and darkness conditions.

The circuit is shown in fig. 13. The primary supply is shown taken straight from the car battery however it is worth noting that in a car there is always the risk of dump voltages up to 120V and it is recommended that some form of protection is included against this.

Under daylight conditions i.e. with sidelights off and T1 not conducting the output of the device is determined by the values of R1, R2 and the photoresitor (PTR). The output voltage is given by

$$V_{out} = V_{ref}\left(1 + \frac{R2}{PTR//R1}\right)$$

If the ambient light intensity is high, the resistance of the photoresistor will be low and therefore V_{out} will be high. As the light decreases, so V_{out} decreases dimming the display to a suitable level.

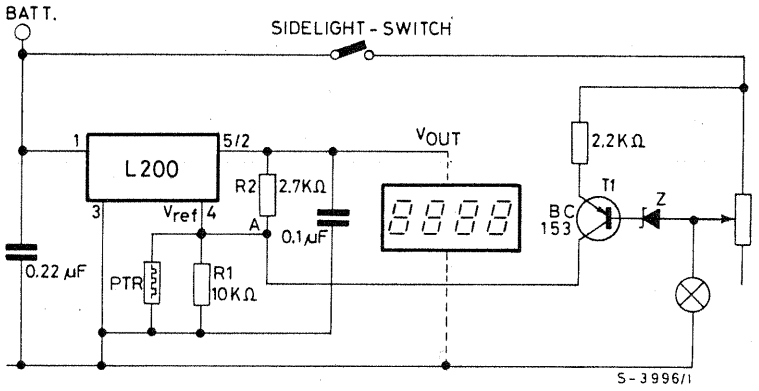


Fig. 15

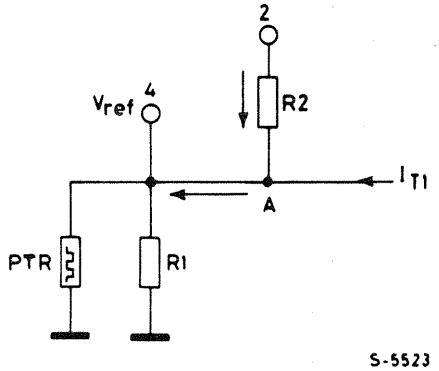
In dusk conditions, when the sidelights are switched on, T1 starts to conduct with its conduction set by the potentiometer. With the potentiometer wiper at its uppermost position the sidelights are at their brightest and current through T1 would be a minimum. With the wiper at its lowest position obviously the opposite conditions apply.

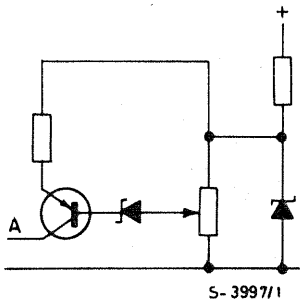
The current through T1 is felt at the summing node A along with the currents through R2 and the parallel network R1, PTR. Since V_{ref} is constant the current flowing through R1, PTR must also be constant. Therefore any change in the current through T1 causes an equal and opposite change in the current through R2. Therefore as I_{T1} increases, V_{out} decreases i.e. as the brightness of the sidelights is increased or decreased so is the brightness of the display.

The values of R2 and PTR should be selected to give the desired minimum and maximum brightness levels desired under both automatic and manual conditions although the minimum brightness under manual conditions can also be set by the maximum current flowing through T1 and, in any case, this should not exceed the maximum current through R2 under automatic operation.

The circuit shown with a small modification can also be used for dimmers other than in a car. Fig. 15 shows the modification needed. The zener diode should have a $V_F \geqslant 2.5V$ at $I = 10\ \mu A$.

Fig. 14





Higher input or output voltages

Certain applications may require higher input or output voltages than the device can produce. The problem can be solved by bringing the regulator back into the normal operating units with the help of external components.

When there are high input voltages, the excess voltage must be absorbed with a transistor. Figs. 16 and 17 show the two circuits:

Fig. 16

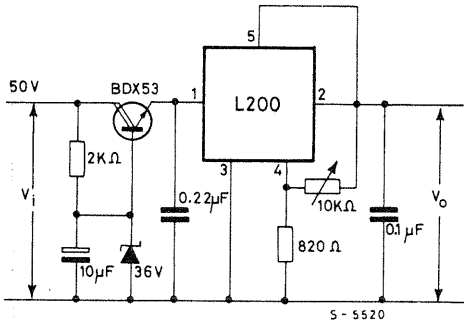
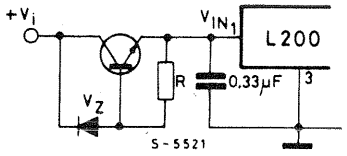


Fig. 17



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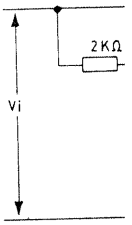


Fig. 19

Fig. 20

The designer must take into account the dissipated power and the SOA of the preregulation transistor. For example, using the BDX53, the maximum input voltage can reach 56V (fig. 16). In these conditions we have 20V of V_{CE} on the transistor and with a load current of 2A the operation point remains inside the SOA. The preregulation used in fig. 16 reduces the ripple at the input of the device, making it possible to obtain an output voltage with negligible ripple.

If high output voltages are also required, a second zener, V_Z, is used to refer the ground pin of an IC to a potential other than zero; diode D1 provides output shortcircuit protection (fig. 18).

Fig. 18

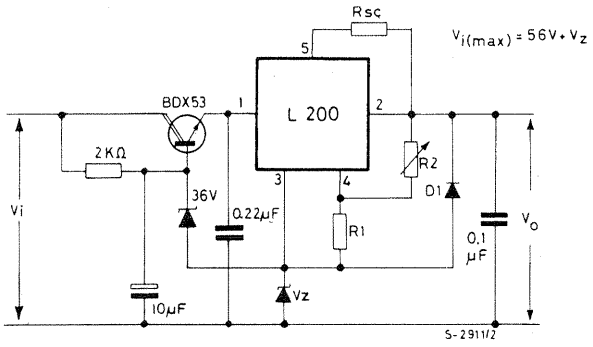
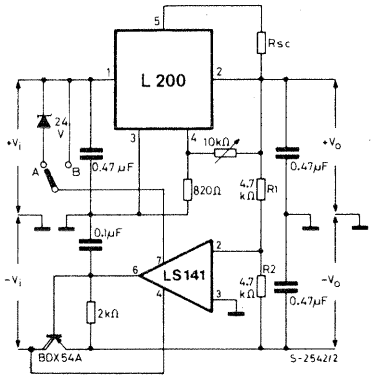


Fig. 19



A: $V_{i(max)} \leq \pm 34V$ $\quad 3 < V_o < 30$

B: $V_{i(max)} \leq \pm 22V$ $\quad 3 < V_o < 18$

Positive and negative voltage regulators

The circuit in fig. 19 provides positive and negative balanced, stabilized voltages simultaneously. The L200 regulator supplies the positive voltage while the negative is obtained using an operational amplifier connected as follower with output current booster.

Tracking of the positive voltage is achieved by putting the non-inverting input to ground and using the inverting input to measure the feedback voltage coming from divider R1-R2.

The system is balanced when the inputs of the operational amplifier are at the same voltage, or, since one input is at fixed ground potential, when the voltage of the intermediate point of the divider goes to 0 Volts. This is only possible if the negative voltage, on command of the op-amp, goes to a value which will make a current equal to that in R1 flows in R2. The ratio which expresses the negative output voltage is:

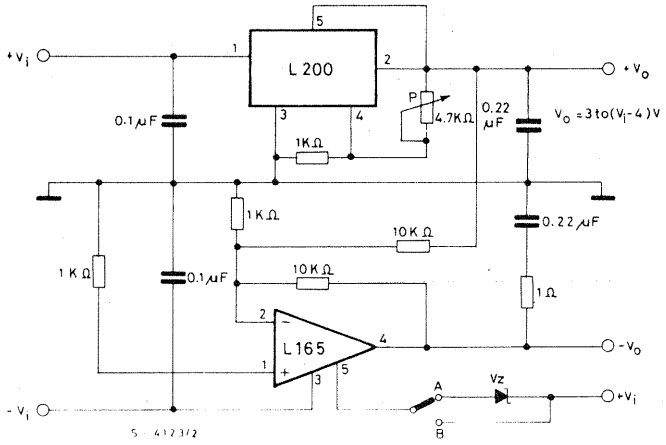
$$V^- = V^+ \cdot \frac{R2}{R1} \qquad \text{(If R2= R1, we'll get } V^- = V^+\text{)}$$

Since the maximum supply voltage of the op amp used is ± 22V, when pin 7 is connected to point B output voltages up to about 18V can be obtained. If on the other hand pin 7 is connected to point A, much higher output voltages, up to about 30V, be obtained since in this case the input voltage can rise to 34V.

Fig. 20 shows a diagram is which the L165 power op amp is used to produce the negative voltage. In this case (as in fig. 19) the output voltage is limited by the absolute maximum rating of the supply voltage of the L165 which is ± 18V. Therefore to get a higher V_{out} we must use a zener to keep the device supply within the safety limits.

If we have a transformer with two separate secondaries, the diagram of fig. 21 can be used to obtain independent positive and negative voltages. The two output diodes, D1 and D2, protect the devices from shortcircuits between the positive and negative outputs.

Fig. 20



A: for $\pm 18V \leq V_i \leq 32V$

Note: V_Z must be chosen in order to verify $2\,V_i - V_Z = 36V$

B: for $V_i \leq \pm 18V$

Fig. 21

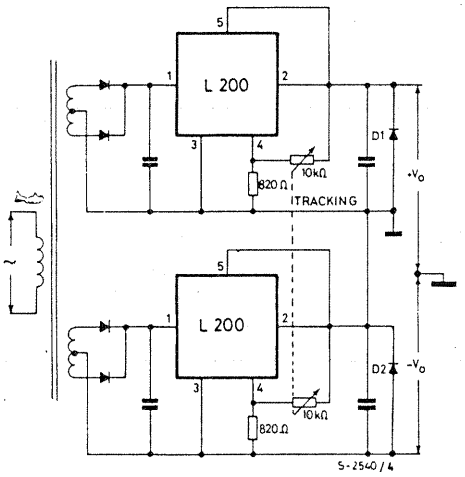
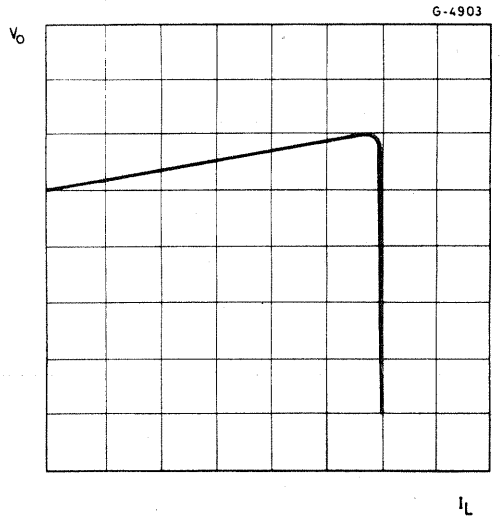


Fig. 23



Compensation of voltage drop along the wires

The diagram in fig. 22 is particularly suitable when a load situated far from the output of the regulator has to be supplied and when we want to avoid the use of two sensing wires. In fact, it is possible to compensate the voltage drop on the line caused by the load current (see the two curves in fig. 23 and 24). R_K transforms the load current I_L into a proportional voltage in series to the reference of the L200. R_K I_L is then amplified by the factor

$$\frac{R2 + R1}{R1}.$$

With the values of R_Z, R2 and R1 known, we get:

$$R_K = R_Z \frac{R1}{R1 + R2}$$

R_Z, R1 and R2 are assumed to be constant.
If R_K is higher than 10 Ω, the output voltage should be calculated as follows:

$$V_o = I_d R_K + V_{ref} \frac{R2 + R1}{R1}$$

Fig. 24

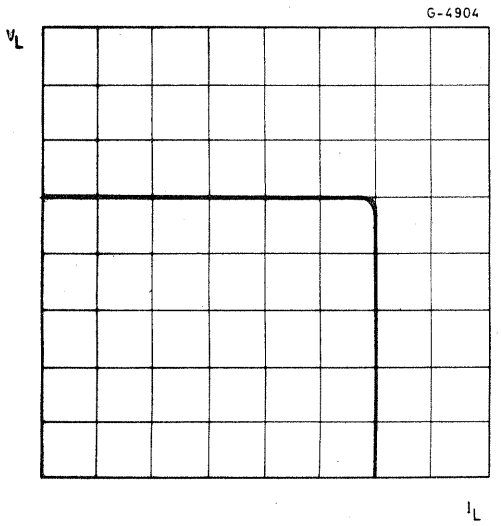
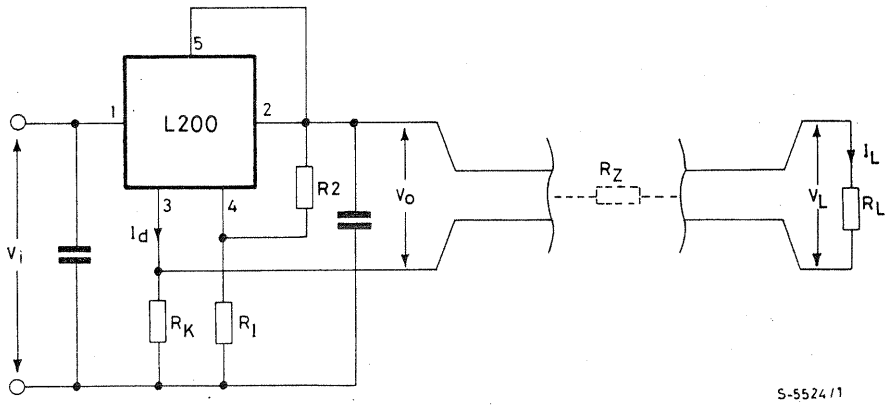


Fig. 22



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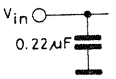

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Motor speed control

Fig. 25 shows how to use the device for the speed control of permanent magnet motors. The desired speed, proportional to the voltage at the terminal of the motor, is obtained by means of R1 and R2.

$$V_M = V_{ref}\left(1 + \frac{R2}{R1}\right)$$

To obtain better compensation of the internal motor resistance, which is essential for good regulation, the following equation is used:

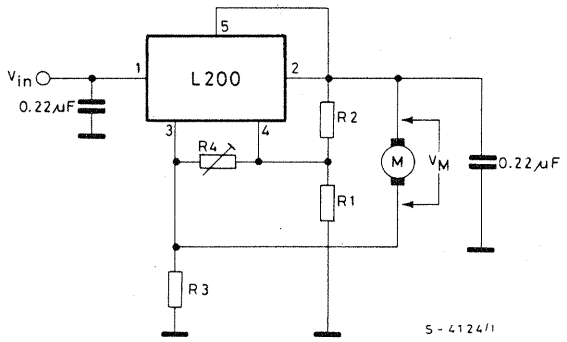
$$R3 \leqslant \frac{R1}{R2} \cdot R_M$$

This equation works with infinite R4. If R4 is finite, the motor speed can be increased without altering the ratio R2/R1 and R3. Since R4 has a constant voltage (V_{ref}) at its terminals, which does not vary as R4 varies, this voltage acts on R2 as a constant current source variable with R4. The voltage drop on R2 thus increases, and the increase is felt by the voltage at the terminals of the motor. The voltage increase at the motor terminals is:

$$V_M = \frac{V_{ref}}{R4 + R3} \cdot R2.$$

A circuit for a 30W motor with $R_M = 4\Omega$, $R1 \doteq 1\,k\Omega$, R2 = 4.3 kΩ, R4 = 22 kΩ and R3 = 0.82Ω has been realized.

Fig. 25



Power amplitude modulator

In the configuration of fig. 26 the L200 is used to send a signal onto a supply line. Since the input signal V_i is DC decoupled, the V_o is defined by:

$$V_o = V_{ref}\left(1 + \frac{R2}{R1}\right)$$

The amplified signal V_i whose value is:

$$G_v = -\frac{R2}{R3}$$

is added to this component. By ignoring the current entering pin 4, we must impose $i_1 = i_2 + i_3$ (1) and since the voltage between pin 4 and ground remains fixed (V_{ref}) as long as the device is not in saturation, $i_1 = 0$ and equation (1) becomes:

$i_2 = -i_3$ with $i_3 = \frac{v_i}{R3}$ (for $X_c \ll R3$) -Therefore:

$v_o = R2\ i_2 = -\frac{v_i}{R3} \cdot R2.$

An application is shown in fig. 27. If the DC level is to be varied but not the AC gain, R1 should be replaced by a potentiometer.

Fig. 26

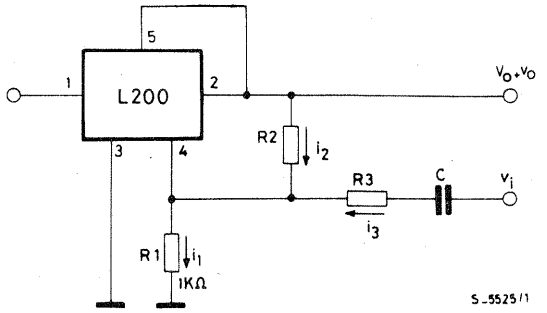
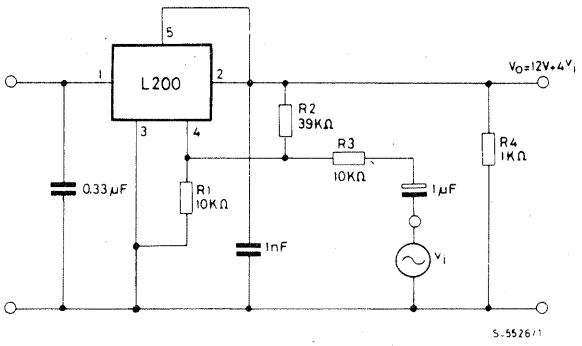


Fig. 27



HIGH CURRENT REGULATORS

To get a higher current than can be supplied by a single device one or more external power transistors must be introduced. The problem is then to extend all the device's protection circuits (short-circuit protection, limitation of T_j of external power devices and overload protection) to the external transistors. Constant current or foldback current limitation therefore becomes necessary.

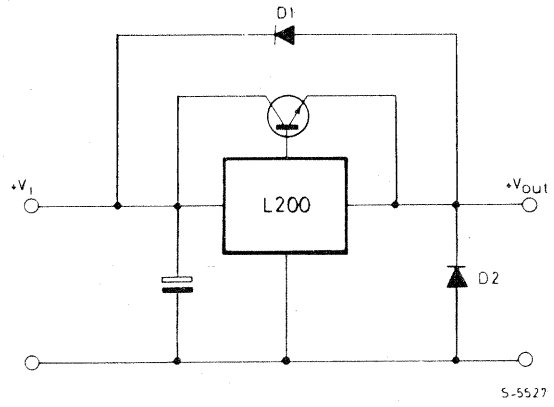
When the regulator is expected to withstand a permanent shortcircuit, constant current limitation becomes more and more difficult to guarantee as the nominal V_o increases. This is because of the increase in V_{CE} at the terminals of the transistor, which leads to an increase in the dissipated power. The heatsink has to be calculated in the heaviest working conditions, and therefore in shortcircuit. This increases weight, volume and cost of the heatsink and increase of the ambient temperature (because of high power dissipation). Besides heatsink, power transistors must be dimensioned for the short-circuit.

This type, of limitation is suited, for example, with highly capacitive loads. Efficiency is increased if preregulation is used on the input voltage to maintain a constant drop-out on the power element for all V_{out}, even in shortcircuit. Foldback limitation, on the other hand, allows lighter short-circuit operating conditions than the previous case. The type of load is important.

If the load is highly capacitive, it is not possible to have a high ratio between I_{max} and I_{sc} because at switch-on, with load inserted, the output may not reach its nominal value.

Other protection against input shortcircuit, mains failure, overvoltages and output reverse bias can be realized using two diodes, D1 and D2, inserted as indicated in fig. 28.

Fig. 28



Use of a PNP power transistor

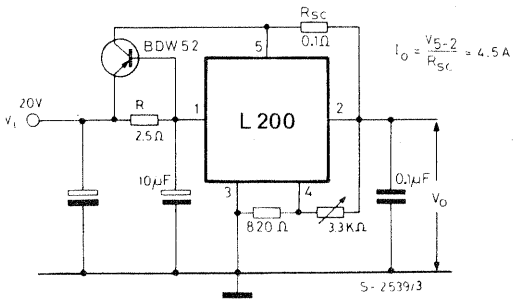
Fig. 29 shows the diagram of a high current supply using the current limitation of the L200. The output current is calculated using the following formula:

$$I_o = \frac{V_{SC}}{R_{SC}} \cong \frac{0.45V}{0.1\Omega} = 4.5A$$

Constant current limitation is used; so, in output shortcircuit conditions, the transistor dissipates a power equal to:

$$PD = V_i \cdot I_o = V_i \cdot \frac{V_{SC}}{R_{SC}}$$

Fig. 29



The operating point of the transistor should be kept well within the SOA; with R_{SC}= 0.1Ω, V_i must not exceed 20V. Part of the I_o crosses the transistor and part crosses the regulator.

The latter is given by: $I_{REG} = I_B + \frac{V_{BE}}{R}$.

where I_B is the base current of the transistor (-100 mA at I_C= 4A) and V_{BE} is the base-emitter voltage (-1V at I_C = 4A); with R = 2.5Ω, $I_{REG} \cong$ ≅ 500 mA.

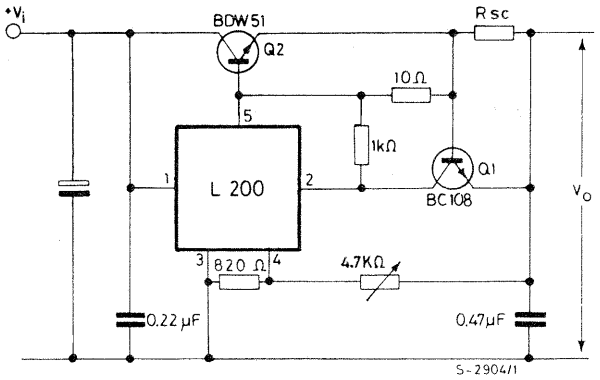
Use of an NPN power transistor

Fig. 30 shows the same application as described in figure 29, using an NPN power transistor instead of a PNP. In this case an external signal transistor must be used to limit the current. Therefore:

$$I_o = \frac{V_{BE\ Q1}}{R_{SC}}$$

As regards the output shortcircuit, see par. 1.5.

Fig. 30



12V 4A Power supply

The diagram in fig. 31 shows a supply using the L200 and the BD705. The 1 kΩ potentiometer, PT1, together with the 3.3 kΩ resistor are used for fine regulation of the output voltage.

Current limitation is of the type shown in fig. 32. Trimmer PT2 acts on strech AB of characteristic. With the values indicated (PT2= 1 kΩ, PT3= 470Ω, R = 3 kΩ), currents from 3 to 4A can be limited. The field of variation can be increased by increasing the value of R_{SC} or by connecting one terminal of PT3 to the base of the power transistor, which, however, provides less stable limitation. If section AB is moved, section BC will also be moved.

The slope of BC can be varied using PT3. The voltage level at point B is fixed by the voltage of the zener diode. The capacitor in parallel to the zener ensures correct switch-on with full load. The BD705 should always be used well within its safe operating area. If this is not possible two or more BD705s should be used, connected in parallel (fig. 33).

Further protection for the external power transistor can be provided as shown in fig. 34. The PTC resistor, whose temperature intervention point must prevent the T_j of the power transistor from reaching its maximum value, should be fixed to the dissipator near the power transistor. Dimensioning of R_A and R_B depends on the PTC used.

Fig. 31

Fig. 32

Fig. 33

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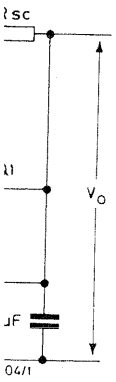

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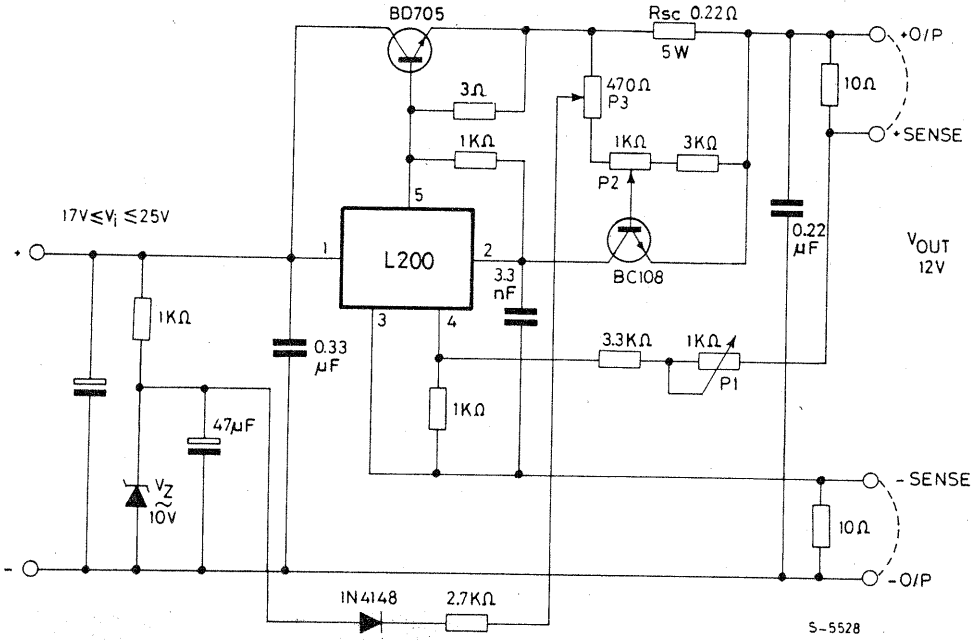


Fig. 32

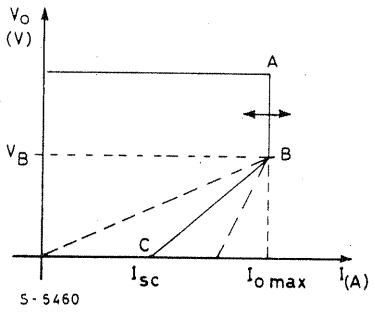


Fig. 34

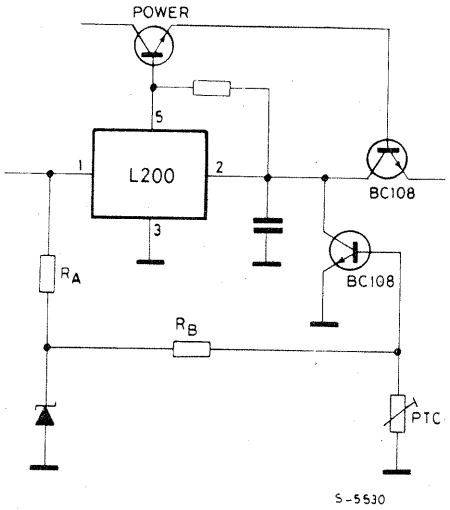
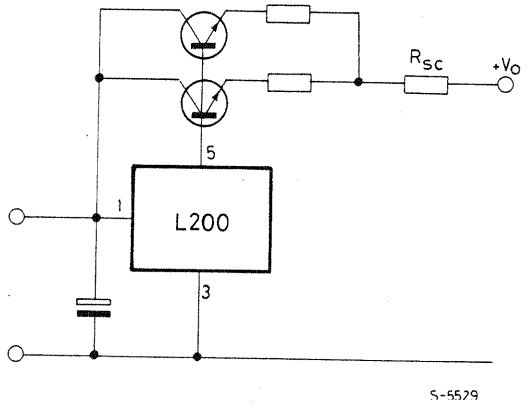


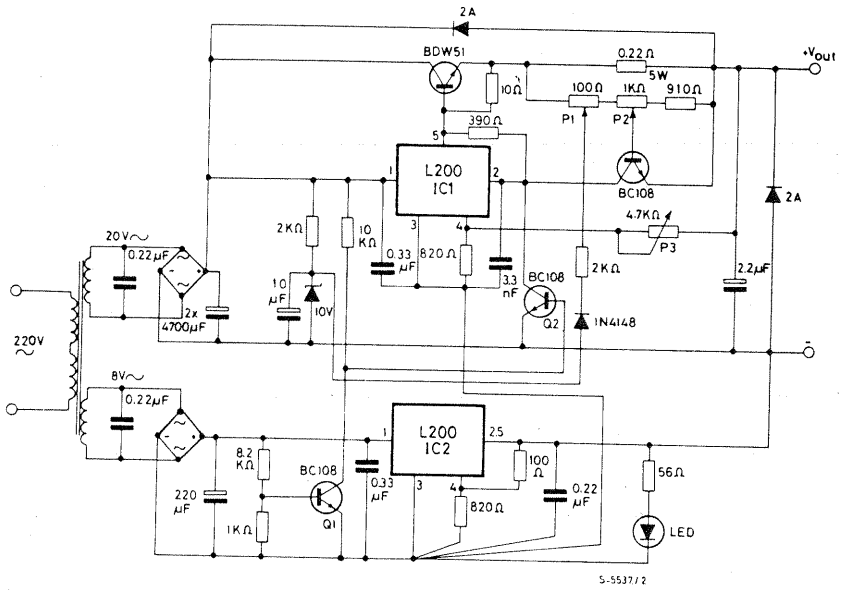
Fig. 33



Voltage regulator from 0V to 16V - 4.5A

Fig. 35 shows an application for a high current supply with output voltage adjustable from 0V to 16V, realized with two L200 regulators and an external power transistor. With the values indicated, the current can be regulated from 2A to 4.5A by potentiometer PT2. PT1, on the other hand, is used for constant current or foldback current limitation. The integrated circuit IC2, which does not require a heatsink and has excellent temperature stability, is used to obtain the 0V output. It is connected so as to lower pin 3 of IC1 until pin 4 reaches 0V. Q1 and Q2 ensure correct operation of the supply at switch-on and switch-off.

Fig. 35



Power supply with V_o= 2.8 to 18V, I_o= 0 to 2.5A

The diagram in fig. 36 shows a supply with output voltage variable from 2.8V to 18V and constant current limitation from 0A to 2.5A. The output current can be regulated over a wide range by means of the op. amp. and signal transistor TR_2. The op. amp. and the transistor are connected in the voltage-current converter configuration. The voltage is taken at the terminals of R3 and converted into current by PT_2.

I_o is fixed as follows:

$$\frac{R4\ I_o}{PT_2} = I_1\ (*) \qquad (**)\ I_{sc} = \frac{V_{SC}}{R2}$$

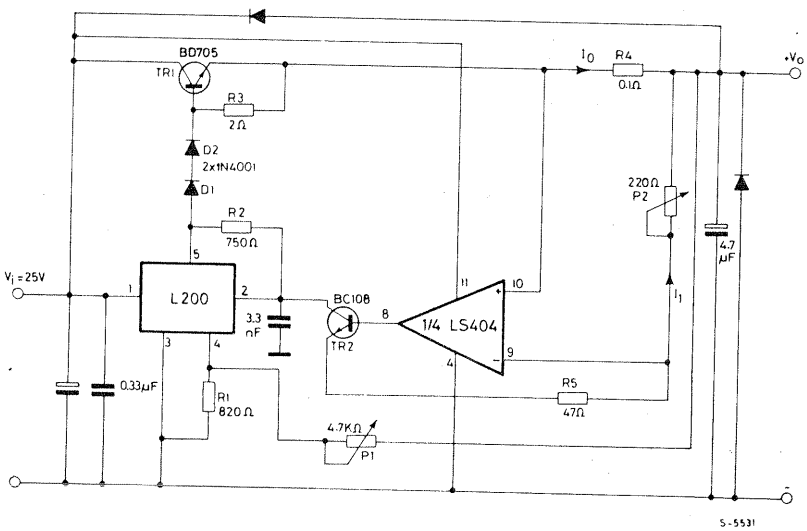
When $I_1 = I_{sc}$, the regulator starts to operate as a current generator. By making (*) equal to (**) we get:

$$\frac{R4\ I_o}{PT_2} = \frac{V_{SC}}{R2}\ ;\ \text{therefore}\ I_o = \frac{V_{SC}}{R2 \cdot R4} \cdot PT_2.$$

Diodes D1 and D2 keep transistor TR_2 in linear condition in the case of small output currents. If it is not necessary to limit the current to zero, one of the diodes can be eliminated: the second diode could also be eliminated if TR_1 were a darlington instead of a transistor.

The op. amp. must have inputs compatible with ground in order to guarantee current limitation even in shortcircuit. With a negative voltage available, even of only a few volts, current limitation is simplified.

Fig. 36



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LAYOUT CONSIDERATIONS

The performance of a regulator depends to a great extent on the case with which the printed circuit is produced. There must be no impulsive currents (like the one in the electrolytic filter capacitor at the input of the regulator) between the ground pin of the device (pin 3) and the negative output terminal because these would increase the output ripple. Care must also be taken when inserting the resistor connected between pin 4 and pin 3 of the device.

The track connecting pin 3 to a terminal of this resistor should be very short and must not be crossed by the load current (which, since it is generally variable, would give rise to a voltage drop on this stretch of track, altering the value of V_{ref} and threfore of V_o.

When the load is not in the immediate proximity of the regulator output "+ sense" and "— sense" terminals should be used (see fig. 37). By connecting the "+ sense" and "— sense" terminals directly at the charge terminals the voltage drop on the connection cable between supply and load are compensated. Fig. 37 shows how to connect supply and load using the sensing clamps terminals.

Fig. 37

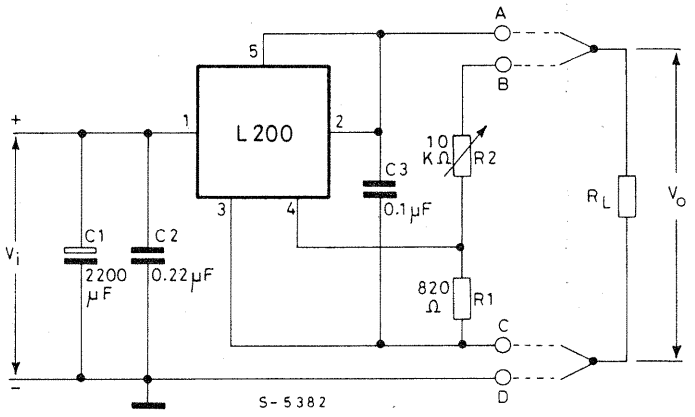
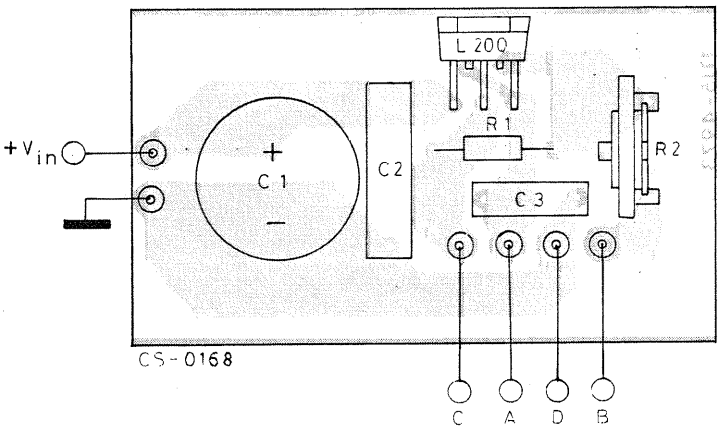


Fig. 38



HEATSINK DIMENSIONING

The heatsink dissipates the heat produced by the device to prevent the internal temperature from reaching values which could be dangerous for device operation and reliability.

Integrated circuits in plastic package must never exceed 150°C even in the worst conditions. This limit has been set because the encapsulating resin has problems of vitrification if subjected to temperatures of more than 150°C for long periods or of more than 170°C for short periods (24 h). In any case the temperature accelerates the ageing process and therefore influences the device life; an increase of 10°C can halve the device life. A well designed heatsink should keep the junction temperature between 90°C and 110°C. Fig. 39 shows the structure of a power device. As demonstrated in thermodynamics, a thermal circuit can be considered to be an electrical circuit where R1, 2 represent the thermal resistance of the single elements (expressed in °C/W);

Fig. 39

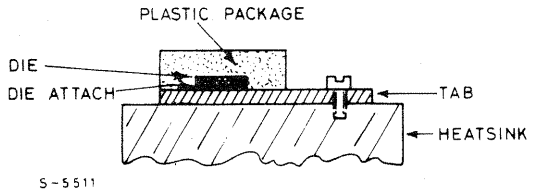
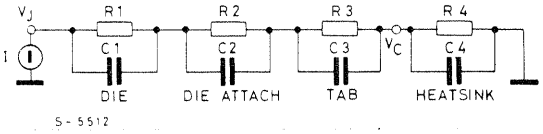


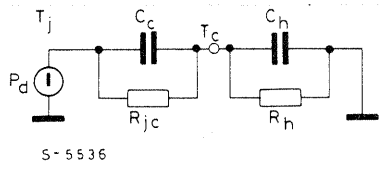
Fig. 40



C1, 2 the thermal capacitance (expressed in °C/W)
I the dissipated power
V the temperature difference with respect to the reference (ground).

This circuit can be simplified as follows:

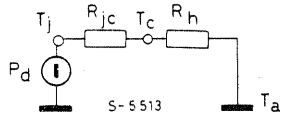
Fig. 41



Where C_e is the thermal capacitance of the die plus that of the tab.
C_h is the thermal capacitance of the heatsink
R_{jc} is the junction case thermal resistance
R_h is the heatsink thermal resistance.

But since the aim of this section is not that of studing the transistors, the circuit can be further reduced.

Fig. 42



If we now consider the ground potential as ambient temperature, we have:

$$T_j = T_a + (R_{jc} + R_h)\, P_D \qquad (1)$$

$$R_h = \frac{T_j - T_a - R_{jc} \cdot P_d}{P_d} \qquad (1a)$$

$$T_c = T_a + R_h \cdot P_d \qquad (2)$$

For example, consider an application of the L200 with the following characteristics:

$V_{in\ typ} = 20V$
$V_o = 14V$
$I_{o\ typ} = 1A$
$T_a = 40°C$
} typical conditions

$V_{in\ max} = 22V$
$V_o = 14V$
$I_{o\ max} = 1.2A$
$T_a = 60°C$
} overload conditions

$P_{d\ typ} = (V_{in} - V_o) \cdot I_o = (20-14) \cdot 1 = 6W$
$P_{d\ max} = (22-14) \cdot 1.2 = 9.6W$

Imposing $T_j = 90°C$ of (1a) we get (from L200 characteristics we get $R_{j-c} = 3°C/W$)

$$R_h = \frac{90 - 40 - 3 \cdot 6}{6} = 5.3°C/W$$

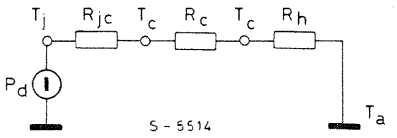
Using the value thus obtained in (1), we get that the junction temperature during the overload goes to the following value:

$$T_j = 60 + (3 + 5.3) \cdot 9.6 = 140°C$$

If the overload occurs only rarely and for short periods, dimensioning can be considered to be correct. Obviously during the shortcircuit, the dissipated power reaches must higher values (about 40W for the case considered) but in this case the thermal protection intervenes to maintain the temperature below the maximum values allowed.

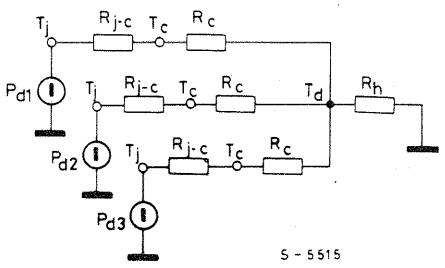
Note 1: If insulating materials are used between device and heatsink, the thermal contact resistance must be taken into account (0.5 to 1°C/W, depending on the type of insulant used) and the circuit in fig. 43 becomes:

Fig. 43



Note 2: In applications where one or more external transistors are used together with the L200, the dissipated power must be calculated for each component. The various junction temperatures can be calculated by solving the following circuit:

Fig. 44



This applies if the various dissipating elements are fairly near to one another with respect to the heatsink dimensions, otherwise the heatsink can no longer be considered as a concentrated constant and the calculation becomes difficult.

This concept is better explained by the graph in fig. 45 which shows the case (and therefore junction) temperature variation as a function of the distance between two dissipating elements with the same type of dissipator and the same dissipated power. The graph in fig. 45 refers to the specific case of two elements dissipating the same power, fixed on a rectangular aluminium plate with a ratio

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Fig. 45

Fig. 46

of 3 between the two sides. The temperature jump will depend on the dissipated power and one the device geometry but we want to show that there exists an optimal position between the two devices:

$$d = \frac{1}{2} \cdot \text{side of the plate}$$

Fig. 46 shows the trend of the temperature as a function of the distance between two dissipating elements whose dissipated power is fairly different (ratio 1 to 4).

This graph may be useful in applications with the L200 + external transistor (in which the transistor generally dissipates more than the L200) where the temperature of the L200 has to be kept as low as possible and especially where the thermal protection of the L200 is to be used to limit the transistor temperature in the case of an overload or abnormal increase in the ambient temperature. In other words the distance between the two elements can be selected so that the power transistor reaches the $T_{j\ max}$ (200°C for a TO-3 transistor) when the L200 reaches the thermal protection intervention temperature.

Fig. 45

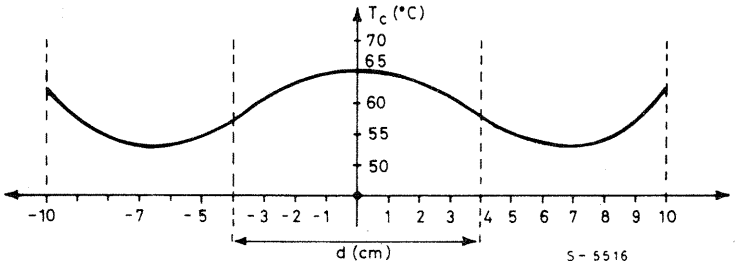
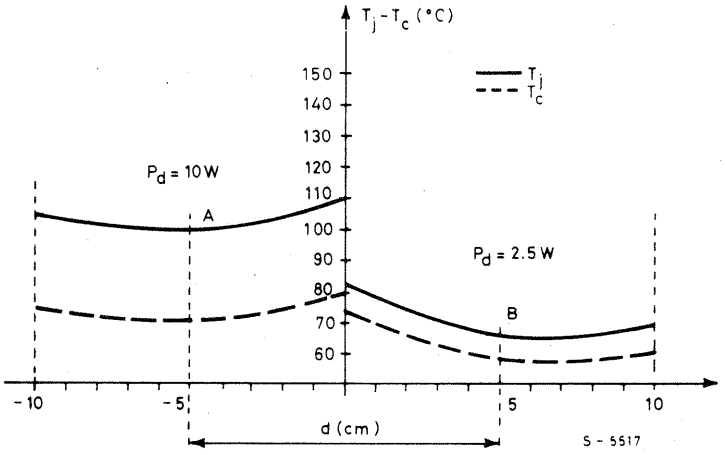


Fig. 46



A : Position of the device with high power dissipation (10W)
B : Position of the device with low power dissipation (2.5W)