## Set 24 : Voltage regulators

This set covers integrated-circuit regulators with and without foldback on the one hand, and discrete-component series and shunt types on the other, but with some special cases in between. The exceptions are two dual-polarity circuits, a switching regulator and one using a current differencing or Norton amplifier.
On card 7 the field-effect transistor was wrongly drawn; page 56 shows the correct wiring, and on card 1 the last equation should have shown a product and not a difference.
Incidentally, on this set and others, unnumbered circuits on the bottom half of the page generally refer to items under circuit modifications and are in the same sequence as the text, unless made otherwise obvious.

# Voltage regulators

The regulator is divided into the reference section and a d.c. power amplifier. These both require supply voltages; the convenience of having a single supply may outweigh the improved stability that can be obtained. The output current can cause the amplifier supply to vary: the source impedance including increased ripple if it is a rectified a.c. supply. Because the current required by the reference circuit is low and constant, it is easier to avoid any serious ripple/regulation effects. It is essential that the d.c. amplifier have (a) an accurately defined voltage gain, (b) a low output resistance, (c) a sufficiently high output current/voltage capability, (d) a temperature drift that is either low or of the appropriate sign and magnitude to compensate for any drift in the reference section.

A simple configuration that meets these requirements in principle is shown in Fig. 1 . The amplifier can be a standard operational amplifier if the output current is not much in excess of 10mA, and single-ended supply operation is permissible in many cases. The method is extended in Fig. 2 to the provision of output voltages that differ from the reference voltage. The output voltage is of opposite polarity to the reference voltage requiring a separate negative supply. The op-amp can be replaced by any circuit meeting conditions (a) to (d) above. Before turning to detailed study of possible configurations it is important to consider an alternative viewpoint.

A discrete component circuit that has all elements of a practical regulator is given in card 3. Three transistors comprise a voltage amplifier of gain $(R_4/R_5+1)$ with high input- and low output-impedance (alternatively $Tr_1$, $Tr_2$ are the error amplifier and $Tr_3$ the series-pass transistor).

A serious problem arises in all regulators with emitter-follower outputs. The minimum input-output differential includes the $V_{be}$ of $Tr_3$ plus the voltage-drop across $R_2$. This figure is markedly increased when $Tr_3$ is replaced by compound transistors for greater output current capability. This property has serious implications for the maximum efficiency of which the circuit is capable, and also for the maximum dissipation the output stage may be called on to tolerate.

A possible solution is to replace $Tr_3$ by a common-emitter p-n-p transistor, driving its base from the collector of $Tr_1$ to restore the feedback condition ($Tr_3$ would then be providing an additional inversion). A simplified form of the circuit for which this is not possible is shown in Fig. 3.) The effective reference voltage in this circuit is $(V_z+V_{be})$ and for best temperature stability, $V_z$ would be chosen to have a drift of $+2\text{mV K}^{-1}$ to cancel the negative $V_{be}$ drift. This circuit is the basis of a large number of commercial regulators, though the functional similarity may be hard to recognize amongst the welter of extra functions such as current limiting variable output voltage etc.

Although the basic form can be designed for output currents of 100mA+, any further increase forces the base current of $Tr_2$ too high – normally $Tr_1$ collector current has to be at least as great as the base current of $Tr_2$. It may not be convenient for the zener current to exceed about 10mA since the regulation is impaired. To keep the zener/error amplifier current low, it is sufficient to replace the output stage by a pair of transistors connected to give increased current gain.

A major problem in the design of voltage regulators is to protect against load resistances falling below specified levels, and the size and cost of transistors, heatsinks and power supplies is dictated by the occasional fault condition rather than by the ratings into any intended value of load. For this reason the technique described as foldback or re-entrant current limiting was devised. The resulting characteristic is shown in Fig. 4(b), with the current falling back to a short-circuit value close to zero as the short-circuit condition is approached. The technique involves a current-limit reference voltage which depends on the output voltage. As soon as the current-limit circuit is activated the output voltage begins to fall simultaneously reducing the current as shown.

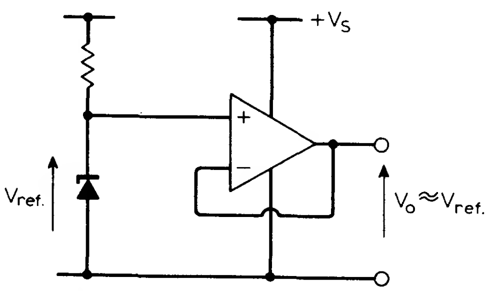


Fig. 1

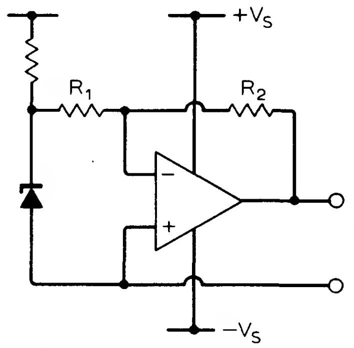


Fig.2

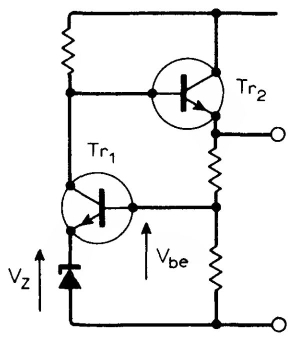


Fig.3

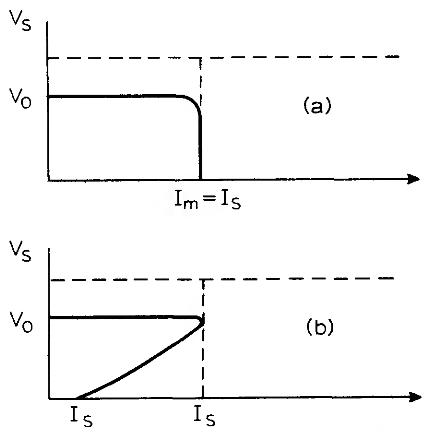
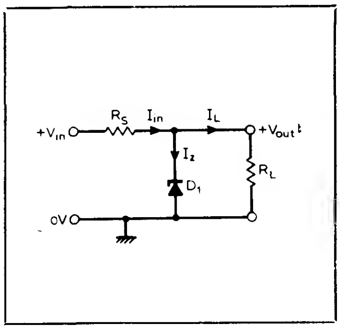


Fig.4

# Zener diode shunt regulator




**Typical performance**
*$I_L$ constant, $V_{IN}$ variable*
$D_1$ BZY96C9V, $I_L$ 50mA
$V_{IN(min)}$ 12V, $V_{IN(max)}$ 18V
$R_S$ 54Ω (39Ω+15Ω)
$R_L$ 195Ω (3×56Ω+27Ω) to set $I_L$ at 50mA

**Measured results**

| $V_{IN}$(V) | $V_{OUT}$(V) | $I_L$(mA) | $I_Z$(mA) | $P_Z$(W) |
|---|---|---|---|---|
| 12 | 9.16 | 48 | 1 | 0.009 |
| 15 | 9.49 | 49 | 48 | 0.46 |
| 18 | 9.72 | 50.5 | 106.5 | 1.03 |

**Typical performance**
*$V_{IN}$ constant, $I_L$ variable*
$D_1$ BZY96C9V, $V_{IN}$ 15V
$I_{L(max)}$ 100mA, $I_{L(min)}$ 80mA
$R_S$ 39Ω+15Ω
$R_L$ 95Ω to set $I_L$ at 100mA
$R_L$ 124Ω to set $I_L$ at 80mA

**Measured results**

| $I_L$(mA) | $V_{OUT}$(V) | $I_Z$(mA) | $P_Z$(mW) |
|---|---|---|---|
| 102 | 9.27 | 7 | 64.9 |
| 80 | 9.56 | 26 | 248.6 |

**Description**
A voltage regulator should ideally provide a form of buffer action which makes its output voltage independent of changes that occur in its input voltage or its load current. The extent to which a particular regulator circuit approaches this ideal usually depends on the complexity of the electronic regulation element used. The simplest form of electronic shunt regulation element is the zener diode which may be considered to consist of an internal reference voltage source ($V_Z$) in series with an internal resistance ($r_z$), both of which have values that depend on the operating point and junction temperature.
The basic form of a zener diode shunt regulator is shown above where the input voltage must be larger than the required regulated output voltage. The input voltage will often be derived from the a.c. mains by rectification and will have a value that varies with mains input voltage and with load current, due to imperfect power supply regulation. The current in $R_S$ is the sum of the load current ($I_L$) and the zener diode current ($I_Z$). If $V_{IN}$ increases, the current in the zener diode and the load increases. But at the same time, a shift occurs in the zener diode operating point causing its internal resistance to fall. Thus the combined effects of the increase in $I_Z$ and the decrease in $r_z$ tends to maintain the output voltage at its former value. Similar but opposite effects occur if $V_{IN}$ decreases. Ability of the circuit to maintain the output voltage depends on the zener resistance and on the temperature coefficient of the zener voltage. The output voltage will not, in general, be equal to the nominal zener voltage because $V_Z$ and $r_z$ have values that depend on $I_Z$ and junction temperature.
In some applications the load current may be virtually constant and in others it may vary over a wide range. If the load current decreases, the current shunted by the zener diode will increase, and vice versa, resulting in a substantially constant output voltage. Protection against excessive load current can be obtained with a fuse, but protection of the zener diode under light loading or open-circuit load conditions must be catered for by choosing a diode that can safely dissipate the power generated when $I_L \to 0$, if there is any possibility of the load being removed.
The design of this shunt regulator therefore becomes a matter of determining the value of $R_S$ and the maximum power dissipated in the zener diode under specified conditions of variable $V_{IN}$ and/or variable $I_L$. Although more precise results can be obtained by measuring and plotting the zener diode characteristics, for all practical purposes the nominal value of $V_Z$ can be used to approximate the value of $V_{OUT}$ in order to determine the component values. The Kirchhof voltage equation for the circuit is

$$V_{IN}=I_{IN}.R_S+V_Z$$
$$\therefore\ R_S=(V_{IN}-V_Z)/(I_Z+I_L),$$
as $I_{IN}=I_Z+I_L$
and $I_Z=(V_{IN}-V_Z)/R_S-I_L$
and the diode dissipation is
$$P_Z=I_Z.V_Z$$
$$=[(V_{IN}-V_Z)/R_S-I_L]\,V_Z$$

Determination of suitable values for $R_S$ and $P_{Z(max)}$ depends on the specification. The value of $R_s$ must be such that the zener current will not fall below some minimum value, $I_{Z(min)}$, required to keep the diode in the breakdown region so that $V_Z$ is maintained. Minimum zener current occurs when $V_{IN}$ is a minimum, $V_Z$ is a maximum and $I_L$ is a maximum, so that

$$R_S=\frac{V_{IN(min)}-V_{Z(max)}}{I_{Z(min)}+I_{L(max)}}$$

Using the nominal zener voltage $V_Z$ and an empirical factor of 10% of $I_{L(max)}$ for $I_{Z(min)}$ gives

$$R_S=\frac{V_{IN(min)}-V_Z}{1.1I_{L(max)}}$$

for the condition where either $V_{IN}$ or both $V_{IN}$ and $I_L$ are variable. When only $I_L$ is variable

$$R_S=\frac{V_{IN}-V_Z}{1.1I_{L(max)}}$$

Having determined $R_S$, $P_{Z(max)}$ can be found from

$$\left[\left(\frac{V_{IN}-V_Z}{R_S}\right)-I_{L(min)}\right]V_Z$$

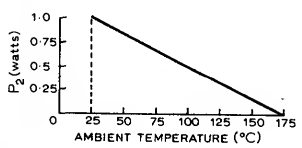
for constant $V_{IN}$ and variable $I_L$.

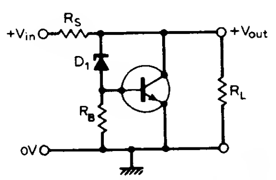
$$\left[\left(\frac{V_{IN(max)}-V_Z}{R_S}\right)-I_L\right]V_Z$$

for constant $I_L$ and variable $V_{IN}$.

$$\left[\left(\frac{V_{IN(max)}-V_Z}{R_S}\right)-I_{L(min)}\right]V_Z$$

for $I_L$ and $V_{IN}$ variable.
A zener diode is then chosen having the desired nominal voltage and capable of safely dissipating this maximum power. It may be necessary to design a heat sink of suitable area and/or to derate the diode's dissipation capability as a function of ambient temperature. Most zener diodes are rated at a temperature of 25°C, a typical derating curve for a 1 watt diode being as shown on this card.
For increased power capability the zener diode can be connected in the base of a power transistor as shown.

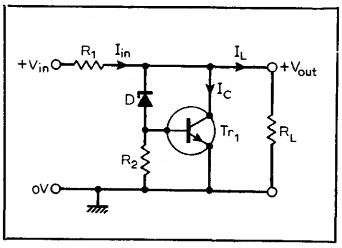



**Further reading**
Zener Diode Handbook, Motorola 1967.
Patchett, G. N. Automatic Voltage Regulators and Stabilizers, chapter 6, Pitman, 1970 (3rd edition).

**Cross references**
Set 23, card 1.
Set 24, cards 3, 4.

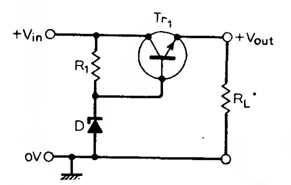
## Simple transistor regulators

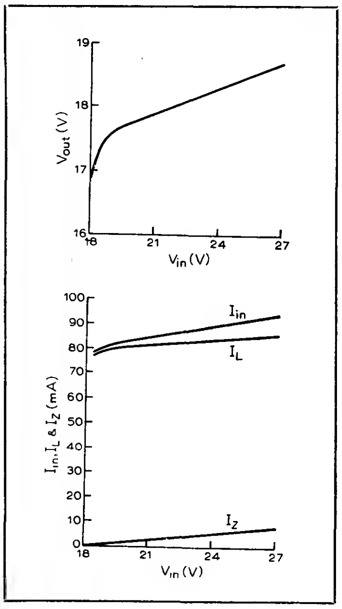



**Typical performance**
$Tr_1$ BFR41, $D_1$ ESM18
$R_1$ 330Ω, 3W; $R_2$ 100Ω
$R_L$ 1kΩ, ½W
$V_{IN}$ 32.5±7.5V
$V_{OUT}$ see graphs opposite

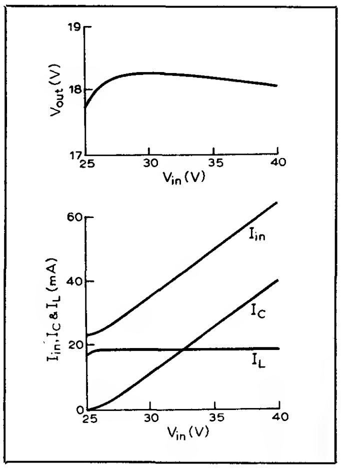
**Description**
Although less efficient than series regulators, the shunt regulator is normally a simpler circuit and is useful where an existing supply is to be used to provide a lower-value regulated output voltage. A simple regulator is shown above which includes a zener diode reference-voltage element and a transistor, in shunt with the load, acting as the regulator element. Note that the circuit is simply a common-emitter d.c. amplifier. The value of $R_2$ is chosen to provide a current in $D_1$ that is greater than the minimum value required to maintain the zener diode in its breakdown region without exceeding the rated dissipation. Output voltage remains essentially constant because the transistor collector current changes as the input voltage and/or the load current changes, causing a corresponding change in the p.d. across $R_1$. Transistor $Tr_1$ must be chosen to accommodate the maximum dissipation that can occur under specified input voltage and load current variations, including open-circuit load if this is a possibility.

**Fig. 3**




**Circuit modifications**
To reduce changes in zener diode current, due to $Tr_1$ base current, cascaded transistors may be used to increase the current gain of the regulating element, as Fig. 1. The base current of $Tr_2$ is then only $\approx I_{B1}/h_{FE2}$ which can be made much smaller than the zener diode current by choice of $R_2$.
An alternative form of simple shunt regulator is Fig. 2 where the zener diode is in series with $Tr_1$ emitter.
A fraction of the output voltage $V_{OUT}R_3/(R_2+R_3)$ is compared with $V_Z$. If $V_{OUT}$ increases the base potential rises causing an increase in collector and emitter currents and hence a larger p.d. across $R_1$ which tends to return $V_{OUT}$ to its previous, lower value. The zener current is now the emitter current of $Tr_1$ which is much larger than the base current, and to make $I_Z$ less dependent on $I_E$ a resistor can be added between




$D_1$ cathode and $V_{OUT}$. Note that $V_{OUT}$ is now adjustable over a wide range, for a given zener voltage by choice of the ratio $R_2$: $R_3$.
The Fig. 3 circuit is a simple transistor series voltage regulator, i.e. the regulating element $Tr_1$ is in series with the load.
Note that the circuit is an emitter follower d.c. amplifier where ideally $V_{OUT}=V_Z$ but in practice $V_{OUT}=(V_Z-V_{BE})$.
If the output voltage tends to decrease due to changes in input voltage or load current, the base-emitter voltage of $Tr_1$ increases causing the transistor to feed a larger current to the load which will tend to restore $V_{OUT}$ to its previous

**Fig. 4**

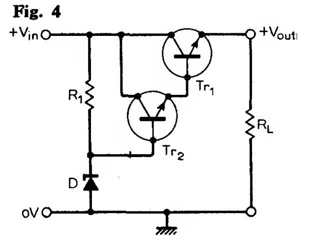



**Fig. 5**

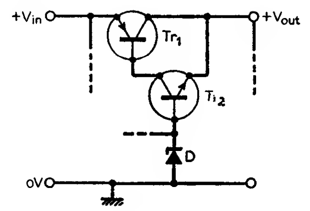



**Fig. 1**

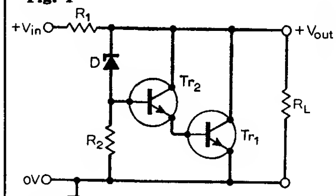
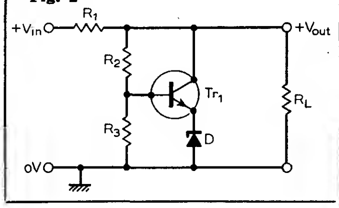



**Fig. 2**



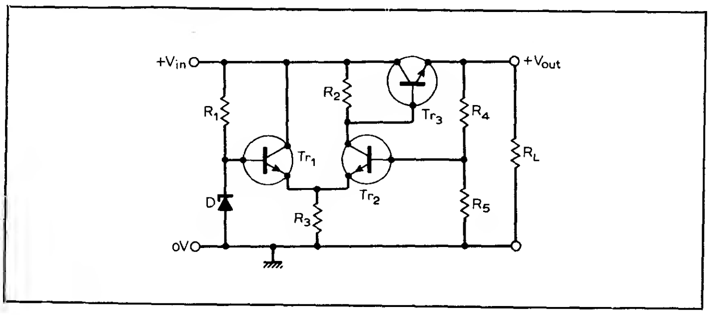

value. The current in the zener diode can be made much larger than the base current of $Tr_1$ by choice of $R_1$ and this current, hence $V_Z$ and $V_{OUT}$, will be subject to variation as $V_{IN}$ changes. The circuit is inherently safe with open-circuit loads but $Tr_1$ must be chosen to dissipate the maximum power generated under $V_{IN(max)}$ and $R_{L(min)}$ conditions.

**Typical performance**
$Tr_1$ BFR41, $D_1$ ESM18
$R_1$ 1kΩ, ¼W; $R_L$ 200Ω, 3W
$V_{IN}$ 22.5±4.5V
$V_{OUT}$ see graphs above
The variation of current in the zener diode with base current can be reduced by replacing $Tr_1$ by a Darlington pair as in Fig. 4, where the base current of $Tr_2$ is then only $I_L/[(1+h_{FE1})(1+h_{FE2})]$.
This principle can be extended to a number of emitter followers or a complementary pair may be used, Fig. 5, to keep $V_{OUT}=(V_Z-V_{BE})$.

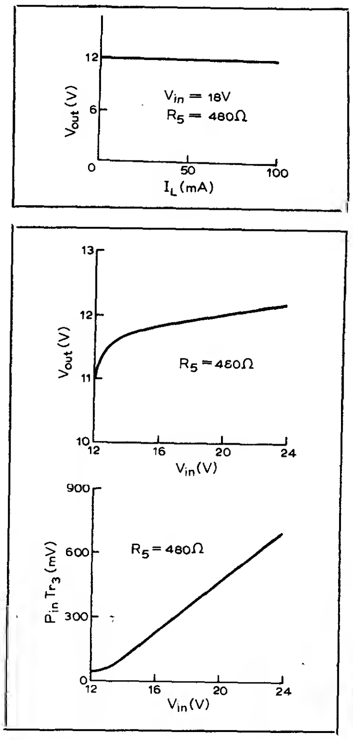
**Further reading**
Patchett, G. N. Automatic voltage regulators and stabilizers, chapter 6, Pitman, 1970 (3rd edition).
Zener Diode Handbook—Motorola, chapter 6, 1967.

**Cross references**
Set 24, card 4.
Set 23, card 1.

## Feedback series regulators



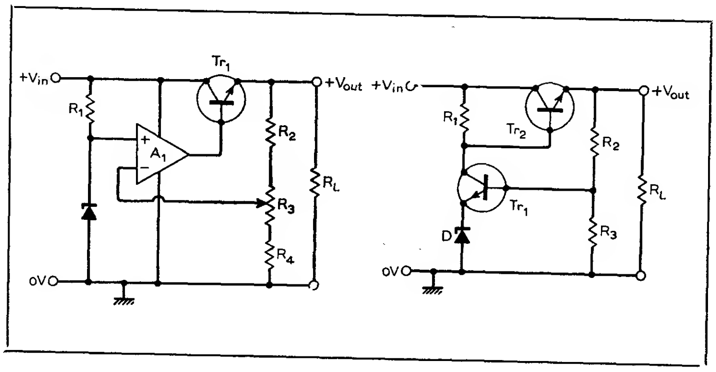

**Typical performance**
$Tr_1$, $Tr_2$ BC125 $Tr_3$ BFR41
$D_1$ BZY88C5V6
$R_1$ 560Ω, $R_2$ 1kΩ
$R_3$ 220Ω, $R_4$ + $R_5$ 1kΩ
$R_L$ 250Ω 1W
$V_{IN}$ 18V ±6V
$V_{OUT}$ see graphs opposite



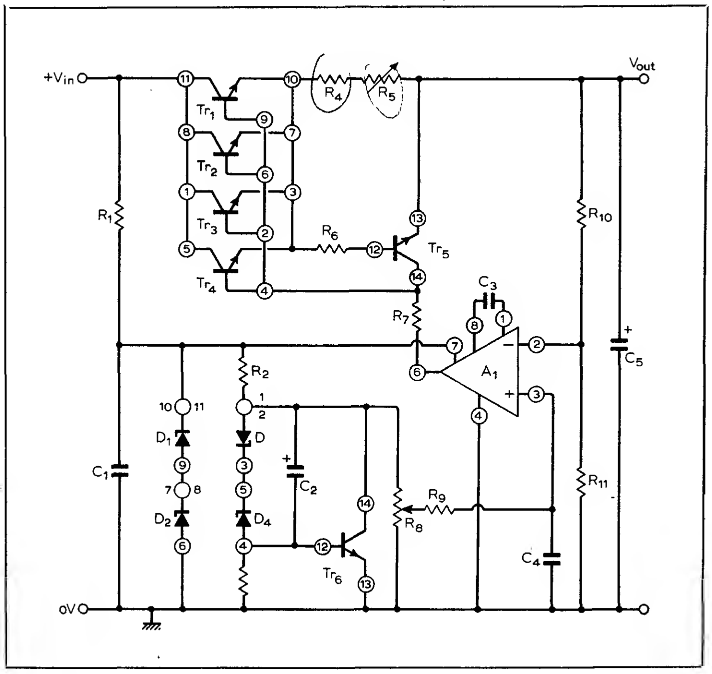

**Long-tailed pair regulator**
A very common type of feedback voltage regulator is shown above where the control amplifier is in the form of a long-tailed pair, or differential-input amplifier containing transistors $Tr_1$ and $Tr_2$. Resistor $R_1$ and $D_1$ act as a simple voltage reference circuit making $Tr_1$ base potential $V_Z$. The base potential of $Tr_2$ is a fraction of the output voltage, determined by the ratio of the resistors in the potential divider $R_4$ and $R_5$, so that the output voltage is continuously monitored. The output from the long-tailed pair is taken from $Tr_2$ collector and controls the base drive to the series transistor $Tr_3$ (an emitter follower). The differential amplifier attempts to keep its two inputs equal by altering the p.d. across the series transistor in order to hold the regulated output voltage constant despite changes that occur in the input voltage or load current. With a load current of 50mA, the circuit shown typically provides a load regulation of about 0.03% and a line regulation of approximately 0.5% for a ±20% change in $V_{IN}$.
For a fixed input voltage the output voltage may be varied conveniently by realizing $R_4$ and $R_5$ in the form of a potentiometer e.g. with $(R_4+R_5)=1k\Omega$ and $R_5$ varied over the range 100 to 900Ω. $V_{OUT}$ may be varied over the range 16.86 to 6.4V.
Frequency stability is important in the design of feedback regulators. Because negative feedback is used in the control amplifier element, a total phase shift around the loop (including the series element) of 180° at high frequencies can result in oscillations unless the closed-loop gain is less than unity. Therefore, at the frequency where the total phase shift is 180°, provision should be made to reduce the closed-loop gain to less than unity. The use of shunt capacitance at the output, or elsewhere in the amplifier section produces the required gain "roll-off" with frequency. The degree of voltage stabilization depends mainly on the stability of the reference voltage, best results being obtained when $Tr_1$ and $Tr_2$ are a matched pair.
The differential-input amplifier is the basis of most operational amplifier designs so the long-tailed pair may be replaced by such an amplifier as shown below. In this arrangement the operational amplifier isolates the zener reference from load changes improving the load regulation. Potentiometer $R_3$ allows the output voltage to vary over a limited range. Typical performance is $A_1$ 741, $Tr_1$ SE3035, D 1N4611, $R_1$ 12kΩ, $R_2$, $R_4$ 1.2kΩ, $R_3$ 2.5kΩ. With $V_{IN}$ of +30V, $V_{OUT}$ may be varied over the range 9 to 25V with load currents up to 100mA. Output impedance is less than 0.1Ω. Useful minimum $V_{IN}$ 20V.
Another type of feedback regulator in common use is the d.c. feedback pair shown below. In this circuit the effective reference voltage is $V_Z + V_{BE}$ of $Tr_1$ and for best results, the temperature stability of the effective reference source should be optimized by choosing a zener diode that provides a temperature drift of approximately +2mV/degC to cancel the negative drift in the $V_{BE}$ of $Tr_1$. In the basic form shown, load currents of up to about 100mA can be accommodated and for higher output currents $Tr_2$ can be replaced by a higher-current-gain transistor pair.

**Further reading**
English, M. Applications for fully compensated op.amp.i.c. *EEE*, January 1969, pp. 63-5.
Potted power, *Design Electronics*, January 1971, pp. 34/5.

**Cross references**
Set 24, cards 1, 4, 5, 6.
Set 23, card 3.
Set 20, cards 1, 2, 4, 10.



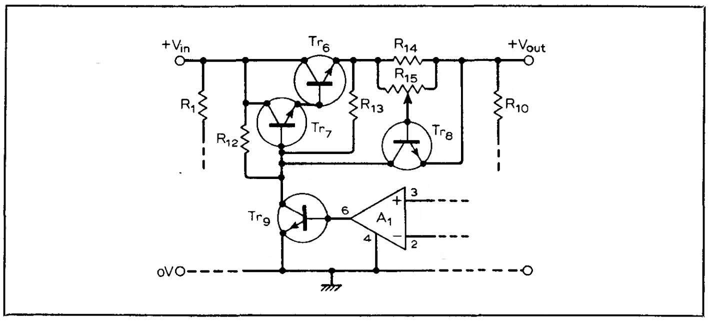
## Bipolar/c.m.o.s. op-amp regulator




**Typical performance**
$A_1$ CA3130
$Tr_1$-$Tr_6$ 1/5×CA3086
$D_1$-$D_4$ 1/5×CA3086
$R_1$ 390Ω, $R_2$ 2.2kΩ, $R_3$ 62kΩ
$R_4$ 3Ω, $R_5$ $R_6$ $R_7$ 1kΩ
$R_8$ 50kΩ, $R_9$ 100kΩ
$R_{10}$ 20kΩ, $R_{11}$ 30kΩ
$C_1$, $C_4$ 10nF, $C_2$ 25μF, 15V
$C_3$ 56pF, $C_5$ 5μF, 25V
With $V_{IN}$ 20V, $V_{OUT}$ variable in range 0 to 13.9V at $I_L$ 40mA
Full load regulation <0.01%
Line regulation 0.02%/V
Standby current 8mA

**Circuit description**
The voltage regulator shown above uses three monolithic integrated circuits. $A_1$ is a bipolar-c.m.o.s. hybrid operational amplifier, $Tr_1$ to $Tr_5$ are contained within one bipolar array package, and $D_1$ to $D_4$ plus $Tr_6$ are contained in another identical package. Diodes $D_1$, $D_2$ and $D_4$ are bipolar transistors with collector and emitter strapped and operating in reverse bias in the breakdown region to serve as zener diodes having a zener voltage of about 7.3V. Diode $D_3$ is forward-biased and consists of a transistor in the same package with its collector and base strapped. Resistor $R_1$ and the series-connected zener diodes $D_1$ and $D_2$ act as a simple shunt regulator across the input to provide a regulated supply of $2V_Z$ for the CA3130 operational amplifier. The output from this part of the circuit also serves as a pre-regulated input to the low-impedance, temperature-compensated voltage reference source consisting of $R_2$, $D_3$, $D_4$, $R_3$ and $Tr_6$—the diodes and transistor being part of the same monolithic structure. The output from this reference source is taken to the non-inverting input of the error amplifier via potentiometer $R_8$ which allows the amplifier's reference to be varied continuously over the range 0V to about 8.3V allowing the output voltage to be controlled over the range 0V to about 13.9V. A fraction of the output voltage is fed to the inverting input of the error amplifier by means of the potential divider $R_{10}$ and $R_{11}$.

Transistors $Tr_1$ to $Tr_4$ are contained in a single integrated circuit package and are all connected in parallel to act as an equivalent series transistor (emitter follower) which is capable of handling the full-load current, when driven from the output of the error amplifier. Transistor $Tr_5$ (in the same package) in conjunction with $R_4$, $R_5$ and $R_6$ serves as a current limiting device. If the load current increases the p.d. across $R_4+R_5$ increases and since the base voltage of $Tr_5$ is held at approximately 600mV above $V_{OUT}$ the base current to $Tr_5$ through $R_6$ increases. Hence the collector current of $Tr_5$ increases, diverting the base current from the series pass transistors. Thus the collector currents of these transistors, and hence the load current, falls back to its previous value. The value of load current at which the current limit becomes operative is set by $R_5$, the maximum limited current being determined by $R_4$, a 3-Ω resistor. Capacitor $C_3$ provides compensation for the operational amplifier the other capacitors serving to remove residual hum at the input and to control the closed-loop gain of the amplifier at high frequencies.



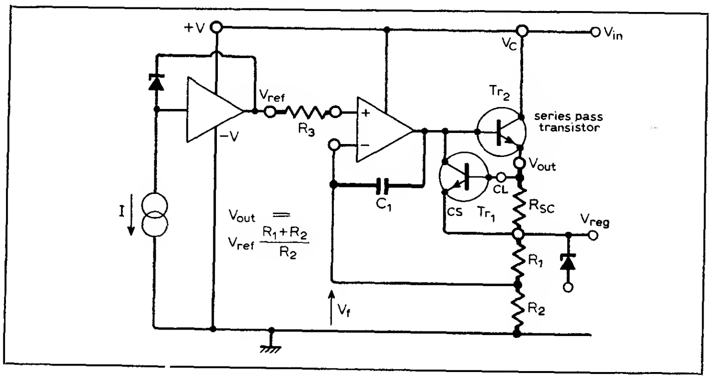

**Modification**
Careful layout of the printed circuit is essential otherwise the circuit may oscillate. This form of circuit may be modified to provide output voltages in the range 100mV to 50V at currents up to 1A with a 55V input by making the following changes. $R_1$ is increased to 4.3kΩ, 1W and a 1kΩ resistor connected between $D_4$ anode and $Tr_6$ base; otherwise the pre-regulator and voltage reference circuits are basically unchanged. The sliding contact of $R_8$ is connected directly to the inverting input of $A_1$. $R_9$ is omitted along with $C_4$ and the compensation capacitor $C_3$ is increased to 1nF. $R_{10}$ and $R_{11}$ are changed to 43kΩ and 8.2kΩ respectively, and their junction taken to the non-inverting input of $A_1$. This inversion of the inputs to $A_1$ is required because of the addition of an inverting current-boosting stage at the output of $A_1$ as shown. Darlington-connected series transistors $Tr_6$ and $Tr_7$ replace $Tr_1$ to $Tr_4$ and $Tr_8$ provides current limiting by adjustment of $R_{15}$.

Typical components are
$Tr_6$ 2N3055, $Tr_7$, $Tr_9$ 2N2102
$Tr_8$ 2N5294, $R_{12}$ 3.3kΩ, 1W
$R_{13}$ 1kΩ, $R_{14}$ 1Ω, $R_{15}$ 10kΩ
$C_1$, $C_5$ 100μF, $C_2$ 5μF
Full load regulation 0.05%
Line regulation 0.01%/V

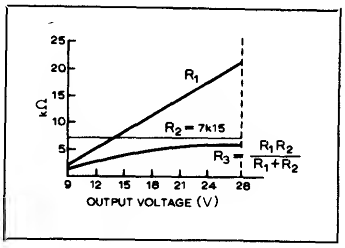
**Cross references**
Set 24, cards 1, 2, 3, 5, 6.
Set 23, card 3.
Set 20, cards 1, 2.

**Further reading**
Solid State Datasheet, RCA no. 817 (CA3130), 1974.
Solid State Databook, SSD-201B, RCA, 1974, pp. 183–8.

## Monolithic regulators—1



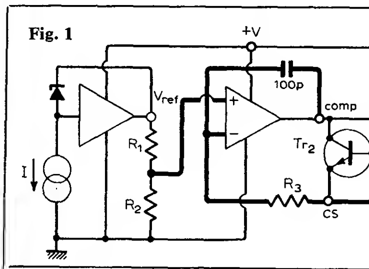

**Typical data**
IC $\mu$A723C or LM723C
Temperature range 0 to 70°C
Line regulation for $V_{IN}$ 12 to 40V 0.1%. For 12-15V, 0.01%
Load regulation 0.03% for 1 to 50mA current range
$R_1$ 7.87k$\Omega$ ±5%
$R_2$ 7.15k$\Omega$ ±5%
$V_{REG}$ +12V, $C_1$ 100pF
Ripple rejection 74dB
Temp. coeff. of 0.003%/degC
$V_{ref}$ 7.15V
Standby current 2.3mA for $V_{IN}$ 30V
Input voltage range 9.5 to 40V
Output voltage range 7 to 37V

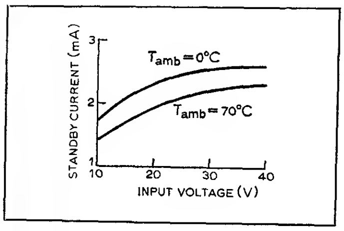



**Circuit description**
The schematic diagram of this regulator package is shown above, with the external components for a high voltage regulator circuit. The series-pass transistor is connected as an emitter-follower, and the amount of feedback to the internal operational-amplifier is defined by $R_1$ and $R_2$. $V_f$ is approximately equal to $V_{REF}$ because the differential input to the op-amp is very small, and hence, as $R_{SC} \ll R_1$ or $R_2$. $V_{OUT} = V_{REF}(R_1+R_2)/R_2$ i.e. the emitter of the series-pass transistor is constrained to be a multiple of $V_{REF}$. Therefore, if the unregulated input $V_{IN}$ increases, this increase must be absorbed by an increase in the collector-emitter voltage of series transistor. Conceptually, if the emitter potential tends to increase, then $V_f$ to the inverting input would increase, which would cause a decrease in potential at the base of the series transistor, and due to emitter-follower action, this opposes the assumed increase.
*Load regulation*. Percentage change in output voltage for a specified load current change.
*Line regulation*. Percentage change in output voltage for a defined change in input voltage. Note—above are defined for a constant junction temp.
*Ripple rejection*. Ratio of pk-pk input ripple voltage to pk-pk output ripple voltage.
*Input-output differential*. Working range of regulator based on difference between supply and regulated voltage.
*Standby current*. Current drain for no load on output or reference.
Note—regulation is sometimes defined on basis of a percentage change in input.
Fig. 1 is a low-voltage arrangement suitable for a 2 to 7V output voltage range $V_{OUT} = V_{REF}R_2/(R_1+R_2)$
For $V_{REG}$ of +5V, $R_1$ 2.15k$\Omega$, $R_2$ 4.99k$\Omega$, $R_3$ 1.5k$\Omega$.
Fig. 2 provides a negative regulated voltage suitable for a −9V to −28V range. Typically $V_{REG}$ −15V, $R_1$ 3.65k$\Omega$, $R_2$ 11.5k$\Omega$, $R_3$ $R_4$ 3k$\Omega$, $R_5$ 2k$\Omega$, $Tr_3$ 2N4898.
An extension to −6V is possible but $V^+$ must be at +3V minimum.
Fig. 3 is similar to Fig. 1 but permits a remote shutdown facility via a logic source.
Current limiting and sensing depends on the value of $R_{SC}$. When the load current is large enough to cause the potential between $C_L$ and $C_S$ to turn on the related transistor, this removes drive current from $Tr_1$ to limit any further increase in output current.
Curves above show typical load regulation and current limiting characteristics for $V_{OUT}$ 5V, $R_{SC}$ 10$\Omega$, $V_{IN}$ +12V.

**Further reading**
Hinatek, E. R. Users Handbook of Integrated Circuits, Wiley, 1973.
$\mu$A723 The Universal Voltage Regulator, Fairchild.

**Fig. 1**



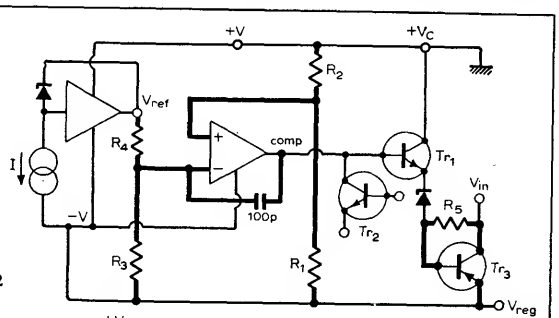




**Fig. 2**

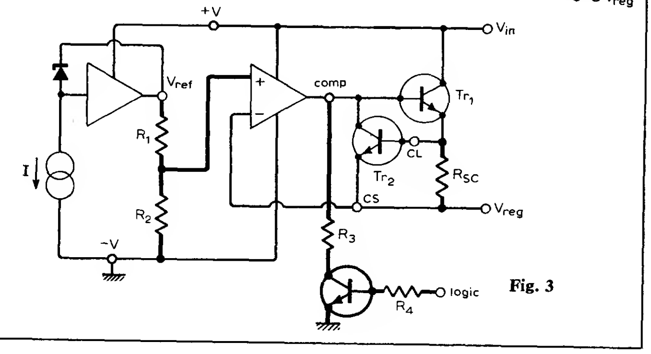
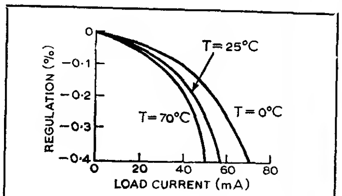
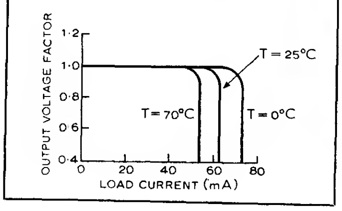



**Fig. 3**



# Monolithic regulators—2

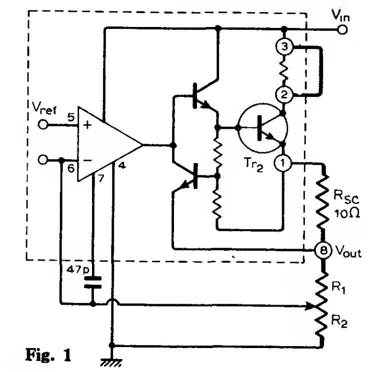



Fig. 1

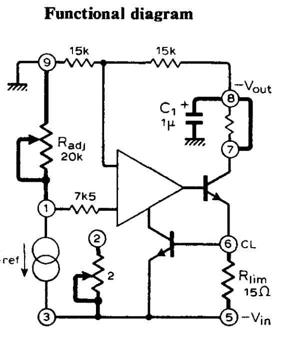



Fig. 2

The schematic diagram of the LM105/205/305 group is within the dashed box. External components provide a basic low current positive regulator circuit.

LM305, Fig 1.
Temperature range: 0 to 70°C
Input voltage range: 8.5 to 40V
Output voltage range: 4.5 to 30V
Output current: 20mA
Load regulation: 0.03% for load current 0 to 12mA
Line regulation: depends on $V_{IN}-V_{OUT}$ differential 0.025%/V
Parallel combination of $R_1$ and $R_2$ should be about 2kΩ
LM305A can provide 45mA.

**Negative voltage regulator** (LM104, Fig. 2, Current reference is temperature compensated. Output voltage programmed by value of $R_{ADJ}$. $R_{LIM}$ provides short-circuit protection. $C_1$ (tantalum) prevents oscillation.
Output current: 25mA
Input range: −50 to −8V
Output range: −40V to −15mV
Typical load regulation: 0.05% from 0 to 25mA
Typical line regulation: better than 0.2% for ±20% input change
$V_{OUT}$: $R_{ADJ}/500$ V
The LM104/LM105 interconnection, Fig. 2, provides a dual polarity tracking regulator.
Using the LM104 as an inverting amplifier i.e. $+V_{OUT}$ at pin 9 appears as $-V_{OUT}$ at pin 8. $V_{IN} \geq \pm 18$V,
$V_{OUT}$ ±15V defined by potential divider chain $R_1$, $R_2$
Output current: 200mA
The RC4195 or MC1468 in Fig. 3 provide a dual balanced ±15V supply in one package, with current capability of around 100mA. Input voltage range 18 to 30V.
The RC4194 is a dual tracking voltage regulator in which the positive and negative output voltages are adjustable over the range 0.05 to ±32V by variation of $R_7$. This should be 2.5kΩ for each volt required.
Input voltage range: 9.5 to 35V
Load regulation (1 to 100mA) 0.001% $V_{OUT}$/mA
Line regulation: For a 10% change in $V_{IN}$ 0.02% $V_{OUT}$
Load current: 100mA
An unbalanced output (+12V, −6V) suitable for comparators is obtainable with $R_0$ of 15kΩ and the addition of $R_B$ of 20kΩ. (LM340) is a three-terminal series positive regulator. It uses an internal temperature independent voltage reference dependent on the predictable gap-energy voltage (card 6, set 23).
Preset output voltages depend on the internal resistor $R_1$.
LM109/309 are earlier versions on the same principle, but designed specifically for +5V logic levels. Fig. 5 is an adjustable output circuit. Capacitors are optional depending on transient response requirement and distance from supply.
Another advantage of this i.c. is the internal circuitry which provides shutdown of the regulator if the die temperature reaches 175°C, thus providing virtually absolute protection.

Fig. 5

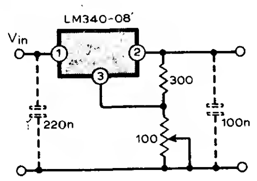



Fig. 2

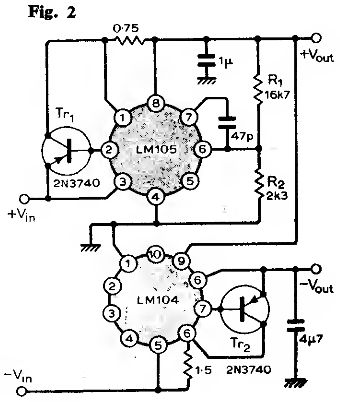
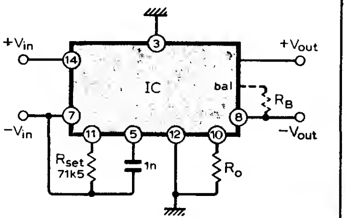



Fig. 3




The current capability of most voltage regulators can be boosted with additional external series transistors. A typical configuration extending the LM105 capability to 2A is shown in Fig. 6.
Foldback current limiting allows control of the current when the output is short circuit.
Appro imately, for a short circuit voltage across $R_{SC} = V_{be} - mV_{be}$. At full load, voltage across $R_{SC}$ is $(1-m)V_{be} + k\,V_{out}$.

Hence $\dfrac{I_{L\,max}}{I_{SC}} = 1 + \dfrac{kV_{out}}{(1-m)V_{be}}$

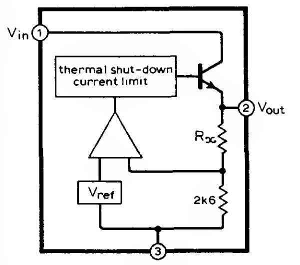
For a specific $V_{out}$, the current ratio is controlled by k. The typical V-I curve is given by

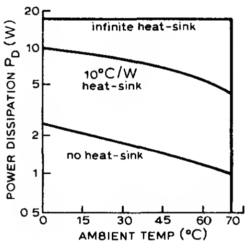
$I_L = \dfrac{(1-m)V_{be} + k\,V_{out}}{R_{SC}}$

**Further reading**
Application notes AN103 (LM340); AN23 (LM105), AN82 (precision tracking regulators), National Semiconductor.
Linear Integrated Circuits Data Book, Motorola, 1972.
Total Linears, Raytheon, 1974.

**Cross references**
Set 24, cards 8, 3, 4
Set 23, card 6
Set 6, cards 2, 10
Set 7, card 11
Set 20, cards 1, 2, 9.

Fig. 4



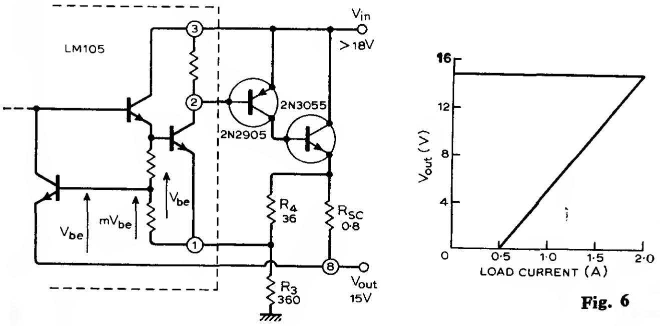
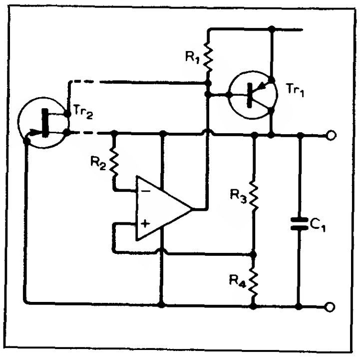




Fig. 6

# Voltage regulation using current-differencing amplifiers

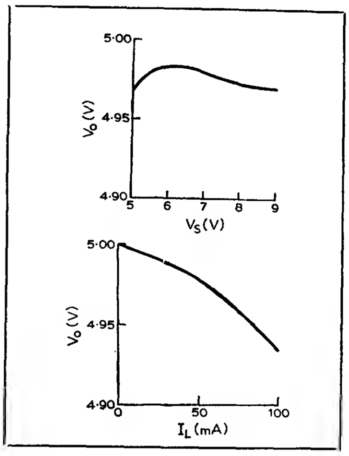



**Typical performance**
$R_1$ 330Ω
$R_2$ 1MΩ
$R_3$, $R_4$ 10kΩ potentiometer set for $V_0$ of 5V. Typically $R_4$=675Ω
$Tr_1$ BC125
$Tr_2$ 2N5457
C 10μF tantalum
V +7V

**Circuit description**
The basic voltage regulators described previously using a current-differencing amplifier had two distinct limitations. The obvious one is the very limited output current available, and this can be overcome by adding an emitter follower inside the feedback loop. This actually increases the second problem—that the minimum value of the supply voltage has to be one or more volts above the regulated output. It is possible to solve both problems while simultaneously improving the regulation against supply changes, if the amplifier is supplied from the regulator output (simultaneously regulating the supply to the three other amplifiers in the package). The trick is to make use of the ability of the output stage to sink current safely even when the output potential is greater than that on the amplifier positive supply terminal (provided the difference does not exceed 5V, breakdown in the internal p-n junctions is avoided). The minimum sink current in this mode is 1.3mA and $R_1$ is chosen so that $Tr_1$ is kept out of conduction when minimum output current is required. In the simplest form shown '$nV_{be}$' biasing is used that fixes the output voltage at $(R_3/R_4+1)V_{be}$ where the $V_{be}$ is that of the internal transistor at the amplifier non-inverting input. $R_1$ provides a small bias current to the inverting input. (Improved regulation would follow from the replacement of $R_3$ by a suitable zener diode.) The main problem remaining is that the circuit is not self-starting since with output temporarily at zero no current flows and the state is held permanently. One solution is to add a junction f.e.t. of low pinch-off voltage and on-current sufficient to bring $Tr_1$ into conduction. Once the output voltage is established, the reverse bias on the f.e.t. gate-source cuts it off and prevents it from disturbing the normal operation.

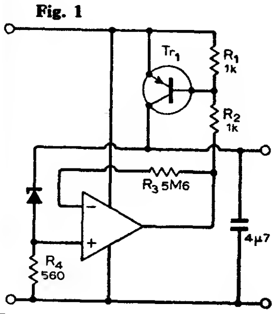
**Component changes**
$Tr_1$: Any silicon p-n-p transistor with suitable current/power rating—circuit can supply up to 200mA but maximum $V_{IN}-V_{OUT}$ rating limited by internal breakdown of amplifier to 5V i.e. 1W dissipation is adequate. BFR81.
$Tr_2$: Junction f.e.t. n-channel. Pinch-off voltage< regulated output. Zero-bias on-current must be sufficient to drive $Tr_1$ into conduction—typically >2mA.
$R_1$: 150 to 390Ω. If resistor is too high the minimum sink current of 1.3mA drives $Tr_1$ into conduction losing control at light loading. If $R_1$ is too low, insufficient drive current is available for $Tr_1$.
$R_3$, $R_4$: In this mode of operation, the potential at the non-inverting input is 0.6V and the ratio of $R_3$: $R_4$ scales this up to $[(R_3/R_4)+1]$ 0.6V. Stability is considerably increased by replacing $R_3$ with a zener diode when $V_0=V_Z+$ 0.6V.
$R_2$: Not critical. Sets operating currents of input transistors. Suitable values 1 to 10MΩ.

**Circuit modifications**
- For increased input-output voltage differential the amplifier is supplied directly from V+. To allow the amplier output to be out of saturation the base of $Tr_1$ is driven through a potential divider. Without this $Tr_1$ could not be driven off. The upper voltage limit is then the rating of the i.c. (36V for the LM3900). All other amplifiers in the package are subject to the full supply voltage variations.




- To increase the supply voltage rating further while retaining a low $(V_{IN}-V_{OUT})$ a second transistor is added such that all terminals of the amplifier are operated at a low voltage while $Tr_1$, $Tr_2$ must be chosen for a suitable voltage rating. An alternative zener circuit is shown in which $R_6$, $R_4$ set the output voltage $R_4V_Z/R_6$ in the absence of $R_7$. A Resistor to ground from either input causes current flow in either $R_4$ or $R_6$, and the resulting contribution to the output voltage has a temperature coefficient which can be used for overall temperature compensation. As shown $R_7$ contributes $-[1+(R_4/R_7)]V_{be}$ to the output.
- To remove the effect of supply variations via $R_1$ a diode network is chosen that ensures self-starting but has $D_3$ dropping out of conduction after starting has been achieved.

**Fig. 1**



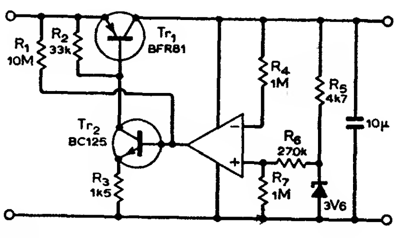




**Fig. 2**

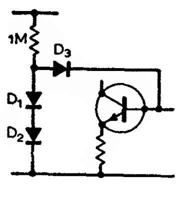
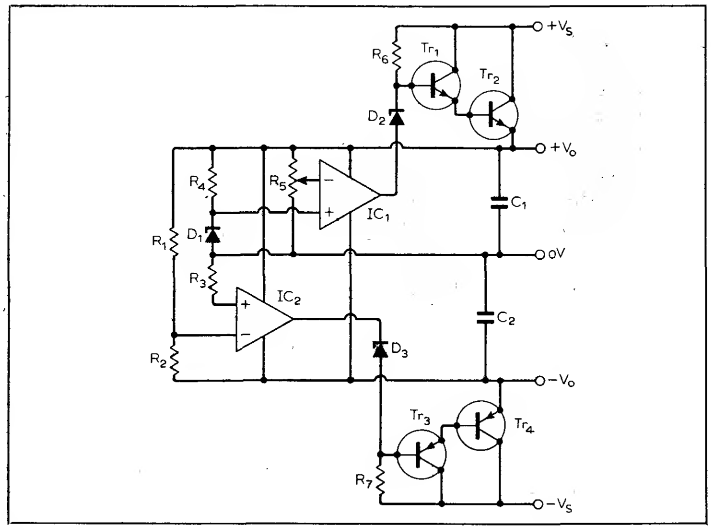



**Fig. 3**

**Further reading**
Frederiksen, T. M., Howard, W. M., Sleeth, R. S. The LM3900, A New Current-differencing Quad of ± Input Amplifiers, National Semiconductor application note AN72

## Dual-polarity regulator




**Typical performance**
$IC_{1-2}$ 741
$Tr_1$ BFR41, $Tr_2$ 2N3055
$Tr_3$ BFR81, $Tr_4$ 2N2955
$D_{1-3}$ 6.8V zener diodes

$R_{1-2}$ 8.2kΩ, $R_3$ 3.9kΩ
$R_4$ 2.2kΩ; $R_5$ 10kΩ
$R_{6-7}$ 22kΩ, $C_{1-2}$ 47μF
$V_S$ ±15V
$V_O$ ±10V, $I_O$ 0-1A

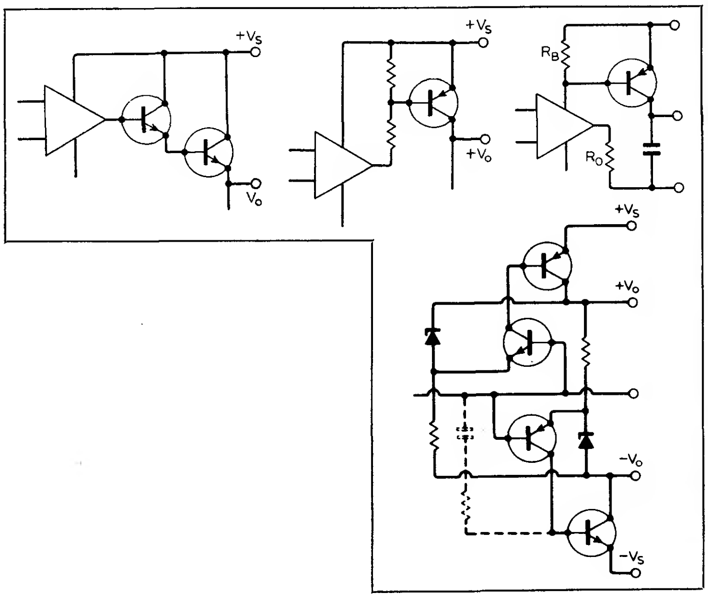
**Circuit description**
In a dual regulator it can be very important that the outputs track well. This can be achieved by having one section dependent on a particular zener diode or other reference element; the other output uses the output of the first as its own reference. Any variation in the zener voltage whether due to supply or temperature changes affects each equally. To maximize the regulation the zener diode and the error amplifiers should if possible be supplied from the regulated outputs. This can complicate the coupling network between each amplifier and the power output stage. The positive regulator compares a variable portion of the output via $R_5$ with the constant voltage across $D_1$. Any difference is amplified by $IC_1$, whose output is coupled via $D_2$ to the Darlington pair composed of $Tr_1$, 2. The negative output is controlled by $IC_2$ via $Tr_{3, 4}$, the amplifier operating in the virtual earth mode with $R_{1, 2}$ defining the inverting gain.

For $R_1 = R_2$ the positive and negative output voltages are equal in magnitude. As shown the positive output is restricted to values greater than the zener voltage, but the negative output can take up values from zero to just short of the negative supply.
The outputs are highly stabilized against both supply and load current changes (typically to within 1 or 2mV) and the stability is limited by that of the zener diode $D_1$.

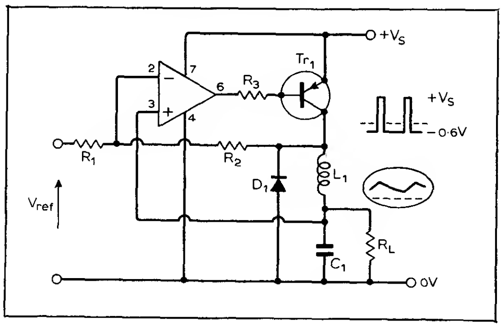
**Component changes**
$IC_{1, 2}$. Most compensated op-amps may be directly substituted. The output stage contributes no additional voltage gain and hence no change in compensation is warranted. $Tr_{1-4}$. The drive transistors are standard silicon medium-power devices and a maximum collector current of a few tens of milliamperes is sufficient for output currents beyond 1A. The power devices may then have to dissipate considerable power under short circuit conditions, i.e. current limiting should be added or adequate heat-sinking provided.
$D_1$: Zener diode with low temperature coefficient for minimum drift.
$D_{2-3}$: Not critical. Included to allow op-amp outputs to remain in linear region while retaining control of output. Diodes can be replaced by resistors typically of same value as $R_{6-7}$.
$R_{1, 2}$: Equal for precise tracking of outputs. 1 to 100kΩ.
$R_3$: Minimizes offset if $R_3 = R_1/R_2$. Can be omitted.
$R_4$: Sets zener diode current to optimum for low drift. 470Ω to 10kΩ.
$R_5$: May be padded out with series resistors where pot. is to provide trimming action only 1 to 100kΩ. Lower range for least offset/drift though overall drift likely to be dominated by zener anyway.
$R_{6,7}$: Set maximum base drive and hence, give coarse limiting of output 1 to 100kΩ.
$C_{1,2}$: Suppress h.f. oscillation. Not critical but must be close to output or load inductance may initiate instability.
$V_S$: Because amplifiers powered from regulated outputs, $V_S$ can be high if transistors have appropriate ratings. Increase $D_{1,2}$ voltages to match.

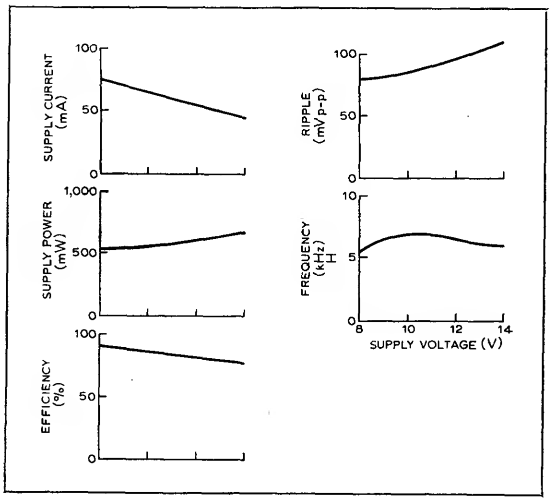
**Circuit modifications**
- The error amplifier outputs may be coupled to the power stage in several ways. Direct coupling reduces the component count but requires that the op-amp be powered from the supply rail. The input-output differential is increased to >3V in many cases.
- To reduce this, the output stage is operated in common emitter (with or without an intermediate driver). The inversion requires the op-amp inputs to be reversed and the resulting circuits are typically non-self-starting and require additional components for starting.
- As in previous power amplifiers the amplifier may drive a dummy load resistor $R_O$, the resulting current bringing the output transistor into conduction when the p.d. across $R_B$ exceeds 0.6V.
- A simple discrete form of the circuit has a good performance with few components but requires to be started either by a CR network together with the switch-on transient or by a separate switch.

**Further reading**
Eckhardt, R. Regulator for op-amps practically powers itself, *Electronics*, Oct. 3, 1974, p. 106.
Holmskov, Ole, Voltage stabilizing a symmetrical power supply, *Wireless World*, May 1975, p. 226.



# Switching regulator




**Typical performance**
$IC_1$ CA3130 (RCA)
$Tr_1$ BFR81
$D_1$ 1N4148
$L_1$ 680$\mu$H
$C_1$ 15$\mu$F
$R_1$ 1k$\Omega$
$R_2$ 470k$\Omega$
$R_3$ 680$\Omega$
$V_{REF}$ 5V
$R_L$ 50$\Omega$
$V_S$ 10V

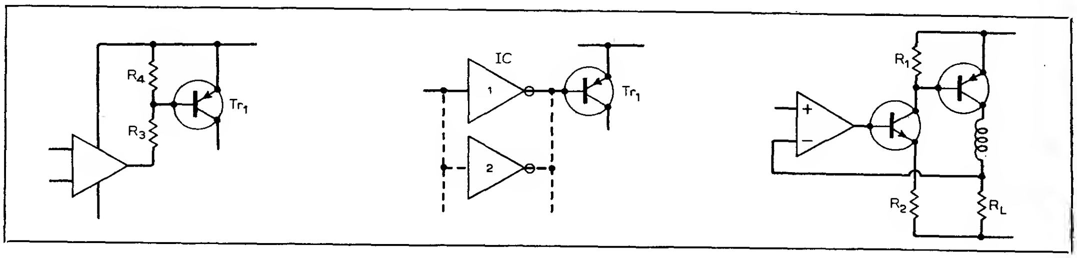
**Circuit description**
Switching regulators are related to Class-D switching amplifiers. The power stage $Tr_1$ conducts for a varying portion of the time. If the switching frequency is high, the current in $L_1$ varies little throughout the cycle, with $D_1$ sustaining the current in the load when the transistor is off. The inverting gain provided by $Tr_1$ reverse the effective polarity of gain at the amplifier inputs; 100% negative feedback is applied from the load to one input and with $L_1$ short circuit, a linear regulator would result were $V_{REF}$ to be fed directly to the other input. A small amount of hysteresis via $R_2$, $R_1$, combined with the $L_1R_L$ creates an astable—the LR equivalent of the standard op-amp CR astable. The load voltage has a similar exponential waveform with a ripple of the order $(R_1/R_2)V_S$ and a mean value of $V_{REF}$ when the hysteresis is small. Power losses include those due to the speed of switching including core losses in $L_1$, and the "d.c." losses such as $V_{ce(sat)}$ for $Tr_1$ and the on voltage of $D_1$. For low output voltages the latter limits the efficiency—between 70 and 90% is common even where the output voltage is $< V_S/2$. As the supply voltage varies the mean current changes in the opposite sense because the mark-space ratio is adjusted automatically via the astable action. Hence the mean power drawn from the supply depends mainly on the power required by the load.

**Component changes**
$IC_1$: This op-amp is particularly suitable for several reasons (i) high input resistance (m.o.s.) allows high $R_2$: $R_1$ ratio without $R_1$ becoming too low. (ii) input common-mode range includes zero line allowing control of output down to zero. (iii) high slew-rate allows switching speeds to be increased to suit optimum frequency-range of ferrite-cored inductor. (iv) c.m.o.s. output stage allows direct coupling to $Tr_1$ if needed with rapid switch-off reducing charge-storage problems.

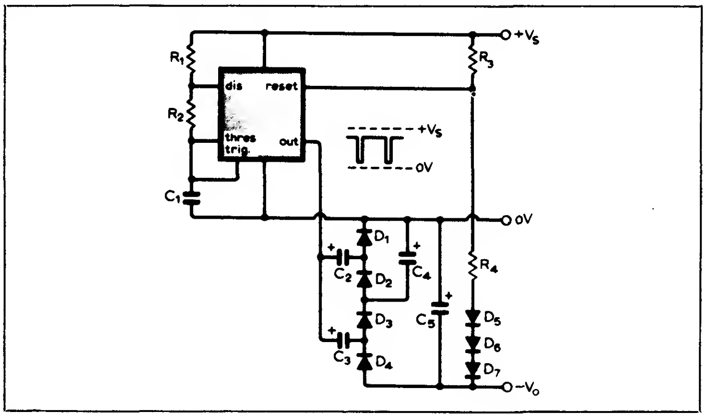
Most other un-compensated op-amps and comparators can be used provided following precautions observed: $V_{REF}$ must lie within input common-mode range; the output may not be able to swing high enough to switch $Tr_1$ off and a resistive network such as $R_3$, $R_4$ may be needed (left); current capability must be sufficient to saturate $Tr_1$ at max. load current—say $I_L/20$.
$Tr_1$ p-n-p silicon, peak current equal to mean load current; low saturation voltage; switching speed fast enough to make rise/fall times much less than the period of waveform; high frequency minimizes dissipation in transistor if above observed.
$D_1$: current rating mean output current; peak inverse voltage rating (p.i.v.) $V_S$; efficiency increased at low output voltages by reducing diode on-voltage (Schottky or germanium diodes if temperature not too high).
$L_1$: Typically 200$\mu$H—10mH depending on current/frequency used. Ferrite cores reduce size provide high Q, low losses; saturation of core at higher currents can inhibit oscillations.
$C_1$: Modifies frequencies for given $L_1$ $R_L$ combination. Not essential to operation, but reduces transients in load.
$R_1$, $R_2$: Ratio sets hysteresis and hence ripple. As ripple is reduced, so is time taken for completion of cycle i.e. frequency increases. By keeping $R_1$, $R_2$ as large as possible injection of switching current into $V_{REF}$ is minimized.
Ratio $R_2/R_1$ typ. 100 to 1,000; high value gives low ripple provided increased frequency does not bring transient problems in.
$R_3$: Not critical. 100$\Omega$ to 1k$\Omega$ with this op-amp.
$V_{REF}$: Equal to required output.
$R_L$: Load currents up to 200mA + possible
$V_S$: +5 to +15V.

**Circuit modifications**
Paralleled c.m.o.s. buffers may be used to boost output drive (centre). See op-amp data sheet. Alternatively use additional transistors (right). Final stage should be common emitter for highest efficiency. $R_1$, $R_2$ 100 to 470$\Omega$. Outputs to 1A.

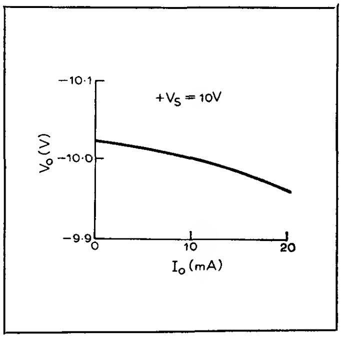
**Cross references**
Set 6, card 7.
Set 24, card 10.
Set 7, card 12.



## Self-regulating d.c.-d.c. converter




**Typical performance**
$IC_1$ 555 timer
$D_{1-7}$ 1N4148
$C_{2-5}$ 47$\mu$F, $C_1$ 0.015$\mu$F
$R_1$ 1.2k$\Omega$, $R_2$ 10k$\Omega$,
$R_{3,4}$ 22k$\Omega$
$+V_S$ +10V
$V_O$ −10V
$I_O$ 0 to −20mA
$V_O/V_S$ ±0.5% for $V_S$ 7 to 14V

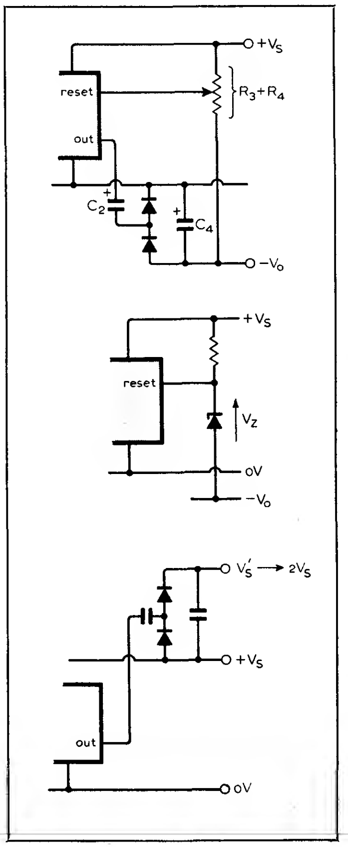



**Circuit description**
Dual-polarity supplies are needed in many systems where only a single supply is initially available. The circuit shown achieves this by acting as a free-running astable oscillator producing an output voltage just less than the supply. This is applied via a diode-capacitor network $D_{1-4}$, $C_{2-5}$ to produce a negative output voltage. Assume ideal diodes, $D_1$ clamps the right hand side of $C_2$ to zero on positive output swings; similarly $D_3$ clamps the right hand side of $C_3$ to $C_4$.

On negative swings, $C_2$ transfers charge via $D_2$ into $C_4$ as does $C_3$ through $D_4$ into $C_5$. Eventually $C_2$, $C_4$ each acquire a p.d. equal to the output swing, while $C_3$, $C_5$ achieve double that value. Two factors reduce this output voltage: losses across the diodes drop the maximum output by about 2V. The timer has a reset terminal; when the potential on this approaches ground the oscillations are inhibited.

A potential divider composed of $R_3$, $R_4$ and $D_{5-7}$ provides a potential at the RESET terminal such that each time the magnitude of the negative output increases, the oscillation is inhibited and the magnitude decreases. The diodes optimize the tracking for $|V_O| = V_S$.

**Component changes**
$IC_1$: The circuit depends on the particular characteristics of the 555 timer available from most i.c. suppliers.
$D_{1-7}$: Not critical. Any fast silicon diodes.
$C_1$: 470p to 0.1$\mu$F. At low frequencies ripple increases and at high frequencies switching losses.
$R_1$: 470 to 10k$\Omega$. Too low a value wastes current.
$R_2$: 2.2k to 22k$\Omega$. Select for frequency of oscillation.
$R_3$, $R_4$: can be replaced by potentiometer for variable output. Value not critical.
$V_S$: +5V to +20V
$V_O$: Up to $2V_S$ in magnitude.



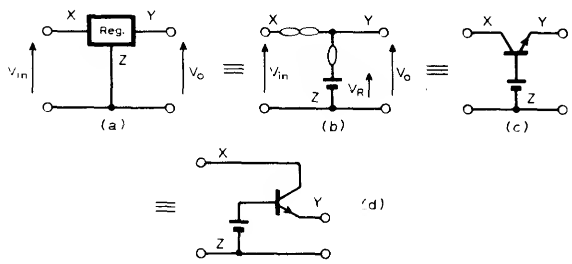

**Circuit modifications**
● For some applications the circuit can be considerably simplified. Where the negative voltage required is less than the positive supply available then the rectifying network can be simplified as shown. If the precise tracking of the two supplies is not important then the compensating diodes $D_{5-7}$ can be omitted. With $R_{3-4}$ replaced by a potentiometer, the result is a convenient circuit for producing, say, −6V from a +12V supply as required by widely used i.c. comparators.
● A second modification allows the negative output to be regulated rather than be proportional to the supply voltage. Replacing $R_4$ by a zener diode, the oscillator is reset each time the negative output pulls the reset terminal close to zero. The resulting ripple characteristics are comparable to the previous circuits. N.B. the regulation and ripple for a single diode pair is significantly better than for the voltage doubler, loads down to 100$\Omega$ being tolerated
● The same oscillator circuit can be used to generate a voltage more positive than the supply as shown. To regulate the output, a separate sensing circuit would be required since the original depended on using the RESET as a virtual earth. The circuit has affinities with certain re-triggerable monostables, and those based on op-amps in which the switching action is controlled by positive feedback can be adapted.
The two functions performed by the timer can be separated, with a clock generator driving a monostable. The latter is gated off each time the required output voltage is exceeded. Where the output swing of the timer is insufficient any of the usual power output stages may be added—Darlington pairs for increased current, complementary common emitter stages for increased output voltage swing. At low supply voltages conventional i.cs are not applicable, and discrete transistor oscillator-switches are required.

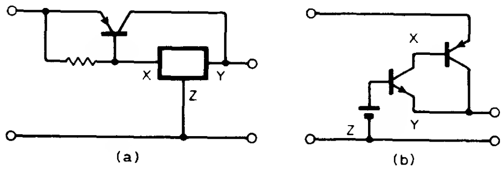
**Further reading**
Gartner, T. IC timer and voltage doubler form a dc-dc converter, *Electronics*, Aug. 22, 1974.

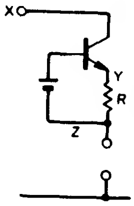
**Cross references**
Set 21, card 9.
Set 10, card 2.
Set 10, card 10.
Set 24, card 9.

The most significant advance in regulator design is arguably the appearance of the i.c. three-terminal regulator. These are now available in a range of voltage and current ratings, but there are bound to be gaps in this range that can be filled by more conventional variable-voltage multi-terminal devices. Alternatively the three-terminal devices can themselves be adapted if their nature is clearly understood. To find a suitable equivalent circuit compare Figs 1(a)–1(d). A nullor (combined with a voltage reference) equivalent to an ideal transistor with infinite voltage and current gain would provide the same performance as an ideal three-terminal regulator, viz that for all conditions for which $V_{IN} > > V_O$ then $V_O = V_R$. This follows from the nullor/ideal transistor properties that a nullator/base-emitter voltage is always zero and the norator/collector-emitter voltage may have any arbitrary value, with the load and supply currents being equal. Just as practical transistors and reference elements require biasing so practical voltage regulators carry a small but finite current in Z rendering the X and Y currents unequal. This current in Z is much less than typical load current and is also kept moderately constant by careful internal design of the i.c. If the equivalent circuit has a potential-divider of low resistance then $V'_O = (R_2/R_1+1)VR$ and the regulator gives a regulated voltage of any required output value $> V_R$ up to the device breakdown limits.

Fig. 3(a) shows a modification that is proposed for increasing the output current. By redrawing as in Fig. 3(b) the circuit is seen to be equivalent to a complementary Darlington pair (again assumed to be ideal) fed from the same voltage reference i.e. the current capability is scaled up by $h_{FE}$ with no change in output voltage. If the load is placed in series with the device as in Fig. 4 but with a dummy resistor R added between Z and Y, then the load current is defined as $V_R/R$, with the addition of the small current in Z mentioned above.

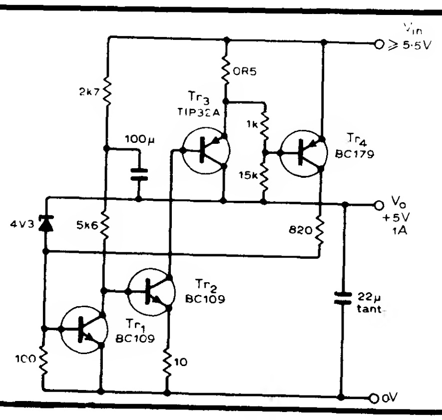



**Reference**

Krause, V.R. Adjustable voltage-switching regulator, *Wireless World*, vol. 82, May 1976, p.80.




As the performance/cost ratio of integrated-circuit voltage regulators continues to improve it is tempting to ignore discrete-component designs altogether.

This is dangerous because the constraints of i.c. processing, for example absence of high performance p-n-p transistors, can bring limitations. In particular for highest efficiency the output transistor has to operate in common emitter, and if a separate power device has to be added the advantages of a single-chip solution are abated. The circuit shown is a neat solution in that the input-output voltage differential can fall to ~0.5V or even towards zero if the current limiting network is eliminated (the 0.5Ω resistor $Tr_4$ etc). The performance claimed for this circuit is particularly good in respect of regulation against supply and load changes, the current-limiting is of the fold-back variety and the transistor types are non-critical.

Mitchell, K.W. High performance voltage regulator, *Wireless World*, vol. 82, May 1976, pp.83/4.