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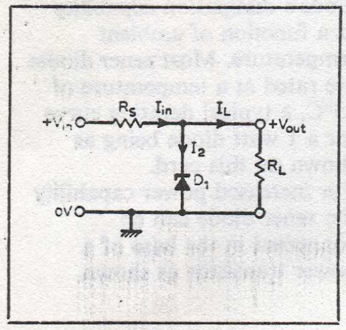
Set 24: Voltage regulators—1

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	0.0092
15	9.49	49	48	0.46
18	9.72	50.5	106.5	1.035

Typical performance
V_{IN} constant, I_L variable
D_1 BZY96C9V, V_{IN} 15V
$I_{L(max)}$ 100mA, $I_{L(min)}$ 80mA
R_S 54Ω 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_S), 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_S 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_S 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

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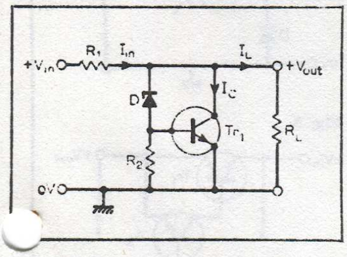
Set 24: Voltage regulators—2

Simple transistor regulators



Typical performance
Tr_1 BFR41, D_1 ESM18
R_1 330Ω, 3W; R_2 100Ω
R_L 1kΩ, $\frac{1}{2}$W
V_{IN} 32.5±7.5V
V_{OUT} see graphs opposite

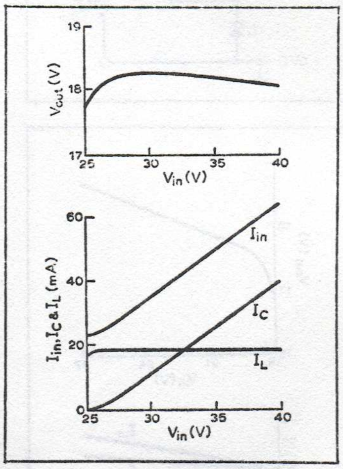


Fig. 1

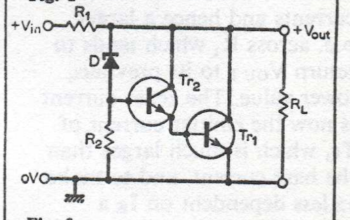
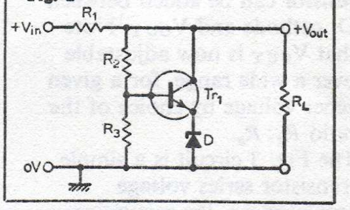


Fig. 2

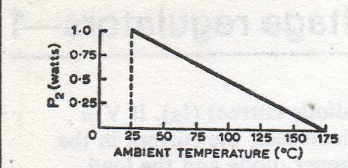


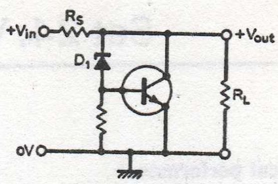
Description

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.

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





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 $I_Z=(V_{IN}-V)/R_S-I_L$
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

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

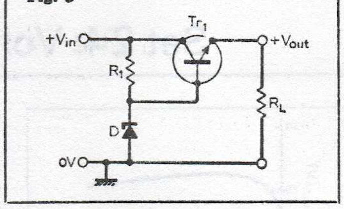
for I_L and V_{IN} variable.
A zener diode is then chosen having the desired nominal voltage and capable of safely dissipating this maximum power. It may be necessary to design a heat sink of suitable area and/or to derate the diode's dissipation capability as a function of ambient temperature. Most 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.

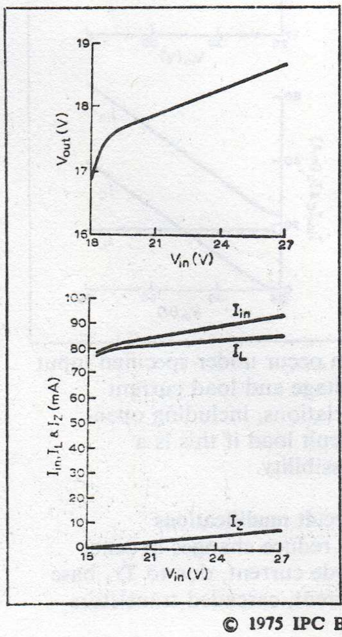
Further reading
Zener Diode Handbook, Motorola 1967.
Patchett, G. N. Automatic Voltage Regulators and Stabilizers, chapter 6, Pitman, 1970 (3rd edition).

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

series with Tr_1 emitter.
A fraction of the output voltage $V_{OUT}R_2/(R_2+R_3)$ 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,

Fig. 3





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 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

Fig. 4

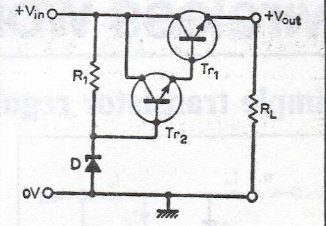
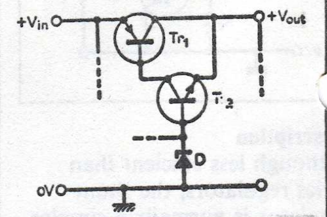


Fig. 5



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.

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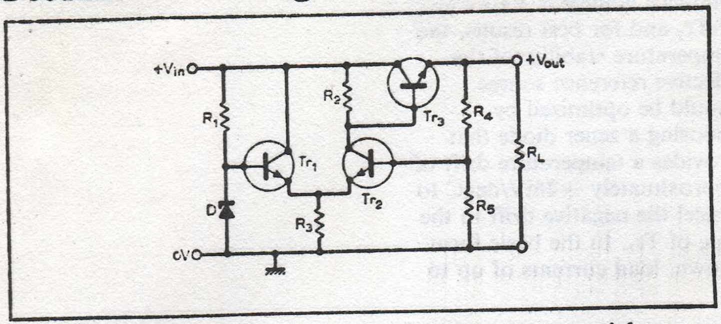
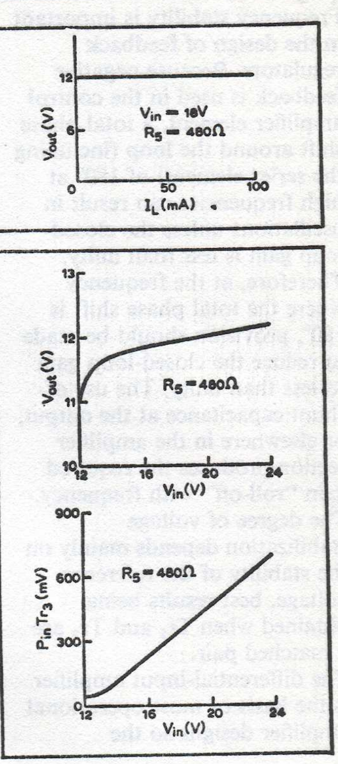
Set 24: Voltage regulators—3

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

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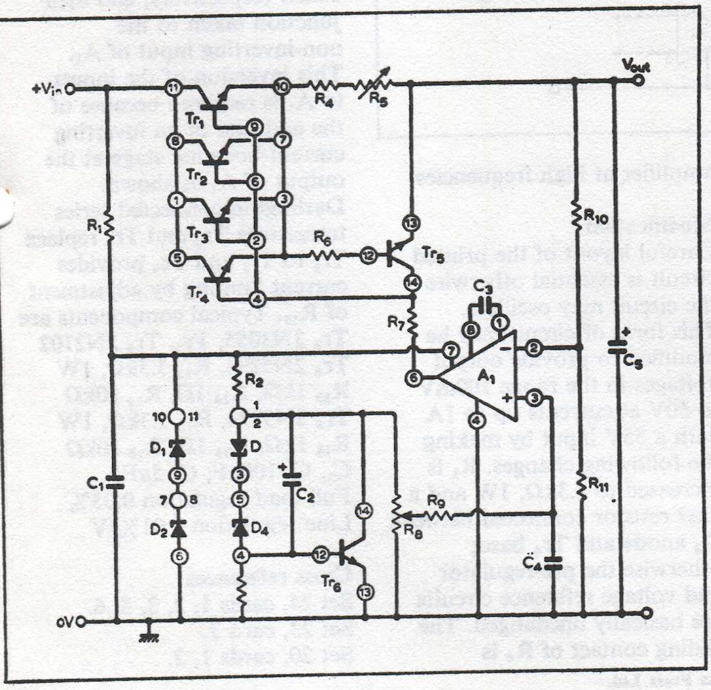
Set 24: Voltage regulators—4

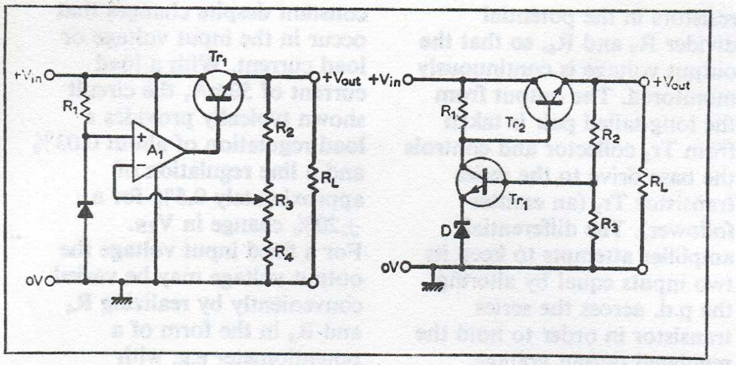
Bipolar/c.m.o.s. op-amp regulator



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

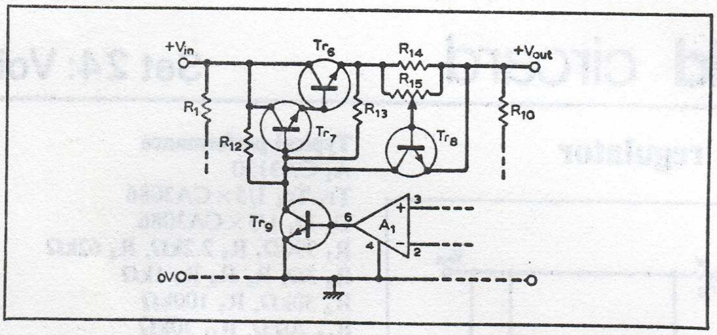
$(R_4+R_5)=1k\Omega$ and R_5 varied over the range 100 to 900Ω. V_{OUT} may be varied over the range 16.86 to 6.4V.
Frequency stability is important in the design of feedback regulators. Because negative feedback is used in the control amplifier element, a total phase shift around the loop (including 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_1 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.

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 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.

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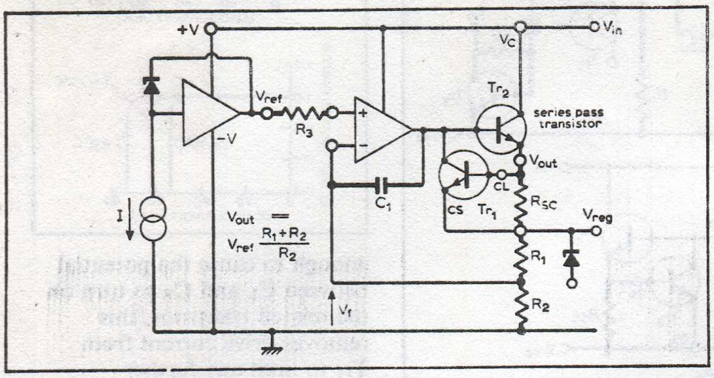
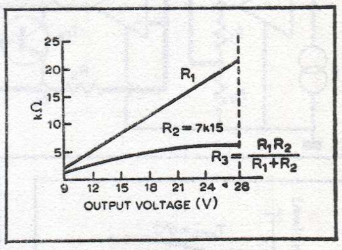
Monolithic regulators—1



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.

Typical data
IC μA723C 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% to 30mA
R_1 7.87kΩ ±5%
R_2 7.15kΩ ±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



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.

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Monolithic regulators—2

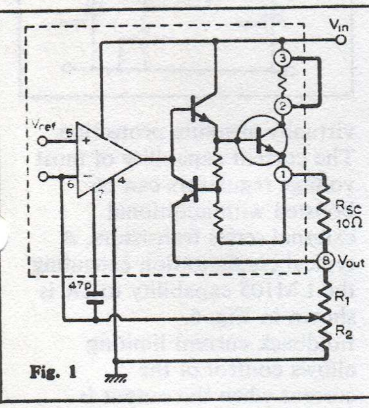
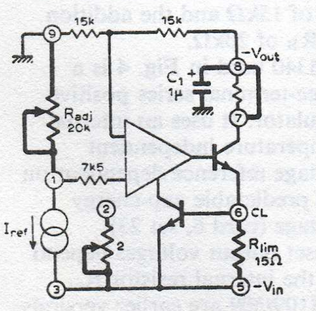


Fig. 1

Functional diagram



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 −V15m
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
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

Fig. 2

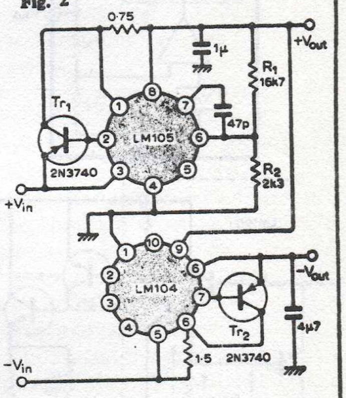


Fig. 3

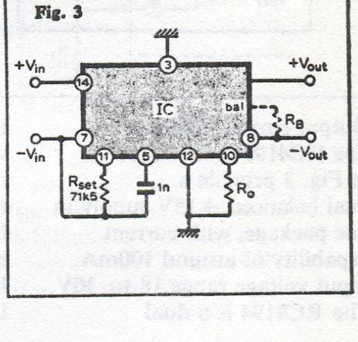
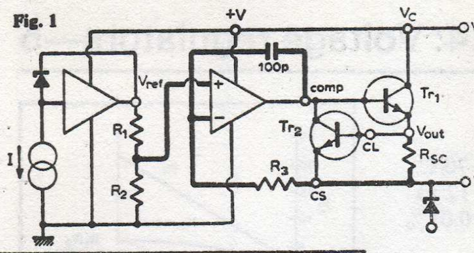
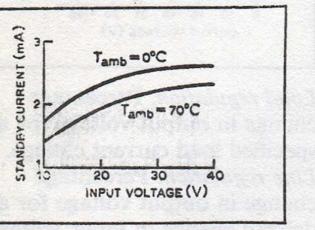


Fig. 1





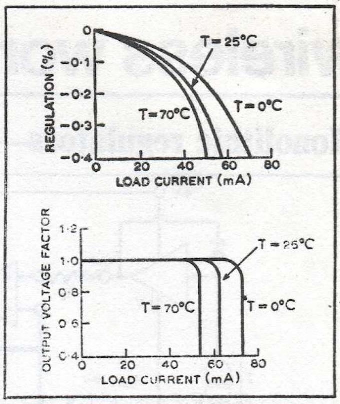


Fig. 2

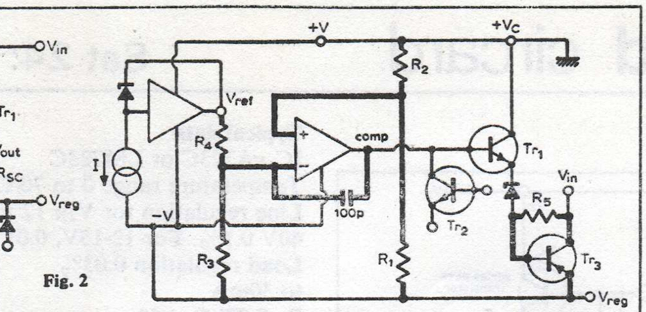
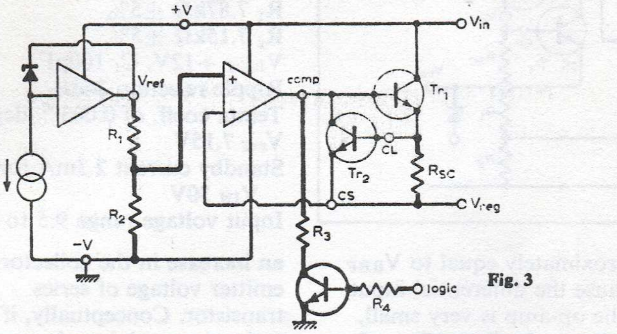


Fig. 3



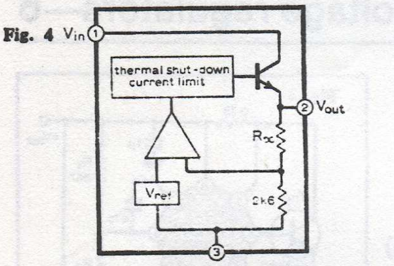
Note—regulation is sometimes defined on basis of a percentage change in input.
Fig. 1 is a low-voltage arrangement suitable for a 2 to 7V output voltage range
$V_{OUT} = V_{REF} R_2/(R_1 + R_2)$
For V_{REG} of +5V, R_1 2.15kΩ, R_2 4.99kΩ, R_3 1.5kΩ.
Fig. 2 provides a negative regulated voltage suitable for a −9V to −28V range.

Typically V_{REG} −15V, R_1 3.65kΩ, R_2 11.5kΩ, R_3 R_4 3kΩ, R_5 2kΩ, Tr_3 2N4898. An extension down 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} 10Ω, V_{IN} +12V.

Further reading
Hinatek, E. R. Users Handbook of Integrated Circuits, Wiley, 1973.
μA723 The Universal Voltage Regulator, Fairchild.

Fig. 4



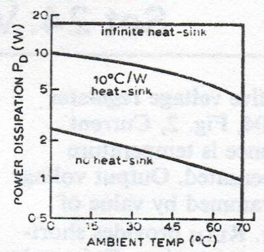


Fig. 6

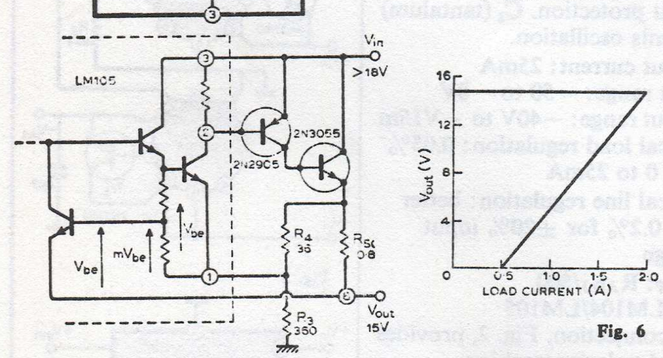
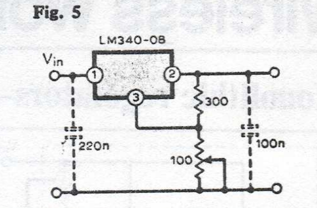
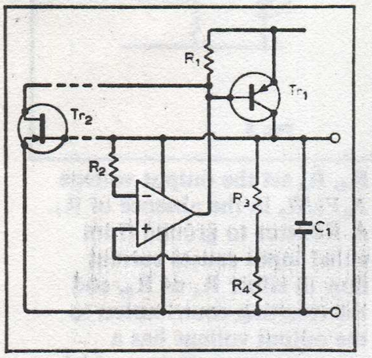


Fig. 5



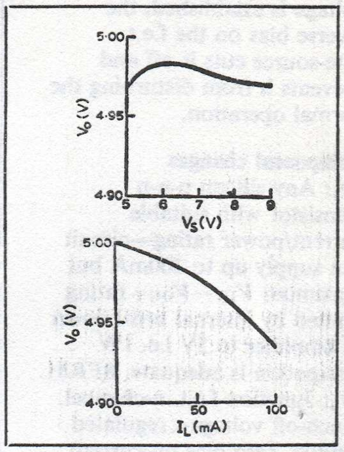
Output current: 200mA
The RC4195 or MC1468 in Fig. 3 provide a dual balanced ±15V supply in one package, with current capability of around 100mA. Input voltage range 18 to 30V. The RC4194 is a dual tracking voltage regulator in which the positive and negative output voltages are adjustable over the range 0.05 to ±32V by variation of R_o. 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_o of 15kΩ and the addition of R_B of 20kΩ.
LM340 used in Fig. 4 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_x.
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.
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.

Voltage regulation using current-differencing amplifiers



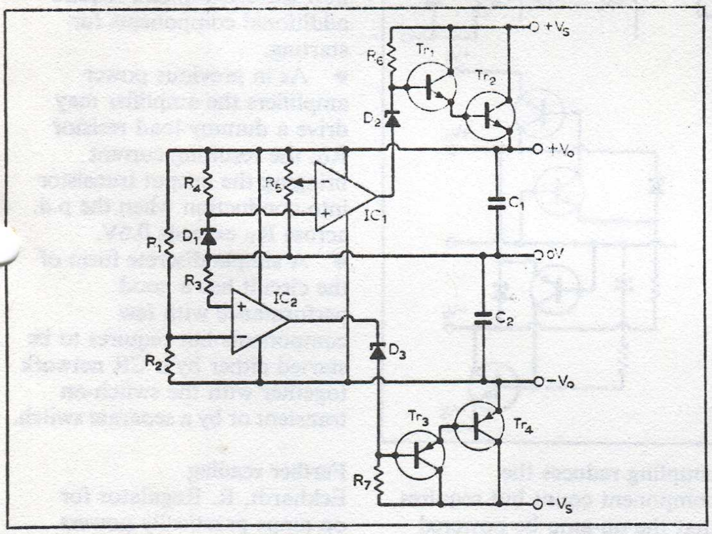
Typical performance
R_1 330Ω
R_2 1MΩ
R_3, R_4 10kΩ potentiometer set for V_o of 5V. Typically R_4=675Ω
Tr_1 BC125
Tr_2 2N5457
C 10μF tantalum
V +7V

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

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

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.

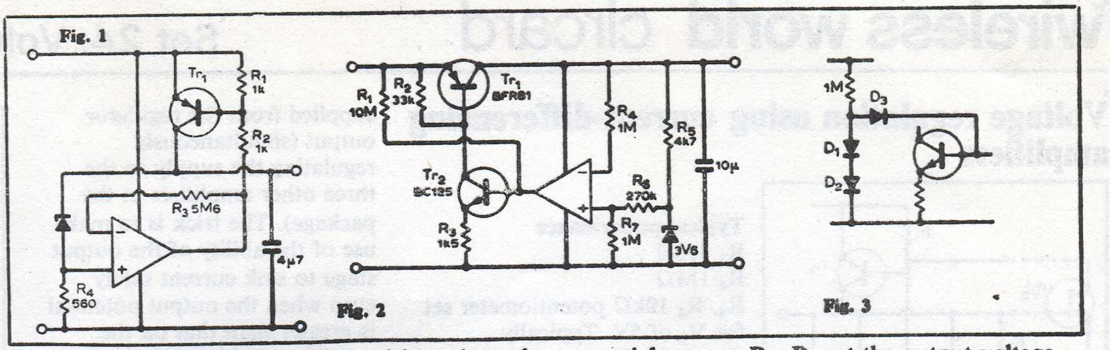
Component changes

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.

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.

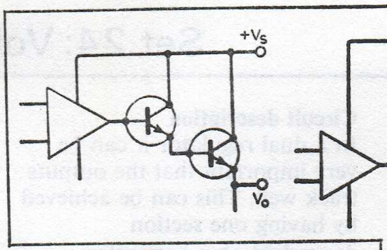
D_{2-3}: Not critical. Included to allow op-amp outputs to remain in linear region while retaining control of output. Diodes can be replaced by resistors typically of same value as R_{6-7}.

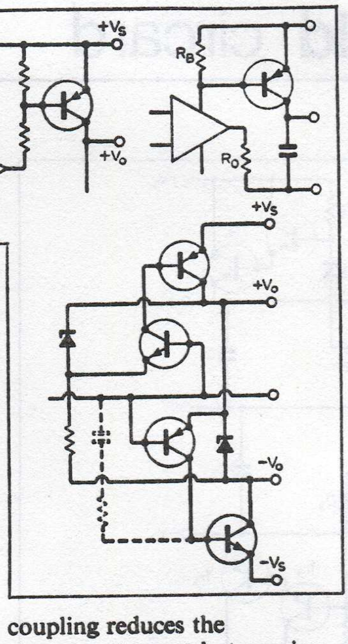
$R_{1,2}$: Equal for precise tracking of outputs. 1 to 100kΩ.

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.

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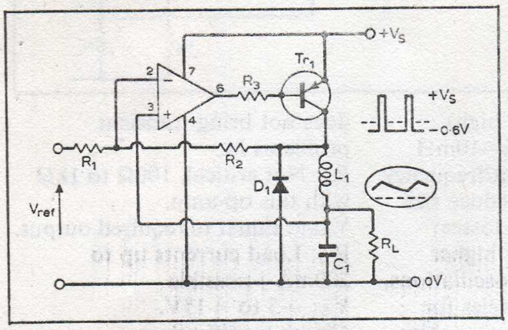
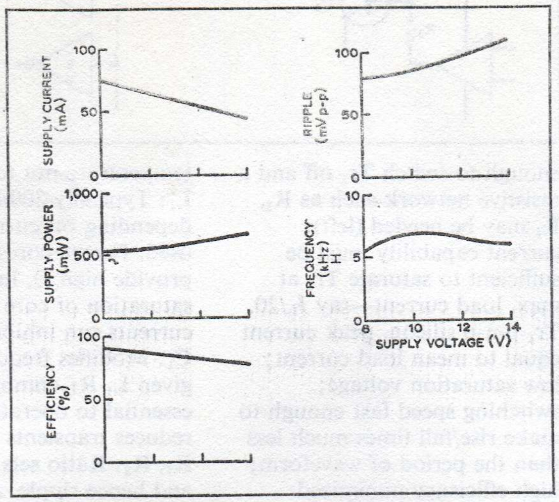
Switching regulator





Typical performance
IC_1 CA3130 (RCA)
Tr_1 BFR81
D_1 1N4148
L_1 680μH
C_1 15μF
R_1 1kΩ
R_2 470kΩ
R_3 680Ω
V_{REF} 5V
R_L 50Ω
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

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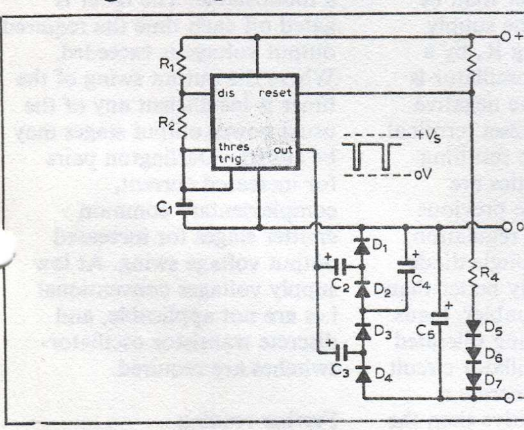
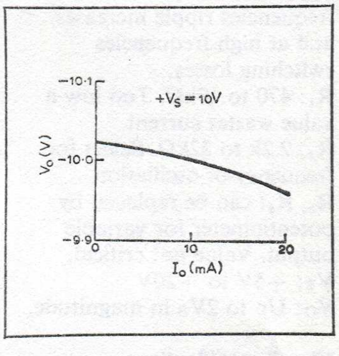
Self-regulating d.c.-d.c. converter



Typical performance
IC_1 555 timer
D_{1-7} 1N4148
C_{2-5} 47μF, C_1 0.015μF
R_1 1.2kΩ, R_2 10kΩ,
$R_{3,4}$ 22kΩ
$+V_S$ +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 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.

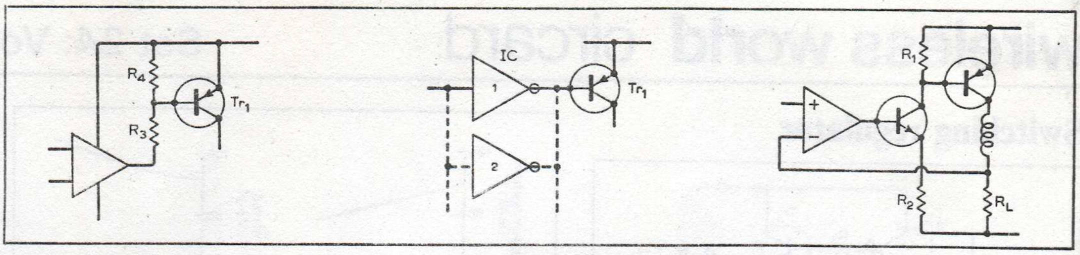


A potential divider composed of R_3, R_4 and D_{5-7} provides a potential at the RESET terminal such that each time the magnitude of the negative output increases, the oscillation is inhibited and the magnitude decreases. The diodes optimize the tracking for $|V_O| = V_S$.

Component changes
IC_1: The circuit depends on the particular characteristics of the 555 timer available from

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



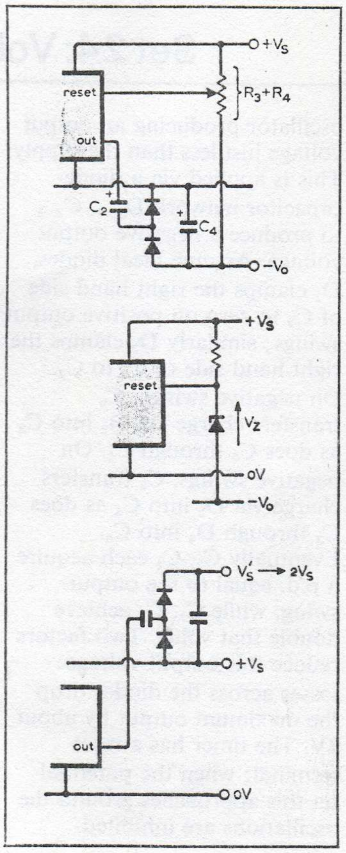
enough to switch Tr_1 off and a resistive network such as R_3, R_4 may be needed (left); current capability must be sufficient to saturate Tr_1 at max. load current—say $I_L/20$.
Tr_1 p-n-p silicon, peak current equal to mean load current; low saturation voltage; switching speed fast enough to make rise/fall times much less than the period of waveform; high efficiency minimized 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 200μH—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. 100Ω to 1kΩ 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 470Ω. Outputs to 1A.

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

most i.c. suppliers.
D_{1-7}: Not critical. Any fast silicon diodes.
C_1: 470p to 0.1μF. At low frequencies ripple increases and at high frequencies switching losses.
R_1: 470 to 10kΩ. Too low a value wastes current.
R_2: 2.2k to 22kΩ. 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 100Ω 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.