Zener diode shunt regulator

**Typical performance**
\( I_{\text{constant}}, V_{\text{IN \, variable}} \)

- \( D_1, \text{BZY96C9V}, \text{IL} \, 50\, \text{mA} \)
- \( V_{\text{IN(min)}} \, 12V, V_{\text{IN(max)}} \, 18V \)
- \( R_4 \, 54\, \Omega \, (39\, \Omega + 15\, \Omega) \)
- \( R_5 \, 195\, \Omega \, (3 \times 56\, \Omega + 27\, \Omega) \) to set \( I_L \) at 50mA

**Measured results**

<table>
<thead>
<tr>
<th>( V_{\text{IN}}(V) )</th>
<th>( I_{\text{OUT}}(\text{mA}) )</th>
<th>( I_L(\text{mA}) )</th>
<th>( P_{\text{D}}(\text{mW}) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>12</td>
<td>9.16</td>
<td>48</td>
<td>0.0092</td>
</tr>
<tr>
<td>15</td>
<td>9.49</td>
<td>48</td>
<td>0.046</td>
</tr>
<tr>
<td>18</td>
<td>9.72</td>
<td>50.5</td>
<td>1.035</td>
</tr>
</tbody>
</table>

**Simple transistor regulators**

**Typical performance**
\( T_{\text{TR41}}, D_1, \text{ESM18} \)
- \( R_1 \, 330\, \Omega, 3\, W; R_4 \, 1000\, \Omega \)
- \( R_2 \, 1\, \Omega, 1\, W \)
- \( V_{\text{IN}} \, 32.5 \pm 7.5V \)
- \( V_{\text{OUT}} \) 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 at the input voltage and/or the load current changes, causing a corresponding change in the p.d. across \( R_2 \). Transistor \( T_{\text{TR1}} \) 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 \( T_{\text{TR1}} \) base current, cascaded transistors may be used to increase the current gain of the regulating element, as Fig. 1. The base current of \( T_{\text{TR2}} \) is then only \( \approx I_{\text{TR1}}/h_{\text{FE2}} \) which can be made much smaller than the zener diode current by choice of \( R_4 \). An alternative form of the 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 $I_L = 0$, 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 Kirchhoff voltage equation for the circuit is:

$$V_{IN} = I_{IN}(R_s + V_Z)$$

where $I_{IN} = (V_{IN} - V_Z)/(R_s + I_L)$ and $V_Z = (V_{IN} - V_Z)/I_L - V_Z$ and the diode dissipation is

$$P_D = I_L V_Z = (V_{IN} - V_Z)R_s - I_L V_Z$$

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

$$R_s = \frac{V_{IN, min} - V_Z}{I_Z, min + I_L, max}$$

Using the nominal zener voltage $V_Z$ and an empirical factor of 10% of $I_L, max$, for $I_L, min$

gives

$$R_s = \frac{V_{IN, min} - V_Z}{1.1 I_L, max}$$

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

$$R_s = \frac{V_{IN} - V_Z}{1.1 I_L, max}$$

Having determined $R_s$, $P_Z$(max) can be found from

$$P_Z(max) = \frac{(V_{IN, max} - V_Z) - I_L, min - V_Z}{R_s}$$

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. Many 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

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

series with $T_3$, emitter.
A fraction of the output voltage $V_{OUT} = V_Z/R_s$ 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 $R_s$, which tends to return $V_{OUT}$ to its previous, lower value. The zener current is now the emitter current of $T_3$, 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 $R_s/R_c$.

The Fig. 3 circuit is a simple transistor series voltage regulator, i.e. the regulating element $T_3$ 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 $T_3$, 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 $T_3$ by choice of $R_s$ 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 $T_3$ must be chosen to dissipate the maximum power generated under $V_{IN, max}$ and $R_L, min$ conditions.

Typical performance
$T_3$, BFR41, D, ESM18
$R_s$, 1kΩ, ½W; $R_c$, 200Ω, 3W
$V_{OUT}$, 22.5 ± 4.5V
$V_{IN}$ see graphs above

The variation of current in the zener diode with base current can be reduced by replacing $T_3$ by a Darlington pair as in Fig. 4, where the base current of $T_3$ is then only

$I_L([1+h_{RE})(1+h_{RE}])$. 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
Feedback series regulators

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 $T_1$ and $T_2$. Resistor $R_1$ and $D_1$ act as a simple voltage reference circuit making $T_1$ base potential $V_Z$. The base potential of $T_2$ 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 $T_2$ collector and controls the base drive to the series transistor $T_3$ (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

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

Typical performance
$A_1$ CA3130
$T_1$, $T_2$ 1/5 x CA3086
$D_1$, $D_2$ 1/5 x CA3086
$R_1$, 390Ω, $R_2$, 2.2kΩ, $R_3$, 62kΩ
$R_4$, 1kΩ, $R_5$, 1kΩ
$R_6$, 50kΩ, $R_7$, 100kΩ
$R_8$, 20kΩ, $R_9$, 30kΩ
$C_1$, 10nF, $C_2$, 25µF, 15V
$C_3$, 56µF, $C_4$, 5µF, 25V
With $V_{IN}$ 20V, $V_{OUT}$ variable in range 0 to 13.9V at IL 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, $T_1$ to $T_5$ are contained within one bipolar array package, and $D_1$ to $D_4$ plus $T_3$ are contained in another identical package. Diodes $D_1$, $D_2$, and $D_3$ 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 $D_4$ 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 $2V_A$ 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 $R_4$, $D_5$, $D_6$, $R_8$, and $T_5$, 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
(R₁ + R₂) = 1kΩ and R₃ varied over the range 100 to 900Ω. VₒU₊ 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 R₃ and R₅ are a matched pair.

The differential-input amplifier is the basis of most operational amplifier designs so the error amplifier via potentiometer R₅ 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 R₃ and R₅.

Transistors Tr₁ to Tr₅ 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 Tr₁ (in the same package) in conjunction with R₃ and R₅ serves as a current limiting device. If the load current increases, the p.d. across R₃ + R₅ increases and since the base voltage of Tr₁ is held at approximately 600mV above VₒU₊, the base current to Tr₅ through R₅ increases. Hence

the collector current of Tr₁ 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 R₅. The maximum limited current being determined by R₃ + R₅, a 3-Ω resistor. Capacitor C₂ 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. R₅ is increased to 4.3kΩ, 1W and a 1kΩ resistor connected between D₃ and Tr₅ base;

otherwise the pre-regulator and voltage reference circuits are basically unchanged. The sliding contact of R₃ is connected directly to the inverting input of A₄. R₅ is omitted along with C₄ and the compensation capacitor C₄ is increased to 1nF. R₉ and R₁₀ are changed to 43kΩ and 8.2kΩ respectively, and their junction taken to the non-inverting input of A₄.

This inversion of the inputs to A₄ is required because of the addition of an inverting current-boosting stage at the output of A₃, as shown. Darlington-connected series transistors Tr₆ and Tr₇ replace Tr₁ to Tr₅ and Tr₆ provides current limiting by adjustment of R₈. Typical components are Tr₁ 2N3055, Tr₅ 2N2102 Tr₂ 2N5294, R₅ 3.3kΩ, 1W R₉ 1kΩ, R₁₀ 10kΩ Tr₆ 2N5294, R₇ 3.3kΩ, 1W R₈ 1kΩ, R₁₁ 10kΩ, R₉ 10kΩ C₈ 100µF, C₉ 5µF

Full load regulation 0.05% Line regulation 0.01%/V

**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.
**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 R₁ and R₂. V₁ is approximately equal to V_{REF} because the differential input to the op-amp is very small, and hence, as R₁ = R₂, V_{OUT} = V_{REF}(R₁ + R₂)/R₁ 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 V₁ 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 50mA
- R₁, 7.87kΩ ±5%
- R₂, 7.15kΩ ±5%
- V_{REF} +12V, C₁ 100pF
- Ripple rejection 74dB
- Temp. coeff. of 0.003%/°C
- 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**

The schematic diagram of the LM104, Fig. 2, Current reference is temperature compensated. Output voltage programmed by value of R_{ADJ}. R_{1M} provides short-circuit protection. C₁ (tantalum) prevents oscillation.

- Output current: 25mA
- Input range: –50 to –8V
- Output range: 0 to 12V
- Typical load regulation: 0.05% from 0 to 25mA
- Typical line regulation: better than 0.2% for ±20% input change

**Negative voltage regulator**
- LM104, Fig. 2, Current reference is temperature compensated. Output voltage programmed by value of R_{ADJ}. R_{1M} provides short-circuit protection. C₁ (tantalum) prevents oscillation.
- Output current: 25mA
- Input range: –50 to –8V
- Output range: 0 to 12V
- Typical load regulation: 0.05% from 0 to 25mA
- Typical line regulation: better than 0.2% for ±20% input change

**Fig. 1**

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 R₁ and R₂ should be about 2kΩ
LM305A can provide 45mA.

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**Set 24: Voltage regulators—5**

**Set 24: Voltage regulators—6**
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
\[ V_{\text{OUT}} = V_{\text{REF}} R_2 / (R_1 + R_2) \]
For \( V_{\text{REF}} \) 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 \( V_{\text{REF}} = -15V \), \( R_1 3.65k\Omega \), \( R_2 11.5k\Omega \), \( R_3 3k\Omega \), \( R_4 2k\Omega \), \( R_5 2N4898 \).
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 \( R_6 \).

When the load current is large enough to cause the potential between \( C_1 \) and \( C_6 \) to turn on the related transistor, this removes drive current from \( T_3 \) to limit any further increase in output current. Curves above show typical load regulation and current limiting characteristics for \( V_{\text{OUT}} 5V \), \( R_6 10\Omega \), \( V_{\text{IN}} +12V \).

Further reading
\( \mu A 723 \) The Universal Voltage Regulator, Fairchild.

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 \( R_6 \). This should be 2.5kΩ for each volt required.
Input voltage range: 9.5 to 35V
Load regulation (1 to 100mA)

Line regulation: For a 10% change in \( V_{\text{IN}} \) 0.02% \( V_{\text{OUT}} \)
Load current: 100mA
An unbalanced output (+12V, -6V) suitable for comparators is obtainable with \( R_6 \) of 15kΩ and the addition of \( R_6 \) 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 \( R_6 \).
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.e. 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**
- $R_1 = 330\Omega$
- $R_2 = 1\,M\Omega$
- $R_3 = 10\,k\Omega$ potentiometer set for $V_o$ of 5V, typically
- $R_4 = 675\,\Omega$
- $R_5 = 2\,N5457$
- 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 $R_4$ is chosen so that $T_1$ is kept out of conduction when minimum output current is required. In the simplest form shown "$V_{ch}$" biasing is used that fixes the output voltage at $(R_4 / R_3 + 1) V_{ch}$ where the $V_{ch}$ 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 $R_4$ 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

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### Dual-polarity regulator

**Typical performance**
- $R_1 = 8.2\,k\Omega$, $R_2 = 3.9\,k\Omega$
- $R_3 = 2.2\,k\Omega$, $R_4 = 10\,k\Omega$
- $R_5 = 22\,k\Omega$, $C_1 = 47\mu F$
- $V_+ = \pm 15V$
- $V_0 = \pm 10V$, $I_o = 0\,\text{-}1\,A$

**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 $R_4$ 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 $TR_2$, the amplifier operating in the virtual earth mode with $R_4$, $IC_2$ defining the inverting gain. For $R_5 = R_3$ 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**
$IC_1$. 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 $T_1$ 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
$T_1$: 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. $T_2$: Junction f.e.t. n-channel. Pinch-off voltage < regulated output. Zero-bias on-current must be sufficient to drive $T_1$ into conduction—typically >2mA. $R_1$: 150 to 390Ω. If resistor is too high the minimum sink current of 1.3mA drives $T_1$ into conduction losing control at light loading. If $R_1$ is too low, insufficient drive current is available for $T_1$. $R_3$, $R_4$: 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 $[(R_3/R_4) + 1] \times 0.6V$. Stability is considerably increased by replacing $R_3$ with a zener diode when $V_s = V_2 + 0.6V$. $R_3$: 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 amplifier output to be out of saturation the base of $T_1$ is driven through a potential divider. Without this $T_1$ could not be driven off. The upper voltage limit is then the rating of the f.e.t. (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 $T_1$, $T_2$ must be chosen for a suitable voltage rating. An alternative zener circuit is shown in which $R_5$, $R_6$, $R_7$ set the output voltage $R_5/V_{SS}R_6$ in the absence of $R_5$.
- A Resistor to ground from either input causes current flow in either $R_4$ or $R_5$, and the resulting contribution to the output voltage has a temperature coefficient which can be used for overall temperature compensation. As shown $R_5$ contributes $-(1 + (R_4/R_5))V_+$ to the output.
- To remove the effect of supply variations via $R_4$, a diode network is chosen that ensures self-starting but has $D_2$ dropping out of conduction after starting has been achieved.

change in compensation is warranted. $T_3$, $T_4$: 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 heatsinking provided. $D_1$: Zener diode with low temperature coefficient for minimum drift. $D_2$: 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 $R_7$. $R_7$: Equal for precise tracking of outputs. 1 to 100kΩ. $R_8$: Minimizes offset if $R_7=R_8$. Can be omitted. $R_9$: 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. $R_{10}$: Set maximum base drive and hence, give coarse limiting of output 1 to 10kΩ. $C_{10}$: Suppress h.f. oscillation. Not critical but must be close to output or load inductance may initiate instability. $V_5$: Because amplifiers powered from regulated outputs, $V_5$ can be high if transistors have appropriate ratings. Increase $D_{11}$ volts 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 inverting 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 $R_8$, the resulting current bringing the output transistor into conduction when the p.d. across $R_8$ 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
Switching regulator

Typical performance
IC: CA3130 (RCA)
Tr, BFR81
D1, IN4148
L1, 680μH
C1, 15μF
R1, 1kΩ
R2, 470kΩ
C2, 680μF
VREF, 5V
R3, 50Ω
V8, 10V

Circuit description
Switching regulators are related to Class-D switching amplifiers. The power stage Tr1 conducts for a varying portion of the time. If the switching frequency is high, the current in L1 varies little throughout the cycle, with D4 sustaining the current in the load when the transistor is off. The inverting gain provided by Tr1 reverse the effective polarity of gain at the amplifier inputs; 100% negative feedback is applied from the load to one input and with L1 short circuit, a linear regulator would result were VREF to be fed directly to the other input. A small amount of hysteresis via R4, R5, combined with the L1,R1 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 (R1/R3)V8 and a mean value of VREF when the hysteresis is small. Power losses include those due to the speed of switching including core losses in L1, and the "d.c." losses such as V0(Eout) for Tr1 and the on voltage of D4. For low output voltages the latter limits the efficiency—between 70 and 90% is common even where the output voltage is < VOUT/2. As the supply voltage varies the mean current changes in the opposite sense because the

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

Typical performance
IC1, 555 timer
D1, D2, IN4148
C1, C2, 47μF, C3, 0.015μF
R1, 1.2kΩ, R2, 10kΩ,
R3, 22kΩ
+V8 = 10V
V0 = -10V
Io = 0 to -20mA

V0/V8 ±0.5% for V8 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 D1 to C4 to produce a negative output voltage. Assume ideal diodes, D1 clamps the right hand side of C4 to zero on positive output swings; similarly D4 clamps the right hand side of C3 to C4. On negative swings, C3 transfers charge via D3 into C4 as does C2 through C4. On negative swings, C1 transfers charge via D1 into C4 as does C2 through D4 into C4. Eventually C3, C4 each acquire a p.d. equal to the output swing, while C1, C2 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 R3, R4, and D3, D4 provides a potential at the RESET terminal such that each time the magnitude of the negative output increases, the oscillations is inhibited and the magnitude decreases. The diodes optimize the tracking for |V0| = V8.

Component changes
IC1: 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: This op-amp is particularly suitable for several reasons
(i)ero input resistance (m.o.s.) allows high R2, R3 ratio without R4 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 TR1 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: VRef must lie within input common-mode range; the output may not be able to swing high enough to switch TR1 off and a resistive network such as R3, R4 may be needed (left); current capability must be sufficient to saturate TR1 at max. load current—say I/20. TR1 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.

D1: Current rating mean output current; peak inverse voltage rating (p.i.v.) Vs; efficiency increased at low output voltages by reducing diode on-voltage (Schottky or germanium diodes if temperature not too high).

L1: Typically 20μH—10μH depending on current/frequency used. Ferrite cores reduce size provide high Q, low losses; saturation of core at higher currents can inhibit oscillations.

C1: Modifies frequencies for given L1, R4 combination. Not essential to operation, but reduces transients in load.

R1, R2: Ratio sets hysteresis and hence ripple. As ripple is reduced, so is time taken for completion of cycle i.e. frequency increases. By keeping R1, R2 as large as possible injection of switching current into VRef is minimized. Ratio R2/R4 typ. 100 to 1,000; high value gives low ripple provided increased frequency does not bring transient problems in.

R3: Not critical. 100Ω to 1kΩ with this op-amp.

VRef: Equal to required output.

R4: Load currents up to 200mA + possible. 

Vs: +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.

R1, R2 100 to 470Ω. Outputs to 1A.

Cross references
Set 6, card 7.
Set 24, card 11.
Set 7, card 12.

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most i.c. suppliers.

D1, D2: Not critical. Any fast silicon diodes.

C1: 470pF to 0.1μF. At low frequencies ripple increases and at high frequencies switching losses.

R1: 470 to 10kΩ. Too low a value wastes current.

R2: 2.2kΩ to 22kΩ. Select for frequency of oscillation. R3, R4: can be replaced by potentiometer for variable output. Value not critical.

Vs: +5V to +20V

V0: Up to 2V 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 D3, D7 can be omitted. With R3=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 R4 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.c.s are not applicable, and discrete transistor oscillator switches are required.

Further reading

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

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