Wireless World Circard

Series 6: Constant-current circuits

Why aren't constant-current circuits used as much as constant-voltage circuits, when for every constant-voltage circuit there is a constant-current dual? For the most part, the reason is probably historical, as the following article points out, and electrical sources have for a long time been constant-voltage. The possibility of using constant-current circuits in place of those designed for constant voltage is interesting. If, for instance, inductors were inconvenient in a certain constant-voltage circuit, the dual would allow capacitors to be used instead, given a suitable constant-current supply.

These cards then give circuits for constant-current supplies, both a.c. and d.c., and one card deals with their use for measuring characteristics of semiconductor devices (card 11, useful for electroplating too).

An extremely simple circuit is given on card 3 for limiting current in the range 1 to 35 mA, together with data obtained as a result of measuring 20 samples of 741 and 301 op-amps in various configurations. The limiting obtained is only approximate and more precise methods of fixing current are given on other cards. How to use voltage regulators in constant-current circuits is explained in card 2, the current-mirror in card 4 and transconductance amplifiers in card 12. The well-known ring-of-two circuit (a.c. card 5, d.c. card 6) is adapted for low-voltage use in card 9.

Constant-current Circuits

For every circuit using a constant-voltage element or sub-section there is a dual circuit based on constant-current properties. That such circuits are less common and often misunderstood is partly for historical reasons, stemming from the lack of sources of electrical energy having constant-current characteristics. One cannot draw out of stores a "5A 250mV-hour battery". A battery that will sustain a constant current into an arbitrary load is not physically realizable, since the electro-chemical processes involved define the e.m.f., the current then being inversely proportional to resistance.

Either capacitors or inductors may be used for temporary storage of energy, but the cost and size penalties of the latter are considerable. Capacitors store charge, having a p.d. proportional to the stored charge, and sustain that p.d. to a first order against varying current drain; until the drain results in a significant loss of charge. Even more important is that the generation, transmission and transformation of a.c. by the Electricity Boards are all constant-voltage processes. After rectification, the only form in which d.c. power can be produced efficiently and with freedom from ripple is as a constant voltage.

Thus all common sources of electrical energy approximate to constant-voltage characteristics and the majority of electronic circuits have been designed for this mode of operation. It is a fascinating thought that there should be as many current-operated circuits as voltage-operated ones, though they may be unfamiliar in shape. For example transistors would have to be operated in series, carrying comparable currents in each device but with progressively increasing p.ds moving from input to output in an amplifier, while the interstage coupling might be inductive. Conversely, where constant-voltage supply circuits use inductors to achieve particular effects, the corresponding circuits using capacitors would be attractive alternatives if constant-current supplies were available.

Each type of constant-current circuit seeks to achieve a constant current against variations in supply voltage, load resistance and ambient temperature as well as against component parameter changes. This will apply whether the supply is in the form of a direct voltage or an alternating current. In the former case, an intermediate step may be used in which the power is converted into a switched waveform before re-conversion to d.c., the method having high efficiency even when the load voltage is much less than the supply voltage (Fig.1). Alternatively the d.c. may be used to power an amplifier which because of the design of its output stage or by virtue of the feedback employed, delivers a current to the load controlled primarily by some signal voltage or current (Fig. 2).

Purely d.c. systems may also fall into this category with the a.c. signal replaced by a direct voltage/current (Fig.3) which can be fixed or variable depending on the application. (Fixed if a constant current is to be forced in a zener diode to define its operating point, variable where used to plot the characteristics of a transistor.) In addition, current control can be achieved by devising a two-terminal circuit to be interposed between source and load (Fig.4). If the circuit has a high dynamic resistance the current is then stabilized against supply and load changes. To apply such a circuit to a.c. supplies involves a number of difficulties, not least that such constant-current action is achieved only for that part of the cycle for which the input amplitude is in excess of some minimum value, typically 2 to 10V. It becomes particularly important to distinguish the parameter of the output whose constancy is being maintained. The peak value will be held constant by a two-terminal device having infinite slope resistance for amplitudes of input above the minimum. In many cases it may be necessary to rectify the applied voltage so that the ciruit deals with a single polarity.

Two further parameters of interest are the r.m.s. and mean-rectified output currents. For both, the rise in current during the pre-limiting region as the input voltage increases causes a rise in the area under the current graphs, i.e. in the mean/r.m.s. current. If the two-terminal network is arranged to have a negative-resistance characteristic then the current can fall back during input peaks. offsetting the tendency for the mean/r.m.s.

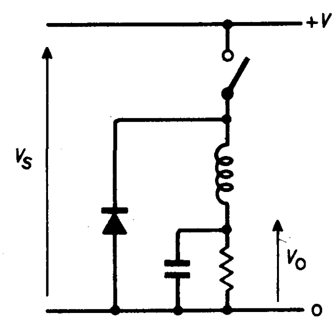

Fig.1. In trying to achieve a current constant against variations in supply voltage, load resistance, ambient temperature, and component parameters an intermediate step is sometimes used where power is switched before reconversion to d.c.

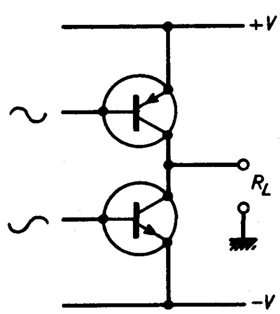

Fig.2. Another method of controlling load current.

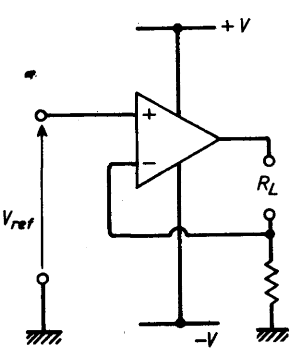

Fig.3. Technique of Fig.2 can be used with an a.c. signal replaced by a d.c. level.

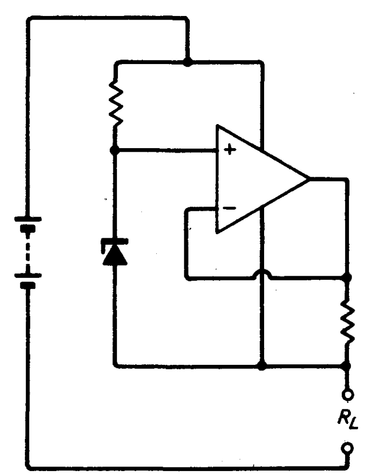

Fig.4. Current control can also be achieved by using Fig.3 as a two-terminal circuit between source and load.

currents to rise (Fig.5). A different value of negative resistance is required for the mean and r.m.s. conditions and it is further dependent on the input waveform. The method has the advantage that it operates on the instantaneous value of input, though methods based on thermistors and thermocouples might be used to monitor r.m.s. current via thermal effects. The necessary feedback would be more easily applied via conventional regulator circuitry and would involve thermal time delays that would not cope with input/load transients.

In the majority of these circuits the reference determining the current will be a voltage such as that developed across a zener diode. Where lower stability is adequate the p.d. across a forward-biased silicon p-n junction has advantages.

The voltage, or some function of it, appears across a resistor defining the current in that resistor. If the load is placed in series with that reference resistor, or in some other circuit path carrying a related current, then load current is fixed. Operational amplifiers have one output terminal committed to ground potential. If the generator representing the output has to appear in series with the reference resistor and load to define the current then a conflict appears (Fig.6). To achieve a current flow to ground the reference resistor (and with it the reference voltage and its associated circuitry) would have to float. As this is inconvenient, circuits based on conventional operational amplifiers may be limited to constant-

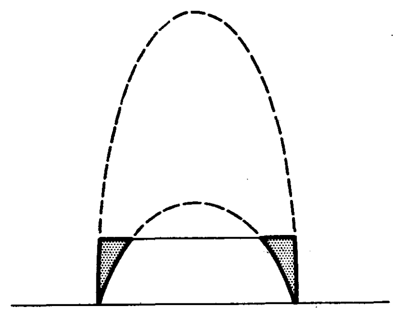

Fig.5. Negative-resistance causes current to fall back during input peaks, offsetting tendency for mean/r.m.s. currents to rise in the pre-limiting period.

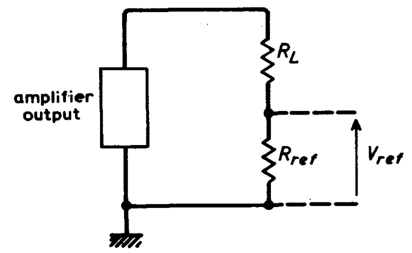

Fig.6. Circuits based on op-amps are generally limited to the case where loads do not require a ground connection.

current operation only with those loads not requiring a ground connection.

Circuit configurations are possible in which negative and positive feedback can be combined to raise the output resistance to very high values. Penalties include relatively poor stability of this output resistance and difficulties in achieving high output currents. This last demand is frequently met by adapting existing voltage regulators with a reference resistor at the normal output voltage terminals and the true load in series with it as outlined above.

Finally, the problem of controlling alternating current may be tackled in a different way by means of thyristor switches. These can be fired at appropriate points on the input waveform such that the mean current in the load is controllable. As the thyristor behaves as an almost perfect switch, no control is exercised over the instantaneous value of current. A filter provides a feedback voltage proportional to the mean current and controls the phase angle of the firing circuit. This phase angle control is quite distinct from the frequency/pulse-width modulation methods that are inherent in the switching amplifiers described earlier, and filtering of the output waveform would not normally be applied. The method would be suitable for such applications as battery charging where the current waveform is uncritical.

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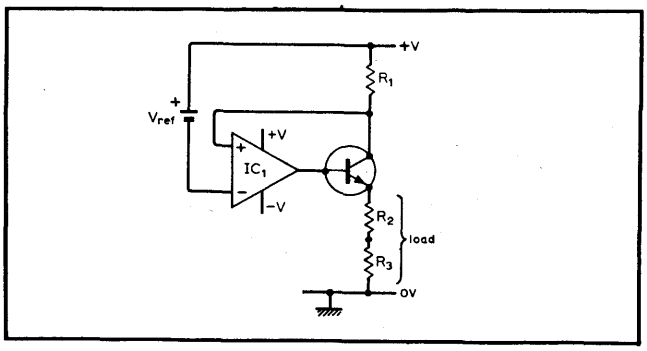
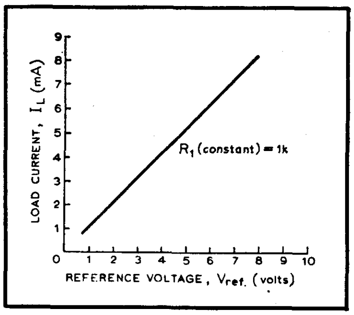
Hybrid constant-current circuit



Typical data

Supply: ±15V
IC_1: 741; Tr_1: BFR41
R_1: 1kΩ ±5%
R_2: 680Ω ±5%
R_3: 100Ω ±0.05%
Maximum current: 8mA.
Note: Careful layout required to avoid r.f. oscillation. Absolute measurement of current obtained using accurate digital voltmeter across R_3. Variation of load current with reference supply shown opposite.



Circuit description

This circuit permits high currents through the load ($R_2 + R_3$ in series), depending on the current capability of the bipolar transistor used. Negative feedback is applied via the operational-amplifier IC_1, the feedback being applied to the non-inverting terminal and being derived from the collector of transistor Tr_1, where inversion has occurred. Load current is essentially defined by V_{ref}/R_1, because the potential difference between inverting and non-inverting inputs of the operational amplifier when the gain is high, is very small. This reference voltage, symbolised by an ideal battery, may simply be a reverse biased zener diode in series with a resistor connected across the d.c. supply, the inverting input being connected to the junction. This has the disadvantage of being uncompensated for temperature variations. If the zener diode has a positive temperature coefficient, this can be offset by connecting a forward-biased silicon diode with a negative temperature coefficient in series. Such a combination is available in a single package to provide a temperature-compensated zener diode.

If the current through R_1 increases, the potential difference across R_1 increases, and the voltage applied at the non-inverting terminal decreases. This change is amplified by the operational-amplifier, and hence the base drive to Tr_1 is reduced, tending to compensate the original increase of the collector current which is approximately equal to the load current. As the gain of IC_1 is high, the input current demanded by this operational amplifier is extremely small, and the feedback also increases the effective output impedance of Tr_1.

Component changes

- IC_1: LM101, Tr_1: 2N2219, Tr_2: 2N3456.
- Supply: useful range down to ±9V. Typically variations of current better than 0.05% over this range, when V_{ref} is independent of the supply.
- If oscillation exists, connect a capacitor across R_1.
- Useful range of R_2: 330Ω to 3.3kΩ. At 2mA load current variations less than 0.05%.
- At 2mA, variations are less than −2% with BFR41 h_{fe} in the range 90 to 220.
- Absolute measurement of current through R_1 and emitter current indicated a variation of around 1.5%.

Circuit modifications

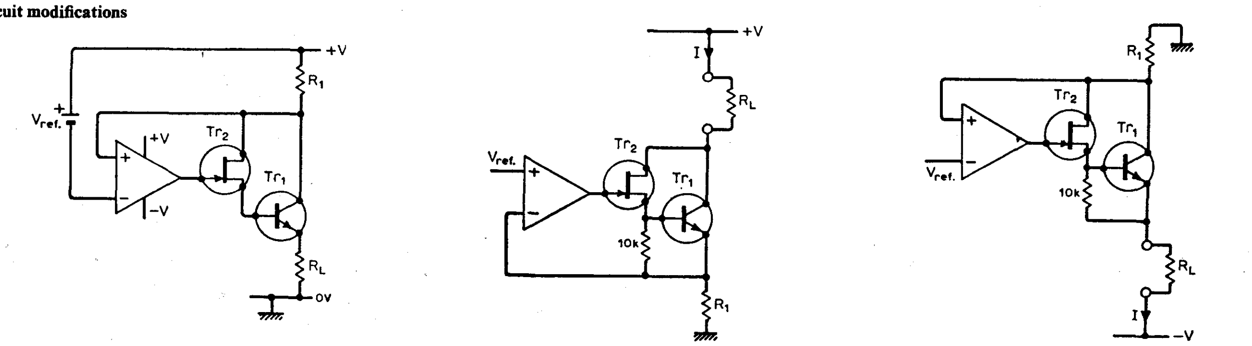
Current through R_1 is defined by V_{ref} in circuit shown left. However in this circuit, the current shunted from the collector to the non-inverting input of the operational-amplifier is considerably less than the original circuit, as the output current demanded from the op-amp is only the gate current of the f.e.t. Tr_2. The f.e.t.-bipolar compound pair has a much higher current gain and the load current is more nearly equal to that defined by V_{ref}/R_1.

- Use f.e.t. 2N5457 to drive the bipolar transistor. Absolute measurement of current through R_1 and emitter currents of Tr_1 now indicate a variation of less than 1%. R_1: 1.1kΩ, R_3: 100. V_{ref} adjusted to give load current of 2mA. R_2 varied from 4.7kΩ (max) down to 10. Current change within 0.01%.
- Alternative arrangement of feedback connection shown centre and right. Circuit in centre uses the output stage as a non-inverting follower allowing feedback to be returned to the inverting terminal of the op-amp. This arrangement is sometimes known as a current sink. Circuit right shows the corresponding current source. This may have both the reference voltage and reference resistor returned to ground or the positive supply rail with the load returned to the negative rail for increased load potential difference.

Further reading

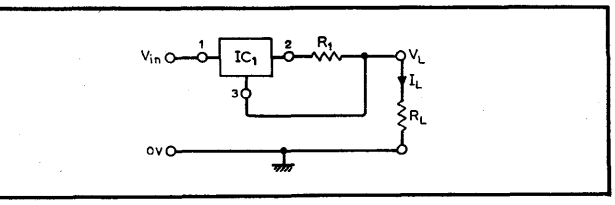
National Semiconductor Linear Applications AN-20.
Hart, B., Current generators using unipolar-bipolar transistor hybrids. *Electronic Applications* 1966 vol. 27, pp. 30-7.
Silicon Zener Diode and Rectifier Handbook, Motorola.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 2

Constant-current use of voltage regulators



Typical performance

V_{IN}: +15V
V_L: 3V
IC: LM309H
R_1: 245Ω ±1%
R_L: 120Ω ±5%
I_L: 25mA

For 1.5V pk-pk input ripple at 100Hz, load current ripple is approx. 16μA pk-pk.
Dynamic output res: 90kΩ
Dynamic to static output resistance ratio: ≈220
For 25mA < I_L 200mA, I_L changes less than 3% for a 100% increase in R_L.

Circuit description

A very simple constant-current generator can be produced by placing a sufficiently large resistance between a constant voltage source and a load. This leads to a requirement of very high source voltages to supply constant currents of only a few mA. This simple approach is normally unacceptable. However, a constant-voltage regulator can be made to provide a constant current into a load, at reasonable voltages, while only carrying a relatively small standing current. The diagram above shows a monolithic voltage regulator connected as a two-terminal constant-current generator. This regulator was designed primarily as a fixed 5-V voltage regulator to supply the widely varying currents in logic circuitry. In the constant-voltage mode, R_1 would be set to zero and terminal 3 connected to ground instead of the output terminal. The circuit thus provides a regulated output voltage between terminals 2 and 3. Inclusion of R_1 between these terminals as shown ensures that it receives a constant voltage from the regulator and therefore carries a constant current which is supplied to the load resistance. (The stability of R_1 determines the stability of I_L.) The load will also carry the quiescent current from terminal 3 but this will normally be much smaller than the current in R_1. This quiescent current places a lower limit on the available output constant current. The voltage regulator chip incorporates a temperature regulator to provide thermal, rather than current, protection. This technique allows a considerable increase in the maximum allowable output current, the device being protected against almost any overload condition.

Component changes

Useful range of V_{IN} + 6 to +35V.
I_L (min) ≈10mA: lower limitation due to quiescent current at regulator terminal 3.
I_L (max) ≈200mA: power dissipation limitation of 2W in regulator without heat sink.
For L_L values of 50, 100 and 200mA typical values of R_1 with V_L = 3V are 109, 51.35 and 25.2Ω respectively.
If regulator is placed some distance from the d.c. supply filter, a capacitor of about 0.1μF may be required between terminal 1 and ground to prevent h.f. oscillation.
For higher output currents, up to about 1A, the LM309H can be replaced by an LM309K.

Circuit modifications

Any voltage regulator that can sustain a constant load voltage at a high current compared with its standing current may be used as a constant-current generator. Circuit shown left is a standard form of voltage regulator using Tr_1 and Tr_2 as a long-tailed pair with Tr_3 and Tr_4 forming a Darlington-connected output transistor. The long-tailed pair compares the reference voltage from the zener diode with the output voltage across a dummy load R_2. If the voltage regulation is good and R_2 is constant then the current in it is constant. The current in the real load R_L is this current plus the currents in the long-tailed pair and reference diode, both of which can be made very much less than the dummy load current. If the "free" collector of Tr_3 and Tr_4 is accessible in the voltage regulator, R_1 may be placed between it and the positive supply, although R_L will not then be referred to ground.
Another floating-load constant-current generator is shown, middle, which applies the principle of series feedback. The p.d. across R_3 is a defined constant voltage and so also is the current in it. This current is virtually identical with that flowing in R_L. Amplifier could be a Darlington-connected pair.
Existing voltage regulators, even of the poorest kind, can be used to provide a constant current, one example being shown right. The zener diode fixed the p.d. across the emitter resistor R_4 and hence the current in R_L. This circuit suffers from the usual problems of matching up the temperature coefficients of the zener diode and transistor.

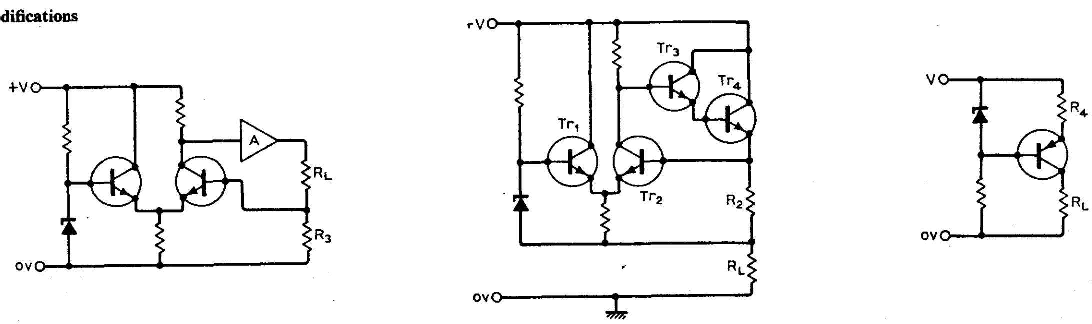
Further reading

Linear Applications Handbook, National Semiconductor, AN 42-1 to 42-6, 1972.
Nowicki, J. R., Power Supplies for Electronic Equipment, vol. 2, chapter 1, Leonard Hill Books 1971.

Cross references

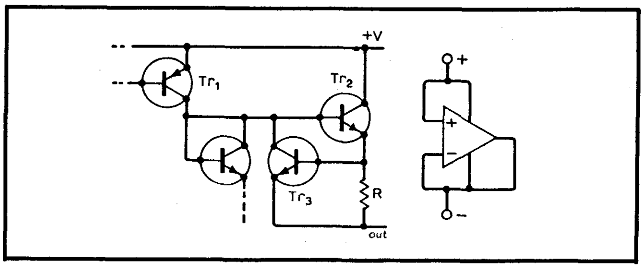
Series 6, cards 13, 10 & 11.

Circuit modifications



Wireless World Circard Series 6: Constant-current circuits 3

Simple current limiting circuits



Typical performance

IC: N574IV (Signetics)
Supply voltage: 10V
Current: 26-30mA
Voltage for limiting: 7-9V
10 samples of other manufacturers 741/748 i.cs gave current range 20-35mA.
10 samples from three manufacturers 301 i.cs gave currents of 15-25mA, but included devices requiring only 2-3V to achieve limiting.

Circuit description

Many i.c. op-amps have protection circuits at their output which limit the current that can flow, even into a short-circuit of the output to either supply line, and regardless of the condition of the input terminals. The current is not defined as precisely as with the other constant-current circuits described on these cards; the limiting action is only intended to be approximate, and generally uses the base-emitter junction of a transistor as the sensing element (e.g. with Tr_3 as in a section of an i.c. shown above). Transistor Tr_2 is one of the output transistors and if the output current flowing through R tries to exceed the value at which the V_{be} of Tr_3 reaches 0.5V, Tr_3 comes into conduction, diverting the base current supplied by Tr_1 and preventing further increase in output. In general, the limit current falls with increasing temperature because the V_{be} of Tr_3 required for conduction falls, and the resistance of R increases with temperature. Such a mechanism is thus not adequate for precision constant current action but can offer good rejection of supply variation including ripple. If an i.c. op-amp having such limiting has its output shorted to one supply line and the inputs connected to the supply lines, in the sense that causes the output to try to drive towards the opposite line, the limiting mechanism comes into play and the complete circuit may be used as a two-terminal device. Placed between source and load, the load current is limited typically to 12-30mA depending on amplifier type for any p.d. across the amplifier above some minimum voltage (5-9V). The max p.d. across the amplifier must not allow the device dissipation to be exceeded, though self-heating minimizes the dissipation by reducing the current.

Component changes

- With output open-circuit the circuit may also draw constant current but of much smaller magnitude. Similarly, connecting output to opposite supply and/or reversing input terminals brings different sections of the circuit into action, i.e. several different current limits can be obtained.

- With typical device from N5741V range, six configurations were tested, as below, with minimum voltage of 8V throughout; tests carried out at 10V and resulting current limits from 0.85 to 30mA obtainable from single device:

inv. input	non-inv. input	output	current (mA) at 10V
+	−	+	30
−	+	−	29
+	−	−	12
+	−	o/c	1.4
−	+	+	0.9
−	+	o/c	0.85

Current reduction 20% for temperature increase of 50 deg C.

Circuit modifications

- The basic idea of using a transistor to monitor the p.d across a current-carrying resistor is also applied in voltage regulators to limit the output current even into a short-circuit load. Here, Tr_3 deprives Tr_2 of base-current, monitoring the p.d. across an external resistor R_2. This allows boosting of the output current via external transistor Tr_1, a variable R giving control of the current limit.

- Limiting by sensing of the collector current of the output stage is also possible. The nature of the drive circuit is often such that a loss of, say, 1V in the collector circuit does not further increase the minimum supply voltage. As shown, Tr_1 is a constant-current stage biased by D_2 acting as a high-impedance load for the error amplifier (not shown). As the output current increases so does the p.d. across R_2 bringing D_1 into conduction and diverting current from Tr_1 i.e. limiting base current of Tr_2.

- In principle simple limiting circuits may be added to any voltage regulator. Shown is a method by which base current is diverted from the series pass transistor by Tr_3 which senses the p.d. across R_4. In this case it is the total current that is limited i.e. load current plus circuit quiescent current.

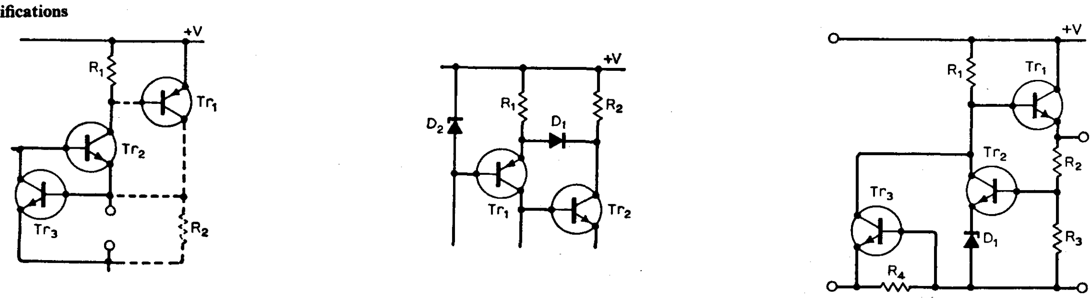
Further reading

Patchett, E. N., Automatic Voltage Regulators & Stabilizers, Pitman 1970, pp. 364-6.

Cross references

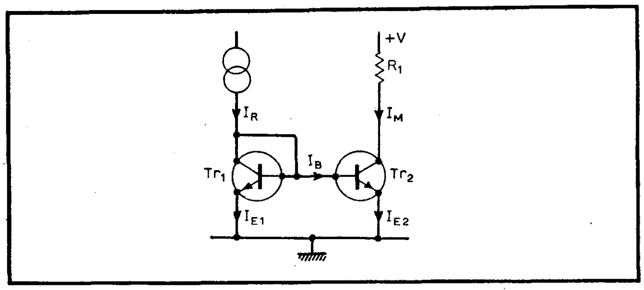
Series 6, cards 2, 5 & 9.

Circuit modifications



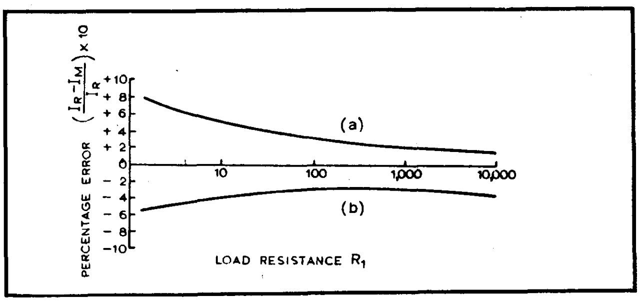
Wireless World Circard Series 6: Constant-current circuits 4

Current mirror



Typical performance

Supply: +6V
Tr_1, Tr_2, part of CA3046
R_1: 0-10kΩ decade resistor, $\pm 0.05\%$
I_R: 0-5mA from commercial current generator, $\pm 0.05\%$
I_M: Calculated from voltage reading across R_1 using five-digit voltmeter
Dynamic output impedance: 2MΩ at 50μA
Curves below show percentage variation of 'mirror' current to reference current for the basic (a) enhanced (b) current mirror circuits, for currents in the range, 1μA to 5mA.
Product of $I_M R_1$ maintained constant.



Circuit description

Circuit configuration is known as a 'current mirror' and is widely used in integrated circuits. If the two transistors Tr_1 and Tr_2 are considered identical so that the base-emitter voltages are the same, then to a first order the collector currents will be the same. Transistor Tr_1 acts as a diode whose forward voltage between base and emitter defines the base-emitter voltage of transistor Tr_2. If Tr_2 has a high current gain, then the reference current I_R will be approximately equal to the collector 'mirror' current I_M.

$$I_R = I_B + I_E = I_{E2}/(1+\beta) + I_{E1} = I_{E2}\left(\frac{1}{1+\beta}+1\right)$$

$$I_M = aI_{E2} = \frac{\beta I_{E2}}{1+\beta}$$

$$I_{E2} = I_R \cdot \frac{1+\beta}{2+\beta};\ \therefore I_M = \frac{\beta}{1+\beta} \cdot I_R \cdot \frac{1+\beta}{2+\beta} \approx I_R.$$

Hence if the reference current is fixed, the collector current of Tr_2 is fixed.

Discrete components are temperature sensitive and the circuit is not reliable with them. Closer matching of the transistor parameters and the facility of compensating changes due to temperature are available, when the transistors are produced on the same monolithic silicon chip. The circuit is thus often used in the reference stage for basic regulator circuits Output impedance is approximately that of a common-emitter configuration; the effective resistance connected across base and emitter is the low dynamic resistance of Tr_1 connected as a diode.

Output resistance characteristic of this circuit is increased considerably by including a diode connected transistor in series with the emitter of Tr_2 as shown below (middle).

Component changes

- Dynamic output impedance reduces to 200kΩ for a current of 500μA, and 90kΩ for load current of 1mA.
- Percentage mirror current error is typically better than 2.5% for I_M = 500μA when R_1 is varied from 0-10kΩ without attempting to maintain V_{ce} of Tr_2 constant.

Circuit modifications

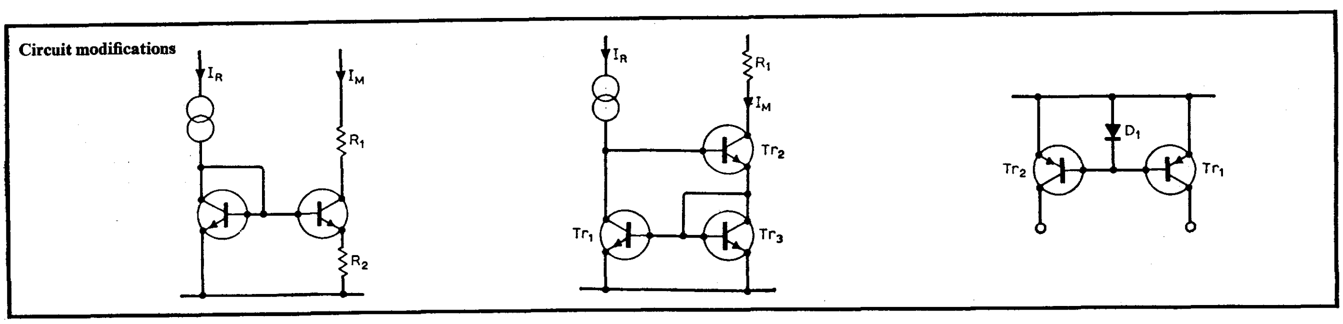
- Output impedance of the current mirror is increased by negative feedback via resistor R_2 (left) but its use should be restricted to currents in the microamp range.
- Higher output impedance obtained using the enhanced circuit shown middle. This requires about 1.2V minimum before control commences as the V_{be} of Tr_2 and Tr_3 must be overcome. The resulting transfer ratio of I_M/I_R can be shown to be $(\beta^2 + 2\beta)/(\beta^2 + 2\beta + 2)$ indicating an improvement dependent on the β^2 term, the $(2\beta + 2)$ term becoming insignificant for high-gain transistors.
- Current mirror, shown right, available within transistor package CA3084. This is a p-n-p version and illustrates the use of the current mirror in establishing multiple current sources. Diode D_1 is a transistor with its base and collector connected. The V_{be} values for each transistor are identical, and hence control of D_1 current ensures first-order constancy of currents in Tr_1 and Tr_2. In practice, the increased number of units of base current degrade the stability if too many stages are controlled.

Further reading

Hart, B. L., Current generators, *Wireless World*, vol. 76, 1970, pp. 511-4.
RCA Solid-State Databook, Series SSD-201, 1973.
RCA Solid-State Databook, Series SSD-202A, 1973, pp. 325-76.

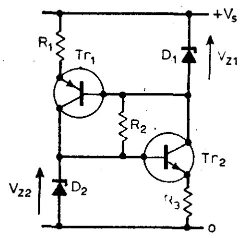
Cross references

Series 6, cards 5 & 12.



Wireless World Circard Series 6: Constant-current circuits 5

Ring-of-two reference



Typical performance

Minimum terminal p.d. $\sim V_{Z2} + V_{Z1} - 0.5\text{V}$

Constant current $\sim \dfrac{V_{Z2} - 0.6}{R_3} + \dfrac{V_{Z1} - 0.6}{R_1}$

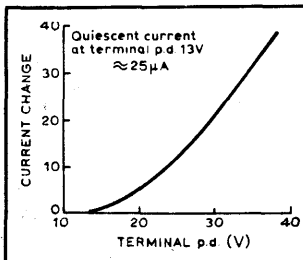
Tr_1: 2N3702

Tr_2: 2N3707

R_1, R_3: 470kΩ; R_2: ∞

D_1, D_2: Reverse biased base-emitter junction at planar transistor e.g. 2S512

Comparable results for currents up to several mA. Self-heating effects significant at higher current.



Circuit description

The ready availability of two-terminal elements which can be placed in parallel with a load to make the load voltage stable is not matched by dual elements for sustaining constant load currents. Constant-current diodes are available but are no match for the variety and performance provided by zener diodes. Two problems have to be overcome in designing a two-terminal constant-current circuit. There will usually be two or more separate paths for current flow and they must either be separately constant or, if variable, such variations must be restricted to a low-current path. A second problem is that the minimum p.d. at which constant-current is achieved must be as low as possible, while the breakdown voltage should be high. The ratio of these p.ds is one guide to the usefulness of the circuit and a ratio of 10:1 or greater is good. The upper voltage is fixed in the present circuit by the V_{cb} breakdown of the transistors and the lower voltage by the sum of the V_z values. The two current paths are separately constant and may be made equal or not as required. Diode D_2 maintains a constant potential at the base of Tr_2 and hence a constant p.d. across R_3 ($V_z - V_{be}$). The resulting constant emitter current ensures that the collector current of Tr_2 and hence the current in D_1 are also constant. Similarly the p.d. across R_1 is defined ensuring the stability of current in D_2. Thus each diode defines the current flowing in the other. The circuit is a form of complementary bistable and precautions must be taken to ensure that the on-state is the only practical one. This may be achieved by a starting resistor R_2 between the bases (or from Tr_2 base to +ve line for example).

Component changes

Tr_1, Tr_2: General purpose silicon e.g. n-p-n. types ME4103, 2N706, BFR41; p-n-p types 2N3702, ME0413, BFR81.

D_1, D_2: Zener diodes 2.7 to 12V. Low voltage units (2.7 to 4.7V) give minimum terminal p.d. and first-order compensation for V_{be} temperature drift. Higher voltage units increase dynamic resistance of circuit. Zeners of breakdown $\approx$6V have low temp. drift, and additional forward-biased diode in series gives temp. comp. (For very low voltage operation see card 9). Diodes need not have equal breakdown voltage. For low currents reverse breakdown in planar transistor base-emitter junctions offers good performance.

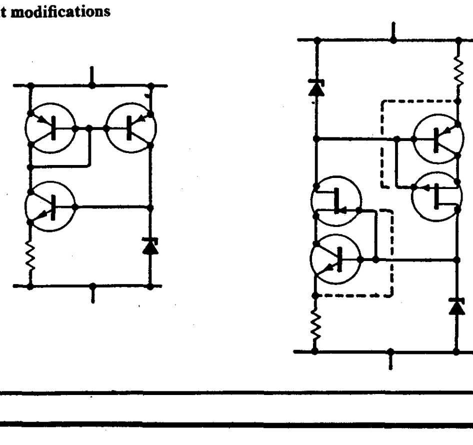
R_1, R_3: 330kΩ 1MΩ. At higher currents, self-heating effects vary current as terminal p.d. changes. At lower currents, low-leakage transistors used for Tr_1, Tr_2. Zeners may be replaced by reverse-biased base-emitter junctions of planar transistors (breakdown voltages typically 5 to 10V, fairly close tolerance for given device type).

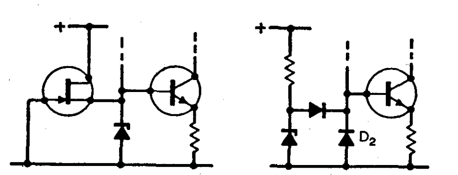
R_2: Typically 330kΩ to 10MΩ. Use highest value that ensures self-starting. 1MΩ adequate with all except high leakage zeners.

Circuit modifications

- To minimize the p.d. at which the circuit achieves constant-current operation, only one half of the circuit has a zener diode. The other half may have the zener replaced by any other element that sustains an approx. constant, p.d. against variation in current. A current mirror in one of its forms allows the circuit to function correctly for a terminal p.d. barely more than the zener voltage. Alternative circuits (card 4) can increase accuracy of current for small increase in minimum p.d.
- For highest dynamic resistance, each transistor may be replaced by cascode or similar circuits while retaining defined V_{be} characteristics of bipolar transistors. Alternative connection for f.e.t. gives higher dynamic resistance but version shown allows f.e.t. to operate with slight forward bias if required, increasing the current capability.
- Circuits are all bistable in form, with a possible non-conducting state. Any resistive start-up circuit degrades dynamic resistance. Use of junction f.e.t. with pinch off between V_{be} and V_z inhibits off-state without contributing current in one state. Identical zener diode with high resistance drive brings D_2 into conduction-preferred method in some i.c. regulators but current in R flows in load if used as two-terminal constant-current circuit.

Circuit modifications





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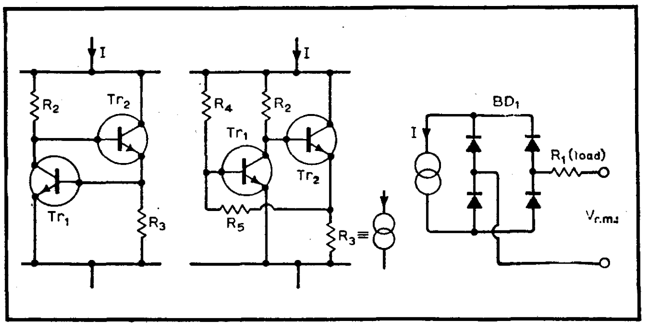
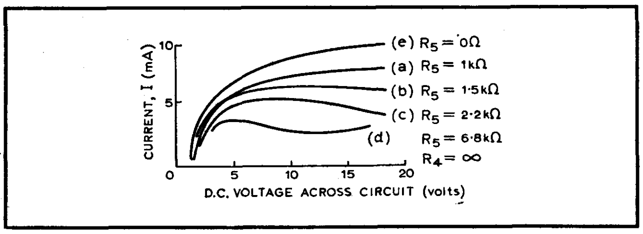
A.C. constant-current circuits



Typical performance

Tr_1, Tr_2: BC125
BD_1: A154
R_1: 47Ω; R_2: 3.9kΩ
R_3: 100Ω; R_4: 47kΩ
R_5: 1.5kΩ
Current constant at 5.8mA ±1% for direct voltage of 6 to 18V.



Circuit description

If a circuit can maintain a constant voltage across a resistor against changes in the supply voltage, then the current flow in this resistor is maintained constant. If this current is greater than any other current in the circuit, then the total current taken from the supply is reasonably constant. A simple circuit that attempts this has the base-emitter of Tr_1 in parallel with a 100Ω resistor R_3, maintaining a current through R_3 of about 6mA with the feedback loop closed via Tr_2. Although the current in Tr_1 varies when the applied voltage varies, this current is appreciably less than that in Tr_2, and so the dynamic impedance of the circuit used as a two terminal element is high. A more complex amplifier, e.g. a Darlington pair, in place of Tr_2 would allow the contribution to total current change, due to the current in R_2, to be very small.
An alternative arrangement is to introduce R_4 and R_5. If supply voltage increases, this potential divider increases p.d. across R_5. The base potential of Tr_1 is substantially constant, and hence p.d. across R_3 must fall, and hence the current i.e. a relatively large increase in the current in R_2 (which is small) is balanced by a small decrease in the relatively large current through R_3. By suitable choice of R_4, R_5, the dynamic resistance can be controlled to be positive or negative, and with a critical value of R_5 is extremely high over a wide range of supply voltage. The operation of the circuit below 5V is non-linear.
When a.c. is to be applied, it may first be rectified so that the circuit sees a unidirectional voltage, but only the peak current can be controlled i.e. currents corresponding to voltages in excess of 5V. To control the r.m.s. value of current, and if the waveshape is unimportant, the negative resistance effect allows the current to fall during the peaks of the applied signal, compensating for the rise during the rest of the cycle. Adjustment is empirical and depends on waveshape, but offers a simple means of controlling current in a resistive load for heating, or the mean charging current in the battery.

Circuit modifications

- A high current gain in the output stage of the simple circuit, allows the bias current to be very small (left) and is therefore also suitable for high current circuits. Also Tr_1 had to act as both an error amplifier and reference against which the current is being compared i.e. the V_{be} of the transistor. To improve this, a zener diode may be added as reference with the transistor primarily performing the function of error amplifier.
- The bias current itself may be made constant if resistors are replaced by elements which are two-terminal constant-current devices (e.g. f.e.t.) which may itself be combined with a better amplifier such as an op-amp, to give improved overall stability.
- The control of alternating currents is possible where devices are available which may be made to directly accept signals of both polarities (right). One practical case is a junction f.e.t. in which a resistor-diode network attached to the gate allows interchangeability of source and drain, e.g. when supply to A is positive, D_2 conducts, clamping gate close to source voltage, and Tr_1 current is near maximum value and unvariant with respect to further increase in supply, As f.e.t s have great variation in pinch-off voltage and 'on' current, equal resistors are connected into source and drain paths, to exercise control over the current.

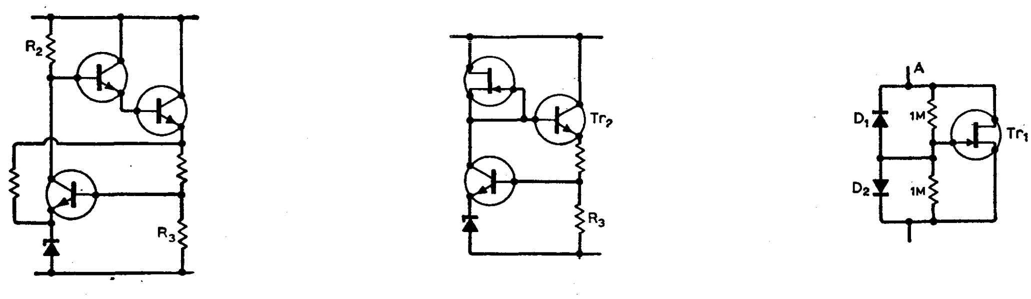
Further reading

Williams, D. A., High-voltage constant-current source, *Wireless World*, Jan. 1972, p. 29/30.
Watson, G. Two transistors equal one, *Electronics*, 6 July, 1962.

Cross references

Series 6, cards 9 & 12.

Circuit modifications



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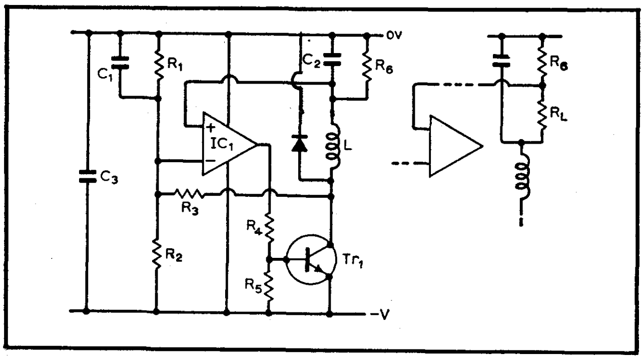
Switching current regulator



Typical performance

IC: 301
Tr: TIP3055
D: 1A 25V diode
C_1: 1nF
C_2: 22μF 6.3V Tantalum
C_3: 22μF 20V Tantalum
R_1: 1kΩ; R_2: 5.6kΩ
R_3: 470kΩ; R_4: 220Ω
R_5: 150Ω
L: 5mH (Ferrite core)*
For $R_6=2\Omega$, $V_s=-10$V
load voltage: 1.2V,
supply current: 150mA
switching frequency: 4kHz
*See component changes
ripple voltage: 100mV
stability: output change $< \pm 4\%$ for supply 5 to 20V
output change $< \pm 1\%$ for load resistance 2 to 15Ω.

Circuit description

The key difference between switching regulators and conventional types lies in the discontinuous operation of power stage which is isolated from the load by an LC network. The power transistor delivers current for short periods to the inductor and during its non-conducting period the current flow in the inductor is sustained through the diode. The resulting voltage step across the inductor (approximately equal to the supply voltage) defines the rate-of-change of current in terms of the inductance. If the period is short enough, the current is relatively constant, and together with the filtering action provided by the capacitor, the ripple voltage across R_1 can be small compared with its mean p.d. The circuit may be alternatively viewed as a simple astable in which the inversion due to the output transistor interchanges the functions of the op.amp. input terminals, while an LR circuit replaces the conventional CR version. Hysteresis provided by R_3 defines the pk-pk swing that will occur across R_6. The smaller this hysteresis, i.e. the larger R_3, the smaller the resulting ripple. This brings with it increased frequency of operation, as the rate-of-change of voltage is a function of L, C_2, R_6 as outlined above. Mean level across R_6 is fixed by that across R_1 and is a fixed fraction of the supply voltage. In most applications this potential divider is replaced by stable reference voltage of suitable value (see cards 5, 9). As shown, the circuit acts as a voltage regulator for a load at R_6. To be used as a constant-current source the load may be placed series with the resistor across which constant p.d. is developed. Switching regulators may be driven by an external oscillator with the internal positive feedback eliminated.

Component changes

L: Frequency of operation is a compromise; too high and amplifier switching times limit performance, too low and increased inductance brings reduced efficiency because of winding resistance. Coils wound on ferrite rings/cores offer wide range of operating frequencies with minimum radiation of switching harmonics if shielded units used. Typical range 200μH to 10mH.
IC_1: Uncompensated op.amp. 748, etc. Possibility of 741,301 compensated amplifiers at low frequency with suitable choice of ferrite.
Tr_1: For currents < 500mA: BFR41, BFY50 with reduced efficiency; somewhat higher frequencies at moderate currents: MJE521.
R_1, R_2, R_3: Set reference voltage/hysteresis. R_1, R_2 replaced normally by separate reference circuit.

Circuit modifications

- To stabilize load voltage/current some stable reference voltage must be added. A simple circuit that allows operation down to very low supply voltages, tolerates high voltages and gives reasonable stability against temperature changes, matches the V_{be} characteristics of a silicon against a germanium transistor. Unselected units give a variation in reference voltage against supply of <2% over the whole supply range of the regulator (e.g. 3 to 20V), and a typical temperature drift of <0.1% per deg. C.
- Output current can be increased by replacing the drive transistor by any high gain combination such as the Darlington pair provided frequency is not too high (charge storage problems) and the increased losses due to saturation are acceptable. At low supply voltages the collector of the first transistor may be returned to the zero line.
- A positive voltage regulator using a standard i.c. is given in the first reference below. It operates at a higher switching frequency and contains its own voltage reference circuit. Pin 6 compares a portion of the output voltage with the internal reference, the error amplifier driving the transistor with positive feedback via pin 6 and defining the hysteresis.

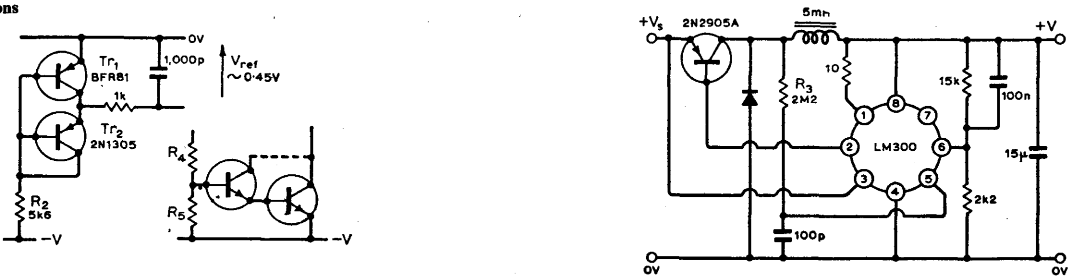
Further reading

Designing Switching Regulators, National Semiconductor application note AN-2, 1969.
Nowicki, J. R., Power Supplies for Electronic Equipment, vol. 2, Leonard Hill, 1971, pp. 153-81.

Cross references

Series 6, cards 8 & 10.

Circuit modifications



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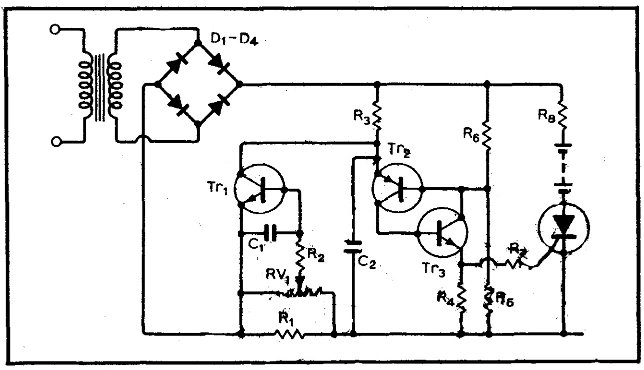
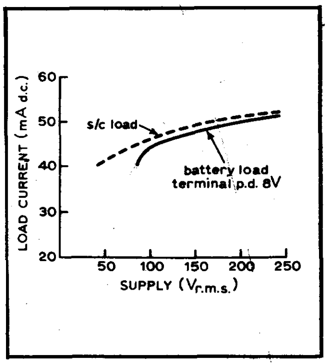
Thyristor control current regulator



Typical performance

T: 240V r.m.s. 50Hz primary
30V r.m.s. secondary
D_1 to D_4: 50V 1A bridge rectifier
Tr_1, Tr_3: BC 125
Tr_2: BC 126
Tr_4: 50V 1A (mean d.c.) thyristor (2N1595 etc)
R_1: 12Ω; R_2, R_5, R_6: 10kΩ
R_3: 150kΩ; R_4: 470Ω
R_7: 100Ω; R_8: 15Ω
C_1: 470μF; C_2: 22nF
Supply: 200V r.m.s.
Battery terminal p.d.: 8V
Charging current set to: 50mA (mean)
Change in current for supply voltage ±25% ≈ ±4%
Change in current for terminal p.d. changed by ±2V ≈ ±0.5%



Circuit description

The circuit consists of four sections: a full-wave bridge-rectified power supply; a thyristor in series with the load with the angle of conduction varying the mean load current; a pulse-generating circuit which delivers a series of pulses to the thyristor starting at a particular instant in each half-cycle and a current-sensing transistor that varies the pulsing circuit to control the mean current via the firing angle. Once the thyristor has fired, the remainder of the circuitry has no influence on the instantaneous current (determined only by the elements in series across the supply: R_1 thyristor, load, R_8). Any increase in the mean current causes the mean p.d across R_1 to increase and via ~~RV_1, smoothed by~~ R_2, ~~C_1~~, bring Tr_1 into conduction. This by-passes some charging current from C_2 delaying the onset of firing of the unijunction equivalent composed of Tr_2, Tr_3, R_5, R_6 (see Series 3, card 4) The minimum p.d. wasted across current-sensing resistor R_1 need only be ≈0.6V, giving good efficiency. Accuracy of control is limited by relatively low gain of control element, its temperature dependence, etc. Adding a zener diode in emitter of Tr_1 and dispensing with RV_1 would define control point more accurately at expense of increased voltage/dissipation in R_1.

Component changes

T, D_1 to D_4: Diodes must carry peak current much greater than mean current where conduction angles are small (high supply voltages, low load voltages) i.e. if mean load current is to be 1A peak currents might have to be >5A. Similarly for transformer, thyristor.

R_1: At max. setting of RV_1, mean voltage across R_1 is 0.6V approx. and mean current = 0.6V/R_1. Setting RV_1 to 50% doubles mean current, and p.d. across R_1, quadrupling power in R_1.
C_1: Smooths bias to Tr_1, 50 to 1000μF low-voltage electrolytic.
R_2: Increased value allows lower C_1 for given smoothing but decreases accuracy of current. Typical range: 2.2 to 47kΩ.
R_3, C_2: To give free running frequency ≫ 100Hz so that firing can occur early in each cycle. R_3: 47 to 470kΩ; C_2: 10 to 100nF.
Tr_2, Tr_3, R_5, R_6: Can be replaced by single unijunction transistor e.g. 2N2646, 2N2160, etc. Any other general-purpose silicon transistors in place of Tr_2, Tr_3.
R_4, R_7: Reduce R_4 to 100 for some unijunctions. R_7 not critical.
Thyristor: Any medium sensitivity, low-voltage thyristor. For higher peak currents reduce R_1, R_8 proportionately. Resistor R_8 can be omitted if very high peaks can be tolerated by thyristor, load.

Circuit modifications

- The supply to the sensing/firing circuits may be limited and/or stabilized by a zener diode to improve control over the firing point, and to protect the circuitry when the thyristor supply is too great. For example, this would be necessary if constant-current action were desired directly from mains with no intervening transformer. Dissipation in R_5 would be high. In this, as in main diagram, a unijunction may be substituted for the complementary bistable.
- Where the circuit is to be used for battery charging, over-voltage protection might be desired. One possibility is to monitor the battery voltage directly (or better via an RC filter to eliminate spikes, as with R_2, C_1 over) using a zener diode or other suitable reference to define onset of conduction in Tr_4. The latter can then be used to raise the potential at the junction of R_5, R_6, delaying and eventually preventing firing. Addition of a series resistor R_9 to the junction of Tr_2 base/Tr_3 collector prevents excessive current flow via Tr_4, Tr_3.
- Alternative coupling methods including pulse transformation, light-emitting diodes, etc., may be used if thyristor is at an inconvenient potential relative to firing circuit.

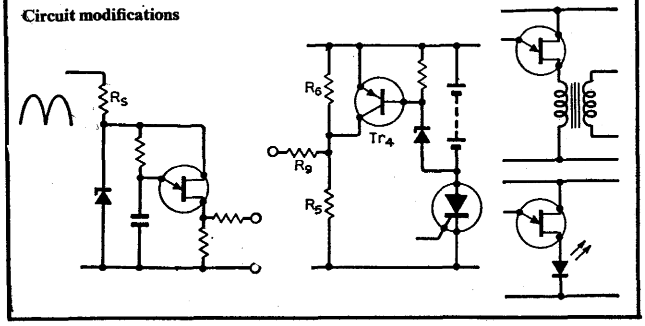
Further reading

Low-cost constant-current battery charger with voltage limiting, *Semiconductors* (Motorola), vol. 3, 1972, no. 1, pp. 15/6.
400V constant-current source, Electronic Circuit Design Handbook, Tab. 1971, p. 298.
Nowicki, J. E., Power Supplies for Electronic Equipment, vol. 2, Leonard Hill, 1971, pp. 182-93.

Cross references

Series 2, card 5. Series 3, card 4. Series 6, card 7.

Circuit modifications



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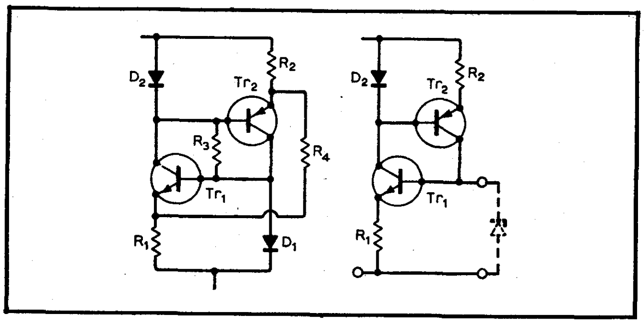
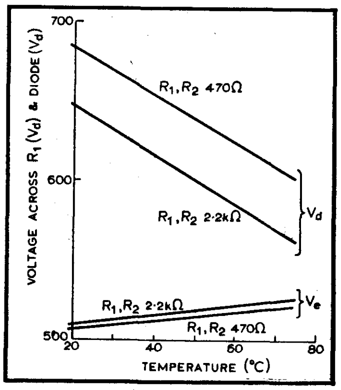
Low-voltage current regulators



Typical performance



D_1: 1S130
D_2: 1S130
Tr_1: 2N1308
Tr_2: 2N404
R_1, R_2: 220Ω
R_3, R_4: ∞. Typically leakage current of Ge transistors sufficient for self-starting. To increase dynamic resistance R_4 may be in range $100R_1$ to $1000R_1$.

Circuit description

The ring-of-two reference (card 5) may be adapted for very low voltage applications by replacing the zener diodes by forward-biased silicon diodes or any other element having dynamic resistance less than static resistance ('amplified' diodes, asymmetric voltage-dependent resistors, gallium arsenide diodes, etc.). The transistors used must then have a V_{be} less than the diode forward voltage drop, and germanium devices are indicated for use with silicon diodes. For optimum temperature compensation with these devices, the p.d. across each emitter resistor should be around 420mV (a figure based on the junction properties of the devices). This is not always convenient to achieve, but stability of 0.1 %/deg. C is normally possible. Leakage currents of the Ge transistors are enough to ensure start-up in most cases and R_3 may be dispensed with. Resistor R_4 may be added to neutralize the effect of R_3 if present, and if absent to control the dynamic resistance of the two-terminal circuit. It bypasses current around the transistors reducing the collector current in each, i.e. opposing the natural tendency for a slight increase in current as the terminal p.d. increases. Dynamic resistance may even be made negative and large if R_4 is reduced sufficiently though over a more limited range of supply voltages than normal. This circuit, as with related circuits on card 5, may be used to supply a constant current to an external zener diode minimizing the total supply voltage required (as compared with its use as a two-terminal circuit interposed between supply voltage and load).

Component changes

D_1, D_2: Any silicon p-n junction including diodes (1N914, etc.) base-emitter junction of transistors (2N3707, BC125, BC126, ME4103, ZTX300, etc.) diode-connected transistor i.e. collector-base short.
Tr_1: n-p-n germanium transistor (OC139, 2N1302, 2N1304, 2N1306, 2N1308).
Tr_2: p-n-p germanium transistor (2N1303, -05, -07, -09, OC42, OC44) for optimum temperature performance with reasonably high gain transistors, diode/transistor combination should result in 400-450mV across emitter resistor.

Circuit modifications

- Diodes may be placed in one limb of the circuit, over-compensating the temperature induced change in Tr_1 V_{be}. By keeping R_1 and R_2 low, resulting decrease in the p.d. across R_1 is insufficient to compensate for the change in the V_{be} of Tr_2. Hence currents in the two limbs change in opposite senses and approximate cancellation is possible. Once this has been achieved, R_1, R_2 may be replaced by a single potentiometer, varying the total current while remaining approximately compensated.
- A different circuit using transistors of only one type is basically a voltage regulator defining the p.d. across a resistor whose current is larger than the remaining circuit currents (similar to card 2). Simplest version defines the current in terms of Tr_1 V_{be} and suffers from variation of current in R_1 as supply varies in addition to temperature dependence (≈0.3 %/deg. C).
- Replacing R_1 by a junction f.e.t. Tr_3 improves the constancy of current against supply voltage while the introduction of D_1 a germanium diode gives first-order temperature compensation.
- With the penalty of higher terminal p.d. better stability is given by the addition of zener diode D_1. Resistors R_2, R_3 compensate for current variations in R_1 by causing the p.d. across R_4 to fall as the supply voltage rises. Typically $R_1 = 10R_4$, $R_2 = 100R_4$, R_3 is varied to optimize slope resistance, but is in the region 0.5 to $5R_4$.

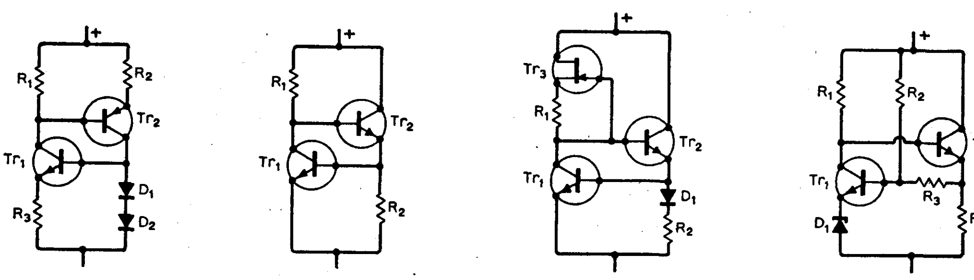
Further reading

Williams, P., Low-voltage ring-of-two reference, *Electronic Engineering*, 1967, pp. 676-9.
Verster, T. C., Temperature-compensated low-voltage reference, *Electronic Engineering*, 1969, p. 65.
Watson, G., Constant-current circuit, *Electronics*, 6th July, 1962, p. 50.

Cross references

Series 6, cards 3, 4, 5 & 6.

Circuit modifications



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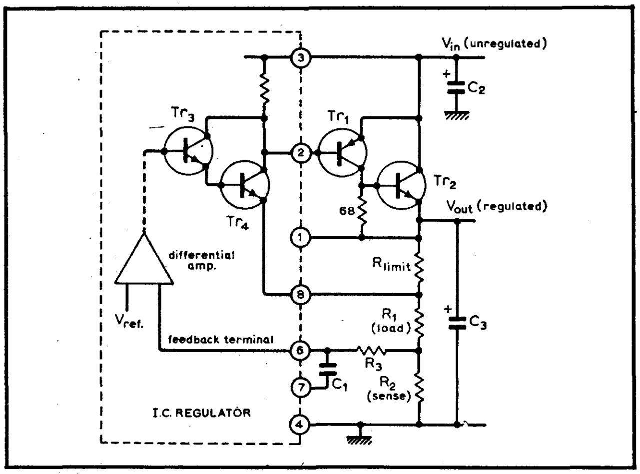
High-power current regulators

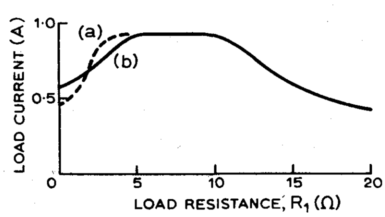


Typical performance

Load current: 0.9A
Unregulated input: 13 to 20V
IC: LM300
Tr_1: BFR81; Tr_2: MJE521
R_{limit}: 1Ω; R_1 (load): 10Ω
R_2: 1.95Ω; R_3: 2.2kΩ

C_1: 47pF
C_2: 1μF (tantalum)
C_3: 4.7μF (tantalum)

Graph shows effect of foldback current limiting on output current when load R_1 is varied (see circuit over, left)



(a) R_{limit} = 1Ω R_4 = 56Ω R_5 = 470Ω
(b) R_{limit} = 0·6Ω R_4 = 56Ω R_5 = 1k5Ω

Circuit description

This is basically a series voltage regulator used as a constant-current source, where the maximum output current depends on the current gain and power rating of the series-pass transistors (Tr_3, Tr_4) connected as a Darlington pair. Further amplification, and thus a greater output current, is available by modifying this series element by connecting two discrete transistors Tr_1 and Tr_2 to give a compound emitter-follower. The p-n-p/n-p-n combination is preferred for an improved temperature coefficient over a straightforward quad emitter-follower.

The essential function of this regulator is that some fraction of the output voltage (or a voltage due to load current through a resistor) is compared with a reference voltage developed within the i.c. regulator. If the output voltage changes, the error signal is amplified and used to compensate for the original change by modifying the drive to the compound emitter-follower. The internal reference voltage is approximately 1.7V, and hence the feedback sense voltage developed across R_2 must approach this value for the desired load current, thus defining R_2. The resistors across the base-emitter terminals of the external transistors cause the operating currents to be raised and improves the stability.

An arrangement for foldback current limiting is shown over (left) and is used to protect the regulator against the load going short-circuit, and limits the current to around 0.5A under this condition. Capacitor C_1 is a frequency compensation capacitor. The additional current gain necessary for the high current regulators may cause h.f. oscillation, eliminated by connecting a tantalum capacitor across the input and the output.

Component changes

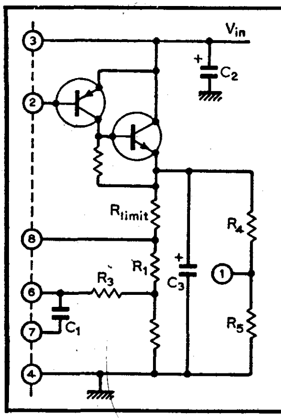
R_1 varies from 1 to 10Ω, current variation within +0.1% over the full range. Regulator may be LM100 or LM305.
Tr_1: 2N3055. Tr_2: 2N2905.
Parasitic oscillations can be suppressed by threading a ferrite bead over the emitter lead of power transistor Tr_2.
Basic voltage regulator normally has its output voltage set by connecting the tap on a potential divider to the feedback terminal. The resistance seen by this terminal should be around 2.2kΩ to minimize drift caused by the bias current at this terminal. This explains the values shown for the voltage regulator divider, and need for R_3 when the i.c. is used as a current regulator, the network equivalents being shown over (middle).

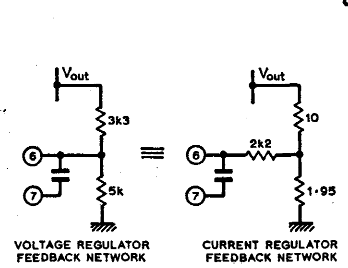
Circuit modifications

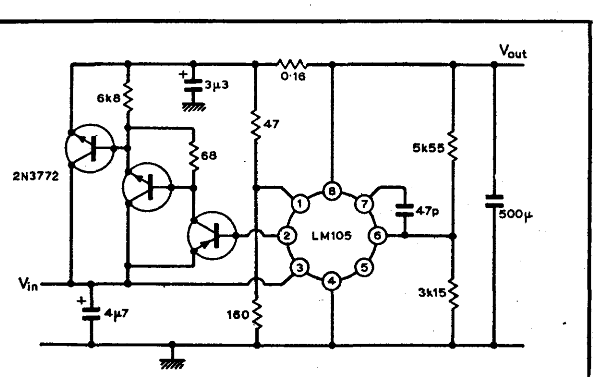
- Foldback current limiting is achieved by connection of resistors R_4 and R_5 (left). This provides protection for the regulator against excessive power dissipation should the load short-circuit, and limits the current to about 0.5A. Limiting starts when the voltage across terminals 1 and 8 exceeds +0.4V, and depends on the potential differences across R_{limit} and R_4. This critical voltage increases the positive bias on a transistor which therefore conducts harder and steers current away from the first transistor of the series element, andhence the load current decreases.
- Very high output currents can be obtained using LM105 or LM305 regulator, and an additional high power transistor. A typical arrangement is shown right to produce 10A, and with foldback current limiting. Input level should be >9V.

Cross references Series 6, cards 2 & 7.

Circuit modifications







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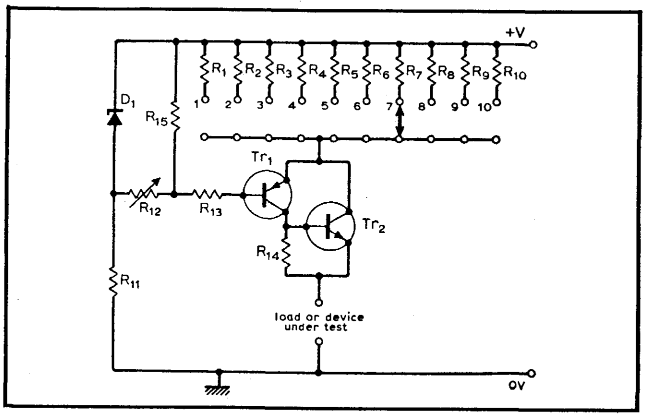
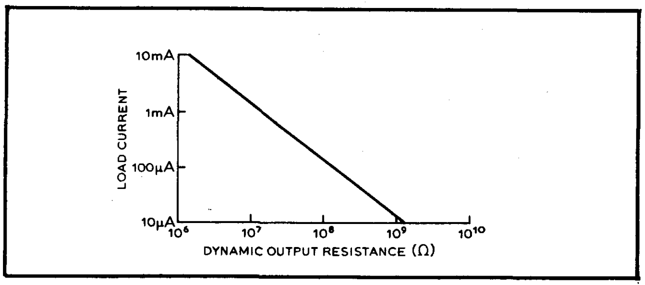
Constant-current applications



Typical performance

Supply: +12V
Tr_1: 2N3702
Tr_2: BFY50
D_1: HS7062
R_1: 560kΩ; R_2: 270kΩ
R_3: 100kΩ; R_4: 56kΩ
R_5: 27kΩ; R_6: 12kΩ
R_7: 5.6kΩ; R_8: 2.7kΩ
R_9: 1.2kΩ; R_{10}: 560Ω
R_{11}: 470Ω; R_{12}: 100Ω
R_{13}, R_{14}, R_{15}: 1kΩ

I_{supply}: 14.5 to 24.3mA. With load of 1kΩ all preset currents within +8% of nominal values and decade values, e.g. 10μA, 100μA, 1mA, 10mA within ±1% of each other. Dynamic output resistance/load current: see graph opposite.



Circuit description

A preset constant current may be used in many instrumentation applications in the same way as a preset voltage. Such a current generator may be used, for example, to test semiconductor devices such as diodes and zener diodes to obtain their current-voltage characteristics; in a zener diode the current may change by a factor or more than 100 with a corresponding voltage change of only a few percent. The circuit shown provides constant currents that are preset within the range 100μA (S_1 in position 1) to 10mA (S_1 in position 10), with an overall stability of less than 1% at any preset value. The accuracy of the preset currents is not so high as preferred-value 5% resistors were used, but can be improved by using selected values. For diode testing over a wide range of currents, the preset currents are chosen to be multiples of 1, 2, 5, 10 to allow rapid construction of a log-scale graph.

The zener diode D_1 sets the base potential of Tr_1 and hence the p.d. across its selected emitter resistor R_1 to R_{10}. Current in the selected resistor is therefore defined as is the current in the load or device under test. Transistors Tr_1 and Tr_2 form a complementary pair, the equivalent compound transistor having a current gain approximately equal to the product of the individual current gains and an input characteristic equivalent to that of Tr_1. The base current of Tr_1 is thus very much less than the load current so that the latter is virtually the same as that defined in the selected emitter resistor. By selecting the emitter resistor to be R_7 the load current can be set to be 1mA by adjustment of R_{12}. Constant currents of 10μA, 100μA and 10mA are then also defined to an accuracy, depending on the tolerances of R_1, R_4 and R_{10} respectively.

Component changes

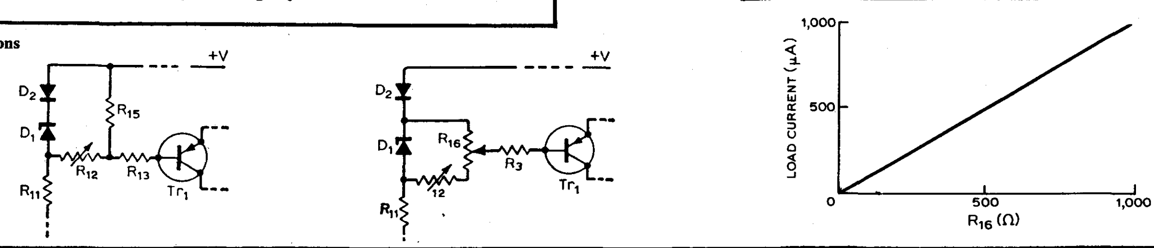
Larger values of constant current can be obtained by changing Tr_1 and Tr_2 to higher power transistors capable of handling the larger currents. The p.d. available at the load terminals can be increased by using a lower voltage zener diode for a given value of +V. The value of +V can be increased, provided that the breakdown voltage of Tr_1 and Tr_2 is not exceeded, to provide higher load voltages at defined currents. If the Tr_1 biasing network is replaced by a simple potentiometer between the supply lines a high output impedance is still obtained but the load current is less stable and the load p.d. will fall as the load current is increased by altering the potentiometer setting.

Circuit modifications

- Errors in the constant currents will be due to drift in the zener diode, drift in V_{be} of Tr_1 and the finite and variable current gain of the compound transistor. In the circuit discussed the zener diode is chosen for low slope resistance to limit dependence on supply voltage. If the circuit is operated from a stabilized voltage supply, the low slope resistance can be abandoned and the zener diode can be chosen to provide best temperature matching. A forward-biased junction diode can then be placed in series with a zener diode to provide temperature compensation for the drift in V_{be} of Tr_1 (see left), where D_1 could be a 5.6V zener and D_2 a BYX22-200.
- In addition to the preset constant currents it is often necessary to provide a current that may be accurately varied over a restricted range. This can be achieved by connecting a potentiometer of the calibrated multi-turn type across the zener diode as shown middle. A graph of the variation in load current achievable using S_1 in position 7 and a 1-kΩ potentiometer is shown right.
- As well as being used for measuring the characteristics of diodes and zener diodes, the unit described may also be used to measure loop resistance by monitoring the load terminal p.d. with a d.v.m. whilst feeding an appropriate constant current to the unknown resistance. By feeding a constant current to the emitter of a transistor and measuring its base current the d.c. current gain can be quickly found. Another application is in electrochemical plating.

Cross references Series 6, cards 1 & 2.

Circuit modifications



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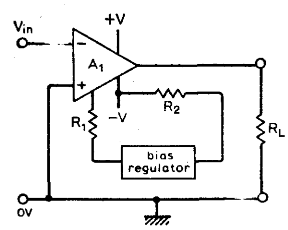
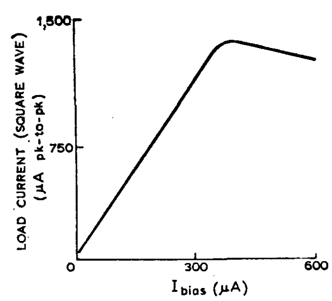
Constant-current amplifiers



Typical performance

Supplies: $\pm 6V$
A_1: $\frac{1}{3}\times$CA3060
(regulator is part of CA3060)
R_1: 53.7kΩ ±1% for I_{bias}: 100μA
R_2: 47kΩ for I_{bias} ≤100μA; R_L=1kΩ
Equivalent source resistance with I_{bias}= 100μA is approx. 264kΩ i.e. load current changes by about 4% for a 1000% increase in R_L.



Circuit description

A class of monolithic amplifiers is now available called operational transconductance amplifiers. This type of amplifier is a novel circuit having similar general characteristics to an operational voltage amplifier except that its gain is better described in terms of a transconductance rather than as voltage gain. Its open-loop voltage gain is equal to the product of its transconductance and the load resistance it feeds.

In the circuit A_1 is one of three transconductance amplifiers in a single package together with a bias regulator. The regulator is supplied from the −V rail through a resistor R_2 and each of the class-A push-pull transconductance amplifiers are biased independently by a suitable resistor R_1. The transconductance of the amplifier is controlled by the bias current i.e. by the value of R_1. For a given input voltage between the inverting and non-inverting inputs the output current is defined by the bias current which can be varied over a wide range.

While the amplifier can be used in its linear mode with various feedback arrangements, the open-loop circuit shown above can deliver a square wave current to the load resistance. The peak-to-peak amplitude of the square wave is under the control of the bias current. As the amplifier has a high output impedance, it may be thought of as being a generator of a current square wave having a definable and constant peak-to-peak value. The circuit can supply an output of around 1V pk-pk into loads of around 10kΩ with an equivalent source resistance of about 260kΩ, provided V_{in} is large enough.

Component Changes

Useful range of supply: ±2.5 to ±7V
Maximum differential input voltage: ±5V
Maximum d.c. input voltage: +V to −V
Useful range of bias current approx: 10μA to 2mA
Maximum bias regulator input current (total for 3 amplifiers): −5mA
Useful frequency range for square wave output current is typically 120kHz.

Circuit modifications

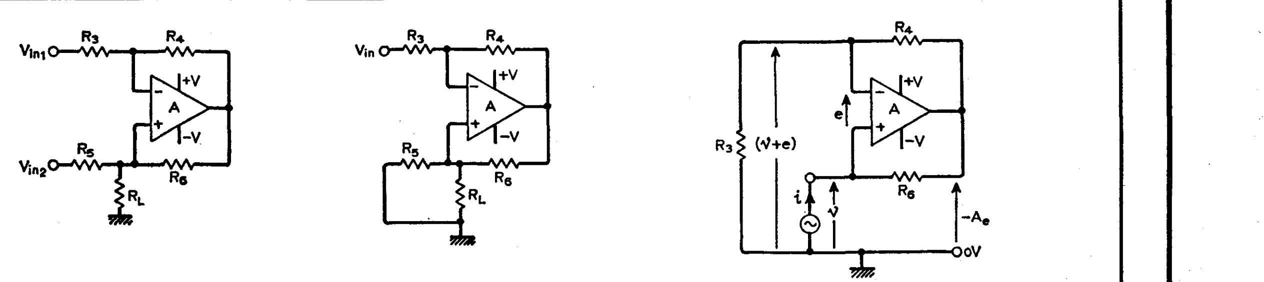
An amplitude-modulated constant-current source is obtained if the modulating voltage source is connected as a floating source in series with R_1 or as a grounded source to the bias terminal through a resistance of the order of 100kΩ. In the first arrangement 100% amplitude modulation of the output square wave is obtainable, whereas the latter connection provides about 30% modulation depth using a 12V pk-pk sine wave source.

Circuit left shows the general form of a circuit, known as the "Howland" circuit, which provides a constant current into the load by virtue of the fact that A, R_4 and R_6 act as a negative impedance converter. As shown, V_{in_2} must supply the short-circuit load current, therefore the circuit is often used in the form shown centre. The high output impedance available at the load terminals can be seen by reference to the diagram on right where R_L has been replaced by a voltage source, V_{in} has been set to zero and R_5 temporarily removed, for analysis. The output impedance at the load terminals is $Z_o = Z_P//R_5$ where $Z_P = v/i$. For simplicity, let $R_3 = R_4 = R_5 = R_6 = R$, then $-Ae = 2(V+e)/R$. Hence $e = -2V/(A+2)$ and $i = (V+Ae)/R = v - [A.2V/(A+2)]$. Thus $Z_P = V/i = (A+2)R/(2-A)$ and $Z_o = Z_P//R = R(A+2)/4$. Therefore, as $A \to \infty$ $Z_o \to \infty$ and a constant current may be fed to R_L.

For an operational amplifier of the 741 type, $A = -jA_o f_o/f$ where A_o and f_o are typically 10^5 and 10Hz respectively. In this case $Z_o \approx -jA_o f_o R/4f$ or $Z_o \approx -j\omega C$ so that Z_o consists of a capacitor $C \approx 2/\pi f_o A_o R$. For $R = 10k\Omega$, $C \approx 64pF$. Thus, the constant load current will be 3dB down w.r.t. its low frequency value at $f = 1/2\pi C R_L \approx A_o f_o R/4R_L \approx 250R/R_L$ (kHz) for a 741-type operational amplifier.

Cross references Series 6, cards 4 & 6.

Circuit modifications

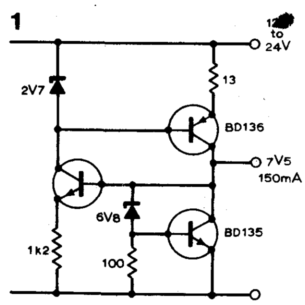


Wireless World Circard

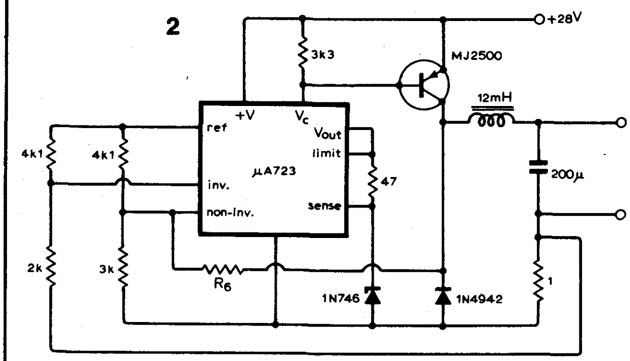
Series 6: Up-date

Constant-current circuits

1. The ring-of-two reference has a complementary pair of transistors driving constant currents into two voltage reference elements such as zener diodes. At higher current levels, one of the zener diodes is replaced by a zener/transistor combination acting as a shunt regulator in which the zener carries only the base current of the n-p-n transistor. Although the circuit as shown is for the regulation of an output voltage, such circuits can be used as two-terminal current regulators i.e. the dual of power zener diodes. The minimum terminal p.d. can be reduced by replacing the 6.8V zener diode by a forward-biased diode or, with further-reduced stability, replacing zener BD135 and 100Ω resistor by a single 15Ω resistor.
Hibbert, G. Avoiding power-supply hum, *Wireless World*, 79, 1456, p.515, (Oct. 1973).



2. Monolithic voltage regulators are readily adaptable to act as switching regulators or to provide a constant current action; the circuit shown combines these functions. The circuit is a form of astable in which R_6 is selected to give a hysteresis of 28mV at the non-inverting input. The internal reference voltage of 7.15V is reduced by the potential divider to give 3V at the non-inverting input i.e. the hysteresis is 1% of the mean value. The network feeding the inverting input sets the p.d. across the 1 Ω resistor to 1V, and the mean current in the load to 1A. The switching mode ensures high efficiency regardless of load p.d. making it a suitable circuit for battery charging of one or more cells. The LC circuit at the output filters the switched output for the power transistor and the switch frequency may rise to 20kHz.
Summer, S. E. Switching regulator produces constant current output, *Electronics*, March 7, 1974, p.114.



3. To control very low currents with high accuracy both the sensing and control elements must have input currents much less than the current to be controlled. For currents at the nA level this implies input currents of order 10pA if accuracies of 1 or 2% is required. The circuit shows one way of achieving this in which a f.e.t. input operation at amplifier compares the p.d. across a low-voltage reference diode with that across a very-high-value resistor. If the f.e.t. is selected for minimum gate leakage current the drain and source currents are equal and the output current is defined at about 1.2nA for the given diode.
Underwood, R. K. New design techniques for f.e.t. op-amps, National Semiconductor application note AN-63, 1972, p.9.