Set 23: Reference circuits

First described by one of the authors a little over a decade ago, the ring-of-two reference has become almost a classic constant voltage reference, with many variants coming to light since (See also set 6, especially card 5). The article reminds readers that zener diodes are not the only reference component; any device with a slope or dynamic resistance sufficiently different from its static resistance can be used. And as well as presenting measured characteristics of four zener types, the set includes characteristics of other kinds of device—v.d.rs, conventional diodes and l.e.ds.
Some circuits shown may be familiar under their alternative names—the amplified diode for the technique of page 40, and a constant current diode for the f.e.t. of page 46. (This last-mentioned page has some changes incorporated from the original card.) And band-gap reference circuits are briefly explained in the article on page 36. (Owners of set 23 of Circards will have noticed that Figs 1 & 2 on page 39 were originally in error on card 2.)

Reference circuits

Some semiconductor devices have highly non-linear characteristics, in which the non-linearity is well-defined with predictable and small dependence on temperature. If a region of the characteristic is found for which the slope resistance is either very much greater than or very much less than the static resistance then the device can be used as a current or voltage reference respectively.

In Fig. 1 (a) there is an extended region over which the voltage varies little for large changes in current. Fig. 1 shows the dual characteristic with current being maintained constant against changes in bias voltage. The most commonly used device belonging to the former category is the zener diode, the reverse characteristic having a sharp breakdown region. There are two physical mechanisms that can control the reverse conduction of a p-n junction: zener breakdown and avalanche breakdown.

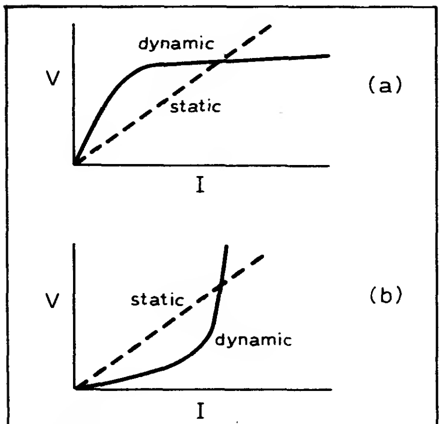


Fig. 1

Zener breakdown is a field effect which dominates for heavily-doped narrow junctions, where even small p.ds of three or four volts can provide a sufficiently intense field for the direct production of hole-electron pairs. The observed characteristics are that the current increases steadily as the operating region is approached, with a rounded knee, and with a temperature drift of about $-2mVK^{-1}$. To a first order the slope resistance of such diodes is inverse to the quiescent current.

At higher p.ds, which can only exist with more lightly doped broader junctions where the zener effect is unable to limit the voltage, thermally generated holes and electrons are accelerated by the field. If the p.d. is large enough some will gain sufficient kinetic energy before colliding with other atoms, to produce further hole-electron pairs by collision. These in turn may generate further pairs and at a particular voltage there is a very sharp increase in current. Below breakdown the current is negligible, while above it the slope resistance is low. The voltage changes with temperature by less than $+0.1\%\ K^{-1}$.

There is an intermediate doping level resulting in breakdown voltages between five and seven volts where both processes contribute significantly to the total current. The proportion is dependent both on the junction and on the current level, but it is possible for diodes between 5.5 and 6.5V to have negligible drift with temperature if biased correctly (lower currents for the higher voltage devices). An identical breakdown occurs in the base-emitter region of a transistor, and planar silicon transistors can be used as low-current zener diodes with good slope resistance. Breakdown voltage for the base-emitter junction is typically 6 to 10V, varying little for a given device. (Breakdown diodes are commonly described as zener diodes regardless of which physical process dominates.)

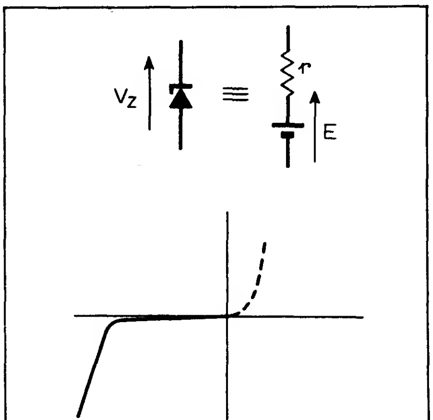


Fig. 2

A simplified equivalent circuit for such a diode if biased into the low slope region is shown in Fig. 2. It consists of a constant e.m.f. in series with a small resistance. The resistance is assumed constant i.e. the characteristic is approximated to by the 'piecewise linear' graph shown. A circuit for a simple zener diode regulator is shown in

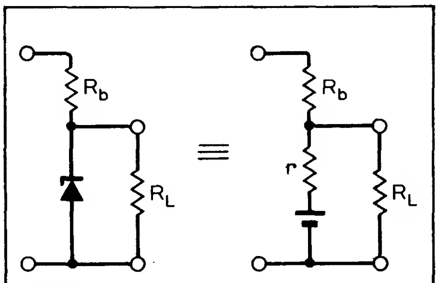


Fig. 3

Fig. 3. For changes in the supply voltage, load current etc the constant e.m.f. may be suppressed e.g. for an input voltage change ΔV_S, the output voltage changes by

$$\left(\frac{\frac{rR_L}{r+R_L}}{Rb+\frac{rR_L}{r+R_L}}\right)\Delta V_S$$

Since $r \ll R_L$ and $r \ll R_b$ are reasonable assumptions for a correctly designed circuit, the result simplifies to $\Delta V_o/\Delta V_S \approx r/R_b$. Similarly the output resistance is $\approx r$.

Where the diode is used simply to produce a stable reference voltage, the load current can usually be arranged to be negligible, or at least reasonably constant. This leaves only supply voltage and temperature variations to be dealt with, though for high-stability designs ageing of the device may be equally important in bringing long-term drift. The two problems require different solutions.

The effect of supply voltage is determined by the circuit design, while temperature effects can be minimized by choosing the right diode. In some cases the reference diode may have one or more forward-biased diodes added in series. By selecting as the reverse-biased diode, one with a breakdown voltage $>7V$, its positive drift can be balanced against the negative drift of the forward-biased diode(s). In the circuits of Figs 4 to 7 the single zener diode could be replaced with any such combination.

Though the diode has a low slope resistance its voltage stability will be ideal if fed from a constant current (Fig 4). A practical way of realizing this is to use a transistor with a fixed base-potential and large emitter resistor. Any variation in supply voltage causes only a small variation in the transistor current and hence a still smaller change in the output voltage. An extension of the method, the ring-of-two reference (Fig. 5) has two zener diodes each controlling the constancy of current fed to the other. In this and other related circuits the variation in output voltage due to supply charges can be reduced to a few tens of microvolts – generally far lower than the variation due to temperature changes.

Most i.c. voltage reference/regulator circuits are based on similar principles while exploiting the matched-characteristics of adjacent transistors as in Fig. 6. The transistors Tr_1 and Tr_2 comprise a current mirror forcing the zener diode current to equal the current in R_E, which in turn is closely defined by the zener voltage. Both circuits contain a positive feedback loop, clamped by the zener, but they are essentially bistable in nature i.e. all devices could remain non-conducting indefinitely. To inhibit this condition the resistor R_S (which can be very much greater than R_E) provides a starting current without significantly impairing the regulation.

Where the reference voltage is of an inconvenient value then a voltage amplifier may be added as in Fig. 7. Further advantages accrue from this approach. The current drawn from the diode is reduced to negligible proportions; the output current capability is increased without forcing the zener to operate at a high current; the output impedance is very low because of the shunt-derived negative feedback; the diode can be biased either from a separate supply, or from the amplifier output provided it is sufficiently greater than the zener voltage. This last method is of the same nature as those adopted in the circuits of Figs 5 and 6 viz that the zener voltage indirectly controls its own bias current. The stability can be extremely high, but the non-conducting state can also occur and may require a separate starting circuit.

Although zener diodes are the most common voltage reference units, they can be replaced by any element conforming to Fig. 2). Examples include forward-biased silicon diodes, assymetric voltage dependent resistors (down to 1V), forward and reverse biased

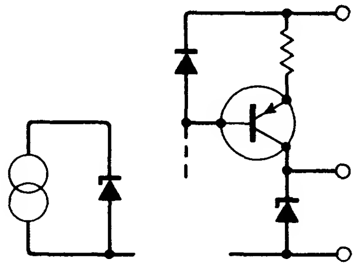


Fig. 4

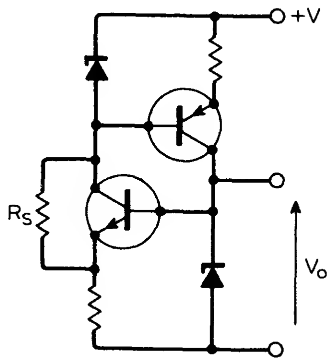


Fig. 5

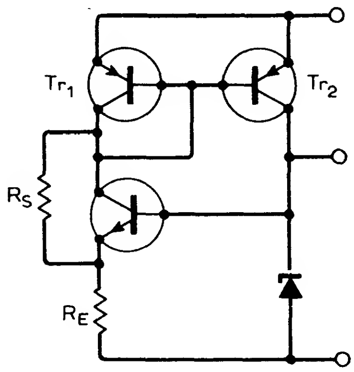


Fig. 6

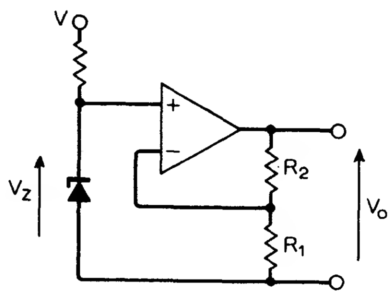


Fig. 7

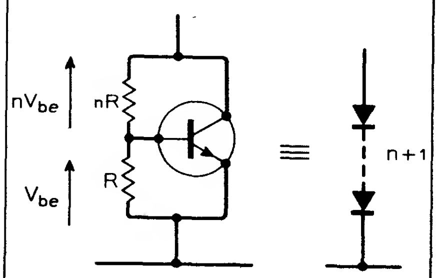


Fig. 8

junctions of transistors. A useful circuit where high stability can be sacrificed in exchange for flexibility is the amplified diode circuit of Fig. 8. If a transistor is biased by a potential divider between collector and emitter then under certain constraints, the terminal p.d. approximates to that of $(n+1)$ diodes in series. The current in the potential divider must be much greater than the transistor base current, but not much in excess of the collector current. Note that n need not be an integer and that by replacing the base-collector resistor with a variable control, we have a simple variable zener diode. The temperature drift is relatively large, about $+0.3\%K^{-1}$, but an overall stability of a few percent is readily achievable under laboratory conditions.

A completely different principle is embodied in the circuit of Fig. 9. While the V_{be} of a transistor falls as the temperature rises, ΔV_{be} between two identical transistors operated at differ-

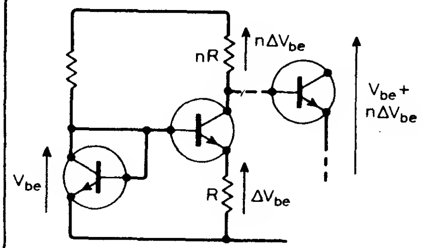
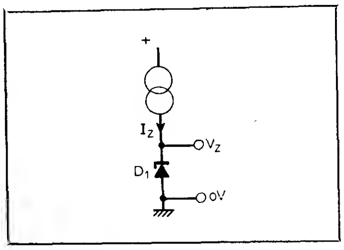


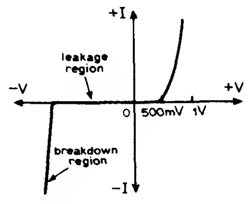
Fig. 9

ent currents has a positive coefficient. The circuit, a much simplified form of that used in recent i.c. regulators, has a terminal p.d. of $V_{be}+n\,\Delta V_{be}$. A study of the transistor equations shows that this sum equals the energy-band gap of silicon at the point where the temperature drifts cancel. This voltage is about 1.23V and is scaled up by suitable amplifying circuits where required. The forward characteristics of devices can reasonly be expected to offer better long-term stabilities than in the breakdown region, and this principle is well-established in i.c. reference circuits of the highest quality.

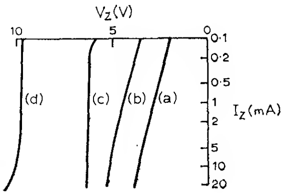
Zener diode characteristics



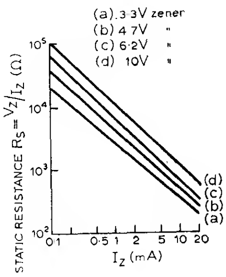
Performance (see graphs)
Constant current source: 0-20mA from commercial generator, $\pm 0.05\%$.
V_z measured with 5-digit d.v.m.
D_1 (a) BZX83C3V3
(b) BZX83C4V7
(c) BZX83C6V2
(d) BZX83C10.



Graph 1



Graph 2

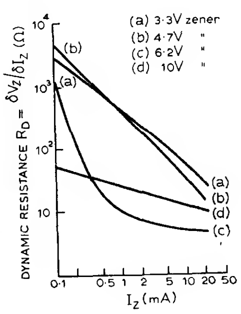


Graph 3

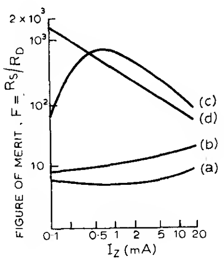
Description
The zener diode exhibits three distinct regions: forward, leakage and breakdown. The forward-bias region is virtually identical with a normal diode, the forward-voltage temperature coefficient for a constant forward current being about −1.4 to −2.0mV/degC. Under reverse bias, up to the breakdown region, a leakage current exists which, although being temperature dependent, is normally less than 1μA over the whole temperature range. When the reverse bias reaches some definite value, which depends on the p-n junction doping levels, the diode current rapidly increases and the breakdown or zener region has been reached. This region is the one used to provide a d.c. reference voltage by supplying the device from a constant-current source, as shown above left. At the onset of breakdown the resistance of the zener diode will be high but as the current increases the number of breakdown sites increases and the resistance falls to a small value. The reverse-breakdown characteristics for several zener diodes are shown at upper right, and their corresponding typical plots of static resistance as a function of zener current are shown lower right. The corresponding plots of slope or dynamic resistance are shown above, extreme left, The curves were obtained by setting I_z to a defined value and then reducing it instantaneously by 20% of the set value to obtain $R_D = \delta V_z/\delta I_z$. A sensitivity factor S of the dependence of zener voltage or current can be defined as $S=(\delta V_z/\delta I_z)/(V_z/I_z)$ and the reciprocal of S used as a figure of merit for the device ($F=1/S=R_S/R_D$), typical plots being as shown above, centre left. This figure of merit refers to the device only and will be degraded to a degree dependent on the circuit in which it is used and the zener current flowing. These figures may be used to assess the ability of the circuit to maintain a desired reference voltage against supply variations. Thus, if I_z changes by $x\%$ due to supply variation the reference voltage will change by approximately $(x/F)\%$ if the current source resistance is much larger than that of the zener diode. This would indicate the use of high-voltage supplies and high-voltage zener diodes operated at relatively low current levels. However, high voltages are not necessarily available and a compromise must normally be made between wasted volts and required stability of the reference voltage. Often the stability of the reference voltage against temperature changes is of prime importance and the choice of zener diode will depend on its temperature coefficient. Typical plots at room temperature are shown above centre right which indicate the use of diodes having a breakdown voltage of about 5V. Note that all these curves indicate that a positive temperature coefficient may be a distinct advantage as all the curves merge at a temperature coefficient exceeding about +4mV/degC. Thus a temperature-stable reference may be produced by replacing the single zener diode with a temperature-compensated reference unit consisting of the reverse-biased zener diode in series with n forward-biased normal diodes which exhibit a negative temperature coefficient of about 2mV/degC. If the zener diode has, for example, a temperature coefficient of +6mV/degC it could be series-connected with three forward-biased diodes.

Further reading
Patchett, G. N. Automatic Voltage Regulators and Stabilizers, 3rd edition, chapter 6, Pitman, 1970.
Buchanan, J. K. *et al.* Zener Diode Handbook, Motorola 1967.

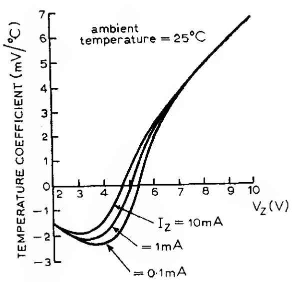
Cross references
Set 23, cards 2, 3, 4



Graph 4

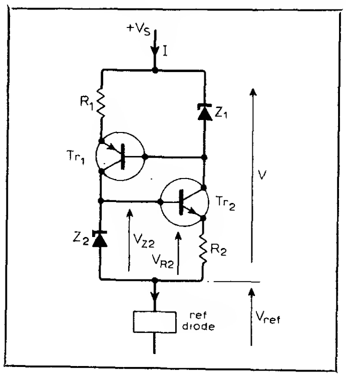


Graph 5



Graph 6

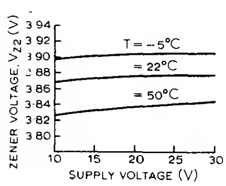
Williams ring-of-two reference

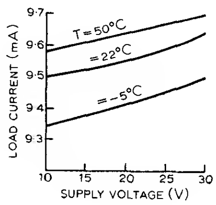


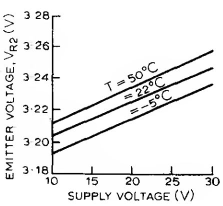
Typical data
Tr_1 BC126, Tr_2 BC125
Z_1, Z_2 BZY88 (3.9V)
R_1, R_2 680Ω
For test requirement, I was determined by measuring voltage across a standard resistance 50Ω±0.05% with a 5 digit voltmeter. Temperature levels obtained in a controlled oven.
Minimum $V_S \approx 10V$.

Circuit description
This circuit may be used to supply a constant current to a high capability reference diode, which is itself used as the voltage reference source. If it is assumed that diode Z_2 provides a constant voltage at the base of transistor Tr_2, then this forces a constant current to flow through the emitter resistor R_2. For reasonably high gain transistors, the collector-current will almost equal the emitter current, and hence the current through Z_1 will be constant. But this diode will maintain a constant potential at the base of transistor Tr_1, which in turn forces a constant current through R_1, Tr_1 to operate the original diode Z_2, which was initially assumed to provide a stable voltage. The total current drawn by the circuit is the sum of the collector currents and is substantially constant.
An increase in the supply voltage V_S largely appears between the collector and emitter of the transistors.
At 22°C, change in V_{Z2} 6.4mV
change in V_s 20V
Stability $\Delta V_S/\Delta V_Z \approx 3300$
At V_S=20V, ΔV_Z= +84mV for overall temperature change from 50°C to −5°C.
Temperature coefficient is −1.5mV/degC.

Note that the graph plots are obtained from a circuit using unselected diodes, and no attempt was made at temperature compensation. Maximum supply voltage allowable will depend on permitted V_{CE} of transistors.
For V_S=20V, I=0.224mA for T=55°C.
$\Delta I/\Delta T \approx 4\mu A$/degC
At 22°C, ΔV_S=20V,
ΔI=128μA
$\Delta I/\Delta V_S \approx 4\mu A/V$
Percentage change ≈ +1.3%
Note. If self-starting difficulties arise, a resistor between bases or resistors across collector-emitter terminals may be used.







Circuit modifications
Stabilization ratio $\Delta V_S/\Delta V_Z$ of 10^5: 1 is claimed for the circuit (Ferranti) using reverse biased base-emitter junctions of transistors ZTX303/300 as reference diodes. Tr_2 ZTX302, Tr_1 ZTX500. Circuit current 1mA for R_1, R_2 6.8kΩ. Supply range 14 to 25V.
The voltage reference circuit of Fig. 1 may provide a stability factor of the order of 10^6 for a voltage range of 20-40V.
Z_1, Z_2 6V planar zeners
Tr_1 2N3702, Tr_2 2N3820
Tr_3 2N3819, Tr_4 2N3707

Circuit Fig. 2 uses forward biased diodes as references to achieve a low voltage reference[2] Stability is maintained down to 1.1V supply. Temperature change compensation is obtained by matching forward voltage drift on the silicon diode against that of base-emitter junction of germanium transistor. The circuit of Fig. 3 includes diode connected transistors to offset the base-emitter voltage variation with temperature[1]. As temperature also affects the transistor common emitter current gains, this effect is minimized by feeding these via op-amps[1]. Notice in Fig. 4 current feedback is to inverting inputs. Also slight variations in base current will still affect the collector currents. The use of junction f.e.ts in Fig. 5 reduces this dependence.

Further reading
Williams, P. Ring-of-two reference, *Wireless World*, July 1967. See also *Proc. IEEE*, January 1968.

References

1 Applied Ideas, *Electronic Engineering*, December 1974.
2 Williams, P. Low-voltage ring of two reference, *Electronic Engineering*, November 1967.
3 Ferranti E-Line Transistor Applications, June 1969.
4 Williams, P. D.C. reference voltage with very high rejection of supply variation, *Proc. IEEE*, January 1968.

Cross references

Set 23, card 1
Set 6, card 5

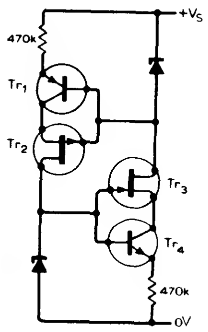


Fig. 1

Fig. 2

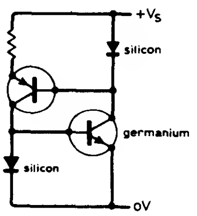


Fig. 3

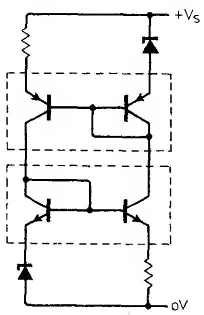


Fig. 4

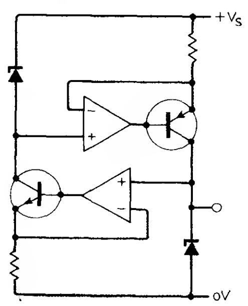
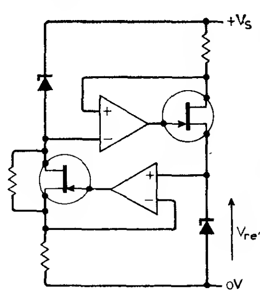
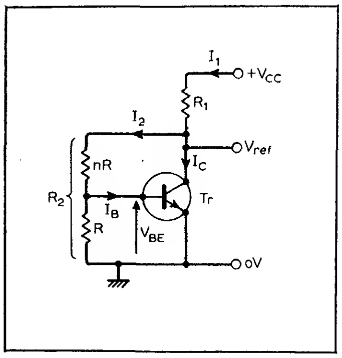


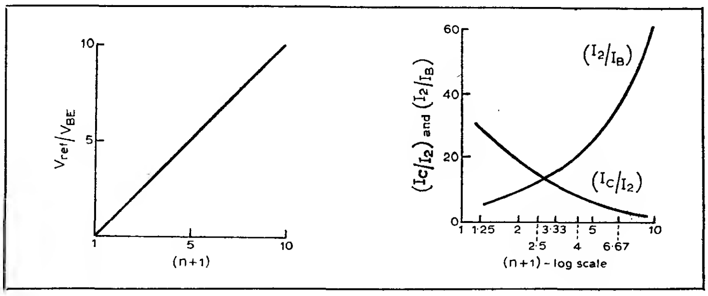
Fig. 5



Variable reference diodes



Typical performance
+Vcc +15V
R_1 4.7kΩ±5%, R_2 10kΩ multiturn ±1%, linearity ±0.1%
Tr_1 BC125
Currents for $n=1$ condition
I_1 2.91mA, I_C 2.77mA
I_2 144.6μA, I_B 16μA
(see graph opposite)
Voltages for $n=1$ condition
V_{BE} 644mV, V_{REF} 1.371V
(see graph opposite)



Description
Although the zener diode is the most common device used to produce a stable reference voltage, may be replaced by any device, combination of devices or circuit that behaves as a two-terminal element having a stable p.d. across it and some internal resistance. Like the zener diode, such elements can normally only produce a definable, fixed reference. Many instances arise, especially under laboratory conditions, where a variable reference voltage is required and which has a range of required values that are not necessarily available from a single device or a combination of devices.
The circuit shown above left is an example of a simple d.c. reference which can often meet these requirements. If the transistor is assumed to have infinite current gain, I_B tends to zero, and the current I_2 in R_2 will produce p.ds across R and *nR* proportional to the current flowing. The p.d. across R is V_{BE}, which is dependent on the transistor current, and the p.d. across resistor *nR* will be *n* times that across R, i.e. nV_{BE}. Hence, the reference voltage will be $V_{BE}+nV_{BE}=(n+1)V_{BE}$. Since the factor $(n+1)$ cannot be less than unity the circuit is normally referred to as the amplified diode or the V_{BE} multiplier. In practice most commonly-available transistors have sufficiently high current gain to allow predictable performance. Departure from the ideal condition of $V_{REF}=(n+1)V_{BE}$ will occur at both high and low current levels, the first because I_B eventually becomes an appreciable part of the current in the bias resistors and the second because the collector current becomes a relatively small part of the total current and ceases to exercise control of the voltage. In most applications *R* will be held constant and *nR* will be a variable resistance used to set V_{REF} to the desired value. The major advantage of this circuit is that V_{REF} can be made almost any number of times greater than V_{BE} including non-integral values. The graphs shown overleaf show the linearity obtainable in practice between desired value of $(n+1)$ and actual value of (V_{REF}/V_{BE}) up to $(n+1)=10$. This graph was obtained by keeping R_2 fixed, by using a potentiometer, and varying the ratio of *nR* to *R*. The second graph overleaf shows the corresponding ratios of collector current to bias-chain current and bias-chain current to base current. Whilst the design aim should be to keep the bias-chain current much greater than the base current and the collector current much greater than the bias-chain current the last requirement is not nearly so important. The temperature dependence of V_{REF} is related to that of V_{BE} of Tr_1 which is typically −2mV/degC so V_{REF} will have a temperature coefficient of approximately −2*n* mV/degC.
The current source to feed the amplified diode can be realized by a current mirror so that three transistors in an i.c. transistor array may be used. At the expense of raising the lower limit of V_{REF} a zener diode with a suitably chosen positive temperature coefficient can be included in series with the emitter as shown in Fig. 1. To provide a negative, variable reference voltage a p-n-p transistor and zener diode are connected as shown in Fig. 2. Circuits of the amplified diode type are available in monolithic, integrated circuit form which can be operated over a wide range of voltages and currents and which have a definable temperature coefficient. An example of this form is the General Electric D13V which is a combination of a Darlington-type transistor pair and a zener diode. The internal circuitry and normal connection arrangements are shown in Fig. 3.

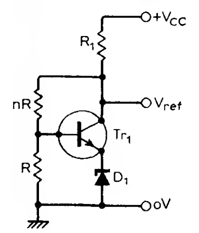


Fig. 1

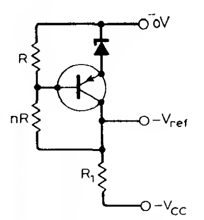


Fig. 2

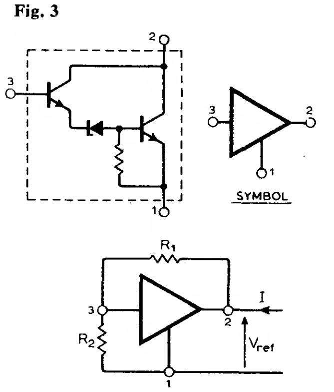
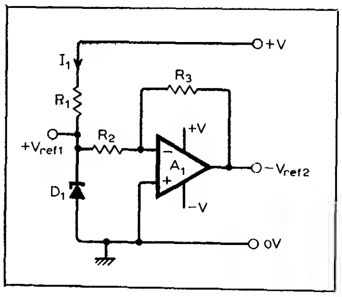


Fig. 3

Further reading
Williams, P. The Amplified Diode, *Design Electronics* January 1968, pp. 32-4.
General Electric D13V data sheet, 1970.
Glogolja, M. Biasing circuit for the output stage of a power amplifier, *New Electronics*, 17 Sept. 1974, pp. 20 24.

Cross references
Set 6, card 4
Set 23, cards 1, 6
Set 7, card 8

Bipolar references

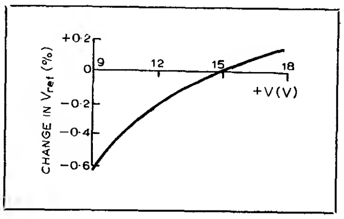


Typical performance
Supply ±15V, $+3.4$mA, -1.8mA
A_1 741, D_1 BZX830, 6.2V
R_1 4.7kΩ $\pm5\%$
R_2, R_3 22kΩ $\pm5\%$
I_1 1.94mA
V_{REF1} V_Z $+6.16$V
$-V_{REF2}$ -6.20V

Circuit description
The reference element is the zener diode D_1, the basic reference circuit consisting of R_1 and D_1 in series, the current in D_1 being determined by R_1 for a given positive supply voltage. In the above arrangement the positive supply to the zener is, for, convenience, made the same as that provided for the operational amplifier A1. The reference voltage V_{REF1}, which is positive and equal to the zener voltage, is fed to the inverting operational amplifier so that $V_{REF2}=-V_{REF1}R_3/R_2$. Thus, with $R_3=R_2$, a simple bipolar reference circuit is obtained with outputs having the same voltage magnitude. In the experimental arrangement, the resistors were $\pm5\%$ tolerance types which were not selected for equality resulting in $V_{REF2}=-1.0065\,V_{REF1}$.
A close match between the reference voltages can be obtained by carefully matching R_2 and R_3. Note that whilst V_{REF1} is fixed for a given zener diode and choice of supply and component values, $-V_{REF2}$ may be varied over a wide range by adjusting the ratio R_3/R_2. However, the values of these resistors should not be so small as to appreciably load V_{REF1}. The negative reference voltage output can be much more heavily loaded than the V_{REF1} output, since the former is available from the low-output-resistance operational amplifier. The change in the values of the reference voltages with temperature changes are essentially due to the zener diode characteristics, since the operational amplifier drift is very small in comparison and with $R_3=R_2$ the resistor temperature coefficients match to maintain a unity gain inversion of V_{REF1}.

Circuit modifications
To allow the V_{REF1} output to be more heavily loaded an operational amplifier voltage follower may be added to the basic circuit as shown in Fig. 1; this reference then being available from a low-output-resistance source.

Although this arrangement makes the current in the zener diode independent of the load currents taken from the V_{REF1} and V_{REF2}, the zener current is still subject to



variations in the positive supply rail voltage. Since an operational amplifier is a device which provides an output voltage that is inherently isolated from supply rail variations, the circuits above can be improved by supplying the direct voltage to R_1 from an operational amplifier instead of directly from the supply rail.
Irrespective of how the zener diode is supplied, its operating current may be more precisely defined and made independent of loading by placing the zener diode in the feedback path of an operational amplifier. This technique also allows the ability to provide a pair of opposite-polarity reference voltages having a summation equal to the zener voltage, including the particular case where they have equal magnitude. The basic form is shown in Fig. 2.
In this circuit R_1 again supplies the current to R_2 and R_3 as well as the zener diode current. Since the junction of R_2 and R_3 is a virtual earth, the output voltages are given by $V_{REF1}=V_ZR_2/(R_2+R_3)$ and $V_{REF2}=-V_ZR_3/(R_2+R_3)$ so with $R_2=R_3$, $V_{REF2}=-V_{REF1}=V_Z/2$. The operational amplifier must be capable of sinking all currents except the load current at the V_{REF1} output. If the operational amplifier is to be an inexpensive type a transistor current booster can be added as shown in Fig. 3.
The operational amplifier now only has to supply the base current to Tr_1. Diode D_2 is included to ensure that the amplifier turns on correctly.
A complete bipolar reference circuit using booster transistors at each output and having the zener current supplied from an operational amplifier is shown in Fig. 4.
In this circuit the magnitude of each output is equal to V_Z.
Typical component values are
$\pm V$ ±15V, A_1, A_2 301A
Tr_1 2N3964, Tr_2 2N2222
D_1 1N829 (6.2 volt zener)
D_2, D_3 1N914, R_2, R_3 6.2kΩ
R_1 826Ω, R_4, R_5 300Ω
R_6 3.1kΩ

Further reading
Miller, W. D. & Defreitas, R. E. Op.amp. stabilizes zener diode in reference-voltage source, *Electronics* Feb. 20, 1975 pp. 101-5.

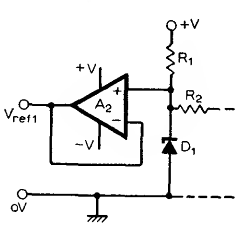


Fig. 1

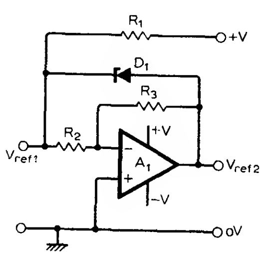


Fig. 2

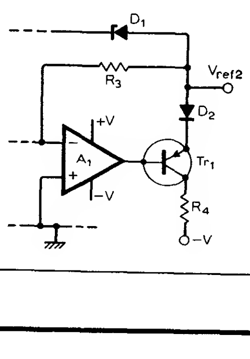


Fig. 3

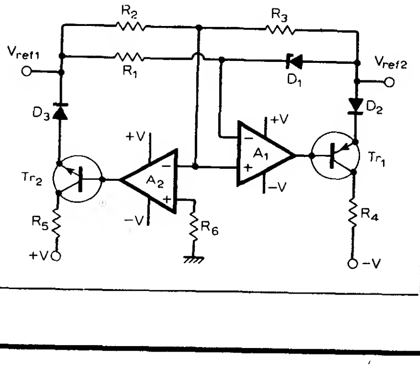
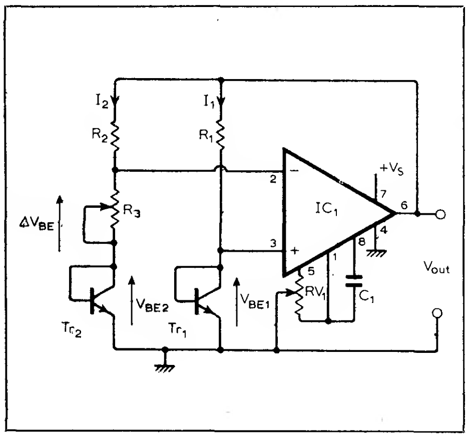


Fig. 4

Low temperature coefficient voltage reference



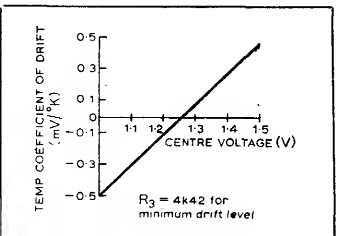
Typical data
Tr_1, Tr_2 Matched pair or $\frac{2}{5}$ CA3086
IC_1 CA3130AT
R_1 4.7kΩ, R_2 47kΩ
R_3 10kΩ (variable)
RV_1 100kΩ
C_1 1000pF
V_s +10V
At ambient temperature:
$V_{out}=600mV+10.(26mV)(2.3)=1200mV$

Circuit Description
Transistors Tr_1 and Tr_2 have identical parameters and hence similar saturation currents whose ratio is therefore temperature independent. The op-amp gain may be considered infinite and therefore the potentials of terminals 2 and 3 can be considered equal, for a finite output, V_{out}.

$$V_{out}=V_{BE1}+I_2R_2$$
$$=V_{BE1}+\Delta V_{BE}R_2/R_3$$

It is arranged by choice of R_1 and R_2, that I_1 is about ten times R_2.
Since $V_{BE}=(KT/_q)\ln(I/I_S+1)$ and $V_{BE}=V_{BE1}-V_{BE2}$, then $V_{out}=V_{BE1}+(R_2/R_3).(KT/q)\ln(I_1+I_{S1}/I_2+I_{S2}).I_{S2}/I_{S1}$
$=V_{BE1}+(R_2/R_3).(KT/q)\ln R_2/R_1$ because I_1/I_2 is in the ratio of R_2/R_1 and $I_{S2}=I_{S1}$. For $R_2/R_1=10$, then V_{out} is approximately defined at 1.2V at room temperature, for $R_2/R_3=10$. See typical data.

The base emitter junction voltage is also approximately given by $V_{go}-CT$ where V_{go} is gap energy voltage at 0 K and C is a constant. When the negative temperature coefficient of V_{BE1} and the positive temperature coefficient of the second term above, cancel, then $V_{out}=V_{go}$, and is then essentially temperature independent. Quoted value at 300K for V_{out} is 1.236V. Although the op-amp used in this circuit has temperature drift in excess of bipolar op-amp, it has the advantage

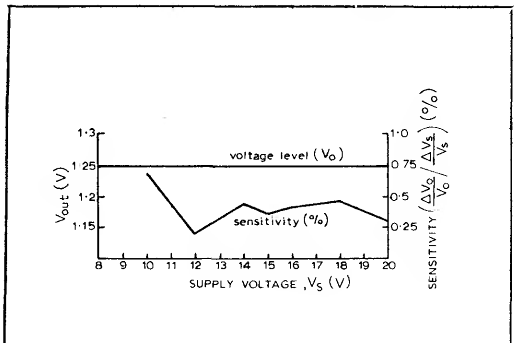


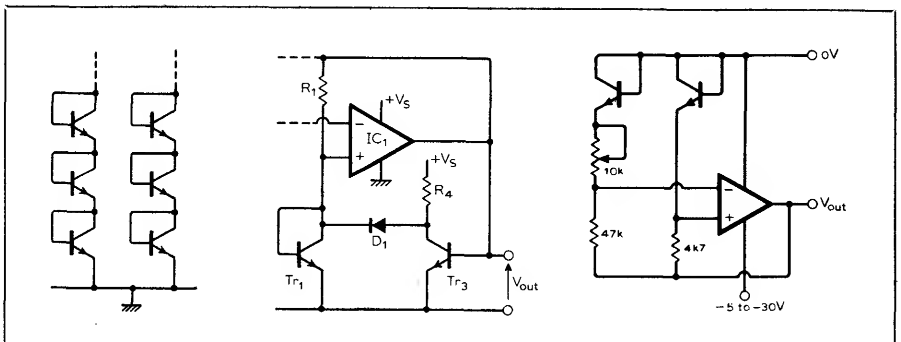
of operating from a single-ended supply and provides the facility of strobing such that a pulsed output is clamped between 0V (due to the c.m.o.s. output stage) and the temperature independent reference level. Centre voltage is the value at ambient, adjusted by R_3. Temperature range imposed on transistor package was +70° to −30°. Minimum drift obtained at $V_{out}\approx 1.25V$.

Component changes
Supply voltage 10V-19V maximum. Change in $V_{out}=0.75\%$.
Op-amp has heavy negative feedback and hence output impedance is low. This allows full current capability of op-amp to be drawn.
Typically
$I_{out}=0$, $V_{out}=1.251V$
$I_{out}=10mA$, $V_{out}=1.249V$
Maximum $I_{out}=22mA$.
Percentage sensitivity graph based on initial supply voltage $V_S=+9V$. Effective over load current range 0 to 20mA.
Bipolar-op-amp using positive and negative power supplies would improve overall drift, once offset is nulled.
Increasing R_1 to 10k, R_2 to 100k demands R_3 increased to 9k to maintain same reference level: i.e. values not too critical provided same ratio maintained.

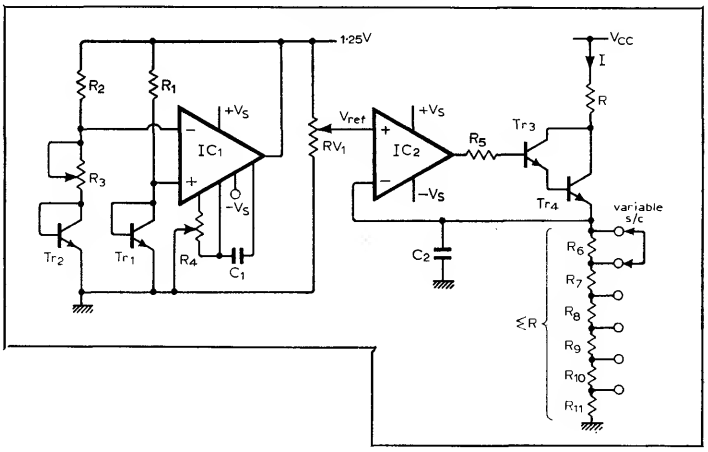
Circuit modification
Transistors can be series connected to increase available reference voltage—see middle e.g. eight diodes per chain will provide output ≈ 10V (Ref. 1). This will also allow a bipolar op-amp to be used from a single-ended supply because the inputs are held well above ground potential. Certain operational amplifiers will operate with inputs close to the most positive supply rail. This permits the dual configuration on diagram extreme right. V_{out} −1.25V.
If self-starting difficulties occur the circuit shown middle right can be used. When the supply is switched on, the output of IC_1 may remain at zero volts and hence there is no supply for transistors Tr_1 and Tr_2. With the addition of diode D_1 and transistor Tr_3, D_1 will conduct if V_{out} is low and Tr_3 is therefore off. This means the collector of Tr_1 will rise and because it is connected to the non-inverting input of IC_1, V_{out} will then increase to bring transistor Tr_3 into conduction. Op-amp CA3130 has a strobe terminal which is connected to the gate of its c.m.o.s. inverter output stage. When this terminal is connected to V_S via an external gating network, e.g. $\frac{1}{3}$ CD4007, the pulsed reference facility is obtained.

Further reading
1. Kuijk, K. E. A precision reference voltage source. IEEE Journal of Solid State Circuits, Vol. SC-8, No. 3, June 1973.





μA—mA/mV—V/calibrator



Typical data
Supply ±10V, IC_1 3130AT
IC_2 741, Tr_1 to Tr_4 use CA3086
R_1 4.7kΩ, R_2 47kΩ, R_3 10kΩ pot.
R_4 100kΩ, R_5 1kΩ, R_6 100kΩ
R_7 10kΩ, R_8 1kΩ, R_9 100Ω
R_{10} 10Ω, R_{11} 1.11Ω, RV_1 10kΩ
C_1 1nF, C_2 10μF tantalum
R variable load. For measurement, standard resistance used (±0.05%) and five digit digital voltmeter across R.
V_{ref} Adjusted for 1.000V

Fig. 1

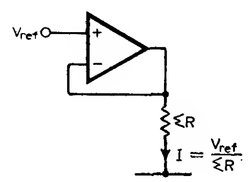
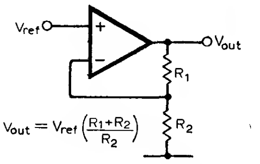
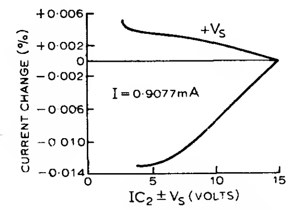
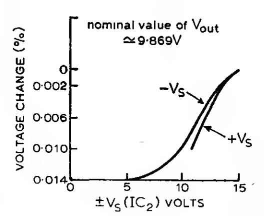


Fig. 2





Circuit description
When used as a current reference the circuit above simplifies to Fig. 1, and as a voltage reference, Fig. 2 shows the relationship between the output voltage and the input reference. The potential difference between the inverting and non-inverting inputs of IC_2 is very small for a high gain amplifier used in this negative feedback mode, and hence V_{ref} appears across the resistor chain R_6 to R_{11}. The current drawn from the V_{cc} supply depends on $V_{ref}/\Sigma R$, giving for the above values, a range from about 10μA to a safe maximum of 10mA (approx.) if R_6 - R_8 are shorted. Operation is such that if I did tend to increase, the voltage across ΣR increases, which would thus tend to increase the voltage applied to the non-inverting terminal of IC_2. This will reduce base drive to transistor Tr_3 and compensate for the assumed increase. Different values of current are achieved by varying RV_1 in association with varying the shorting points across ΣR.



Resistor R_{11} is chosen to provide an integer-multiplier for the voltage calibrator function. In this case the junction of R_6 and R_7 is connected to the inverting input of IC_2. This provides a ratio of exactly ×10 for the values chosen, and a case for 1% tolerance resistors is justifiable for an accurate source. Continuous variation of the output from 0-10V is available for this arrangement.

Component changes
Graph over indicates percentage current variation (100k and 10kΩ short-circuited) within ±0.01% for power supply variations of IC_2 up to 50% in $\pm V_S$.
Range of V_S 3.5 to 20V for above setting. Minimum value depends on current required and load R.
Graph left shows change of output voltage for 30% change in $\pm V_S$ is within −0.01%.
Change in V_S of IC_2 changes power dissipation of its input transistors and hence changes chip temperature. This slightly alters offset voltage due to unbalanced heating of input pair.

Fig. 3

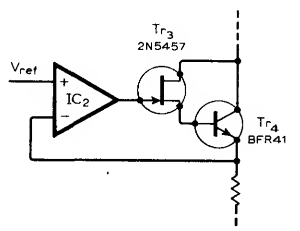


Fig. 5

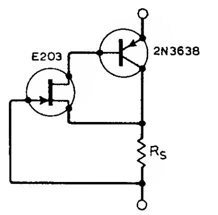
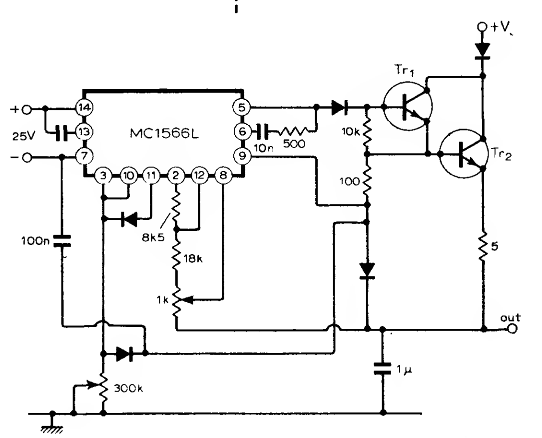


Fig. 4



Value of R_5 not critical because drive current is a small percentage of load current e.g. R_5 1kΩ, I_B 0.3μA, I_{LOAD} 10mA, R_5 100kΩ, I_B 0.365μA

Circuit modifications

- Drive current error can be minimized by using a f.e.t./bipolar combination as Fig. 3. Current capability can be increased provided appropriate transistors used for Tr_3 Tr_4 combination e.g. Darlington package.
- IC MC1566L allows a variable constant current adjustable from 200μA to 100mA (depending on rating of output transistor Tr_2).

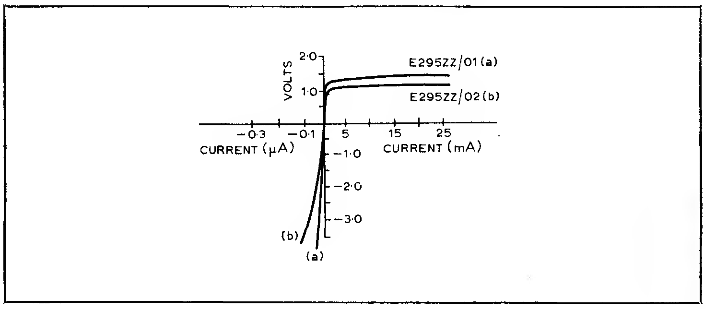
A simpler current reference, programmed by resistor R_S, is shown right and uses a combination of bipolar p-n-p and n-type junction f.e.t. to provide a low cost arrangement with claimed drift better than 0.03%/degC at 20mA.

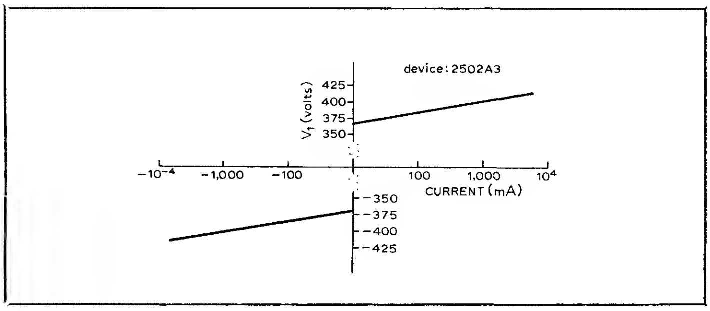
References
1. IC constant-current source at 750V, *Electronic Engineering*, May 1974.

Further reading
CA3130 Data Sheet 817, RCA.
Constant-current regulator 10m to 100mA, *Electronic Engineering*, July 1973.

Cross references
Set 23, cards 3, 5
Set 6, card 1

Non-zener device characteristics





1. Asymmetric voltage dependent resistors—low voltage
Non-linear resistors are made of certain polycrystalline materials having a voltage-current relationship given by

$V = CI^{\beta}$

where C and β are constants. This can be expressed as

$\log V = \beta \log I + \log C$

giving on log-log paper a straight line of slope β and intercept C.

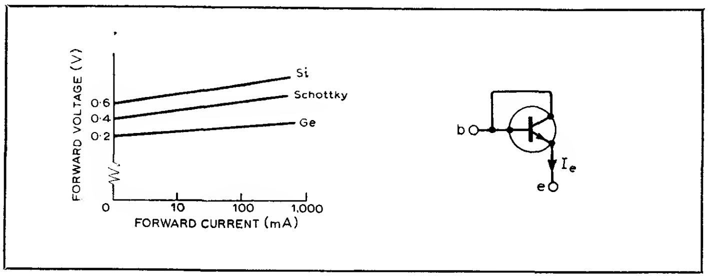
In the case of asymmetric devices β and C change with current direction and a zener diode type characteristic can be obtained as shown above. The particular devices shown have knee voltages intermediate between that of zener diodes and that of Si or Ge diodes. The temperature coefficient of forward voltage for both these types is −0.2% per degC maximum.

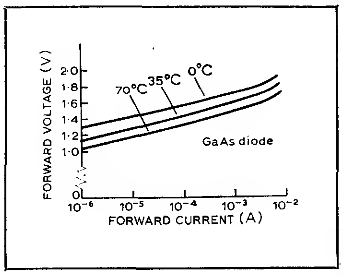
2. Symmetrical voltage dependent resistors—high voltage
These devices have the same form of relationship as that of the asymmetric device but are essentially simpler in that C and β do not change with current direction. They are frequently used as transient suppressors or over voltage protection devices in high power systems but may be used to regulate supplies. They are also used in some control applications where the non-linear characteristic is used intentionally. The characteristics shown are those obtained for a Z502A3 metal-oxide varistor. The temperature stability is claimed to be excellent, our device at 500μA on test giving a temperature coefficient of −0.13mV per degC.

3. Semiconductor diodes
All semiconductor diodes are governed by the same form of exponential equation over a wide range of currents. Below is shown the characteristic curves that are obtained as a result of replotting current on a log. scale. The slopes differ according to the material used as do the approximate constant voltages obtained across the diodes when conduction starts. The voltage obtained falls in general by 60mV for each decade drop in current although falls of up to 120mV per decade of current can occur at very high or very low values of current. Again for semiconductor junctions the temperature drift is of the order of −2mV per degC, no matter the material, although this increases at lower current densities.
The range of voltages to be expected from Si diodes is 0.5 to 0.8V and from Ge diodes is 0.1 to 0.3V. Schottky diodes are intermediate.
The diode connected transistor shown above will exhibit an exponential relationship between I_e and V_{be} over a much wider range of currents than is usual with simple diodes. The exponential relationship falls off at very low and very high currents due to loss of gain since basically it is I_c and V_{be} which are exponentially linked and the relationship between I_e and V_{be} being exponential depends on I_b being negligible.

4. Light emitting diodes
Light emitting diodes are semiconductor diodes which tend to be used for their light emitting properties rather than for their other properties which are identical in form to those of any semiconductor diode.

They happen to have larger knee voltages than Si etc diodes, values in the range from 1.5V to 2.2V being common for medium current operation (several mA usually). This knee value is reduced by the normal 60mV per decade of current. Low current operation may extinguish the light but does not alter the simple exponential action. As before the temperature drift is −2mV per degC.
The graph shows the results from a GaAs l.e.d. Higher knee voltages are obtained with the other common l.e.d. material viz GaAsP.





Compensated reference circuits

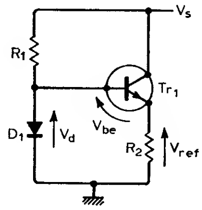


Fig. 1

Fig.2(a)

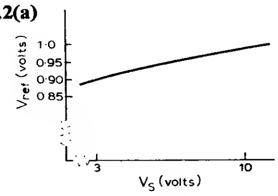


Fig.2(b)

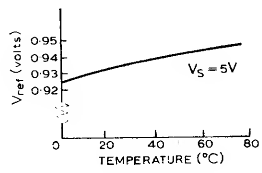
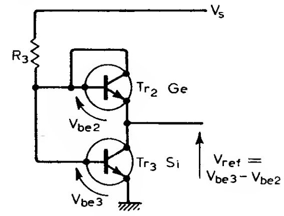


Fig. 3



Circuit 1

Components

R_1 470Ω, R_2 220Ω
Tr_1 BC125
D_1 5082-4850 (Hewlett Packard)
Performance—see graphs 1 and 2

Description

V_{ref} is given by $-V_{be}+V_d$ i.e. $-0.7+1.6=0.9$V to a first order approximately. The variation of V_{ref} with respect to supply voltage changes is shown in Fig.2(a). From this we observe a variation of approximately 10% in V_{ref} over the range shown so that compensation with respect to V_s is poor. The graph is approximately linear with slope +12mV per volt of V_s. Fig.2(b) shows the variation of V_{ref} with temperature, V_s being 5V. The graph is linear in the range shown with slope +0.27mV/degC. The variation of an independent p-n junction with temperature is approximately −2mV/degC. Clearly there is an element of temperature compensation involved, arising from the fact that V_{ref} is the difference in two junction voltages. In this case the effect of temperature variation on V_{be} is greater than the effect on V_d. For any junction the variation of junction voltage with temperature is dependent on the voltage itself and is approximately +3μV/degC for each change of +1mV. Hence, increasing the l.e.d. current and/or decreasing the transistor current could improve the drift due to temperature. This can be done by decreasing R_1 and/or increasing R_2.

Component changes

For any given V_s, R_1 and R_2 are not critical if only first-order temperature compensation is required. R_1 effectively controls the l.e.d. current since the transistor base current will be negligible and obviously R_1 should not be so high as to prevent l.e.d. conduction. Likewise R_2 carries the transistor current and should not be so low as to cause saturation of the transistor. Hence R_1 and R_2 are dictated by the particular l.e.d. and Tr_1 used.
Most l.e.ds and Si transistors will give the same reference voltage but if the transistor is replaced by a Ge transistor V_{ref} will become 1.3V approximately.

Circuit 2

The first circuit is not well compensated for supply voltage changes. In Fig. 3 Tr_2 is a Ge transistor and Tr_3 is a Si transistor and, hence, V_{ref} is approximately 0.4V—giving the circuit the added attraction of being a very low voltage reference. Ref. 1 quotes regulation 1% over supply current ranges of 100:1; the reason for this excellent regulation is that both transistors carry the same current so that the effects on V_{be} due to different currents (or different values of V_s) are the same for both transistors.
For the same reasons as before the circuit is also temperature compensated and indeed if the difference in the extrapolated band gaps at 0K is 0.43V complete temperature compensation is obtainable (Ref. 2). Achieving this requires selection of appropriate transistors.

Fig. 4

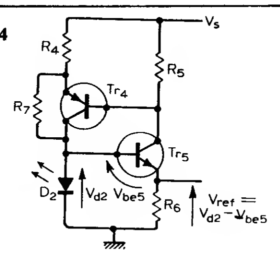
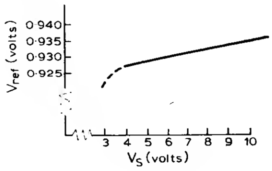


Fig. 5



Circuit 3

Components

R_4, R_6 220Ω
R_5 560Ω, R_6 22kΩ
Tr_4 BC126, Tr_5 BC125
D_2 5082-4850 (HP)

Performance

Fig. 5 shows the regulation achievable with this circuit viz 1.5mV per volt of V_s in the range 4 to 10V. For lower values of V_s saturation of the transistors occurs.

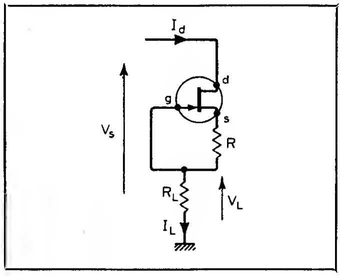
Description

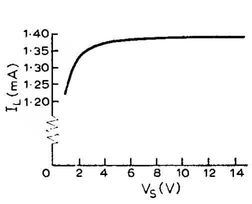
Fig. 4 shows a different approach to achieving simultaneous temperature compensation and supply voltage insensitivity. In this case the l.e.d. and transistor of Fig. 1 are incorporated in a ring-of-two reference circuit (Ref. 4). As in Fig. 1, V_{ref} is given by $V_d - V_{be}$ so that similar ambient temperature compensation is effected. In addition however, the current through Tr_4 is largely independent of the supply voltage so that effects due to varying V_s are minimal. The action of the ring-of-two circuit is independent of V_s and is briefly as follows. Suppose D_2 is made to conduct (insured by presence of R_4 and R_7). Then V_{d2} is approximately 1.6V and Tr_5 will conduct with V_{be} approximately 0.7V. The current in R_7 is, therefore, defined by $V_{d2}-V_{be5}$ and this current all flows through R_5 (ignoring base currents). This defines the voltage across R_5 and this voltage and the V_{be} of Tr_4 defines the current in R_4 which is the current in D_2 (ignoring current in R_7). Hence, once D_2 conducts the current through it is fixed and so are all the other currents. Hence $V_{d2}-V_{be5}$ is supply insensitive. Only the transistor collector-emitter voltages are supply dependent.

References

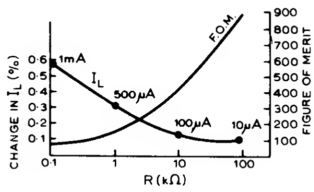
1 Very low voltage d.c. reference, P. Williams, *Electronic Engineering*, June 1968, pp. 348-349.
2 National Semiconductor LM311 voltage comparator data sheets.
3 Set 6, card 5

Simple current reference

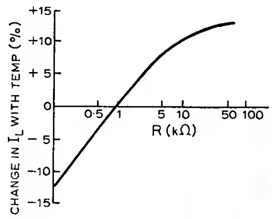




Graph 1



Graph 2



Graph 3

FET 2N5457

Performance

Graph 1 was obtained with $R_2 = 10\Omega$ and $R = 0$

Graph 2 was obtained with $R_2 \approx 1k\Omega$. V_s was set at 10V and I_L noted; V_s was then changed to 15V and the new I_L noted. From this was obtained the graph of % change in I_L for various values of R, the marked values of I_L on the graph being those at $V_s \approx 10V$. The figure of merit (f.o.m.) shown is defined as the slope resistance $\Delta V_s/\Delta I_L \div$ static resistance (V_s/I_L at $V_s=10V$).

For a perfect current source this should be ∞ (see main text). As R increases the quiescent current I decreases but the f.o.m. increases, so we see clearly that the arrangement works best as a current source at low values of I_L. Note that from the graph of % change in I_L we can deduce the static and slope resistances which ranged from $8.1k\Omega$ and $794k\Omega$ to $1M\Omega$ and $10^3M\Omega$ respectively over the range of R shown.

Graph 3 shows the percentage change in I_L, as a result of a temperature change from −5°C to +70°C, for various R. This was obtained with $R_L=1k\Omega$ and $V_s=11.4V$ (a choice intended to produce 10V across R_L with $R=0$, but of no material significance). In this respect it should be noted that although V_s is constant neither V_{dg} not V_L is constant because of the varying I_L. The main point, however, is that a temperature independent condition is achieved at approximately $R=1k\Omega$ for the particular f.e.t. that we used. Whilst graph 3 shows the effect of a large temperature change, results were taken at intermediate temperatures. These indicated that to a first approximation, for any given R, the change in I_L is proportional to the change in temperature.

Circuit description

No matter the value of R, the gate current in a f.e.t. is negligible so that I_L and I_d can be equated. With $R=0$ V_{gs} is zero and graph 1 is simply the normal f.e.t. characteristic for this value of V_{gs}. With increasing R more and more negative current feedback is used and consequently, the circuit behaves more like a current source with the ensuing drop in output as graph 2 shows.

Some explanation of the figure of merit used seems in order because some people use different figures of merit (Ref. 1 & 2) and notably slope resistance is at least implied as being all important. Consider Fig. 1 and the device characteristics shown, all passing through the same operating point, X. From this most people would agree that
OCD = perfect voltage source
OFG = imperfect voltage source
OAB = perfect current source
OHJ = imperfect current source
OE = indeterminate case.

From this we conclude that for a current source then (1) the slope resistance must be greater than the static resistance (HJ closer to horizontal than OE) (2) given 1 then the greater the slope resistance the better.

From this, one is tempted to conclude that slope resistance is all important. Consider, however, Fig. 2 where we have the same slope resistance but different static resistances at points P and Q. Is the device whose characteristic is ORPQ a better current source at P or at Q? At P we find that the angle between the given characteristic and the indeterminate case (OP) is β and clearly $\beta > \gamma$ i.e. at P we have a better current source. We could propose the figure of merit $(\alpha+\beta)\div\alpha$, which becomes larger as the device approaches a perfect current source but is rather devoid of meaning. However is we take $\tan(\alpha+\beta)\div\tan\alpha$ we obtain the same overall picture with the merit that $\tan(\alpha+\beta)\div\tan$ = slope resistance ÷ static resistance. Since slope resistance and static resistance are meaningful we adopt this as our figure of merit. The inverse of this we would adopt for a voltage source since we want any f.o.m. to become bigger the more the source resembles a perfect source.

With regard to the temperature dependence of the circuit one should note that to a first order approx. the temperature independent condition is, in any f.e.t, obtained when $V_{gs}=V_P+0.63$

For our circuit this can be achieved when

$$R=\frac{|V_p-0.63|}{I_0}\times\frac{V_p^2}{0.63^2}$$

where V_p is the pinch-off voltage and $I_0=I_d$ when $V_{gs}=0$. The usefulness of these formulae is restricted because of the large range of V_p and I_0 quoted by the manufacturer for any device and the difficulty of measuring both for any single device. However, Graph 3 has been plotted over a very large range of R and consequently tends to exaggerate the effect of temperature.

References

1 I.R.C. zener diode handbook.
2 Motorola zener diode handbook.

Fig. 1

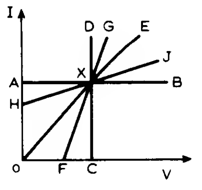
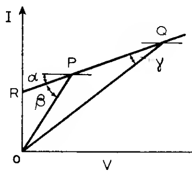


Fig. 2



Reference circuits

Fig. 1

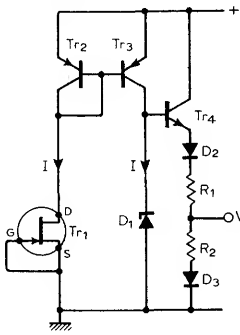
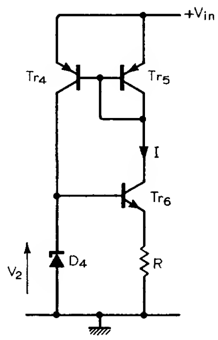


Fig. 2

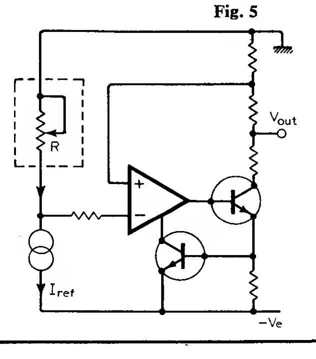


Circuit description

Internal reference circuits are based on the simpler voltage reference shown in Fig. 1, found in the low-cost LM376 voltage regulator. The self-starting technique used here employs a f.e.t. Tr_1, with gate-source connected to ensure a constant current from drain to source. Tr_1 will draw current initially via the base emitter-junctions of transistors Tr_2 and Tr_3. These transistors are therefore turned-on, and due to the current-mirror connection the current I through avalanche diode D_1 is defined, and hence its voltage. This turns on transistor Tr_4 and now produce a current through the D_2, R_1, R_2, D_3 chain. The circuit has no feedback loop, and an alternative is shown in Fig. 2. As it stands, this is not inherently self-starting, but with Tr_4 conducting, the voltage V_z developed across avalanche diode D_4, will maintain a constant current through resistor R. The collector current I of Tr_6 is mirrored in the collector of Tr_4 again to maintain a constant current through D_4 thus completing the loop. This is shown more detailed in Fig. 3, reference circuit contained in the LM100 voltage regulator. The multi-collector transistor Tr_8 in conjunction with Tr_7, Tr_9 forms a closed loop. The V_{BE}'s of these transistors will be essentially similar, and their collector currents closely related, but dependent on relative collector area, but if one collector current is stabilized the others must be. If I is assumed to increase, the base drive of Tr_9 would increase and hence divert current away from Tr_7, since current through R is constant $(V_X - V_Y)/R$. This would imply a reduction in the base current of Tr_8, which negates the original assumption. D_5 in association with $2V_{BE}$ drops across Tr_8, Tr_7 provide self-starting. The reference terminal can have a 0.1µF to ground to bypass possible noise from zener diode D_7. An alternative low impedance voltage reference point is available if the output and feedback terminal via a 2.2kΩ resistor. The internal circuit in principle is shown in Fig.4 where 5 and 6 are the base terminals of transistors connected as a long-tail pair. Diode connected transistors Tr_{11}, Tr_{12} and resistors R_1, R_2, R_3 provide a temperature compensation which is optimized to about 1.8V. Fig. 5 is a function block for the LM104 negative voltage regulator which uses an internal current reference network. An external 2.4kΩ resistor provides a current I_{REF} of 1mA. When this conducts through an external programming resistor R, the output voltage available is regulated up to zero volts.

Type	Nominal ref. voltage (V)	Standby current (mA)	Temperature range (°C)	Input voltage range (V)	Temperature stability (%)
LM100	1.7	1	−55 to 125	8.5 to 40	0.3
LM300	1.7	1	0 to 70	8.5 to 30	0.3
LM305	1.7	0.8	0 to 70	8.5 to 50	0.3
LM304	—	1.7	0 to 70	−8 to −40	0.3
LM309	5	0.5	0 to 70	7 to 35	—
LM723	7.15	1.3	−55 to +125	9.5 to 40	coeff. 0.002%
LM723C	7.15	1.3	0 to 70	9.5 to 40	coeff. 0.003%
MC1723	7.15	2.3	0 to 75	9.5 to 40	coeff. 0.002%
MCC1463	−3.5V	—	0 to 75	up to −40	
MCC4060A	4.1	3.7	−10 to +75	9 to 38	coeff. 0.003%

Fig. 5



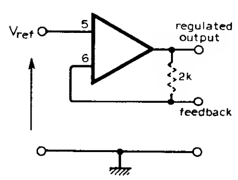
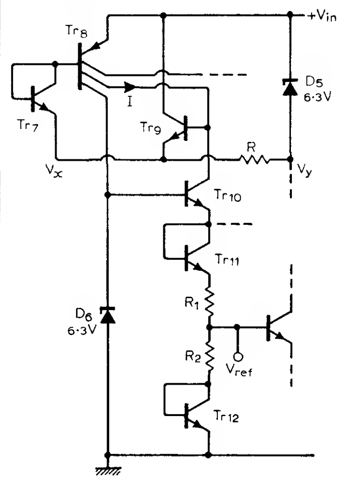


Fig. 4

Fig. 3



If the output voltage is a linear function of temperature, then the temp-coefficient of voltage is a constant, and the total change in voltage (often called the stability) is the product of temperature coefficient and the total temperature change $dV = (\delta V/\delta T)dT$

The temperature coefficient of voltage changes with temperature because it results from a number of different sources of drift, and the actual $\delta V/\delta T$ at a given temperature might be greater or smaller than the average value implied by overall stability figure.

Further reading

Linear applications—National Semiconductor Corporation.

Linear Integrated Circuits—National Semiconductor Corporation.

Cross reference

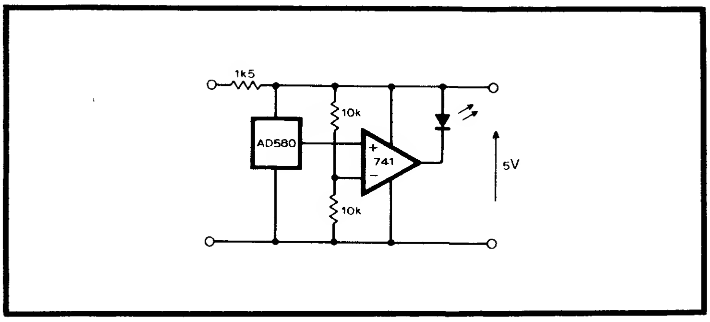
Set 23, card 1.

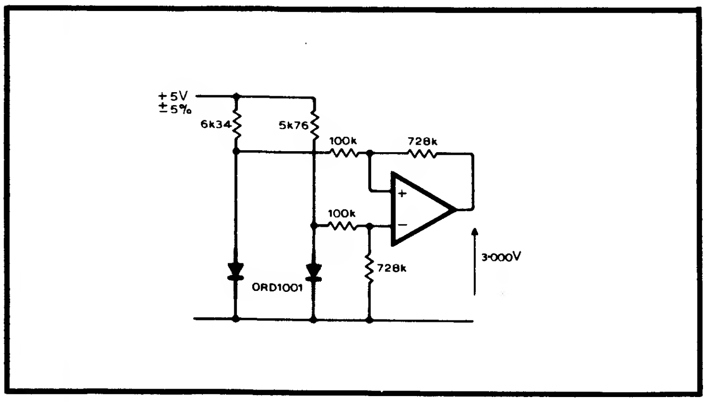
wireless world circard

The strongest trend in reference designs in recent years has been that toward band-gap techniques. This is particularly true for low-voltage applications where zener diodes cannot be used. When dissimilar junctions or similar junctions with different current densities are used to generate a reference voltage, then temperature-independence is observed when that voltage is related to the band-gap voltages of the semiconductor materials. Two related approaches led to three-terminal reference elements as in the circuit shown or two-terminal devices such as LM113 or ZN423. These last-mentioned have a terminal voltage equal to the band-gap of silicon, about 1.2V. The three terminal devices can accept a wide range of input voltages with a fixed output voltage scaled up from the band-gap voltage internally. The output of the device shown is 2.5V but 5V and 10V versions have also been reported. The circuit shows a simple means of converting to other output voltages with the additional advantages that (i) both reference unit and amplifier are supplied from the stabilized output markedly reducing supply dependence (ii) visual indication is provided by the l.e.d. whose main purpose is to keep the op-amp output terminal within its linear range.

Reference

Jung, W. C. Programmable voltage reference is stable yet simple, *EDN*, Nov. 5, 1975, pp. 99/100.





One special device that has been reported, contains a pair of silicon p-n junctions. One is so heavily doped that its band-gap voltage is reduced to that of germanium. The voltage difference is then temperature stable at around 0.41V, the difference between silicon and germanium band-gap voltages. The configuration of the circuit is not particularly novel, but the temperature coefficient is < 10p.p.m. K^{-1}, over the full military temperature range.

The Si-Ge band-gap circuit using only two transistors is the simplest capable of negligible variation against both supply and temperature changes. It requires a match between the Si and Ge transistors such that ΔV_{BE} equals the band-gap difference of ~0.43V. This is an acceptable limitation in return for its simplicity and very low voltage requirement (<1V). If the transistor current densities can be adjusted then any pair of transistors can be used without prior matching. The circuit exploits this idea with the 20k potentiometer being adjusted to force 0.43V across the 100Ω resistor hence meeting the stability condition. Although the example shown is for a 5V output this can be reduced, and there is the further advantage that the temperature coefficient can be made either positive or negative as required by readjusting the current densities.

References

Verster, T. C. Regulated low-voltage power supply with controllable temperature coefficient, *Rev. Sci. Instrum.*, vol. 44, August 1973, pp. 1127/8.

Dubow, J. Low-voltage reference is ultra-stable, *Electronics*, Feb. 7, 1974, pp. 133/4.