A Simple Three-Terminal IC Bandgap Reference

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Abstract—A new configuration for realization of a stabilized bandgap voltage is described. The new two-transistor circuit uses collector current sensing to eliminate errors due to base current. Because the stabilized voltage appears at a high impedance point, the application to circuits with higher output voltage is simplified. Incorporation of the new two-transistor cell in a three-terminal 2.5-V monolithic reference is described. The complete circuit is outlined in functional detail together with analytical methods used in the design. The analytical results include sensitivity coefficients, gain and frequency response parameters, and biasing for optimum temperature performance. The performance of the monolithic circuit, which includes temperature coefficients of 5 ppm/°C over the military temperature range, is reported.

I. INTRODUCTION

THE REQUIREMENT for a stable reference voltage is almost universal in electronic design. The temperature-compensated avalanche breakdown diode fills many of the needs, but cannot be used with low voltage supplies and often suffers from long-term stability problems. Use of a transistor base emitter diode temperature compensated to the bandgap voltage of silicon is a technique which overcomes some of the avalanche diode limitations. Bandgap circuits can be operated from low voltage sources and depend mainly upon subsurface effects which tend to be more stable than the surface breakdowns generally obtained with avalanche diodes.

The conventional three-transistor bandgap cell works well for very low voltage two-terminal or "synthetic Zener diode" requirements. The three-transistor cell is less flexible in three-terminal applications and in circuits where the desired output is not an integral multiple of the bandgap voltage.

The two-transistor cell presented here is simpler, more flexible in three-terminal applications, and eliminates sources of error inherent in the three-transistor cell. The two-transistor cell offers separate control over output voltage and temperature coefficient in a circuit using only a single control loop.

The new bandgap circuit has been used as the basis of a monolithic three-terminal reference circuit supplying a stable 2.5-V output and operating down to 4-V input.

II. CONVENTIONAL CIRCUIT

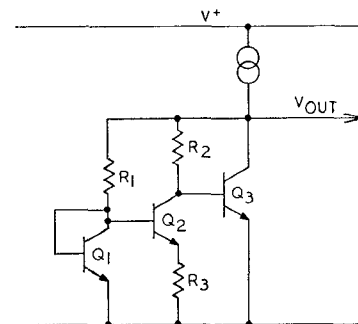Conventional bandgap circuits are based on the concept illustrated in Fig. 1. The transistors Q_1 and Q_2 are

Fig. 1. Conventional bandgap circuit.

operated at different current densities to produce temperature proportional voltages across R_3 and R_2. A third transistor Q_3 is used to sense the output voltage through R_2. As a result, Q_3 drives the output to a voltage which is the sum of its V_{BE} and the temperature-dependent voltage across R_2. When the output voltage is set to approximate the bandgap voltage of silicon, the voltage across R_2 will compensate the temperature coefficient of V_{BE}, and the output voltage will be temperature invariant [1]. To minimize the output voltage temperature coefficient the collector current of Q_3 must be made proportional to temperature, as are the currents in Q_1 and Q_2. This large temperature-dependent current at the point where the stabilized voltage appears makes it inconvenient to produce an output greater than the bandgap voltage. Higher voltages can be generated by stacking several junctions to produce, in effect, several circuits like Fig. 1 in series [2].

The theory used to predict the temperature behavior of circuits like Fig. 1 neglects the effect of base current flowing in R_1 and R_2. The variability in this current due to processing and temperature effects on h_{FE} gives rise to an output voltage error and drift. This effect is particularly severe when the current in Q_2 is made much smaller than currents in Q_1 and Q_3 to produce the required current density difference. Use of "Super Beta" processing to reduce this problem results in low-voltage transistors not suitable for a three-terminal reference.

An additional temperature stability problem arises out of the nonlinearity and nonuniformity of the temperature characteristics of diffused resistors. The nonlinearity cannot easily be compensated, and the nonuniformity cannot be accommodated in the design.

The idealized circuit shown in Fig. 2 minimizes the difficulties of obtaining outputs above the bandgap voltage, reduces the problem of h_{FE} variability to one of α match, and can be implemented with thin-film resistors on the monolithic chip to virtually eliminate nonlinear
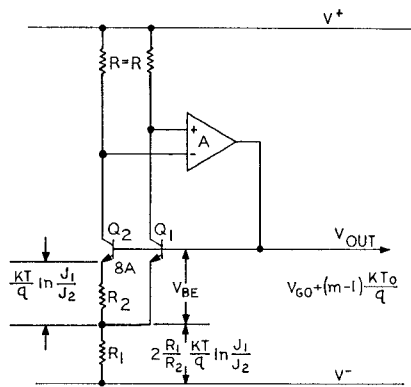
Fig. 2. Idealized circuit illustrating two-transistor bandgap cell.

temperature coefficients of resistance (TCR) as an error factor. The circuit uses two transistors and collector-current sensing to establish the bandgap voltage. The voltage appears at active transistor (as opposed to diode-connected) bases, so that it is a straight forward and simple matter to obtain overall output voltages above the bandgap voltage.

III. BASIS OF THE NEW CONFIGURATION

### A. Generating the Bandgap Voltage

In the circuit of Fig. 2 the emitter area of Q_2 is made larger than that of Q_1 (by a ratio of 8-to-1 in the example given). When the voltage at their common base is small, so that the voltage drop across R_2 is small, the larger area of Q_2 causes it to conduct more of the total current available through R_1. The resulting imbalance in collector voltages drives the op amp so as to raise the base voltage. Alternatively, if the base voltage is high, forcing a large current through R_1, the voltage developed across R_2 will limit the current through Q_2 so that it will be less than the current in Q_1. The sense of the collector voltage imbalance will now be reversed, causing the op amp to reduce the base voltage. Between these two extreme conditions is a base voltage at which the two collector currents match, toward which the op amp drives from any other condition. Assuming equal α or common-base current transfer ratio for Q_1 and Q_2, this will occur when the emitter current densities are in the ratio 8-to-1, the emitter area ratio.

When this difference in current density has been produced by the op amp, there will be a difference in V_{BE}, between Q_1 and Q_2, which will appear across R_2. This difference will be given by the expression

$$\Delta V_{BE} = \frac{kT}{q} \ln \frac{J_1}{J_2}. \tag{1}$$

Since the current in Q_1 is equal to the current in Q_2, the current in R_1 is twice that in R_2 and the voltage across R_1 is given by

$$V_1 = 2 \frac{R_1}{R_2} \frac{kT}{q} \ln \frac{J_1}{J_2}. \tag{2}$$
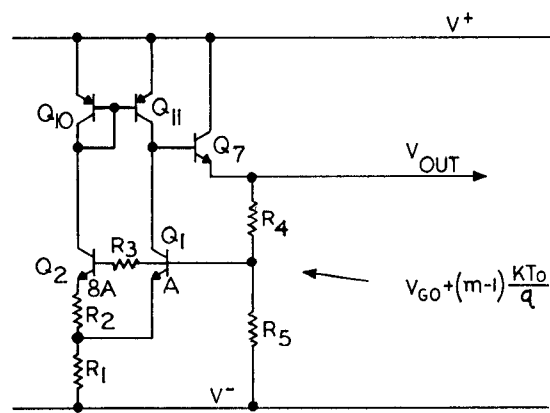
Eliminate error due to base current in R_4 by setting:

$$R_3 = \frac{R_2}{R_1} \frac{R_4 R_5}{R_4 + R_5}$$

Fig. 3. Simplified circuit for developing higher output voltages.

Assuming that the resistor ratio and current density ratio are invariant, this voltage varies directly with T, the absolute temperature. This is the voltage which is used to compensate the negative temperature coefficient of V_{BE}.

The voltage at the base of Q_1 is the sum of the V_{BE} of Q_1 and the temperature-dependent voltage across R_1. This is analogous to the output voltage of the conventional bandgap circuit and can be set, by adjustment of R_1/R_2, to a temperature stable value, as described in the Appendix.

B. Increasing the Stabilized Output Voltage

Assuming that the amplifier of Fig. 2 has sufficient gain, it will balance the collector currents of Q_1 and Q_2 despite an additional voltage drop added between its output and the common-base connection. This additional drop will not affect the base voltage which results in collector current balance. If the voltage is introduced by means of a resistive voltage divider, the op amp output voltage will be proportional to the common-base voltage.

The circuit of Fig. 3 uses an active load to sense the collector current of Q_1 and Q_2 more directly. The function of the op amp is replaced by Q_{10}, Q_{11}, and Q_7. The p-n-p transistors form a simple current mirror which takes the difference of the collector currents of Q_1 and Q_2. This difference current drives the base of Q_7 which supplies the circuit output voltage. This voltage is divided by R_4 and R_5 and applied to the base of Q_1. The sense of the signal to Q_7 drives Q_1 and Q_2 to minimize the collector current difference. By designing the circuit to stabilize the base voltage at the bandgap voltage the output will be stabilized at a higher voltage. Since the output voltage depends upon R_4 and R_5 it can be set to any convenient value and need not be an integral multiple of the bandgap voltage.

C. Base Current Error Correction

In the circuit of Fig. 3 the base current of Q_1 and Q_2 must flow through R_4. This current will require an increase above the nominal output voltage to bring the base of Q_1 to the proper level. This increase will be an h_{FE} dependent output voltage error which will vary from lot to lot and drift with temperature. The effect can be minimized by using relatively low values for R_4 and R_5, or R_3 can be added to compensate the effect. The proper value of R_3 is given by the following analysis.

To simplify the analysis, neglect the effects of finite h_{FE} and output conductance in Fig. 3 to idealize the performance of the Q_{10}, Q_{11}, Q_7 amplifier function.

If E is taken to be the circuit output voltage in the absence of base current for Q_1 and Q_2, then E' resulting from considering R_3 and the two base currents is given by

$$E' = E + R_4(i_{b1} + i_{b2})$$
$$- i_{b2}R_3(2R_1/R_2)(1 + R_4/R_5). \quad (3)$$

This relation contains a term due to the base currents through R_4 and an offsetting term due to reduction of ΔV_{BE} by base current through R_3. If E' is set equal to E, (3) can be reduced to a constraint on R_3. Expressing the relationship between the base currents in terms of a parameter $P_1 = i_{b1}/i_{b2}$ permits (3) to be reduced to

$$R_3 = (P_1 + 1)R_2R_4R_5/2R_1(R_4 + R_5). \quad (4)$$

In the case shown in Fig. 3 the collector currents and hence the base currents are assumed to match, making P_1 equal to 1 and resulting in the reduced expression shown in the figure. The general form of (4) is useful in circuits where the current density ratio is controlled by forcing unequal collector currents, rather than by emitter area ratios.

IV. PRACTICAL REALIZATION OF THE CONCEPT

Although the circuit of Fig. 3 can be used in some simple applications a number of factors limit its applicability. Base width modulation of Q_1 and Q_2, finite output impedance of Q_{11} and the finite h_{FE} of Q_7, Q_{10}, and Q_{11} all combine to raise the circuit's dynamic impedance and to degrade its input-voltage rejection. The configuration shown in Fig. 4 reduces these problems and has been built and tested in monolithic form.

The circuit elements shown in Fig. 4 correspond, roughly, to similarly numbered elements in the other figures. The basic two-transistor bandgap cell consisting of Q_1, Q_2, R_1, and R_2 is the same in all the figures. In Fig. 4 the current mirror transistor Q_{10} and Q_{11} are bootstrapped to the output voltage to improve the supply voltage rejection of the circuit. Degeneration resistors R_7 and R_8 have been added to raise the output impedance of Q_{11} and improve the emitter current match of the pair. To minimize the effect of p-n-p h_{FE} Q_8 drives the common bases of Q_{10} and Q_{11}. The output control voltage is picked off at the collector of Q_{11} by Q_9. A level translator consisting of Q_{12}, R_6, and Q_3 then applies the output
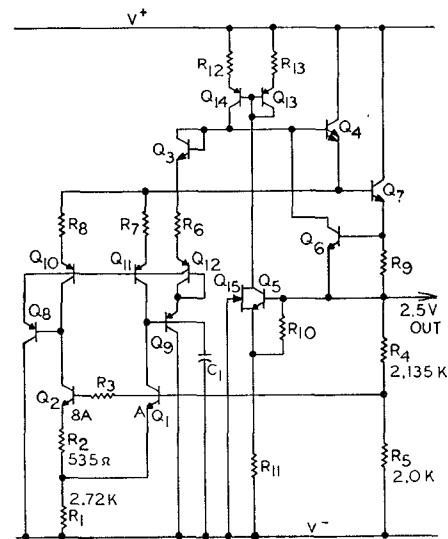

Fig. 4. Complete circuit of the monolithic three-terminal reference chip.

signal to the base of Q_4. This transistor forms a Darlington connection with Q_7, the output transistor, and provides the current-mirror bootstrap voltage as well.

The circuit, as shown in Fig. 3, has a stable "off" or no-current state. An epitaxial layer FET, Q_{15}, is incorporated into the circuit of Fig. 4 to provide starting. The FET insures that a minimum current flows into the current mirror Q_{13} and Q_{14} even when the base voltage of Q_1 and Q_2 is zero. This current is "reflected" by Q_{13} and Q_{14} to drive Q_4 and turn on the circuit. Once the circuit is on, the collector current of Q_{14} becomes nearly equal to the currents in Q_{10} and Q_{11}. As a result, the V_{BE} of Q_3 is equal to that of Q_4 which has twice the emitter area and supplies roughly twice the current. Moreover, the voltage drop across R_6 (which is equal in resistance to R_7 and R_8) is made equal to the voltage drops across R_7 and R_8. The emitter of Q_{12} is the same size as the emitters of Q_{10} and Q_{11} so that they all operate at the same current density and provide nearly equal emitter voltages for Q_8 and Q_9. The collector of Q_{12} is split to provide equal emitter currents for Q_8 and Q_9.

The operating bias level for Q_{14} is controlled by Q_5. This transistor is matched to Q_1 and has its emitter current forced by R_{10} and R_{11}. These resistors are in inverse ratio to R_4 and R_5 so that their open-circuit equivalent voltage with respect to the output is the same as the voltage across the base emitter of Q_1 in series with R_1. Their parallel resistance is twice the resistance of R_1 so that the current in Q_5 is matched to the current in Q_1. The fraction of the current bypassed by Q_{15} has a negligibly small effect on the V_{BE} of Q_5 and hence on the total current forced by R_{10} and R_{11}. The total current through Q_5 and Q_{15} drives Q_{13}. The resistors R_{12} and R_{13} raise the output impedance and improve the current matching of the simple current mirror Q_{13} and Q_{14}.

Output current limiting is provided by R_9 and Q_6.

$$g_m = \frac{di_o}{dE}\Big|_{i_o=0} = \frac{1}{R_1}\left(\frac{\ln(J_1/J_2)}{2+\ln(J_1/J_2)}\right)$$

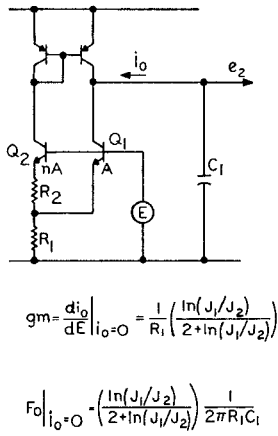$$F_0\Big|_{i_o=0} = \left(\frac{\ln(J_1/J_2)}{2+\ln(J_1/J_2)}\right)\frac{1}{2\pi R_1 C_1}$$

Fig. 5. Transconductance and frequency compensation model.

V. FREQUENCY COMPENSATION

The amplifier in this circuit operates in a closed loop to regulate the output voltage. A composite junction-MOS capacitor C_1 is used to control the open-loop cross-over frequency and stabilize the closed-loop response. The analytical basis for this compensation is illustrated by Fig. 5. When the two-transistor bandgap cell is operated into a current mirror an output current is produced whenever the common-base voltage departs from the nominal voltage determined by the current density ratio and by R_1 and R_2. The change in this current as a function of the departure of base voltage from its nominal value has the dimensions of transconductance and can be used as such in design. The following incremental approximation gives a simple result which is more than adequate for most design procedures.

Incremental changes in the base voltage of Q_1 give rise to changes in collector current which can be approximated by the ratio of the voltage change to R_e, the incremental emitter resistance. This same voltage increment also drives R_2 and the incremental impedance of Q_2. If the transistors are operating at equal currents, the two R_e terms will be equal, making the total effective resistance in the Q_2 branch higher. This will result in a lower incremental current in Q_2. Equating the incremental base voltage changes gives

$$\Delta i_{e_1} R_{e_1} = \Delta i_{e_2}(R_{e_2} + R_2). \tag{5}$$

Substituting for R_e and for R_2 in terms of the voltage across it and current through it converts (5) to

$$\Delta i_{e_1}\frac{kT}{qi_{e_1}} = \Delta i_{e_2}\left(\frac{kT}{qi_{e_2}} + \frac{kT}{qi_{e_2}}\ln\frac{J_1}{J_2}\right). \tag{6}$$

A second approximation made is that the total incremental current is due to the voltage change across R_1 resulting from a voltage change ΔE at the common bases. That is

$$\Delta i_{e_1} + \Delta i_{e_2} = \frac{\Delta E}{R_1}, \tag{7}$$

which neglects R_e as compared with the value of R_1. After eliminating common factors and assuming $i_{e_1} = i_{e_2}$, the combination of (6) and (7) can be manipulated to form

$$\frac{\Delta i_{e_1} - \Delta i_{e_2}}{\Delta E} = \frac{1}{R_1}\left(\frac{\ln(J_1/J_2)}{2+\ln(J_1/J_2)}\right). \tag{8}$$

The difference in the Q_1, Q_2 collector currents is the output current i_o and is approximately given by the difference of the emitter current increments. Taking (8) to the limit at $i_o = 0$ yields

$$\frac{d_{i_o}}{dE}\Big|_{i_o=0} = \frac{1}{R_1}\left(\frac{\ln(J_1/J_2)}{2+\ln(J_1/J_2)}\right). \tag{9}$$

With a current density ratio of 8:1, the term $\ln J_1/J_2$ is approximately 2, so that the "transconductance" of the entire circuit is approximately $1/(2R_1)$.

A capacitive load on the current output of the circuit in Fig. 5 will give a 6 dB/octave rolloff of voltage transfer ratio. The frequency at which the capacitive reactance equals the transconductance will be the unity-gain frequency of the simple circuit. This is given by the expression for F_0 in the figure. In the circuit of Fig. 4, the loop attenuation due to R_4 and R_5 reduces the overall unity-gain frequency by the ratio of the bandgap voltage to 2.5 V, which is approximately two.

The transconductance can also be used to estimate a low frequency "gain." In the simple circuit of Fig. 3, the gain is expressed as the ratio of the voltage at the base of Q_7 to a small-signal input applied to the base of Q_1 at balance. Using a value of 3 kΩ for R_1 and estimating the output impedance of Q_{11} at about 300 kΩ gives a gain of about 50. In the monolithic circuit, the effective open-loop gain is increased several orders of magnitude by the bootstrap connection to the current mirror.

VI. MONOLITHIC CIRCUIT PERFORMANCE

The circuit of Fig. 4 is shown in Fig. 6 as it appears in monolithic form. Several diffusion lots have been made and measurements of these units indicate the typical properties given by the following table.

Typical Reference Circuit Parameters
(−55 to +125°C)

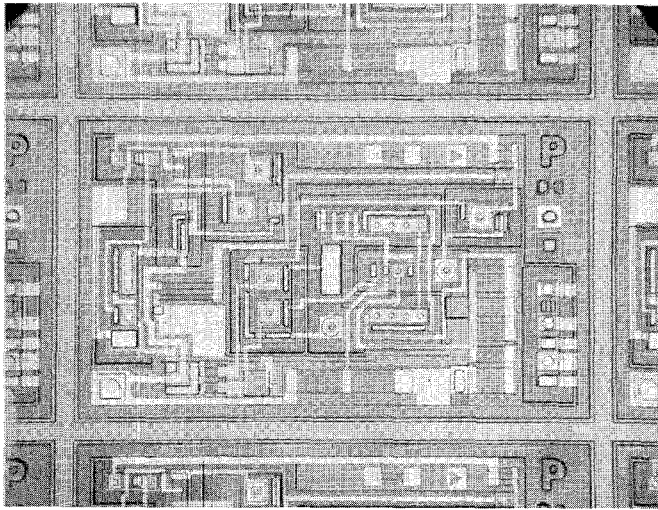|	
Output voltage	2.5 V ±2 percent
Minimum input voltage	4 V
Load regulation, 0 to 10 mA	3 mV
Supply rejection, 4.5 to 7 V	0.25 mV
Supply rejection, 7 to 30 V	0.25 mV
Standby current	1 mA
Output voltage temperature coefficient (γ) $\left(\gamma = \frac{(V_{max} - V_{min})}{(V_{nominal})(\Delta T)}\right)$	5 to 60 ppm/°C

The observed variation in temperature coefficients

Fig. 6.   Illustration of 37 × 62 mil monolithic die showing thin-film resistors and balanced thermal layout.

arises mainly from variability in absolute voltage at the base of Q_1 and variation in the coefficient m (see Appendix). Units showing a very constant temperature coefficient generally show a good correlation between absolute output voltage and drift.

Several units showing large but constant temperature coefficients have been adjusted to very small final temperature coefficients. This has been done by measuring the temperature coefficient calculating the ideal zero temperature coefficient voltage correction and laser-trimming R_1 or R_2 as required. Other units which are closer to the nominal output voltage exhibit very stable temperature characteristics initially. In extremely low-drift units, the performance of the basic cell appears to be masked by other drifts arising in the remainder of the circuit. These residual drifts are on the order of 2 to 4 ppm/°C over a temperature range of −55 to +125°C.

Some diffusion lots have shown a greater curvature in the temperature characteristic. These units exhibit the roughly parabolic temperature characteristic, which is implied by (14) of the Appendix. They show good temperature performance around a peak which occurs at a temperature related to initial output voltage. At temperature extremes the temperature coefficient may increase to 60 ppm/°C or more, if the peak is not centered in the −55 to +125°C range.

The elapsed time since obtaining the first completed units has not been sufficient to accumulate long-term drift results. Accelerated life tests have been made at high temperatures to uncover any gross drift problems. The temperature stability and monitoring equipment have not been adequate to determine the ultimate stability of the device. Examination of data taken over the course of 1000 hours at +125°C does not reveal any trends or systematic drifts at the 100 ppm level, which approximates the repeatability of the measurements.

APPENDIX

THEORETICAL TEMPERATURE BEHAVIOR

A. Optimum Cell Voltage

The optimum voltage at the base of Q_1 is approximately the bandgap voltage. A general analysis of the two-transistor bandgap cell yields a more-exact result and some insight into the residual temperature drift in an optimally adjusted circuit.

As a matter of convenience, the circuit description has involved the assumption that the required current density ratio is established by the use of emitter area ratio alone. The circuit can also be based on different collector currents and equal areas or a combination of unequal areas and currents.

The equation describing the voltage across R_1 can be generalized by including a parameter P_1 which is the ratio of emitter currents i_{e_1}/i_{e_2}, as follows:

$$V_1 = (P_1 + 1) \frac{R_1}{R_2} \frac{kT}{q} \ln \frac{J_1}{J_2}. \tag{10}$$

An expression given by Brugler [3] has been modified by the addition of a current-dependent term to give the V_{BE} of Q_1 as

$$V_{BE_1} = V_{go}\left(1 - \frac{T}{T_o}\right) + V_{BE_o} \frac{T}{T_o}$$
$$+ \frac{mkT}{q} \ln \frac{T_o}{T} + \frac{kT}{q} \ln \frac{J}{J_o} \tag{11}$$

where V_{go} is the bandgap voltage of silicon and V_{BE0} is the value of V_{BE_1} at a reference temperature T_o.

Since the voltages across both R_1 and R_2 are proportional to temperature, it follows that the current and the current density in Q_1 are also proportonal to temperature. Therefore, the current density ratio in the last term of (11) can be equated to a temperature ratio as $J/J_o = T/T_o$. This relation can be used to reduce the sum of V_1 and V_{BE_1} to the form

$$E = V_{go} + \frac{T}{T_o}(V_{BE_o} - V_{go}) + (m - 1) \frac{kT}{q} \ln \frac{T_o}{T}$$
$$+ (P_1 + 1) \frac{R_1}{R_2} \frac{kT}{q} \ln \frac{J_1}{J_2}. \tag{12}$$

This represents the stable voltage established at the base of Q_1 in the circuit of Fig. 1. Differentiating this result twice with respect to temperature yields

$$\frac{dE}{dT} = \frac{1}{T_o}(V_{BE_o} - V_{go}) + (P_1 + 1) \frac{R_1}{R_2} \frac{k}{q} \ln \frac{J_1}{J_2}$$
$$+ (m - 1) \frac{k}{q} \ln \frac{T_o}{T} - 1 \tag{13}$$

and

$$\frac{d^2E}{dT^2} = -(m - 1) \frac{k}{q} \frac{1}{T}. \tag{14}$$

Equating the first derivative to zero results in the equation

$$V_{BE_o} + (P_1 + 1) \frac{R_1}{R_2} \frac{kT_o}{q} \ln \frac{J_1}{J_2} = V_{go} + (m - 1) \frac{kT_o}{q}. \tag{15}$$

The left side of (15) is the value of E at $T = T_o$. This means that, in principle, if the base voltage of Q_1 is set to $V_{go} + (m - 1) kT_o/q$ at temperature T_o, the temperature coefficient of the output voltage will be zero. Assuming values of m greater than one in (14) implies, however, a nonzero temperature coefficient at temperatures other than T_o. Experimental data indicate that values of m as low as 1.2 have been achieved.

Examination of (13) and (15) shows that departures from the optimum output voltage will result in an approximately constant temperature coefficient. The magnitude of this coefficient will be the absolute error voltage divided by the absolute temperature. For example, a 3 percent absolute voltage error at 300 K will result in a 3 percent/300 K = 100 ppm/°C temperature coefficient.

B. Effects of Resistor Temperature Coefficients

Both the common and differential temperature coefficients of resistance of R_1 and R_2 enter into the overall output voltage stability. The differential TCR may be treated by assuming that the total change in resistor ratio is due to change in R_1. Since the remainder of the circuit forces a predetermined current in R_1, its temperature coefficient of resistance is translated to an error voltage in equal proportion. The effect on the overall output voltage is reduced by the fact that the voltage across R_1 accounts for only a fraction of the total output. This fraction is variable, but is about 1/2 at room temperature. Therefore, the differential TCR of R_1 and R_2 appears as about half the equivalent proportional change in the output voltage.

The common TCR affecting both resistors equally does not affect the voltage across R_1, which depends only upon their ratio. It does affect the total current in the transistors, however, and therefore affects the V_{BE} of Q_1.

This effect can be evaluated by expressing V_{BE_1} as

$$V_{BE_1} = \frac{kT}{q} \ln \left(\frac{i_{e_1}}{I_s}\right). \tag{16}$$

Differentiating

$$\frac{dV_{BE_1}}{dR_2} = \frac{kT}{q} \frac{I_s}{i_{e_1}} \left(\frac{-i_{e_1}}{R_2 I_s}\right) = \frac{-kT}{qR_2} \tag{17}$$

and

$$\frac{dV_{BE_1}}{dR_2} R_2 = \frac{-kT}{q} \tag{18}$$

which at room temperature is about -26 mV or $-260 \mu V/$percent. Dividing this coefficient by the cell's output voltage of about 1.22 V gives a relative coefficient of about -2.13×10^{-4}/percent of output voltage change as a function of relative resistance change. This is a reduction in sensitivity of about 47 times.

Despite this reduction, the TCR of diffused resistors is so large, nonlinear, and nonuniform that it still represents a serious temperature drift. The monolithic circuit described here uses thin-film-on-silicon resistors with a total TCR of about -60 ppm so that the 47 times reduction makes the uncorrectable component of TCR sensitivity less than 1 ppm/°C.

The effects of the TCR of other resistors, or of ratios such as R_4 and R_5, are readily evaluated by standard analytical techniques. The low differential TCR inherent in thin-film resistor pairs keeps these effects on the output voltage at a minimum.

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