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A TEMPERATURE-COMPENSATED, LOW-VOLTAGE REGULATOR

David C. Auth

30 April 1964

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David C. Auth

FOR THE COMMANDER:
Approved by

P. E. Landis
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CONTENTS

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ABSTRACT

A solid-state, low-voltage regulator accurate to 0.1 percent with temperature stability has been developed. Employing thermistor compensation, forward-biased diodes, and matched silicon transistors, it represents a significant advance in the state-of-the-art. The circuit discussed herein yields 1.500 ± 0.001 v at 1 to 15 ma with input voltages of 8 to 12 v over the temperature range -10° to $+40^{\circ}C$.

Circuit design and derivations for temperature compensation using thermistors and silicon resistors are covered in detail along with general design criteria for low-voltage regulators.

1. INTRODUCTION

When at the request of NASA,HDL undertook to develop a durable solid-state voltage regulator of 0.1-percent accuracy over the temperature range -10° to $+40^{\circ}C$ with an output of 1.5 v, it was apparent that several innovations were necessary. Regulation to 1 percent is generally considered sufficient. Zener diodes, standard cells, and constant-current sources are the usual reference elements for the regulator designer.

The lower limit of breakdown voltage for zener diodes is 3.3 v. Unless use is made of a difference voltage between two zener diodes (involving a floating ground), the zener diode is not suitable for low-voltage ($\sim$ 1.5 v) regulation. With the further restriction of durability, standard cells and constant-current sources are eliminated also.

The need for a new voltage reference became increasingly apparent. The stabistor (i.e., forward-biased silicon reference diode) with a voltage range of 0.2 to 1.4 v was selected. The selection involved a serious compromise in performance. The stabistor exhibits a higher 3 to 6X) dynamic impedance than that of the zener diode; and consequently, its voltage varies more with equal current fluctuations. In addition, the stabistor has a relatively high negative temperature coefficient of voltage. These problems were solved by the use of sufficient amplification in the regulator feedback loop and by temperature compensation via a thermistor in the sampling stage.

2. THE SOLID-STATE VOLTAGE REGULATOR

The solid-state voltage regulator can be of two general types, series or parallel. The parallel type is highly inefficient because it relies upon a shunting current path (in parallel with load) to increase the IR drop in the voltage source and hence to effect voltage regulation. For this reason, the series type was used in the circuit described herein.

In the series regulator (fig. 1), a variable resistance is placed in series with the load. This series resistance is then varied according to the variations of load current and voltage input. A transistor is used as the series resistance and its control (resistance) is based upon the output voltage level. The series transistor must be able to dissipate power equal to the difference in voltage of the input and output voltage times the maximum load current i.e.

$$[E_{IN} - E_{OUT}] \; [I_L \text{ (max)}] = \text{power dissipated} \tag{1}$$

The series transistor is controlled by feedback from the sensing stage of the regulator, where a comparison is made between a known portion E_S of the output voltage E_O and the reference voltage E_{REF}:

$$E_S = E_O \left(\frac{R_{10}}{R_{10} + R_9} \right) \tag{2}$$

If, for example, the output voltage tends to increase, the base-emitter bias E_{BE} of the series transistor Q_1 is made to decrease, thereby raising the resistance (collector-emitter) of Q_1 and lowering the output voltage.

3. OBTAINING ACCURATE LOW-VOLTAGE OUTPUT

3.1 Differential Amplifier (ref 1) and Sampling Circuit

In the circuit of figure 1, Q_3 begins to turn on when $E_S > E_{REF}$ and regulation of the output voltage occurs. The output voltage is related directly to E_S by the equation

$$E_O = E_S \left(\frac{R_{10} + R_9}{R_{10}} \right) \qquad \text{(cf eq 2)} \tag{3}$$

It is desirable to have the output voltage depend on E_{REF} in the following manner

$$E_O = E_{REF} \left(\frac{R_{10} + R_9}{R_{10}} \right) \tag{4}$$

However, it is apparent from figure 1 that

$$E_S = E_{REF} + E_{BE} \; (Q_3) \tag{5}$$

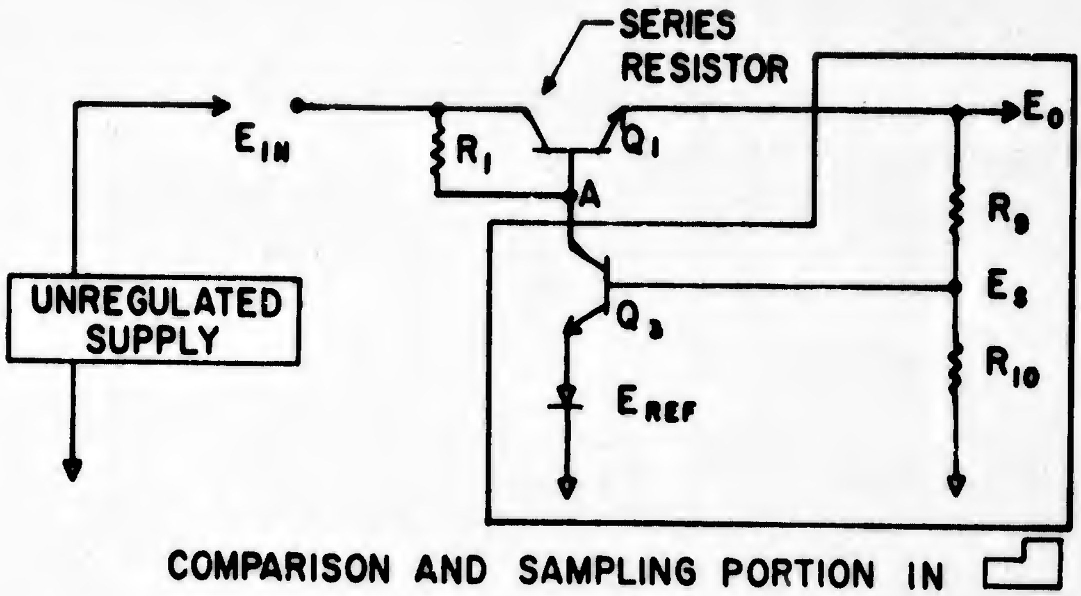

Figure 1. A series-type, solid-state, voltage regulator where Q_1 is the series control element.

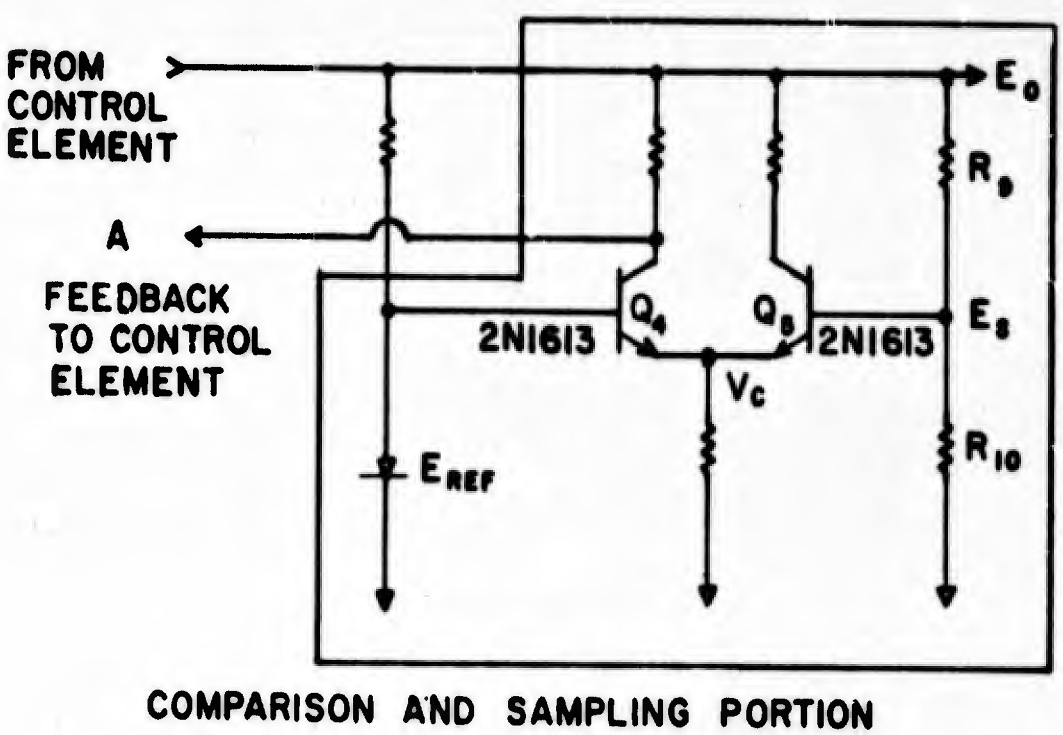

Figure 2. Differential amplifier comparator circuit.

where $E_{BE\ (Q_3)}$ is the base-emitter voltage of transistor Q_3. If

$$E_{BE\ (Q_3)} \ll E_{REF},\ E_{REF} \approx E_S \tag{6}$$

If equation 6 is applicable, equation 4 holds as a result of equation 3. In low-voltage regulators $E_{REF} \sim E_{BE} \sim 0.5$ v necessarily. In this case it is evident that equation 6 cannot hold because the restriction $E_{BE\ (Q_3)} \ll E_{REF}$ cannot be maintained. The output voltage for figure 1 is a function of $E_{BE\ (Q_3)}$ and E_{REF}; i.e.,

$$E_0 = \left[E_{REF} + E_{BE\ (Q_3)}\right] \left[\frac{R_{10} + R_9}{R_{10}}\right] \tag{7}$$

Thus, any variation of $E_{BE\ (Q_3)}$ due to current variation or temperature change (ref 2) will alter the output voltage. If equation 5 is modified as a consequence of changes in the sampling circuit to a differential amplifier scheme as shown in figure 2, then

$$E_S - E_{BE\ (Q_4)} = E_{REF} - E_{BE\ (Q_5)} \tag{8}$$

and if, at balance $E_{BE\ (Q_4)} = E_{BE\ (Q_5)}$, equation 8 reduces to equation 6 and equation 4 holds with no E_{BE} dependence. The circuit described by equation 8 (fig. 2) solves the above difficulty.

Since it is necessary that $E_{BE\ (Q_4)} = E_{BE\ (Q_5)}$, the choice of transistors Q_4 and Q_5 is critical. A matched pair of silicon planar transistors (2N1613) was selected; these transistors possess leakage currents (I_{CBO}) in the nanoampere range and E_{BE} tracking an order of magnitude better than do mesa units. The use of a differential amplifier with matched transistors is extremely important for highly stable, low-voltage regulators.

3.2 Feedback and Amplification

As prevously mentioned, sufficient amplification is required to offset the high (relative) dynamic impedance of the stabistor. The amplification is accomplished in two stages from point A (fig. 3) to the base of the series transistor Q_1. One stage consists of transistor Q_2 cascaded with the series transistor; the other stage is a simple

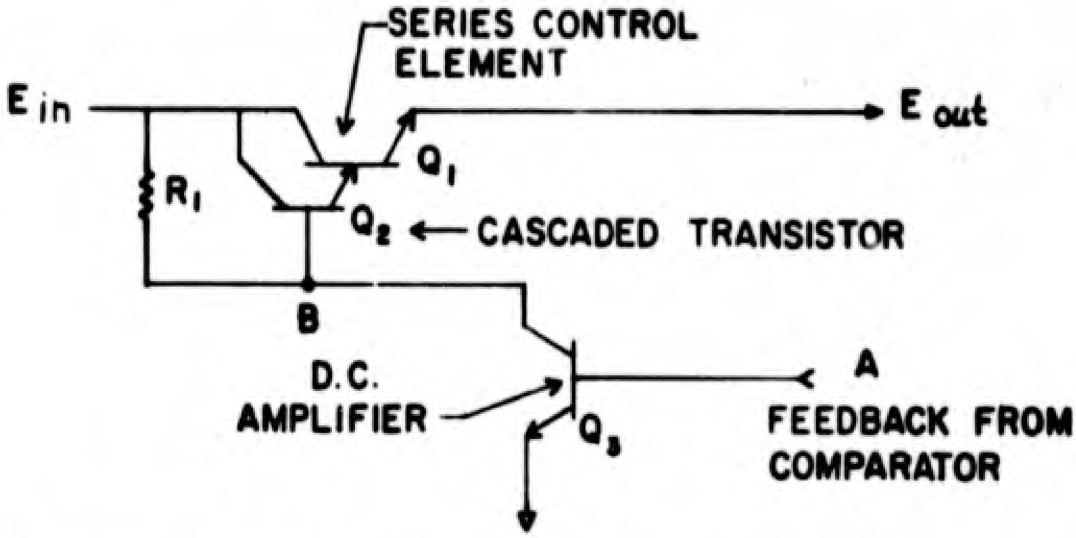

Figure 3. Feedback amplification with control element.

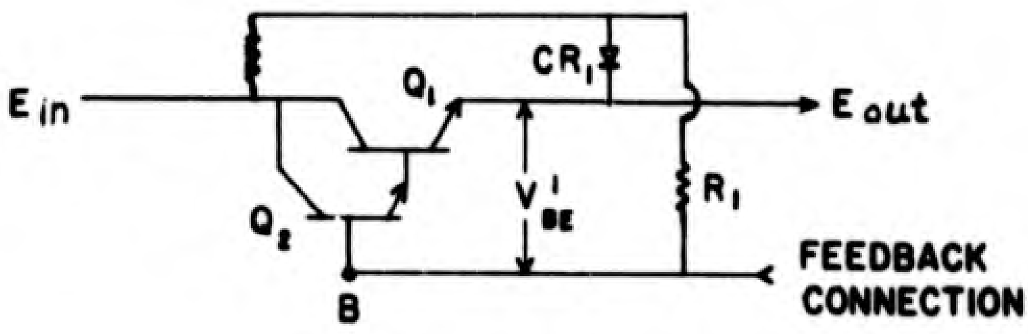

Figure 4. Preregulator circuit with series transistor.

d-c amplifier. The cascaded transistor, commonly termed a β multiplier (ref 3), greatly increases the effective β (amplification) of Q_1, the effect being a large change in the resistance (collector-emitter) of Q_1 with small voltage changes at point B. The d-c amplifier is a direct-coupled transistor Q_3 that operates into a relatively high load impedance (R_1) as a result of the presence of Q_2, which provides the bulk of the E_{BE} to Q_1.

3.3 Control Element with Preregulator

In the circuit of figure 4, it can be seen that the bias voltage applied to Q_2 and Q_1 is related to the voltage drop across Q_1 (collector-emitter) that corresponds to $E_{IN} - E_{OUT}$. Hence, since E_{OUT} remains constant (or very nearly so), the E_{BE} of Q_1 increases with an increase in E_{IN}. This tends to lower the resistance of Q_1 instead of raising it as would be desired with larger E_{IN}. By placing a stabistor in parallel with the Q_2 and Q_1 base-emitter cascade, the E_{BE} of Q_2 and Q_1 remains nearly constant with respect to $E_{IN} - E_O$. However, E_{BE} of Q_1 and Q_2 is still variable depending upon the feedback voltage at point B. The diode (stabistor) CR_1 (fig. 4) then tends to regulate the bias applied to Q_1; this diode configuration is referred to as a preregulator.

4. TEMPERATURE COMPENSATION

4.1 Compensation Technique

The stabistor obeys the diode equation (ref 4) to a good approximation; hence, for $V_D > \frac{kT}{q} = \frac{1}{40}$ v at 25°C

$$V_D = AkT \ \ln \frac{I}{I_S}$$

where

A = constant

k = Boltzmann constant

T = absolute temperature

$-q$ = charge of electron

I = diode current

I_S = saturation current

V_D = voltage drop across diode

Assuming that all terms remain constant except V_D and T, it is readily apparent that V_D is linearly dependent on T. The dependence is about 0.28 percent/°C for the stabistor at room temperature. Actually, I_S will change slightly with temperature. (A, k, q, and I remain constant with temperature change.) However, after the logarithmic operation is performed on I_S, its effect is negligible or nearly so with temperature variation.

From equation 4, $E_O = E_{REF} \dfrac{R_{10} + R_9}{R_{10}}$, it is evident that the negative temperature coefficient of E_{REF} will lower the output voltage with increasing temperature. It is equally apparent that a lowering of R_{10} will increase the output voltage. If R_{10} is temperature dependent with a negative coefficient, it is possible for the temperature effects of R_{10} and E_{REF} to exactly cancel, or provided the devices possess similar temperature dependence;

$$E_O = \text{constant} = E_{REF} \left\{ \frac{R_{10} + R_9}{R_{10}} \right\} = E_{REF}\,(1 - \alpha t) \left\{ \frac{R_9 + R_{10}\,(1 - \beta t)}{R_{10}\,(1 - \beta t)} \right\} \tag{10}$$

where R_{10} = resistance of R_{10} at 25°C.

$-\alpha$ = temperature coefficient of voltage for E_{REF} = - 0.28 percent/°C for stabistor

$-\beta$ = temperature coefficient of resistance for R_{10}

E_{REF} = reference voltage at 25°C

t = temperature change from 25°C

α is given and β must be determined for equation 10 to be consistent. Cancelling terms and expanding we have:

$$R_{10} + R_9 = \frac{R_{10} + R_9 - \beta t R_{10} - \alpha t\, R_9 - \alpha t\, R_{10} + \alpha t\, \beta + R_{10}}{1 - \beta t}$$

multiplying through by $(1 - \beta t)$ and cancelling:

$$-\beta t\, R_9 - \beta t\, R_{10} = -\beta t\, R_{10} - \alpha t\, R_{10} + \alpha t\, \beta t\, R_{10}$$

then,

$$-\beta t\, R_9 = -\alpha t\, R_9 - \alpha t\, R_{10} + \alpha t\, \beta t\, R_{10}$$

and

$$-\beta t\ (R_9 + \alpha t\ R_{10}) = -\alpha t\ (R_{10} + R_9)$$

$$\therefore\ \beta = \frac{\alpha(R_{10} + R_9)}{\alpha t\ R_{10} + R_9} \tag{11}$$

if $\alpha t\ R_{10} \ll R_9$

$$\beta \approx \frac{R_{10} + R_9}{R_9}\ (\alpha) \tag{12}$$

4.2 Using Thermistors

While thermistors (ref 5) do have a negative temperature coefficient of resistance, the thermistor resistance varies as exp E_g/kT (ref 6) where E_g is a constant. This exponential dependence upon T must be linearized if a thermistor is to be used to compensate for a linearly varying reference. The combining of a thermistor with a temperature-stable resistor (metallic film) in a parallel configuration as shown in figure 5B tends to linearize the temperature dependence of the compound resistance. This is illustrated graphically in figure 5 where a 400-ohm precision resistor has been added in parallel with a 100-ohm (25° C) thermistor. The deviation from a straight line in the range of -10° to +40°C is very slight--less than 10 percent maximum. In this particular case, the net linear temperature coefficient is 3.08 percent/° C (γ) at 25°C. In order to use this parallel combination for temperature compensation where a precise value of β is required, the temperature coefficient γ must undergo a "dilution" of effectiveness. This may be accomplished by the addition of a series resistor with the temperature sensitive parallel combination (fig. 5C). The total resistance (R_{10}) must have an effective temperature coefficient equal to β as given in equation 11, which demands an equation of the form

$$(1 - \gamma\ t)\ R_T + R_D = R_{10}\ (1 - \beta t) \tag{13}$$

where

R_D = diluting resistance

R_T = resistance of parallel combination

$R_{10} = R_D + R_T$

then $\quad -\gamma\ t\ R_T + R_T + R_D = R_{10} - \beta\ t\ R_{10}$

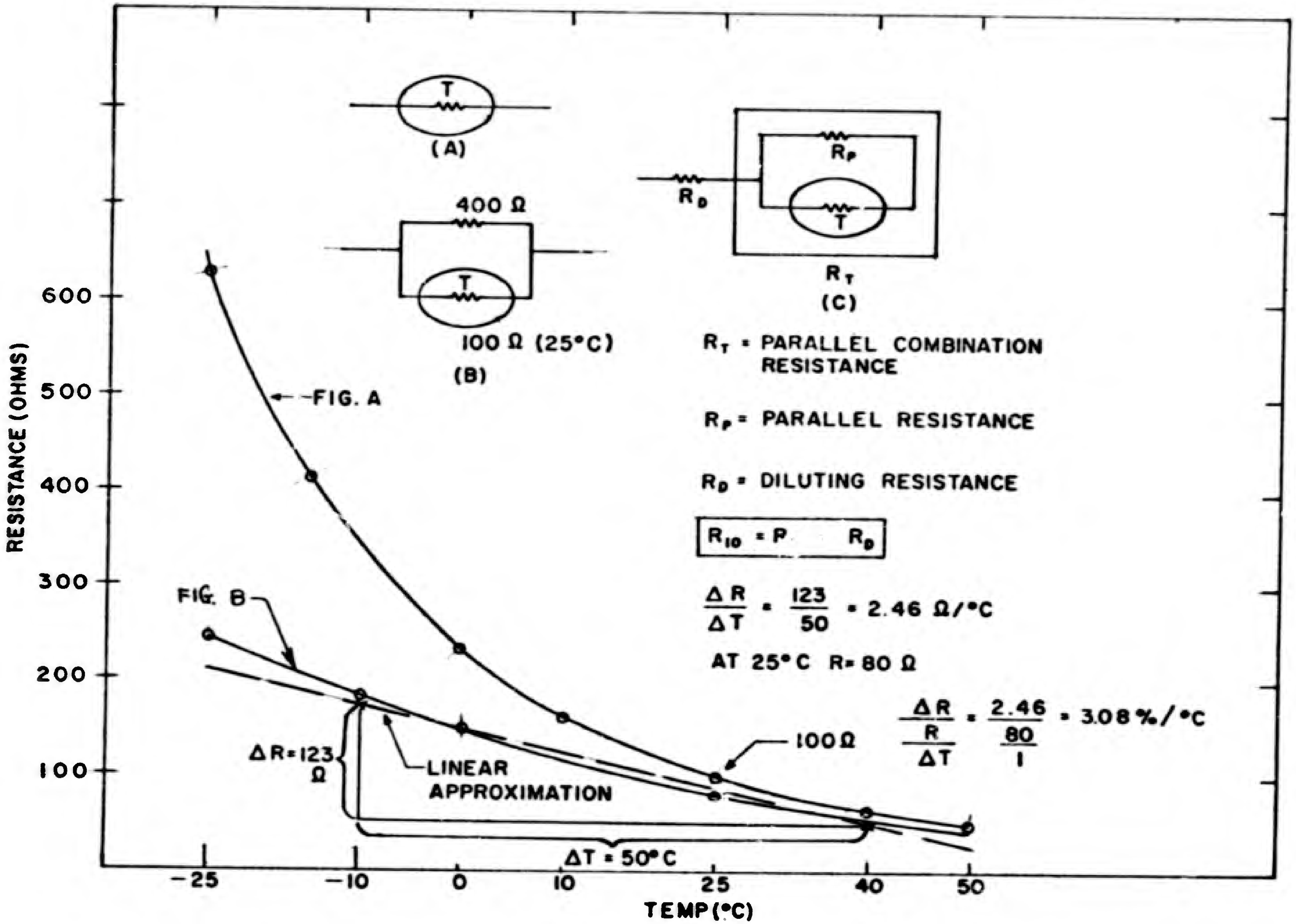

Figure 5. Resistance versus temperature of precision thermistor with and without parallel linearizing resistor.

Since $R_{10} = R_D + R_T$

$$-\gamma R_T = -\beta R_{10} = -\beta(R_T + R_D)$$

$$\frac{-\gamma}{-\beta} = \frac{R_T + R_D}{R_T} = 1 + \frac{R_D}{R_T}$$

so
$$\frac{R_D}{R_T} = \frac{\gamma}{\beta} - 1 \text{ or } R_D = R_T\left(\frac{\gamma}{\beta} - 1\right) \quad (14)$$

Thus, the diluting resistance is defined in terms of the known quantities R_T, γ, and β.

For a typical situation, R_D will be 10 times R_T; hence, the variation of the temperature coefficient of R_{10} from a straight line is less than 1 percent maximum. If the reference diode is strictly linear with temperature, the deviation of R_{10} from a straight line will induce a change in the output voltage of the same order of magnitude as the deviation. In actual tests the deviation was found to be ~ 0.1 percent for the output voltage. This better-than-expected stability may be attributed to the use of a stabistor that exhibited a slightly nonlinear temperature coefficient; resulting in a fortuitous coincidence of effects yielding the excellent temperature stability shown in figure 8 of section 5.

4.3 Using Silicon Resistors

A second order coincidence in temperature dependence is not essential to achieve accurate compensation (~ 0.1 percent). The silicon resistor can be made (by appropriate doping, etc.) to possess a positive, linear coefficient of temperature dependence. It may be used directly in the upper portion of the sampling circuit (R_9 in fig. 6) without the addition of a parallel linearizing resistor. The silicon resistor is inserted in the position of R_9 instead of R_{10} (as for thermistor compensation) because of its positive rather than negative temperature coefficient. The derivation for positive temperature coefficient device compensation is similar to that used in equation 11. The result is

$$\delta = \left\{\frac{E_0}{E_0 - E_{REF}} \left(\frac{\alpha}{1 - \alpha}\right)\right\} \quad (15)$$

where

δ = temperature coefficient of R_9

α = stabistor temperature coefficient (absolute value)

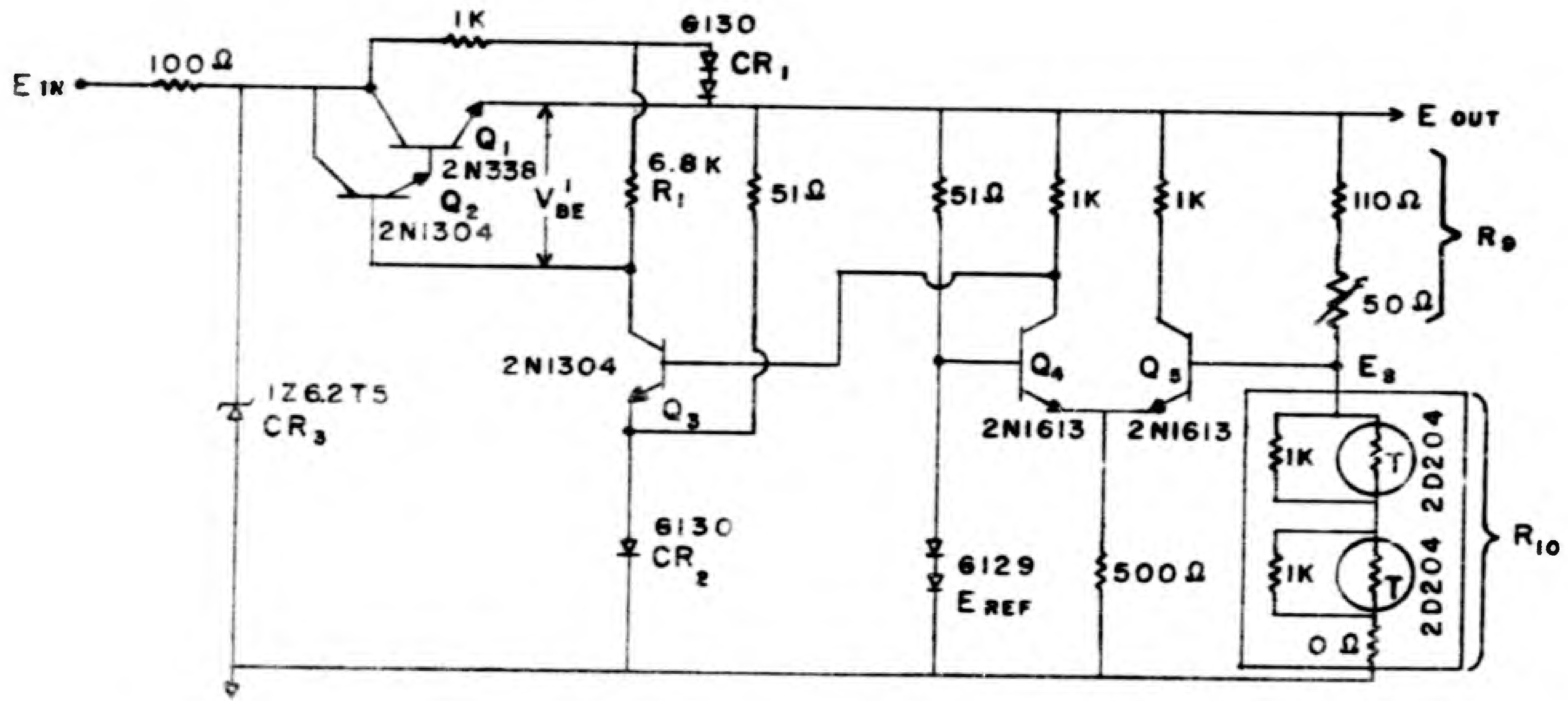

Figure 6. Complete voltage regulator design.

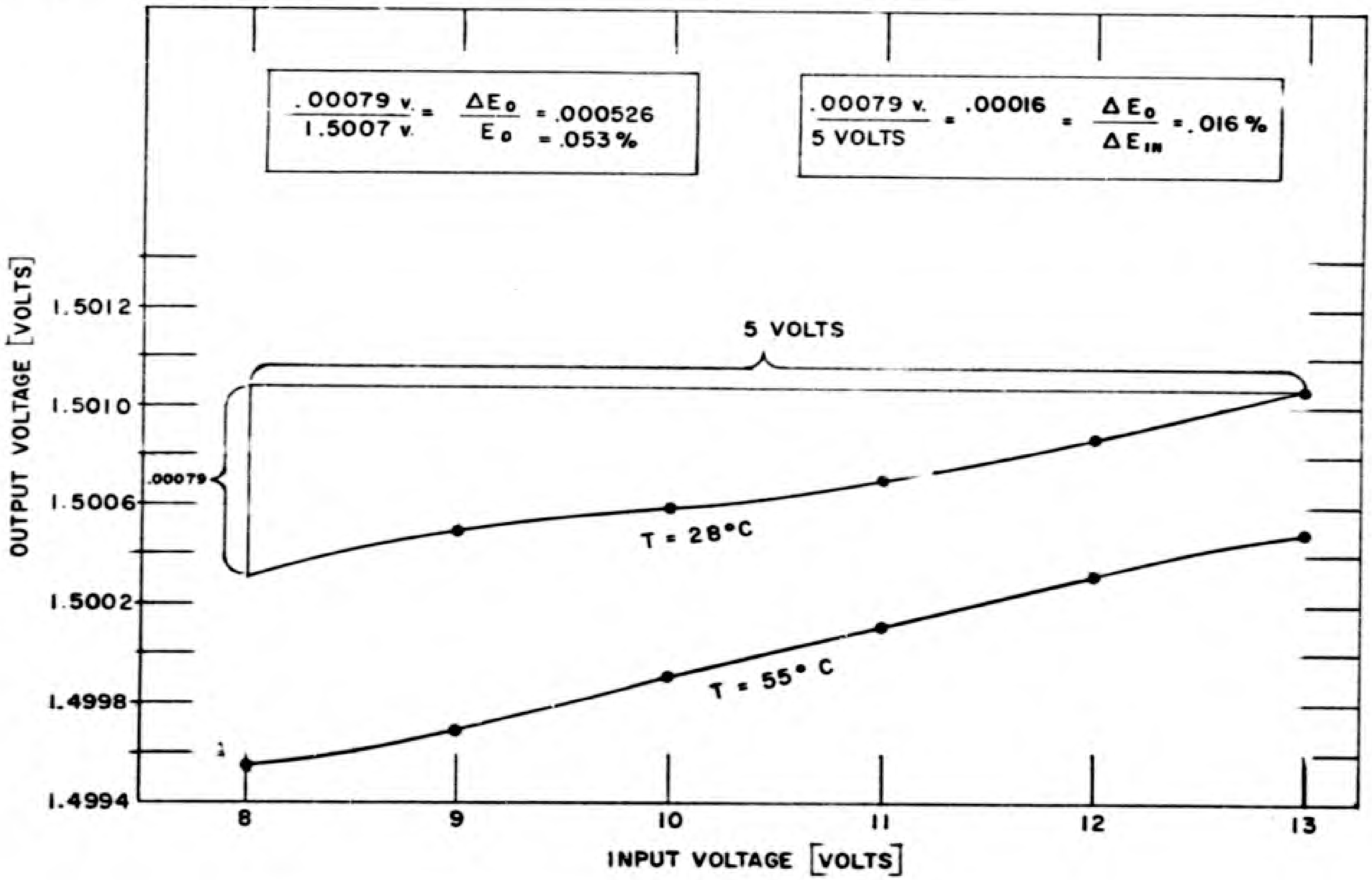

Figure 7. Operating characteristics of 1.5-v regulator.

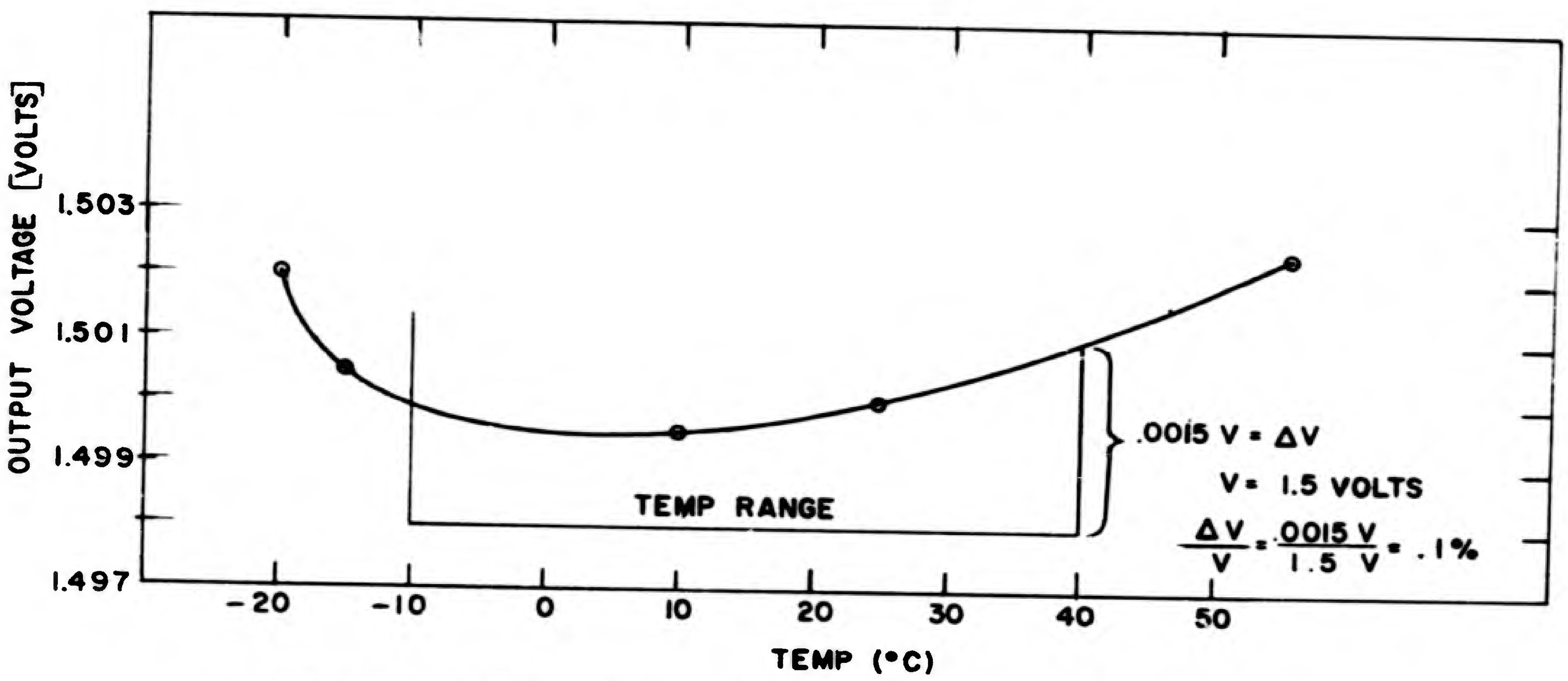

Figure 8. Output voltage versus ambient temperature for 1.5-v regulator.
[INPUT VOLTAGE = 10 v.]

The silicon resistor temperature coefficient may also be diluted by adding a series resistor similar to that used for thermistor dilution (fig. 5C). Analogous to equation 14, the diluting resistance R_D for silicon resistor compensation is given as

$$R_D = R_S \left(\frac{\eta}{\delta} - 1\right) \qquad (16)$$

where

$$R_9 = R_D + R_S \text{ at } 25^{\circ}C$$

η = silicon resistor temperature coefficient

R_S = silicon resistor resistance at $25^{\circ}C$

In using silicon resistor compensation, it is necessary to have a linearly varying reference diode (with temperature). Stabistors may be obtained that exhibit linear characteristics with temperature variation. By using linear devices in both the reference diode and the compensating portions of the circuit, it would be expected that even greater accuracy and stability than that illustrated in figures 7 and 8 would be achieved.

4.4 Further Considerations

In addition to extensive concern over the temperature stability in the sampling-comparator stage of the voltage regulator, some attention was necessarily directed to the remainder of the circuit. In general, small variations due to temperature changes could be ignored completely since the regulator would compensate fully, provided the comparator portion of the circuit was stable with temperature. However, when very gross changes occurred, the circuit was unable to compensate. e.g. in initial testing, it was found that the output voltage increased by 2.5 v (166 %) when the regulator was operated with a $60^{\circ}C$ ambient temperature. It was readily apparent that the series transistor (Q_1) was the source of the anomalous behavior, it being a germanium transistor with high leakage (I_{CBO}) at temperatures significantly higher than room temperature. Upon substituting a silicon transistor for Q_1 the output voltage returned to the expected 1.5-v level. Since the effects of Q_2 and Q_3 (see fig. 6 of sec 5) were less pronounced than that of Q_1, they were retained as germanium transistors with no deleterious effect; this is a noteworthy example of the circuit's ability to compensate for lesser fluctuations due to temperature change.

Since the series transistor dissipates some power, it is advisable that it be mounted on a sufficient heat sink. Also, the entire comparator-sampling portion of the circuit should be on a common heat sink, so that Q_4 and Q_5 track each other closely and R_9,

R_{10}, and E_{REF} perform at identical temperatures. If the circuit were to be potted, it would be desirable to maintain close proximity of sampling components and as much isolation of the series transistor as possible.

5. COMPLETE CIRCUIT

Figure 6 represents a combination of figures 2, 3, and 4, with some additions. Zener diode CR_3 was added to improve the overall regulation and bring the input voltage to a level ($\sim$ 6 v) desirable for optimum circuit performance. This addition was necessary because of the special requirement of an input voltage of 8 to 12 v. As can be seen in figure 7, the regulation $\frac{\Delta E_{IN}}{\Delta E_{OUT}}$ of the complete circuit is less than 200 parts per million. Diode (stabistor) CR_2 was added to raise the level of the emitter of Q_3, so that it would operate in a linear region. The double-diode configuration of CR_1 and E_{REF} was adopted to provide a voltage level more appropriate to the particular requirements. Addition of a diode doubles the voltage level of a single diode, but retains the other diode characteristics as if the combination were a single unit. The thermistors added to the comparator-sampling portion of the regulator provide the temperature compensation for the temperature-sensitive, reference diode. The thermistors are in a double configuration to obtain the proper level of resistance. Since the output voltage is nearly equal to the reference voltage, the addition of a series-diluting resistor was unnecessary to attain satisfactory results (see fig. 8). A small variable resistance was placed in the upper leg of the output voltage divider, which makes possible a final adjustment of the output voltage level which had very little dependence ($\sim$ 0.1mv) upon load currents of 1 to 15 ma.

6. CONCLUDING REMARKS

It has been demonstrated that highly accurate, low-voltage regulation is possible with a completely solid-state design. Temperature-compensating circuitry is necessarily involved--the price to be paid for highly specialized requirements. It should be noted that the temperature stability of the regulator described herein is of the same order of magnitude as that of a standard cell for the temperature range under consideration. It is recommended that further attempts at temperature-stable low-voltage regulators make use of silicon resistors and linear voltage references to eliminate the necessary coincidence of second-order effects. The design criteria given herein will serve to direct design of a regulator utilizing linear devices.

Efficiency is obvously low, but where a highly precise voltage is required as a reference level with power output (for example in telemetry systems) the inefficiency can easily be overlooked, since the load current may vary typically from 1 to 15 ma. Thus, the overall power requirements are less than 1/2 w.

The circuit techniques discussed in this report offer the designer several new additions to the state-of-the-art. Included in these are: thermistor and silicon resistor linearization and dilution; use of forward-biased reference diodes for highly accurate, low-voltage regulation; design equations and considerations for utilization of temperature-dependent resistors for device compensation.

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David C. Auth

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A solid-state, low-voltage regulator accurate to 0.1 percent with temperature stability has been developed. Employing thermistor compensation, forward-biased diodes, and matched silicon transistors, it represents a significant advance in the state-of-the-art. The circuit discussed herein yields 1.500 ± 0.001 v at 1 to 15 ma with input voltages of 8 to 12 v over the temperature range -10^{o} to $+40^{o}C$.

Circuit design and derivations for temperature compensation using thermistors and silicon resistors are covered in detail along with general design criteria for low-voltage regulators.

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TR-1222, 30 April 1964, 18 pp text, 8 illus, DA-1P523801A300 AMCMS Code 5523.11.62400, HDL Proj. No. 96300 UNCLASSIFIED Report

A solid-state, low-voltage regulator accurate to 0.1 percent with temperature stability has been developed. Employing thermistor compensation, forward-biased diodes, and matched silicon transistors, it represents a significant advance in the state-of-the-art. The circuit discussed herein yields 1.500 ± 0.001 v at 1 to 15 ma with input voltages of 8 to 12 v over the temperature range -10° to +40°C.

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