NASA TECHNICAL REPORT

NASA TR R-271

c. 1

NASA TR R-271

ABSOLUTE TEMPERATURE MEASUREMENT AT MICROWAVE FREQUENCIES

by George G. Haroules and Wilfred E. Brown III

Electronics Research Center
Cambridge, Mass.

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION • WASHINGTON, D. C. • SEPTEMBER 1967

ABSOLUTE TEMPERATURE MEASUREMENT

AT MICROWAVE FREQUENCIES

By George G. Haroules and Wilfred E. Brown III

Electronics Research Center
Cambridge, Mass.

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

For sale by the Clearinghouse for Federal Scientific and Technical Information
Springfield, Virginia 22151 – CFSTI price $3.00

ABSOLUTE TEMPERATURE MEASUREMENT* AT MICROWAVE FREQUENCIES

By George G. Haroules and Wilfred E. Brown III

ABSTRACT

The measurement of absolute temperature under field operating conditions is usually achieved at the expense of tedious calibration using carefully controlled cryogenic devices. In this report, a technique is described by means of which such measurements can be made using only conventional laboratory hardware. Two Dicke-type radiometers have been developed and constructed which embody this technique.

The measurement system that has been developed establishes the rf zero of the radiometer and determines the absolute temperature while operating in a rf-balanced condition. A simplified laboratory calibration of the device and the effect of gain stability on measurement accuracy, while observing temperatures near absolute zero, are discussed.

Field measurements of sky temperature using standard gain horns and a 28-foot parabolic antenna were conducted at frequencies of 8 and 15 gHz. The experimental results of the program are compared with theoretical predictions of sky temperature for various atmospheric opacities.

I. INTRODUCTION

The ability to obtain a precise measure of absolute flux density at microwave frequencies requires an antenna, the gain of which is accurately known, and a radiometer capable of measuring signal power at the antenna output in units of equivalent absolute temperature. Most radiometric methods for absolute temperature measurement use a calibrated source of noise power and require a series of precise rf attenuation measurements which represent the major source of instrument calibration error. In addition, for the measurement of small temperatures near zero, cryogenic equipment is frequently incorporated in the measurement system, introducing the added complexity of interrelated circuit components that operate at markedly different thermometric temperatures.

The purpose of this report is to describe a new radiometric technique for the absolute measurement of temperatures at microwave frequencies. This technique eliminates radiometer calibration errors associated with rf attenuation measurements. In addition all components of the instrument operate near the ambient temperature. The technique takes advantage of the inherent ability of a radiometer to measure noise temperature differences precisely. A known source of noise power is required only for laboratory adjustment and calibration of the instrument.

---

* Presented at URSI (International Scientific Radii Union), Commission I. Stanford, University, Stanford, California, December 8, 1966.

II. REVIEW OF CONVENTIONAL METHODS AND TECHNIQUES

The performance requirements on a radiometer may be described in terms of its output indicator. There are two:

1. The output indicator scale must be calibrated in degrees Kelvin.

2. The output indicator values must be referred to absolute zero.

The more common techniques for absolute microwave temperature measurement will be reviewed in terms of the degree to which they meet these two requirements.

First consider the Dicke radiometer in its most conventional form. A simplified functional block diagram of the unit is shown in Figure 1. Its most significant feature is the rf modulator at the receiver input which acts as a single-pole, double-throw switch. The modulator provides an amplitude-modulated noise signal input to the radiometer which is proportional to the temperature difference between the noise powers presented to the input ports of the modulator. The modulation component is amplified at the rf frequency, detected, and then further amplified by a narrow band amplifier at the modulation frequency. The signal is then synchronously detected to provide an output dc voltage to the indicator system, proportional to the input temperature difference. Post-detection integration and dc amplification of the output voltage is frequently introduced between the detector output and the indicator system input. The ability to obtain an absolute measure of an input signal in terms of its noise temperature is determined by the circuitry preceding the modulator.

That portion of the radiometer following the modulator output is common to several modes of radiometer operation. In the discussion which follows we will make the reasonable assumption that the noise power signal level is sufficiently low relative to the receiver noise power level to assure a square law detector response.

The sensitivity of a Dicke radiometer is conventionally described by a mathematical statement of the criterion of detectability. The minimum detectable signal is defined as one in which the mean deflection at the output indicator is equal to the standard deviation of the fluctuations about the mean deflection of the indicator. As shown by Dicke (ref. 1), the minimum detectable signal is related to receiving system parameters in the form:

$$\overline{\Delta T}_{\eta} = \frac{\beta T_{\eta}}{\sqrt{(\Delta\nu)\,(t)}} \tag{1}$$

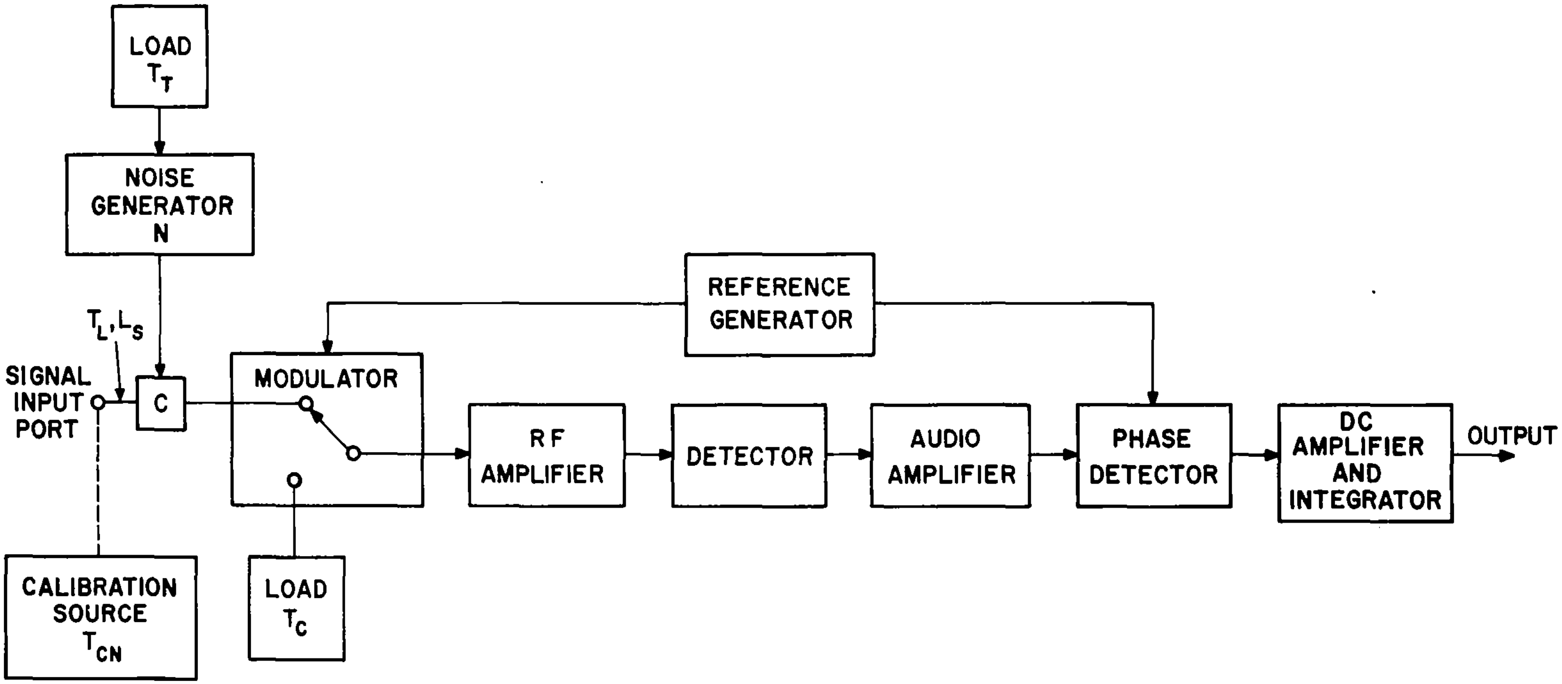

FIGURE 1
BASIC DICKE RADIOMETER

where

T_η = the receiving system noise temperature

$\Delta\nu$ = the predetection receiver bandwidth

t = the post-detection integration time constant

β = a constant of proportionality.

The value of β is determined by other receiver system parameters. For most systems the value of β falls between 1 and 2. The receiver noise temperature may be derived from a measure of system noise figure F by the relation:

$$T_\eta = (F - 1)\, T_o \qquad (2)$$

where T_o = 290 °K.

The ability of the Dicke radiometer circuit (shown in Figure 1) to provide an accurate measure of temperatures near absolute zero in terms of previously stated criteria is limited by

1. RF attenuation of the signal by passive circuitry between the input signal port of the radiometer and the input signal port of the modulator

2. Receiver gain variations

3. Noise radiated into the signal transmission line by the passive components which constitute the transmission path.

Attenuation of the signal is associated with absorption in the resistive losses of the walls of the transmission line and direct loss of that portion of the signal coupled to the internal load of the directional coupler C. Noise radiated into the transmission line consists of two components:

1. Re-radiation from the resistive losses in the walls of the transmission line

2. A noise component originating in the thermal resistive load T_T and injected into the transmission line via the side arm of the directional coupler (noise generator off).

Receiver gain variations introduce a fluctuating component at the output indicator. These fluctuations are superimposed on the statistical fluctuations associated with the inherent receiver system noise. The amplitude of this component is the product of fractional gain variations and the amplitude of the noise power difference at the input ports to the modulator and may be expressed as (ref. 2):

$$\left|\Delta T_g\right| = \frac{\Delta G}{G_o}\left|T'_a - T'_c\right| \tag{3}$$

where

$\frac{\Delta G}{G_o}$ = fractional gain variation

T'_a = effective temperature of the "signal noise power" at the signal port of the modulator

T'_c = effective temperature of the comparison noise power at the comparison port of the modulator.

Signal attenuation by the transmission line and directional coupler provides a signal level referenced to the modulator input of

$$\text{attenuated signal} = \frac{T_a}{L_s}\left(1 - \frac{1}{C}\right) \tag{4}$$

where

C = coupling value

L_s = transmission line loss in db

T_a = absolute temperature of the input signal.

The composite noise component radiated into the transmission line by resistive losses in the walls of the line and by the load at a temperature T_t is given by

$$\left(1 - \frac{1}{L_s}\right) T_L\left(1 - \frac{1}{C}\right) + \frac{T_t}{C} = \text{noise added} \tag{5}$$

where T_L is the thermometric temperature of the transmission line. The effective temperature of the comparison noise source at the comparison port of the modulator is determined in a similar manner by the attenuation and noise radiation of the transmission line connecting the comparison source output to the modulator input.

From the foregoing it is apparent that the effective "signal noise power" temperature T_a referenced to the input port of the modulator will be:*

$$T'_a = \frac{T_a}{L_s}\left(1 - \frac{1}{C}\right) + \left(1 - \frac{1}{L_s}\right) T_L\left(1 - \frac{1}{C}\right) + \frac{T_t}{C} \tag{6}$$

and the effective temperature of the comparison noise source referenced to the input comparison port of the modulator will be:

$$T'_c = \frac{T_c}{L_c} + \left(1 - \frac{1}{L_c}\right) T_L. \tag{7}$$

The accuracy obtained in the measurement of a noise signal at an equivalent absolute temperature T_a is determined by the accuracy to which the transmission line attenuation, directional coupler value, and the thermometric temperature of the individual passive components can be measured. The significance of these factors in determining the accuracy of an absolute temperature measurement is best illustrated by a numerical example. Typical system parameters are presented in Table I.

The performance characteristics of the system described by the parameter values in Table I are summarized in Table II.

An inspection of Table II shows the following:

1. The sensitivity of the radiometer is limited by receiver gain variations, not by receiver noise temperature.

2. The temperature presented at the input comparison port of the modulator is equivalent to the temperature of the comparison noise source, since the interconnecting transmission line is at the same temperature as the source.

3. The attenuating effect of the transmission line resistive losses and coupler on the input signal is small in comparison with the noise added by these losses and by the load T_t.

---

*Terms such as signal transmission line loss following the coupler are not included for the purpose of brevity. Their introduction would further complicate the expression but not affect the result.

TABLE I

TYPICAL SYSTEM PARAMETERS

Parameter	Value
Signal temperature, T_a	1 ^oK absolute
System noise figure, F	7 db
Receiver gain stability, $\frac{\Delta G}{G_o}$	1/100
Predetection bandwidth, $\Delta \upsilon$	1 gHz
Post-detection, time constant, t	10 seconds
Comparison load temperature, T_c	290 ^oK
Thermometric temperature of passive components, T_L	290 ^oK
Directional coupler value, C	10 db
Load termination temperature, T_t	290 ^oK
Signal path transmission line loss, L_s	0.5 db

TABLE II

SUMMARY OF PERFORMANCE CHARACTERISTICS

Parameter	Eq. No.	Value (^oK)
System sensitivity, ^oK rms (β = 2)	1, 2	0.02
Signal level at modulator input, ^oK	4	0.81
Noise added to signal, ^oK	5	52.20
Effective signal temperature at modulator, ^oK	6	53.01
Effective comparison temperature at modulator, ^oK	7	290.00
Input temperature difference at modulator, ^oK	6, 7	236.99
Output indicator fluctuation level due to receiver gain variations, ^oK	3	2.370

4. The major source of calibration error is associated with the added noise introduced by the passive circuit components in the signal transmission line and by receiver gain variations.

Calibration of an instrument of this type requires extremely careful measurement of transmission line losses and the directional coupler value (refs. 3, 4). An error of 0.1 db in the measurement of the coupling value of the directional coupler would introduce an error of 7.5 °K in the calculated value of the "added noise component". Even if attenuation and coupling measurements can be made to an accuracy of 0.1 db, however, the noise measurement is complicated by the requirement that this be done over the full predetection bandwidth of the receiver and that their integrated effect on a broadband noise signal be introduced in the calibration.

The effect of receiver gain variations can be made negligible by direct reduction of the rf unbalance at the modulator input. Appendix B contains a more complete discussion of the effect of receiver gain variations. Historically, this has been achieved by use of a comparison noise power source, the temperature of which is nearly the same as the signal temperature. A resistive load immersed in a liquid helium bath is frequently used as the comparison noise source for low-level signal temperature measurements. If the modulator is maintained at ambient temperature, the noise contribution of the inter-connecting transmission line must be included in the calibration procedure. Composite resistive losses are carefully measured and the thermometric temperature along the transmission line simultaneously monitored at several points to establish the integrated noise contribution of the several segments of the line which are at different temperatures. This transmission line is frequently insulated and evacuated to prevent internal condensation. Although the introduction of a liquid helium load as the comparison noise source in a Dicke radiometer reduces the effect of receiver gain variations when measuring low level noise signals, its use does not affect or reduce the previously discussed instrument calibration errors associated with signal transmission line components. The addition of an appropriate amount of noise in the signal transmission line provided by a gas discharge noise achieves the same result insofar as reducing the effect of receiver gain variations (ref. 2). Knowledge of the noise power temperature added by the gas discharge depends on the accuracy to which the directional coupler value is known and on the effective temperature of the noise power output of the gas discharge (refs. 5, 6). Both are determined by a series of rf-attenuation measurements.

A minor modification of the basic Dicke circuit is shown in Figure 2. This circuit is identical to that of Figure 1 with the addition of an input switch S and a calibrated noise source T_{c2} at liquid helium temperature. Here again the instrument calibration error is determined by the accuracy of attenuation and coupler value measurements. The differential deflection at the output indicator is recorded when switch S is connected to the signal input port. Since the noise added by the transmission line and the load T_t via the side arm coupler is constant for either position

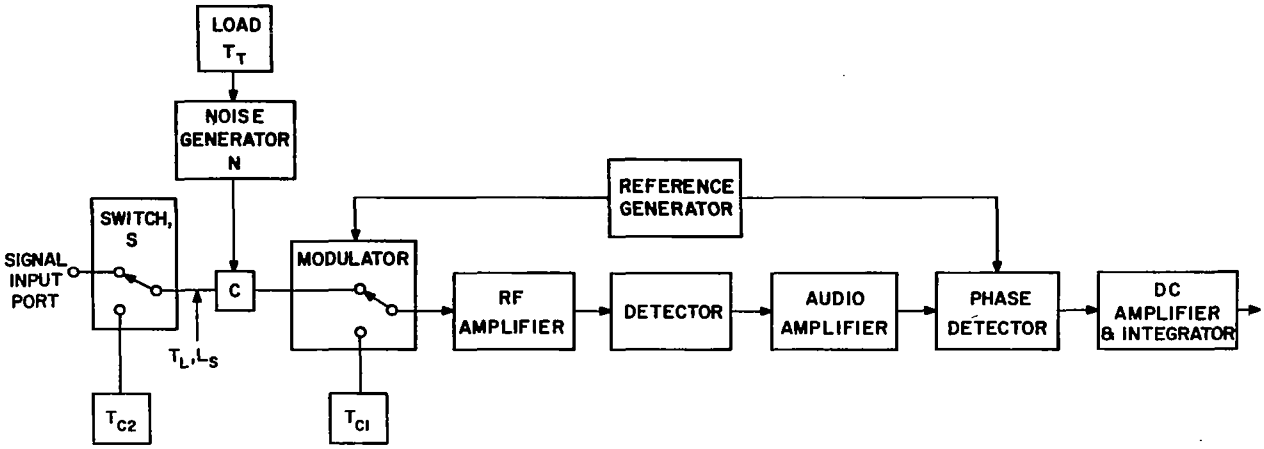

FIGURE 2

DICKE RADIOMETER WITH SINGLE LOAD INPUT CALIBRATION

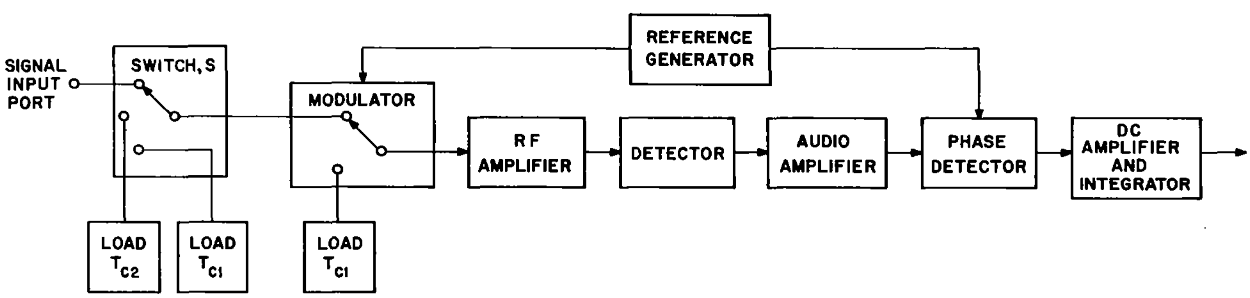

FIGURE 3

DICKE RADIOMETER WITH DUAL LOAD INPUT CALIBRATION of switch S, it acts as a fixed noise bias. Hence, the recorded differential deflection at the output indicator only needs to be corrected for the measured attenuation of the line and the signal loss to the internal coupler load.

A further modification of the Dicke circuit is shown in Figure 3. A three-port, single-pole switch S and two calibrated noise sources are introduced at the radiometer input port. One noise source is at the same temperature as the comparison source T_{c1}. The coupler and associated gas discharge noise circuitry are not used in this circuit. The gain calibration of the instrument is provided by the temperature difference between the sources T_{c1} and T_{c2}. Computation of the noise bias component introduced by the transmission line loss determines the error associated with establishing the output indicator value relative to absolute zero. The accuracy is determined by measurement of the associated attenuation value. Correction for signal attenuation is not required since the percent attenuation of the transmission line is common to the signal source as well as the sources T_{c2} and T_{c1}. From an instrumentation standpoint the system is quite complex; it requires a minimum of two cryogenic baths, i.e., the two resistive loads at a common temperature can be placed in one bath. An added complexity is the need for thermometric temperature monitoring of the various parts of the rf circuitry contained outside of either bath.

In summary, conventional techniques for absolute temperature measurement at microwave and millimeter frequencies involve complex instrumentation and require precise knowledge of passive rf component characteristics. The magnitude of calibration errors is determined by the accuracy to which resistive and coupler losses in the signal path can be measured. The magnitude of these errors establishes the ability to refer the output indicator reading to absolute zero. While the calibration of the output indicator scale in degrees Kelvin is usually obtained with a high degree of accuracy, the absolute zero reference point is not. Common techniques other than those described represent combinations or minor variations of the circuits shown in Figures 1 through 3 or some step in the calibration procedure for these circuits.

The techniques discussed previously are generally applicable to radiometers with receiver noise temperatures of the order of 1000 $^{\circ}$K or greater. For low noise receiving systems with cooled parametric amplifiers or masers, the "Y factor method" of receiver noise temperature measurement is commonly employed to obtain an absolute measure of signal temperature since, in this case, the signal temperature may represent a large fraction of the receiver noise temperature. The accuracy of the Y factor method of measurement is directly determined by the ability to calibrate accurately the attenuators used in the measurement.

Radiometers which employ masers are subject to two additional sources of error. There are the effect of maser gain variations on the noise temperature of the following receiver stages and the possible deviation of the detector from a square law characteristic. The latter is a potential source of error when the signal temperature is comparable to the maser noise temperature. As noted previously, this is of little concern in systems with noise temperatures of the order of 1000 $^{\circ}$K or more.

III. THE METHOD OF CALIBRATED RF INPUT ZERO TRANSLATION

A radiometer input circuit which provides a relative zero and operates in a near balanced condition was suggested in 1965 (ref. 7). The novel technique to be described here extends these features to absolute temperature measurements. A simplified block diagram of the circuit is shown in Figure 4. The advantages of this circuit are that (1) it eliminates the need for cryogenic equipment after calibration has been completed, (2) it does not require rf attenuation measurements for absolute temperature calibration, and (3) it places all rf components within an environmental enclosure at a temperature slightly higher than the highest anticipated ambient operating condition. A small section of the input signal transmission line passing through the wall of the environmental enclosure is made of thermal insulating material, such as gold-plated stainless steel, in order to minimize temperature variations along the line between outside ambient and the internal environmental temperature.

Calibration of the instrument proceeds as follows:

1. The scale value of the output indicator is calculated by relating the measured negative deflection from the indicator zero reference to the absolute temperature difference between the thermometric temperature of the environmental enclosure T_e and the known temperature of the noise calibration source T_{cn}. The gain of the receiver is carefully measured at this point to assure that all further adjustments are made at the same gain.

2. The gas discharge noise source N_1 is ignited and attenuator A is adjusted to provide a positive output indicator deflection equivalent to the known absolute temperature of the noise calibration source. The indicator scale units are those provided by the prior calibration of the "negative deflection". It is important to assure that the receiver gain remains the same during this step in the procedure as when the "negative deflection" was measured. Gain stability may be checked by extinguishing the gas discharge and noting that the amount of negative deflection is identical to that previously measured. If not, the re-receiver gain is appropriately adjusted at some point in the circuit following the output of the rf modulator. If the coupler is located forward of the switch, a separate calibration signal is required which may be derived from the same noise source. The level of this signal should be sufficient to measure receiver gain variations.

3. The value of attenuator A is set permanently and the radiometer calibration is complete.

A detailed analysis of the RF input zero translation is contained in Appendix A.

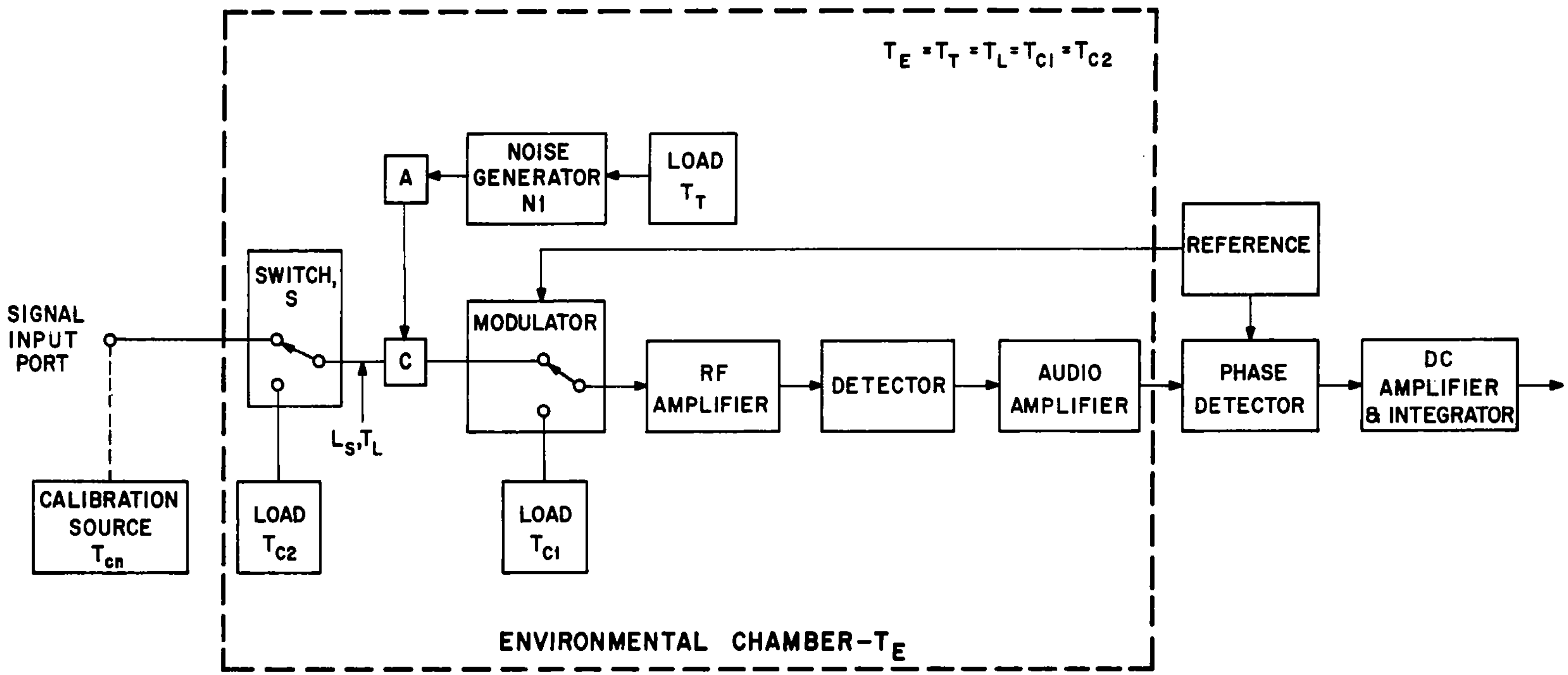

FIGURE 4

METHOD OF CALIBRATED RF INPUT ZERO TRANSLATION FOR ABSOLUTE CALIBRATION OF A RADIOMETER

IV. MEASUREMENT OF A TEMPERATURE NEAR ZERO

The operating procedure for measurement of a signal input temperature near absolute zero is as follows:

1. The indicator zero reference point is established with switch S connected to load T_{c2}.
2. The gas discharge noise source is ignited and the receiver gain adjusted to obtain a positive deflection equivalent to the thermometric temperature of the environmental enclosure in equivalent scale units as determined by laboratory calibration.
3. Switch S is connected to the signal input port of the radiometer.
4. The absolute temperature of the signal is indicated directly as a positive deflection relative to the indicator zero reference point which is absolute zero.

The rf input components of a radiometer may be arranged in various circuit configurations and obtain the same result as long as the fundamental steps in the calibration procedure can be performed. The significant features of the technique include:

1. Containment of all critical rf input circuit components in a common environmental enclosure at a near ambient temperature.
2. Introduction of a switch at the signal input port of the radiometer to provide a measure of the indicator zero reference when the switch is connected to an internal source of noise at the temperature of the environmental enclosure.
3. Provision for injection of a noise signal in the signal transmission line which translates the indicator zero reference to absolute zero.

V. LIMITATIONS AND APPLICATIONS

Applicability of the technique to specific system configuration may be limited by physical considerations, i.e., the size and location of other components in the vicinity of that point in the signal transmission line at which an absolute measure of signal temperature is desired. The most frequent compromise will be addition of an rf transmission line between the desired point of measurement and the input signal port of the radiometer. In this case care should be taken to maintain the added section of transmission line at a near constant temperature by use of appropriate heating and insulating techniques. During laboratory calibration, the negative deflection at the output indicator corresponding to the temperature difference between the calibrated noise source T_{cn} and the environmental enclosure T_e is obtained by interconnection of a thermal load at the temperature T_e. This is followed by the calibrated noise source at the temperature T_{cn} to the signal input port of the added section of transmission line.

The absolute measurement of noise signals at a temperature sufficiently above absolute zero to make the effect of receiver gain variation noticeable requires adjustment of the indicator zero reference level to a temperature near the anticipated signal temperature. This can be accomplished in several ways. A relatively simple circuit is shown in Figure 5. The circuit is calibrated in accordance with the procedure previously described during an actual operational measurement. The positive indicator deflection corresponding to the absolute temperature of the input signal is nulled at the output indicator by ignition of gas discharge noise source N_2 and adjustment of attenuator A_2. The corresponding absolute temperature displacement of the output indicator zero reference is then obtained by a series of comparative measurements of the output indicator deflection with noise source N_1 "on" and N_2 "off" and with N_1 and N_2 both "on". The final measurement of signal temperature is obtained from the small residual deflection of the output indicator with both noise sources ignited, combined with knowledge of the negative translation of the absolute zero indicator reading.

Care should be taken to fix firmly the position of attenuator A_2 during the field measurement to provide an opportunity to repeat the measurement of the absolute zero translation value.

For a signal temperature near the ambient temperature of the environmental enclosure, both gas discharges N_1 and N_2 are extinguished during the period of measurement. Gas discharge N_1, in accord with the basic procedure, is used for calibration of receiver gain. For signals well above the ambient temperature of the enclosure, operational measurements are made with gas discharge noise source N_1 "off" and source N_2 "on". Attenuator A_2 is adjusted to provide an approximate null output. The corresponding value of the indicator zero reference point is obtained in a manner similar to that previously described for signal temperatures for absolute zero and the temperature of the environmental enclosure. In this case, however, the negative suppression of the indicator reading for absolute zero is the sum of the noise temperature injected by N_2, referenced to the comparison port of the modulator, plus the temperature of the environmental enclosure.

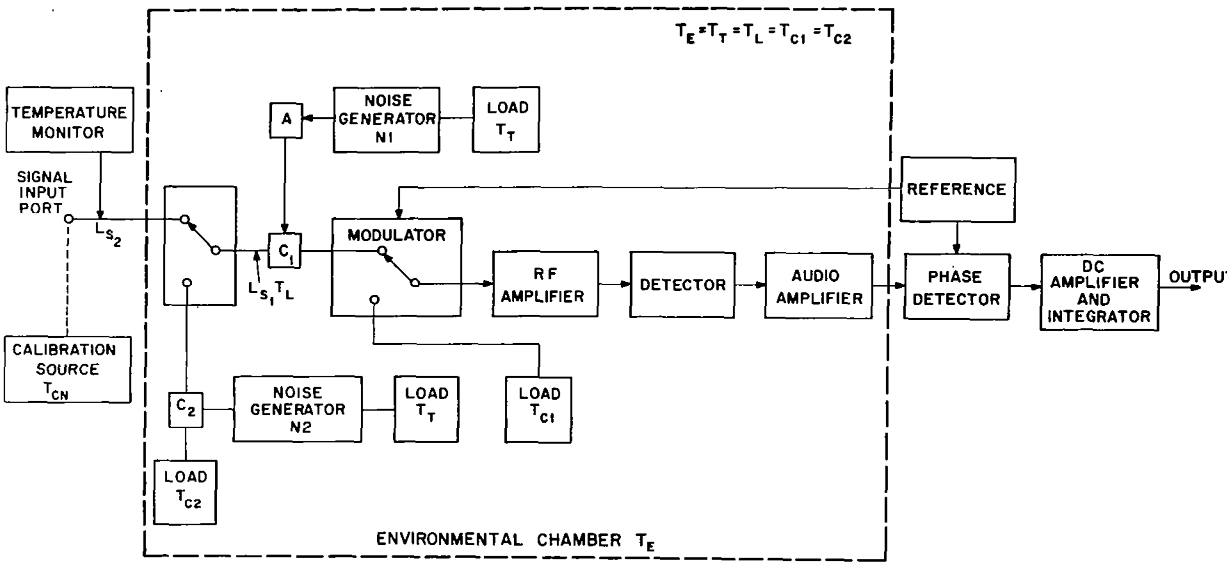

FIGURE 5
MODIFIED APPROACH TO ABSOLUTE CALIBRATION OF MICROWAVE RADIOMETER

VI. EXPERIMENTAL VERIFICATION

The radiometer technique that has been described was tested at 8 and 15 gHz. The functional components were connected as shown in Figures 6 and 7. Photographs of the laboratory radiometers are shown as Figures 8 and 9. Gas discharge noise sources were used throughout the calibration procedure to monitor receiver gain. The noise injection level of these sources relative to the comparison port of the modulator at frequencies of 8 gHz at 15 gHz were 306 and 302 $^{\circ}$K, respectively. The measured sensitivity of the 8-gHz radiometer was 0.04 $^{\circ}$K rms and the 15-gHz radiometer was 0.06 $^{\circ}$K rms for a post-detection integration time constant of 1 second. The corresponding peak-to-peak fluctuation level at the output indicator was of the order of 0.25 and 0.37 $^{\circ}$K, respectively, for the 8-gHz and 15-gHz radiometers. Temperature changes introduced in the waveguide by the heat from the switch coil in its energized position were easily detected.

The reference noise source used for calibrating the radiometer was a resistive waveguide load immersed in a bath of acetone and dry ice. The thermometric temperature of the load was monitored by three thermocouples separately located at the bottom, midway, and at the top of the waveguide section.

The output indicator used in the calibration was a Sanborn strip chart (type 7716A). A preset and fixed value of zero suppression was introduced in one recorder channel to provide a full-scale expansion of the upper 20 percent of the total output voltage obtained when the gas discharge noise tubes were ignited. Under this condition, a 2.5-mm deflection was equivalent to a 1 percent change in receiver gain. The peak-to-peak statistical fluctuation level was slightly less than 0.5 mm on this scale. By using three channels of the strip chart recorder, each preset with a fixed-zero suppression, both positive and negative deflections were independently measured without recourse to recorder adjustment and possible sources of error otherwise with attenuator settings of the recorder.

An environmental temperature enclosure was not used in the laboratory evaluation program. The value of such an enclosure, however, was quite evident. Minute temperature differences between the loads T_{c1} and T_{c2} were easily detected during that step in the procedure in which the indicator zero reference point was established. The value of a self-latching feature for switch S_1 was also apparent.

Once calibrated, the radiometer provided consistent results over a long time period. Excluding problems associated with the lack of a thermal environmental enclosure and heating of the guide by switch S, the major source of calibration error was traceable directly to the accuracy with which the calibrated noise source temperature was known. This error is common to any method of absolute calibration. Considerable technique development effort has been reported on this subject (refs. 6, 7). Appropriate design considerations are well documented.

Field tests of the calibrated radiometers were made using a small horn pointing toward zenith. The average sky temperature obtained during 3 days of

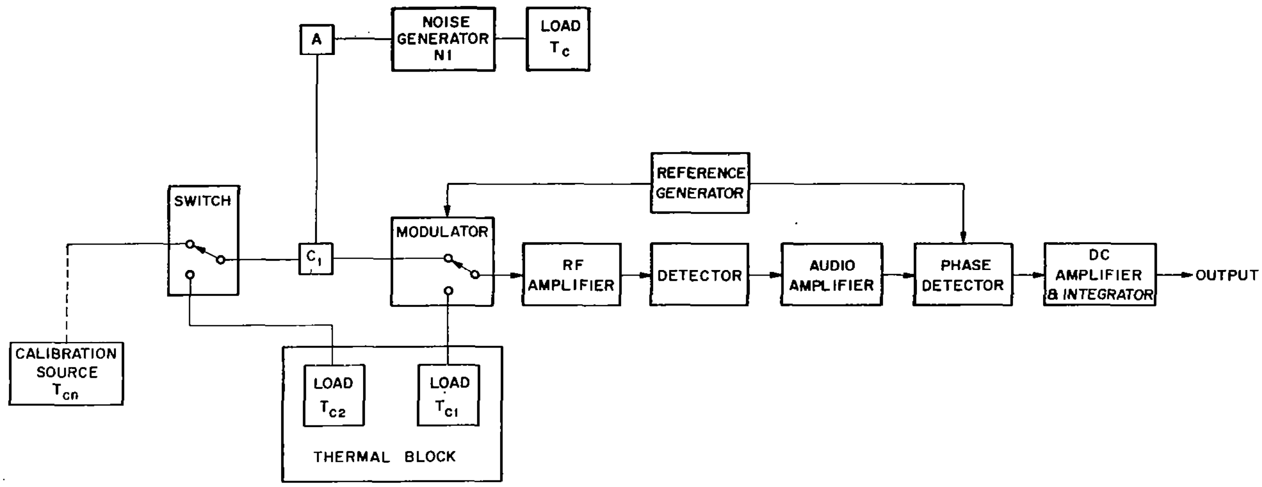

FIGURE 6
8 GHz SIMPLIFIED FUNCTIONAL BLOCK DIAGRAM
OF LABORATORY BREADBOARD TEST SYSTEM

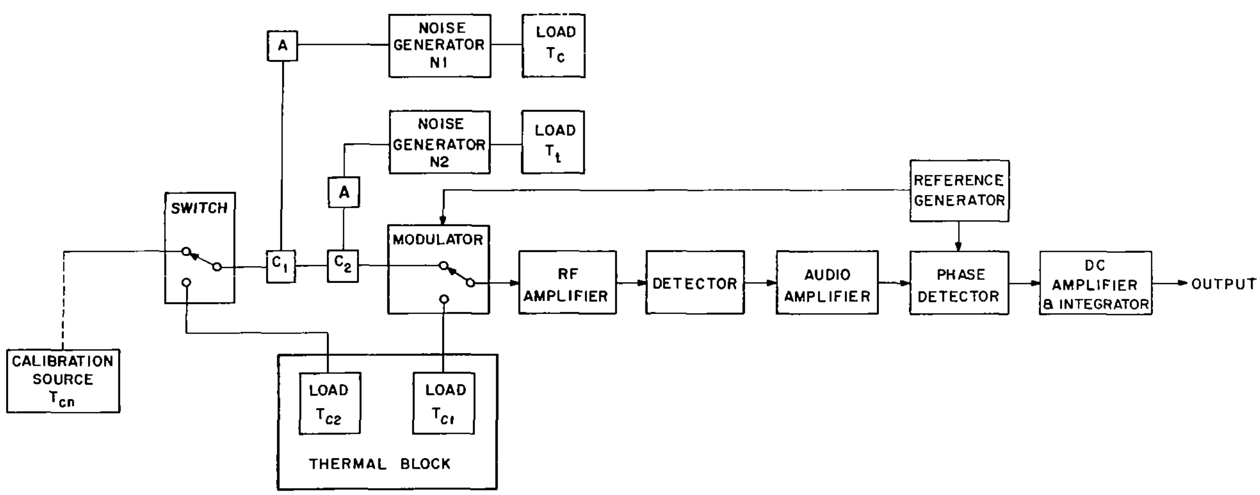

FIGURE 7
15GHz SIMPLIFIED FUNCTIONAL BLOCK DIAGRAM
OF LABORATORY BREADBOARD TEST SYSTEM

observation under "clear" weather conditions was 11 $^{\circ}$K at a frequency at 15 gHz and 7 $^{\circ}$K at a frequency of 8 gHz, a reasonable value for days of moderate humidity during the month of August in Boston. The instruments were also used to obtain a measure of the noise temperature of a 28-foot parabolic antenna with the feed at focus. Appendix C derives the theoretical method of calculating sky temperature from the radiative transfer equations. The measured temperature with the antenna pointing toward the zenith averaged 75 $^{\circ}$K at a frequency of 15 gHz and 68 $^{\circ}$K at a frequency of 8 gHz during a 4-day period of night time observations at each frequency. This value was considered in reasonable agreement with the anticipated performance of the antenna.

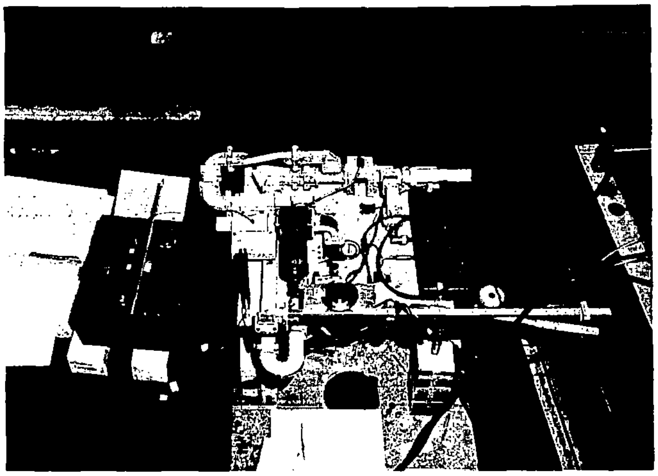

FIGURE 8
LABORATORY CALIBRATION OF 8-GHz RADIOMETER

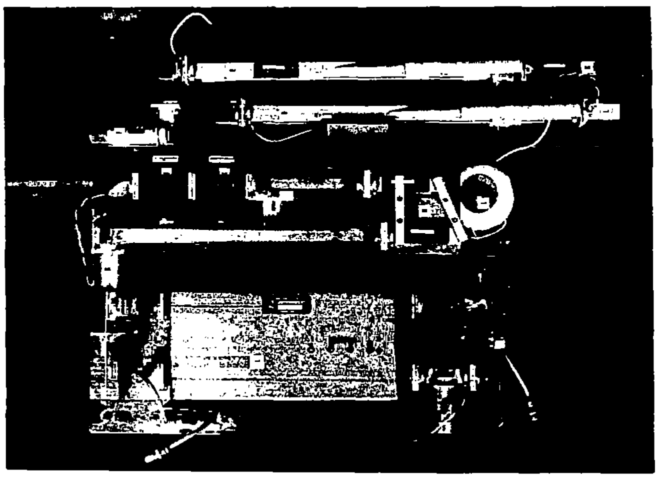

FIGURE 9
15-GHz RADIOMETER

VII. CONCLUSIONS

The radiometric technique of calibrated input rf zero translation for the absolute measurement of temperatures at microwave frequencies is based on the following fundamental considerations.

1. The measurement of small temperatures near zero is dependent on system sensitivity, not a low noise receiving system.

2. Operational use of a resistive load immersed in liquid helium as a comparison noise source is not required for the precise measurement of signal temperatures near absolute zero.

3. The linearity of the output temperature indicator of a radiometer to an internal calibration signal is determined by the detector law as well as the receiver gain at the time the calibration signal is injected. Knowledge of the detector law is a critical prerequisite to any calibration procedure. Both linear and square law detectors provide a square law response to input noise signals, the effective temperature of which is small compared with the system noise temperature. The higher the receiver noise temperature, the more certain is the square law response of the detector to low level signals.

4. The output noise power level of a well-designed gas discharge noise source is remarkably stable when power supply current is well regulated.

5. The ability of a sensitive radiometer to measure differential performance of rf components to a higher degree of accuracy than their absolute characteristics.

6. A calibration procedure which establishes the radiometer zero and permits absolute temperature measurements while the radiometer is in an rf-balanced condition.

REFERENCES

1. Dicke, R. H.: The Measurement of Thermal Radiation at Microwave Frequencies. Rev. Sci. Instr., vol. 17, 1946.

2. Drake, F. D.; and Ewen, H. I.: A Broad-Band Microwave Source Comparison Radiometer for Advanced Research in Radio Astronomy. Proceedings of the IRE, vol. 46, No. 1, January 1958, and Strum, P. D.: Considerations in High-Sensitivity Microwave Radiometry. Proceedings of the IRE, vol. 46, No. 1, January 1958.

3. Ohm, E. A.; and Snell, W. W.: A Radiometer for a Space Communication Receiver. B. S. T. J., September 1963.

4. Penzias, A.; and Wilson, R.: Measurement of the Flux Density of Cas A at 4080 Mc/s, Ap. J., vol. 142, No. 3, 1965.

5. White, W. D.; and Greene, J. G.: On the Effective Noise Temperature of Gas Discharge Noise Generators. Proceedings of the IRE, vol. 44, 1956.

6. Mumford, W. W.: A Broad-Band Microwave Noise Source. B. S. T. J., vol. 28, 1949.

7. Ewen, H. I.: State of the Art of Microwave and Millimeter Wave Radiometer Sensors. International Symposium on Electromagnetic Sensing of the Earth from Satellites, November 1965.

8. Wulfsberg, K.: Apparent Sky Temperature at Millimeter Wave Frequencies. Phys. Sci. Res., Paper No. 38, Air Force Cambridge Research Laboratories, July 1964.

APPENDIX A

ANALYSIS OF RF ZERO TRANSLATION

Since all rf components are at the same thermometric temperature T_e, including the resistive loads T_{c1} and T_{c2}, the effective temperature at either input port of the modulator is equal to the environmental temperature of the enclosure T_e when switch S is connected to T_{c2}. With the switch in this position, the output indicator reading corresponds to a balanced rf input condition, hereinafter referred to as "the indicator zero reference". Under this condition, the noise temperature at the signal input port to the modulator is expressed as:

$$\frac{T_e}{L_s}\left(1 - \frac{1}{C}\right) + \left(1 - \frac{1}{L_s}\right) T_e \left(1 - \frac{1}{C}\right) + \frac{T_e}{C} = T_e \qquad \text{(A-1)}$$

or

$$\frac{T_e}{L_s}\left(1 - \frac{1}{C}\right) + T_r = T_e \qquad \text{(A-2)}$$

where T_k is a constant noise temperature bias component and T_{c1}, T_{c2}, T_{t2}, and T_1 are all equal to T_e, the temperature of the environmental enclosure. It is important that we know the relative receiver gain at the time of laboratory calibration with a known source of noise power.

When the calibrated source of noise power at an equivalent temperature T_{cn} is connected to the input signal port of the radiometer via switch S, the effective temperature of noise referred to the signal input port of the modulator is:

$$\frac{T_{cn}}{L_s}\left(1 - \frac{1}{C}\right) + T_k \qquad \text{(A-3)}$$

From equations (A-2) and (A-3), the negative deflection of the output indicator obtained when switch S is connected to the radiometer input signal port may be expressed in the form:

$$\text{Deflection} = \alpha\,(T_e - T_{cn}) \qquad \text{(A-4)}$$

where α is a constant given by

$$\alpha = \frac{C - 1}{CL_s} . \tag{A-5}$$

There are two effects on any signal input temperature when referred to the signal input port of the modulator; a fractional change α in amplitude and the addition of a constant noise bias term equal to T_k*. Since the noise bias term is constant the deflection at the output indicator system (equation A-4) is always directly proportional to the absolute temperature difference between the thermometric temperature of the environmental enclosure and the absolute temperature of the input signal. It is not important that we directly measure the constant of proportionality.

Verification of the calibration is obtained by connecting the input switch S to the internal resistive load T_{c2}. The output indicator deflection will be positive and equal to the absolute temperature of the environmental enclosure T_e. If the gas discharge noise source is extinguished, the output indicator reading will return to the indicator zero reference.

The amount of noise injected by the gas discharge, referenced to the signal input port of the modulator is obtained from the expression:

$$\alpha T_{cn} + T_k + T_{ni} = T_e + \alpha T_{cn} \tag{A-6}$$

or the calibrated rf input zero translation noise T_{ni} is given by:

$$T_{ni} = T_e - T_k \tag{A-7}$$

*The value of T_k in the signal path may not be exactly equal to the analogous T_k in the comparison path. This does not affect the accuracy of the calibration since it is the value of T_k in the signal path which is accommodated by the calibration procedures.

The following should be emphasized in regard to the calibration procedure:

1. The measurement of signal transmission line resistive losses L_S is not required for absolute calibration.

2. Precise knowledge of the directional coupler value C and attenuator A are not required for calibration.

3. A cryogenic load is required, but only for laboratory calibration of the instrument. This load is not required or made an integral part of the operational system.

4. Calibration accuracy is determined by the ability to measure differential deflections at the output indicator to a high percentage accuracy.

5. Since the calibration error is directly dependent on the ability to measure the "negative deflection" of the output indicator to a fraction of a percent during the first step in the procedure, it is desirable that the calibrated source of noise power be at the temperature difference, as large as practical, relative to the environmental temperature of the enclosure. For this reason, a resistive load immersed in a liquid helium bath is recommended for the calibration.

APPENDIX B

GAIN STABILITY

The technique of "gain modulation," introduced approximately 1 decade ago, involves the adjustment of receiver gain in synchronism with the modulation frequency (Figure B-1) to provide an equivalent level of noise at the input to the envelope detector during both portions of the switch cycle. In effect, this technique provides a convenient adjustment of the temperature level at either input port without adding noise or changing the temperature of the comparison noise source. The gain modulator technique is sensitive to changes in system noise figure, and must be used with this in mind.

Improvement in receiver gain stability can usually be achieved by the use of well-regulated power supplies, control of the equipment temperature environment, or introduction of temperature compensation networks. More sophisticated techniques for gain stabilization include a variety of noise feedback AGC networks. Comparison of the voltage output from the envelope detector, with an external voltage reference during that half cycle of the modulation frequency in which the signal input of the radiometer is terminated in the comparison load, provides a convenient measure of receiver gain variations. The difference signal derived in this manner can be used in a feedback network to control the receiver gain to a preset level. As in the case of the gain modulator, this technique is also susceptible to changes in system noise figure.

Other techniques for gain stabilization include the introduction of a low-level-modulated noise signal either in quadrature with the radiometer modulation frequency or at some other audio frequency which is then synchronously detected to derive an AGC feedback control voltage.

An additional limitation imposed on radiometer sensitivity in the true sense of detection capability is "baseline drift," defined as drift in the output indicator zero level. Baseline drift may be associated with spatial variations in the background temperature which are detected by the antenna. In radio astronomy applications, spatial temperature variations of this type may be associated with the background cosmic noise level in the vicinity of a celestial source being measured or, alternatively, if the antenna is tracked on target in a sidereal rate, the baseline drift may be associated with variations in the spillover or back lobe source contributions arising as a consequence of antenna motion relative to the local coordinate framework of reference while the antenna is fixed in inertial space. Instrument sources of baseline drift are most frequently associated with dc amplifier instability, differential reradiation temperature contributions from rf losses forward of the switch input ports, reference load temperatures stability, and system noise figure variations when a gain modulator is used.

A typical example of the effect of baseline drift on radiometer sensitivity is shown in Figure B-2 which is an actual strip chart recording of a tunnel diode radiometer operating in a TRF mode at a frequency of 8 gHz. Referring to Figure B-2, the full-scale deflection is 10 $^{\circ}$K; hence each millimeter of deflection is equivalent to 0.2 $^{\circ}$K. The chart speed was 0.2 mm/sec.

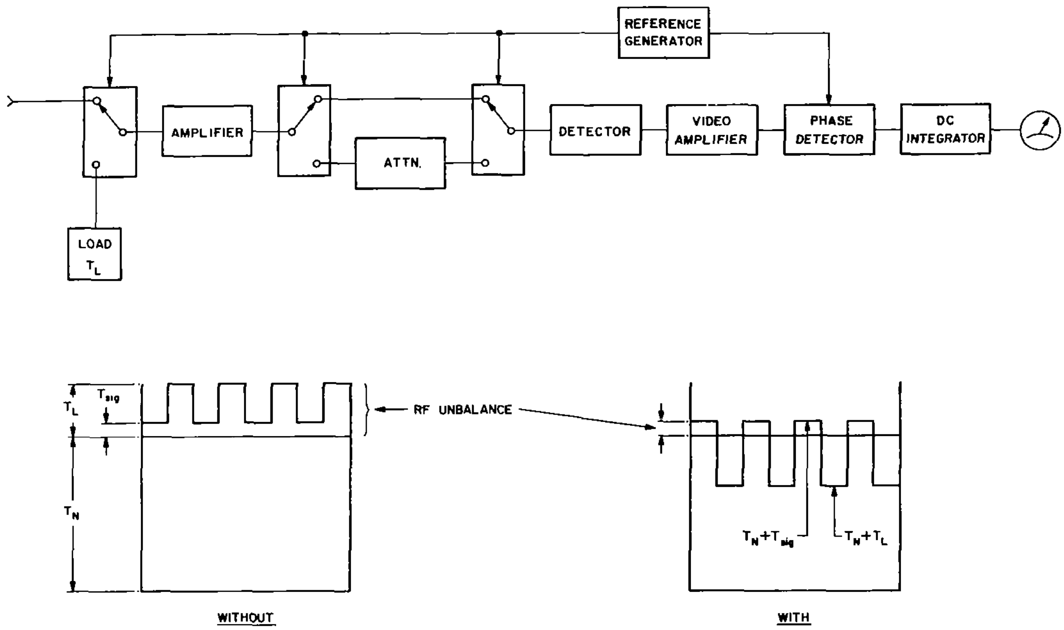

FIGURE B-1

THE GAIN MODULATION TECHNIQUE, TO REDUCE EFFECT OF GAIN VARIATIONS

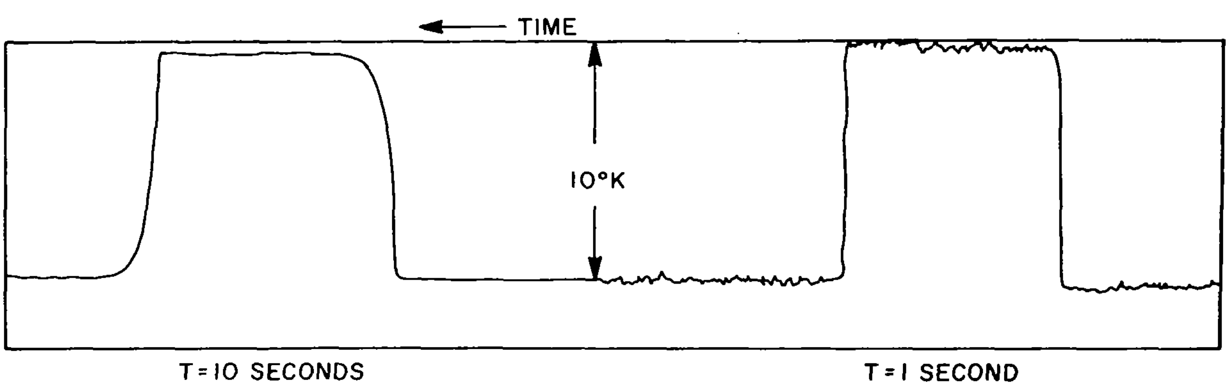

FIGURE B-2

AN EXAMPLE OF BASELINE DRIFT

The total time period shown on the figure is 21 minutes. During the total test period shown, a baseline drift of 2 mm was experienced corresponding to 0.4 ^{o}K. Recalling that the peak-to-peak fluctuation level is approximately six times the rms value, the record shown for the 1-second time constant indicates a rms sensitivity of the order of 0.06 ^{o}K. The rms sensitivity for the 10-second integration time constant would be approximately 0.02 ^{o}K.

It is apparent that the sensitivity of this system for observing periods in excess of a few minutes would be completely determined by baseline instability rather than by the rms level of output indicator fluctuations associated with internal receiver noise and the instantaneous value of the predetection bandwidth, i.e., the magnitude of the baseline drift in 0.5 hour is more than an order of magnitude greater than the rms sensitivity obtained with the 10-second post-detection integration time constant.

Though not associated with baseline drift, it is of interest to note in reference to Figure B-2, the difference in the amplitude response to the calibration signal for the two integration time constants shown. The 10-second integration network shows a 4 percent smaller response than that obtained with the 1-second time constant. This feature is noted here merely to point some of the finer details in radiometer design and operation which must be considered. A functional unit as simple as an RC integration network must be carefully normalized during laboratory adjustment, or alternatively, the operator must be particularly cautious of the output recorder scale calibration differences between modes of operation which represent the slightest difference in signal path circuitry.

APPENDIX C
SKY TEMPERATURE

THE ANTENNA

A microwave radiometric temperature measuring system consists of an antenna, amplifier, detector, integration network, and output indicator.

As a receiving device, the antenna serves to select electromagnetic waves from particular directions with a particular polarization and to present at the antenna output terminals the power extracted from these waves. An antenna is a linear device; thus, the energy which is delivered to the receiver has the same frequency characteristics as the energy extracted from the electromagnetic wave.

The amount of energy absorbed by an antenna and presented at the input terminals of the receiver depends upon the orientation of the antenna, the polarization of the wave, and the impedance match of the receiving system. Since all antennas are polarized, regardless of design, the maximum amount of energy accepted by an antenna, from a randomly polarized wave, is one-half of the total energy content of the wave. The only way in which all polarizations can be accepted from a randomly polarized wave is by using two separate receivers in conjunction with two antennas of complementary polarization. The outputs of the two receivers must then be combined after detection.

If we assume that the antenna is perfectly matched and that the incoming wave is randomly polarized with a flux density of S, then the absorbed power P_a is given by the expression:

$$P_a = 1/2\ SA \tag{C-1}$$

In the above expression, the flux density S of the radiation is assumed to be from a source of small angular size and is measured by the flow of energy from the source through unit area in the wave front at the observing point. If energy dE in the frequency range d ν flows through area dA in time dt (where dt is long compared to the period of one cycle of the radiation), then the flux density S is given by the expression:

$$S = \frac{dE}{dA\ d\nu\ dt} \tag{C-2}$$

which has the dimensions of power per unit area per unit bandwidth.

The antenna receiving cross section A will depend upon the direction in which the plane wave is incident on the antenna. In general:

$$A = A(\theta, \phi) \tag{C-3}$$

where θ and ϕ are the spherical angles representing the direction of incidence of the electromagnetic wave.

ANTENNA EQUILIBRIUM

Consider a transmission line, one end of which is terminated with a matched load and the other end of which feeds an antenna in an absorbing medium. If we were to replace the antenna by its equivalent two-terminal network, and assume that it is a purely resistive impedance and equivalent to the load impedance, then a transmission line terminated in a matched antenna may be treated in a manner similar to a transmission line terminated with a resistive load.

Let there be enough absorbing medium to completely absorb any radiation from the antenna. The medium and the matched termination will then be at the same temperature T.

From Johnson noise power considerations, the termination must radiate a power $kT\,d\nu$ to the antenna. If the antenna, in turn, did not accept $kT\,d\nu$ of radiation from the medium and transfer this power to the load, there would be a net transfer of thermal energy from one region to another at the same temperature without application of work, in violation of the second law of thermodynamics. This would indicate that in the microwave and millimeter portion of the spectrum, the power delivered to the receiving system input by an antenna immersed in an absorbing medium at temperature T is independent of the frequency of observation.

This conclusion can also be reached by noting that the medium appears as a blackbody to the radiation resistance of the antenna, i.e., it absorbs all incident radiation, and since it is at a uniform temperature, it radiates in accord with Planck's Law:

$$B d\nu = \frac{2 h \nu^3}{c^2} \frac{d\nu}{e^{h\nu/kT} - 1} \tag{C-4}$$

where

h = Planck constant

k = Boltzmann constant

c = velocity of light.

and the brightness, B, equals the power per unit area per unit solid angle per unit bandwidth.

From the definition of flux density S, it is evident that:

$$S = \int B \, d\Omega . \tag{C-5}$$

This definition of flux density holds for any source of radiation over all solid angles.

It would appear from equation (C-4) that the power received from a blackbody, measured at the input terminals of the receiving system, would be frequency-dependent. The equivalence of blackbody radiation and Johnson noise power appears inconsistent. The answer to this paradox lies in the characteristic frequency response of any antenna system and the frequency characteristic of blackbody brightness in the millimeter and microwave portions of the spectrum. This can be most easily seen by noting that the power absorbed by an antenna operating in the frequency range $d\nu$, from a randomly polarized source, is:

$$P_r = 1/2 \, A(\theta, \phi) \, B \, d\nu \, d\Omega . \tag{C-6}$$

Hence, for an extended source:

$$P_r = 1/2 \iint A(\theta, \phi) \, d\Omega \, B \, d\nu . \tag{C-7}$$

If the extended source is a blackbody, the power received by the antenna would then be expressed in the form:

$$P_r = 1/2 \iint A(\theta, \phi) \, d\Omega \, \frac{2 h\nu^3}{c^2} \qquad \frac{d\nu}{e^{h\nu /kT} - 1} \tag{C-8}$$

In that portion of frequency spectrum where the energy of the photon h is much less than the random thermal energy per degree of freedom at temperature T, the expression for blackbody brightness (equation (C-4)) takes the simplified form:

$$B \, d\nu \approx \frac{2 k T}{\lambda^2} \, d\nu \qquad \text{(Rayleigh Jeans)} \tag{C-9}$$

where λ is the wavelength of observation.

Consequently:

$$P_r = \frac{kT}{\lambda^2} \, d\nu \int A(\theta, \phi) \, d\Omega \qquad \text{(C-10)}$$

Recalling that the average effective cross-section area for any antenna may be expressed in the form:

$$\overline{A} = \frac{1}{4\pi} \int A(\theta, \phi) \, d\Omega = \frac{\lambda^2}{4\pi} \qquad \text{(C-11)}$$

We immediately arrive at the conclusion that the power received by an antenna immersed in a blackbody at temperature T is frequently independent and equivalent to the Johnson noise power $kT \, d\nu$.

As a consequence, the power received by a microwave or millimeter radiometer is conventionally described in terms of equivalent temperature units.

Since most sources of thermal radiation are not, in fact, blackbodies, the "signal temperature" measured by the radiometer refers to the power level that would be received from a blackbody at a temperature capable of producing an equivalent power level at the output terminals of the antenna.

RERADIATED TEMPERATURES

Because of the general thermodynamic relationship which exists between absorption and emission of any body at any frequency, the emission characteristics of the constituents must also be considered when describing the atmospheric effects.

The emission characteristics of any physical body at a fixed frequency may be compared to those of a blackbody at the same temperatures. In the microwave region, the noise energy emitted by a blackbody is given by the Rayleigh-Jeans Law:

$$B(\nu) = \frac{2\,kT}{\lambda^2} \qquad \text{(C-12)}$$

where

$B(\nu)$ = blackbody brightness

ν = frequency

T = absolute temperature, ^{o}K

λ = wavelength of observation

k = Boltzmann's constant (1.38054×10^{-23} erg) / °K).

For an infinitesimal optical depth, d τ , the radiative transfer differential equation is:

$$\frac{d}{d\tau}\left[I(\nu)\right] = -I(\nu) + B(\nu) \tag{C-13}$$

where I(ν) is the intensity of the radiation. Analogous to the temperature dependence of noise energy (equation (C-13)), one may, in the microwave region, relate the intensity of radiation, I(ν), received from a particular direction to an equivalent temperature T_b(ν), called the thermal noise temperature or the brightness temperature, by the following relation:

$$I(\nu) = \frac{2k\,T_b\,(\nu)}{\lambda^2} \tag{C-14}$$

Solution of equation (C-14) for the entire atmosphere is:

$$I(\nu) = I_m\,(\nu)e^{-\tau} + \int_0^{\tau} B(\nu)e^{-\tau}\,d\tau \tag{C-15}$$

where I_m (ν) is the unattenuated brightness of a discrete source (such as the sun or moon) and τ is the optical depth of the entire atmosphere, which can be expressed as:

$$\tau = \int_0^{\infty} \gamma(h)\,dh \tag{C-16}$$

where

$\gamma(h)$ = absorption coefficient at height h

dh = incremental height.

Substituting the expressions of I(ν) and τ we obtain:

$$T_b\,(\nu) = T_b m\,(\nu)e^{-\tau} + \int_0^{\tau} T\,(\tau)\,e^{-\tau}\,d\tau. \tag{C-17}$$

The first term of equation (C-17) represents the contribution to the brightness temperature of an extraterrestrial discrete source; the second term represents the contribution to the total brightness temperature from the atmosphere itself. It should be noted that this expression for the source and sky brightness temperatures may be integrated in closed form, only for the most simple of atmospheric models.

SKY TEMPERATURE AS A FUNCTION OF ZENITH ANGLE

If measured by a hypothetical antenna of infinitely narrow beamwidth, the brightness temperature of the atmosphere will depend upon the spatial orientation of the antenna (satisfactorily represented by the initial zenith angle, θ , of the antenna) and the frequency of the antenna operation.

Since the total path length through the atmosphere increases with zenith angle, a corresponding increase in absorption and radiation occurs. For a plane earth, the path length is proportional to the secant of the zenith angle, and if the atmosphere is horizontally stratified:

$$\tau(\theta) = \tau \sec \theta. \tag{C-18}$$

The secant law is widely used in absorption calculations for the case of the spherical earth down to zenith angles of about 75 degrees. The relation can, however, also be used at larger angles with fairly good accuracy because of the increasing predominance of low altitude absorption with increasing zenith angle.

The apparent sky temperature at a zenith angle θ is given by:

$$T_s(\theta) = T_m(1 - e^{-\tau \sec \theta}) \tag{C-19}$$

where

T_m = the effective average temperature along the raypath through the atmosphere.

Average values of $T_s(\theta)$ were measured by Wulfsberg in 1964 at A. F. C. R. L., Bedford, Mass. for frequencies of 15, 17, and 35 gHz, respectively. Equation (C-19) has been plotted in Figure C-1 for various values of τ where T_m = 262 $^{\circ}$K.

National Aeronautics and Space Administration
Electronics Research Center
Cambridge, Massachusetts, May 1967
125-21-02-26-25

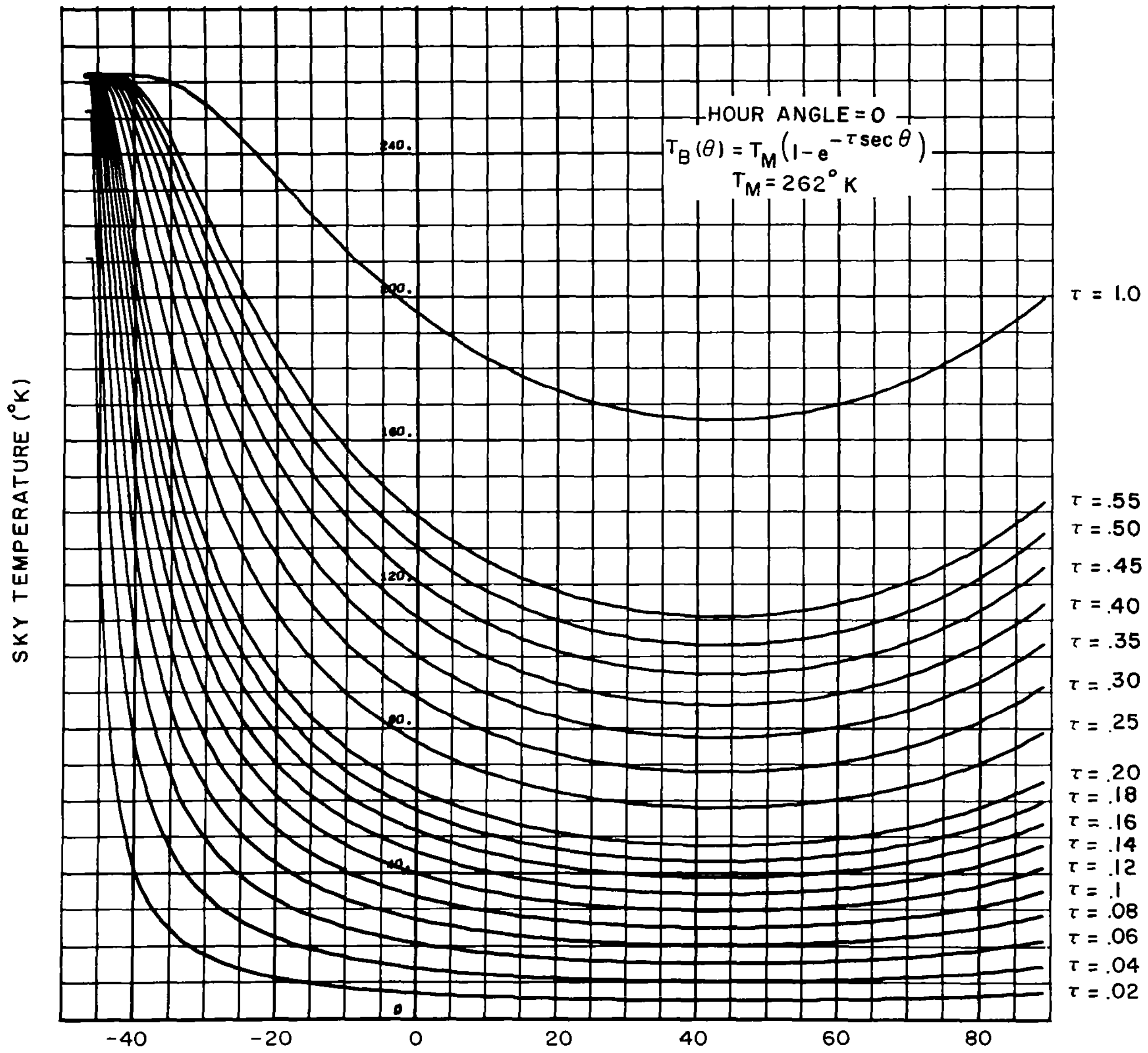

FIGURE C-1

DECLINATION (FOR BOSTON)

NASA-Langley, 1967 —— 7