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A SWITCHED LOAD RADIOMETER

Torleiv Orhaug and William Waltman

ABSTRACT

An analysis of the stability of a switched radiometer with a general detector law is carried out. The method of gain modulation as a way of balancing the radiometer is introduced, and it is shown that in this mode of operation the system is not completely insensitive to variations in radiometer noise temperature. The effect of instabilities in a system having non-linear elements in front of the gain modulator is discussed. A detailed investigation is carried out for the two cases when the comparison switch is located at the front-end and at the IF-level in a mixer-type radiometer, and it is shown that even though the suppression of instabilities is greatest in a front-end switched system, switching at the IF-level may have advantages. The corrections necessary for departure from a square law detector are given. A highly stable switched load continuum radiometer which has been constructed using the principle of gain modulation is discussed. Apart from the front-end switch and the gain modulator, which are described in detail, the radiometer has been entirely assembled from commercially available equipment. Records of receiver performance when the radiometer is used both as total power and as a switched load radiometer are discussed.

I. INTRODUCTION

One of the main problems in high sensitivity radiometer technique is to lower the effect of instabilities in the system so that the output fluctuations are governed by noise fluctuations. The principle solution is to remove the DC-term in the radiometer's output, which is determined by the total system noise temperature, and there are three ways in which this can be done. The first is to rapidly compare the signal with a constant comparison temperature source (the celebrated switching technique introduced by Dicke (1946)). In the second method the comparison between the signal source and the comparison source is done continuously and the difference is measured, but this method can only be used for observing spectral features (the DC-comparison technique introduced by Selove (1954)). In the third method the signal is compared with the comparison source using a correlation technique. (This method was first published by Tucker (1955)). The last method has widely been used for interferometric observations, but many technical difficulties arise in the application to a single dish type telescope. The switching technique is, therefore, still the most important principle for continuum radiometers using a single antenna. Even though the importance of instabilities in the system has been pointed out (J. C. Greene (1957)) a detailed treatment of the effect of all the parameters in the system has never been published. An attempt to such a detailed investigation is

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made in this report, and section II contains the results of this analysis. A general detector law is assumed, and the effect on the output signal of a number of sources of instability in the system is treated. A switched load radiometer for continuum observations has been developed, using some of the principles in section II, and this radiometer is described and discussed in section III. The few components which are not commercially available are described in detail, and test records of the receiver's performance are discussed.

II. STABILITY ANALYSIS OF A SWITCHED RADIOMETER

In this section we shall analyze the stability of a switched radiometer in order to determine the output fluctuations that can be expected when changes occur in various radiometer parameters. It is well known that the statistical noise fluctuations are only one of the many sources of more or less gaussian variation of a radiometer's output. Random and non-random variations of the important parameters of a radiometer such as gain, bandpass characteristics and receiver noise temperature due to such causes as voltage

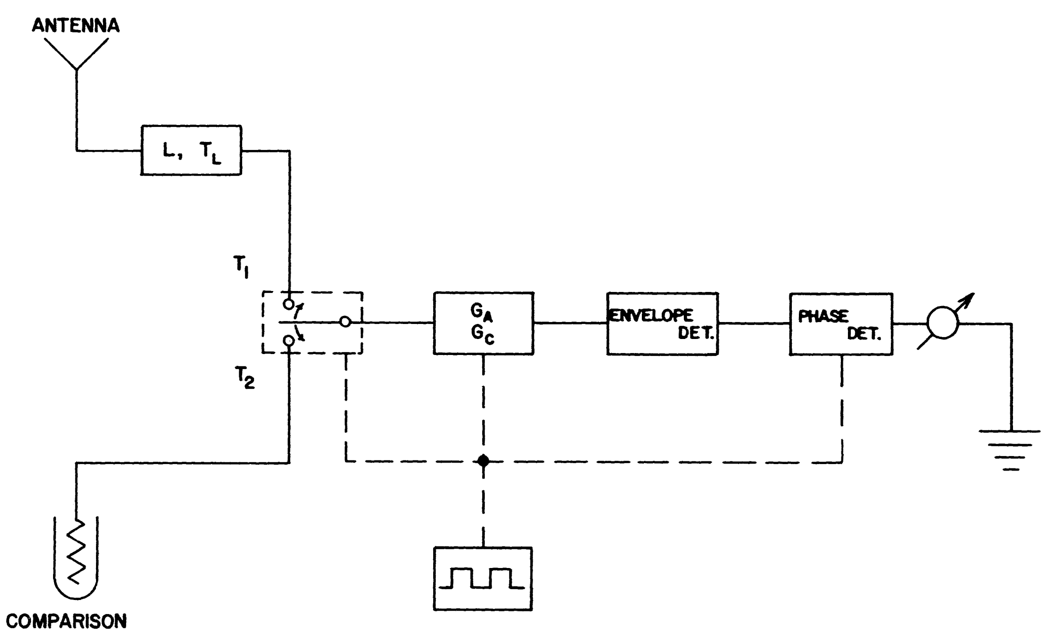

Fig. 1.— Principle arrangement of a switched receiver.

changes and ambient temperature changes often cause corresponding output variations which add to the true noise fluctuations. The ultimate goal in building any radiometer for which the noise temperature, pre-detection bandwidth, and post-detection integration time are already fixed, is to arrange that no other effects make the output fluctuations sensibly greater than those set by these parameters. If such a goal is achieved, the receiver is said to be "noise limited" and if the goal is not achieved, the receiver is said to be "stability limited". The most important step in this direction was the introduction of the switched comparison technique by Dicke, and most of the radiometers developed since that time make use of this technique in one form or another.

1. <u>Derivation of the Instability Equation.</u>

For the treatment of the stability problem of the receiver, the principle diagram of Figure 1 will be used. The antenna signal T_A is assumed to be fed through a lossy pad of loss L and temperature T_L, and this will later on enable us to place the comparison switch either before or after the microwave mixer, which will be assumed to be the front-end of the radiometer. After passing through the pad, the antenna signal is periodically compared with a comparison temperature source with the same impedance as the signal source. The temperature of the comparison source is taken to be T_C.

The input terminals to the switch have the following equivalent temperatures (Figure 1)

$$T_1 = \frac{T_A}{L} + (1 - 1/L)T_L$$

$$T_2 = T_C$$

In order to investigate the gain modulation technique for receiver balancing, we assume that the total receiver gain from the output of the comparison switch to the envelope detector is G_A and G_C for the two switch positions, respectively. In other words, the gain of the receiver is switched simultaneously with the front-end switch. The input power to the detector is given by

$$\begin{aligned} P_1 &= k(T_1 + T_R)BG_A \\ &= k\left[\frac{T_A}{L} + (1 - 1/L)T_L + T_R\right]BG_A \end{aligned} \tag{1a}$$

and

$$\begin{aligned} P_2 &= k(T_2 + T_R)BG_C \\ &= k(T_C + T_R)BG_C \end{aligned} \tag{1b}$$

where we have assumed a perfect switch (no insertion loss and infinite isolation) and where k is Boltzman's constant, B is the overall noise bandwidth and T_R is the noise temperature of the parts of the receiver after the switch.

We assume that the envelope detector obeys a general power law,

$$V = C_1 P^{\beta}$$

where V is the DC-output voltage from the detector, P is the input power to the detector and C_1 and β are constants. The DC-outputs for the two switch positions are therefore

$$V_1 = C_1 \left[k(T_1 + T_R) B G_A \right]^{\beta}$$

$$V_2 = C_1 \left[k(T_C + T_R) B G_C \right]^{\beta}$$

and the DC voltage v derived from the phase sensitive detector is then

$$v = C_2 (V_1 - V_2)$$

where C_2 is a constant. Inserting the expressions for V_1 and V_2 we obtain

$$v = C_1 C_2 (kBG_A)^{\beta} \left[(T_1 + T_R)^{\beta} - K^{\beta} (T_C + T_R)^{\beta} \right] \tag{2}$$

where

$$K = \frac{G_C}{G_A} \tag{3}$$

is the gain ratio (gain modulation factor).

In order to study the effect on the output signal of changes in the receiver parameters G_A, T_R, T_L, L, K, B, and T_C we will assume that these parameters are subject to small changes ΔG_A, ΔT_R, etc., and we will refer the corresponding changes in the output to the equivalent input signal change, ΔT_A, which would produce the same output changes. Therefore, in terms of equivalent input temperature change ΔT_A, the variation in output signal caused by small variations in the u_1, u_2 ... u_7 is

$$\left(\frac{\partial v}{\partial T_A} \right)^2 \overline{(\Delta T_A)^2} = \sum_{u_1 \, .. \, u_7} \left(\frac{\partial v}{\partial u} \right)^2 \overline{(\Delta u)^2} \tag{4}$$

where $u_1 = G_A$, $u_2 = T_R$, $u_3 = T_L$, $u_4 = L$, $u_5 = K$, $u_6 = T_C$, and $u_7 = B$. In Equation (4) we have added the mean square values of the fluctuation terms. This will give the correct result if the variations in the parameters u_1 ... u_7 are statistically independent. This is, however, a point of uncertainty. Part of the variations are, for instance, temperature dependent, and in this case we should expect a high correlation between variations of several of the parameters in Equation (4). For correlated changes, the output fluctuations are added linearly, thereby increasing the output fluctuations, compared to the case where the variations are uncorrelated. The correlated case represents the worse possible case; for zero or partly correlated variations, the total output fluctuations are less than for the correlated case. Under the assumption of statistical independence, and omitting output fluctuations caused by the total noise temperature of the system (see III, 4), the total receiver output fluctuations are given by the instability equation

$$\begin{aligned}
\overline{(\Delta T_A)^2} = {} & L^2(T_1 + T_R)^2 \left[\frac{T_C + T_R}{T_1 + T_R}\right]^{2\beta} \left[\left(\frac{T_1 + T_R}{T_C + T_R}\right)^{\beta} - K^{\beta}\right]^2 \overline{\left(\frac{\Delta G}{G}\right)^2} \\
& + L^2\left[1 - K^{\beta}\left(\frac{T_C + T_R}{T_1 + T_R}\right)^{\beta - 1}\right]^2 \overline{(\Delta T_R)^2} \\
& + (L-1)^2\, \overline{(\Delta T_L)^2} + (T_L - T_A)^2 \overline{\left(\frac{\Delta L}{L}\right)^2} \\
& + L^2(T_1 + T_R)^2 \left(\frac{T_C + T_R}{T_1 + T_R}\right)^{2\beta} \left[\left(\frac{T_1 + T_R}{T_C + T_R}\right)^{\beta} - K^{\beta}\right]^2 \overline{\left(\frac{\Delta B}{B}\right)^2} \\
& + L^2(T_1 + T_R)^2 \left(\frac{T_C + T_R}{T_1 + T_R}\right)^{2\beta} K^{2\beta} \overline{\left[\frac{\Delta K}{K}\right]^2} \\
& + L^2 K^{2\beta}\left(\frac{T_C + T_R}{T_1 + T_R}\right)^{2(\beta - 1)} \overline{(\Delta T_C)^2}
\end{aligned} \tag{5}$$

2. Discussion of the Instability Equation.

We will now discuss the influence of variations in the various receiver parameters based on Equation (5).

a. The gain variation term.

The quantity multiplying $\overline{\left(\frac{\Delta G}{G}\right)^2}$ in Equation (5) can be made zero and the receiver thus made insensitive to gain variations only if

$$K = \frac{T_1 + T_R}{T_C + T_R} \tag{6}$$

and this condition holds irrespective of the detector law. For an ordinary receiver where $G_A = G_C$ and therefore $K = 1$, this means that the inputs to the switch should be balanced, or $T_1 = T_2$ to ensure insensitivity to gain changes. For unbalanced inputs, on the other hand, Equation (6) gives the condition for the gain unbalance in order to produce a balanced output. The stability of a receiver balanced by adding noise to the channel of lowest noise temperature has been successfully demonstrated by Drake and Ewen (1958). The other method of reducing the effect of gain variations is according to the result above to alter the gain of the receiver in synchronism with the switch, so that the output is balanced for the two positions of the switch. In this method, which was first introduced by the Ewen-Dae Corporation, the ratio of the gain is given by Equation (6).

b. Change of receiver temperature T_R.

Equation (5) shows that any of the following possible conditions will give a system independent of possible changes in T_R, the noise temperature of the receiver after the switch:

a) Square law detector ($\beta = 1$), $K = 1$, $T_1 \neq T_2$. This means that the gain is balanced. In this case the system is not balanced at the input ($T_1 \neq T_2$).

b) Any detector law (β arbitrarily), $K = 1$, $T_1 = T_2$. The system is insensitive to noise temperature changes for any detector law provided that the gain is balanced and the input is balanced ($T_1 = T_2$).

c) If the condition

$$K = \left(\frac{T_1 + T_R}{T_C + T_R}\right)^{\frac{\beta-1}{\beta}} \tag{7}$$

is fulfilled, then the system also is insensitive to changes in noise temperature. This is, however, an impractical case, because this gain unbalance is not the same as the unbalance which gives cancellation of gain instability.

When the receiver is operated with a gain unbalance according to Equation (6) in order to obtain insensitivity to gain changes, the system is sensitive to noise temperature changes. A change in noise temperature of ΔT_R is equivalent to an input signal temperature change ΔT_A, where

$$\Delta T_A = L(1 - K)\, \Delta T_R \ .$$

It should be pointed out, however, that in any switched system there is a difference in gain in the two switch positions. The magnitude of the difference depends upon the difference in impedance in the two switch positions and may be made smaller by using an isolator between the switch and the receiver. This means that in any system the second term in Equation (5) will be of some significance, even for a standard type radiometer with no gain modulator and using a square law detector.

In the total power receiver (a receiver where the DC signal from the envelope detector is directly recorded) the dependence upon the changes in noise temperature is ΔT_R, irrespective of the detector law. We therefore notice that in the switched receiver using a gain modulator producing balanced output, the dependence has been reduced by a factor (1 - K) compared with the total power case. For an 800 °K receiver connected to a 50 °K antenna and using a comparison load at 300 °K, the factor K = 0.77, and therefore

$$\Delta T_A = 0.23\, \Delta T_R .$$

For the same receiver with no gain modulator, having a gain unbalance of K = 0.95, the dependence upon receiver temperature changes would be

$$\Delta T_A = 0.05\, \Delta T_R .$$

assuming a square law detector.

c. Changes of the temperature and the attenuation of the loss L.

The system is always sensitive to changes in the temperature T_L of the pad in front of the switch as long as $L \neq 1$, and the same is true for changes in the loss if the source temperature is different from the temperature of the loss.

d. Changes in the noise bandwidth B.

Since the receiver gain G_A and the effective noise bandwidth B only enter through their product in the expression giving the output DC-voltage (Equation (2)), the instability term for variations in the bandwidth B is identical to the unstability term for variations in the gain (Equation (5)). The system is thus insensitive to changes in bandwidth if it is balanced, either by adding noise at the front-end making $T_1 = T_C$ or by introducing a gain unbalance according to Equation (6).

e. Changes in gain ratio K.

If the gain ratio K is adjusted for a balanced condition, then the output variation caused by a change in the gain ratio is proportional to

$$L(T_1 + T_R)\left(\frac{\Delta K}{K}\right) \tag{8}$$

that is, it is proportional to the total temperature to the input of the switch.

3. Discussion of Non-Linear Effects.

If a switched receiver is balanced using the gain modulation technique, then there are no changes of the output signal due to variations in receiver parameters (gain, detector characteristics, etc.) for those parts of the receiver following the gain modulator. This is not necessarily true for the parts of the receiver which are located between the switch and the gain modulator. In order to show that non-linear elements between the switch and the gain modulator may produce instabilities, we assume that the gain is a function of the total noise level T, say g(T) -- then we have for the two positions of the switch

1. Input level: $T_1 + T_R$; gain: $G_1 = g(T_1 + T_R)$

2. Input level: $T_2 + T_R$; gain: $G_2 = g(T_2 + T_R)$

Prior to an observation, the system is balanced, which means that

$$(T_1 + T_R) \cdot G_1 = (T_2 + T_R) \cdot G_2 \tag{9}$$

giving the initial gain ratio for balance,

$$K_o = \frac{G_2}{G_1} = \frac{T_1 + T_R}{T_2 + T_R} \quad . \tag{10}$$

The gain function and the balanced conditions are depicted in Figure 2.

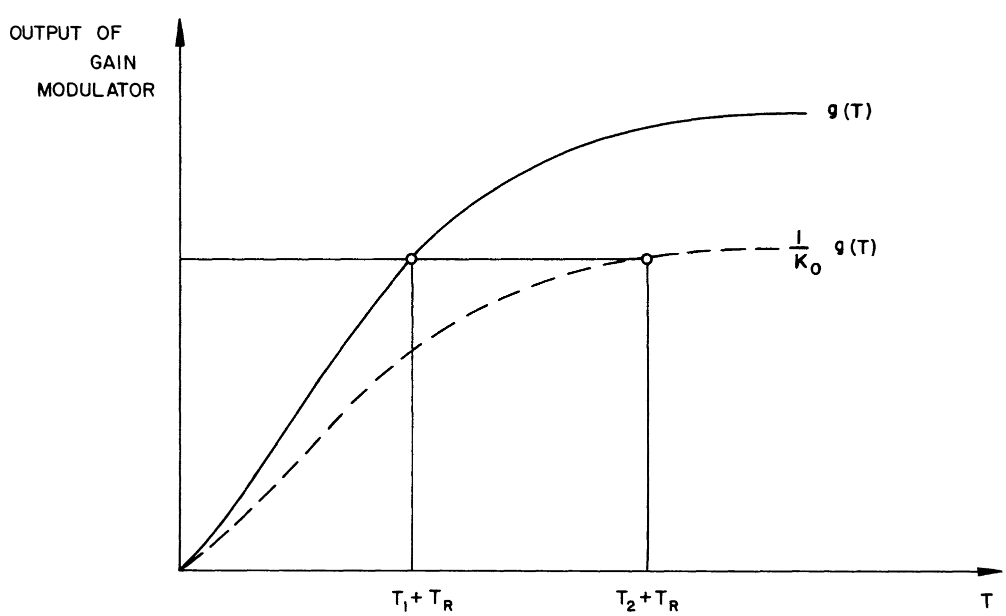

Fig. 2. — The effect of the gain modulator in balanced condition.

We now assume that the gain function g(T) has changed to g(T) $[1 + \psi(T)]$ where $\psi(T)$ is a function of T. Even if the system were balanced prior to the change of the gain function, the system is now unbalanced, and the output unbalance corresponds to an input signal ΔT in channel 1,

$$\Delta T = [T_1 + T_R]\,[K - K_o]$$

where K and K_o are the actual and the initial gain ratios, respectively. The new gain ratio is

$$K = K_o \frac{1 + \psi(T_2 + T_R)}{1 + \psi(T_1 + T_R)}$$

giving

$$\Delta T = K_o[T_1 + T_R]\frac{\psi(T_1 + T_R) - \psi(T_2 + T_R)}{1 + \psi(T_1 + T_R)} \tag{11}$$

The system is therefore independent of variations in the gain function provided that

$$\psi(T_1 + T_R) = \psi(T_2 + T_R)$$

This condition is fulfilled, for instance, if $\psi(T)$ = constant, and in this case the variation of the gain function takes place with no change in <u>shape</u>. In order to avoid this non-linear effect, the gain modulator should be placed electrically as near the switch as possible,

<u>4. Front-End Switching</u>.

From a stability point of view, the switch should be placed as near the signal source (antenna) as possible. Dependent upon the observing frequency, this may be difficult, because of the limited possibility of obtaining fast acting switches with low insertion loss and wide bandwidth.

If we assume the switch to be at the front-end, we can neglect the loss in the signal channel between the source and the switch, and the instability terms are then, for the square law case ($\beta = 1$), omitting the bandwidth variation

$$\overline{(\Delta T_A)^2} = \Big[(T_A + T_R) - (T_C + T_R)K\Big]^2 \overline{\left(\frac{\Delta G}{G}\right)^2} + (1 - K)^2\,\overline{(\Delta T_R)^2}$$

$$+ (T_C + T_R)^2 K^2 \overline{\left(\frac{\Delta K}{K}\right)^2} \tag{12}$$

For a crystal mixer receiver of the type described in section II, the noise figure is

$$F = L_{CR}(t_{CR} + F_{IF} - 1)$$

where L_{CR} is the loss of the crystal, $t_{CR} = \frac{T_{CR}}{T_o}$ is the noise ratio of the crystal and F_{IF} is the noise figure of the IF-amplifier following the crystal mixer. In terms of noise temperature we can write

$$T_R = \left\{ L_{CR} \left[\frac{T_{CR}}{T_o} + F_{IF} - 1 \right] - 1 \right\} T_o$$

or

$$T_R = L_{CR} \left[T_{CR} + T_{IF} \right] - T_o$$

where T_{CR} and T_{IF} are the noise temperatures of the mixer crystal and the IF-amplifier, respectively. If we only consider the instability term caused by variation in receiver noise temperature, then this term can be shown to be

$$\overline{(\Delta T_A)^2} = L^2_{CR} (1 - K)^2 \left[(T_{CR} + T_{IF})^2 \overline{\left(\frac{\Delta L_{CR}}{L_{CR}} \right)^2} + \overline{(\Delta T_{CR})^2} + \overline{(\Delta T_{IF})^2} \right] \tag{13}$$

where we have assumed a square law detector ($\beta = 1$), no loss in front of the switch ($L = 1$ and $\Delta L = 0$), and the system balanced according to Equation (6). We will later compare this expression with the corresponding one for a system having the switch placed between the mixer and the IF-amplifier.

It is obvious from the previous discussion that the receiver at the input to the switch should be as near balance as possible in order to increase the overall stability. This is not difficult to obtain with a receiver having a relatively high noise temperature. A receiver in the microwave region having a noise temperature of 1000 °K connected to an antenna having a temperature of 50 °K, will give $K = 0.81$ if the comparison load is at room temperature. For receivers with lower noise temperature (parametric amplifiers, masers) the same difference between antenna and load temperature would give a value of K much smaller than unity and therefore would make the system very sensitive to variations in receiver temperature. In this case, a comparison channel with lower noise temperature is needed (comparison antenna or cooled termination).

5. IF-Switching.

If a microwave switch is not available, or if for some reason switching at the front-end is not preferable, the stability of the total power receiver can be greatly improved by using a switch at the IF-level. In this case the switch should be placed between the mixer and the first IF-amplifier. The instability terms caused by variations in mixer- and IF-parameters are now

$$\overline{(\Delta T_A)^2} = (T_{CR} - T_A)^2 \overline{\left[\frac{\Delta L_{CR}}{L_{CR}} \right]^2} + (L_{CR} - 1)^2 \overline{(\Delta T_{CR})^2} + L^2_{CR} (1 - K)^2 \overline{(\Delta T_{IF})^2} \tag{14}$$

The input to the switch is in this case

$$T_1 = \frac{T_A}{L_{CR}} + (1 - \frac{1}{L_{CR}}) T_{CR}$$

and for an ordinary crystal mixer, $T_{CR} \simeq 1.2 \cdot 290$ °K and $L_{CR} \simeq 4$, giving $T_1 \simeq 282$ °K. The equivalent temperature of the mixer output is therefore not too far from room temperature, as was pointed out by Ewen (3). This means that a comparison load at room temperature gives nearly balanced condition at the input to the switch.

A comparison between Equations (13) and (14) gives the following results:

1. The instability of the front-end switched system with respect to variation in crystal mixer loss is decreased by a factor

$$\frac{1}{L_{CR}(1 - K)} \frac{T_{CR} - T_A}{T_{CR} + T_{IF}}$$

   as compared to the IF-switched system. Assuming the following values: $L_{CR} = 4$, $K = 0.80$, $T_C = 290$ °K, $T_A = 50$ °K, $T_{CR} = 350$ °K, and $T_{IF} = 90$ °K, the fluctuations at the output of the front-end switched system are 82 per cent of the fluctuations as compared to the IF-switched system, when variations in mixer loss are considered.

2. The instability of the front-end switched system with respect to variation in crystal noise temperature is decreased by a factor

$$\frac{1 - 1/L_{CR}}{1 - K}$$

   as compared to the IF-switched system.

   Assuming the same values for L_{CR} and K as used above, we find that the ratio of the output fluctuations for the IF-switched system as compared to the front-end switched system is 5.35. This ratio will be greater than unity as long as

$$KL_{CR} > 1 .$$

3. The instability of the front-end switched system with respect to variation in IF-noise temperature is equal to the instability of the IF-switched system.

6. Output Non-Linearity Caused by Detector Characteristics.

One complexity in highly accurate radiometric observations is the non-linearity of the receiver response caused by departure of the detector from a true square law. The input power to the detector for the two switch positions is, according to Equations (1a) and (1b) given by

$$P_1 = k(T_1 + T_R)\, BG_A$$

$$P_2 = k(T_2 + T_R)\, BG_A K$$

We again assume a general power law detector

$$P = C_3 V^{\alpha} \tag{15}$$

where $\alpha = 1/\beta$ compared to the previous expression.

If the receiver is balanced prior to observation, then

$$K = \frac{T_1 + T_R}{T_2 + T_R} \; .$$

The temperature in channel 1 (antenna channel) is now assumed to increase by an amount ΔT, giving rise to a corresponding increase in output voltage, ΔV, and

$$G_A k(\Delta T + T_R)B = C_3(V + \Delta V)^{\alpha}$$

giving

$$\frac{\Delta T}{T_1 + T_R} = \left(1 + \frac{\Delta V}{V}\right)^{\alpha} -1 \; . \tag{16}$$

Series expansion of Equation (16) or differentiation of Equation (15) now gives the relation between <u>small</u> input temperature increases and corresponding output level change,

$$\left[\frac{\Delta T}{T_1 + T_R}\right]_{\Delta T \rightarrow 0} = \frac{\alpha}{V} \cdot \Delta V$$

showing that there exists a linear relationship between input and output for small input changes, and the proportional factor is dependent upon the detector law and the signal level at the detector. If we assume that it is possible to find a proportional factor for small input changes by calibration at the front-end, then it is of interest to find the factor μ, which we define as the ratio of the proportional factor for arbitrary input changes to the proportional factor for small input changes,

$$\mu = \frac{\frac{\Delta T}{T_1 + T_R}}{\left[\frac{\Delta T}{T_1 + T_R}\right]_{\Delta T \to 0}} = \frac{\left[1 + \frac{\Delta V}{V}\right]^{\alpha} - 1}{\alpha \frac{\Delta V}{V}} \tag{17a}$$

and the factor μ thus gives the correction necessary when going from small input changes to large or arbitrary ones. The correction factor μ was first given by Seeger, et al (1956).

It may be more convenient to relate the correction factor to the input temperature rather than to the output detector levels. We have then

$$\frac{\Delta T}{T_t} = \frac{\Delta V}{V}$$

where $T_t = T_1 + T_R$ giving for

$$\mu = \frac{\left[1 + \frac{1}{\alpha}\frac{\Delta T}{T_t}\right]^{\alpha} - 1}{\frac{\Delta T}{T_t}} \ . \tag{17b}$$

A graph showing μ as a function of α and $\Delta T/T_t$ is given in Figure 3.

III. DESCRIPTION OF A SWITCHED LOAD CONTINUUM RADIOMETER

In the design of high frequency radiometers for continuum observations the basic requirements are reliability, stability, and low noise temperature. In addition, ease of construction and moderate cost are quite desirable. Except for low noise temperatures, these are principally characteristics of the back-end part of the receiver, which is normally identical for all operating frequencies. It is therefore possible to design a standard back-end which may be combined with various front-end units to provide many combinations of noise figure, price, complexity, and operating frequency.

The design which has been developed and which is presented here is quite satisfactory in many cases in its basic form but is also well suited to many types of modifications to meet specific observing requirements. Standard, commercially built units have been used wherever possible. This simplifies development and construction of the radiometer and facilitates the stocking of identical replacement units for servicing.

The stability, cost, and overall performance of the radiometer are the result of the overall design and of the characteristics of the commercial items used. The design described in this paper provides a complete radiometer, including recorder, for a low cost. It exhibits gain stability on the order of 1 per cent for normal ambient conditions, base line stability of better than 1 °K over 24 hours, and has a bandwidth of 8 mcs, and a noise temperature less than 1000° at L-band. It is suitable for any frequency from .3 to 2 Gc.

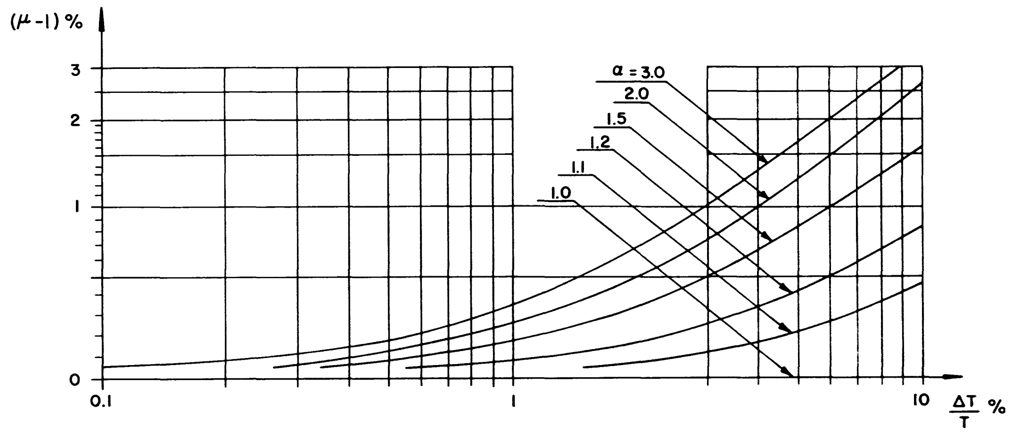

Fig. 3. — Graph showing μ as a function of $\frac{\Delta T}{T_t}$ for various values of α.

1. Description of the System.

A block diagram of the receiver is shown in Figure 4. Table 1 lists the components used and their manufacturers. All the items are commercially available except for the gain modulator, integration network, reference generator and front-end switch which were designed and built at the Observatory.

Items 1-13 form the back-end part of the receiver and with the exception of the local oscillator are independent of the radiometer frequency. Items 14-18 form the front-end of the receiver and are normally located near the focal point of the antenna.

Except for the addition of a means of balancing, the radiometer is of the conventional switched type originated by Dicke (1946). The receiver is balanced by means of the gain modulator, an attenuator which is switched between two settings in synchronism with the front-end switch and phase detector. For balance, it is set to provide equal noise levels at the detector for both switch positions. It is described more fully in section III 3. The output of the crystal diode detector is fed through a 400 cycle tuned amplifier to the phase detector which produces a DC output proportional to the difference in detected noise level in the two switch positions. An R-C integration network provides time constants from 2 to 60 seconds for the recorder output and an additional output with a 2-second time constant to drive a digital output system.

2. Discussion of the Front-End Switch.

The front-end switch may be represented by a passive three port network feeding into

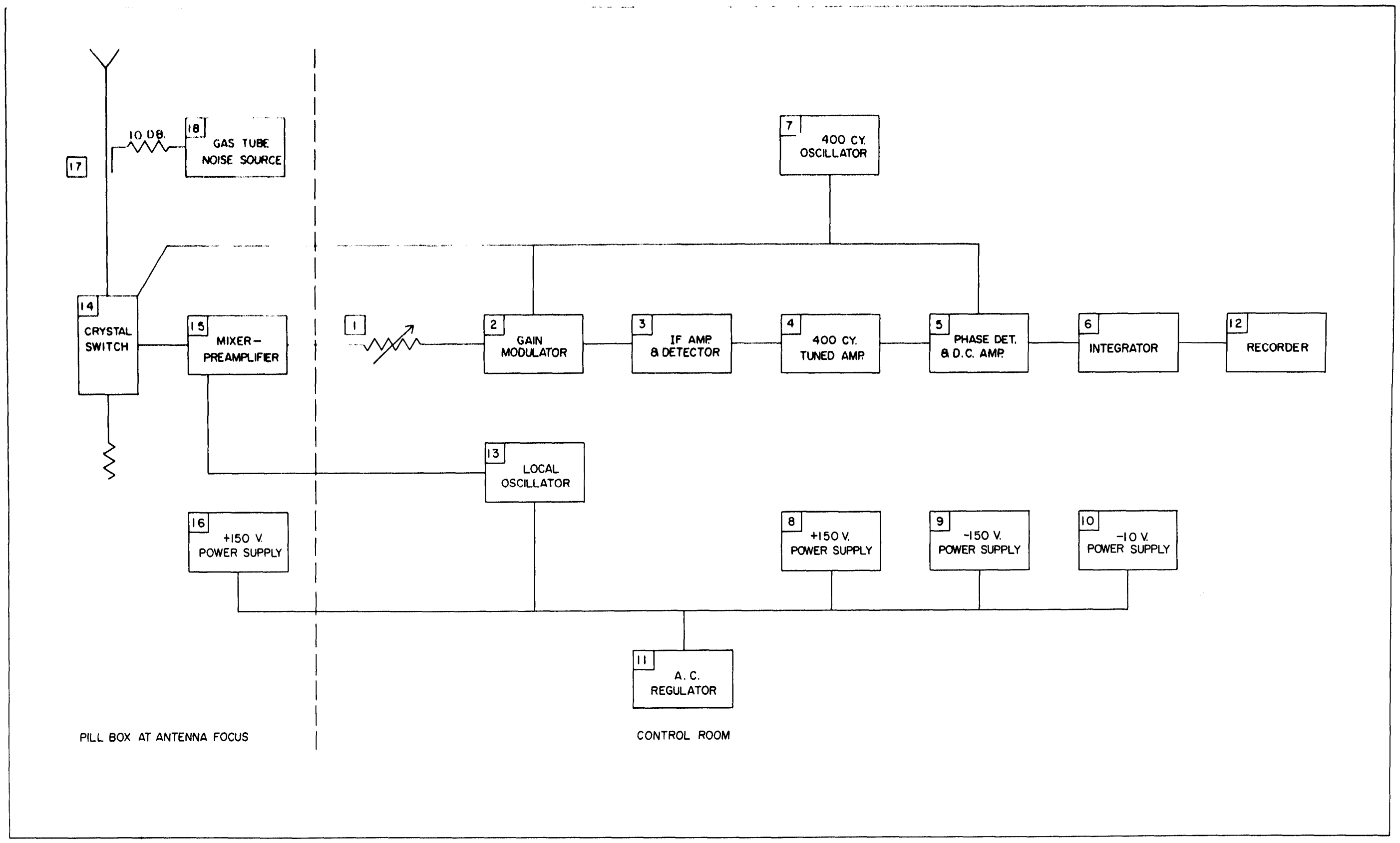

Fig. 4. — Block diagram of a standard switched radiometer.

TABLE 1

RECEIVER COMPONENTS

No.	Item	Company	Model
1.	IF attenuator	Kay	20-0
2.	Gain modulator	NRAO	
3.	Main IF amplifier	LEL	IF20B
4.	Tuned 400 cy amplifier	White	212
5.	Phase detector and amplifier	Sanborn	850-1200
6.	Integration network	NRAO	
7.	400 cps oscillator and amplifier	NRAO	
8.	+150 v. DC supply	AC-DC Electronics	RPM 200-150
9.	-150 v. DC supply	AC-DC Electronics	RPM 50-150
10.	-10 v. DC supply	Sorensen	QM9-.22
11.	AC voltage regulator	Sorensen	1001A
12.	Recorder	Sanborn	152-100B and 140-2900Z amplifier
13.	Local oscillator	General Radio	1217-A with 1201-B power supply
14.	Front-end switch	NRAO	
15.	Mixer-preamplifier	LEL	LAC-3
16.	Power supply	AC-DC Electronics	RPM 50-150
17.	20 db directional coupler	Narda	3002-20
18.	Noise generator	AIL	7010
19.	Two relay racks	Par	PX 6724A

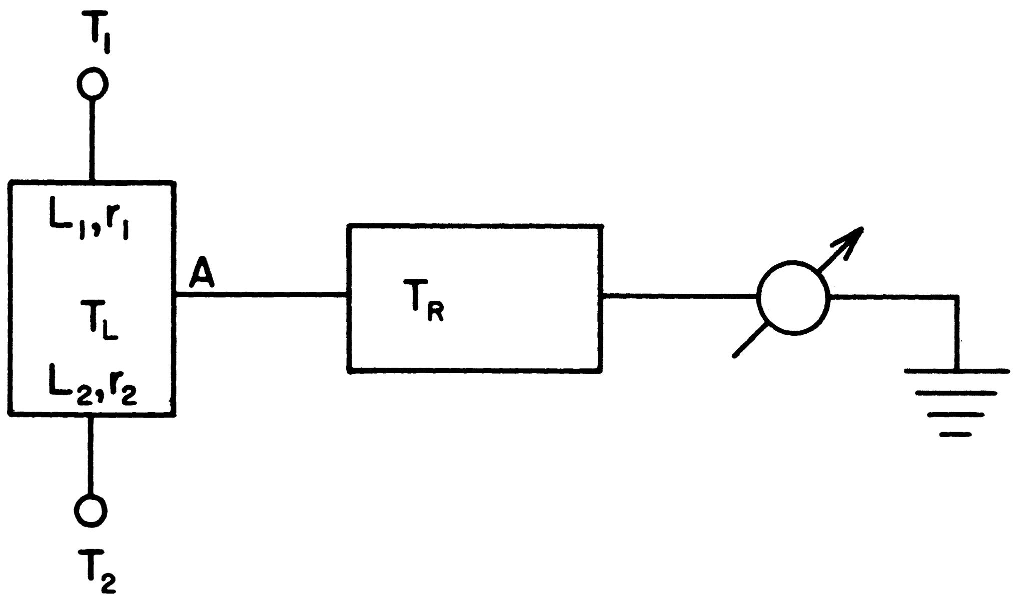

Fig. 5. — Representation of a front-end switch.

a receiver with noise temperature T_R as shown in Figure 5. When the switch is in position I it has losses L_1 and r_2L_2 from inputs 1 and 2 to the output. In position II these become r_1L_1 and L_2, respectively. The loss temperature of the switch is T_L.

Assuming a square law detector, the receiver output with the switch in position I, referred to point A, is

$$T_I = \frac{T_1}{L_1} + \frac{T_2}{r_2L_2} + \left(1 - \frac{1}{L_1} - \frac{1}{r_2L_2}\right)T_L + T_R \tag{18a}$$

In position II the output referred to the same point is

$$T_{II} = K\left[\frac{T_1}{r_1L_1} + \frac{T_2}{L_2} + \left(1 - \frac{1}{r_1L_1} - \frac{1}{L_2}\right)T_L + T_R\right] \tag{18b}$$

In this expression K is the gain modulation factor as defined in section II 1. The output of the phase sensitive detector will be proportional to

$$T_I - T_{II} = \left[1 - \frac{K}{r_1}\right]\frac{T_1}{L_1} - \left(K - \frac{1}{r_2}\right)\frac{T_2}{L_2} + \left[\left(K - \frac{1}{r_2}\right)\frac{1}{L_2} - \left(1 - \frac{K}{r_1}\right)\frac{1}{L_1}\right]T_L$$
$$+ (1 - K)T_L + (1 - K)T_R . \tag{19}$$

From (18a) and (19) we obtain the total effective receiver temperature referred to input 1.

$$T_{tot} = \left(\frac{r_1}{r_1 - K}\right)\left[T_1 + \frac{L_1}{r_2 L_2} T_2 + \left(L_1 - 1 - \frac{L_1}{r_2 L_2}\right) T_L + T_R L_1\right].$$

For a receiver with a symmetric switch and balanced by the addition of noise to one input channel, we have $r_1 = r_2 = r$, $L_1 = L_2 = L$, $K = 1$, $T_1 = T_2 + \Delta T$, giving

$$T_{tot} = \left(\frac{r}{r - 1}\right)\left[\left(1 + \frac{1}{r}\right) T_2 + \Delta T + \left(L - 1 - \frac{1}{r}\right) T_L + T_R L\right].$$

From this it can be seen that a switch with a switching ratio, r, of 10 db will give an increase of from 12 to 15 per cent in effective receiver temperature over a perfect switch, depending upon the value of T_R. For a receiver using gain modulator balancing and an ambient temperature load, we have $T_2 = T_L = T_o$ giving

$$T_{tot} = \left(\frac{r}{r - K}\right)\left[T_1 + (L - 1)T_o + LT_R\right].$$

In this case the increase for a 10 db switching ratio is about 11 per cent.

For an insertion loss L = .4 db = 1.1, the increase is 10 per cent + 30°, corresponding to as much as a 20 per cent increase in the noise temperature of a low noise receiver. For a passive power dividing network, $1 + \frac{1}{r} \leq L$. Therefore, an insertion loss of less than .4 db insures a switching ratio of greater than 10 db. Thus, the important parameter for the front-end switch is the insertion loss.

The front-end switch used in the radiometer described in this report is a diode switch of conventional design. As shown in Figure 6, it uses 1N270 diodes as shunt switching elements at the end of quarter wavelength lines. With a positive drive voltage, diode A is back biased. Its high resistance is transformed to a low value at point A by the quarter wavelength line L_A. This low shunt impedance reflects power entering at input 1, but is transformed to a high impedance at C and so does not affect the power entering at input 1. Diode B is forward biased, appearing as a high shunt impedance at point B. With a negative drive voltage the situation is reversed. The exact lengths of the diode lines are adjusted for maximum isolation, cancelling the shunt capacitance of the back biased diodes. The stubs S_A and S_B are adjusted for minimum insertion loss, cancelling the resulting susceptance due to the forward biased diodes. All lines are 50 ohms impedance, except for the diode lines which are made with higher impedance to increase the effectiveness of the forward biased diodes.

The shunt stubs S_A and S_B may be omitted with only a slight increase in insertion loss if an external return exists for the drive current. The 300 ohm resistors in the drive circuit provide a back bias of 20 v. for a forward current of 60 ma. Such switches have been built with quite satisfactory results for several frequencies from 600 mcs to 1420 mcs. The measured performance for a 750 mcs switch is shown in Figure 7. A problem of considerable importance for very low noise receivers is that the conducting diodes may generate as much as 100 to 200 °K of excess noise, as referred to the output of the switch.

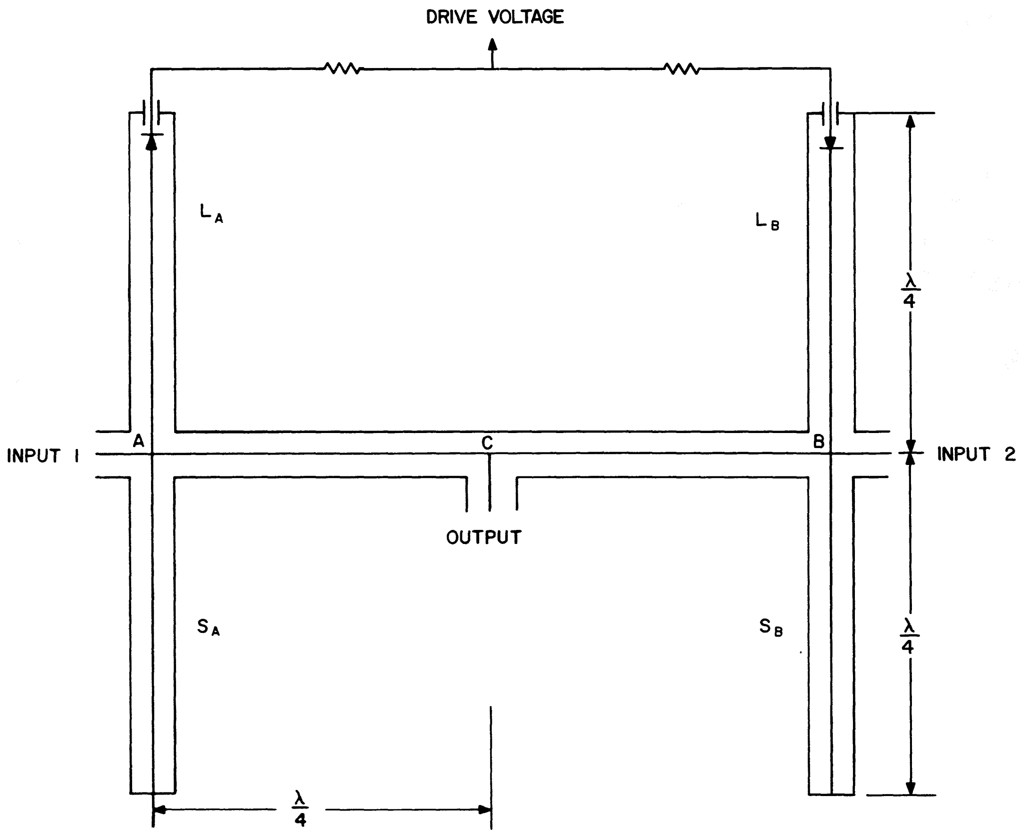

Fig. 6. — Schematic diagram of the front-end switch.

3. Discussion of the Gain Modulator.

In section II we introduced and discussed the gain modulation technique as a way of having the radiometer working under balanced conditions. As we have seen earlier, the gain modulation does not have to be introduced at the front-end, but may be introduced at a later stage, at least as long as the system is linear in power. One therefore chooses to make the modulation of the gain at the IF-frequency, and in Figure 8 is shown a scheme which has been used by NRAO. The IF-signal (30 mc) is fed to a tuned IF-stage (6136), and the plate circuit of this stage is loaded by two identical attenuator

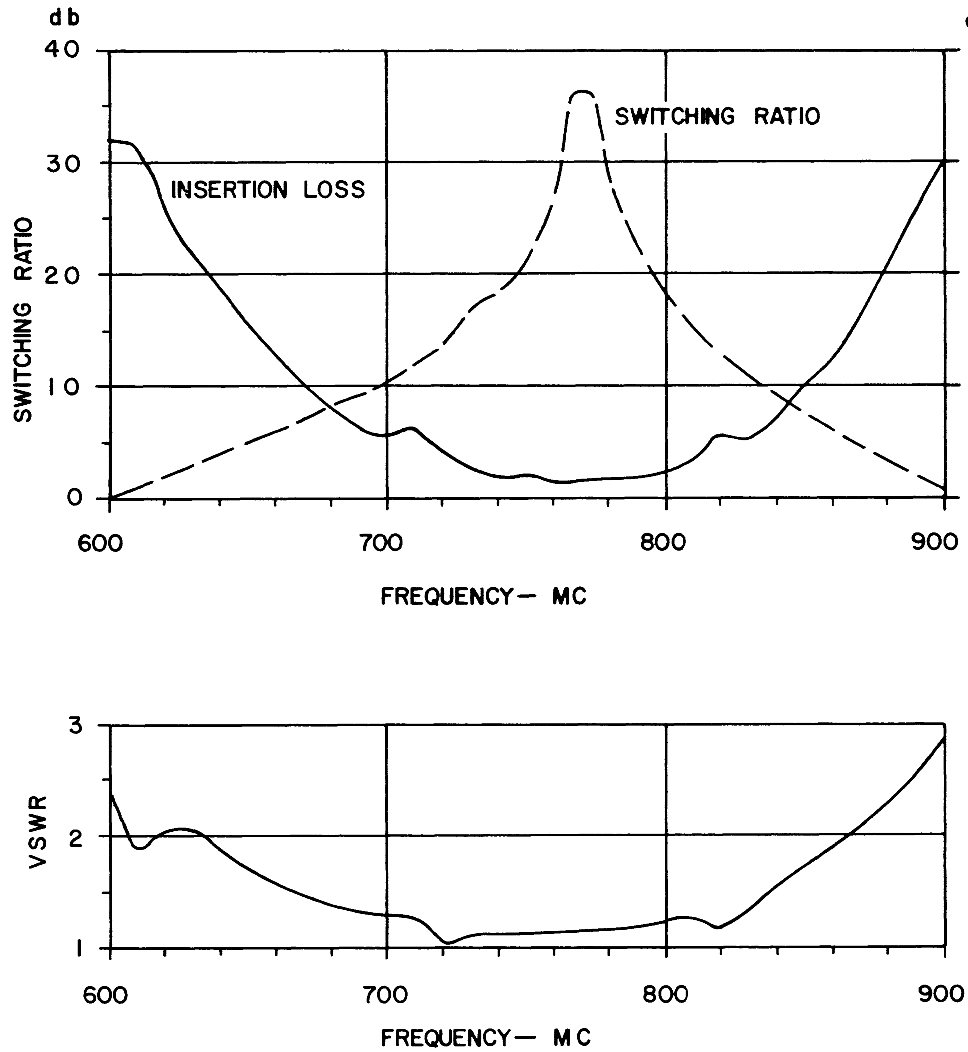

Fig. 7. — Measured characteristics for a 750 mc switch.

circuits. The attenuator chains consist of 11 series connected 12 ohm resistors in series with a variable 37 ohm resistor. By means of a diode switch at the input and at the output of the attenuator circuits, the IF-signal can be coupled to either of the two attenuators. The attenuation can be changed continuously and therefore all values of relative gain can be obtained. By switching the attenuators in synchronism with the front-

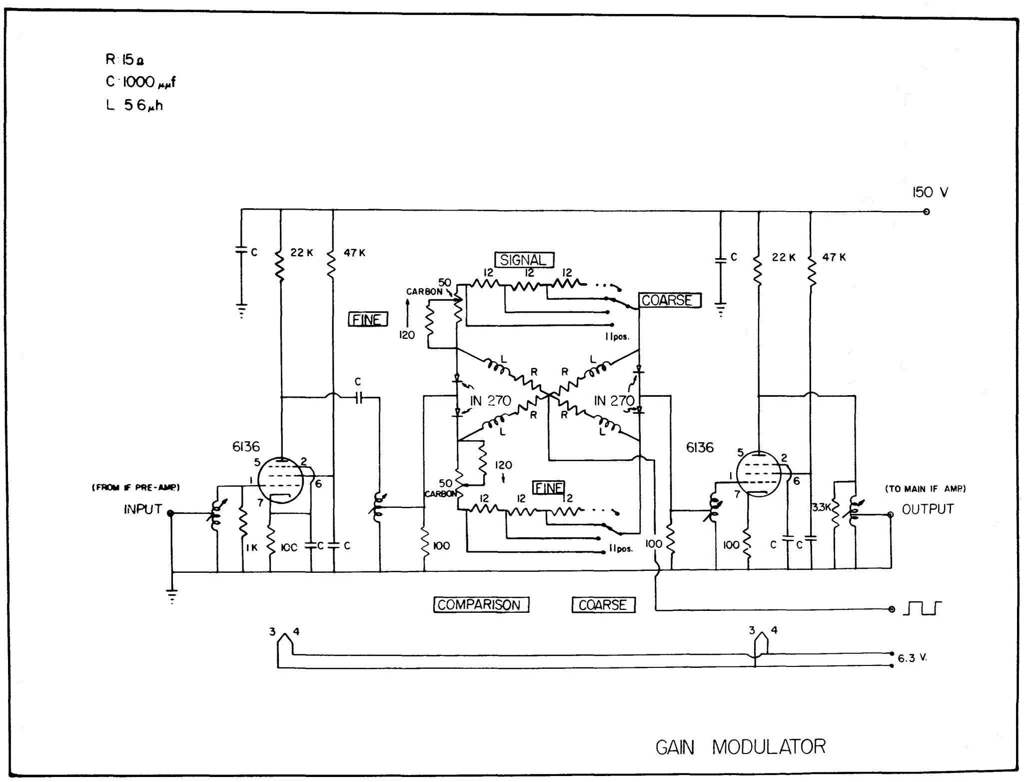

Fig. 8.— Schematic diagram of the gain modulator.

end switch, the signal from the comparison channel can be attenuated relative to that from the signal channel, and in this way a balanced condition according to Equation (7) can be produced. The gain of the circuit according to Figure 8 is approximately 1, and the 1 db bandwidth is 10 mc. From a stability point of view it is of interest to know the variation of gain as a function of the drive current for the switching diodes. Such a measurement is shown in Figure 9, and it is seen that for a total switch current of more than 30 mA, the gain is rather insensitive to changes in drive current. Experience with the gain modulation technique has not shown any problem in keeping the gain unbalance $K = G_C/G_A$ constant over a considerable length of time (days). Even if the amplitude of the reference voltage varies, the amount of relative change in gain in the two positions is the same if the system is linear and therefore the gain unbalance is not affected. A change in the gain unbalance K would, of course, have the most serious effect upon the total stability, as can clearly be seen from the fifth term of Equation (5). Another important parameter affecting the system stability is the symmetry of the switching voltage. Variation in the symmetry directly changes the output of the phase detector and may be an important source of instability.

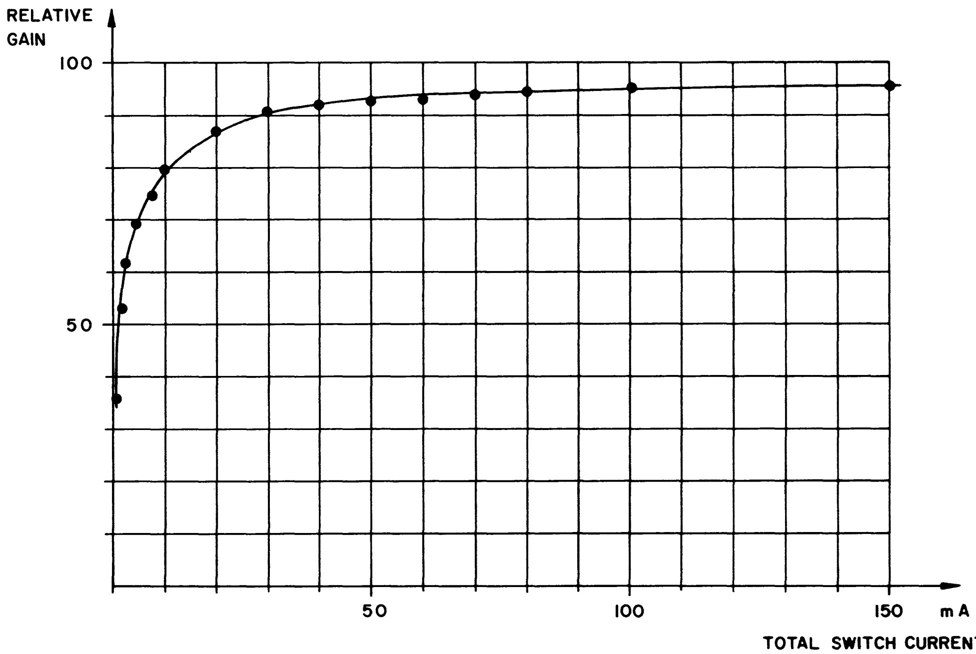

Fig. 9. — Variation of gain as a function of total drive current.

If the square wave is unsymmetrical and the on-off times have a ratio of φ, then the balance of the receiver is determined by

$$K_2 = \varphi K$$

where K is the previous gain unbalance and where we have assumed no audio filter. Therefore, a change in the symmetry affects the receiver performance in the same way as does a change in the gain unbalance K. It is, of course, possible to use this principle for producing a balanced condition, and one advantage of this approach, compared to the present one discussed, would be that no additional components are needed in the signal path; the balance would be produced at the switch itself.

The gain modulation principles are believed to be a very easy and convenient way of balancing a radiometer. Because the modulation of gain can be done at the IF-level, the gain modulator can physically be located in the back-end part of the receiver, and consequently there is no need for a remote control. The principle can best be used where the amount of unbalance is relatively small as was pointed out in section III.

The gain modulation principle has also the advantage that the system noise temperature is not increased by adding noise to the system; actually an artificial cold load is created. Therefore, with no loss in sensitivity, the comparison channel may have a noise temperature higher than the signal channel.

4. Discussion of Receiver Performance.

In section II the variations in output signal caused by different sources of instability in gain, receiver temperature, etc., were discussed. The output noise fluctuations have been treated by various authors (Goldstein (1957), Galejs (1957), Graham (1958)), and in the case of a total power receiver the relative sensitivity (output rms noise fluctuations equalized to variation in input temperature) can directly be derived using the power spectrum of the post-detection noise signal as found by Rice (1944), giving

$$\Delta T = T_{tot}\sqrt{\frac{2b}{B}} \tag{21}$$

where b and B are the post-detection and predetection equivalent noise bandwidths, T_{tot} is the total receiver system temperature ($T_R + T_A$). It should be noticed that both the linear and the quadratic detector give the same sensitivity for small signals. Usually, a simple RC network is used as an integration network, and the corresponding power spectrum is then

$$G(\omega) = \frac{1}{1 + \omega^2\tau^2}$$

where the integration time is $\tau = RC$. The corresponding noise bandwidth is obtained by integrating the power spectrum, giving

$$b = \frac{1}{4\tau}$$

and the sensitivity of the total power receiver is therefore in terms of predetection bandwidth and post-detection integration time

$$\Delta T = \frac{T_{tot}}{\sqrt{2B\tau}} . \tag{22}$$

The switched receiver has a sensitivity which is dependent upon the type of modulation and the type of filtering of the switched component. For square wave modulation and a narrow band filter centered at the switch frequency, the sensitivity has been shown to be

$$\Delta T = \frac{8}{\pi}\, T_{tot}\sqrt{\frac{b}{B}} \tag{23}$$

or in terms of time constant for an output RC filter

$$\Delta T = \frac{4}{\pi} \frac{T_{tot}}{\sqrt{B\tau}} . \tag{24}$$

The NRAO uses a digital output system, consisting of a voltage to frequency converter, a frequency counter, a scanner coupler, and an output punch. The frequency counter actually counts the instantaneous frequencies (proportional to instantaneous DC-voltage) during a specified time (integration time) after which it is reset to zero, and therefore acts as a finite memory type of integrator. The power spectrum for this type of low-pass filter can be shown to be

$$G(\omega) = \left[\frac{\sin \frac{\omega\tau}{2}}{\frac{\omega\tau}{2}}\right]^2$$

and the corresponding noise bandwidth is

$$b = \frac{1}{2\tau} .$$

The switched receiver, when using the finite memory type of integrator, therefore, has the sensitivity

$$\Delta T = \frac{8}{\pi\sqrt{2}} \frac{T_{tot}}{\sqrt{B\tau}} . \tag{25}$$

If we only take into account output fluctuations caused by noise and by gain instability, the total fluctuations for a switched receiver are

$$\sqrt{\overline{(\Delta T_A)^2}} = T_{tot} \left\{ \frac{\gamma^2}{B\tau} + \overline{\left(\frac{\Delta G}{G}\right)^2} \left[\frac{\Delta T_{unbal.}}{T_{tot}}\right]^2 \right\}^{1/2}$$

where γ is a constant dependent upon receiver characteristics. If we assume the unbalance to be η times the rms fluctuations, we have

$$\sqrt{\overline{(\Delta T_A)^2}} = \frac{\gamma}{\sqrt{B\tau}} T_{tot} \left[1 + \overline{\left(\frac{\Delta G}{G}\right)^2} \left(\frac{\eta}{T_{tot}}\right)^2\right]^{1/2} . \tag{26}$$

In Figure 10 are shown records taken with a 750 mc radiometer, and both total power and switched records are shown. The two records clearly show the increased stability of the switched receiver and at the same time also show the decreased sensitivity as regard to noise fluctuations.

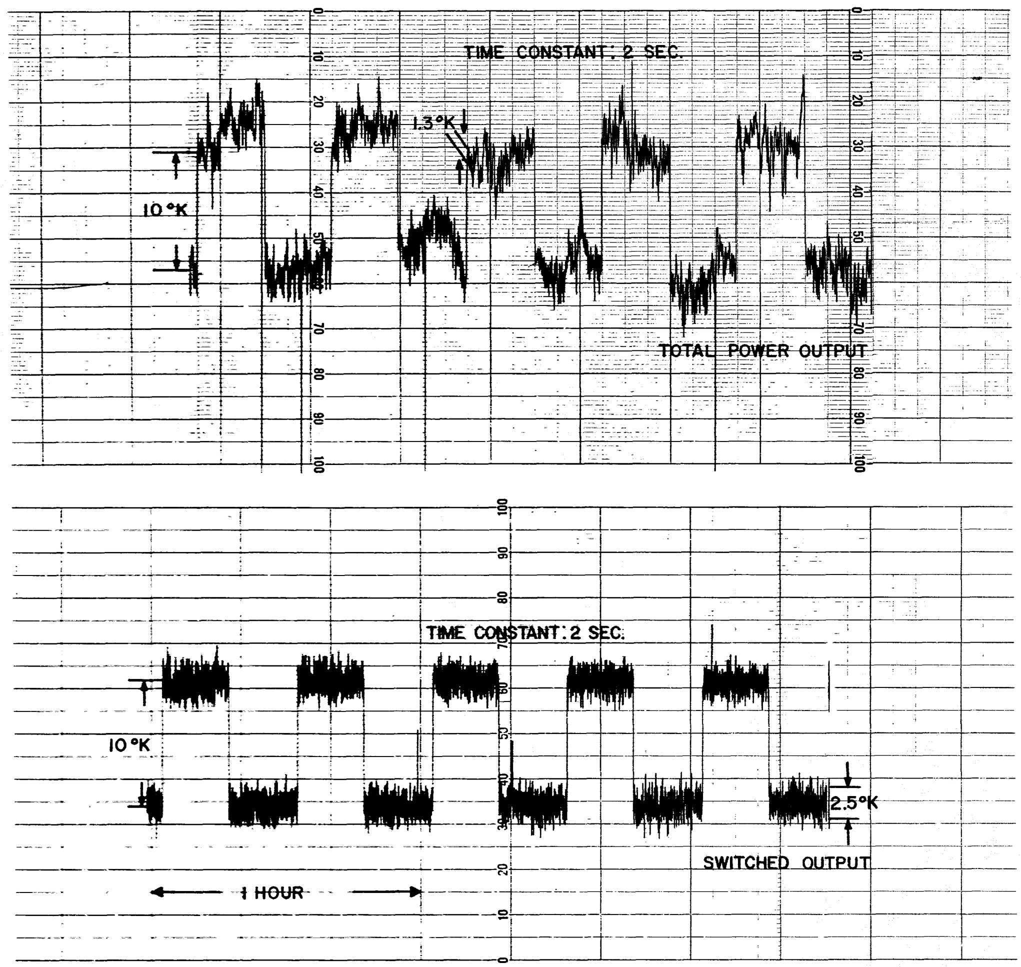

Fig. 10. — Test records taken with a 750 mc radiometer having a total noise temperature of 1400 °K (including 300 °K termination). The upper record is a total power record, taken with the switch locked in one position. The noise fluctuations are almost entirely hidden by short and long term instabilities. The lower record is a switched record. Calibration signals of 10 °K are shown on both records.

In order to test the performance of the switched receiver for variation in system gain for different values of output unbalance, Figure 11 shows a record taken with different output unbalances. For each value of gain unbalance, a 3 db change in gain was introduced by means of the attenuator in the IF-channel.

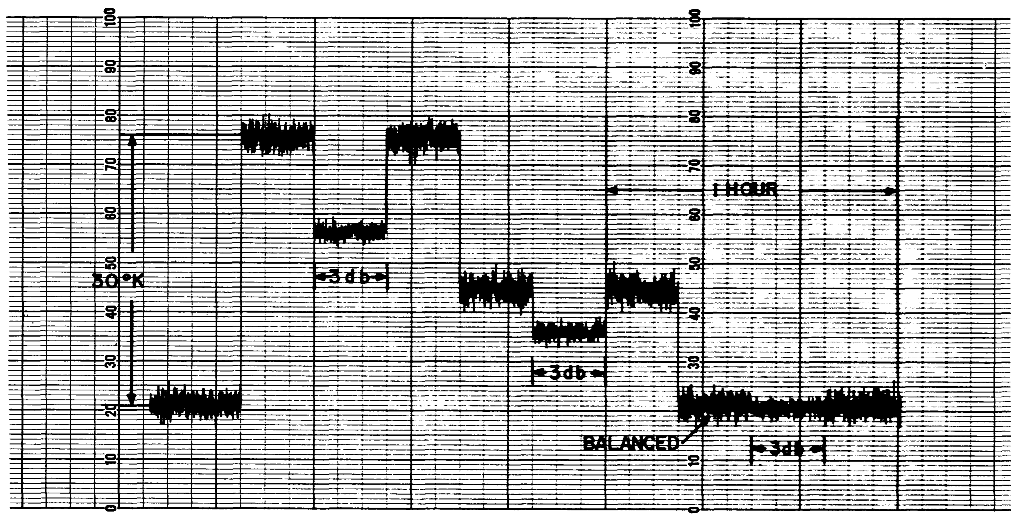

Fig. 11.— Response of a switched load receiver to a 3 db change in IF gain in a balanced condition and for unbalances of 15 °K and 30 °K, τ = 2 sec. T_R = 1400 °K.

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