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THE NRAO 50-CHANNEL SPECTRAL LINE RECEIVER

Michael Balister

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THE NRAO 50-CHANNEL SPECTRAL LINE RECEIVER

M. Balister

1. Introduction

This report describes a 50-channel receiver with individual channel bandwidths of 100 kHz which was developed at the National Radio Astronomy Observatory for spectral line work. The receiver was first used for 6 cm excited hydrogen line work on the 140-foot telescope during September 1967.

The filter bank may be used as a set of 50 contiguous filters or alternatively it may be split into two sets of 25 filters for use with separate RF front ends. Integration of an on-line computer into the system allows measurement and correction for system gain and zero drifts and provides an immediate calibrated display on a memory oscilloscope. The data may be also printed out or recorded on magnetic tape for further analysis.

This receiver may be used with most of the NRAO front-ends and either frequency or load switching may be employed.

2. Description of the Filter Receiver

2.1 General

A block diagram of the receiver is shown in Figure 1. The filters lie in the frequency range of 3-8 MHz; each consists of a double tuned LC circuit which has a 3 dB bandwidth of 100 kHz.

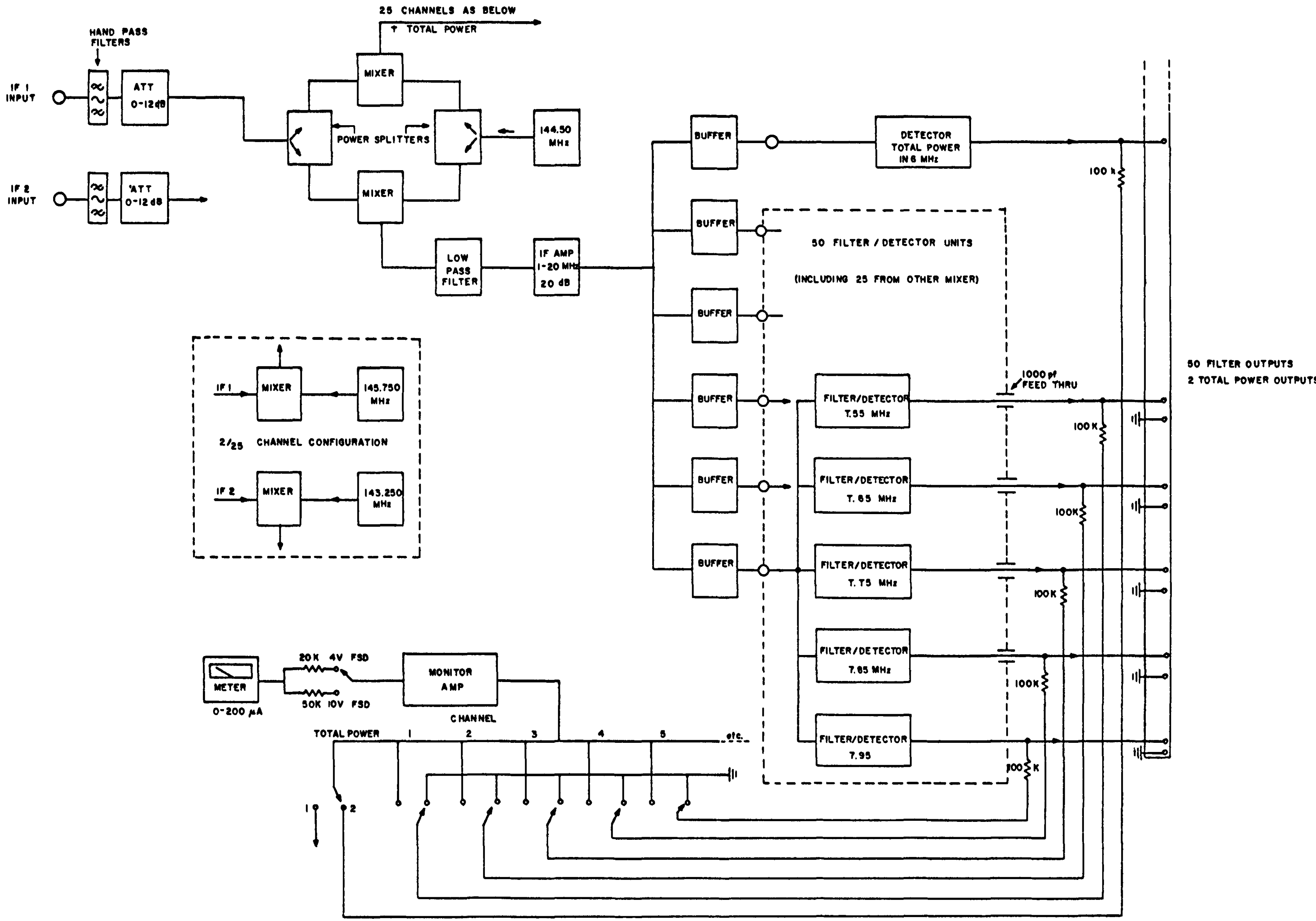
The receiver is normally used as a contiguous comb of filters covering an IF frequency range of 147.5 to 152.5 MHz. The IF center frequency of 150 MHz is converted to 5.5 MHz by use of a 144.5 MHz local oscillator. Figure 2 lists the filter center frequency for each channel before and after conversion.

In the alternative mode the filters are split into two and each set covers an IF frequency range of 148.75 to 151.25 MHz. The local oscillator frequency is 143.25 MHz for channels 1-25 and 145.75 MHz to channels 25-50. Figure 2 lists the IF center frequency of channels 1-50 for this alternate configuration.

The filter receiver is constructed so that it can be installed in a standard 19" rack; the height of the front panel is 7". A photograph of the completed receiver is shown in Figure 3.

2.2 Mixer and Power Splitting Section

The mixer and power splitting section consists of two identical circuit boards; the circuit is shown in Figure 4. When the receiver is used in the 50-channel mode the two inputs are fed via a power splitter from a common IF input. The local oscillators are derived via a power splitter from a single 144.5 MHz oscillator.



BLOCK DIAGRAM OF 50 CHANNEL RECEIVER (100 KHz)

FIGURE 1

D.M.H.
5/14/68

Channel No.	Center Frequency of Filter (MHz)	IF Center Frequency	
		50-Channel (MHz)	2/25 Channel (MHz)
1	3.05	147.55	148.8
2	3.15	147.65	148.9
3	3.25	.75	149.0
4	3.35	.85	.1
5	3.45	.95	.2
6	3.55	148.05	.3
7	3.65	.15	.4
8	3.75	.25	.5
9	3.85	.35	.6
10	3.95	.45	.7
11	4.05	.55	.8
12	4.15	.65	.9
13	4.25	.75	150.0
14	4.35	.85	.1
15	4.45	.95	.2
16	4.55	149.05	.3
17	4.65	.15	.4
18	4.75	.25	.5
19	4.85	.35	.6
20	4.95	.45	.7
21	5.05	.55	.8
22	5.15	.65	.9
23	5.25	.75	151.0
24	5.35	.85	.1
25	5.45	149.95	.2
26	5.55	150.05	148.8
27	5.65	.15	.9
28	5.75	.25	149.0
29	5.85	.35	.1
30	5.95	.45	.2
31	6.05	.55	.3
32	6.15	.65	.4
33	6.25	.75	.5
34	6.35	.85	.6
35	6.45	.95	.7
36	6.55	151.05	.8
37	6.65	.15	.9
38	6.75	.25	150.0
39	6.85	.35	.1
40	6.95	.45	.2
41	7.05	.55	.3
42	7.15	.65	.4
43	7.25	.75	.5
44	7.35	.85	.6
45	7.45	.95	.7
46	7.55	152.05	.8
47	7.65	.15	.9
48	7.75	.25	151.0
49	7.85	.35	.1
50	7.95	.45	.2

Figure 2 — Filter and IF Frequencies for Both Configurations

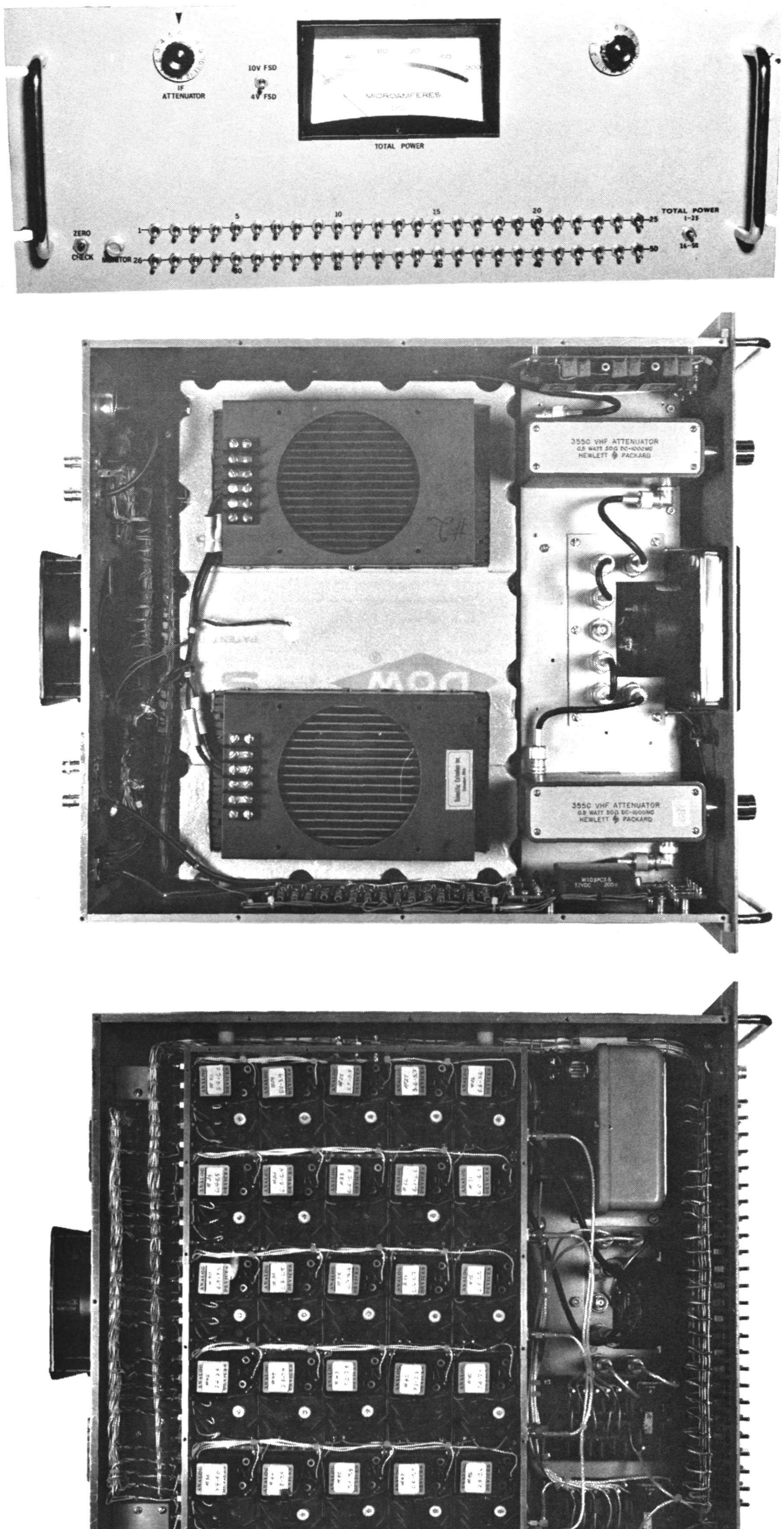
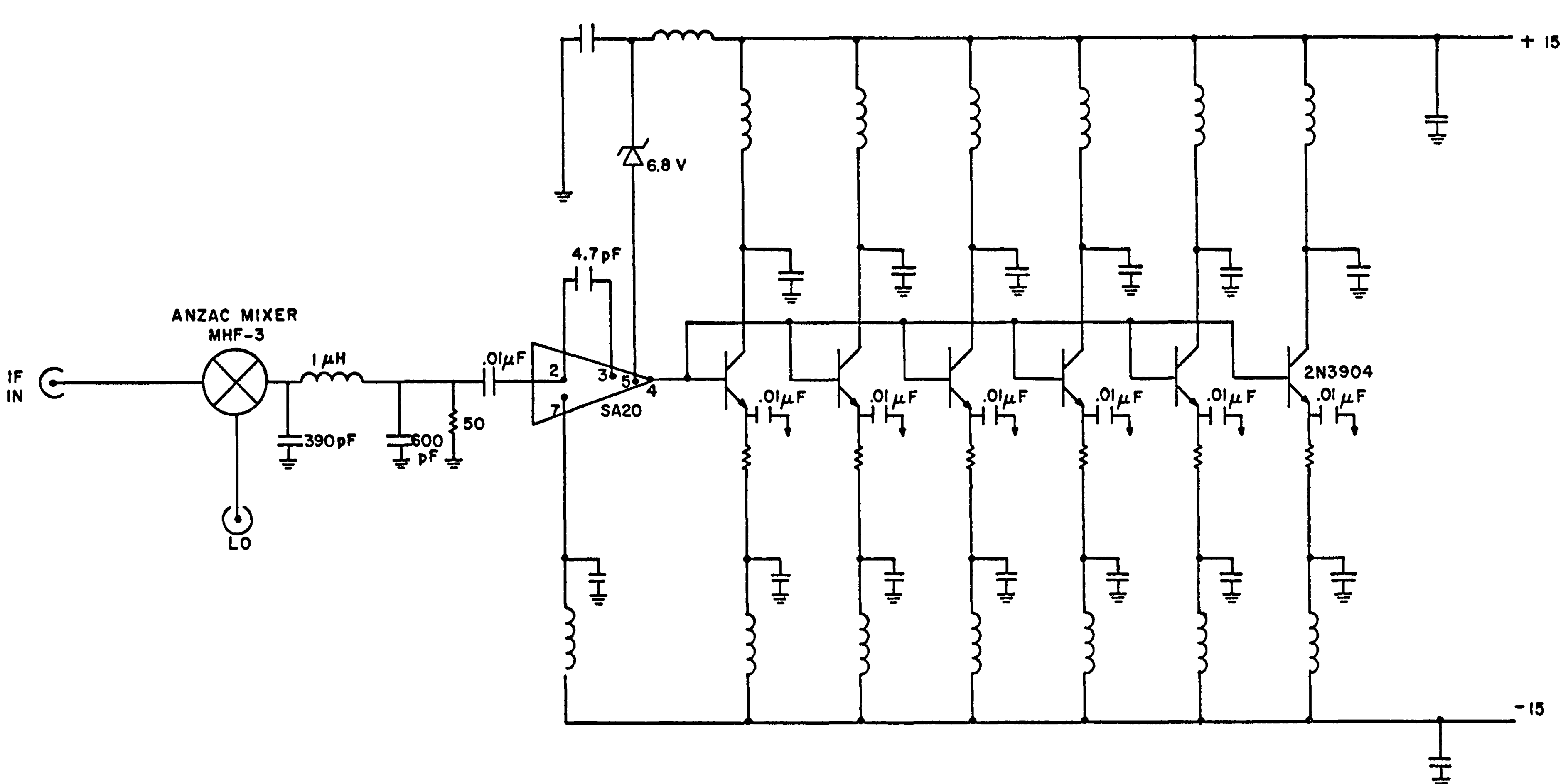


Figure 3 — Three Views of the Receiver



THIS CIRCUIT DRIVES 5X5 CHANNELS + 1 TOTAL POWER CHANNEL

UNMARKED CAPACITORS —— 0.22 μF

UNMARKED CHOKES ———— 22 μH

IF POWER SPLITTER FOR 50 CHANNEL RECEIVER

FIGURE 4

In the alternative configuration the IF inputs are connected to separate RF front-ends; the local oscillator signals are derived from two oscillators (143.25, 145.75 MHz).

The IF level can be set to the optimum level by means of a common 0-12 dB attenuator in the 50-channel case, or by two separate 0-12 dB attenuators in the alternate case.

Each mixer/power splitter board has six buffered outputs; five outputs drive five filters (25 in all) and the sixth drives a detector which gives the wide band total power. The mixer is an Anzac model MHF-3 which has a minimum LO, RF to IF isolation of 25 dB. The mixer is followed by a low pass filter which has a 3 dB bandwidth of 11 MHz. A Sylvania SA20 integrated circuit RF amplifier follows with a gain of 20 dB; the output drives six emitter followers in parallel.

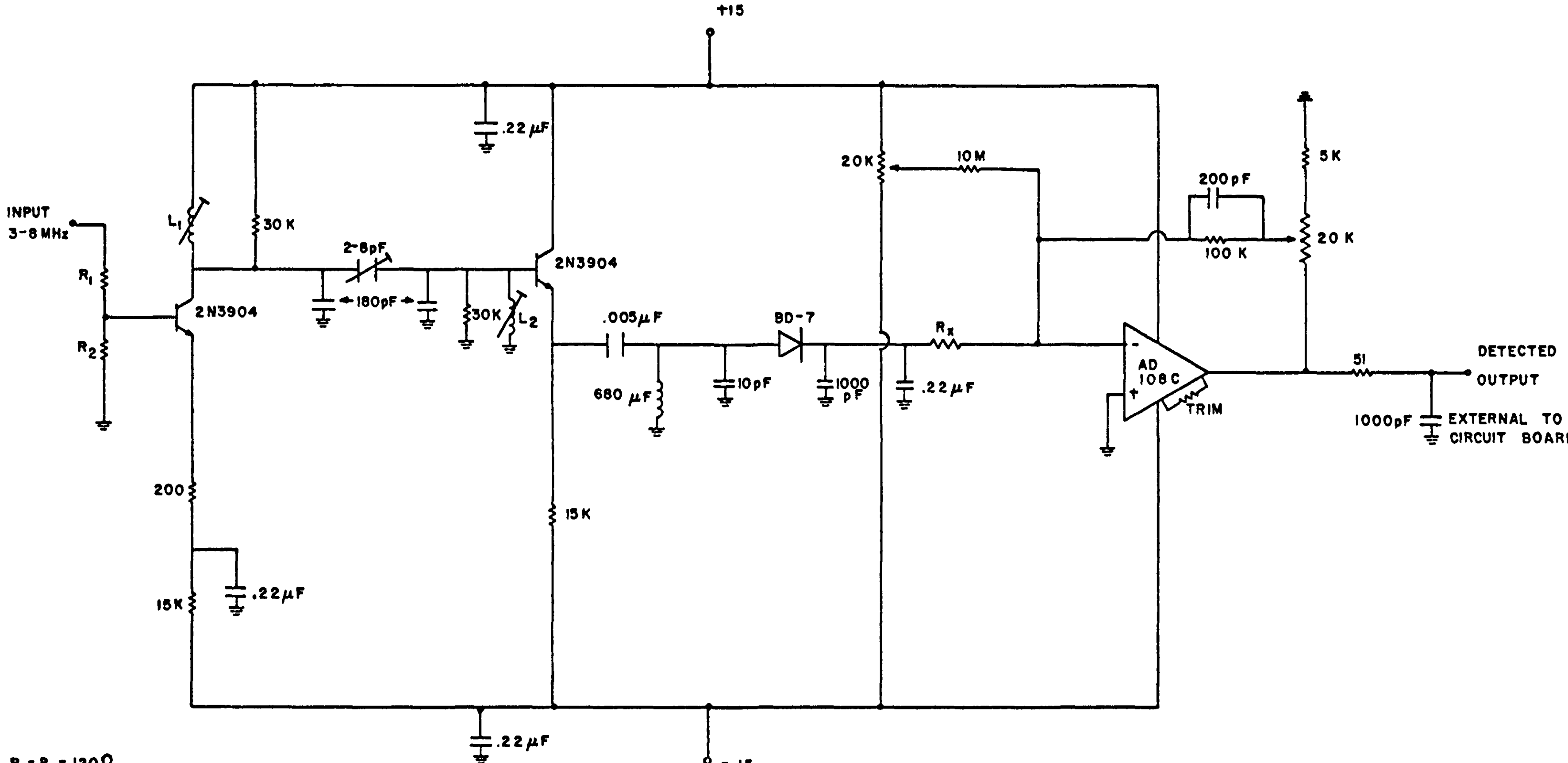
2.3 Filter Bank and Detector Unit

The circuit of a filter and detector unit is shown in Figure 5. These units are driven in groups of five from a single emitter follower; an input attenuator is included to give some isolation between channels. The resistor values in the input attenuator are changed for filters 21-50 in order to increase the gain of these channels to compensate for the decrease in gain of the filters with increasing frequency.

The filter consists of a double tuned circuit which was originally designed to be critically coupled giving a 1 dB bandwidth of 100 kHz. The receiver was subsequently tuned so that the 3 dB bandwidth was 100 kHz and consequently the coupling is less than critical. Figure 6 shows the bandpass characteristics of typical channels. The filters have an overall bandpass temperature coefficient of about 1 kHz/°C, largely as a result of using inductors with a ferrite core material. The moderate temperature coefficient was overcome by stabilizing the temperature of the box containing the filters to 25 ± 0.1 °C. Section 4 describes the temperature controller.

Several types of detectors were investigated to determine their suitability as square law detectors. The germanium back diode was finally selected because of its good square law performance, and low temperature coefficient (≈ 0.1%/°C). The BD7 diode has a maximum reverse peak point of 0.01 mA; less good back diodes (approaching tunnel diode) were tried with less success. With the BD7 diode the deviation from square law was generally less than ± 2% up to about -14 dBm. R_x was selected to give best high level performance. The value was generally in the range of 390-470 ohms. The normal operating point for the detectors was set to about -20 dBm which is equivalent to -37 dBm in 100 kHz at the filter input. A few typical detector responses are shown in Figure 7.

An Analog Devices 108C operational amplifier was used to amplify the detected signal. The 108C has an input drift of 5 μV/°C. The differential current drift may be ignored because of the low source resistance. The gain control (see Figure 5) was set to give a 2 V DC output with -20 dBm at the detector. The AC component, due to noise



R_x SELECTED FOR BEST SQ LAW CHARACTERISTICS IS 430Ω

FILTERS 1-21 $R_1 = R_2 = 120\Omega$

FILTERS 21-50 $R_1 = 62\Omega$ $R_2 = 180\Omega$

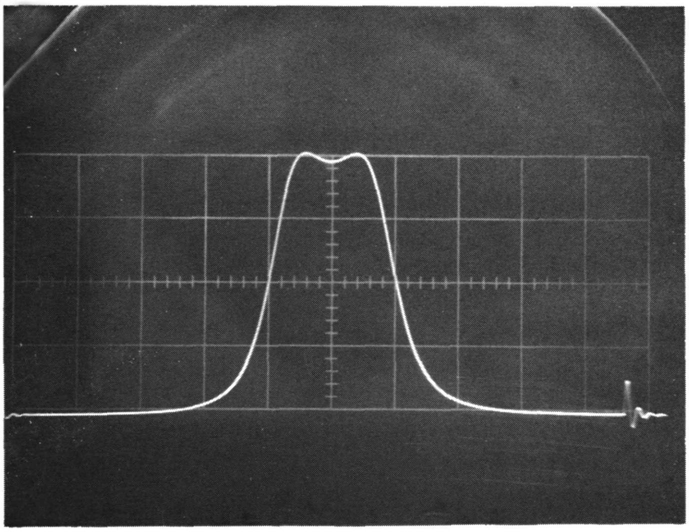
CHANNEL	$L_1 = L_2$	
1-7	15 μH	
8-17	10 "	
18-24	6.8 "	NYTRONICS
25-34	4.7 "	WEE V-L INDUCTOR
35-42	3.3 "	
43-50	2.2 "	

FILTER-DETECTOR UNIT OF 50 CHANNEL RECEIVER

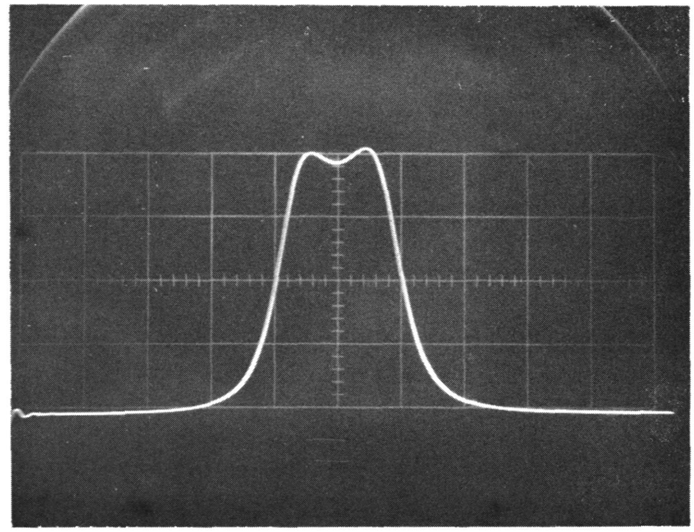
FIGURE 5

Horizontal Scale: 1 cm $\equiv$ 50 kHz

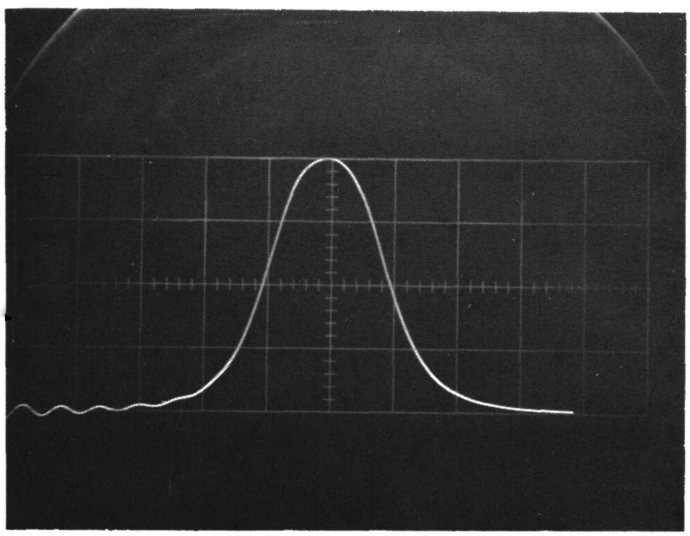
Channel 1



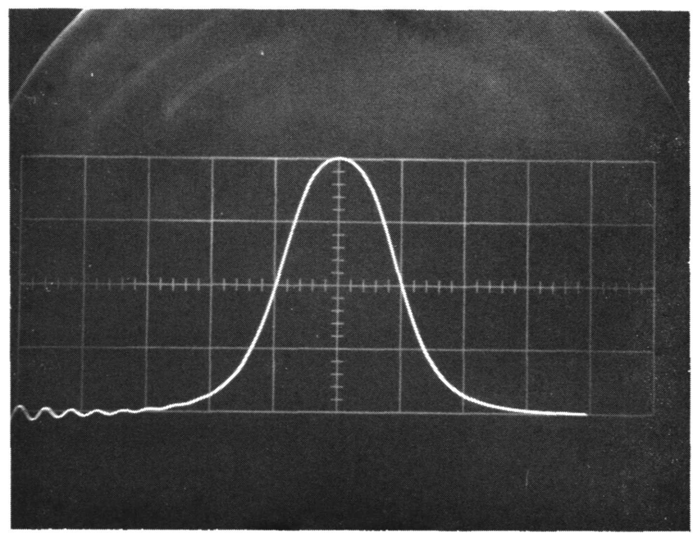
Channel 10



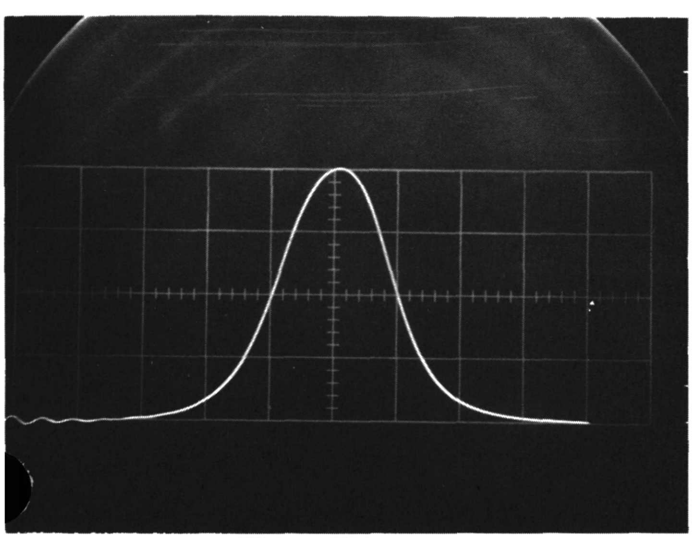
Channel 20



Channel 30



Channel 40



Channel 50

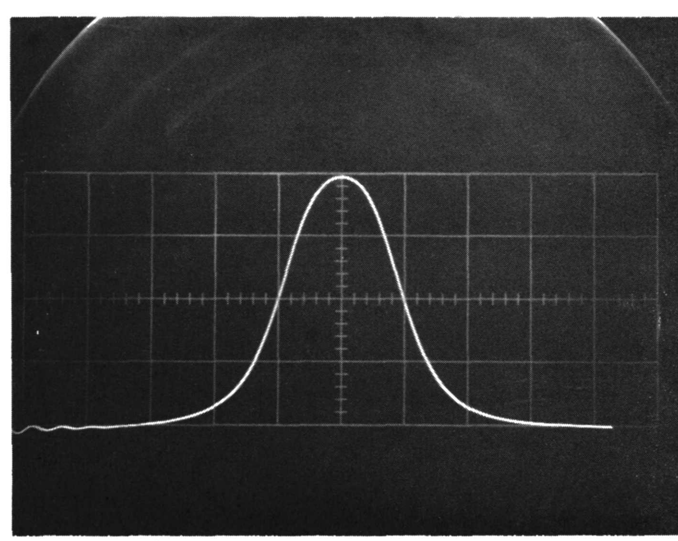


Figure 6 — Filter Frequency Responses

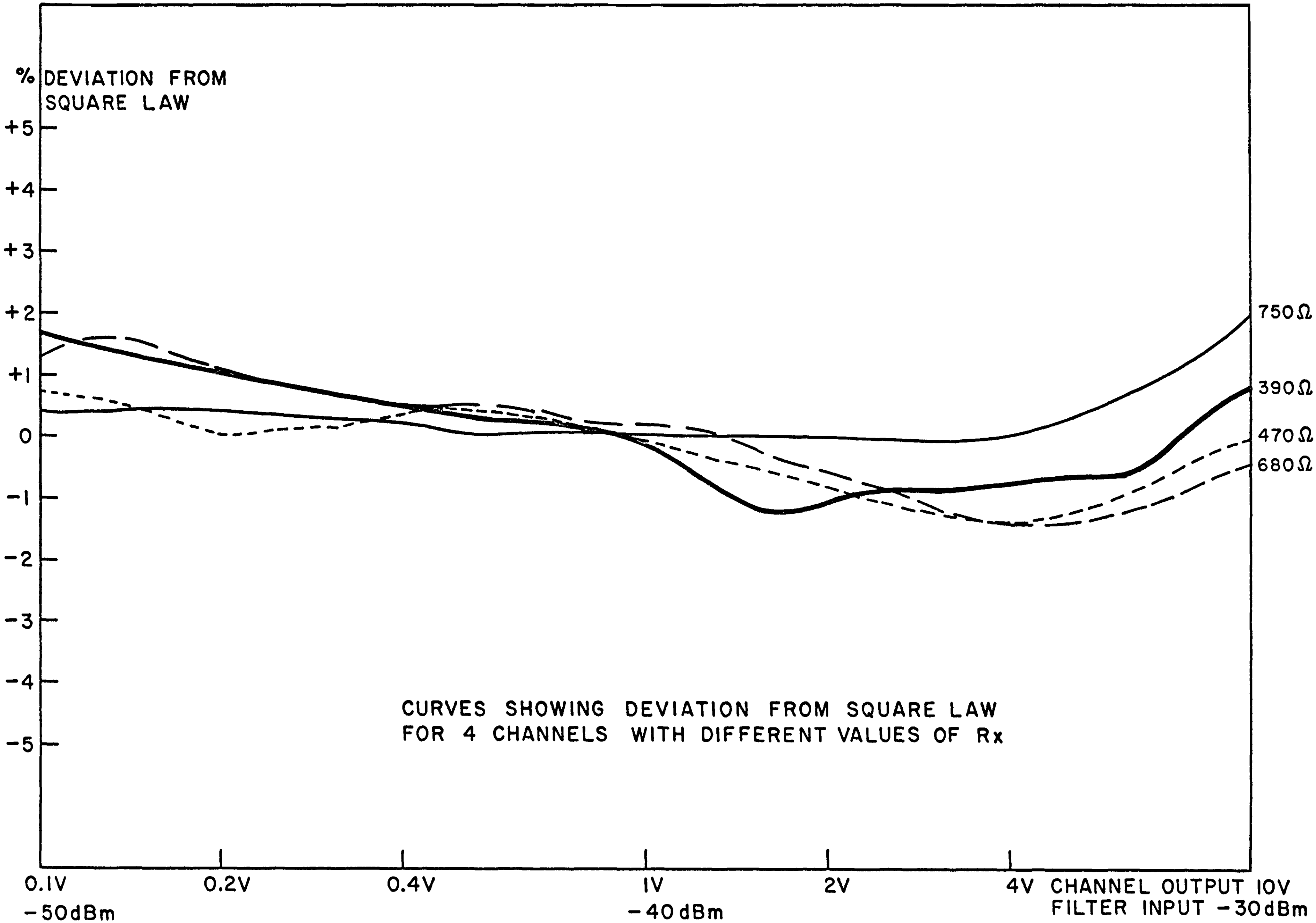


FIG. 7 TYPICAL DETECTOR RESPONSES

which is superimposed on this DC level, is approximately 0.5 V RMS. This corresponds to a video bandwidth of 3 kHz which was checked by measurement. The zero offset (the output voltage with no signal input) is adjusted to zero by the other potentiometer on the board. In the completed system the zero offsets were all within ± 20 mV. These residuals were removed by the computer program.

The unit is build on a 3" x 2" printed circuit board; a photograph is shown in Figure 8.

2.4 Total Power Detectors and Monitor Circuits

The sixth buffer in each of the mixer and power splitting sections drives a wide band detector; the circuit is shown in Figure 9. The total power channels need no extra RF gain since the IF bandwidth of 6 MHz compensates for the loss in filter gain of 20 dB. The front panel meter is driven by an operational amplifier (Figure 9) which may be connected by means of the front panel switches to any of the 50 channels and also the total power outputs.

2.5 Second Local Oscillators

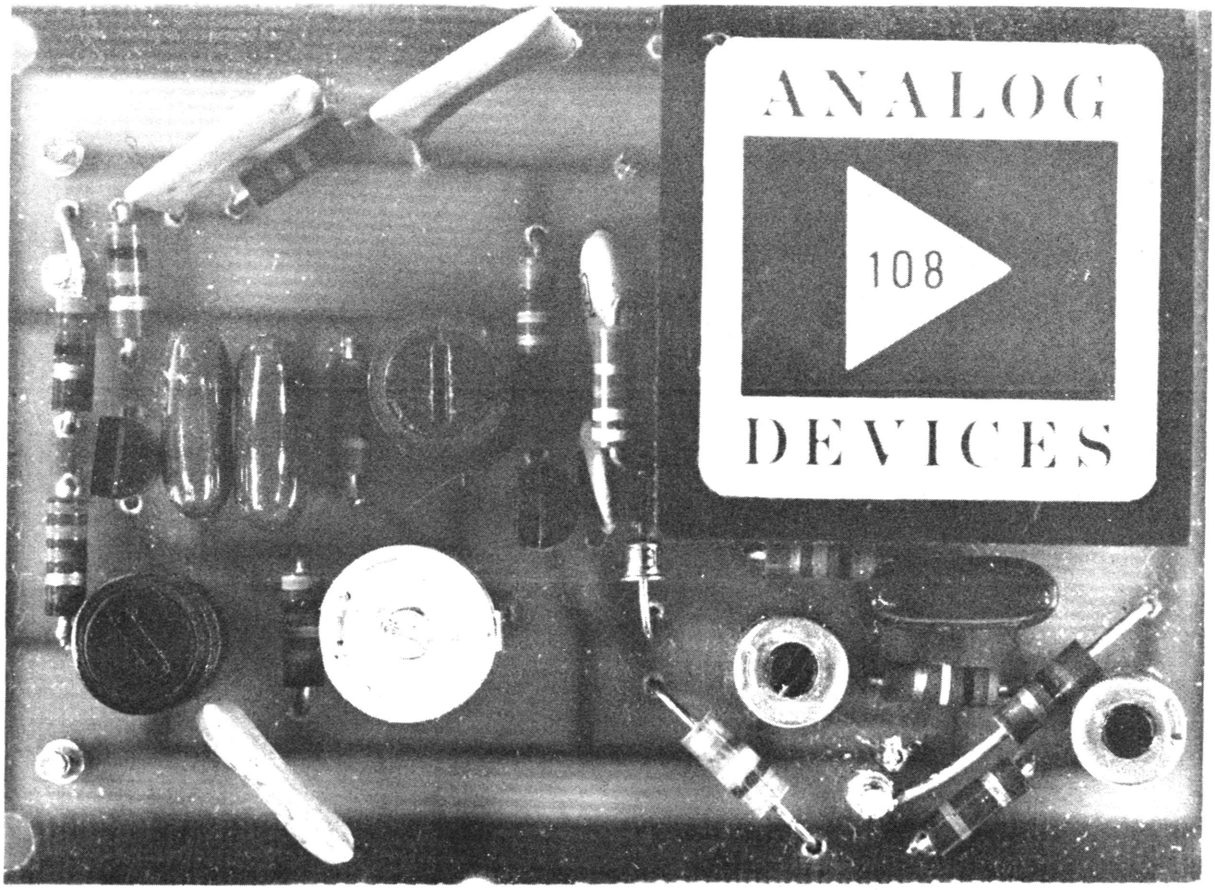
The circuit of the second local oscillator which converts the IF frequency of 150 MHz down to the 3-8 MHz range of the filters is shown in Figure 10. We call this the second local oscillator to avoid confusion with the first local oscillator which converts the line frequency to 150 MHz.

The circuit consists of three stages, oscillator, times three multiplier, and power amplifier. The output power is about 10 mW, and each oscillator can be tuned to within a few hundred cycles of the correct output frequency. The long-term stability is better than 1 kHz. The oscillators are not temperature compensated or stabilized since the individual filter bandwidth is 100 kHz.

2.6 Power Levels in the Receiver and Image Rejection

Figure 11 gives the approximate power levels of the signals at various stages in the line receiver. The variation in filter/detector gain can be corrected for by the gain adjustment on each board.

The line receiver has an image response which is relatively close to the IF frequency of 150 MHz because of the low frequency of the lower channels. The image band or bands depend on the particular mode in which the receiver is used; these are shown in Figure 12. In order to remove signals at the image frequency, two input filters are used. These have a 1 dB bandwidth of 6 MHz and 5 sections, resulting in a minimum image rejection of 25 dB which is adequate for most programs.



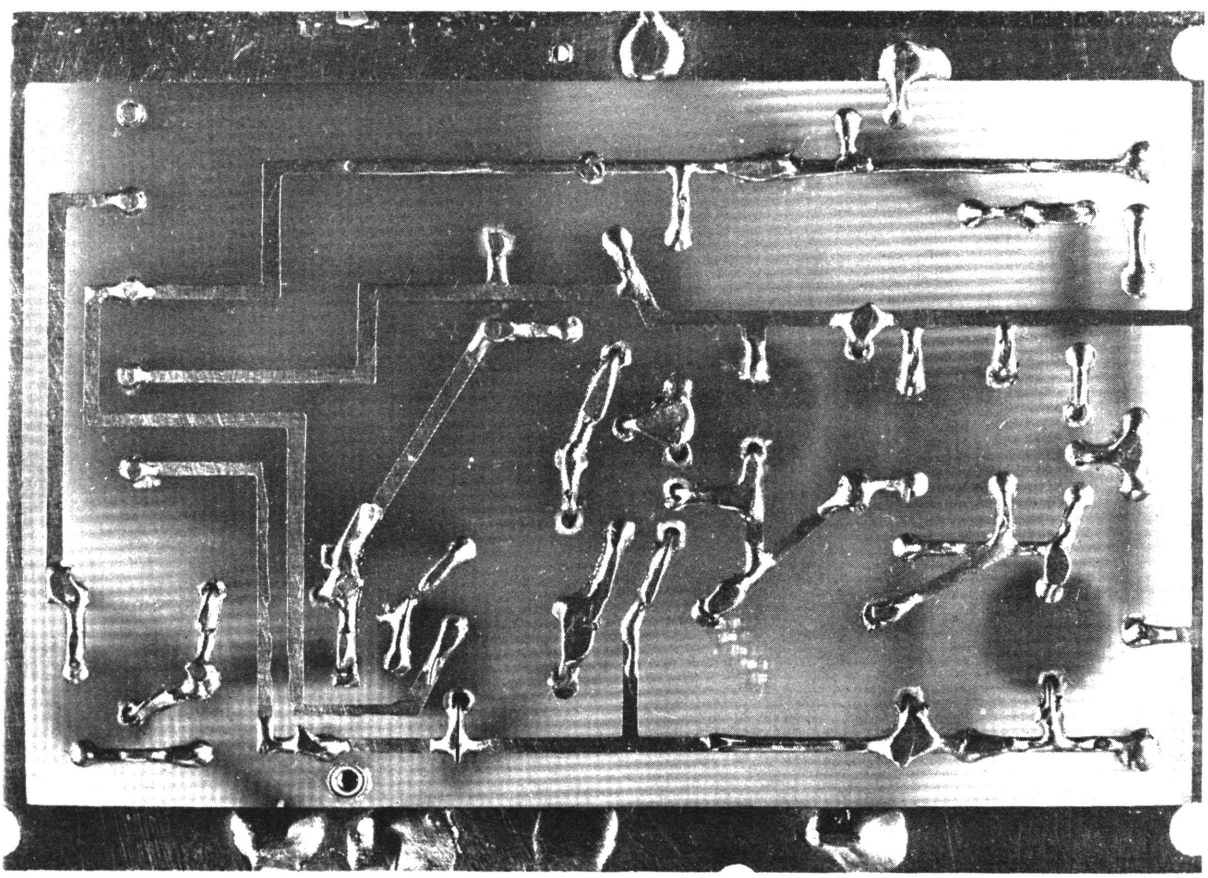
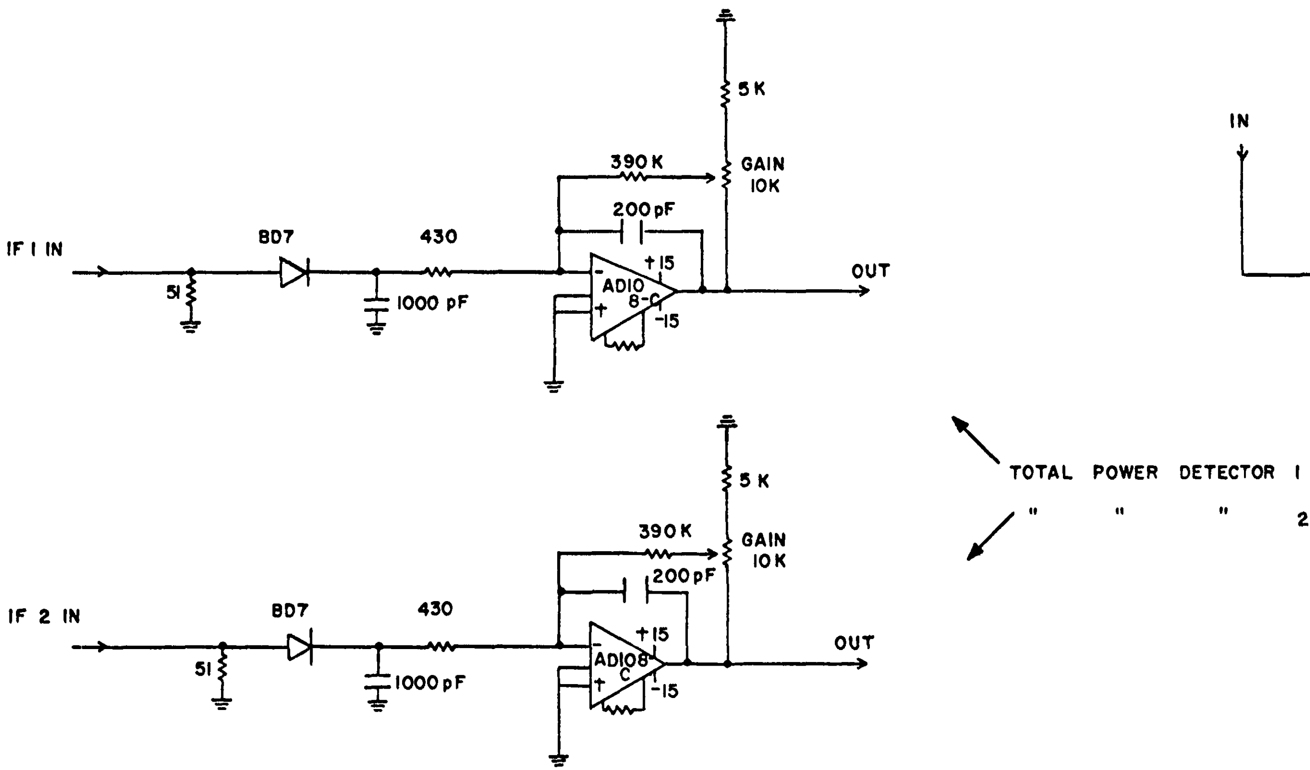
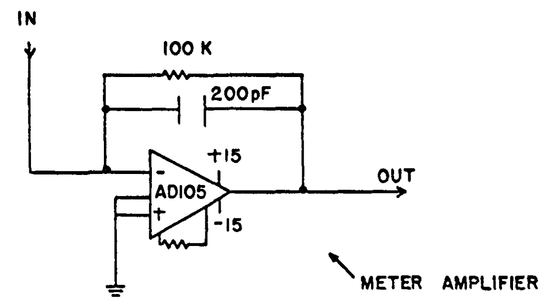
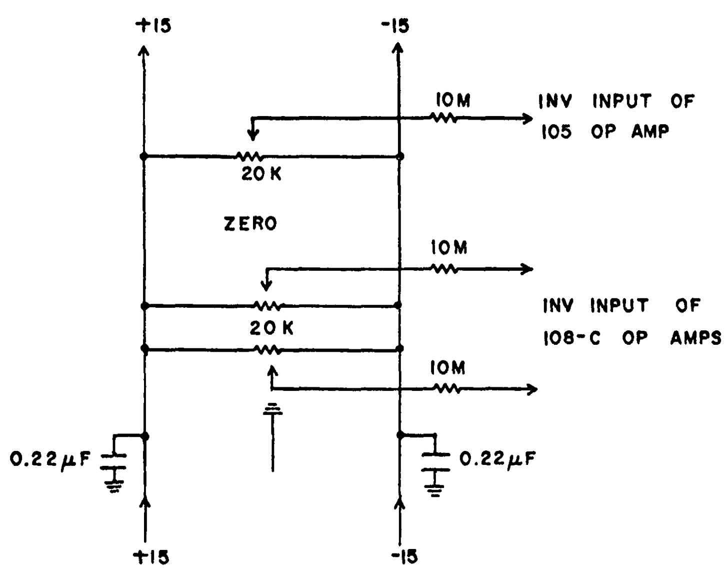


Figure 8 — Photograph of Filter-Detector Circuit Board

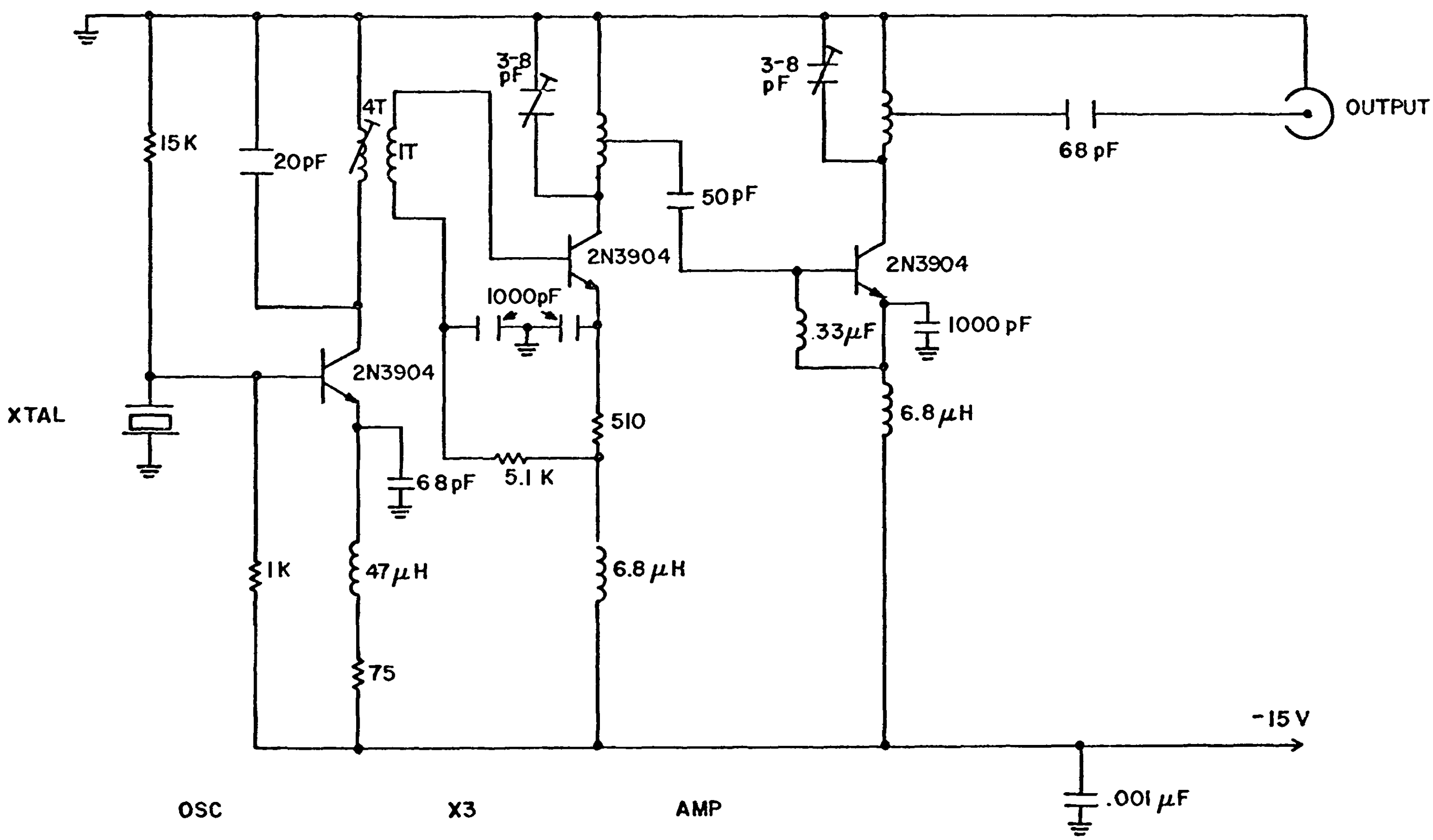






TOTAL POWER DETECTORS
AND MONITOR CIRCUIT

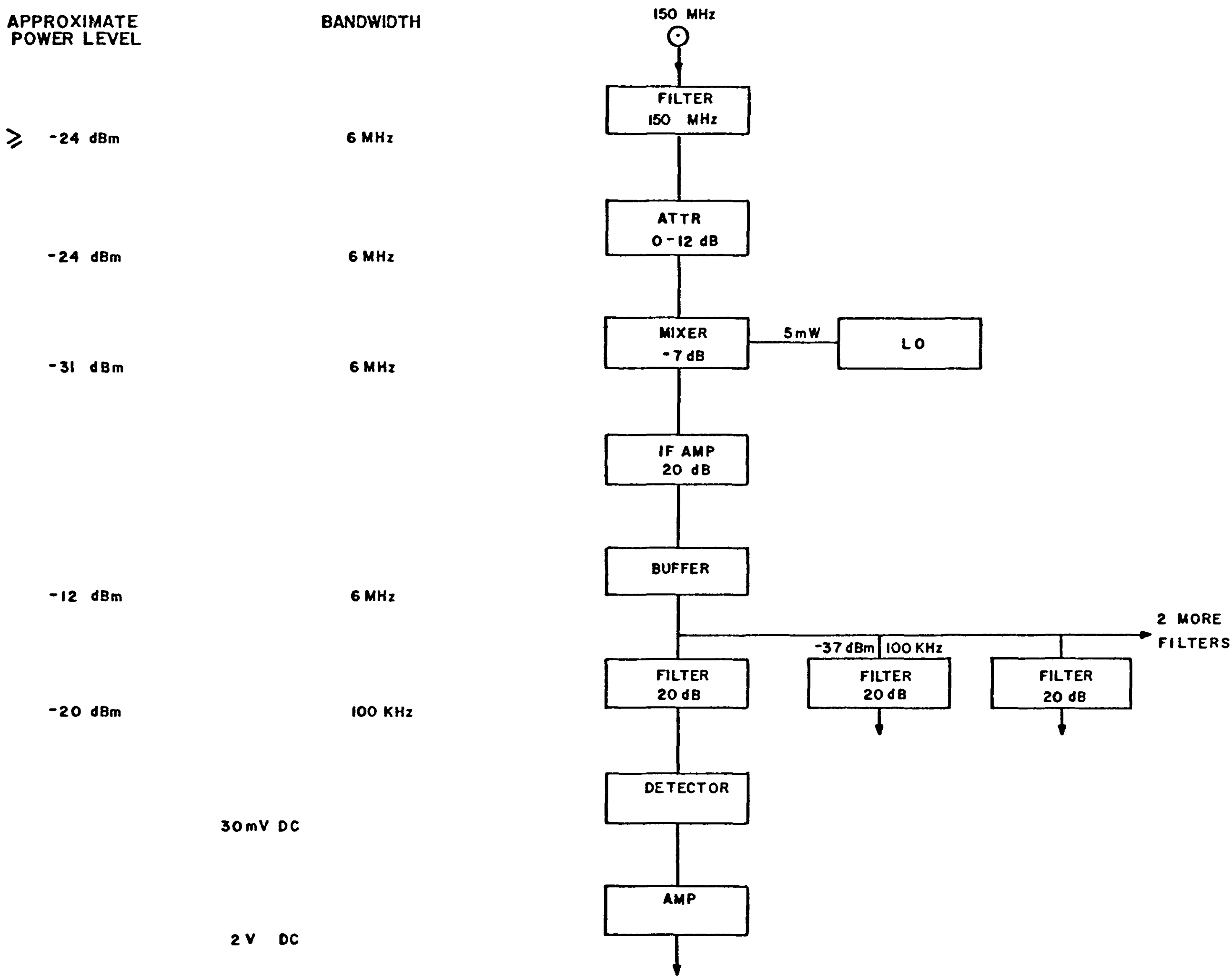
FIGURE 9



RECEIVER CONFIGURATION	XTAL FREQ MHz	OUTPUT MHz
2/25 CHANNEL	48.5833 47.7500	145.750 143.250
50 CHANNEL	48.1667	144.500

SECOND LOCAL OSCILLATOR

FIGURE 10



POWER LEVELS IN 50 CHANNEL RECEIVER

FIGURE 11

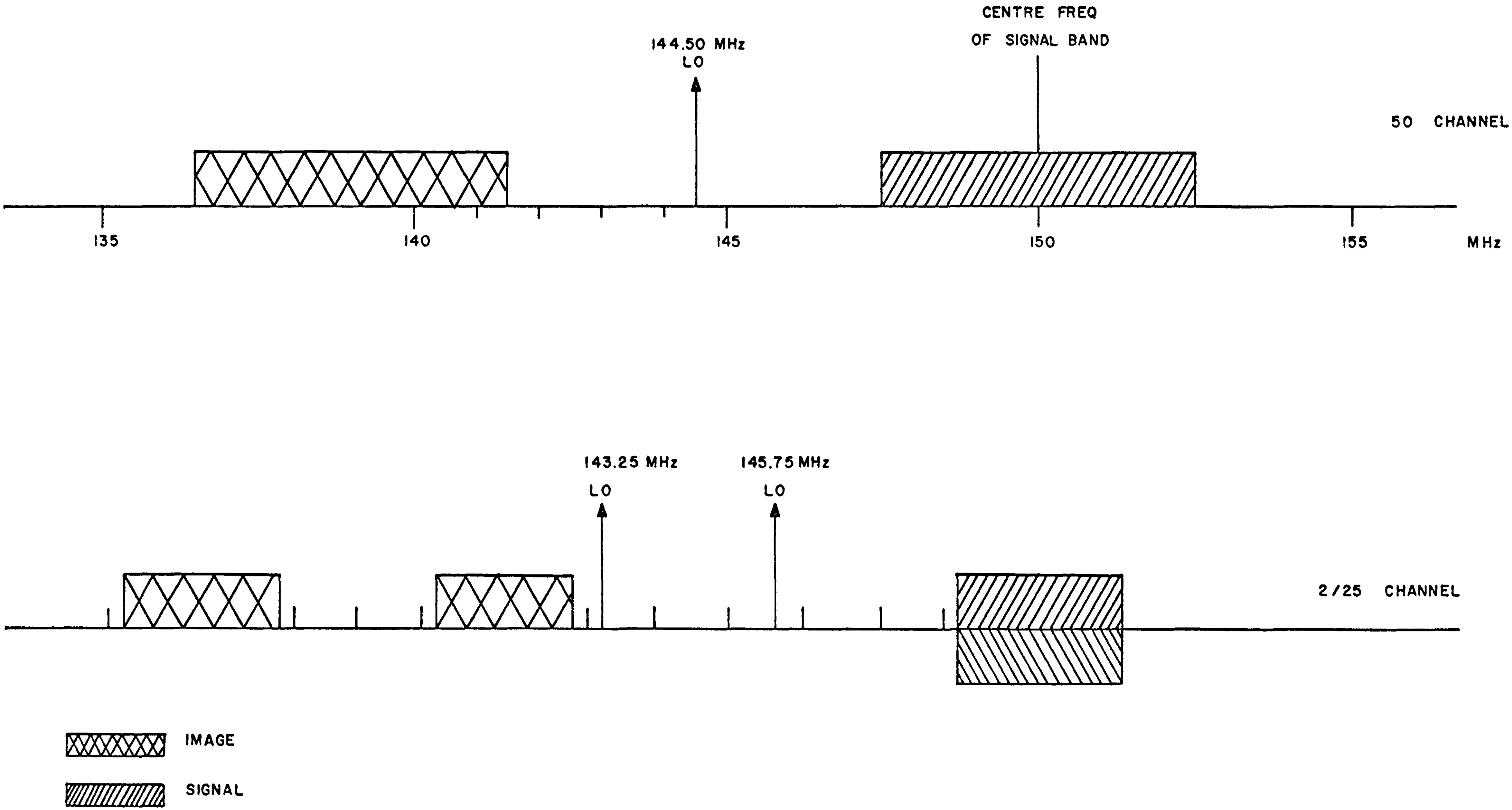


DIAGRAM SHOWING IMAGE RESPONSES FOR
BOTH CONFIGATIONS OF 50 CHANNEL RECEIVER

FIGURE 12

3. Integration of the Line Receiver with the DDP-116 Computer

3.1 General

The integration of an on-line computer into the system allows rapid measurement and correction for system gain and zero drifts and also provides convenient calibrated data displays. Figure 13 shows the block diagram of the computer radiometer system which was used for excited hydrogen line work on the 140-foot telescope in September 1967. The computer furnishes control bits for frequency switching, calibration noise control, integrator resetting, and a bit which removes the IF signal for measurement of DC zero errors. The system shown in Figure 13 is frequency switched at 10 cps; the outputs from all 50 channels are integrated for 45 ms and then held for 4 ms. During the hold time each channel is sampled by a multiplexer and A/D converter. The integrators are then reset by a 1 ms pulse. Figure 13 shows how these events are related in time to the 10 cps reference signal. The calibration noise tube is also fired at regular intervals for 50 ms during a signal frequency period. The computer accummulates many 50 ms integrations of three variables for each of the 50 channels. The variables and their relation to the receiver inputs are as follows:

$$A = G\,(T_L + T_s)$$

$$B = G\,T_s$$

$$C = G\,(T_L + T_s + T_{cal})$$

where G is the total system gain.

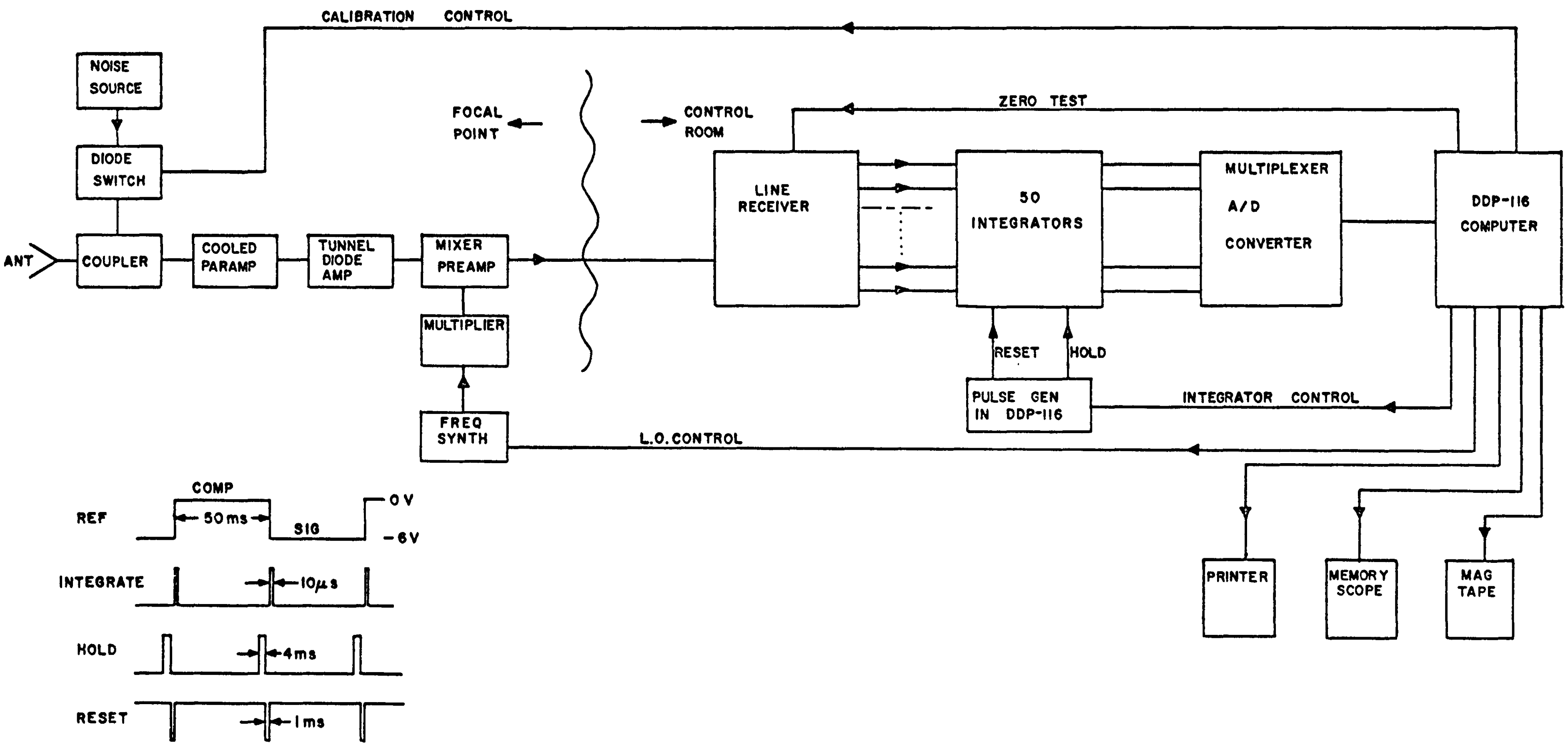
T_L is the line temperature spectrum to be measured.

T_S is the system noise temperature away from the line frequency and includes the continuum temperature of the source.

T_{cal} is the calibration signal noise temperature.

The following steps are taken by the computer to determine T_L in terms of T_{cal}:

$$GT_L = A - B.$$



FRONT END— 50 CHANNEL RECEIVER—COMPUTER INTERFACE

FIGURE 13

This is normalized by dividing by B to give

$$\frac{T_L}{T_s} = \frac{A - B}{B}$$

This is done for each of the 50 channels.

T_s in terms of T_{cal} is

$$T_s = \frac{B}{C - A} T_{cal}$$

This is also done for each of the 50 channels and T_s is then averaged over the 50 channels.

Using this average value for T_s, we then have

$$T_L = \frac{A - B}{B} \left(\frac{B}{C - A}\right)_{av} T_{cal}$$

for each of the 50 channels.

This procedure assumes that T_s is constant from one channel to another although it may vary with time; this is valid if the parametric amplifier bandwidth is large with respect to the 5 MHz being analyzed.

The selection of the integration times spent measuring A, B, C to get the best estimate of T_L depends on the frequency and time dependence of G and T_s and also the magnitude of T_{cal} and T_s.

As previously mentioned for the first and also subsequent experiments with this receiver, we spent 50 ms on A and B repetitively, except for every 2 seconds when 50 ms are spend on C. During the first experiment the system noise temperature was 100 °K and the calibration signal value was 4 °K. This enabled a value for T_s to be determined with small error resulting in a peak-peak noise on T_L which was equal to the theoretical figure for a Dicke radiometer. The largest error in the system is due to the detectors not being exactly square law. For example, if there is an unbalance in the total power between signal and comparison periods due to a change in gain or antenna temperature, there will be an error in the unbalance which is due to variations in detectors between individual channels. This error will have a maximum magnitude of about ± 2% of the unbalance in degrees, provided the unbalance is small compared to T_s. As a consequence, it is important to minimize any gain unbalance between signal and comparison bands. In the frequency switching case this is relatively easy to do if one has control of the RF amplifier bandpass shape. Unbalances for a 100 °K system are typically 1 °K or less, giving rise to baseline fluctuations due to imperfect detector laws of about ± 0.01°.

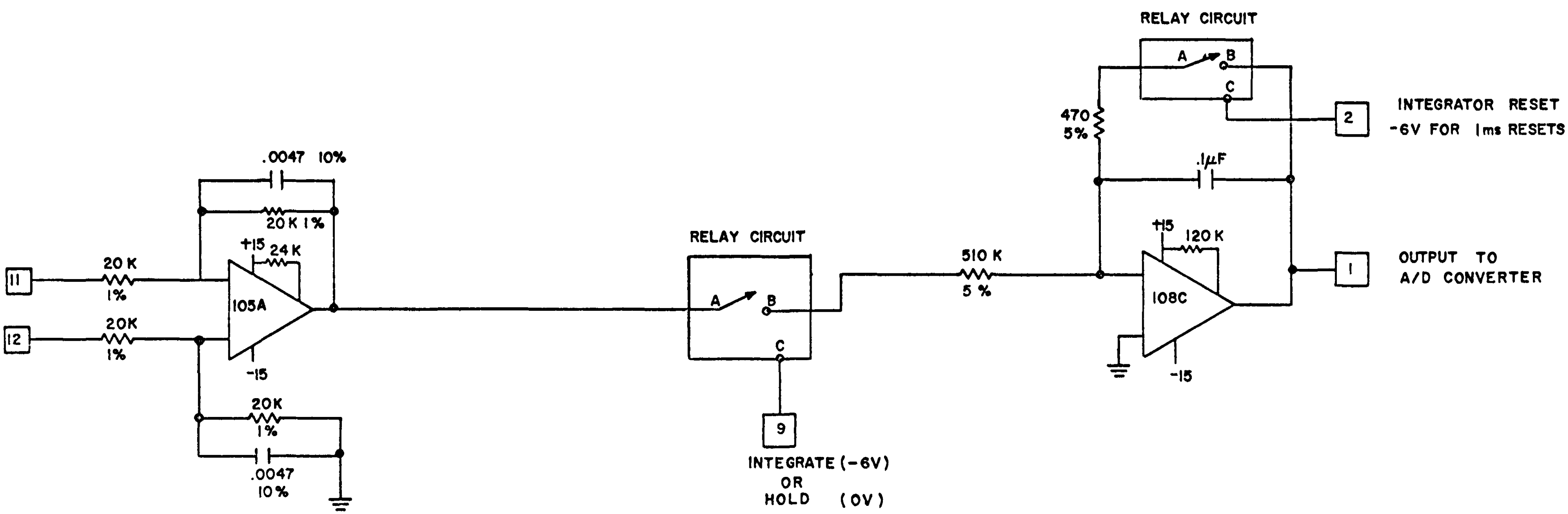
In the load switching case, especially when a 300 °K load is used, a gain modulator must be used to reduce the unbalance close to zero. The computer makes the same computation to determine T_L in the load switching case.

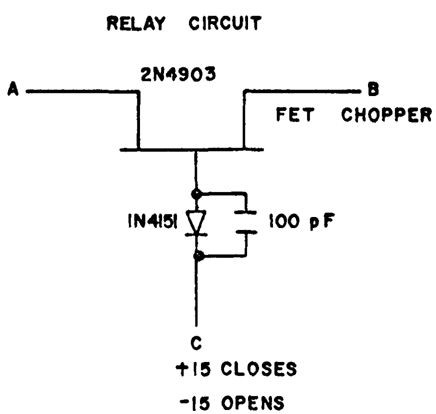
3.2 Integrators

The output from each channel in the line receiver goes to a differential input integrator card which is shown in Figure 14. These are installed in the rack which contains computer ancillary equipment. The circuit is self-explanatory; two FET switches control the integrate, hold and reset functions of the integrator. The differential input is used to minimize pick-up on the cables between the computer and the receiver.

3.3 Zero Checking

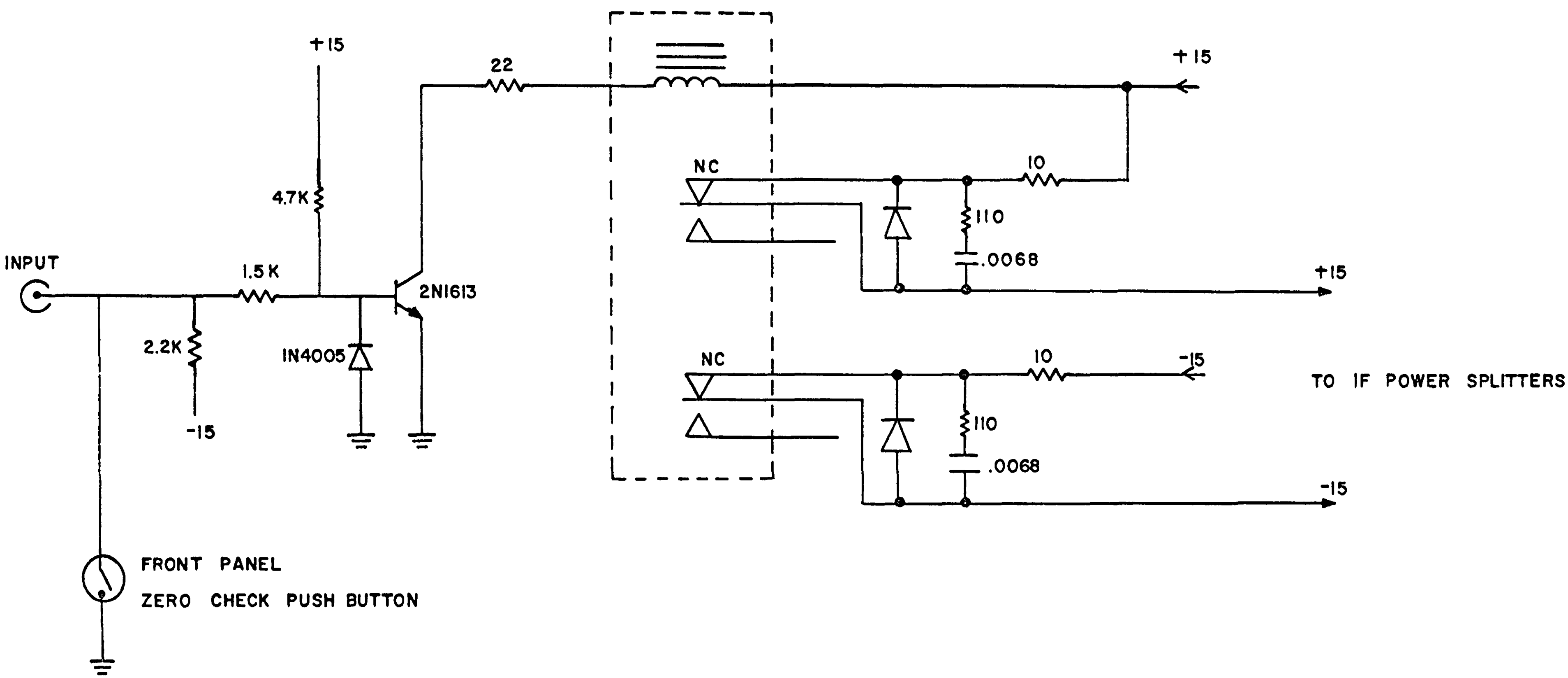
At the commencement of every scan the supply voltage to the IF amplifier in the line receiver is removed for 1 second. This is done so that the zero offsets for each channel may be measured. These values are stored and subsequent measurements of A, B and C are corrected for these offsets. The offsets are small (typically ± 10 mV) and have long time constants (thermal) and consequently need not be constantly checked. Figure 15 gives details of the zero checking circuit in the line receiver.





DIFFERENTIAL INPUT INTEGRATOR CIRCUIT

FIGURE 14



-6V OR OPEN CIRCUIT IS NORMAL— RELAY UNENERGIZED
0V OR SHORT CIRCUIT CHECKS ZEROS

ZERO CHECK CIRCUIT

FIGURE 15

4. Temperature Controller for the Filter Bank

The thermoelectric coolers are manufactured by Scientific Columbus, Inc., model number TCP-6-40F. Two of these units are used, one on each side of the filter package. One of these coolers would have been sufficient if the full current capability of the unit had been used. However, it was considered that the design of the temperature controller would be easier if two units were run in series with a maximum current of 10 amps. If the direction of the current through the coolers is reversed, they become heaters.

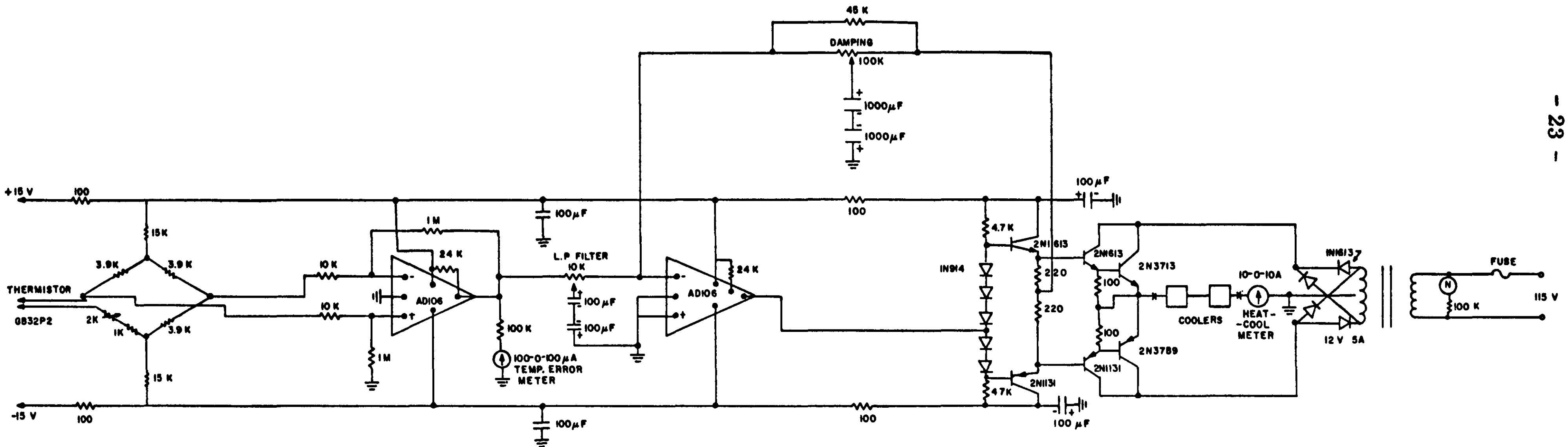
The circuit for the temperature controller is shown in Figure 16. A photograph of the unit is shown in Figure 17.

The temperature sensor is a thermistor which is incorporated into a DC bridge. The error voltage is amplified by an operational amplifier which drives another operational amplifier and a temperature error meter. This meter reads approximately ± 5 °C from the preset temperature which is about 25 °C.

The second operational amplifier drives an amplifier which can drive current in either direction through the heater/cooler. This is accomplished by connecting an NPN and PNP power transistor in series between ± 5 V supply rails. The coolers are connected between the junction of the transistors and ground. The driving circuit for these transistors, which are both heat sink mounted, is conventional.

The second operational amplifier is used to stabilize the servo loop. The input low pass network is used to filter out any 60 cps and noise which may be present at the input. The damping adjustment has maximum effect in the center of the potentiometer.

The temperature controller maintained the temperature of the filter package to within a few tenths of a degree Centigrade of the preset value of 25 °C.



CIRCUIT OF TEMPERATURE CONTROLLER

FIGURE 16

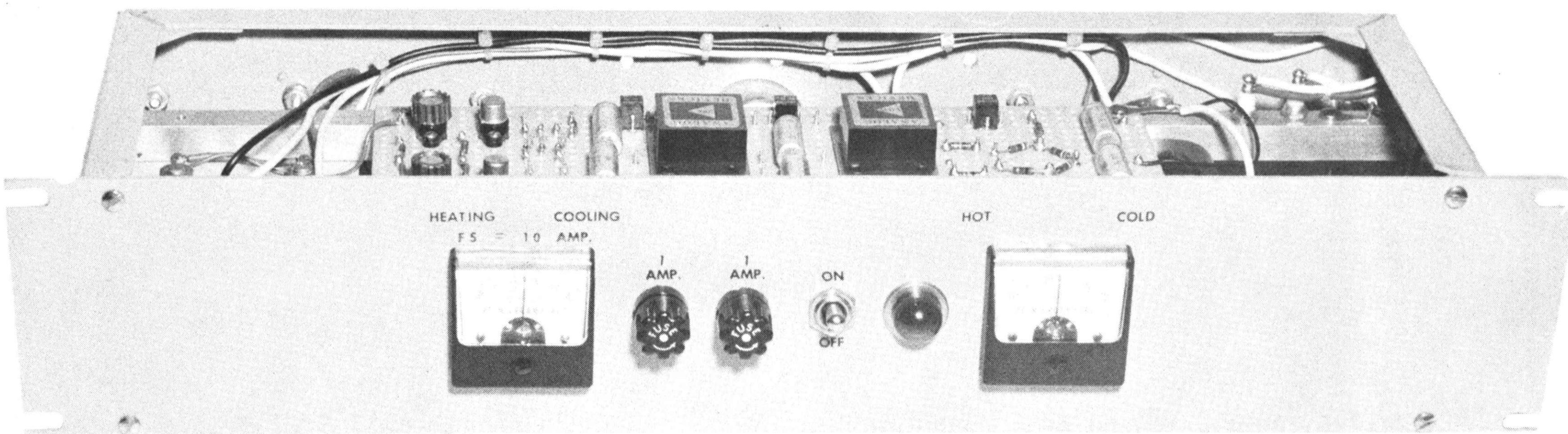


Figure 17 — Temperature Controller for Filters

5. Observations Made With the Receiver

The first program using the line receiver was an excited hydrogen line survey made by Mezger, Burke, Wilson and Reifenstein in September 1967. The system used is shown in Figure 13. The single-stage parametric amplifier was manufactured by AIL and has 17 dB gain and 200 MHz bandwidth centered at 5000 MHz. The total system noise temperature was approximately 90 °K of which 15° was second stage contribution, 10° was waveguide losses, and approximately 15° was the antenna temperature at zenith. The amplifier was cooled by an A. D. Little 204LS refrigerator.

Figure 18 shows an on-off composite profile of M17 and shows three spectral lines — H 137β, H 109 α, and He 109 α. The patch of apparent noise at V(LSR) of 135 km/sec is due to the fact that the β line scans were set up so that the line would appear in the comparison band as well; consequently, its (negative apparent profile) signal is averaged in with the otherwise flat baseline of the M17 H 109 α peak. The source of the flat region between the hydrogen and helium 109 α line is unknown.

Figure 19 shows an on-off profile of the H 109 α line in a weaker source which is identified by its galactic coordinates G30.33-0.15. The on source integration time was 60 minutes.

One problems which became apparent when this receiver was used was that there was a serious perturbation of the spectrometer baseline caused by interference of signals reflected at the antenna feed and the surface of the parabolic reflector. Figure 20 shows an example of this effect. S. Weinreb has shown that the total perturbation of the baseline ΔT is given by

$$\Delta_T = 2|\Gamma||\Gamma'|(T_1 - T_A)\cos\left(\frac{4\pi Lf}{c} + \varphi\right)$$

where $|\Gamma|$ is the reflection coefficient of the feed.

$|\Gamma'|$ is the reflection coefficient of the antenna.

T_1 is the temperature of the noise radiated by the receiver.

T_A is the antenna temperature of the source.

L is the focal length of the telescope.

f is the observation frequency.

c is the velocity of light.

φ is an unknown phase angle.

The period of the effect is fixed for a particular telescope and is independent of the observation frequency; for the NRAO 140' telescope it is 7.5 MHz.

The effect does not cancel out in on-off measurements on strong continuum sources because of its dependence on T_A.

The effect can be minimized by

1) Careful matching of feed and parametric amplifier combination.

2) Averaging spectrum taken at focal positions $\frac{\lambda}{8}$ and $-\frac{\lambda}{8}$ from the nominal focal positions.

3) Placing scattering material at the reflector vertex.

6. The author thanks E. C. Reifenstein for permission to use Figures 18, 19 and 20 in this report.

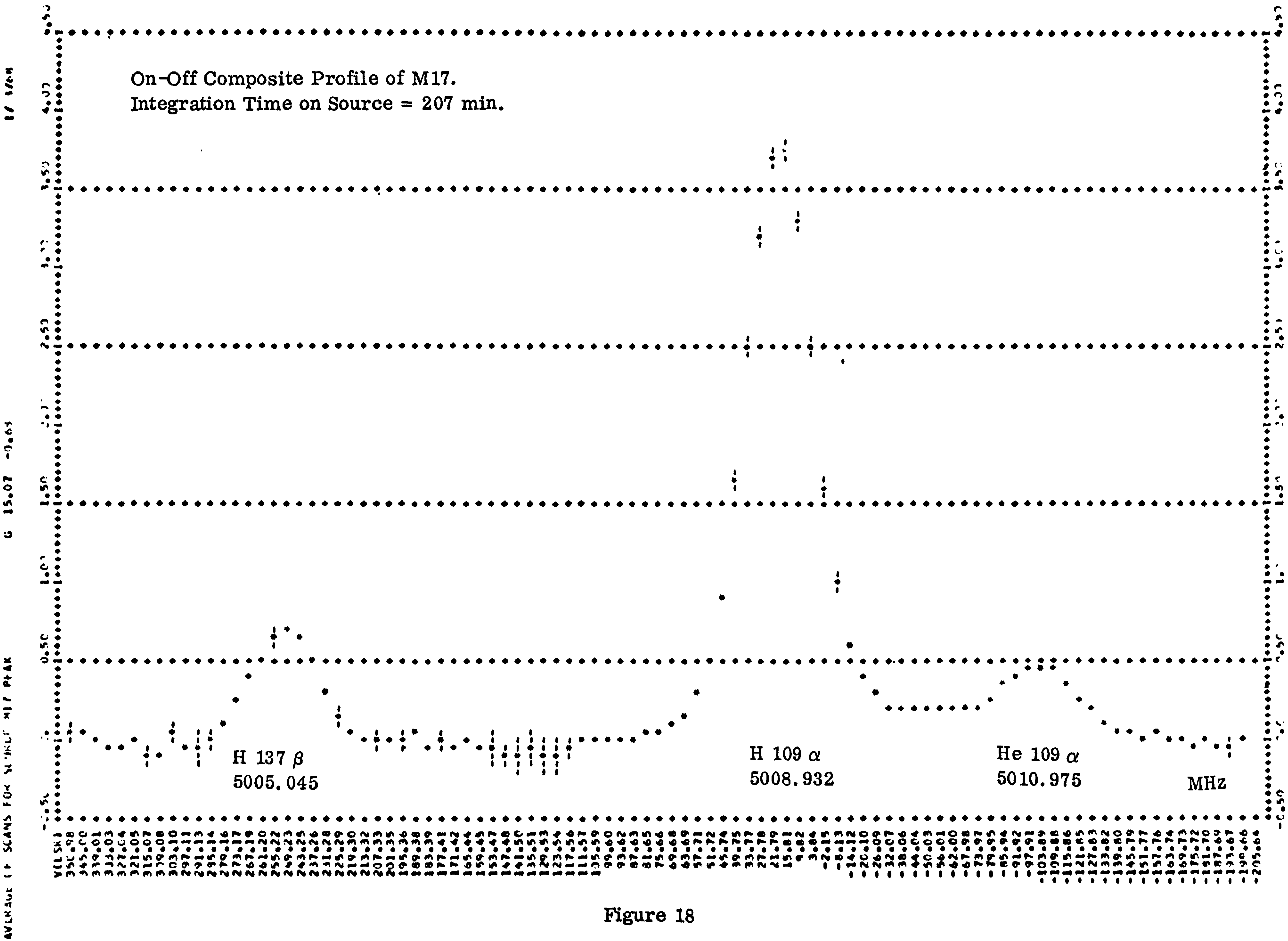

On-Off Composite Profile of M17.
Integration Time on Source = 207 min.
H 137 β
5005.045
H 109 α
5008.932
He 109 α
5010.975
MHz

Figure 18

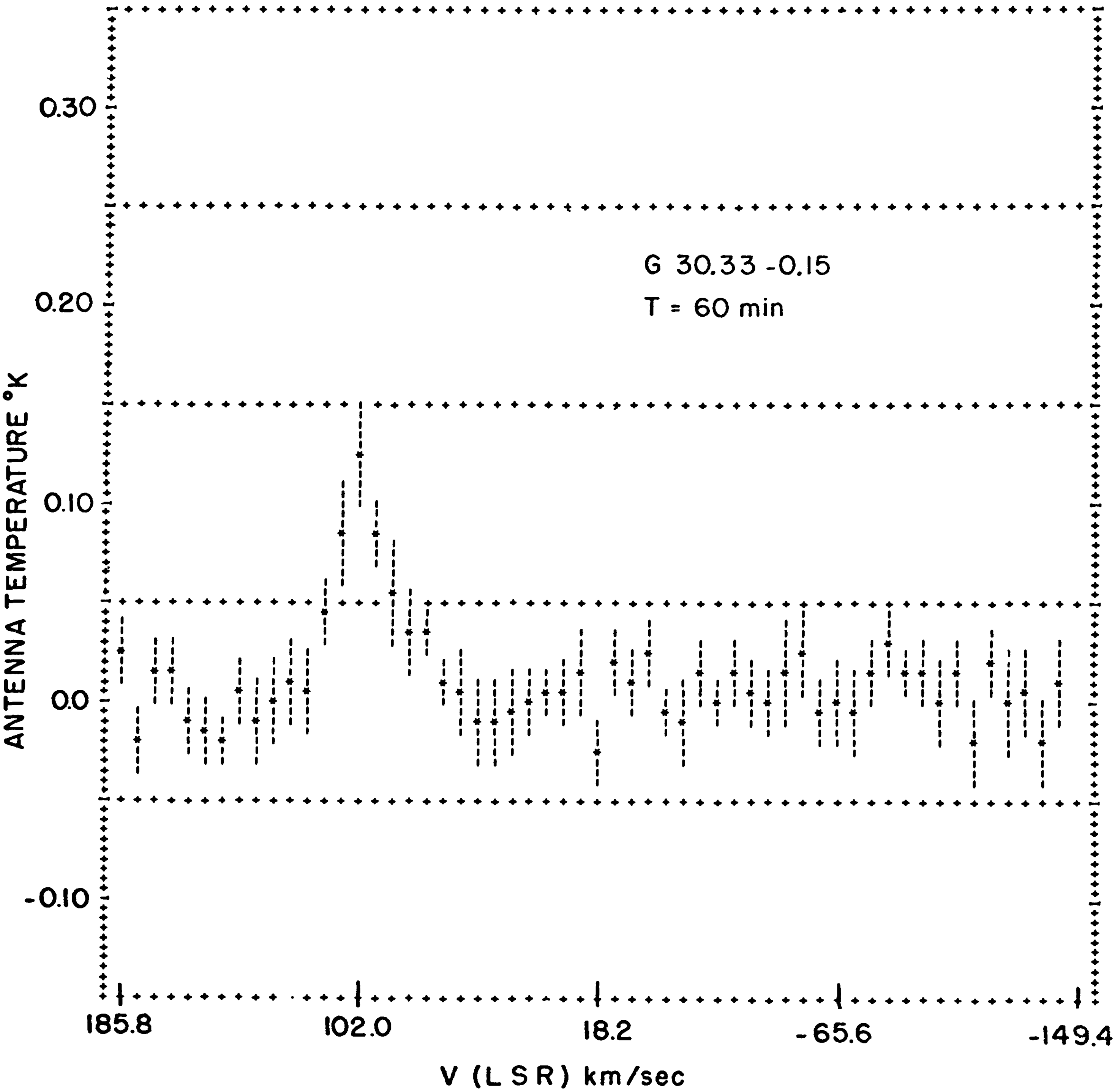


Figure 19 — Weak Source Profile

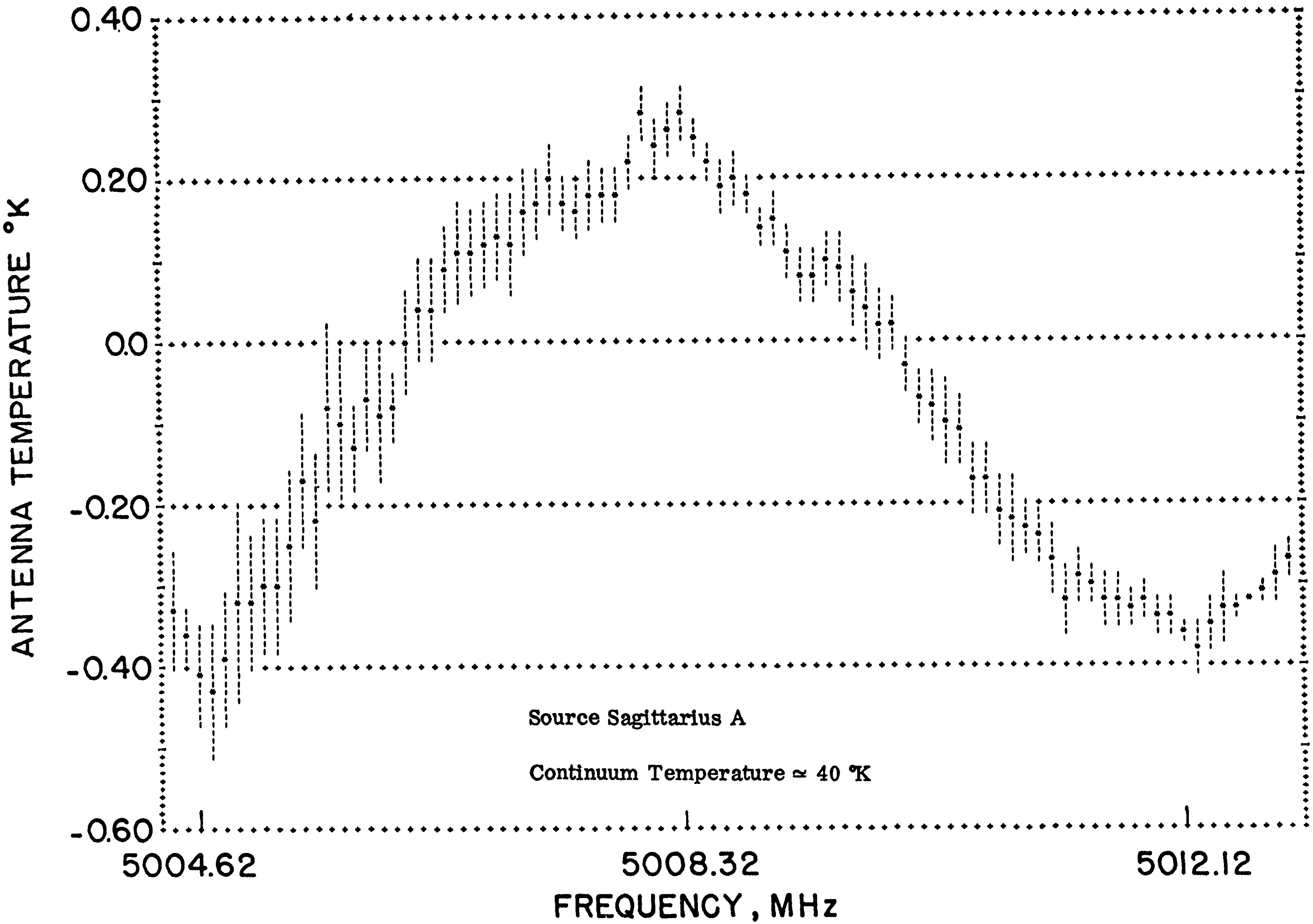


Figure 20 — Antenna-Feed Interference Spectrum