An experimental portable c.w. intercept receiver
Hendry, A.; Poirier, A.L.; Royer, G.M.

Publisher's version / Version de l'éditeur:
https://doi.org/10.4224/21274203
Report (National Research Council of Canada. Radio and Electrical Engineering Division : ERB), 1964-06

ERB-675

SECRET

COPY NO. 5

NATIONAL RESEARCH COUNCIL OF CANADA
RADIO AND ELECTRICAL ENGINEERING DIVISION

AN EXPERIMENTAL PORTABLE C. W. INTERCEPT RECEIVER

A. HENDRY, A. L. POIRIER, AND G. M. ROYER

OTTAWA
JUNE 1964 NRC # 35671

ABSTRACT

An experimental portable c.w. intercept receiver for battlefield use is described. The sensitivity of the receiver varies between 30 and 45 db below 1 milliwatt per square metre in the frequency range 7.0 to 15.5 Gc/s. The receiver is also responsive to pulsed radiation of the same average power level. It is completely self-contained and utilizes temperature-compensated transistor circuitry. An audio tone warns the operator when he is under radar surveillance. Circuit schematics, details of the microwave components, and data on receiver performance are included in the report.

CONTENTS

	Page No.
Introduction	1
Detection Range of Intercept Receivers	2
Description of Receiver System	4
Antenna	5
Modulator-Detector Unit	6
Description of Circuitry Preamplifier Main Amplifier and Audio Circuits Oscillator Circuits Modulator-Driver	9
Discussion of Circuitry	12
Performance of Circuits	14
Setup Procedure	15
Discussion of Other Systems Tried	16
Performance of Receiver with C.W. Radiation	16
Performance of Receiver with Pulsed Radiation	17
Acknowledgment	19
References	19

FIGURES

Fig. 1 Detection range nomogram

Fig. 2 Block diagram of receiver

Fig. 3 H-Plane antenna patterns

Fig. 4 E-Plane antenna patterns

Fig. 5 Sectional view of modulator and crystal mount

Fig. 6 Frequency response characteristic of detector

Fig. 7 Modulation efficiency

Fig. 8 Frequency response characteristic of modulator-detector unit

FIGURES (cont'd.)

Fig. 9 Preamplifier schematic diagram

Fig. 10 Amplifier schematic diagram

Fig. 11 Schematic diagram of 8 kc/s and 100 kc/s oscillators

Fig. 12 Temperature drift of 100 kc/s oscillator

Fig. 13 Modulator-Driver schematic diagram

Fig. 14 Variation of receiver sensitivity with frequency

Fig. 15 Variation of receiver sensitivity with temperature

Fig. 16 Receiver performance with pulsed radiation

PLATES

Plate I Exterior view of receiver

Plate II Antenna, varactor mount, and crystal mount

AN EXPERIMENTAL PORTABLE C.W. INTERCEPT RECEIVER

- A. Hendry, A.L. Poirier*, and G.M. Royer* -

INTRODUCTION

This report describes a portable intercept receiver designed for passive detection of c.w. and f.m. c.w. battlefield surveillance radars. The receiver, which has been developed for the Canadian Army, is designated as the "C.W. Micradet Receiver". ("Micradet" is an acronym for MICrowave RAdiation DETector.)

The primary purpose of the receiver is to warn infantry patrols that they are under enemy radar surveillance in order that they may take appropriate evasive action; e.g., that they remain motionless or under cover when illuminated by the radar, and proceed when the illumination has ceased. A secondary purpose is to provide means for approximate determination of the azimuth of an enemy radar.

The design of this receiver is based upon a proposal made by Canadian Arsenals Limited [1]. Work on the project was begun at CAL but owing to the impending closure of the Instrument and Radar Division of CAL, the project was transferred to the National Research Council where the main part of the work was done with assistance from the Defence Research Telecommunications Establishment, Ottawa, Ont.

The work described in this report is an extension of work previously reported by the National Research Council on the problem of detection of battlefield surveillance radars [2].

In the development of this receiver, the design goal has been to provide a system capable of detecting contemporary f.m. c.w. radars at distances at least as great as the maximum ranges of such radars against a single person. For increased usefulness, the detection range of the intercept receiver should be at least equal to the maximum range capability of the radars against which it is likely to be employed. Table I shows characteristics of two battlefield surveillance radars which are currently in use.

---

*Defence Research Telecommunications Establishment, Shirley Bay, Ottawa, Ont.

Table I — Characteristics of C.W. Radars

Radar	AN/PPS-7	Olifant
Country of origin	USA	France
Transmitter power	12 milliwatts	15 milliwatts
Antenna gain	27 db	27 db
Frequency range	8.5 - 9.6 Gc/s	8.9 - 9.6 Gc/s
Detection range	1000 metres (personnel) 2000 metres (vehicles)	1200 metres

For intercept receiver design purposes, a "standard" radar characterized by the following parameters has been chosen:

Transmitter power	15 milliwatts
Antenna gain	27 db
Frequency	X-band (8.2 - 12.4 Gc/s)
Detection ranges:	
Personnel	1000 metres
Vehicles	2000 metres

DETECTION RANGE OF INTERCEPT RECEIVERS

The theoretical free-space maximum range at which radiation may be detected is given by:

$$R_{max} = \left(\frac{P\, G_t}{4 \pi F_{min}} \right)^{\frac{1}{2}} \text{ metres,}$$

where P = radiated power (watts)
G_t = transmitting antenna gain (ratio)
F_{min} = minimum detectable flux (watts/metre^2).

In terms of the minimum detectable signal power, S_{min} (watts), the range is given by:

$$R_{max} = \left(\frac{P G_t A_e}{4\pi S_{min}}\right)^{\frac{1}{2}} \text{ metres,}$$

where A_e = effective receiving antenna capture area (metres^2) and S_{min} is the minimum detectable power level referred to the antenna terminals. A_e may, typically, be 50 to 60% of the actual area of the antenna aperture. A_e is related to the antenna gain by

$$A_e = \frac{G_r \lambda^2}{4\pi}$$

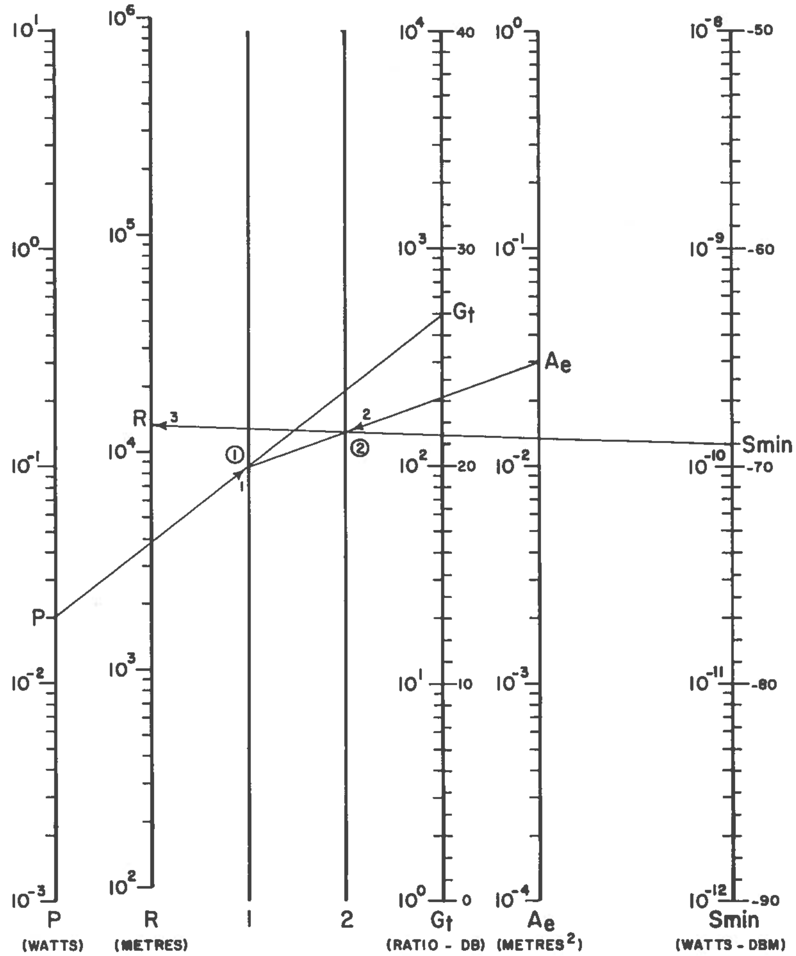
where G_r is the receiving antenna gain. (Note that G_r is a function of λ, the wavelength.)

The preceding relationship of R_{max} to P, G_t, A_e and S_{min} is displayed in Fig. 1. To use this nomogram, a line is drawn from the Point P (corresponding to the power) to the Point G_t (corresponding to the antenna gain of the transmitting source). The intersection of this line with the vertical line No. 1 provides a point to be used in the next step of calculation. To this latter point a line is drawn from the point A_e (which corresponds to the receiving antenna effective area). This latter line intersects with vertical line No. 2. Finally, a line is drawn, from the appropriate point S_{min} on the right-hand scale, through the intersection point No. 2; this line terminates on the range scale, yielding the desired range. (See the sample calculation, which has been made for $P = 2 \times 10^{-2}$ watts, $G_t = 27$ db, $A_e = 3 \times 10^{-2}$ metres^2, and $S_{min} = -69$ dbm, yielding $R = 1.4 \times 10^4$ metres.)

Alternatively, the chart may be used to find the required system sensitivity S_{min}, for a given set of values of P, G_t, A_e, and R_{max}.

The required sensitivity for the detection of the "standard" radar at 2000 metres range is $F_{min} = 1.49 \times 10^{-4}$ milliwatts/metre^2, i.e., -38.3 dbm/metre^2, while a sensitivity of -32.3 dbm/metre^2 is required in order to obtain 1000 metres range.

$$R = \sqrt{\frac{P G_t A_e}{4\pi S_{min}}}$$



P = TRANSMITTED POWER
Gt = TRANSMITTING ANTENNA GAIN
Ae = RECEIVING ANTENNA EFFECTIVE AREA
Smin = MINIMUM DETECTABLE SIGNAL

Fig. 1 Detection range nomogram

Two values of effective antenna area have been considered for this receiver; viz., 0.001 metres^2 and 0.01 metres^2. The former antenna would be approximately the size of the horn antenna used by Pulfer[2]. With such an antenna, the receiver circuitry would have to be responsive to a power level of -68.3 dbm at the antenna terminals, in order to achieve 2000 metres range performance, while with the larger antenna, a sensitivity of -58.3 dbm would suffice.

On the basis of preliminary measurements, and experience with other c.w. detection systems [3], a minimum detectable power level of -60 dbm was expected to be achievable. Consequently, a horn antenna having an aperture area of at least 0.01 metres^2 was specified for the receiver.

DESCRIPTION OF RECEIVER SYSTEM

The experimental model of the receiver (Plate I) is completely self-contained in a molded plastic case and may be either hand-held or mounted on a tripod. The principle of operation may be understood by reference to the block diagram (Fig. 2). Energy delivered to the receiver by the antenna is "chopped" at 92 kc/s by a varactor-modulator located in the waveguide just beyond the throat of the horn. A broad-band crystal detector, based on a design by Staniforth [4] forms an integral part of the modulator-crystal mount assembly. (See Plate 2 which shows the horn antenna, the varactor mount, and the crystal mount.)

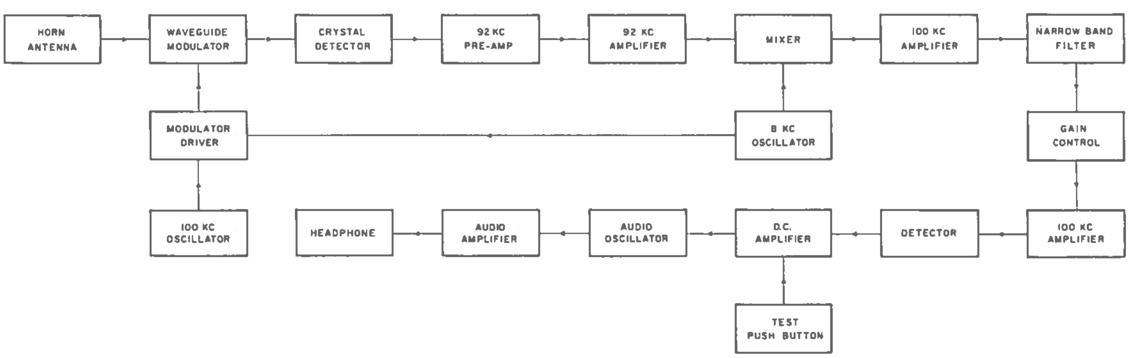


Fig. 2 Block diagram of receiver

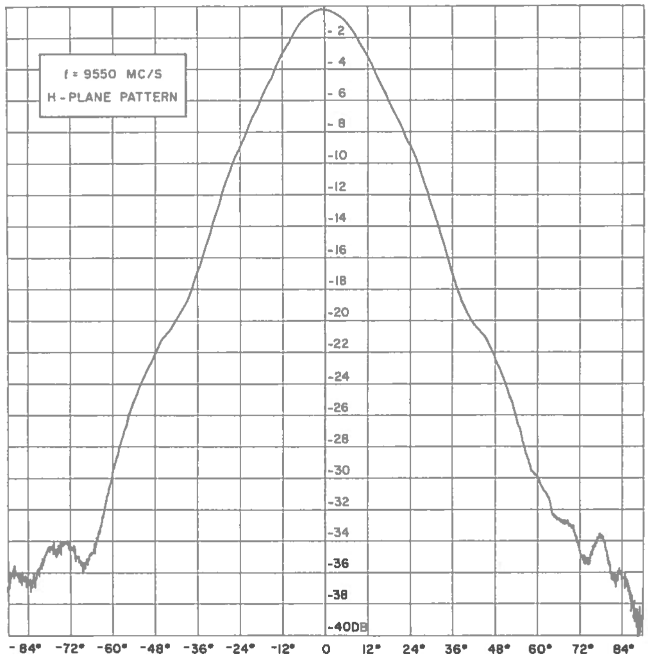
When r-f energy is incident on the antenna, a 92 kc/s output is thus obtained from the crystal detector. This output is amplified first in the low-noise preamplifier, then in a 3-stage 92 kc/s amplifier, then converted to 100 kc/s for further amplification, and finally rectified in the detector circuit. The resulting d-c voltage is used to trigger an audio oscillator whose output, after amplification, is fed to the operator's headset. The 92 kc/s modulator output, which is applied to the varactor, is generated by mixing the output of a very stable 100 kc/s oscillator with that of a free-running 8 kc/s oscillator and selectively filtering the difference frequency as the output.

A receiver bandwidth of 6 c/s is obtained through the use of a magnetostrictive band-pass filter of this bandwidth in the 100 kc/s amplifier. In order to keep the wanted signal within the passband of this filter, it is necessary that the 100 kc/s oscillator frequency must be within ±3 c/s of the filter centre frequency over the range of ambient temperatures in which the receiver is expected to function. To assure this, the 100 kc/s oscillator employs a similar filter in its feedback path. (The two filters were supplied as a matched pair by the manufacturer, Spectran Electronics Corporation.)

This scheme follows closely the proposal made by CAL; however, a number of changes to their scheme were found to be necessary. These are more fully described in the detailed description of the circuitry and microwave components which follows.

ANTENNA

The horn antenna which is shown in Plate II, was fabricated from sheet brass. (Consideration has been given to use of metallized laminate board in future development in order to reduce weight.) The horn has an aperture 9.5 × 12.1 cm, and is 10.8 cm long. H-plane and E-plane antenna patterns, taken at 9550 and 15,000 mc/s are shown in Figs. 3 and 4, respectively. The half-power beamwidths at 9550 mc/s are 20° and 23° in the E- and H-planes, respectively. (In the normal carrying position, the E-plane is vertical.) Because of the low ratio of horn length to aperture (required in order to avoid excessive length), the side-lobe level in the E-plane is rather high, the first side lobes being about 6 db below the main beam at X-band. However, since there is no appreciable dip in response between the side lobes and the main beam, the resulting pattern is quite satisfactory for the intended use. The gain of the antenna is 18.4 db at 9550 mc/s.



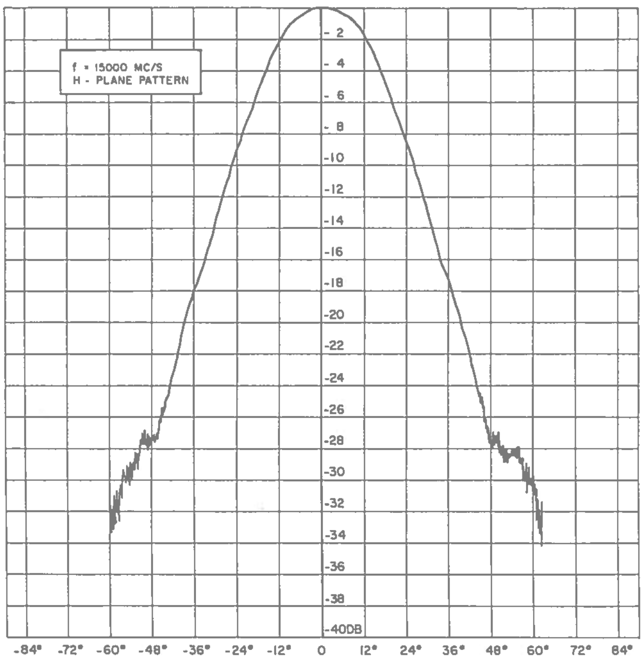


Fig. 3 H-Plane antenna patterns

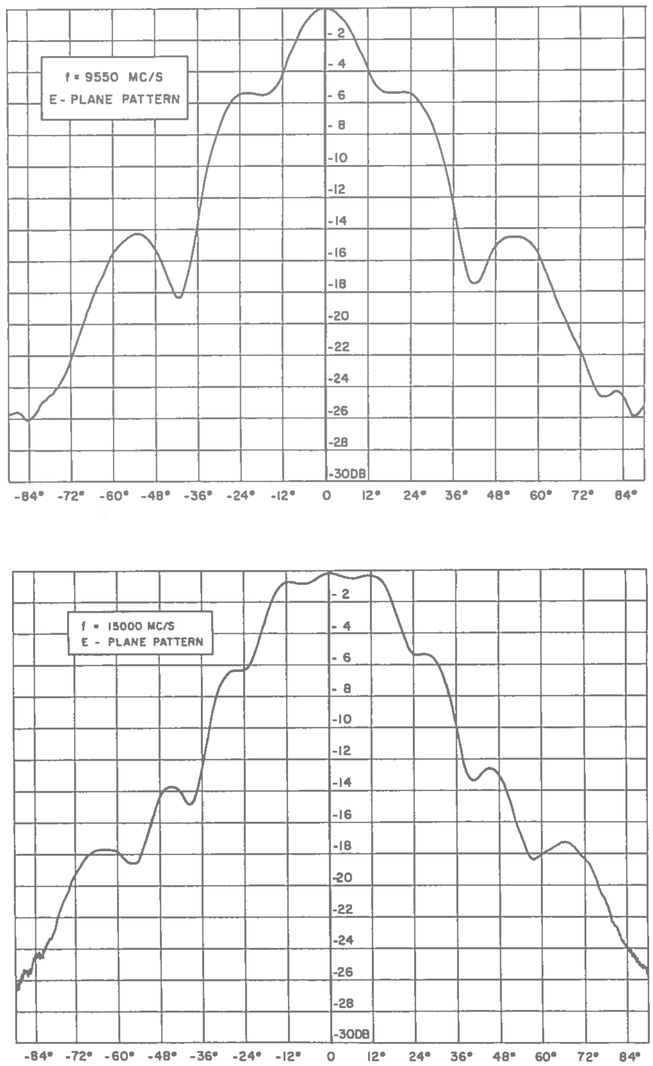


Fig. 4 E-Plane antenna patterns

MODULATOR-DETECTOR UNIT

The modulator-detector unit (Fig. 5) consists of a Type-MA4554 diode, which modulates the incoming signal, followed by a Type-1N78B crystal detector. The modulator and detector diodes are located in a section of ridged waveguide. Modulation is achieved by varying the bias on the varactor, which, in turn, changes the strength of the signal reaching the detector. Between the ridged waveguide and the horn there is a matching section with a tapered ridge and a section of X-band waveguide which terminates the horn.

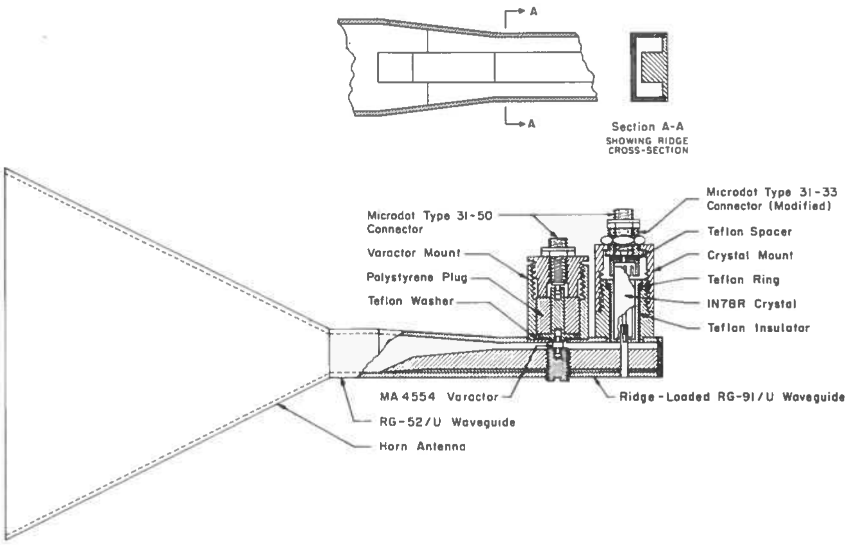


Fig. 5 Sectional view of modulator and crystal mount

The cutoff frequency in the X-band waveguide is 6.56 Gc/s, whereas in the ridged waveguide it is 4.95 Gc/s. To keep reflections small, the tapered ridge in the matching section was designed so that the'cutoff frequency in it varies approximately linearly with distance in the direction of propagation.

The dimensions of the detector mount and ridged waveguide are essentially the same as those used by Staniforth [4] for a broad-band crystal mount. There are several reasons for using ridged waveguide in preference to rectangular waveguide without a ridge.

a) Ridged waveguide can be designed to have a wider frequency bandwidth than rectangular waveguide. In our case the cutoff frequency for the TE_{10} mode is 4.95 Gc/s and for the TE_{20} mode is 14.4 Gc/s. This makes it possible to operate at frequencies further from cutoff, where the characteristic impedance and guide wavelength vary more slowly with frequency.

b) The waveguide impedance can be varied by changing the dimensions of the ridge.

c) The fields in the waveguide are concentrated in the region between the top of the ridge and the top of the guide. The modulating diode is located in this gap, which should help to control the amount of energy reaching the detector.

d) In our case the gap was approximately equal to the length of the modulating diode. This made it possible to make the leads from the diode short, and thus minimize lead inductance.

Figs. 6, 7, and 8 indicate the characteristics of the modulator-detector unit. To make the measurements for these figures, the horn was removed and microwave energy was fed into the X-band section of waveguide. The bias conditions on the diodes and the peak-to-peak modulating voltage were equal in magnitude to those actually used in the receiver. The input power was square-wave-modulated at 100 c/s, while the modulating voltage applied to the varactor was a square-wave at 1000 c/s. This resulted in a detected output of the form shown in Fig. 8. However, the actual modulating waveform used in the receiver is approximately a sine wave. It was observed that, at some frequencies, sine-wave and square-wave modulating waveforms of the same peak-to-peak voltage did not produce the same peak-to-peak video voltages at the detector. This result occurred at frequencies where the maximum and minimum modulating voltages do not produce, respectively, the maximum and

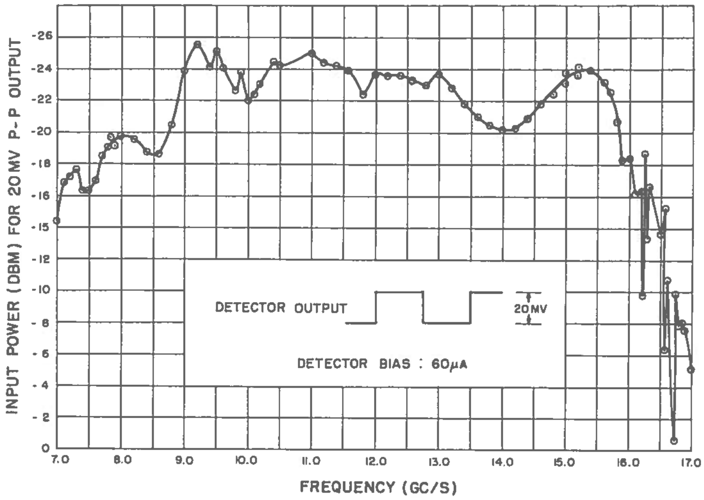


Fig. 6 Frequency response characteristic of detector

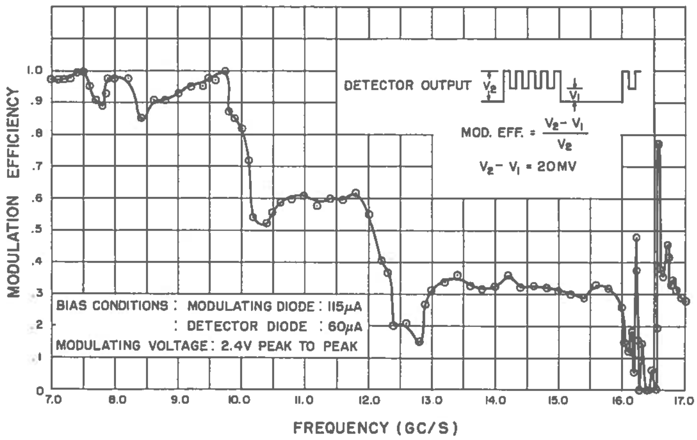
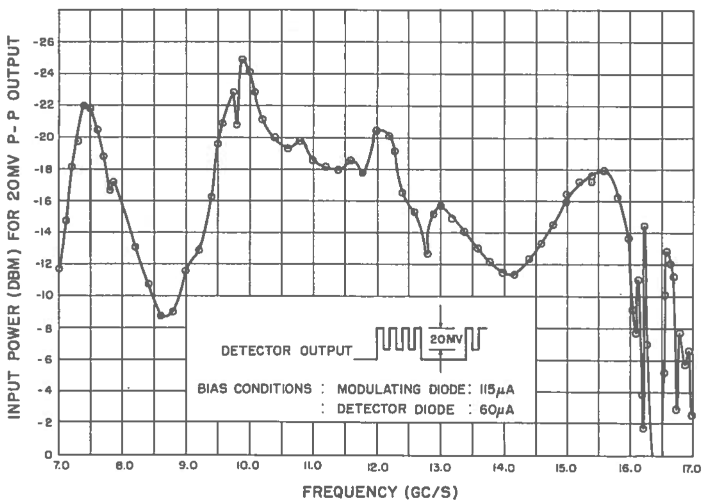


Fig. 7 Modulation efficiency



SECRET

Fig. 8 Frequency response characteristic of modulator-detector unit minimum detected voltages. The above condition was present mainly above 16 Gc/s and therefore the figures should still be valuable in assessing the modulator-detector characteristics.

Fig. 8 shows the over-all performance of the modulator-detector unit. This curve was obtained by measuring the input power required to produce a modulated output of 20 mv peak-to-peak. It should be noted that the input power recorded in the figures was the peak power, not the average power. By comparing the modulator-detector characteristics (Fig. 8), with the C.W. Micradet Receiver sensitivity curve (Fig. 14) it can be seen that the major variations with frequency correspond. Differences in the curves are caused mainly by the characteristics of the horn.

Fig. 6 shows the characteristics of the detector alone. To obtain this curve, the modulating diode was removed and square-wave-modulated microwave energy was fed in at the input. The input power shown is the peak input power required to give a 20 mv peak-to-peak detected voltage.

By comparing Figs. 6, 7, and 8, it can be seen that:

a) The drop in sensitivity of the modulator-detector at about 8.5 Gc/s is not caused by the detector or poor modulation efficiency. It is probably, therefore, due to an increase in the minimum insertion loss of the modulator.

b) Between 10 Gc/s and 15.5 Gc/s, Fig. 6 shows that, with the modulating diode removed, the input power required to produce a 20 mv peak-to-peak output is fairly constant. A greater power is required to produce a modulated output of 20 mv peak-to-peak. This power difference can be explained by the fact that the modulation efficiency is somewhat less than 1.0, and, as in (a), that there is insertion loss due to the modulator.

c) Between 16 Gc/s and 17 Gc/s both detector sensitivity and modulation efficiency drop off and become very frequency-sensitive. This is probably caused by excitation of the TE_{20} mode, resulting in a weak field near the centre of the waveguide where the modulating and detecting diodes are located.

d) The sensitivity of the modulator-detector drops off below 7.5 Gc/s because we are approaching the cutoff frequency of the X-band section of waveguide.

During initial work several different types of modulating diodes, modulator mounts, and waveguide configurations were tried. The modulating diode used in the receiver was a Type-MA4554 varactor diode. Other diodes tried were:

a) Type-MA4571 P-I-N diode
b) Type-MA4557 varactor diode
c) Type-1N263, 1N23B and MA4125A detector diodes in X-band waveguide

A section of ridged waveguide, with a narrower ridge than that finally used, was tested. The narrower ridge resulted in higher waveguide impedance, and it was thought that this might improve the modulation efficiency. Also, a cylindrical bypass capacitor for the modulator, instead of the disc type shown in Fig. 5, was tried. Over the whole frequency band, however, none of the above changes resulted in better performance than that of the final design shown herein.

DESCRIPTION OF CIRCUITRY

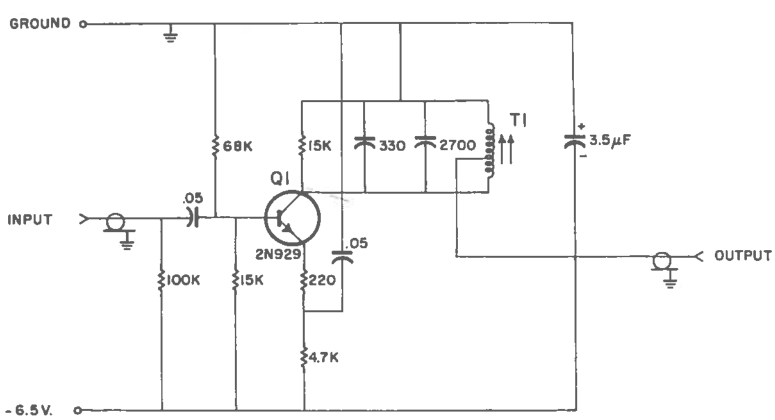


Fig. 9 Preamplifier schematic diagram

Preamplifier

The 92 kc/s component of the signal at the output of the crystal detector is amplified in a low-noise frequency-selective transistor preamplifier (Q1), and the amplified signal is applied to the input of the main amplifier. A schematic diagram of the preamplifier circuit is shown in Fig. 9. The preamplifier is tuned in the collector circuit of the transistor. The inductor which forms part of the resonant collector circuit is tapped and serves as an impedance-matching coupling transformer. The bandpass of the preamplifier is 6 kc/s, and the voltage-gain is 15 db.

Main Amplifier and Audio Circuits

The main amplifier (Fig. 10) is a high-gain 92-kc/s tuned amplifier. It consists of three 92 kc/s amplifier stages (Q2 to Q4), a frequency-converter (Q5), the narrow-band 100 kc/s magnetostrictive filter, and three 100 kc/s amplifier stages (Q6 to Q8), (see Fig. 10). The bandwidth of the main amplifier is 6 c/s and the maximum voltage gain is 129 db. The gain is controlled by a potentiometer which is part of the coupling circuit between Q6 and Q7. The audio-frequency circuit consists of a threshold circuit (Q11), a d-c amplifier (Q12), a gated audio-frequency oscillator (Q13), and an audio-frequency amplifier (Q14 and Q15).

Amplified signals in the collector circuit of Q8 are rectified and applied to the audio threshold circuit. When the rectified voltage exceeds 0.6 volt, the excess voltage is amplified in Q11 and Q12 and applied to the base of Q13. Q13, which in the absence of a detected signal is unbiased and therefore cut off, is gated-on by the applied voltage and oscillates at a 3 kc/s rate. The 3 kc/s signal is amplified in Q14 and Q15 and applied to the operator's headset.

Oscillator Circuits

Schematic diagrams of the 100 kc/s and 8 kc/s oscillator circuits are shown in Fig. 11. The frequency of the 100 kc/s oscillator is stabilized by a magnetostrictive filter which is matched to the one used in the main amplifier circuit. The center frequencies of the two magnetostrictive filters are identical at 68°F, and track each other within a 0.05 c/s /°F change in temperature. Fig. 12 shows the temperature drift of the 100 kc/s oscillator.

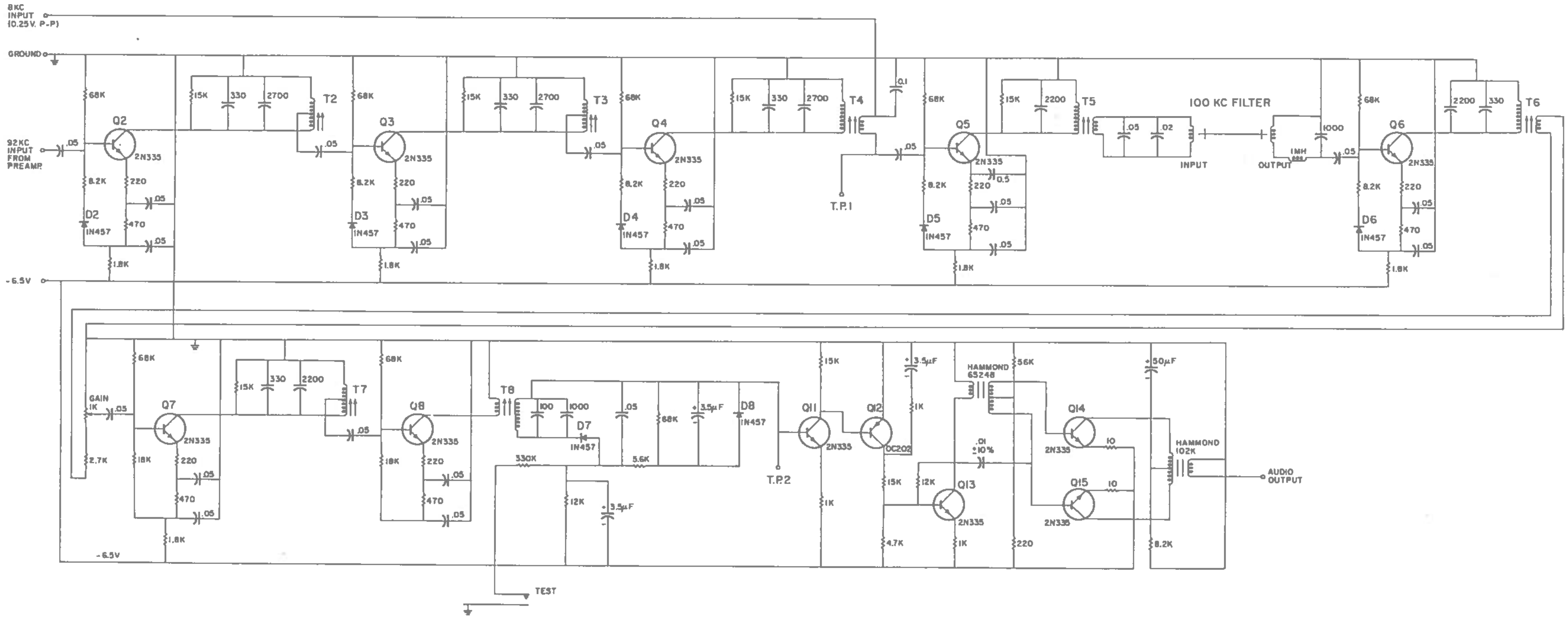
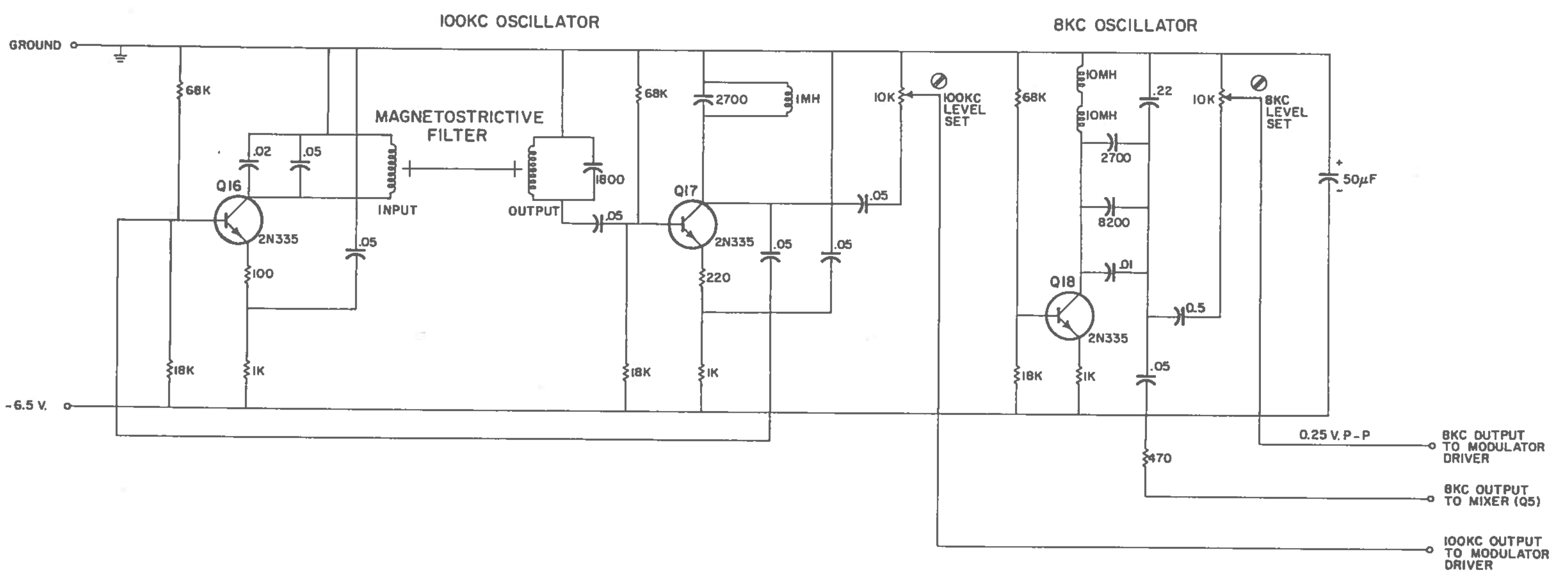


Fig. 10 Amplifier schematic diagram

SECRET



SECRET

Fig. 11 Schematic diagram of 8 kc/s and 100 kc/s oscillators

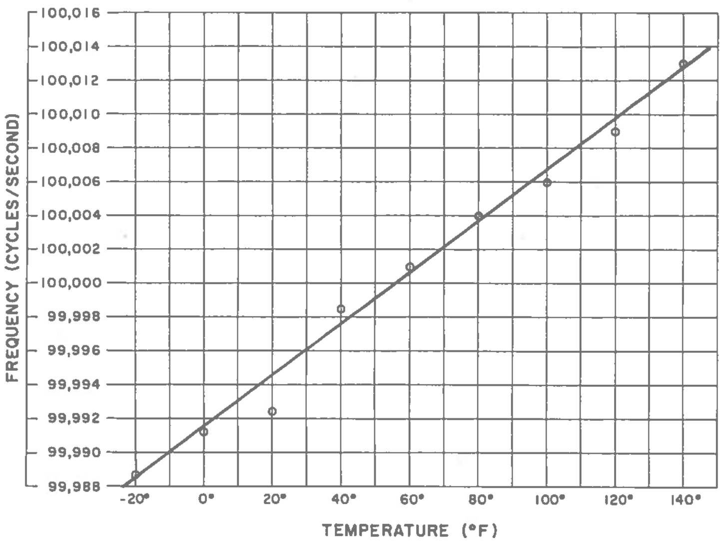


Fig. 12   Temperature drift of 100 kc/s oscillator

The frequency of the 8 kc/s oscillator is determined by the resonant circuit of Q18. The 2700 and 8200 mmfd capacitors are fixed trimming capacitors and were selected to tune the oscillator to within 20 c/s of 8 kc/s at 68°F.

The output of the 100 kc/s oscillator is coupled through a level-setting potentiometer to the 100 kc/s input of the modulator-driver. The output of the 8 kc/s oscillator is coupled through a level-setting potentiometer and an emitter follower (Q10) to the modulator-driver (Fig. 13), and through a limiting resistor to the mixer transistor (Q5) in the main amplifier circuit.

Modulator-Driver

The 92 kc/s modulation voltage which drives the varactor-modulator is produced in the collector circuit of the modulator-driver transistor (Q9). The transistor is driven by the 100 kc/s and 8 kc/s sinusoidal voltages, and the 92 kc/s component of the resultant collector current is selected by resonating the collector circuit to that frequency. The inductor which forms part of the resonant circuit is the primary winding of the modulator-driver output transformer. The secondary winding of the transformer is coupled to the varactor.

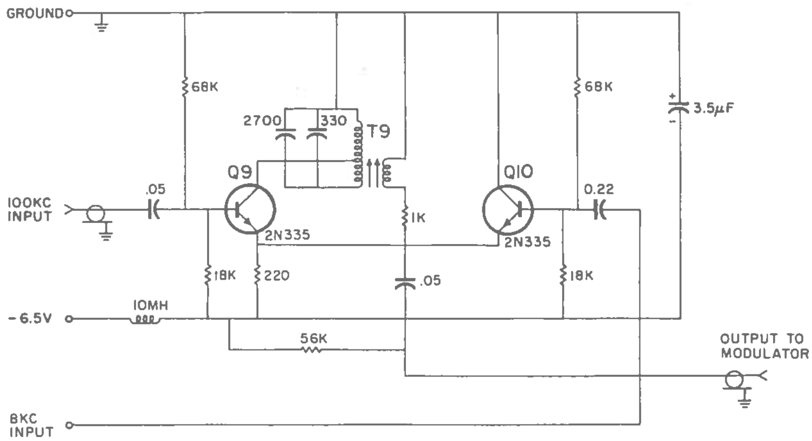


Fig. 13 Modulator-Driver schematic diagram

DISCUSSION OF CIRCUITRY

The simplest method of monitoring r-f energy in a wide band of frequencies is to use a wide-band antenna followed by a wide-band r-f detector and some warning indicator. The sensitivity of such a system is limited by the large low-frequency component of the video noise of the r-f detector, and by instability of the d-c amplifier or of any other circuit which is required to process the detected signal. A more sensitive method, and the one which was used in this intercept receiver, is to amplitude-modulate the incoming r-f signal, detect it, and amplify the modulation component of the detected signal in a narrow-band amplifier tuned to the modulation frequency. The amplified signal is then rectified and applied to the gated audio-frequency oscillator circuit. The sensitivity of such a system is limited by the cross-coupling between the modulator and the preamplifier circuits, and by the r-f detector noise in the passband of the amplifier. Cross-coupling between the modulator and the preamplifier is minimized by shielding. The noise level in the passband of the amplifier is minimized by making the bandwidth of the amplifier very narrow. However, reducing the bandwidth of the amplifier increases the response time of the system. As system response time is an important consideration, the choice of amplifier bandwidth requires a compromise between noise discrimination and system response time. Another consideration which affects both the noise level and the cross-coupling is the choice of modulation frequency. The spectrum of the r-f detector noise rises at low frequencies and introduces more noise into a low-frequency than into a high-frequency bandpass amplifier. The effects of cross-coupling, however, are greater at high than at low frequencies. The selection of a modulation frequency is thus a matter of choice between the relative importance of these two conflicting parameters.

Because the bandwidth of the amplifier is very narrow, the noise energy in its passband is very low, and the voltage-gain required for the gated audio oscillator to operate is very high; i.e., approximately 127 db. Superheterodyne techniques were used in the design of the intercept receiver to obtain such a high voltage gain in a stable amplifier system. The incoming signal is first amplified at 92 kc/s, then frequency-converted, and amplified at 100 kc/s. Instability due to cross-coupling between the high and low level amplifier stages was thus eliminated. The 100 kc/s frequency was chosen because of the availability of the narrow-band filters at that frequency. 100 kc/s was also the original choice of modulating frequency [1]; however, this frequency was lowered to 92 kc/s when difficulty was experienced in obtaining a stable 100 kc/s amplifier having the required over-all gain. 92 kc/s was chosen so that there would be no appreciable overlap of the two passbands involved (i.e., with the 100 kc/s amplifier passband), and so that the harmonics of the auxiliary frequency (8 kc/s) would lie outside the two passbands.

After preliminary tests under field conditions it was decided that temperature-compensation of the critical circuits was required. Three separate effects combined to make the amplifier gain temperature-sensitive. The first was the effect of a change in temperature on the gain of the transistors. The second was the variation of the transfer characteristics of the magnetostrictive filter with changes in temperature. These two effects were partly compensated for by inserting temperature-compensating diodes in some of the transistor bias circuits. The third was caused by gain variations due to detuning of the amplifier resonant circuits. This effect was reduced by increasing the bandwidth of the tuned circuits by resistive loading.

The efficiency of a low-level crystal detector can be improved by the application of a small amount of forward bias. The 100,000 ohm resistor which connects the crystal input of the preamplifier circuits (Fig. 9) to the negative battery supply line is included for this purpose. The applied bias improves the sensitivity of the detector by approximately 10 db. Similarly, an improvement in performance was noted when a forward bias was applied to the modulation varactor. However, the improvement consisted of a reduction in modulator drive required to obtain a given depth of modulation. This reduction in drive requirement also reduced the effects of cross-coupling between the modulator-driver and the preamplifier circuit. The 56,000 ohm varactor bias resistor is connected at the output of the modulator-driver circuit shown in Fig. 13.

A test circuit has been included in the receiver. The circuit includes a push button which, when depressed, adds an increment of voltage to the detected voltage already applied to the audio threshold circuit, and thereby causes the audio oscillator to operate on smaller noise signals. Depressing the button increases the percentage of time that the indicator is triggered by the noise, and can be used as a rough indication of the condition of the receiver. The test circuit does not test the performance of the modulator, the crystal, or the 100 kc/s oscillator. It does, however, indicate that all the other circuits are operating. A similar test can be performed by advancing the amplifier gain control until the audio oscillator is triggered by the noise. However, when the receiver is in use, this control requires careful setting and the inclusion of the test push button eliminates the need for disturbing this setting.

The audio level is not adjustable. It can, however, be changed by replacing a resistor in the collector circuit of the audio output amplifier with a resistor of larger or smaller value until the required audio output level is obtained.

PERFORMANCE OF CIRCUITS

The maximum available voltage gain of the amplifiers is 160 db. However, when the gain control and its associated 2700 ohm resistor are inserted in the circuit, as shown in Fig. 10, the amplifier gain is reduced to 144 db. The loss in gain is partly due to the shunting effect of the resistors, and partly due to their voltage divider configuration.

The over-all bandwidth of the amplifier is 6 c/s. However, the bandwidth of each of the stages which forms part of the amplifier with the exception of the narrow-band stage and the output detector stage, is 3 kc/s or greater. The bandwidth of the detector stage is 1 kc/s.

The voltage required to operate the threshold circuit varies from approximately 0.8 volt at -20°F to 0.55 volt at 140°F. This change in threshold sensitivity is due to a change in the internal properties of the threshold transistor itself.

SETUP PROCEDURE

The setup procedure for the C.W. Micradet Receiver can be broken down into two parts. The first consists of internal adjustments which are normally performed by technical personnel, and the second of an external adjustment which is performed by the operator.

The internal adjustments consist of adjusting the level of the 8 kc/s and 100 kc/s drive voltages to the modulator-driver. The following procedure is recommended for making these adjustments. The 8 kc/s and 100 kc/s level-setting potentiometers (Fig. 11) are first set to their minimum positions. The amplifier gain control is then adjusted so that an intermittent tone is heard in the headset. The 8 kc/s level-control is now adjusted to its half-way point, and the 100 kc/s drive voltage increased until a change in the intermittency of the tone is heard in the headset. The change is due to cross-coupling between the modulator and preamplifier circuits. The 100 kc/s drive voltage is now reduced slightly. A carrier frequency in the frequency range of the receiver is now fed into the antenna and the level of the carrier adjusted until the tone in the headset becomes nearly continuous. The level of the 8 kc/s drive and of the carrier are now readjusted until the receiver sensitivity is maximum. The 8 kc/s drive voltage is left at that level and the same procedure repeated for the 100 kc/s drive. These two adjustments are not very critical and can be made fairly rapidly.

The amplifier gain is the only adjustment that is under the operator's control. Optimum sensitivity is obtained when the gain of the receiver is so adjusted that amplified noise is just below the triggering threshold of the audio oscillator. Operation at this point may be obtained by first setting the gain control to yield an intermittent output in the absence of any incident radiation. The gain control is then backed off until the intermittent note just ceases. The push-to-test control produces a predetermined change in the triggering threshold, and may be used to assist in setting the correct gain. In the absence of any c.w. radiation, if the gain is properly set, an intermittent note will be heard when the push button is depressed, but no note will be heard when it is released.

DISCUSSION OF OTHER SYSTEMS TRIED

Some preliminary tests were made on a circuit configuration which used 100 kc/s noise obtained from the output stage of the narrow-band amplifier, to drive the modulator-driver. The reception of a c.w. signal by the antenna caused a 100 kc/s signal component to be added to the noise. As the received signal power increased, the 100 kc/s signal component became more and more predominant. The circuit behaved like an oscillator in which the r-f signal completed the feedback path. Owing to lack of time, the tests were brief and inconclusive, and since insufficient noise voltage was available for complete modulation, the sensitivity was less than that obtained by using the separate 100 kc/s oscillator. Replacing the 100 kc/s oscillator by the noise signal would eliminate the need for the frequency-stabilizer filter. Also, if the decrease in sensitivity that accompanies an increase in the bandwidth of the narrow-band amplifier can be tolerated, the remaining magnetostrictive filter could be replaced by a crystal. The decrease in sensitivity that can be expected with this change is approximately 3 db. As the magnetostrictive filters are much larger than a crystal, the reduction in size may be an acceptable substitute for the loss in sensitivity.

Another circuit variation that was tried consists of combining the 8 kc/s and 100 kc/s voltages directly in the varactor-modulator. The advantage of this variation is that the effects of cross-coupling between the modulator and the preamplifier are reduced owing to the elimination of the 92 kc/s modulation voltage. A reduction in the effects of cross-coupling may be an important consideration if an r-f crystal detector with lower noise is used. However, the efficiency of the modulator was reduced; hence this method of modulation was not adopted in the present receiver.

PERFORMANCE OF RECEIVER WITH C.W. RADIATION

The performance of the receiver is summarized in Figs. 14 and 15, which describe the variation of sensitivity with frequency and temperature, respectively.

The minimum detectable flux was determined using standard-gain antennas fed by signal generators to cover the frequency range 7.0 to 17.0 Gc/s. The transmitter and receiver were set up on an antenna pattern range which was essentially clear of obstructions so that conditions approximating free-space propagation were obtained.

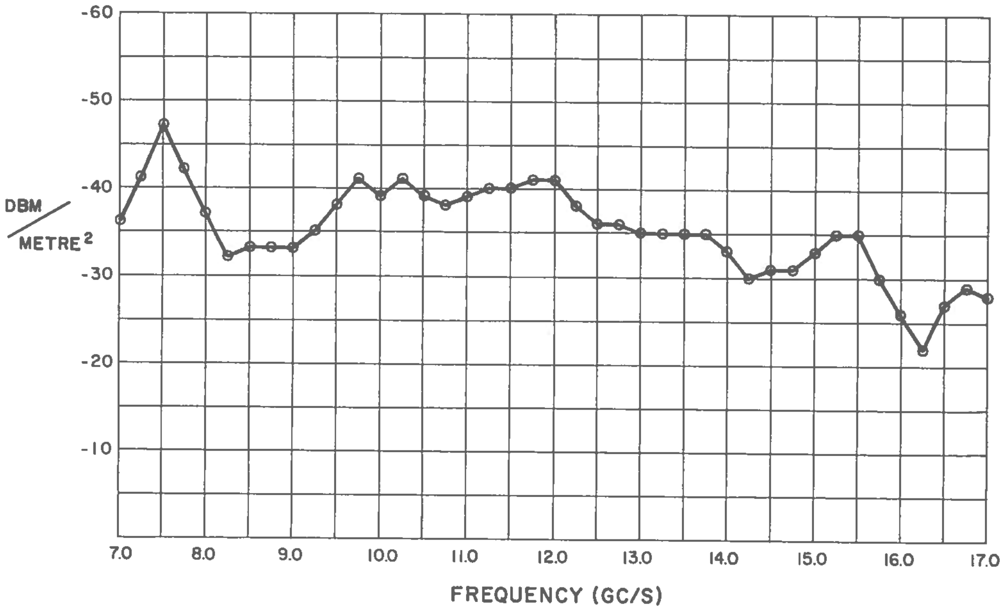


Fig. 14 Variation of receiver sensitivity with frequency

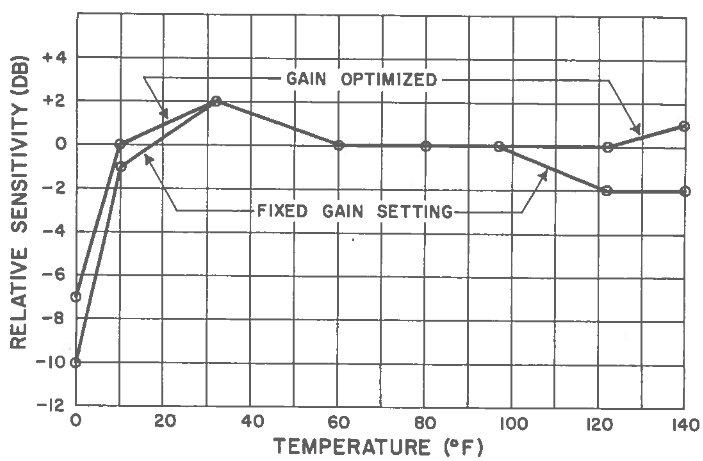


Fig. 15 Variation of receiver sensitivity with temperature

SECRET

The power delivered to the transmitting antenna was adjusted so as to yield a nearly continuous tone in the headset. (Such a level is some 2 db above the level at which some indication of signal is obtained; the repeatability of the readings was about $\pm\frac{1}{2}$ db.)

The results indicate that the desired sensitivity of -38 dbm/metre^2 has been achieved over most of X-band except for the low end, and that useful sensitivity has been achieved from 7.0 to 15.5 Gc/s. (Below 7 Gc/s the sensitivity falls very rapidly because cutoff in the ridged waveguide section is being approached, while between 17 and 20 Gc/s, where some readings were taken, the sensitivity fluctuates widely, but is at best much poorer than at 17 Gc/s.)

The sensitivity of the receiver at various ambient temperatures is illustrated in Fig. 15. There is little degradation (2 db maximum) in sensitivity from +10°F to +140°F; however, below +10°F the performance is degraded. It is expected that the degradation can be minimized by more elaborate temperature compensation of the equipment.

The other main technical characteristics of the receiver are:

Case size:	13.3 cm W. × 11.4 cm D. × 18.4 cm L
Weight:	1.9 Kg., including battery and headset
Power Supply:	Self-contained 6.5 V. mercury battery
Battery Life:	Estimated to be at least 100 hours at room temperature

PERFORMANCE OF RECEIVER WITH PULSED RADIATION

Although the receiver was constructed principally to provide an intercept capability against radiation which has no amplitude modulation, it also provides detection capability against pulsed radiation. To determine its performance when illuminated by such radiation, pulse trains of varying pulse widths and repetition frequencies were supplied to the receiver from a 15 Gc/s signal generator. Pulse widths from 0.5 to 100 μseconds, and pulse repetition frequencies from 1000 to 50,000 per second were used, as well as square waves of various frequencies.

The results obtained indicate that the receiver is, in most cases, responsive to the average power contained in the pulse train; i.e., the minimum detectable peak power level at a given frequency is larger than the corresponding minimum detectable c.w. power level by the factor $\frac{1}{Du}$, where Du is the duty cycle of the pulsed radiation. However, in the region of certain discrete p.r.f.'s which appear to be harmonically related to the internal frequencies of the receiver (e.g., 2 and 4 kc/s, being sub-harmonics of the 8 kc/s oscillator frequency), the response of the receiver is erratic, and it may even become completely inoperative at certain frequencies. In addition, trouble was experienced with overloading of the amplifier at low duty cycles. This can be overcome by lowering the gain of the 92 kc/s amplifier. A 1000 ohm resistor connected across the output of Q1 eliminates most of the overloading, while not impairing the c.w. capabilities of the receiver. A similar result could, no doubt, be obtained by locating the gain control nearer the input of the amplifier chain.

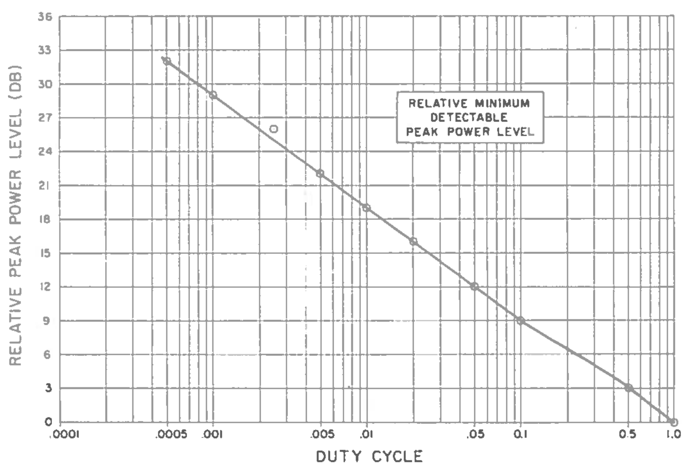


Fig. 16 Receiver performance with pulsed radiation

In Fig. 16 the performance is plotted as a function of the duty cycle. The points plotted in this figure were obtained by averaging a number of measurements taken with differing p.r.f.'s and pulse lengths to yield the same duty cycle. All of these measurements were taken at p.r.f.'s well removed from the critical values referred to above. Thus, provided that the pulse length lies within the range mentioned, and with the precaution described above, the minimum detectable flux in the case of a pulsed radar may be determined from Fig. 14, if the average rather than the peak power is used in the flux computation.

ACKNOWLEDGMENT

The authors are indebted to Messrs. D.T. Bradley, C.A. Epp, and G. Gratton for assistance in the construction of the receiver described in this report.

REFERENCES

1. Canadian Arsenals Limited. Technical proposal on a video receiver for detection of c.w. and pulsed radar transmissions. Ref. File DRBS 3803-27-1 (D. Phys. R.) July, 1963 (Secret)

2. J.K. Pulfer. A miniaturized crystal-video receiver. NRC Report ERB-577, January 1961 (Confidential)

3. A. Hendry. A broad-band c.w. intercept receiver. NRC Report ERB-643, August 1963 (Confidential)

4. A. Staniforth. A broad-band crystal mount, 10.5 kmc to 20 kmc. IRE Trans., MTT-8:464, 1960

Plate I — Exterior view of receiver

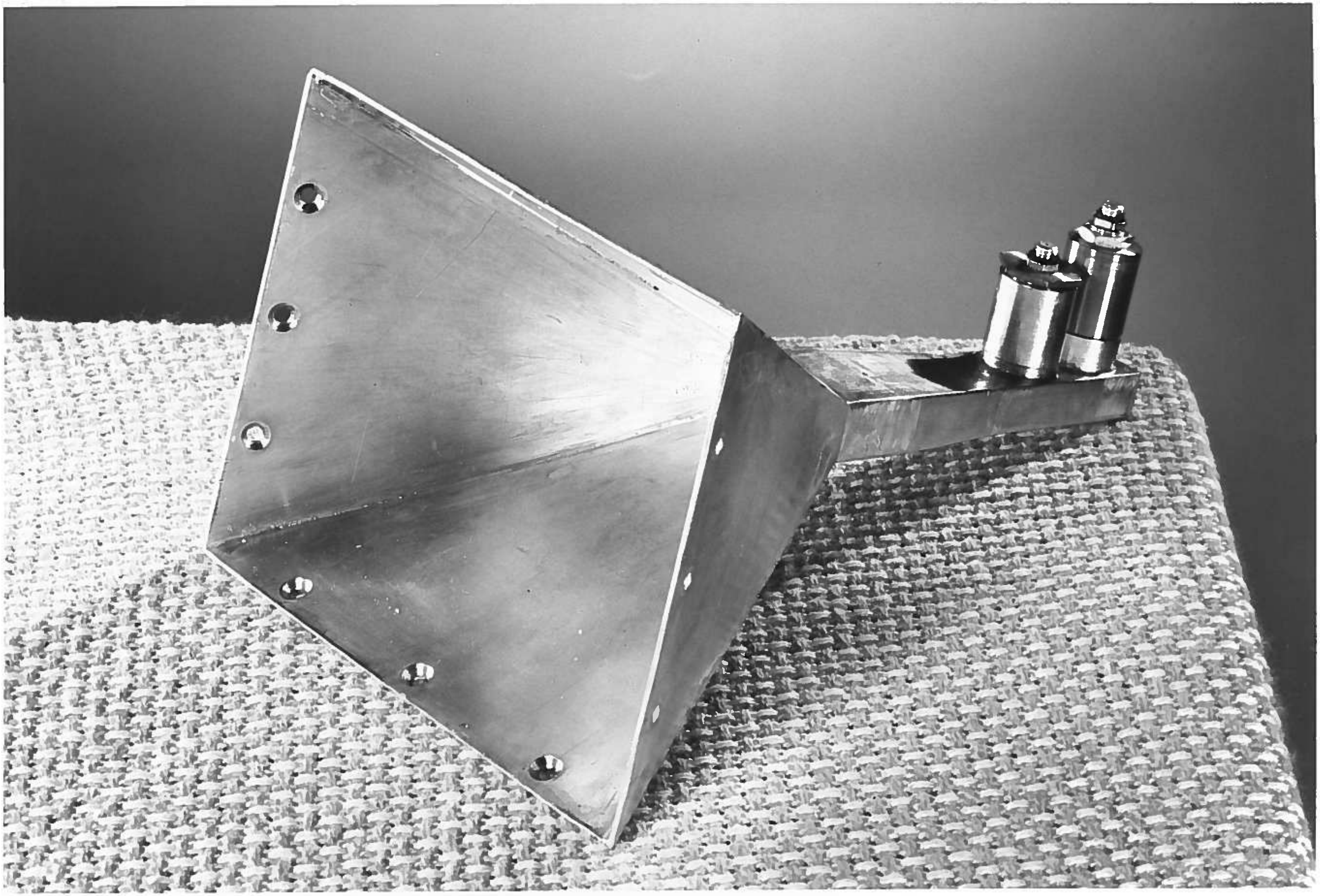


Plate II — Antenna, varactor mount, and crystal mount