[redacted]

REPORT NO. R-2509

DECLASSIFIED

NRL-54

REPORT NO. R-2509

DATE APRIL 24, 1945.

SUBJECT

DECLASSIFIED by NRL Contract Declassification Team
Date: 7 SEP 2016
Reviewer's name(s): A. THOMPSON, P. HANNA
Declassification authority: NAVY DECLASS MANUAL, 11 DEC 2012, 08 SERIES

REPORT ON

METHODS OF REDUCING THE MINIMUM OBSERVABLE

SIGNAL IN THE PRESENCE OF NOISE

DECLASSIFIED: By authority of
5000A January 1958
Entered by: E. Bliss Code 2027

FR-2509

NAVAL RESEARCH LABORATORY

BELLEVUE, D. C.

DISTRIBUTION STATEMENT A APPLIES
Further distribution authorized by UNLIMITED only.

U. S. GOVERNMENT PRINTING OFFICE 16—21906

DECLASSIFIED

320 - retain

[redacted]

RETURN TO RSR FILES
File under Folder 201

[redacted]

5330

24 April 1945 6 DECLASSIFIED NRL Report R-2509

373 6/4/45 NB R

NAVAL RESEARCH LABORATORY
WASHINGTON 20, D. C.

Radio Division - Consultant Group

METHODS OF REDUCING THE MINIMUM OBSERVABLE SIGNAL IN THE PRESENCE OF NOISE

by

O. D. Sledge

Approved by:

John M. Miller - Head, Consultant Group

A. Hoyt Taylor
Superintendent, Radio Division

A. H. Van Keuren, Rear Admiral, USN (Ret.)
Director, Naval Research Laboratory

Title Page	- 1 sheet (a)
Abstract	1 sheet (b)
Table of Contents	- 1 sheet (c)
Text	- 12 sheets
Plates	- 3 sheets

CLASSIFICATION CHANGED TO UNCLASSIFIED
BY AUTHORITY OF NRL Bibliography 3400
Reference Authority
ON Sept 1949
(DATE)
Jack M. Hayes
Signature of Custodian

UNCLASSIFIED

[redacted]

DECLASSIFIED

ABSTRACT

Three proposed methods of reducing the minimum observable signal are evaluated. It is found that two of the methods namely, the "stenode" receiver and the pre-detection "Type A" presentation, offer no appreciable reduction in the minimum observable echo of the conventional radar system.

Experiment indicates that homodyne detection, the third proposed method, offers from 5 to 15 decibels of reduction in the minimum observable echo of the conventional radar receiver. The amount of this reduction depends upon the i-f bandwidth of the receiver and the pulse width and repetition rate of the signal appearing at the input of the receiver. Unfortunately, the homodyne reduction in the minimum observable echo is found to be realizable only in a radar system operating on stationary targets.

TABLE OF CONTENTS

C[redacted]

1. INTRODUCTION

1-1. It has been known for some time that a linear detector discriminates against a weak modulated signal in the presence of a considerably stronger signal [1],[2]. This effect manifests itself as a reduction in the signal-to-noise ratio upon detection if the signal is weaker than the rms noise at the input of the detector. W. H. Jordan [3] has treated the problem theoretically and found that the suppression of the signal by the noise in a linear detector is given approximately by,

$$\text{DB Suppression} = 7 + 20 \operatorname{Log}_{10} \left(\frac{N}{S}\right) \qquad \left[\frac{N}{S} > 1\right] \tag{1}$$

where (N) and (S) are respectively the rms noise and rms signal at the detector input. It is observed that the loss in signal is a function only of the signal-to-noise ratio at the input of the detector. Also it is noticed that the suppression of the signal is quite appreciable, being thirteen decibels for a signal equal to one half of the rms noise.

1-2. The minimum observable signal (pulse) in the presence of noise is greatly effected by the suppression effect in the linear detector. Therefore any means by which the suppression of the signal in a linear detector may be reduced or eliminated will correspondingly reduce the minimum observable signal. It is the purpose of this study to evaluate several proposed methods of reducing the minimum observable signal in the presence of noise.

2. THE MINIMUM OBSERVABLE SIGNAL

2-1. From extensive measurements using a synthetic radar system, A.V. Haeff[4] arrived at the following empirical expression for the minimum observable signal (pulse) in the presence of fluctuation noise:

$$\text{Min. Obs. Signal (rms)} = knB^{\frac{1}{2}} \left(1 + \frac{1}{BT}\right)\left(\frac{1670}{r}\right)^{1/6} \tag{2}$$

where (k) is a constant of the receiver; (n) is the rms noise per unit bandwidth at the receiver input; (B) is the noise bandwidth of the predetection portion of the receiver; (T) is the pulse width of the input signal; (r) is the repetition rate of the signal pulse. Notice that the minimum observable signal is independent of the bandwidth of the amplifier following the detector.

2-2. For a fixed repetition rate and pulse width of the input signal the smallest value of the minimum observable signal obtains when the bandwidth (B) is equal to the reciprocal of the pulse width of the input signal. The optimum bandwidth is not at all critical, for Eq. (2) shows that the minimum observable signal increases only 1.2 decibels as the bandwidth is varied from its optimum value to either one third or to three times the optimum value

3. PRE-DECTECTION "TYPE A" PRESENTATION

3-1. By observing the output of the i-f amplifier of the receiver on a "Type A" indicator, the linear detector and its suppression effect are eliminated; and as a result one might expect to reduce the minimum observable signal obtainable with a conventional post-detection "Type A" presentation.

[redacted]

However, with a fixed amount of noise power at the output of the i-f amplifier of the receiver, it is found experimentally that the minimum observable signal is not appreciably altered when one changes from a post-detection to a pre-detection "Type A" presentation. Apparently the reduction in the minimum observable signal obtained by the elimination of the detector is approximately offset by an increase in the minimum observable signal which is probably a result of the reduction in the integration of the signal on the screen of the cathode-ray tube which accompanies the change from a post-detection to a pre-detection type of presentation.

4. THE STENODE RECEIVER

4-1. The percent of energy in noise which is transmitted through a band-pass network is proportional to the bandwidth of the network; on the other hand the percent energy of a pulse that is transmitted through such a network is not a linear function of the bandwidth. Fig. 1, Plate 1 shows the percent energy of rectangular and triangular shaped pulses, which is passed through an ideal band-pass filter, plotted as a function of the product of the half-amplitude pulse width and the bandwidth. The percent energy functions of Fig. 1 were derived by Dr. B. Salzberg of this Laboratory.

4-2. From the percent energy curves, it is noticed that for a given pulse width the percent energy transmitted is approximately a linear function of the bandwidth provided the bandwidth is less than one fourth the reciprocal of the pulse width. In radar practice, where a reasonably good pulse shape in the output of the i-f amplifier is required, the i-f bandwidth is made about twice the reciprocal of the pulse width of the input signal. Reference to the percent energy curves of Fig. 1 shows that by the use of such a receiver design the noise energy is increased relative to that of the signal. If the bandwidth is restricted to less than one fourth of the reciprocal of the pulse width, the signal-to-noise energy ratio at the input of the detector is approximately its maximum value. This condition suggests that the suppression of the signal-to-noise ratio in the detector of a conventional radar receiver may possibly be reduced by the use of a "stenode" type of receiver where the i-f bandwidth is less than one fourth of the reciprocal of the pulse width. The transmission characteristics of the i-f amplifier of the "stenode" receiver must be equalized in the low-pass amplifier following the detector in order to restore the overall transmission characteristics of the receiver to approximately those of a conventional radar receiver. The equalization of the phase as well as the amplitude characteristics of the i-f amplifier is necessary in order to keep the pulse shape and the resolution approximately those of a conventional radar system. A brief treatment of the method of equalization is given in Appendix I of this report.

4-3. A receiver of 100 kc i-f bandwidth and a complementary video (post-detector) amplifier which restored the overall bandwidth to 500 kc was compared with a receiver possessing a 500 kc i-f bandwidth and a video amplifier of 4 Mc bandwidth. With a fixed amount of noise power and a two microsecond pulse at the input of the two receivers, it was found that the minimum observable signal was approximately the same for the two receivers.

4-4. It is concluded that the "stenode" receiver offers no appreciable reduction in the minimum observable signal which is obtainable with a conventional receiver possessing a bandwidth equal to the reciprocal of the pulse width of the input signal and a flat video amplifier.

5. HOMODYNE DETECTION

5-1. The Origin and Definition of Homodyne Detection

5-1-1. It is well known that the envelope of the amplitude modulated wave appearing at the output of any frequency converter has approximately the same form as that of the modulated wave which drives the converter provided the heterodyne voltage is large in comparison with the input signal voltage. This statement holds true for any number of input signals if the signal frequencies all fall well within the pass-band of the converter circuit. As a result of this linear action of the frequency converter, the signal-to-noise ratio at the output is equal to that at the input of the device if the noise generated in the converter tube is negligible. This fact suggests that one might eliminate the reduction in the signal-to-noise ratio which occurs in a detector, when the signal is weaker than the noise, by heterodyning the signal to zero difference frequency which process might be called homodyne detection.

5-1-2. In Appendix II, it is shown that the reduction of the signal-to-noise ratio in the detection of signals weaker than the noise is eliminated in any general detector by the injection into the detector input of a homodyne voltage which is large relative to the noise voltage and which is in time phase with the signal carrier component.

5-2. Experimental Measurements and Results

5-2-1. In order to study the reduction in the minimum observable signal obtained by the introduction of a homodyne voltage into the input of a diode detector, the experimental arrangement of Fig. 2, Plate 2 was set up. In the course of the measurements, the following procedure was used. The noise level was adjusted to the point where the desired level of signal was discernible fifty percent of the time on the A-scope with no homodyne voltage applied to the detector. The level of the output of the signal generator was recorded in decibels. Leaving the noise level unchanged, a homodyne voltage which was about twenty times the rms noise voltage was introduced into the detector input circuit. The phase of the homodyne voltage was adjusted to give a maximum signal-to-noise indication on the A-scope. The signal level was then reduced until the pulse was again discernible fifty percent of the time in the noise. This reduction in signal level represented the reduction in the minimum discernible signal produced by changing from conventional diode detection to homodyne detection.

5-2-2. Measurements of the reduction of the minimum observable signal produced by the introduction of a homodyne voltage into the

[redacted]

input circuit of a diode detector yield the results shown in Fig. 3, Plate 3. It is seen from Fig. 3 that the homodyne reduction in the minimum observable signal increases with the pulse width and the repetition rate of the input signal if the i-f bandwidth is kept constant. The video bandwidth when varied from 0.02 to 2.0 megacycles results in only a small change in the homodyne reduction of the minimum observable signal. Experiment also shows that the signal level at the detector input has only a slight effect on the results.

5-3. The Application of the Homodyne Detection to Radar

5-3-1. Unfortunately the successful use of homodyne detection is difficult in radar practice because one is usually more interested in moving targets than in stationary ones. The echo from a moving target has a different carrier frequency from that of the incident pulse; this is known as the Doppler Effect. The carrier frequency of the reflected pulse is given by:

$$f = f_o \left[\frac{2V_T}{V_P} + 1 \right] \qquad (3)$$

where (V_T) is the component of the target velocity taken in the direction from the target to the radar antenna; (V_P) is the velocity of propagation of the incident wave; and (f_o) is the carrier frequency of the incident signal. The existence of this difference in frequency prevents the generation of a satisfactory homodyne voltage at the input of the detector of the radar receiver.

5-3-2. It might seem possible that an appreciable fraction of the homodyne reduction in the minimum observable signal could be obtained if the "Doppler" frequency difference were not too large. However, experiment shows that this is not the case; no discernible reduction in the minimum observable signal is obtained when the heterodyne (formerly homodyne) frequency differs from the carrier frequency of the echo by as little as ten cycles per second. Two precisely matched quartz crystals were used in this experiment to determine the frequencies of the simulated echo and the heterodyne voltage. Capacitance loading of one crystal produced the small difference frequency.

5-3-3. For a stationary target one might use a CW oscillator and a keyed amplifier as the radar transmitter, as was done in the experimental set-up of Fig. 2, in order to obtain true homodyne detection and the accompanying reduction in the minimum observable signal.

5-3-4. It has been suggested that one might derive the homodyne voltage from the echo itself by means of a highly selective filter such as the quartz crystal type which has its pass-band centered on the carrier frequency of the echo as it appears in the i-f amplifier of the receiver. This is not a satisfactory method of the generation of the homodyne power because noise frequency components as well as the carrier frequency of the echo appear at the output of the selective filter. Even in more optimistic

[redacted]

of arrangements with regard to pulse repetition rate, pulse width, i-f bandwidth, and filter bandwidth, the rms noise and the amplitude of the echo carrier frequency are of the same order of magnitude at the output of the selective filter. For example, suppose we have given that the i-f bandwidth is 200 Kc; the filter bandwidth is 100 cycles; the pulse width is ten microseconds; and the pulse repetition rate is 1000 pulses per second. Since the minimum observable signal amplitude and the rms value of the noise appearing at the detector input are of the same order of magnitude, we shall assume that at the filter input the pulse amplitude is equal to the rms value of the noise. The amplitude of the signal carrier frequency at the input of the filter is equal to the product of the pulse width, the repetition rate, and the pulse amplitude or one percent of the rms noise in this example. In passing through the selective filter, the signal carrier amplitude is assumed unattenuated; but the noise is reduced by a factor which is equal to the square root of the ratio of the filter bandwidth to the i-f bandwidth. In the example chosen the noise reduction factor of the filter is $\sqrt{5}/100$; hence the rms noise at the filter output is $\sqrt{5}$ times the amplitude of the signal carrier at the same point in the circuit.

5-4. Conclusions on Homodyne Detection

5-4-1. It is concluded that by the use of homodyne detection instead of conventional diode detection the minimum observable signal may be reduced by 5 to 15 decibels depending on the i-f bandwidth of the receiver and the pulse width and repetition rate of the signal input to the receiver. Practical considerations lead one to believe that homodyne detection in a radar receiver is a satisfactory means of reducing the minimum observable echo from a stationary target; however, homodyne detection cannot be applied successfully to a radar system operating on moving targets.

6. ACKNOWLEDGMENT

6-1. The methods of reducing the minimum observable signal discussed in this report were suggested by Dr. B. Salzberg of this Laboratory. The writer wishes to express his appreciation to Dr. B. Salzberg for valuable discussions concerning this research project.

[redacted]

APPENDIX I

1. EQUALIZATION OF BAND PASS RESPONSE CHARACTERISTICS WITH COMPLEMENTARY LOW PASS NETWORKS

1-1. The admittance of the simple band-pass network of Fig. 4 is:

$$Y = jC\omega + \frac{1}{jL\omega} + \frac{1}{R} \tag{1a}$$

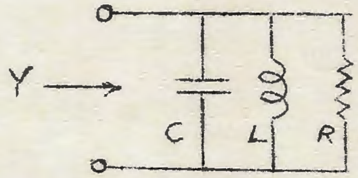

Fig. 4

The relative admittance of the network is:

$$YR = 1 + R\left(jC\omega + \frac{1}{jL\omega}\right) = 1 + j\frac{R}{L\omega}\left(LC\omega^2 - 1\right) \tag{1b}$$

Let $\omega_o^2 = \frac{1}{LC}$ then Eq. (1b) becomes:

$$YR = 1 + j\,RC\omega_o\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right) \tag{1c}$$

$$YR = 1 + j\,RC\omega_o\, d\,\frac{(d+2)}{(d+1)} \tag{1d}$$

$$YR = 1 + j\,RC\omega_o\, d\left[2 - d + d^2 - d^3 + \ldots\ldots\right] \tag{1}$$

where $d = \frac{\omega - \omega_o}{\omega_o}$

approximately,

$$YR \cong 1 + j\,RC\omega_o\, d\,(2 - d) \quad \text{if} \quad \left[d^2 \lll (2 - d)\right] \tag{1e}$$

Further approximation leads to the expression:

$$YR \cong 1 + j\,2RC\omega_o\, d = 1 + j2RC(\omega - \omega_o) \quad \text{if} \quad \left[d \ll 2\right] \tag{1f}$$

1-2. The relative admittance of the simple low-pass network of Fig. 5 is given by the expression,

[redacted]

$$YR = 1 + j\,RC\,\omega \tag{2}$$

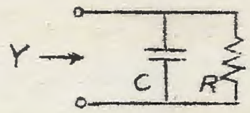

Fig. 5

1-3. Suppose these networks are used as the plate loads of pentode amplifier stages; then the value of (R) for the band-pass and low-pass stages should be equal if the same voltage gain per stage is to be had in the two types of amplifiers.

1-4. Assume that the excitation of the band-pass amplifier is an amplitude modulated wave with a carrier frequency equal to the resonant frequency ($\omega_o/2\pi$) of the band-pass network, and that the low-pass amplifier is driven by a voltage having the same wave-form as the amplitude envelope of the excitation of the band-pass amplifier. Further assume that the ratio of the highest modulation frequency to the carrier frequency is much less than two. Then it follows from Eq. (1f) and (2) that the transmission characteristics of the modulation frequencies are the same in the band-pass and low-pass stages provided the capacitance in the low-pass network is twice that in the band-pass network. In other words, the modulation wave undergoes the same frequency and phase distortion in passing through a stage of either the band-pass or the low-pass amplifier. This fact makes it possible to correct narrow band-pass networks with complementary low-pass networks on a one to one basis.

1-5. Let (Y_1R_1) be the relative admittance of the low-pass network which simulates the transmission effects of a narrow band-pass network on the modulation envelope. Further, let (Y_2R_2) be the relative admittance of the complementary low-pass network which is to equalize the response of the narrow band-pass network to a response equivalent to the response of a reference network possessing a relative admittance (Y_oR_o). It follows that

$$(Y_1R_1) \cdot (Y_2R_2) = Y_oR_o \tag{3}$$

Now if the low-pass networks are of the type depicted in Fig. 5, we have from (3) that

$$Y_2R_2 = \frac{1 + j\,R_oC_o\,\omega}{1 + j\,R_1C_1\,\omega} = \frac{1}{1 + j\,R_1C_1\,\omega} + \frac{j\,R_oC_o\,\omega}{1 + j\,R_1C_1\,\omega} \tag{4}$$

1-6. Consider the network of Fig. 6. The admittance of this network is:

[redacted]

$$Y_3 = \frac{1}{R_3 + j\,L_3\omega} + \frac{1}{R_4 + \frac{1}{j\,C_4\omega}} \tag{5}$$

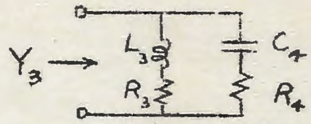

Fig. 6

The relative admittance of the network of Fig. 6 may be written in the form

$$Y_3R_3 = \frac{1}{1 + j\,\frac{L_3\,\omega}{R_3}} + \frac{j\,R_3C_4\,\omega}{1 + j\,R_4C_4\,\omega} \tag{6}$$

Comparison of Eqs. (4) and (6) shows that the network of Fig. 6 may be used as the complementary network provided that

$$Y_2R_2 = Y_3R_3 \tag{7}$$

or

$$\begin{bmatrix} R_2 = R_3 & R_3C_4 = R_oC_o \\ R_1C_1 = R_4C_4 = \frac{L_3}{R_3} & \end{bmatrix} \tag{8}$$

1-7. In the application of such a complementary network, unavoidable circuit capacitances will upset the correcting effect of the network unless the admittance of the complementary network (Y_2) is made larger than the admittance of the circuit capacitance over the frequency range of interest.

[redacted]

APPENDIX II

1. THE USUAL REDUCTION IN THE SIGNAL TO NOISE RATIO IN THE DETECTION OF SIGNALS WEAKER THAN NOISE IS ELIMINATED IN ANY GENERAL DETECTOR BY THE USE OF A HOMODYNE VOLTAGE LARGE RELATIVE TO THE NOISE.

1-1. Assume the detector current is given by the polynomial

$$i = \sum_{k=o}^{K} a_k e^k \tag{1}$$

where a_k = parameters of the detector and its associated circuits.
e = input voltage to the detector.
K = positive integer > 1.
$k = 0, 1, 2, 3, 4 - -$

1-2. Let the input to the detector be:

$$e = S(t) \cos \theta + H \cos (\theta - \phi) + N \cos \left[\theta - (\delta + \phi)\right] \tag{2}$$

$$e = S \cos\theta + H \cos\theta \cos\phi + H \sin\theta \sin\phi + N \cos\theta \cos(\delta + \phi)$$
$$+ N \sin\theta \sin(\delta + \phi) \tag{2a}$$

$$e = \left\{\left[S + H \cos\phi + N \cos(\delta + \phi)\right]^2 + \left[H \sin\phi + N \sin(\delta + \phi)\right]^2\right\}^{\frac{1}{2}}$$
$$\cos (\theta - \psi) \tag{2b}$$

$$e = H\left[\left[\frac{S}{H}\right]^2 + 1 + \left[\frac{N}{H}\right]^2 + 2 \frac{S}{H} \cos\phi + 2 \frac{SN}{H^2} \cos(\phi + \delta) + 2 \frac{N}{H} \cos\delta\right]^{\frac{1}{2}}$$
$$\cos (\theta - \psi) \tag{2c}$$

$$e = E \cos (\theta - \psi) \tag{3}$$

where

$$\psi = \tan^{-1}\left[\frac{H \sin\phi + N \sin(\delta + \phi)}{S + H \cos\phi + N \cos(\delta + \phi)}\right] \tag{3a}$$

$$E = H\left[1 + \left[\frac{S}{H}\right]^2 + \left[\frac{N}{H}\right]^2 + 2 \frac{S}{H} \cos\phi + 2 \frac{SN}{H^2} \cos(\phi + \delta) + 2 \frac{N}{H} \cos\delta\right]^{\frac{1}{2}} \tag{3b}$$

H = constant amplitude of the homodyne voltage
$S(t)$ = amplitude envelope function of the input signal
$N(t)$ = amplitude envelope function of the input noise
$\delta(t)$ = phase function of the input r-f noise
$\theta = 2\pi$ times the product of the signal carrier frequency and time
ϕ = constant phase displacement between the signal and the homodyne voltage

[redacted]

1-3. The detector current is from Eqs. (1) and (3)

$$i = \sum_{k=0}^{K} a_k E^k \cos^k (\theta - \psi) \tag{4}$$

Contribution to the detector video output is obtained only from the terms of Eq. (4) in which (k) is an even integer; hence we may write the expression for the detector output voltage in the form,

$$e_d = \sum_{m=1}^{M} a_{2m} A_{2m} E^{2m} \tag{5}$$

where A_{2m} is the constant term of the finite Fourier series representing the function $\cos^{2m} (\theta - \psi)$ and M is a positive integer equal to or greater than unity. Substitute the expression for the amplitude envelope (E) of the input from Eq. (3b) into Eq. (5) and obtain:

$$e_d = \sum_{m=1}^{M} a_{2m} A_{2m} H^{2m} \left[1 + \left[\frac{S}{H}\right]^2 + \left[\frac{N}{H}\right]^2 + 2\frac{S}{H}\cos\phi + 2\frac{SN}{H^2}\cos(\phi + \delta) + 2\frac{N}{H}\cos\delta \right]^m \tag{6a}$$

$$e_d \cong \sum_{m=1}^{M} a_{2m} A_{2m} H^{2m} \left[1 + 2\frac{S}{H}\cos\phi + 2\frac{N}{H}\cos\delta \right]^m \text{ if } \left[\begin{array}{l} \left[\frac{N}{H}\right]^2 \ll 1 \\ \left[\frac{S}{H}\right]^2 \ll 1 \end{array} \right] \tag{6b}$$

Further approximation yields:

$$e_d = \sum_{m=1}^{M} a_{2m} A_{2m} H^{2m} \left[1 + 2m\left(\frac{S}{H}\cos\phi + \frac{N}{H}\cos\delta \right) \right] \text{ if } \left[\frac{(m-1)(N+S)}{H} \ll 1 \right] \tag{6}$$

From inspection of Eq. (6), we see that the signal-to-noise ratio of the detector output is given by:

$$\frac{\text{Signal}}{\text{Noise}} = \frac{S \cos\phi}{N \cos\delta} \tag{7}$$

From Eq. (7) it is seen immediately that in order to obtain a maximum signal-to-noise ratio in the output of the detector, the homodyne must be either in phase or out of phase with the input signal; i.e., $|\cos\phi| = 1$. Eq. (7) shows further that with proper homodyne phasing relative to the signal the detector output signal-to-noise ratio is greater than the input signal-to-noise ratio (S/N).

[redacted]

REFERENCES

(1) R. T. Beatty, Apparent Demodulation of a Weak Signal by a Stronger One. "Experimental Wireless and Wireless Engineering", Volume 5, page 300, June 1928.

(2) F. E. Terman, Apparent Selectivity Obtained by Linear Detection, "Radio Engineering", page 319, McGraw-Hill Book Company, Inc., New York.

(3) W. H. Jordan, "Action of Linear Detector on Signals in the Presence of Noise", MIT Report No. 61-23.

(4) A. V. Haeff, "Special Report on Minimum Detectable Radar Signal and Its Dependence upon Parameters of Radar Systems", NRL Report RA 3A 208A.

[redacted]

DECLASSIFIED

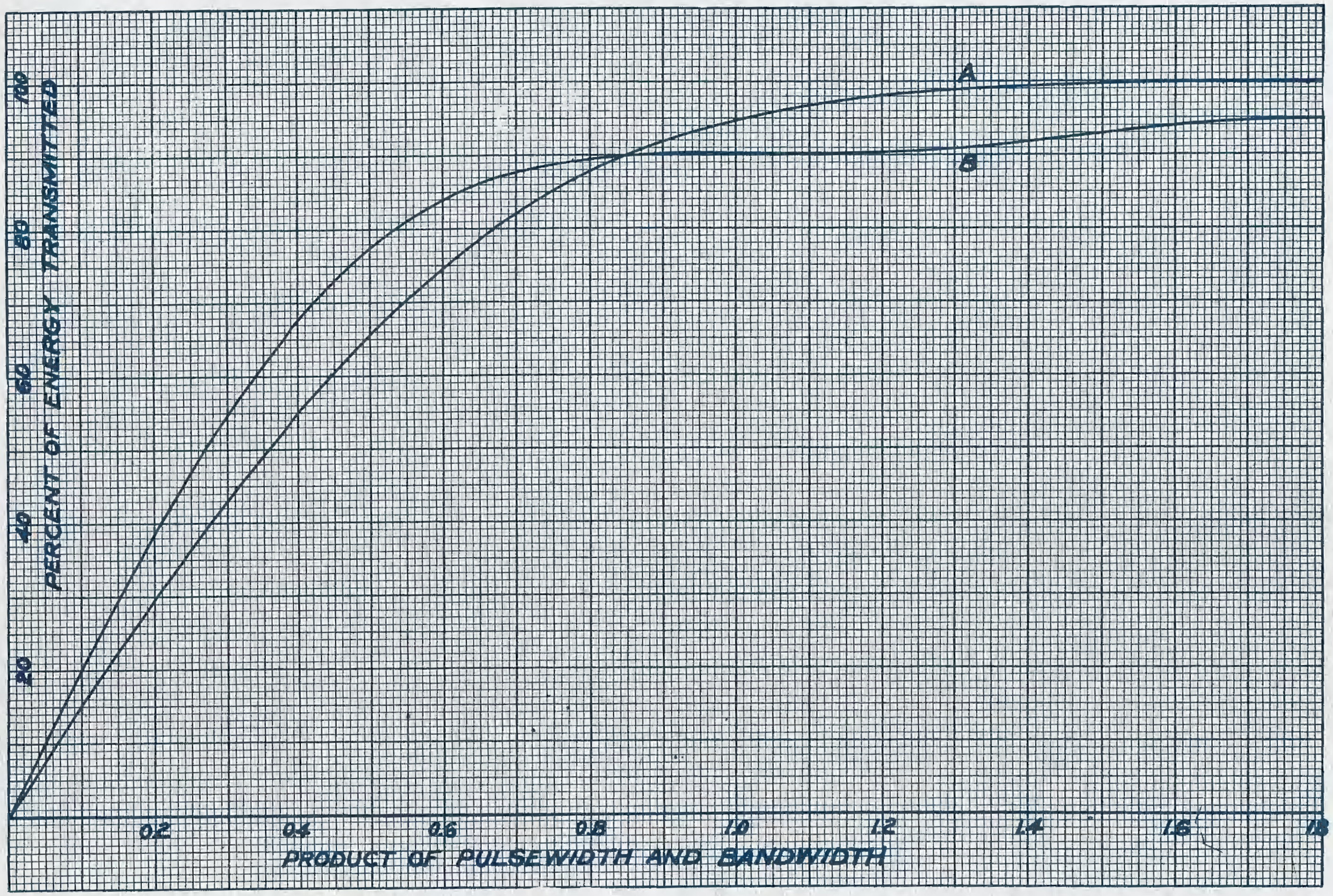

FIGURE 1 — PERCENT OF ENERGY OF PULSE TRANSMITTED THROUGH IDEAL BAND-PASS FILTER

A — TRIANGULAR INPUT PULSE   B — RECTANGULAR INPUT PULSE

PULSEWIDTH MEASURED AT ONE HALF AMPLITUDE

PLATE 1

DECLASSIFIED

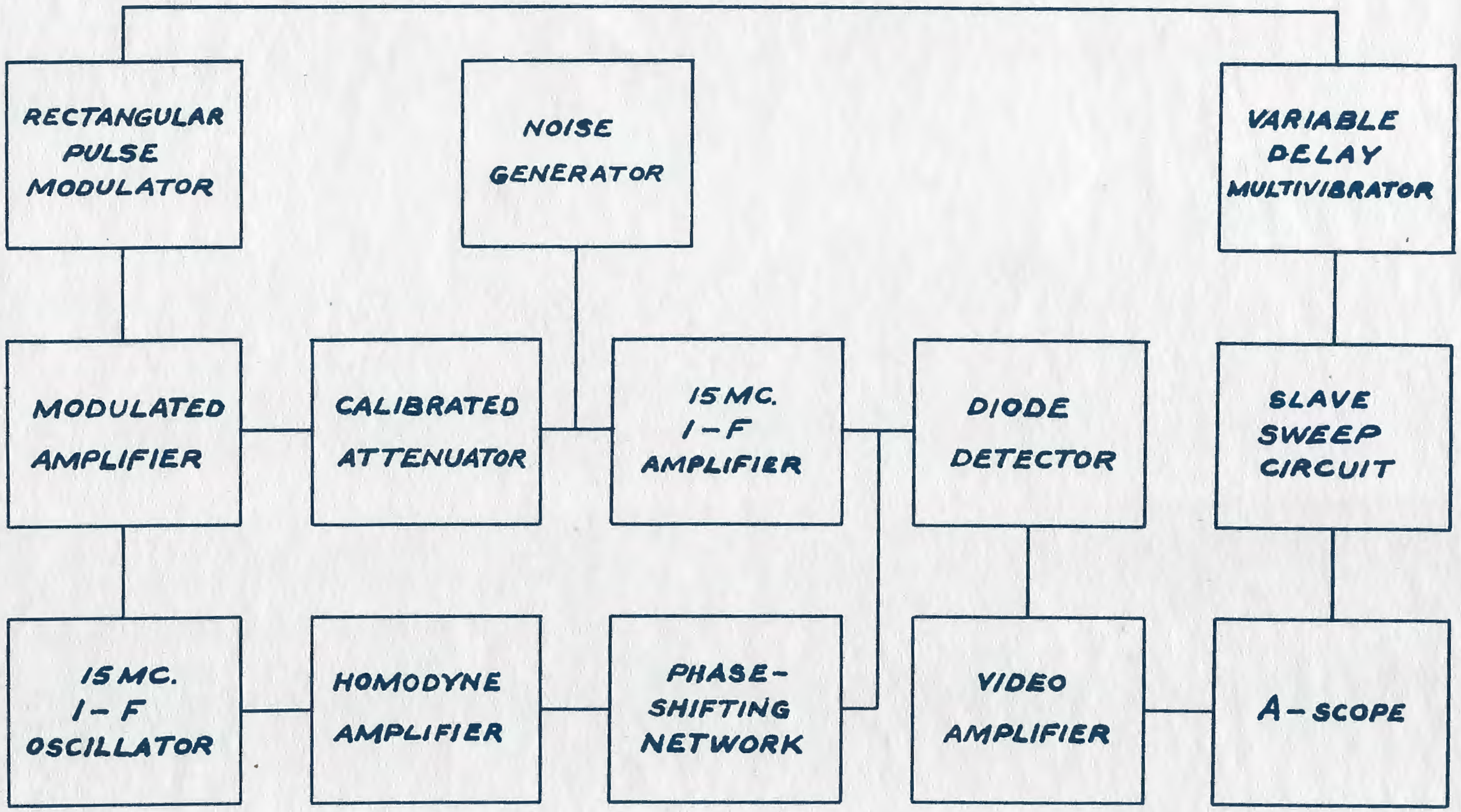

FIGURE 2 — THE EXPERIMENTAL ARRANGEMENT FOR MEASUREMENT OF THE REDUCTION IN THE MINIMUM OBSERVABLE SIGNAL PRODUCED BY THE CONVERSION FROM DIODE TO HOMODYNE DETECTION

PLATE 2

DECLASSIFIED

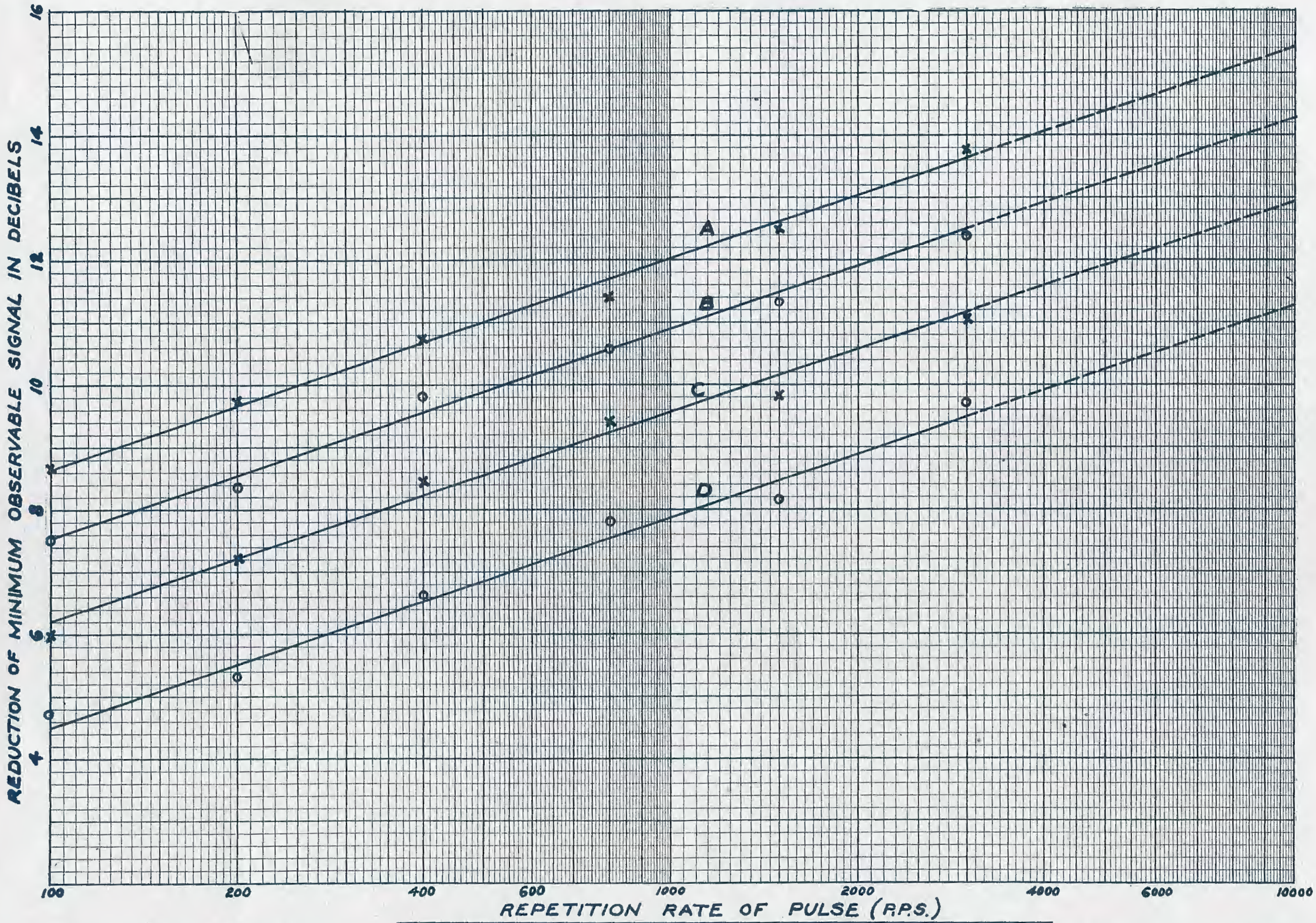

FIGURE 3 —THE REDUCTION OF THE MINIMUM OBSERVABLE SIGNAL DUE TO HOMODYNE DETECTION

I-F BANDWIDTH = 0.30 MC. RANGE SWEEP = 25 NAUTICAL MILES

PULSEWIDTH AT INPUT OF I-F AMPLIFIER

A — 20 μ-SEC. B — 10 μ-SEC. C — 5 μ-SEC. D — 1.7 μ-SEC.

DECLASSIFIED