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MASSACHUSETTS INSTITUTE OF TECHNOLOGY

LINCOLN LABORATORY

QUARTERLY PROGRESS REPORT

DIVISION 3 - COMMUNICATIONS AND COMPONENTS

15 April 1953

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CAMBRIDGE MASSACHUSETTS

FOREWORD

This is the ninth report on the classified program of research on Communications and Components which was started in the Research Laboratory of Electronics, Massachusetts Institute of Technology, in the late Fall of 1950, and is now being continued by Division 3 of the Lincoln Laboratory.

Further progress in the development and production of audible radar and specialized VHF scatter-communication facilities for PROJECT COUNTER CHANGE is reported in the sections describing the work of Groups 31 and 36.

The work reported herein was completed during the period from 1 January 1953 through 31 March 1953.

W. H. Radford
Division Head

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PROCESSING AND PRESENTATION OF INFORMATION

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PROCESSING AND PRESENTATION OF INFORMATION

GROUP 31

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I. INTRODUCTION

Group 31 is developing techniques and equipment directed toward improving the processing and display of information in air defense systems. In recent months, the activity has been concentrated on two major programs: the development of equipment for a radar alerting network in the far North; and a study of system organization directed toward optimizing the processing of data presented by the Land display.

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II. DEVELOPMENT OF RADAR EQUIPMENT FOR A DISTANT EARLY-WARNING LINE

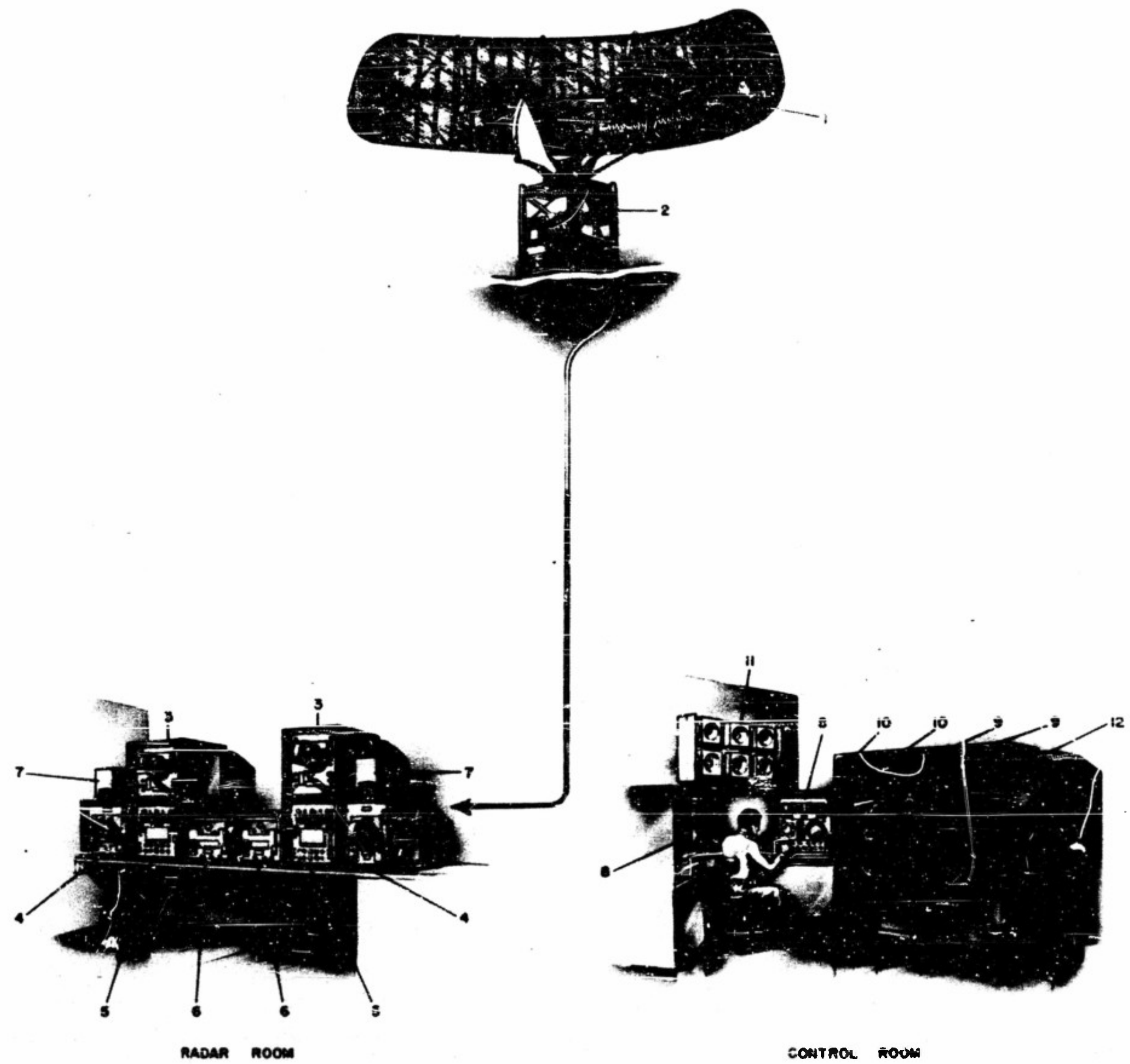
A. Arctic Alerting Radar

The Quarterly Progress Report of 15 January 1953 contained a discussion of the work directed toward supplying alerting equipment for an experimental distant early-warning (DEW) line. This work has continued at an increasing pace. At present, the principal endeavors are production engineering, system checkout, and flight tests. The system is now designated as Automatic Alerting Radar X-1.

Figure 31-1 is an artist's conception of a typical alerting station. Two complete radars and alarm systems (Radalarms) are to be installed at each northern site to insure continuous operation. Figures 31-2 and 31-3 show the arrangement of the components in the Radalarm units. Each unit contains six range gates and automatic alarms, which provide for monitoring six adjustable annular rings centered at the radar. The basic Radalarm components are installed in two 7-foot relay cabinets.

1. Status

Because of the "crash" nature of the development and production program (the program is only now four months old), compromises have been made in system performance and component fabrication. The completed production units, however, appear to be sufficiently



1. ANTENNA AS-348/TPS-1D
2. ANTENNA BASE AB-221/TPS-1D
3. RADAR RECEIVER-TRANSMITTER RT-212/TPS-1D (MOD A)
4. RADAR MODULATOR MD-144/TPS-1D (MOD A)
5. POWER SUPPLY PP-674/TPS-1D (MOD A)
6. SIGNAL COMPARATOR CM-36/TPS-1D (MOD A)
7. LOCAL OSCILLATOR U 501A
8. AZIMUTH RANGE INDICATOR ID-141/TPS-1D (MOD A)
9. RANGE GATE CABINET X-1
10. AUDIO-ALARM CABINET X-1
11. LOUD SPEAKER PANEL X-1
12. TEST EQUIPMENT X-1

Fig.31-1. Typical station arrangement (Automatic Alerting Radar X-1).

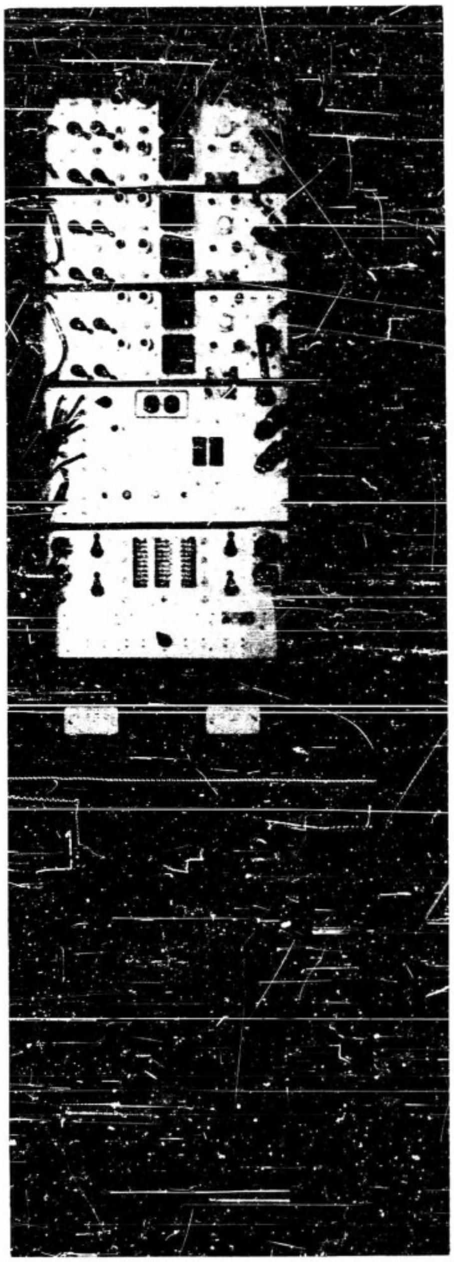


Fig.31-2. Range gate cabinet, X-1.

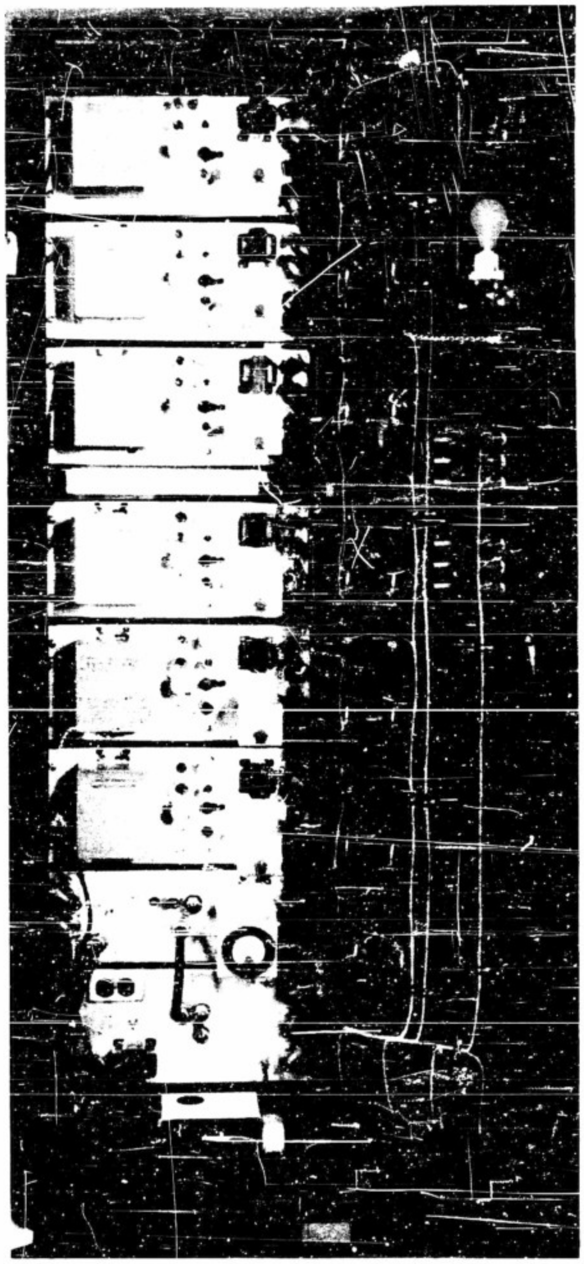


Fig.31-3. Audio alarm cabinet, X-1.

GROUP 31

carefully constructed to withstand shipping and usage under the conditions likely to be encountered in the northern sites. Lincoln Laboratory is providing enough spare parts and components to permit continuous operation of the equipment for a period of approximately one year.

Two modified AN/TPS-1D radars were shipped to the Western Electric Company test sites in Illinois early in March, and six additional modified radars will be shipped about 1 April. Two Radalarm X-1 systems will be shipped about 10 April. The six remaining Radalarm X-1 systems are scheduled for delivery to the Western Electric Company on 1 May 1953.

2. Flight Test

The experimental alerting system at Building 22, M.I.T., has been superseded by a more finished installation at the Lexington Field Station. A flight-test program, for component shakedown as well as system-performance data, was initiated approximately 6 weeks ago at the Field Station. Flight-test data indicated that the AN/TPS-1D stalo did not have sufficient stability to eliminate spurious responses from strong fixed echoes at long ranges. Laboratory tests indicated that the instability was due to both mechanical and electrical inadequacies in the design of the stalo. The difficulties were eliminated by additional filtering in the power supply, by reworking of the stalo cavity to improve mechanical tolerances, and by placing the unit in an external shock-mounted box.

The flight-test program is now evaluating the coverage of the automatic-alerting radar system. The calculated performance of the equipment being shipped for DEW installations is shown in Fig. 31-4. Because of the necessity for eliminating signal components at multiples of the repetition rate, the system is not sensitive to Doppler velocities that correspond to frequencies near 400 and 800 cycles per second. The sensitivity of the system as a function of Doppler frequency is shown in Fig. 31-5. The blind velocities can be determined from this graph; the Doppler shift at L-band is approximately 3.8 cycles per mile per hour of radial target velocity.

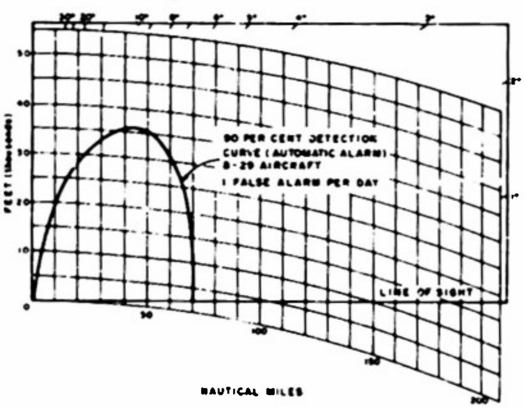


Fig. 31-4. Calculated coverage of Automatic Alerting Radar X-1 over poor soil.

3. Instruction Book

A radar instruction book, covering the radar modifications and the operation of the Radalarm system, has been prepared by Jackson and Moreland, Engineers, under subcontract, and will accompany the equipment into the field.

H. G. Weiss
J. Freedman
R. F. Meyer
F. R. Naka
R. M. R. Olivier
M. J. Levine
M. Winick

4. Investigation of the Performance of the Automatic-Alarm Components

Empirical tests are now being made to determine the effect of varying the threshold-to-noise ratio and signal-to-noise ratio on the false-alarm rate and the target-detection probability of the automatic alarm circuits. At present, simulated signal and noise inputs are being employed.

Supplementary measurements will be made shortly when two narrow-band filter banks are added to the system. One of the filter banks was constructed in the laboratory. The other was obtained from Motorola, Inc. The latter consists of 40 tuned vibrating reeds ("vibrasponders"), each with an amplitude-sensitive contact. The set of reeds covers the same band of frequencies as the laboratory filter bank, i.e., 400 to 600 cps.

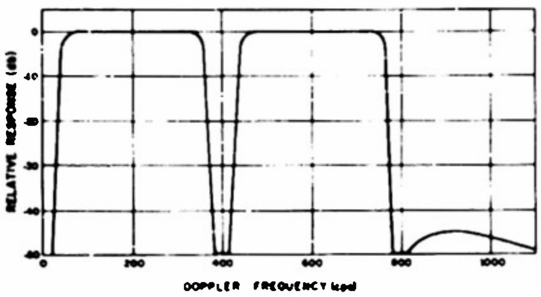


Fig.31-5. Filter characteristic (automatic alerting system).

A block diagram of the measurement equipment is shown in Fig. 31-6. The variable-level limiter shown in the block diagram is employed in the determination of the necessary dynamic range of the amplifiers that drive the alarm circuits. Dynamic range is especially important when narrow-band filters are employed, for in that case only a small portion of the input power spectrum passes through each filter.

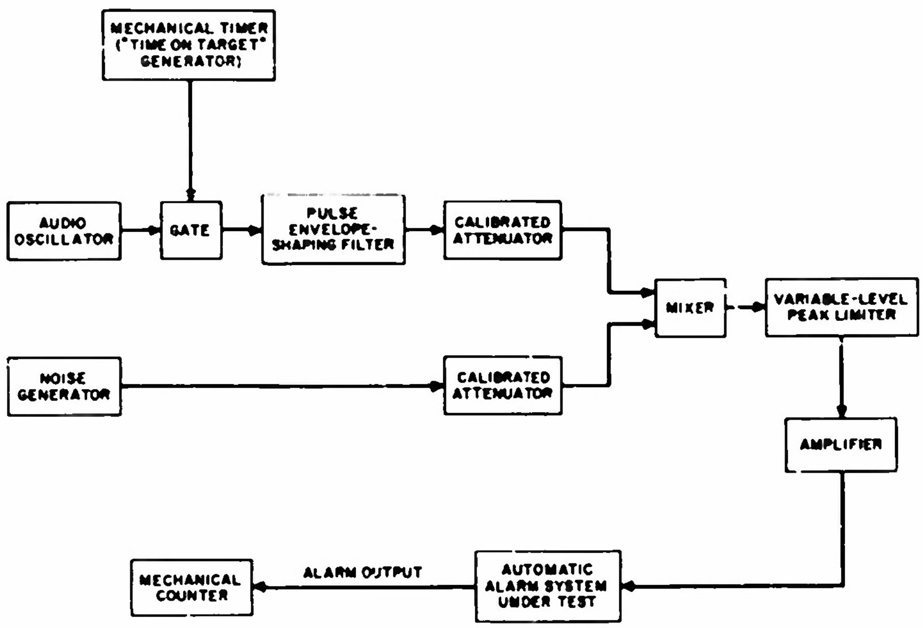


Fig.31-6. Block diagram of automatic alarm system measursment apparatus.

The bank of 20 narrow-band filters built in the laboratory is now ready for testing. A preamplifier and power amplifier have been constructed to drive the filters, and a simple relay circuit has been added as an alarm detector. The alarm threshold level is adjusted by varying the gain of the preamplifier.

Since the last Quarterly Progress Report, several changes have been made in the circuit of the filter bank. In place of the diode matrix, the arrangement shown in Fig. 31-7 has been employed for peak amplitude detection. The use of the set of cathode followers shown in

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Fig. 31-7 avoids completely the loading effect of the diode matrix on the tuned circuits. The reduced loading, in turn, makes it possible to use capacitors with higher dissipation factors and thereby to dispense with the polystyrene condensers that, we became convinced, were too bulky for use in the filter bank.

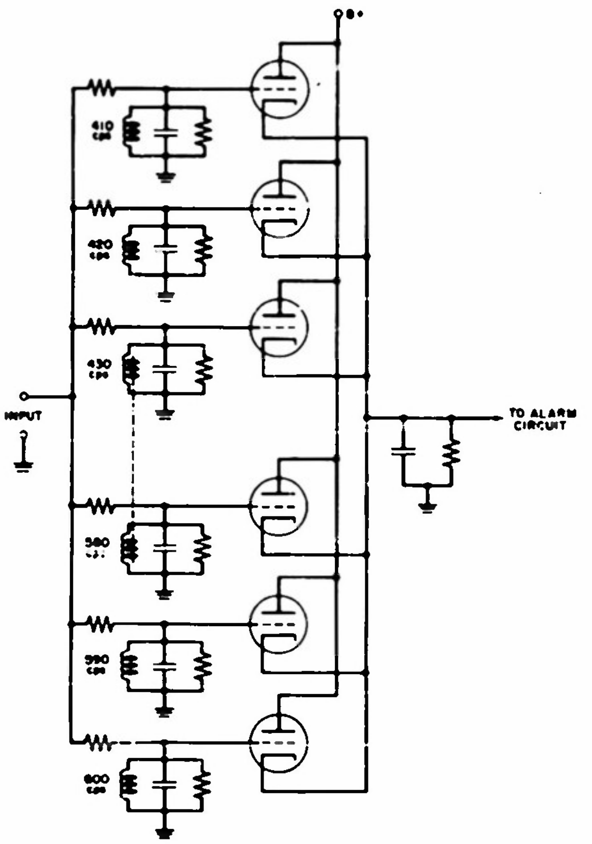

Fig.31-7. Narrow-band filter bank.

R. P. Sallen
E. L. Key

B. <u>CW Fence Radar</u>

The CW Fence Radar (also called Fluttar or Type F Radar) consists of a transmitter and receiver separated by approximately 30 miles. It operates in the range 400 to 600 Mc/sec. On reflection from a moving aircraft, the signal, shifted in frequency (Doppler effect), mixes at the receiver with the direct signal from the transmitter to produce an amplitude-modulated signal. The side bands pass through selective filters and activate an alarm. A description of this system is included in the Lincoln Laboratory Quarterly Progress Report of Division 3, dated 15 January 1953.

Because intensive field tests of this system will be conducted shortly by the Bell Telephone Laboratories and the Western Electric Company, our Fluttar activity will be limited in scope.

A Fluttar link, operating on 408 Mc/sec, was established between M.I.T. and Rockport, Mass., a 32-mile baseline, in December 1952. Although the actual strength of the signal at the receiving location was considerably less than anticipated, it was possible to detect aircraft passing through the link. It was established that the percentage modulation was over 50 in most cases when the aircraft was directly over the baseline. The low signal-to-noise ratio, and the high noise level at the receiving location, made quantitative tests difficult.

Because of the generally unsatisfactory results, it was decided to explore the possibility of changing the location of the receiver site to a more favorable environment. A site situated in the Blue Hills in Canton, Mass., and presently under the control of AFCRC, was selected. The Blue Hills site has an elevation of 685 feet. Two portable antennas, each having 11 db gain, were used to measure the signal strength. An adequate signal level of 40 microvolts was measured at this site. Flight tests have not yet been resumed.

Alternate sites, with possible baselines of 30 to 36 miles, are being investigated north of Boston. It is anticipated that one such site will be in operation during the latter part of May.

E. A. Slanina
E. Caldwell
H. G. Weiss

GROUP 31

C. Listening Performance of Human Operators

Work continues on two possible means of improving detectability of a signal representing a target in one of the range gates of the audible radar.

In the Quarterly Progress Report of Division 3, dated 15 January 1953, it was reported that spatial separation of the loudspeakers fed individually by the several range gates of the radar failed to produce significant improvement in detectability of 500-cycle signals. Tests have now been completed with signals of various frequencies. The previous findings hold throughout the audio range. Further tests point to sound reflection within the listening room as the disruptive factor. It would probably not be reasonable, however, to construct anechoic listening rooms for the audible radars; one of the main hopes for aural radar presentation is eventual simplicity of equipment.

Means for presenting the signals from various range gates in separate frequency bands have been devised and are ready for testing. In addition, laboratory setups are being prepared for studies of:

(1) The effects of variations of antenna rotation rate and gate length;

(2) The relative efficiencies of automatic alarms and human listeners working with identical signals;

(3) The effect of frequency modulation of an acoustic signal on its detectability.

J. F. Bennett
W. P. Harris

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III. INFORMATION-PROCESSING SYSTEMS

Analysis of the data obtained from the series of experiments on information processing systems described in previous Quarterly Progress Reports has been continued. The principal functions of the intelligence section of an Air Defense Direction Center – detection, identification, tracking, and cross-telling – have been evaluated. Further analyses of the latencies and durations of these operations, of information-flow characteristics, of the decision processes, etc., are being made. A 16-mm sound motion picture has been prepared to illustrate the organization and operation of both information processing systems.

James W. Degan
Bert F. Green, Jr.
William J. McGill

IV. DESIGN AND EVALUATION OF DISPLAYS

A. Airborne Photographic Target-Position Indicator

Photographs of the display of an AN/APS-20A ground-stabilized PPI (AN/APA-81) were made on a flight of 2 hours' duration in a P2V aircraft borrowed specifically for the purpose from an operating squadron at the Naval Air Station, Quonset Point, Rhode Island. The exposure plan of these pictures was such as to produce an exact simulation of the Land display.

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Photographs were taken of two or three scans per exposure. The 75-mile range position was used throughout the tests. The positive film, printed from the developed negative, was projected in a Stereo Vivid Projector (dual lens but less Polaroid filters) with Wratten filters Types A and B. Several frames were thereby superimposed, half of them in red and half in green, and the stabilized blips from stationary reflectors appeared yellow.

Study of the results showed that several improvements could be made: the use of shorter ranges would make moving targets more distinct; exposure should be reduced; background light, which was unexpected and detrimental, should be reduced; and the film-advance technique should be improved. Nevertheless, the results showed promise; moving-target indication by color (red and green separated for moving targets) was noted over water; and integration of very slow-moving boats in light sea clutter produced yellow blips.

When the Navy P2V aircraft assigned to Lincoln Laboratory is equipped with ground-stabilization gear, additional flight tests will be made.

E. Caldwell

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B. Flying-Spot Scanner

The flying-spot scanner is a device that provides simulated radar signals from positive film transparencies.

Efforts have been continued to obtain photographs of suitable quality for use in the scanner. With care, it is possible to obtain from a standard plan-twelve indicator photographs that yield simulations useful in all but the most critical applications.

With photographs obtained from a special indicator-camera setup, it appears to be possible to obtain simulations that are very close to "live" radar. Construction of a special setup can be avoided by modifying the scanner to permit its use for photography as well as for reproduction. The modifications are to be incorporated in the production model of the scanner.

Consideration has been given to the use of feedback around the scanner to control picture quality. The feedback is from the output of the video stage to the intensity grid of the cathode-ray tube. Analysis shows that useful control of the photographic Gamma and noise can be gained. Experimental work will wait upon the delivery of the first scanner.

The laboratory model of the scanner is on loan to the Naval Research Laboratory where it is to be used as a source of radar signals in a system operational evaluation program.

T. Randell

C. Polar-Coordinate Grids

In the Quarterly Progress Report of Division 3, dated 1 October 1952, a study was reported concerning the speed and accuracy of reading the positions of targets using polar-coordinate grids of various designs. A second study is now in progress. Five grids, all

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variants of a basic grid with twelve azimuth lines at 30° intervals and three range rings at 50-mile intervals, are being compared.

B. F. Green, Jr.
Lois Anderson

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D. Data Links from ADDC's to ADCC's

At the request of Group 21 of the Lincoln Laboratory, we have started an experiment to determine the relative merits of telefax transmission and voice-telling for the purpose of reporting data from an Air Defense Direction Center (ADDC) to an Air Defense Control Center (ADCC). Telefax presents a picture on a scope that is viewed by a plotter at a large vertical plotting board. The plotter looks at the scope and plots the positions of aircraft on the plotting board. No voice communication is involved in this system. In the alternative system, a teller at the ADDC voice-tells the aircraft positions to the plotter at the ADCC. Both these systems have been simulated in the laboratory. Data are being obtained on the number of tracks that can be handled adequately under each system.

B. F. Green, Jr.
Edna Wilson

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V. CAPABILITIES AND LIMITATIONS OF THE HUMAN OPERATOR

A. Low-Probability Watch

We have continued work on the low-probability watch along the lines described in the Quarterly Progress Report of Division 3, dated 15 January 1953. Two additional experiments have been run, and a more detailed analysis of the previous data has been made.

The recent experiments demonstrate clearly that the rate at which signals are presented has an important effect on the percentage of signals detected. Figure 31-8 shows the percentage of signals detected at three average rates of presentation: 7.5, 60, and 480 signals per hour.

In the earlier experiments, we confirmed Mackworth's finding* that more signals are detected if rest periods are introduced to break up a long watch. We became interested in the effects of frequent brief interruptions. Figure 31-9 shows the marked improvement that resulted when 90-minute watches were interrupted for 30 seconds every 5 minutes. During each 30-second interruption, the observer's display was turned off. Turning off the display 10 per cent of the time increased the detection of signals in the last hour of the test by almost 20 per cent.

*N. H. Mackworth, Researches on the Measurement of Human Performance, Privy Council, Medical Research Council, Special Report Series No. 268, H.M.S.O. (London), 1950.

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In analyzing the data from the several experiments, we have been particularly interested in finding out what factors govern the incidence of false reports. The following obser-

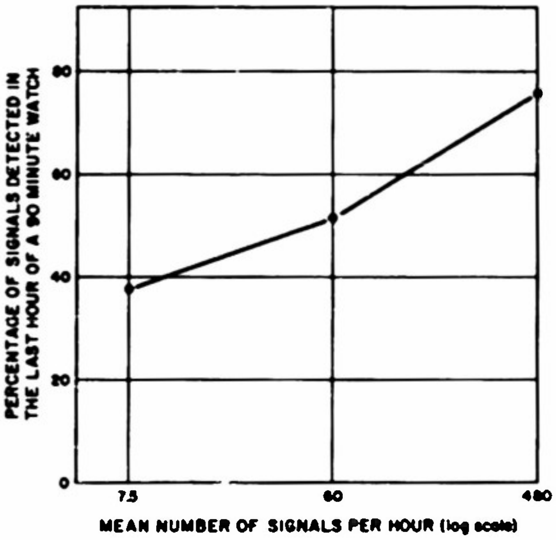

Fig.31-8. Detection as a function of the mean rate at which signals are presented.

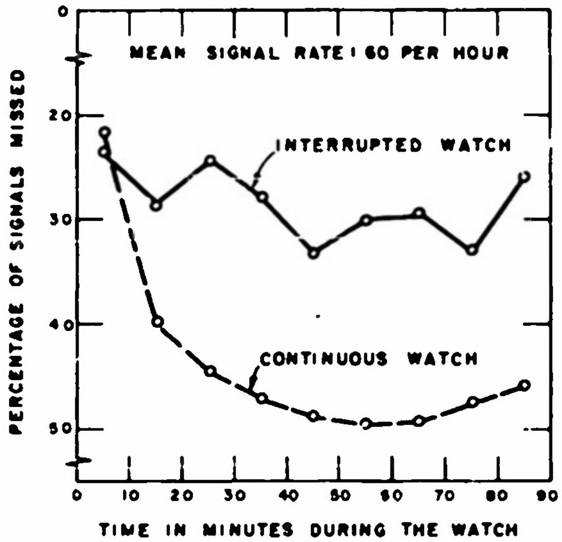

Fig.31-9. The effect of frequent brief interruptions on signal detection in a 90-minute watch.

vation seems to be a step in the desired direction: *the observer's average latency of response to lights presented at irregular intervals during a watch is closely related not only to the percentage of signals correctly reported but also to the incidence of false reports.* As the observer becomes less alert, his reaction time increases, the number of correct reports decreases, and the number of false reports increases. The observer continues to report, but the correlation between the reports and the signals decreases.

In this first phase of the investigation, the intent has been to survey rather quickly the effects of several experimental conditions in order to outline the general problem. A report of the first-phase findings is in preparation.

H. M. Jenkins

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B. Analysis of Sequential Effects in Auditory Judgments

When responses occur one after another in sequence, earlier parts of the response sequence may affect the parts that come later. This research was designed to explore the application of a new analytical technique to the sequential effects that occur in a series of auditory judgments.

A listener learned to identify pulses of tones of four frequencies, 890, 925, 970, and 1005 cycles per second, presented one at a time in random order. After the listener had learned to make accurate identifications, a masking noise was turned on, and he attempted to identify a long sequence of tone pulses heard against the noise background. These identifications were

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analyzed by an application of "multivariate transmitted information."* Multivariate transmitted information breaks down into a set of transmission components that have the properties of Shannon's measure of transmitted information. Thus "transmission" from the tone pulses to the listener's identifications, and "transmission" from previous responses to the listener's identifications could be measured simultaneously. It was possible to compute a pair of weights that showed the average effects of the stimulus and of the last identification in determining any given identification.

Results showed that the listener's identifications started to become uncorrelated with the auditory stimulus at a signal-to-noise ratio considerably higher than that at which the tone pulses disappeared into the noise. As the information transmitted from the tone pulses to the identifications decreased, there was a small but consistent increase in the information transmitted from the last response. When the tone pulses disappeared into the noise, practically all the transmitted information came from the last response, but the effect was not large.

W. J. McGill

*See Technical Report No. 38 of the Air Force Human Factors Operations Research Laboratories.

RADAR TECHNIQUES

GROUP 32

RADAR TECHNIQUES
GROUP 32

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Schneider, W. P.

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I. PULSED-DOPPLER RADAR

All major components of the pulsed-Doppler radar, with the exception of the indicating system, have been constructed and have been given preliminary tests. Details of assembly and associated problems are under consideration at the present time; comprehensive tests of the major components are in progress.

The RF section is complete and has been operating successfully for several weeks. To facilitate flight tests, clutter targets within range of the radar are being located and plotted on a map of the surrounding area.

W. P. Schneider

A. Audio-Frequency Section

Audio sections of three types have been completed. They comprise: (1) a wide-band unit with limiting amplifier; (2) a narrow-band unit with limiting amplifier, and (3) a wide-band unit with no amplifier. The first two of these were described in the 15 January 1953 Quarterly Progress Report of Division 3; the third is identical with the first except for the omission of the amplifier. The purpose of the third type is to permit a study of the signals without the modifications introduced by the limiting amplifiers.

The characteristics of the clutter filter and the 3000-cps low-pass filter are shown in Figs. 32-1 and 32-2 respectively. These curves are in satisfactory agreement with the

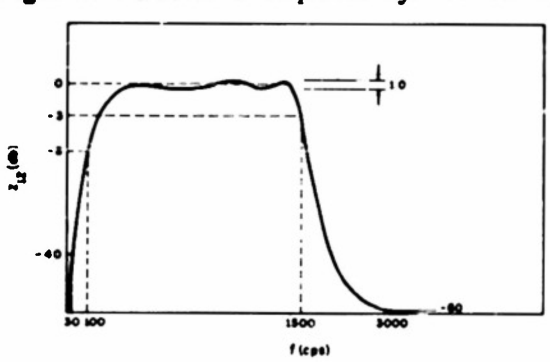

Fig. 32-1. Clutter filter.

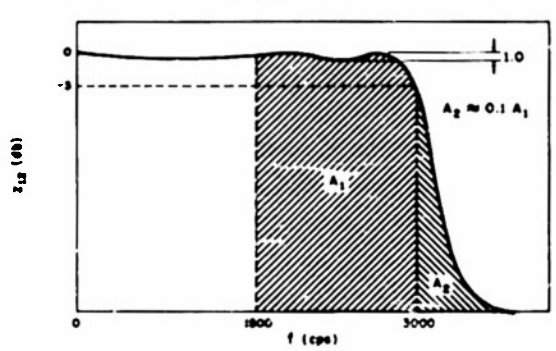

Fig. 32-2. Low-pass filter.

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theoretical characteristics.

Experimental studies to determine quantitatively the performance of the audio units are in progress. Since the system is nonlinear, signal-to-noise measurements do not provide complete information; hence, direct experimental measurements of such quantities as the relations among the probability of detection, false-alarm interval, clutter-to-signal ratio, and signal-to-noise ratio are planned.

P. J. Caruso, Jr.

B. Doppler-Filter Reed Banks

The development of the oscillator employing a reed as the frequency-determining element, described in the last Quarterly Progress Report, has been temporarily suspended. Work has proceeded essentially along two lines. Magnetic damping (with eddy-current loss) is being investigated theoretically and experimentally. This method of damping looks promising although it may involve an intricate mechanical arrangement. Work is continuing on the experimental side, and a new idea, which will reduce the detuning, will be tried.

Three banks of one-cycle bandwidth reeds are being constructed. These reeds measure about $0.002 \times 1/16 \times 3/8$ inch, resonate about 400 cps, have bent tips for visual output, and are tuned by end loading with solder. Magnetic coupling between reeds has been greatly reduced by using a stainless steel which is only slightly magnetic.

Richard K. Bennett

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II. NOISE RADAR

This quarter has been devoted to building the noise-generating and receiving parts of the radar system. Referring to the block diagram (Fig. 32-3), the status of each component is as follows.

The phase-shifting noise source and narrow-band random noise source (described in earlier reports) have been built and tested. The third random-noise generator shown may be set for any one of the indicated bandwidths. It has been built and will be tested shortly.

Similarly, the gated mixer, limiter, timer, TR box, 100-kcps bandwidth receiver, and multipliers have been built and are being tested.

A commercially available reed frequency meter has been modified to permit its use as a frequency mixer as well as a Doppler indicator. These changes also increased the sensitivity of the meter. The construction and preliminary testing of a prototype power amplifier for driving the reed bank have been completed. This amplifier has a pass band from 0.5 cps to over 1000 cps and requires a special transformer. Construction of three filter-amplifier units incorporating this power amplifier will begin very shortly.

The three IF amplifiers shown are being constructed.

A prototype delay-line amplifier has been built. Construction of the complete set of amplifiers will start soon in anticipation of delivery in about two weeks of the quartz delay units.

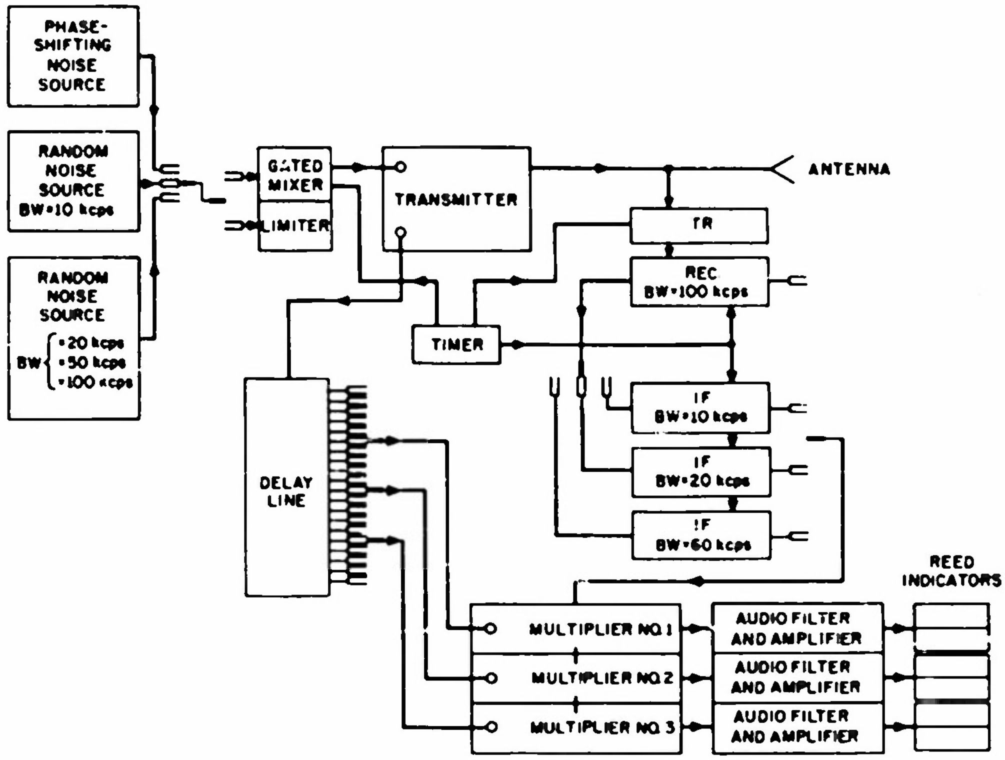

Fig. 32-3. Block diagram of noise radar.

It is planned to conduct the first experiments with three multiplier-filter-reed channels as shown in Fig. 32-3. The three channels may be connected to any of the range taps available on the delay line. The multipliers are to be of the limiting type described and discussed in the last Quarterly Progress Report.

L. G. Kraft
Philip Fire
E. M. Howlett

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III. TEST EQUIPMENT

A. Spectrum Analyzer

As an aid to investigating the output of the pulse-Doppler radar, a specialized

GROUP 32

spectrum analyzer has been designed and built. The primary feature of the equipment is a high-Q single-tuned LC filter having essentially the same band-pass characteristics as the mechanical reed filter. A simplified block diagram of the analyzer is shown in Fig. 32-4. The unit contains two filters which may be tuned independently, with the half-power bandwidth of each being continuously variable from 10 cps to 30 cps. The filters may be connected directly to the input for the purpose of investigating sidebands about 10 kcps or to the output of the suppressed carrier modulator when the input spectrum falls about DC. The output of the filters is indicated on both peak and rms type meters. The peak meter circuit is such that the peak output of either filter or the difference of the peak outputs can be read. The rms meter is a thermocouple vacuum-tube voltmeter which has an external input for use as a separate meter for nonsinusoidal waveforms (i.e., noise, pulses, etc.).

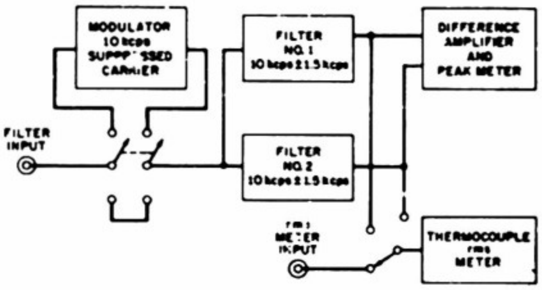

Fig. 32-4. Spectrum analyzer block diagram.

The analyzer is now being calibrated and tested.

Hayes Penfield

B. Doppler-Signal Simulators

Modifications of the Type C delay line* for use as a Doppler-signal simulator are continuing. The crystal mounting has been redesigned and built. The testing of this unit awaits completion of a pulsed oscillator which is under construction.

Philip Fire

IV. THEORETICAL STUDIES

A. The Probability of False Alarm

Let $u(t)$, $-\infty < t < +\infty$, be a stochastic process with the following properties:

(1) Temporally stationary, (that is, the probabilities defining $u(t)$ are invariant under translation of t);

(2) Markoffian, (that is, the future depends upon the present, but is independent of the past);

(3) For $t_1 \neq t_2$, $[u(t_1), u(t_2)]$ has a nonsingular two-dimensional Gaussian distribution.

Then the only physically reasonable $u(t)$ has correlation function $\sigma_o^2 \exp[-\beta|t|]$, where $\beta > o$ and

$$\sigma_o^2 = E\left(\left\{u(t) - E[u(t)]\right\}^2\right) \quad ,$$

which is independent of t by (1) and for $t_1 < t_2 < \ldots < t_n$, $[u(t_1), u(t_2), \ldots, u(t_n)]$ is distributed according to an n-dimensional Gaussian law with common mean, $m = E[u(t)]$, and common variance, σ_o^2.

Conditions (1), (2), (3) allow $u(t)$ as a model for random voltages. It is of interest in questions about false alarms to know the distribution of the maximum of $u(t)$, that is,

*H. B. Huntington, *Components Handbook*, M.I.T. Radiation Laboratory Series, Vol. 17 (McGraw-Hill, Inc., New York, 1949), pp. 228-30.

$$\text{Prob}\left\{\begin{matrix}\text{Max } u(\tau) < c \\ 0 \leq \tau \leq t\end{matrix}\right\} .$$

The determination of this distribution has been reduced to solving an ordinary, second-order differential equation of Sturm-Liouville type on $(-\infty, \infty)$. This differential equation is being studied.

E. J. Akutowicz

B. Dead-Velocity Control

The analysis of the method (see 1 October 1952 Quarterly Progress Report of Division 3) for the control of the dead-velocity situation in radar systems by transmission of pulses spaced alternately by times T_1, T_2, where $T_1 < T_2$, is now complete. The success of this method depends on choosing carefully the first two $n_{critical}$ values, $C_1 = m + p$, and $C_2 = 3(m + p)$. Since the positive odd integer p turns out to be the factor by which the first blind speed in the ordinary single-pulse system is multiplied (provided the criterion mentioned below for choosing C_1 is used), we wish to choose p as large as possible. However, if p is chosen too large, zeros from the sinusoidal envelope factor will lie between C_1 and C_2, and the improvement obtained will depend not on p but on the location of the first such zero. Hence, the improvement may be given by a factor that is much less than p. To avoid this sort of trouble, it is pointed out that we should choose C_1 so that $C_1 << s_1/3$, where s_1 is the first sinusoidal zero. Even if this criterion is used, the question of the amplitudes of the spectral lines that will be passed by the filter is one that should be considered in each individual case. We have worked out a number of such examples.

Now consider the positive integer m. The larger m, and hence $T = T_1 + T_2$, is made, the greater will be the density of spectral lines. Thus, there is the possibility that, if we make m too large, certain of these spectral lines may lie quite close to the noncritical zeros. In general, it is best to make m as small as possible (although this is not an absolute criterion). Again, several examples have been worked out that show the effect of different values for m.

A further consideration regarding the practicability of this system is the following. The suppression of the $(m + p)^{th}$ line depends on maintaining the ratio $(T_1/T) = (1/2k)$ at a precise value. Several examples have been completed. The conclusion from these is that the drift question is no real obstacle.

Detailed analyses will be made of other proposals for control of the dead-velocity situation.

A. F. Bartholomay

LONG-RANGE COMMUNICATIONS

GROUP 33

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LONG-RANGE COMMUNICATIONS

GROUP 33

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Granlund, J.
Gray, D.J. (USAF)
Gurske, V.P.
Johnson, A.H.
Kelly, E.J.
Lacey, G.G.
Lephakis, A.J.
Loewenthal, M.
Mentzer, J.R.
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Morrow, W.E., Jr.
Portmann, P.A.
Quinn, W.T., Jr.
Rasmussen, T.S.
Reisener, W.C., Jr.
Ring, R.M.
Smith, E.A.
Stone, M.L.
Stutt, C.A.
Wang, S.C.
Wilkinson, J.E.
Wood, G.D.

I. INTRODUCTION

The work reported here consists of both experimental and theoretical research on ionospheric and tropospheric radio propagation, principally throughout the upper part of the radio spectrum, and the practical application of this knowledge to problems in air defense systems.

The research includes:

(a) Investigation of ionospheric radio propagation, primarily at HF and VHF, at frequencies normally above the median ionospheric maximum-usable-frequency (MUF) as determined by current standard methods;

(b) Investigation of tropospheric radio propagation, primarily at UHF and SHF to distances well beyond the horizon;

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* * *

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(c) Development of HF systems for the detection of targets below line-of-sight.

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Some of the work formerly handled by Group 33, including the development of high-powered transmission systems, has been transferred to Group 36, Systems Engineering, which was established in January 1953. During the past quarter, several members of Group 33 worked temporarily in Group 36 on tasks for PROJECT COUNTER CHANGE. This diversion of effort necessitated curtailment of modulation studies started earlier in Group 33 under the VHF "scatter" propagation program.

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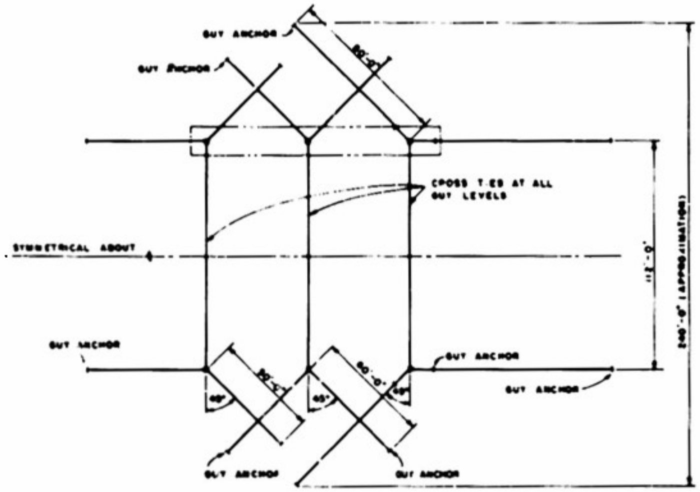
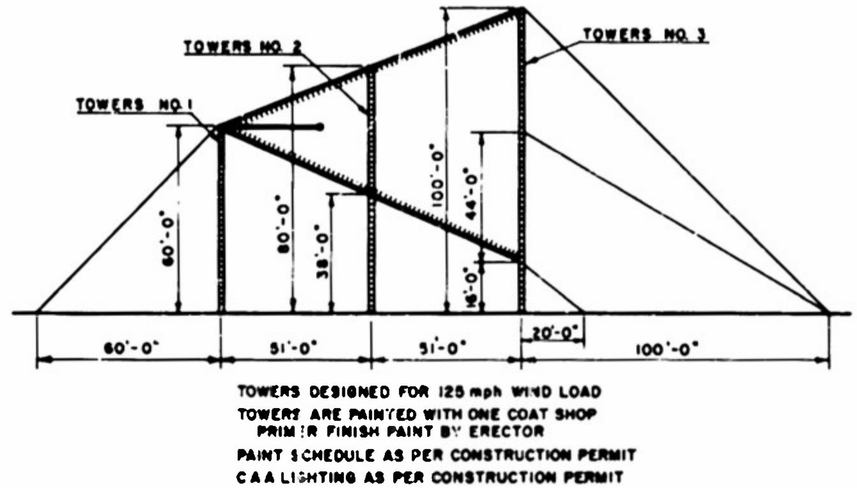
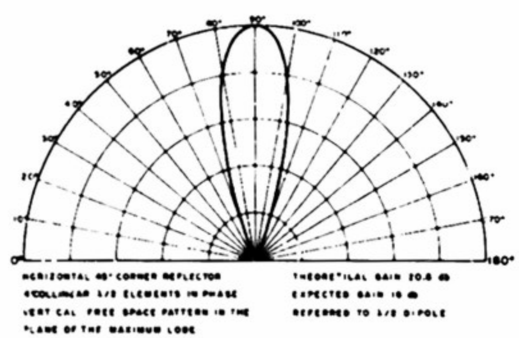


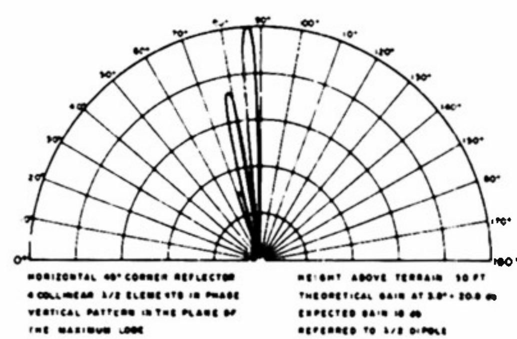
(a) Plan arrangement of towers.



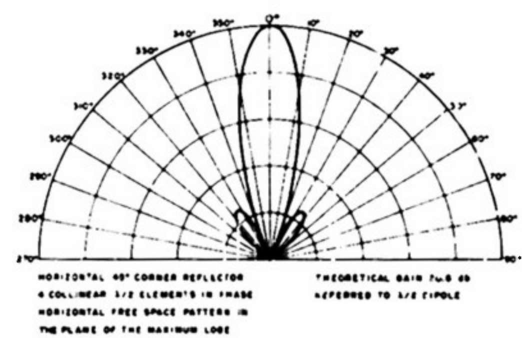
(b) Elevation of towers.



(c) Four horizontal, collinear λ/2 elements in phase. Computed vertical free space pattern in the plane of the maximum lobe.



(d) Four horizontal, collinear λ/2 elements in phase. Height above terrain, 150 feet. Computed vertical pattern in the plane of the maximum lobe.



(e) Four horizontal, collinear λ/2 elements in phase. Computed horizontal free space pattern in the plane of the maximum lobe.

Fig.33-1. Goose Bay (Labrador) 45° corner-reflector receiving antenna.

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II. RADIO-PROPAGATION RESEARCH

A. Radio Propagation at HF

The propagation research at HF during this quarter was confined for the most part to the Alpine (New Jersey) - Goose Bay (Labrador) 23-Mc circuit. At Round Hill, reception and recording of 15- and 18-Mc signals transmitted from Cincinnati were discontinued 1 February 1953 because of the pressure of other work. Arrangements have been made with the Collins Radio Company to provide special transmissions at 27.75 Mc beamed toward Sterling, Va., and Round Hill, Mass., and these transmissions are scheduled to begin in April 1953.

1. Experimental Studies

a. Alpine-Goose Bay (23 Mc)

Transmission and reception facilities described in the 1 October 1952 Quarterly Progress Report of Division 3 were operated throughout this quarter.

The five-element Yagi antenna intended as an interim transmitting antenna at Alpine was destroyed in an ice storm on 11 January before installation had been completed.

Fabrication of a two-Yagi stacked array was rushed to completion and the array shipped to Alpine. This antenna is to be located at a mean height of 240 feet at the end of a cross-arm on an existing tower at Alpine. Comparative performance of this array and that of the existing compromise rhombic antenna will be made during April.

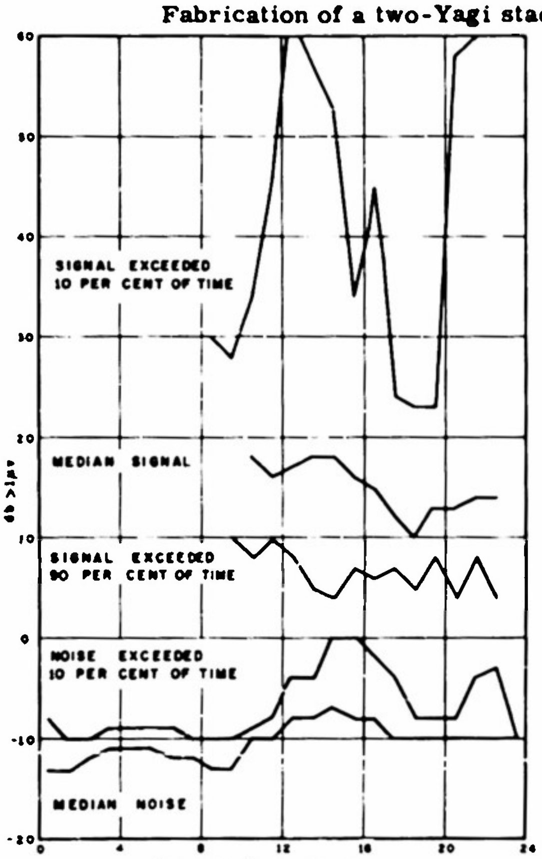

Fig. 33-2. Hourly median values of signal and noise levels, 22.8585 Mc, December 1952 (Alpine-Goose Bay).

Elevation and plan views of the horizontal 45° corner-reflector receiving antenna at Goose Bay prepared by the RAF are given in Figs. 33-1(a) and 33-1(b). This antenna is located on a cliff at a height of approximately 100 feet above the reflecting plane. Antenna patterns, computed by the E. C. Page Consulting Radio Engineers, are given in Figs. 33-1(c) through 33-1(e). Computed gains are given as 20.8 db (perfect ground) and 18 db (with expected losses) at 3.8° vertical angle, both with respect to a $\lambda/2$ dipole.

As received at Goose Bay, median values of 22.8585-Mc signals from Alpine, and noise level for each hour have been obtained for the months of December 1952, January and February 1953. Median values of signal level and noise level, together with values of signal level exceeded 10 per cent and 90 per cent of the time, are plotted in Figs. 33-2 through 33-4.

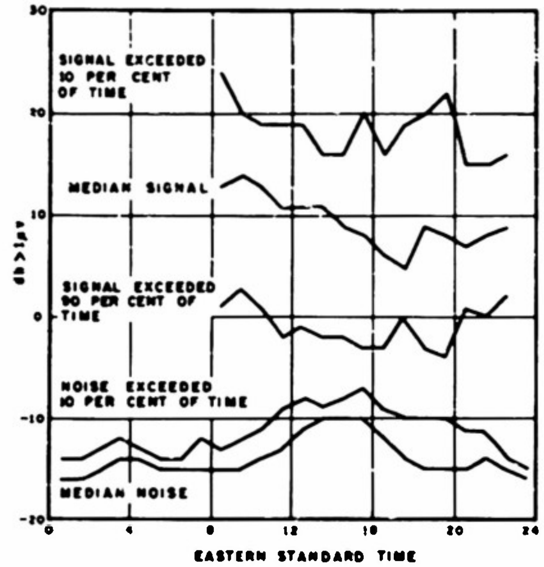

Fig. 33-3. Hourly median values of signal and noise levels, 22.8585 Mc, January 1953 (Alpine-Goose Bay).

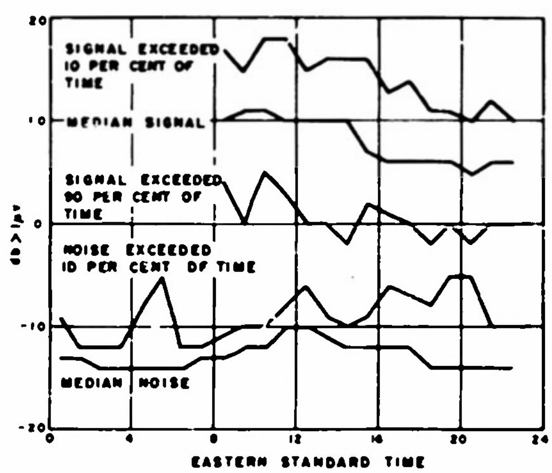

Fig. 33-4. Hourly median values of signal and noise levels, 22.8585 Mc, February 1953 (Alpine-Goose Bay).

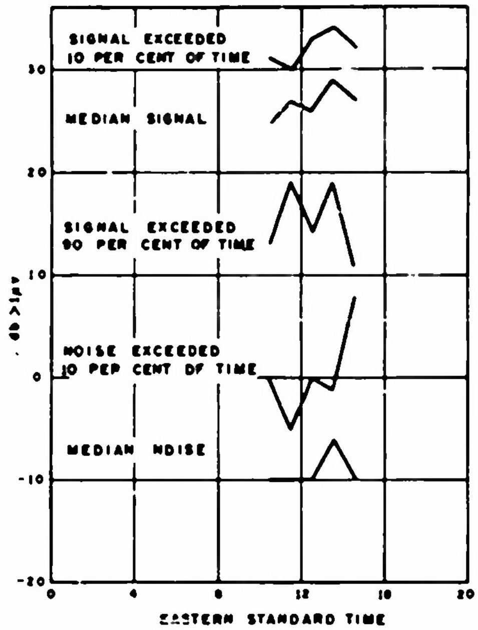

Fig. 33-5. Hourly median values of signal and noise levels, 17.795 Mc, January 1953 (Cincinnati-Round Hill).

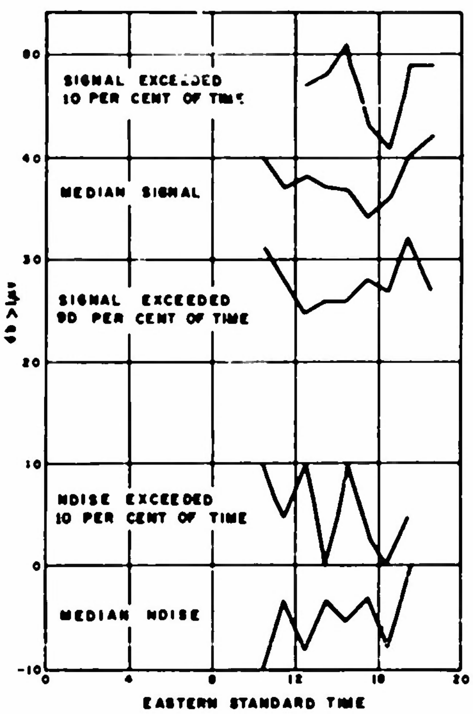

Fig. 33-6. Hourly median values of signal and noise levels, 15.250 Mc, January 1953 (Cincinnati-Round Hill).

b. Cincinnati-Round Hill (15 and 18 Mc)

Transmission and reception facilities of this circuit were described in the previous Quarterly Progress Report. Reception and recording at Round Hill of 15.250-Mc and 17.795-Mc signals radiated from Cincinnati were terminated on 1 February 1953 because of the pressure of other work at the Round Hill Field Station. Median values of recorded field intensity and noise level for the month of January 1953, together with values of field intensity exceeded 10 per cent and 90 per cent of the time, are plotted in Figs. 33-5 and 33-6.

W. G. Abel

B. Radio Propagation at VHF

The propagation research at VHF has been reoriented during this quarter to allow more experimental effort to be applied to other programs. Continuous 24-hour daily recording of signals from Cedar Rapids was discontinued on 1 February 1953, after completion of a full year of continuous recording. Periodic signal-level recordings were made during teletype tests. Detailed analysis of the time trends and distribution characteristics of 50-Mc data recorded over a full year was continued during this quarter.

Further development on the all-electronic commercial glow-transfer counter version of the signal level vs time (SLVT) indicator has been curtailed, as was further work on two-frequency measurements and the 49.6-Mc pulse receiver. Some effort continued at a reduced level on space-diversity and teletype studies.

1. Experimental Studies

a. Time Trends and Distribution Characteristics of 50-Mc Transmissions

Transmission and reception facilities described in the 30 April 1952 Quarterly Progress Report were operated continuously during this quarter until 1 February 1953.

Monthly medians of these received signals for December 1952 and January 1953, together with mass plots of fE_S observed at Washington, D.C., for the same months, are shown in Figs. 33-7 and 33-8.

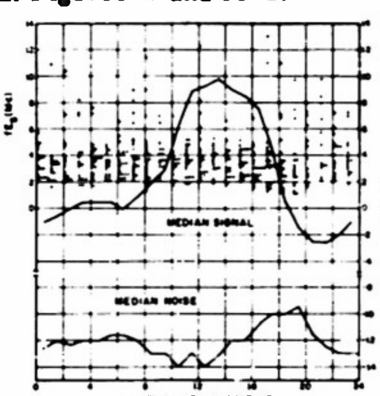

Fig. 33-7. Hourly median values of signal and noise levels (solid lines), 49.6 Mc, December 1952 (Cedar Rapids-Round Hill). Also shown is a mass plot (dots) of fE_s observed on the same date at Washington, D.C.

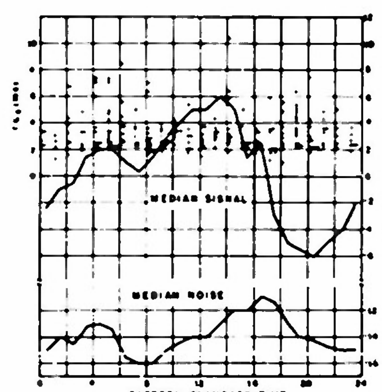

Fig. 33-8. Hourly median values of signal and noise levels (solid lines), 49.6 Mc, January 1953 (Cedar Rapids-Round Hill). Also shown is a mass plot (dots) of fE_s observed on the same date at Washington, D.C.

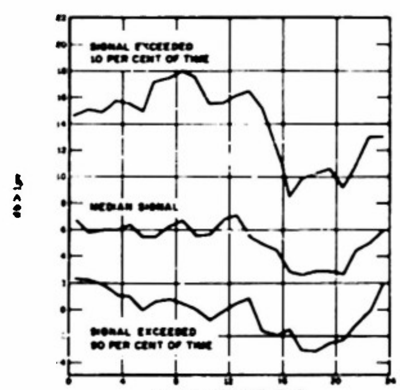

Fig. 33-9. Hourly median values of signal levels, determined from SLVT indicator, 49.6 Mc, September 1952 (Cedar Rapids-Round Hill).

GROUP 33

Signal-level distributions for the month of September 1952 were determined from data obtained during operation of the SLVT indicator, described in the 1 October 1952 Quarterly Progress Report of Division 3. Median signal levels, together with signal levels exceeded 10 per cent and 90 per cent of the time, are shown in Fig. 33-9.

W. G. Abel
J. H. Chisholm
G. D. Wood

b. Space-Diversity Studies

The array of five 5-element Yagi antennas described in the preceding Quarterly Progress Report was installed during this quarter. Considerable difficulty was encountered in obtaining a feed line and switching system that would operate properly under changing weather conditions. The feed system was redesigned to eliminate the original tubular dielectric lines and switches in favor of a separate open-wire strain feeder from each of the antennas back to the receiver building. Tests with a signal generator placed at a distance of approximately one mile indicated that reception on each antenna system was practically identical. A twin-beam oscilloscope and dual-channel receiver with common local oscillators was installed to allow a photographic study of the instantaneous phase relation of 50-Mc signals on any pair of the spaced Yagi antennas. Further study is planned of the median gain of combinations of pairs of Yagi antennas in the array by utilizing the time-totalizing equipment previously used in the recording program.

J. H. Chisholm
J. E. Wilkinson
G. D. Wood

c. Comparative Teletype Studies

A National Company triple-diversity radio teletype terminal, Type AN/FRR-24, shown in Fig. 33-10, was installed at Round Hill during this quarter. Operation of this equipment, the CV/TRA-7 dual-diversity military-type equipment, and other special teletype equipment developed by this Laboratory, is scheduled for the next quarter. Comparative tests of these teletype systems were made over a three-day period in late March, utilizing the two broadside antennas. Faulty operation of the teletype equipment at the transmitter was discovered in these tests and the results were not considered valid. An adequate monitoring system at Cedar Rapids is being planned to overcome this difficulty.

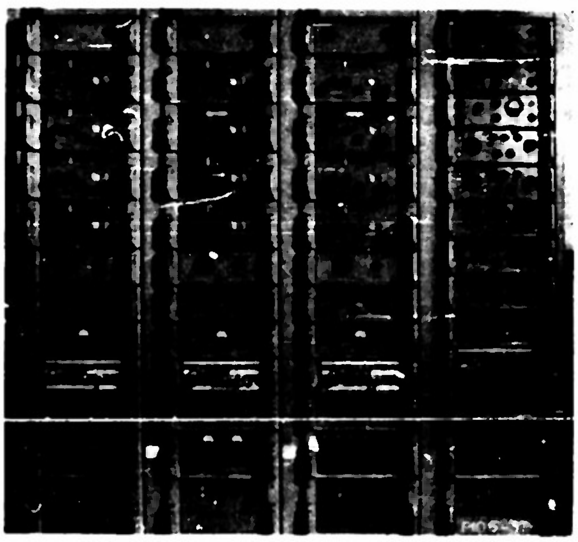

Fig. 33-10. National Company type AN/FRR-24 triple-diversity receiver, used with additional frequency converters for 49.6-Mc reception.

J. H. Chisholm
T. J. Fitzsimmons
G. D. Wood

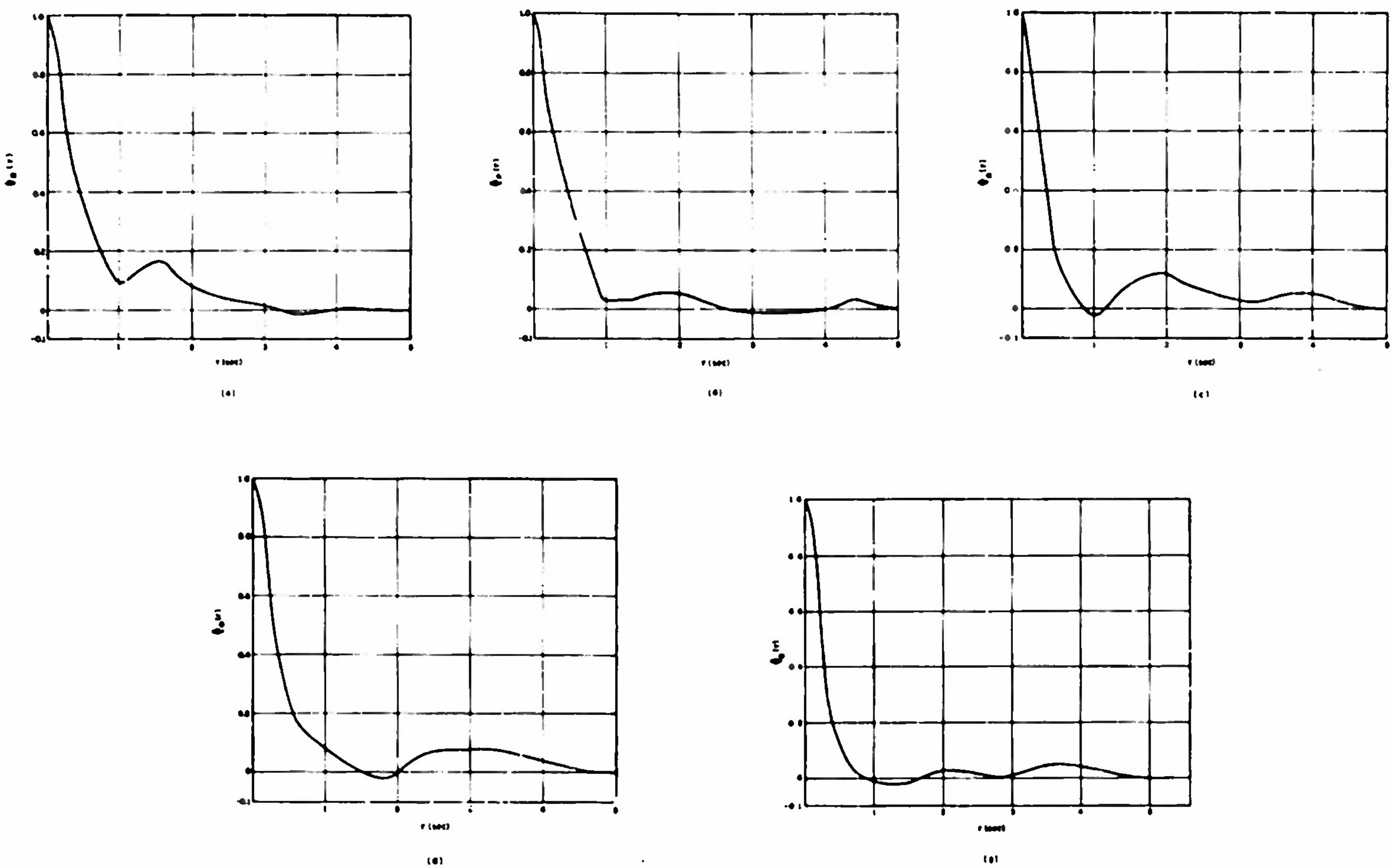

Fig. 33-11. Autocorrelation functions of the amplitude of VHF (49.6-Mc) signals (Cedar Rapids-Round Hill) taken on 4 December 1952 at the following times (EST): (a) 1120 hours, (b) 1430 hours, (c) 1530 hours, (d) 1550 hours, (e) 1610 hours.

GROUP 33

2. Theoretical Studies

a. Autocorrelation Studies on VHF Signals

A major breakdown of the digital electronic correlator occurred at the beginning of this period, which necessitated the complete reconstruction of the 45-tube multiplier chassis. The new chassis employs the same basic circuit design as the original unit, but it has been completely repackaged to facilitate maintenance. A reset circuit which is extremely useful for locating faults in the multiplier and integrator circuits has been incorporated in the new unit. After an extensive period of testing, the correlator was again placed in service.

In the interval during which reconstruction was proceeding on the multiplier, a program of computation of the correlation functions was started on the Servomechanisms Laboratory's mechanical correlator.* This machine requires input data recorded on Brush oscillograph paper. Without employing frequency division, the nominal range of the oscillograph extends from DC to 100 cycles per second. The paper record is fed through the computer by an automatic drive. Conversion of the record is accomplished by two operators who track the oscillograph trace with the aid of two linkage-operated styli. The motions of the styli are coupled to two mechanical integrators in such a way that one integrator computes the mean value of the signal while the other computes the mean of the product of the two tracking signals. Various values of τ are obtained by displacing the position of one stylus with respect to the other.

To date, five autocorrelation functions of the received amplitude of the VHF signal have been computed on the analogue correlator. The data were taken on the Cedar Rapids-Round Hill link. Steps of τ of 1/5 sec were taken, with a maximum value of 4 to 5 sec. The results are shown in Figs. 33-11(a) through 33-11(e).

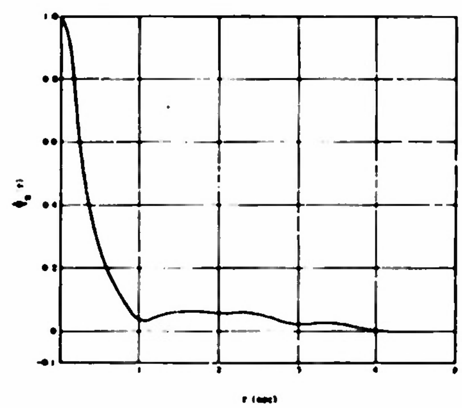

Fig. 33-12. Average of correlation functions of amplitude of received 49.6-Mc signals shown in Fig. 33-11.

The variations in the curves are probably attributable to the different times at which the original data were taken. The receiving equipment was operated from a gasoline engine-generator power supply when the data from which the curve in Fig. 33-11(a) was obtained were taken.

The average of the curves in Figs. 33-11(a) through 33-11(e) is shown in Fig. 33-12. A power spectrum of this curve is being computed.

More comprehensive results of the autocorrelation analysis of VHF signals are given in Lincoln Laboratory Technical Report No. 21, dated 18 May 1953.

A. J. Lephakis
M. L. Stone

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*N. Zabusky, "The Mechanical Correlation Computer," Engineering Report No. 32, Servomechanisms Laboratory, M.I.T. (11 October 1951).

C. Radio Propagation at UHF and SHF

Construction of the experimental equipment designed for an investigation of the practical utility of tropospheric propagation of UHF and microwave frequencies at distances well beyond the horizon is nearing completion. Preliminary operation of a 425-Mc system began on 27 March over a 162-mile path between Alpine and Round Hill. Operation will continue on a 0900 to 1600 hours schedule over a five-day week for the initial studies. Microwave pulse experiments over a 188-mile path from Holmdel (New Jersey) to Round Hill are expected to begin in April. Plans to supplement these microwave pulse experiments with a system that will permit multiplexing of several voice channels were initiated in this quarter in collaboration with the Bell Telephone Laboratories. The Laboratory received shipment in February of the components of a crystal-controlled CW microwave radar system through the cooperation of the Air Force Cambridge Research Center. This system, developed by the Sperry Gyroscope Company, incorporates a 600-watt CW klystron amplifier operating at a frequency of 5050 Mc. This equipment is being modified to provide a transmitter capable of FM modulation over moderate bandwidths with a high order of frequency stability. The Bell Telephone Laboratories is assisting in this program by furnishing components of the standard AT & T Type TD-2 microwave relay terminal equipment, and by modifying the final multiplier stages of the equipment to operate with the 5050-Mc transmitter and receiver. It is expected that this system will be placed in operation in the next quarter if two special multiplier klystrons on order with the Sperry Gyroscope Company are received on schedule.

1. Experimental Studies

a. UHF Experiments (425 Mc)

The 425-Mc transmissions at Alpine are provided by a 15-kw crystal-controlled resnatron transmitter operated by Major E. H. Armstrong. The antenna for this transmitter is a 16-foot diameter paraboloidal antenna with a numerical power gain of 300 above a half-wave dipole. The transmitting antenna is mounted on a rotatable mount on a 100-foot steel tower and is directed toward Round Hill on a bearing of 75°. The transmitter is capable of FM modulation and pulse-amplitude modulation. Initial operation began 27 March utilizing 1000-cycle tone with ±20 kc deviation frequency modulation, and was received at Round Hill on two identical Motorola FM receivers described in the preceding Quarterly Progress Report. The receiving antennas were two small horizontally polarized, corner-reflector antennas. One of the two antennas is mounted in fixed position and the other is mounted on a carriage which permits a variable horizontal spacing up to 50 feet normal to the propagation path. The mounting arrangement is shown in Fig. 33-13. The signals received on these antennas (approximately 9 db gain) during the daily period 0900 to 1600 hours had hourly-median values of 3 to 10 μv across the 50-ohm receiver input. These values are approximately 70 to 60 db, respectively, below free space for this 162-mile path, assuming the gain of the transmitting and receiving antennas is fully realized. Preliminary recording of the detector current (first limiter grid current in the FM receiver) for the two systems with an antenna spacing of 20 feet indicates a high degree of correlation in

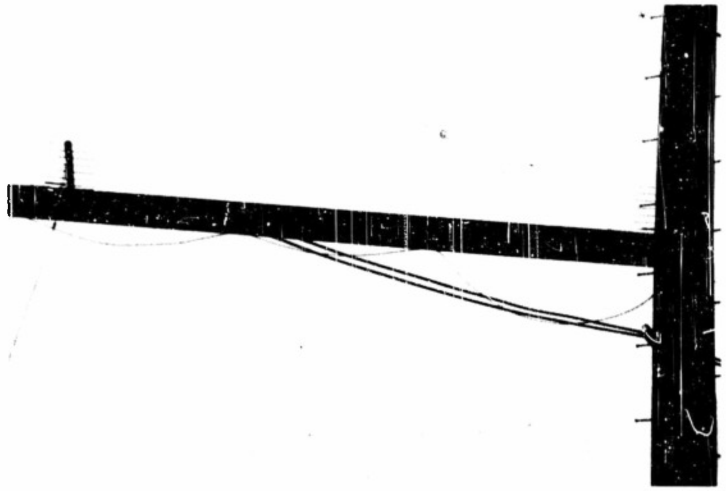

Fig.33-13. Corner-reflector antennas, 425 Mc, mounted for space-diversity reception studies at Round Hill (Alpine-Round Hill circuit).

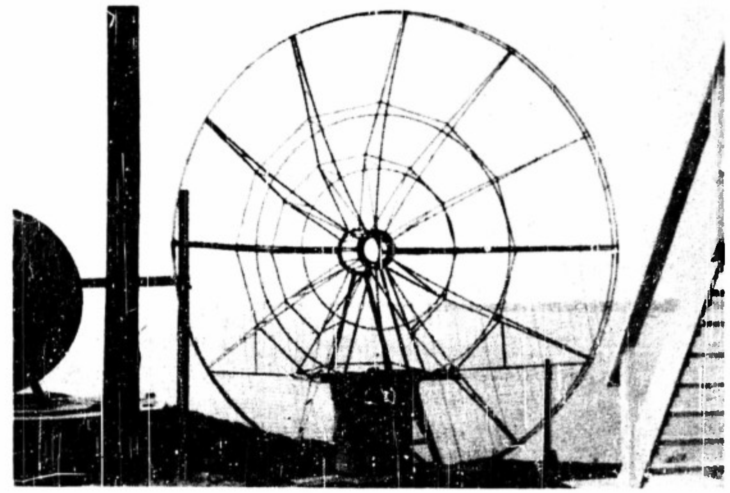

Fig. 33-14. Paraboloidal antenno, 17-foot diameter, installed at Round Hill for 425-Mc signal propagation studies (Alpine-Round Hill circuit).

Fig.33-15. Rear view of the 28-foot diameter paraboloidal antenna installed at Round Hill for SHF signal propagation studies.

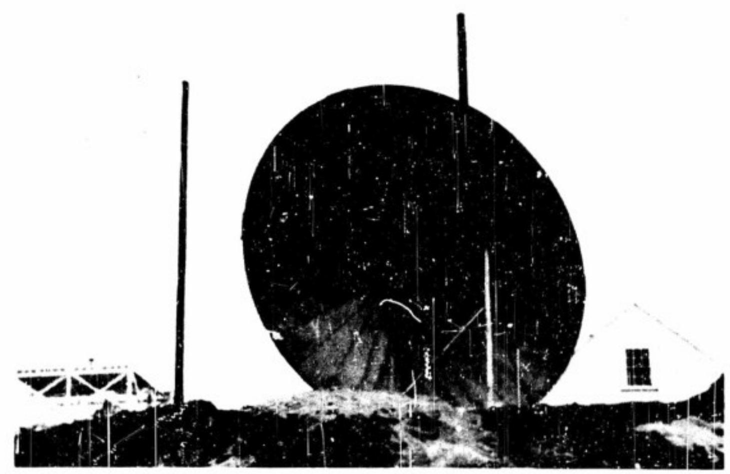

Fig. 33-16. Front view of the 28-foot diameter paraboloidal antenna installed at Round Hill for SHF signal propagation studies.

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the fading. More extensive studies of space diversity, with the two receivers arranged for operation with common local oscillators, is scheduled for the next quarter.

One of two 17-foot paraboloidal antennas was received from the Boston Naval Shipyard through the cooperation of the Bureau of Ships, USN. This antenna and pedestal was installed on Round Hill (as shown in Fig. 33-14) for a comparative study of the gain characteristics of a range of antenna sizes. The antennas available are the small corner reflectors, 17-foot diameter paraboloid, and the 28-foot diameter paraboloid described in the SHF experiments. Antenna feeds suitable for 425-Mc operation of the latter two antennas are under construction and will be completed in early April.

J. H. Chisholm
T. J. Fitzsimmons
P. A. Portmann
G. D. Wood

b. SHF Experiments (3650 Mc)

The pulse experimental system described in the preceding Quarterly Progress Report is expected to be complete and in operating condition in April. The two 28-foot paraboloidal antennas were completed and delivered in March. Antenna foundations and the steel supporting frames were installed at Crawford Hill (New Jersey) and at Round Hill. Both antennas were assembled and erected in late March. Rear and forward views of the assembled antenna at Round Hill are shown in Figs. 33-15 and 33-16.

The modified SV radar equipment and its associated synchronization circuits have been completed and tested in the laboratory. Some difficulty in the subdivider system was experienced, and it was found to result from the circuits containing germanium diodes. These circuits were redesigned to give more reliable operation. A special pulse receiver shown in Fig. 33-17 was constructed for reception at Round Hill. This receiver was designed to

Fig. 33-17. Special pulse receiver for reception at Round Hill, 3650 Mc, of transmissions from Crawford Hill (modified SV radar).

incorporate an AVC circuit, to provide a large dynamic range for recording purposes and to operate with a photographic pulse-shape recording system. The two Western Electric 100-kc frequency standards were operated and compared throughout the quarter. A faulty thermostat was found in one of the units and replaced. It was found that the stability of these standards could be further improved by mounting them in enclosed insulated cabinets. Latest tests of the units thus mounted indicate relative stability of one cycle at a 100-kc frequency over a one-hour period, which indicates that use of the standards at each end of the test circuit would permit displaying a pulse in a 5-μsec gate for periods of the order of 30 minutes without adjustment.

J. H. Chisholm
P. A. Portmann
W. C. Reisener, Jr.

c. Modulation tests at 5050 Mc over the Crawford Hill-Round Hill 190-Mile Path

During the last quarter, a program has been initiated in collaboration with the Bell Telephone Laboratories and with equipment manufactured by the Sperry Gyroscope Company and furnished by the Air Force Cambridge Research Center for the purpose of setting up suitable equipment to study propagation conditions well beyond the radio horizon at SHF (5050 Mc). The equipment is intended to provide information on the suitability of long-range voice and teletype communications at SHF, thus complementing the pulse-propagation studies being made at 3650 Mc. It is expected that the 28-foot paraboloidal antennas and the recording equipment set up for the 3650-Mc tests will also be used for the 5050-Mc studies.

AFCRC has made available a Sperry Doppler-radar system which will be adapted to the propagation studies. This unit consists of a crystal-controlled multiplier chain, microwave klystron amplifiers, and a crystal-controlled receiver. For the purpose of the tests, this unit will be modified at the Lincoln Laboratory to operate with an AT&T TD-2 microwave relay system and a 70-Mc FM modulator which can accommodate a number of voice and teletype channels. A block diagram of the proposed transmitter is shown in Fig. 33-18. When completed, this transmitter is expected to have a CW power output of 600 watts and a frequency stability of ±3 parts in 10^6.

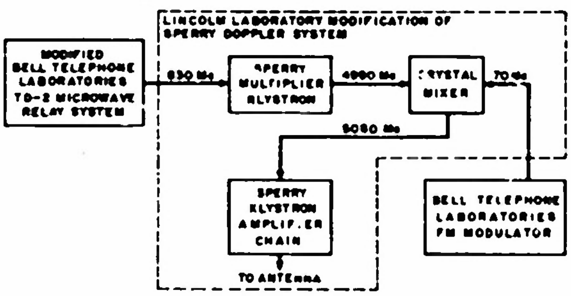

Fig. 33-18. Block diagram of 5050-Mc communication test transmitter (Crawford Hill).

Work on the system at BTL and Lincoln has been proceeding satisfactorily, and the unit is expected to be ready for operation in April of this year. Thus far, modifications of the TD-2 system at BTL have indicated that this equipment will operate satisfactorily. Work at this laboratory has been directed toward assembly of mounting racks and panels and provision of additional power supplies for an extra klystron amplifier. A modified crystal modulator system has also been designed and constructed. The klystrons and crystal modulator are shown in Fig. 33-19.

Plans to modify the system to obtain greater frequency stability, if this proves to be necessary, have been formulated. The use of a highly stable system would permit a substantial

Fig. 33-19. Klystrons and crystal modulator for 5050-Mc communication test transmitter (Crawford Hill).

reduction in the bandwidth of the receiving equipment, with a corresponding increase in signal-to-noise ratio.

J. H. Chisholm
P. A. Portmann

2. Theoretical Studies: Comparison of Bilinear Model with Measurements at VHF

a. Bilinear Model Computations vs Measurements at VHF

Progress with calculations based on the bilinear profile for the index of refraction of the normal atmosphere has continued to indicate that experimental observations on the average strength of weak nonoptical fields at VHF and microwave frequencies can be explained as a normal property of the earth's inhomogeneous layer of air dielectric, without assumption of omnipresent turbulence up to several miles in the troposphere. For a frequency of 50 Mc, it has been possible to calculate both the real and imaginary parts of the complex characteristic values corresponding to a linearly stratified atmosphere of finite thickness, given surface index and standard index gradient. Further, assuming conventional antenna heights of 500 and 30 feet, only about 24 modes appear to make important contributions to the expression for the field strength at 50 Mc. Figure 33-20 shows a calculated sum of 24 modes at various distances from 5 to 400 miles from the 500-foot transmitting antenna to a 30-foot receiving antenna at 50 Mc. A dotted curve by Saxton between 60 and 300 miles, which represents his average curve drawn

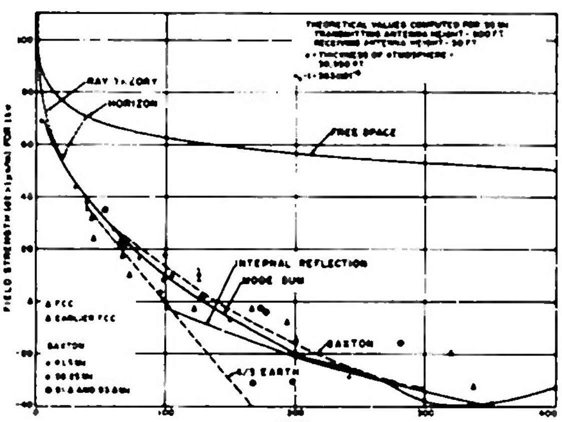

Fig. 33-20. Comparison of VHF measured field strengths with computed theoretical values (bilinear model, 24 modes). Measured data of Saxton in England (41.5, 58.25, 91.5 and 93.8 Mc) and FCC.

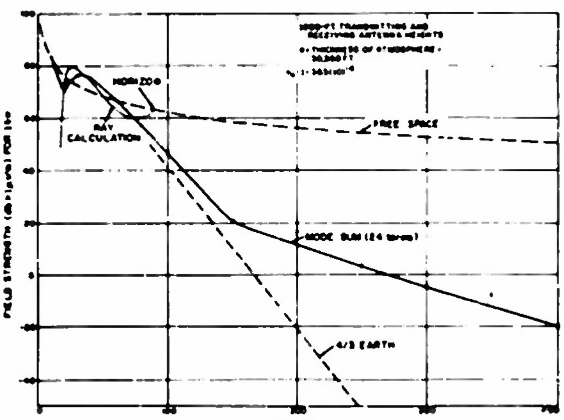

Fig. 33-21. Computed theoretical VHF field strengths (bilinear model, 24 modes).

through long-term experimental observations in England on FM and TV stations and long-term average observations on United States FM and TV stations are also shown. The agreement of the mode-sum curve and Saxton's experimental-average curve is very good. Note that the mode-sum curve in the region of the horizon slopes more steeply than it does further out, in approximate agreement with the conventional 4/3 earth curve. Within the horizon, the mode-sum curve joins on to the values calculated on the basis of direct and earth-reflected rays.

Figure 33-21 shows a similar calculation of a sum of the same 24 modes for 50-Mc frequency, but for 1000-foot antenna heights. Within the horizon, the mode-sum curve oscillates about the free-space value, and has a maximum and a minimum at about the ranges indicated by the direct and reflected-ray calculations shown by the dotted line. The way the bilinear mode sum agrees with simple ray calculations of the lobe structure within the horizon, and the rather sharp drop-off just beyond the horizon, as well as the more slowly attenuating fields well beyond the horizon, constitutes support for believing that the model reasonably represents the effect of the normal troposphere on propagation out to about 400 miles. The conventional 4/3 earth curve is not needed for the calculation at any distance. The conventional curve does follow the experimental values for a short distance beyond the horizon – namely, where the field is still so strong that the feeble internal-reflection effect of the atmosphere may be neglected compared to the predominant upward leakage of energy in the near-shadow region of the earth bulge. The conventional 4/3 earth theory fails well beyond the horizon in the deep shadow because it neglects completely the downcoming partial reflections produced by the inhomogeneous troposphere. The 4/3 earth notion can be used to correct the diffracted field for the refractive effect of the normal troposphere, but not for the reflective effect.

Although the bilinear profile seems to explain the average fields out to about 400 miles, it is believed that beyond this range the calculated values will be too low, because the model assumes no air above 30,000 feet; and neglected partial reflections occurring at greater heights will become more important as the intersection of the transmitter and receiver horizons occurs higher in the air.

T. J. Carroll
R. M. Ring

b. Variations of 49.6-Mc Signal Strength With Surface Refractive Index

In previous reports, variations of 49.6-Mc signal strength have been compared with sporadic-E ionization, or more specifically with fE_s. On the basis of the observed correlations, sporadic-E ionization is suggested as one probable mechanism for the propagation of these 49.6-Mc signals.* To see whether the troposphere plays a more important part in the propagation mechanism controlling the nighttime behavior of the long-distance 49.6-Mc signal, it was suggested that tropospheric surface index changes, both at receiver and transmitter, should be examined and compared with the observed fields. Comparisons of this type have recently been made by Pickard and Stetson** both at 44.1 Mc and 92.1 Mc over a 167-mile path. The surface index

*See Quarterly Progress Report, 30 January 1952, Research Laboratory of Electronics and PROJECT LINCOLN.
**G.W.Pickard and H.T.Stetson, J. Atmos. and Terr. Phys. 1, 32-6 (1950).

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of refraction at one or both terminals was shown to have a strong correlation with the received fields.

The 49.6-Mc fields received at Round Hill from Cedar Rapids (horizontal polarization, path length approximately 1068 miles) were compared with the surface refractive index at Iowa City (Iowa) which is close enough to Cedar Rapids to have very similar surface weather. Weekly-median fields at Round Hill during the nighttime interval 2000 to 0330 hours EST* were compared with the corresponding weekly numerical averages of the surface refractive index** at Iowa City during the nighttime interval 1900 to 0300 CST. The results are shown in Fig. 33-22 for the period 1 February to 15 August 1952. An examination of this figure shows rather poor correlation during February and March; but, beginning with April and extending into August, a much better correlation seems evident.

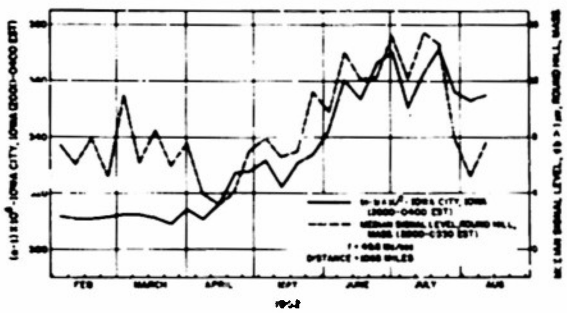

Fig. 33-22. Comparison between weekly nighttime values of 49.6-Mc signal levels received at Round Hill from Cedar Rapids and nighttime averages of surface refractive index at Iowa City.

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III. RADAR SYSTEMS FOR TARGET DETECTION BELOW LINE-OF-SIGHT

A. Ground-Wave Radar Systems

1. Noise-Modulated System

The major part of the construction by the Theodore Loranger Company of New Bedford for the Round Hill Field Station has been completed. The transmitter-receiver building is finished and is presently in use. The broadside array of eight vertical semirhombics is sufficiently well advanced so that definite plans for impedance, phase, and pattern measurements can be made for the next quarter.

The Radio Engineering Laboratories Model 520 FM broadcast transmitter, modified for use in the HF band, has been delivered, and is now installed in the transmitter-receiver building at Round Hill.

Progress on design and construction of the noise modulators, receiver components, delay lines, and reed-indicating devices for this system is reported by Group 32 in this report.

In addition to the noise equipment, a crystal-controlled pulse modulator and coherent

*Because of polarization switching, horizontal-horizontal signals were recorded only for half-hour periods beginning on the hour.

**The index was calculated from the formula given in *Propagation of Short Radio Waves* (edited by D.E. Kerr, McGraw-Hill, Inc., 1951, Eq.(3), p. 13) in conjunction with hourly values of surface weather data supplied by the United States Weather Bureau, Washington, D.C.

detector are being designed and built to permit this system to be used as a pulse radar. In addition to obvious radar applications, a provision for pulse modulation will facilitate studies of fixed echoes around the site and Doppler studies of sea-clutter and back-scatter returns.

A study of alternate methods for display of the Doppler frequencies is also being initiated. An aural presentation of the Doppler frequencies may offer considerable advantages over reed devices, but because of the very low Doppler range (0 to 40 cps in the HF band) the Doppler frequencies cannot be heard directly, so that some system, such as frequency multiplication or modulation of an audio carrier, must be used.

2. High-Power Pulse System

Under a Lincoln Laboratory subcontract, the Raytheon Manufacturing Company is installing transmitting and receiving equipment and providing other services for the placing of the Air Force Field Station at Plum Island in an operational condition. Originally scheduled to begin 1 January 1953, the actual starting date was unavoidably delayed until 23 March 1953.

Installation of equipment will proceed in parallel with impedance, phase, and pattern measurements of the antenna during the next quarter.

Constructions of the coherent detector for use with the high-power pulse system has been completed at the Lincoln Laboratory, and it is now being aligned and tested preparatory to installation at Plum Island.

B. Ionospheric Radar System

As stated in the last Quarterly Progress Report, the transmitting and receiving equipment for the high-power pulse system will be used for the first experiments on a one-hop ionospheric system. A suitable antenna for general use in an ionospheric radar will not be available for a considerable time, and is the bottleneck in performing such experiments at an early date. It should be possible, however, to perform experiments of limited scope during the summer months with the present ground-wave radar antenna and with the present frequency assignments by utilizing one-hop E-layer transmission. The possibility of constructing an adequate array for F_2-layer transmission through use of simple and easily constructed elements is also being investigated.

C. A. Stutt
W. T. Quinn, Jr.

COMMUNICATIONS TECHNIQUES

GROUP 34

COMMUNICATIONS TECHNIQUES

GROUP 34

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Turin, G.L.
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I. EXPERIMENTAL TWO-CHANNEL NOMAC SYSTEMS

A. High-Frequency NOMAC Teletype System

During the period 26 January to 6 February, the performance of a NOMAC teletype system and a standard frequency-shift-keyed teletype system were compared on a 24-hour basis over a 230-mile path from Coles Signal Laboratory, Fort Monmouth, New Jersey, to Building B, Lexington, Mass., on a frequency of 5.4325 Mc. The results obtained with both systems were unsatisfactory because of strong local interference. Pressure of other work has precluded further testing.

B. M. Eisenstadt
P. L. Fleck, Jr.
C. A. Wagner

B. VHF-NOMAC-FSK Teletype System[1]

During the week of 11 January, the VHF system was put into operation over the Cedar Rapids to Round Hill link. This preliminary test indicated the feasibility of using noise modulation over a path using scatter propagation. Several modifications were then made. The major change was to replace the 125-kc band shaping filters, which were Jacobian elliptic filters, with 6-pole Butterworth filters. At the same time, the VHF overtone crystal for the front end was received from the factory and installed.

Operational tests in the first week of February showed satisfactory performance; the relatively large error rate, about one per cent, was attributed to fading. Hence, the receiver was adapted for dual space-diversity reception. During the second week of March, the dual-diversity receiver was completed and tested. It was found that dual diversity reduced the percentage error in received copy to one-fifth to one-fourth of the value without diversity. In the receiver, two methods of diversity combining were tried: the first was the addition of the rectified keying waveforms; the second, the addition of the keying tones at the output of the first averaging filter. Both methods gave about the same reduction in percentage error.

R. S. Berg
W. C. McLaughlin
R. A. Paananen
G. L. Turin

C. NOMAC Systems for PROJECT COUNTER CHANGE

Equipment for six NOMAC systems for PROJECT COUNTER CHANGE is being built; four will be completed by 1 May 1953 and the other two by 1 June 1953. Further discussion of this work will be found in the report of Group 36.

R. S. Berg
W. C. McLaughlin
R. A. Paananen
G. L. Turin

GROUP 34

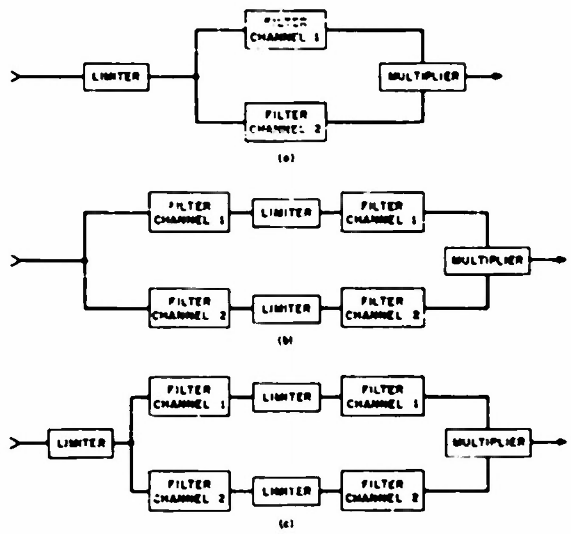

Fig. 34-1. Diagram of limiting schemes.

D. UHF NOMAC Data Link

The experimental data link was subjected to interference tests. The tests were inconclusive, since the asymmetry of some components caused different rates of error in the two channels and denied us a satisfactory way of evaluating our results. For this reason, we are revising the data link to improve the components.

We are now building input filters for the revised system. When the system works, we shall compare the limiting schemes shown in Fig. 34-1 with various kinds of jamming action. These results will be compared with theory.

B. Eisenstadt
P. Fleck, Jr.
C. Wagner

II. EXPERIMENTAL ONE-CHANNEL NOMAC SYSTEMS

A. Stored-Signal System

The experimental system has been completed and its performance is being studied. Figure 34-2 shows a comparison of theoretical and experimental data on loss in output as a function of time displacement between transmitted and stored signals. The comparison has been made for two different stagger-tuning configurations of the ringing circuits whose combined output forms the signal. The theoretical behavior is described by Eq. (34-35) on page 55 of this report.

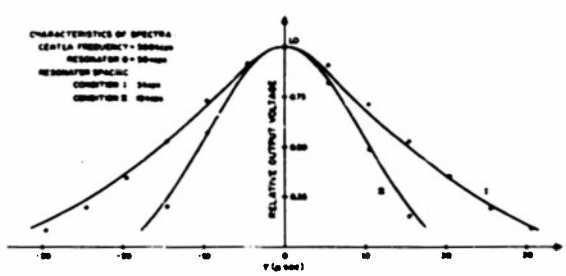

Fig. 34-2. Loss of output vs time desynchronization for two spectrum shapes.

For condition I, the ratio of output-to-input (S/N) has been measured at 22.3 db, compared with a theoretical figure of 23.8 db.

P. E. Green

III. STUDIES OF PROPAGATION AND CHANNEL DISTURBANCES AND NOMAC SYSTEMS

A. Communication System Synthesis for Multipath Propagation

In the preceding report,[1] mathematical synthesis of a representative path was studied as a first step toward treatment of the general time-varying multipath problem. The component path was assumed to be of fixed delay, but with a strength Rayleigh-distributed and simple Markoff. Study of the transmission of white Gaussian symbols over such a path was found to involve considerable analytical difficulty.

Re-examination of the mechanism of single-path fading,[2] much as in recent studies of ionospheric scatter transmission, has led to a more realistic formulation of the component path, and has also simplified the mathematical expressions. The path model is now defined by the following mathematical operations upon the transmitted signal. First, it is still assumed that the transmitted signal and additive channel noise are confined to a common frequency band W wide, so that it is possible to conduct the sampling-point type of entropy analysis usually employed in the study of continuous waveforms. The transmitted waveform may be specified by the amplitudes of sine and cosine components (or, equally well, the amplitude of the waveform and its derivative) at these sampling points, which are spaced in time 1/W apart. This gives 2TW degrees of freedom in a transmission nominally limited to band W and time T. At each sampling point, the resultant of the sine and cosine components is also adequate specification, and path-strength fluctuations are considered to be the result of separate random multiplication of the projections of the resultant on two axes 45° on either side of it. If the two multipliers are taken to be Gaussianly distributed independent random processes with identical Gaussianly shaped autocorrelation functions, the model satisfies the theory of random scatterers postulated by Ratcliffe, and it agrees with the observed Rayleigh distribution of the received envelope when a sine wave is transmitted. It is important to note that, with this revised model, the received carrier phase is random, unlike the rather unrealistic constant time delay assumed in the study of the Rayleigh multiplier. Following the "complex" multiplication, channel noise is, of course, added to complete the progression from transmitted to received signals.

Preliminary studies of the revised model of the path have been undertaken, assuming a binary transmission of sine and cosine components; that is, the amplitudes of the components are $\pm a$, with equal probability. Let i and j represent two successive sampling points, and x be the projection of the transmitted resultant on one of the 45° axes, these axes being fixed at the transmitter. Let the random multipliers have unit mean-square, and the additive noise have power σ_n^2. Let the projection of the received resultant on one of its 45° axes be W, with the condition that both W_j and W_j are the projections of the two successive resultants on one of the axes aligned at 45° to the first resultant. Thus at the receiver the axes are re-aligned from one sampling point to the next. Although strictly possible, the main purpose of this artificiality is to enable the two components to be treated independently in probability and entropy calculations. We find, if the signal and noise are independent from point to point,

$$P(W_j/W_i, x_i, x_j) = \sqrt{\frac{\sigma_n^2 + x_i^2}{2\pi[\sigma_n^2(\sigma_n^2 + x_i^2) + \sigma_n^2 x_j^2 + x_i^2(1-\phi^2)]}} \times \exp\left\{ - \left[\frac{\left(W_j \sqrt{\sigma_n^2 + x_i^2} - \dfrac{W_i x_i x_j \phi}{\sqrt{\sigma_n^2 + x_i^2}} \right)^2}{\sigma_n^2(\sigma_n^2 + x_i^2) + \sigma_n^2 x_j^2 + x_i^2 x_j^2 (1-\phi^2)} \right]^{-2} \right\}. \tag{34-1}$$

and

$$P(W_j/W_i) = \frac{\exp\left[-\frac{(W_j - kW_i)^2}{2(\sigma_n^2 + a^2)(1-k^2)}\right] + \exp\left[-\frac{(W_j + kW_i)^2}{2(\sigma_n^2 + a^2)(1-k^2)}\right]}{2\sqrt{2\pi(\sigma_n^2 + a^2)(1-k^2)}} \quad , \tag{34-2}$$

where

$$k = \frac{\phi a^2}{\sigma_n^2 + a^2} \quad , \tag{34-3}$$

and ϕ is the autocorrelation function of the multipliers, evaluated at $1/W$. $P(W_j/W_i, x_i, x_j)$ is the conditional probability density of W_j, knowing x_i, x_j, and W_i, while $P(W_j/W_i)$ is that knowing W_i alone. Similar equations hold for the components on the other 45° axis.

Now

$$H(W) = -\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} P(W_i, W_j) \log P(W_j/W_i)\, dW_k dW_j \quad , \tag{34-4}$$

and

$$H_x(W) = -\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} P(W_i, W_j, x_i, x_j) \log P(W_j/W_i, x_i, x_j)\, dW_i dW_j\, dx_i dx_j \quad . \tag{34-5}$$

Using the results from Sec. B (pp. 53, 54) of this report, we obtain the channel capacity:

$$C = H(W) - H_x(W)$$
$$= \log_e 2 - \frac{2}{\pi\alpha} + \frac{1}{\pi}\left[\frac{2}{\alpha^2} - 1\right]\tan^{-1}\alpha - \begin{cases} \displaystyle\sum_{n=2}^{\infty} \frac{(-1)^n \tan^{-1}\frac{\sqrt{\beta}}{(2n-1)}}{\pi n(n-1)\frac{\sqrt{\beta}}{\alpha}} & \text{for } \beta > 0 \\ \displaystyle\sum_{n=2}^{\infty} \frac{(-1)^n \tanh^{-1}\frac{\sqrt{-\beta}}{(2n-1)}}{\pi n(n-1)\frac{\sqrt{-\beta}}{\alpha}} & \text{for } \beta < 0 \end{cases} \quad , \tag{34-6}$$

where $\beta = \alpha^2 - 4n(n-1)$ and

$$\alpha = \frac{\sqrt{(\sigma_n^2 + a^2)^2 - \phi^2 a^4}}{\phi a^2} = \frac{\sqrt{(1+R)^2 - \phi^2 R^2}}{\phi R} \quad , \tag{34-7}$$

with $R = a^2/\sigma_n^2$ the power signal-to-noise ratio.

In Fig. 34-3, C is plotted as a function of R and ϕ. Also, for comparison there are plots of C for no scatter, both for binary transmission and for the optimum Gaussian symbol transmission. It is apparent that scatter severely affects the available channel capacity when the receiver is operating on the simple Markoff basis specified by Eq. (34-1) and Eq. (34-2). Below R = 0.5, the channel capacity is found to be roughly proportional to R in the nonscattering case, and to ϕ^2R^2 in the case of scatter. Furthermore, in this region these results apply regardless of the form of the amplitude distribution of the transmitted resultants, provided they still assume only the four directions. On the other hand, the upper asymptotes indicated depend in general on the statistics of the transmission. For the binary transmission indicated, maximum efficiency is realized for $\phi \gtrsim 1$ at $R \gtrsim 1$.

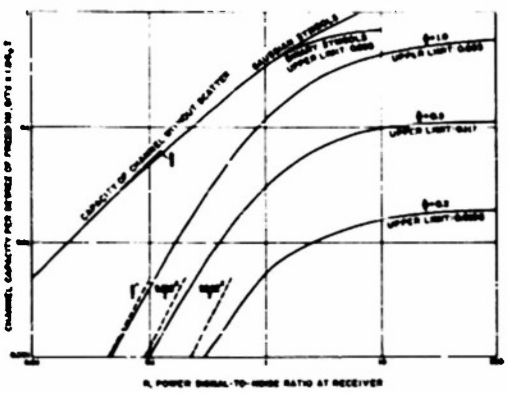

Fig. 34-3. Channel capacity of a binary system for scatter and noise in channel.

Optimizing the transmitted symbols for given R and ϕ involves a difficult integral equation, and no general results have been obtained from this problem. Nevertheless, the new path model permits several integrals to be obtained in closed form where they could not be before. This is mainly due to the Gaussian multiplicative disturbance convoluting nicely with the Gaussian additive disturbance.

It is proposed to investigate higher-order probability-computing processes at the receiver, in the belief that improvements in channel capacity over those indicated in Fig. 34-3 can be achieved.

R. Price

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B. <u>The Effect of Channel Perturbations on NOMAC Systems</u>

The problem under consideration is to evaluate the effects of certain general types of channel perturbations on the performance of NOMAC systems. A formulation of one specialization of this problem is given in the previous Quarterly Progress Report, dated 15 January 1953, to which we refer. With a slight change in notation we have, as we had there, an expression for the output Y of the correlation detector

$$Y = \int_0^T \left[\int_0^{\gamma T} A(t, \tau)\, s(t-\tau)\, d\tau \right] s'(t)\, dt \quad ,$$

where $A(t, \tau)$ is in general a random function characterizing the behavior of the channel, T is the time length of a symbol, γT is the time interval over which a single impulse at the transmitter is stretched out at the receiver, $s(t)$ is the random function representing the transmitted signal, and $s'(t)$ is equal either to $s(t)$ or to another random function with the same statistics as $s(t)$. We want first to know the probability law for the random variable Y, or, failing that, at least its first and second moments.

One special case which seems to be of interest is where $A(t, \tau)$ is not actually random, so that the perturbation of the signal is only a "smearing," as might be caused by a nonrandom scattering of the propagated signal. The previous report contained some calculations about this

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case. Another calculation has shown the following. Suppose the transmitted signal can be written

$$s(t) = \sum_{k=W_1}^{W_T} \left[x_k \cos k\omega_o t + y_k \sin k\omega_o t \right] \quad ,$$

where x_k and y_k are normalized, independent Gaussian variables; suppose $A(t, \tau) = A_o$ = const. Then the probability law of the random variable Y is given by the density function

$$f(Z) = \sum_n^+ \left\{ \frac{\exp\left[- \frac{Zn\omega_o}{\sin n\omega_o \gamma T}\right]}{\frac{\sin n\omega_o \gamma T}{n\omega_o} \prod_{k \neq n} \left(1 - \frac{n\omega_o \sin k\omega_o \gamma T}{k\omega_o \sin n\omega_o \gamma T}\right)} \right\} \text{for } Z > 0 \quad ,$$

and

$$f(Z) = \sum_n^- \left\{ \frac{\exp\left[- \frac{Zn\omega_o}{\sin n\omega_o \gamma T}\right]}{\frac{\sin n\omega_o \gamma T}{n\omega_o} \prod_{k \neq n} \left(1 - \frac{n\omega_o \sin k\omega_o \gamma T}{k\omega_o \sin n\omega_o \gamma T}\right)} \right\} \text{for } Z < 0 \quad ,$$

where $Z = 2Y/AT$ and $\sum_n^+$ is the summation over all n between W_1 and W_2 for which $n\omega_o/\sin n\omega_o \gamma T$ is negative, and $\sum_n^-$ is a similar summation for $n\omega_o/\sin n\omega_o \gamma T$ positive. It can also be shown that, if $A(t, \tau)$ is allowed to fluctuate with mean value A_o, the same probability density f(Z) is approached as the fluctuations become arbitrarily fast.

Another special case is that in which the smearing time γT is zero so that A(t) is simply a random multiplier. Then

$$Y = \int_o^T A(t)\, s(t)\, s'(t)\, dt \quad .$$

If now we let A(t) be a stationary Gaussian process of moving averages with mean value $A_o > 0$ and correlation function $\exp[-|\alpha t|] + A_o^2$ and restrict s(t) to be a given signal function, Y is a Gaussian random variable whose mean and variance are easily calculated.

We may show that symptotically for large T the variance of Y is

$$2\int_o^T (T - u) \exp[-\alpha u]\, \overline{s^2(u)\, s'^2(0)}\, du \quad ,$$

if reasonable statistical assumptions are made about the class of signal functions [s(t)]. The integral is elementary for either of the representations of the signal that have been used above and in the previous report.

W. L. Root

IV. STUDIES OF CODING AND DATA HANDLING

A. Error Correction for NOMAC Systems

A new type of error-correcting code has been proposed. It is designed specifically for use with a two-symbol system using correlation detection.

Let the two possible transmitted symbols be denoted by $x_1(t)$ and $x_2(t)$, and suppose that the receiver has available replicas of each. If $x_1(t)$, say, is transmitted and corrupted by channel noise to $x_1(t) + n(t)$, the receiver (ideally) forms the two finite-time correlations

$$M_1(T) = \frac{1}{T}\int_0^T [x_1(t) + n(t)]\, x_1(t)\, dt \qquad (34\text{-}8)$$

and

$$M_2(T) = \frac{1}{T}\int_0^T [x_1(t) + n(t)]\, x_2(t)\, dt \quad . \qquad (34\text{-}9)$$

The receiver then selects $x_1(t)$ or $x_2(t)$ as the transmitted symbol according to whether $M_1(T)$ or $M_2(T)$ is larger (maximum probability detection).

A finite probability of error per symbol arises as a result of the possibility that $M_2(T)$ is less than $M_1(T)$ even though $x_1(t)$ was sent. For, as a result of the finite time of integration, $M_1(T)$ and $M_2(T)$ are random variables with means

$$S = \lim_{T\to\infty} \frac{1}{T}\int_0^T x_1^{\,2}(t)$$

and zero, respectively, and variances

$$\sigma_{M_1}(T) = \frac{1}{2TW}\,(2S^2 + SN) \qquad (34\text{-}10)$$

and

$$\sigma_{M_2}(T) = \frac{1}{2TW}\,(S^2 + SN) \quad , \qquad (34\text{-}11)$$

where N is the power of a Gaussian noise of bandwidth W. Equations (34-10) and (34-11) can be derived in an elementary fashion from an expression given in a paper of Davenport, Johnson, and Middleton.[3] The probability of error per symbol P is then just

$$P = \int_{-\infty}^{+\infty} \text{prob}(M_2 = x)\,\text{prob}(M_1 < x)\, dx \quad , \qquad (34\text{-}12)$$

which can be evaluated as*

$$P = \frac{1}{2}\,(1 - \text{erf}\, a) \quad , \qquad (34\text{-}13)$$

where

$$a = \frac{S}{\sqrt{2(\sigma_1^{\,2} + \sigma_2^{\,2})}} \quad . \qquad (34\text{-}14)$$

Our error-correcting code operates as follows: The usual parity-check digit is appended to each transmitted word. (A word is a group of symbols considered as a unit, such as a command in a data link.) As each symbol arrives at the receiver, it is tentatively decided that

*One method was suggested by Dr. I.S.Reed, Group 24.

the transmitted symbol is the one corresponding to the larger of the integrals M_1 and M_2. However, each of the correlation outputs is stored until the parity-check digit arrives. If examination of that digit indicates an error in the word, the symbol with the least separation ΔM between M_1 and M_2 is taken to be wrong and is corrected. (The symbol in question may be the parity-check digit itself.) In this way, it is hoped to get a single-error-correcting code which, though it cannot correct all single errors, may be as effective as a Hamming code for a fixed data rate, since only a single additional digit is needed.

Calculations of the reduction of the per-word probability of error to be achieved by use of this code are in progress and appear to be formidable but perhaps not prohibitive. We hope to master the computational difficulties. As a preliminary result, we have calculated the probability that, if one of the two symbols is known to be wrong, the wrong symbol is the one with the least separation ΔM. That conditional probability is given by

$$\frac{1}{2\sqrt{\pi}\,PQ}\int_{-a}^{\infty} \exp[-v^2][\mathrm{erf}(v+2a) - \mathrm{erf}\,a]\,dv \quad , \qquad (34\text{-}15)$$

where $Q = 1 - P$, and P and a are given by Eqs. (34-13) and (34-14).

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B. Binary Coding

We continue the problem posed before.[4] A variable V, with values v_i, is often coded as a set of binary digits; during transmission or handling, one (or more) of these digits is mistakenly altered, so that v_j is received as v_k, for example. $|v_j - v_k|$ is called the error and depends on which digits have been altered. If the probability of altering one digit in a v_i is small, the probability of altering two is much smaller, and we therefore restrict our problem to that of trying to find the coding that minimizes the mean-square error, say, accruing to altering all possible digits one at a time.

As a starting point, we set $V = [v_i] = [0, 1, \ldots A]$, and the number of binary digits N, and each of the 2^N sets of binary digits x_i, so that for a set $[a_{ij}]$ of binary digits

$$x_i = \sum_{j=1}^{N} a_{ij} 2^{j-1}$$

Clearly, $2^N > A$. A function f maps v_i onto the x_k, so that $f(v_i) = x_k$; the inverse function ϕ is such that either $\phi_f(x_k) = v_i$ if $f(v_i) = x_k$, or, if not, ϕ is unspecified.

What we want to do, according to our formulation, is find a function f that minimizes

$$T_{A,N}(f) = \frac{1}{2^N}\sum_{i=1}^{2^N}\sum_{j=1}^{N}\left\{ v_i - \phi[f(v_i) + (1 - 2a_{ij})\,2^{j-1}]\right\}^2 \quad . \qquad (34\text{-}16)$$

where, if $\phi f(v_i) + (1 - 2a_{ij})\,2^{j-1}$ is unspecified by its definition, it is set equal to v_i. We set

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$T^*_{A,N}$ as the least $T_{A,N}$.

We have not been able to find any simple expression for $T^*_{A,N}$ in terms of A and N; a number of results is listed below, whose proofs are too tedious for this report:

$$T^*_{A,N} = 0 \qquad \text{if } A \leq 2^{N-1} \quad .$$

$$T^*_{2^N,N} \leq \frac{1}{3}(2^{2N} - 1);$$

or, more generally,

$$T^*_{A,N} \leq \frac{2}{3}(A - 2^{N-1})(2^{N+2} - 1) \qquad \text{for } 2^{N-1} \leq A \leq 2^N \quad .$$

$$T^*_{A,N} \leq \begin{cases} 4(A - 2^{N-1})(2^{N/2} - 1) & \text{for N even} \quad , \\ 3(A - 2^{N-1})[2^{(N+1)/2} - \frac{4}{3}] & \text{for N odd,} \end{cases}$$

where $2^{N-1} \leq A \leq 2^{N-1} + k$, for $k \leq N - 1$.

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V. INFORMATION THEORY AND NOMAC SYSTEMS

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A. Bounds on Entropy

Shannon[5] has shown that, for a one-dimensional probability distribution p(x), of standard deviation σ^2, the maximum entropy that can be achieved is log $\sqrt{2\pi e \sigma^2}$, and that it is achieved when p(x) is a Gaussian distribution. The entropy H is defined for a continous probability distribution p(x) as:

$$H = -\int_{-\infty}^{+\infty} p(x) \log p(x)\, dx \quad , \tag{34-17}$$

while:

$$\sigma^2 = \int_{-\infty}^{+\infty} x^2 p(x)\, dx - \left[\int_{-\infty}^{+\infty} x p(x)\, dx\right]^2 \quad . \tag{34-18}$$

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We can establish a lower bound to the entropy with the inequality[6]

$$-\int_{-\infty}^{+\infty} p(x) \log p(x)\, dx \geq -\log \int_{-\infty}^{+\infty} p^2(x)\, dx \quad . \qquad (34\text{-}19)$$

Thus:

$$-\log \int_{-\infty}^{+\infty} p^2(x)\, dx \leq H \leq \log \sqrt{2\pi e\left\{\int_{-\infty}^{+\infty} x^2 p(x)\, dx - \left[\int_{-\infty}^{+\infty} x p(x)\, dx\right]^2\right\}} \quad . \qquad (34\text{-}20)$$

These bounds become useful when Eq. (34-17) is difficult to evaluate in closed form. The closeness of the bounds is indicated in the following table for a few elementary distributions, all of zero mean and variance σ^2.

TABLE 34-1

Distribution Form: $p(x)$	$-\log\int_{-\infty}^{+\infty} p^2(x)\, dx$	$H = -\int_{-\infty}^{+\infty} p(x) \log p(x)\, dx$	$\log\sqrt{2\pi e\left[\int_{-\infty}^{+\infty} x^2 p(x)\right] dx}$
Gaussian: $a \exp[-bx^2]$	$\log \sqrt{4\pi\sigma^2}$	$\log \sqrt{2\pi e\sigma^2}$	$\log \sqrt{2\pi e\sigma^2}$
Exponential: $a \exp[-bx]$	$\log \sqrt{8\sigma^2}$	$\log \sqrt{2e^2\sigma^2}$	$\log \sqrt{2\pi e\sigma^2}$
Uniform: $\begin{cases} a; & -b \leq x \leq b \\ 0 & \text{elsewhere} \end{cases}$	$\log \sqrt{12\sigma^2}$	$\log \sqrt{12\sigma^2}$	$\log \sqrt{2\pi e\sigma^2}$

It is also interesting to note that the maximum value of the lower bound,

$$-\log \int_{-\infty}^{+\infty} p^2(x)\, dx \quad ,$$

for a given σ^2, is $\log \sqrt{125/9\ \sigma^2}$, achieved by a distribution that is an inverted parabola between finite bounds and zero outside.

Since the logarithm is a convex function, Eq. (34-19) is an equality only if $p(x)$ is everywhere 0 except for an interval or set of intervals on which $p(x)$ is constant. Thus Eq. (34-20) never strictly determines H except for the degenerate cases where $p(x)$ is a delta function and $H = -\infty$, or where $p(x) \to 0$ everywhere and $H \to +\infty$.

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B. The Entropy of the Sum of Two Displaced Gaussian Distributions

When binary digits are transmitted through Gaussian channel noise, the probability distribution of the levels observed at the receiver is of the form

$$p(x) = \frac{1}{2\sqrt{2\pi}} \exp[-\frac{(x-a)^2}{2}] + \exp[-\frac{(x+a)^2}{2}] \quad , \qquad (34\text{-}21)$$

where the pulses have amplitudes $\pm a$, and the noise has unit variance. We wish to evaluate the entropy

$$H = -\int_{-\infty}^{+\infty} p(x) \log p(x)\, dx \quad . \qquad (34\text{-}22)$$

By a perturbation method, it has been found

$$H = \log 2\sqrt{2\pi e} - \frac{\sqrt{2}}{\pi} a \exp[-\frac{a^2}{2}] + (2a^2 - 1)\, \mathrm{erf}(-a)$$

$$- \sum_{n=2}^{\infty} \frac{(-1)^n \exp[2a^2 n(n-1)]\, \mathrm{erf}(-a)\,(2n-1)}{n(n-1)} \quad , \qquad (34\text{-}23)$$

where

$$\mathrm{erf}(s) = \frac{1}{\sqrt{2\pi}} \int_s^{\infty} \exp[-\frac{y^2}{2}]\, dy \quad . \qquad (34\text{-}24)$$

If $a \geqslant 1$, the series converges rapidly, and a good approximation is

$$H \approx \log 2\sqrt{2\pi e} - 0.79788\, a \exp[-\frac{a^2}{2}] + (2a^2 - 1)\, \mathrm{erf}(-a) + \frac{\exp[4a^2]\, \mathrm{erf}(-3a)}{2}$$

$$+ \frac{\exp[-\frac{a^2}{2}]}{a}\left(0.010003 - \frac{0.000456}{a^2} + \frac{0.000062}{a^4}\right) \qquad (34\text{-}25)$$

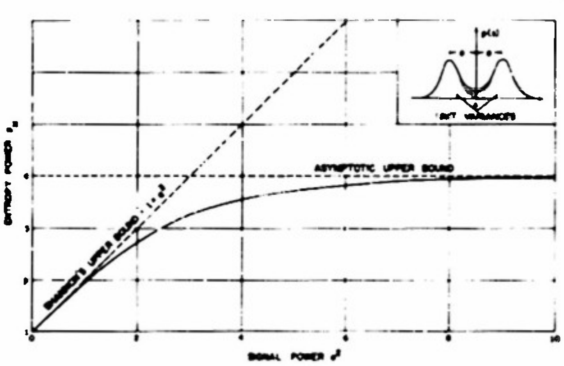

Fig. 34-4. Entropy power of two displaced Gaussian distributions.

In Fig. 34-4, the entropy power p_H of p(x) is plotted as a function of a^2, the signal power. Entropy power is defined as the power of a Gaussian distribution having the same entropy as that given, and is related to H by

$$p_H = \frac{\exp[2H]}{2\pi e} \qquad (34\text{-}26)$$

Indicated also in Fig. 34-4 are two upper bounds. One is readily obtained for infinite separation of the peaks, while the other was

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established by Shannon[7] as the sum of noise and signal powers. It is especially significant to note the convergence of p_H to the latter, for this demonstrates the absorption property of Gaussian noise, also stated by Shannon.[7] It is seen that in this case the signal-to-noise ratio may be as high as unity with no appreciable loss in entropy power from the use of a non-Gaussian signal distribution.

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C. <u>Output Signal-to-Noise Ratio for Arbitrary Gaussian Signal and Additive Noise</u>

A more general expression than that previously available[8] has been obtained for the output signal-to-noise ratio for cross-correlation detection of a signal in additive Gaussian noise where the amplitudes are assumed Gaussianly distributed. In the block diagram of Fig. 34-5, $S(\omega)$, $N_1(\omega)$, and $N_2(\omega)$ are the power-density spectra of the signal and the two independent noise disturbances. The quantity $H(\omega)$ represents a linear perturbation causing one signal to differ from the other, and is assumed transformable and time-invariant. The effect of quasi-stationary multipath conditions or of a state of desynchronization between the two signals can be studied by suitably specifying $H(\omega)$.

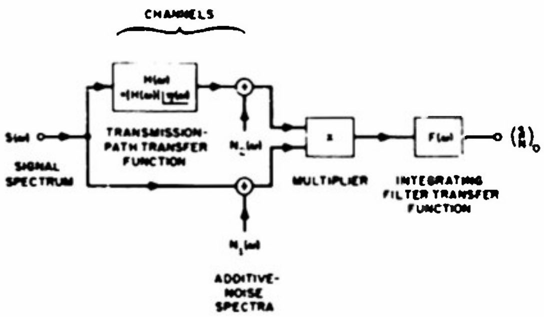

Fig. 34-5. Model of correlation detection system for additive channel noise and linear channel distortion.

The expression for output signal-to-noise ratio is

$$\left(\frac{S}{N}\right)_o = \frac{1}{W_f}\left[\int_o^\infty S|H| \cos\eta \, d\omega\right]^2 \left[\int_o^\infty S^2|H|^2 \cos^2\eta \, d\omega + \frac{1}{2}\int_o^\infty N_1 S|H|^2 \, d\omega + \frac{1}{2}\int_o^\infty SN_2 \, d\omega + \frac{1}{2}\int_o^\infty N_1 N_2 \, d\omega\right]^{-1} , \quad (34\text{-}27)$$

if the integrating filter characteristic $F(\omega)$ is a low-pass function, and W_f is the effective bandwidth of the filter in radians per second. For the condition in which the spectra of the two signals are deliberately separated by a difference frequency $\omega = \Delta$ and the averaging filter is tuned to this frequency,

$$\left(\frac{S}{N}\right)_o = \frac{1}{W_f}\left[\left(\int_o^\infty S|H| \cos\eta \, d\omega\right)^2 + \left(\int_o^\infty S|H| \sin\eta \, d\omega\right)^2\right]\left[\int_o^\infty S^2|H|^2 \, d\omega + \int_o^\infty SN_2 \, d\omega + \int_o^\infty SN_1|H|^2 \, d\omega + \int_o^\infty N_1 N_2 \, d\omega\right]^{-1} , \quad (34\text{-}28)$$

in which it is implicit that in performing each integration the spectra are first brought into coincidence by shifting one of them by Δ.

Several special cases of these equations are of interest:

(1) If the signal and noises have exactly the same spectral shape [and if $H(\omega) = 1$], then

$$\left(\frac{N}{S}\right)_o = \frac{W_f}{W_s} K\left[1 + \frac{1}{2}\left(\frac{N}{S}\right)_1 + \frac{1}{2}\left(\frac{N}{S}\right)_2 + \frac{1}{2}\left(\frac{N}{S}\right)_1\left(\frac{N}{S}\right)_2\right] \qquad (34\text{-}29)$$

for low-pass integration and

$$\left(\frac{N}{S}\right)_o = \frac{W_f}{W_s} K\left[1 + \left(\frac{N}{S}\right)_1 + \left(\frac{N}{S}\right)_2 + \left(\frac{N}{S}\right)_1\left(\frac{N}{S}\right)_2\right] \qquad (34\text{-}30)$$

for band-pass integration. Here

$$K = \frac{\int_0^\infty S^2(\omega)\, d\omega}{S_{max}(\omega)\int_0^\infty S(\omega)\, d\omega} \qquad (34\text{-}31)$$

is a form factor of the signal-power density spectrum. Several values of this constant are listed in Table 34-II. W_s is the effective bandwidth of the signal

$$W_s = \frac{1}{S_{max}} \int_0^\infty S(\omega)\, d\omega \quad . \qquad (34\text{-}32)$$

(2) If both channel noises are white and have bandwidths W' larger than the signal bandwidth (and if $H(\omega) \equiv 1$), the two expressions become

$$\left(\frac{N}{S}\right)_o = \frac{W_f}{W_s}\left[K + \frac{1}{2}\frac{N_{01}}{S_{max}} + \frac{1}{2}\frac{N_{02}}{S_{max}} + \frac{1}{2}\frac{N_{01}}{S_{max}}\frac{N_{02}}{S_{max}}\frac{W'}{W_s}\right] \quad .$$

and

$$\left(\frac{N}{S}\right)_o = \frac{W_f}{W_s}\left[K + \frac{N_{01}}{S_{max}} + \frac{N_{02}}{S_{max}} + \frac{N_{01}}{S_{max}}\frac{N_{02}}{S_{max}}\frac{W'}{W_s}\right] \quad . \qquad (34\text{-}33)$$

respectively, where N_{01} and N_{02} are the power densities of the two noises.

(3) If $H(\omega)$ represents a pure time shift τ, so that $\eta(\omega) = \omega\tau$, then the output signal power is

$$\left[\int_0^\infty S(\omega) \cos \omega\tau \, d\omega\right]^2 \qquad (34\text{-}34)$$

if low-pass integration is used, and

$$\left[\int_0^\infty S(\omega) \cos \omega\tau \, d\omega\right]^2 + \left[\int_0^\infty S(\omega) \sin \omega\tau \, d\omega\right]^2 \qquad (34\text{-}35)$$

for band-pass integration.

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TABLE 34-II

Type of Spectrum	$S(\omega)$	K
Rectangular	$\begin{cases} 1 \text{ for } \omega \text{ in } W \\ 0 \text{ otherwise} \end{cases}$	1
Triangular	$\begin{cases} 1 - \frac{\lvert\omega - \omega_c\rvert}{\frac{W}{2}} & \text{for } \lvert\omega - \omega_c\rvert < \frac{W}{2} \\ 0 \text{ otherwise} \end{cases}$	$\frac{2}{3}$
Gaussian	$\exp[-(\omega - \omega_c)^2]$	$\frac{1}{\sqrt{2}}$
Exponential	$\exp[-\lvert\omega - \omega_c\rvert]$	$\frac{1}{2}$
1^{st} order Butterworth ("Optical" or "Cauchy")	$\frac{1}{1-(\omega-\omega_c)^2}$	$\frac{1}{2}$
n^{th} order Butterworth	$\frac{1}{1-(\omega-\omega_c)^{2n}}$	$\frac{\Gamma(\frac{2-1}{2n})}{\Gamma(\frac{1-1}{2n})}$

These expressions are also correct for non-Gaussian signals. From the first of these, it is observed that the output power decreases as the square of the signal autocorrelation function, as would be expected. If the signal is a band-pass function, this correlation function takes the form of a cosine wave whose frequency is equal to the band center and whose amplitude decreases with τ.

The behavior with band-pass integration is quite different, because of the presence of the added term. For example, if $S(\omega)$ is a band-pass spectrum, even about some center frequency, the loss in output with delay τ is the same as though $S(\omega)$ were imagined centered at zero frequency, that is, the oscillations of output power with τ observed with low-pass integration are missing. This represents a much greater latitude in synchronizing the two multiplier inputs when band-pass integration is used.

This second expression is of some historical interest since its appearance antedates by thirty years the first generally known use of the correlation function.[9] The normalized expression is identical to that used by Michelson[10] in 1891 to relate the frequency spectrum of a light source[$S(\omega)$]to the output of an interferometer as a function of path difference (τ).

The first denominator term of the $(S/N)_o$ expression, representing the self-noise, is independent of τ for band-pass integration; but for low-pass integration the factor $\cos^2 \omega\tau$ causes this term to decrease by one-half as τ increases from zero to a sufficiently large value.

These results have been discussed in detail in an internal memorandum, "NOMAC Output Signal-to-Noise Ratios for Arbitrary Gaussian Signal and Noise," dated 24 March 1953.

P. F. Green, Jr.

D. NOMAC Systems Employing Randomly Phase-Modulated Carriers

A block diagram of the basic NOMAC system is shown in Fig. 34-6. The relation between the channel signal-to-noise ratios and the signal-to-noise ratio at the output of the averaging filter has been studied by Fano for the case of sinusoidal signals, and for the case of Gaussian signals with "optical" spectra.[8] Results for the case of Gaussian signals with arbitrary spectra are reported by Green in the preceding section.

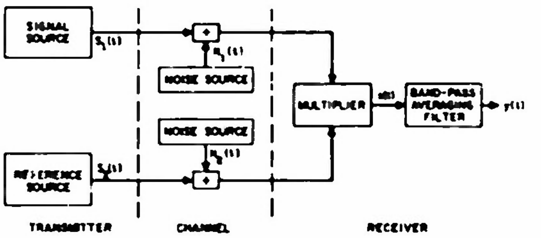

Fig. 34-6. The basic NOMAC system.

The analysis has been extended to a more inclusive class of signals: those in which a carrier is randomly modulated both in amplitude and phase.[11] In particular, it was assumed that the signal $S_1(t)$, could be expressed as

$$S_1(t) = A(t)\cos[pt + \phi(t)] \quad ,$$

and that the reference, $S_2(t)$, differed from $S_1(t)$ only by a frequency shift of amount $q/2\pi$. It was further assumed that $A(t)$ and $\phi(t)$ had only low-frequency components, and that they were statistically independent, but otherwise arbitrary, random variables. Under these assumptions, the multiplier output due to the signal alone is (considering only difference frequency terms)

$$x(t)_s = \frac{A^2(t)}{2} \cos qt \quad .$$

It should be noted that this term is randomly modulated only in amplitude, and not in phase. The signal-to-noise power ratio at the output of the multiplier may be expressed as

$$\left(\frac{S}{N}\right)_x = \frac{1}{\dfrac{\sigma^2(A^2)}{\overline{A^2}^2} + \left(\dfrac{N_1}{S} + \dfrac{N_2}{S}\right) + \left(\dfrac{N_1}{S}\right)\left(\dfrac{N_2}{S}\right)} \quad ,$$

where $\sigma^2(A^2)$ is the variance of the square of the amplitude modulation, N_1 and N_2 are the noise powers in channels 1 and 2, respectively, and where S is the signal power input (assumed to be the same in both channels). When the signal and reference are Gaussian, $A(t)$ is Rayleigh, and hence $\sigma^2(A^2)/\overline{A^2}^2 = 1$. This agrees with the previously reported results.

When the signals are randomly phase-modulated carriers (e.g., clipped narrow-band Gaussian noise), $A(t)$ is a constant, and hence $\sigma^2(A^2)/\overline{A^2}^2 = 0$. Two cases are of particular interest:

(a) That of a stored-reference system in which $N_2 = 0$. In this case

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$$\left(\frac{S}{N}\right)_x = \left(\frac{S}{N}\right)_i$$

for all values of the input channel signal-to-noise power ratio, $(S/N)_i$.

(b) That of a transmitted-reference system in which $N_1 = N_2 = N_i$. In this case

$$\left(\frac{S}{N}\right)_x = \frac{\left(\frac{S}{N}\right)_i}{2 + \left(\frac{N}{S}\right)_i} ,$$

and hence

$$\left(\frac{S}{N}\right)_x \to \begin{cases} \left(\frac{S}{N}\right)_i^2 & \text{as } \left(\frac{S}{N}\right)_i \to 0 \ . \\ \frac{1}{2}\left(\frac{S}{N}\right)_i & \text{as } \left(\frac{S}{N}\right)_i \to \infty \ . \end{cases}$$

These results are plotted in Fig. 34-7, along with corresponding results for Gaussian signals.

It may be seen, from Fig. 34-7, that the randomly phased carrier system behaves essentially the same in the small input signal-to-noise ratio region as the Gaussian-signal system. However, in the large input signal-to-noise ratio region, the Gaussian-signal system has a maximum multiplier-output signal-to-noise power ratio of unity, while that ratio increases linearly with the input ratio in the randomly phased carrier system. This is due to the absence of self-jamming effects, at the multiplier output, in the latter system.

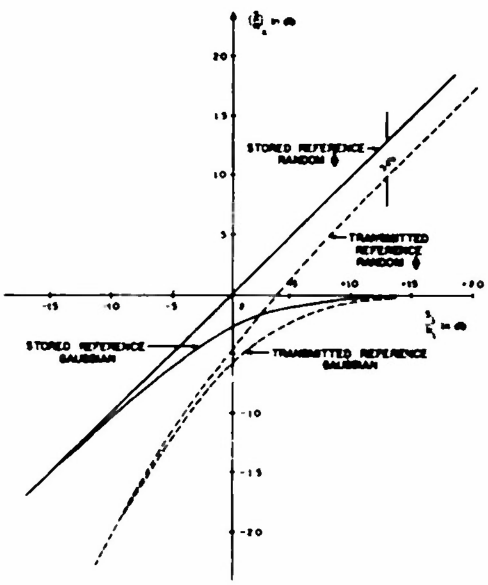

Fig. 34-7. Multiplier output signal-to-noise power ratios.

The above results refer to the multiplier output. It may be shown[11] that the signal-to-noise power ratio at the output of the averaging filter is (in the general case)

$$\left(\frac{S}{N}\right)_c = \frac{\frac{W_s}{KW_f}}{\frac{\sigma^2(A^2)}{\overline{A^2}^2} + \left(\frac{N_1}{S} + \frac{N_2}{S}\right) + \left(\frac{N_1}{S}\right)\left(\frac{N_2}{S}\right)}$$

when the channel signals and interfering noises have the same spectral shapes, and when the effective channel signal bandwidth W_s is much larger than the effective bandwidth of the averaging filter W_f. K is the spectral shape form factor introduced by Green.

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E. Probability of Error in Frequency-Shift NOMAC Systems

Figure 34-8 is the block diagram of the receiver of a typical frequency-shift NOMAC system. The power-density spectra shown in this figure are those that appear at the indicated

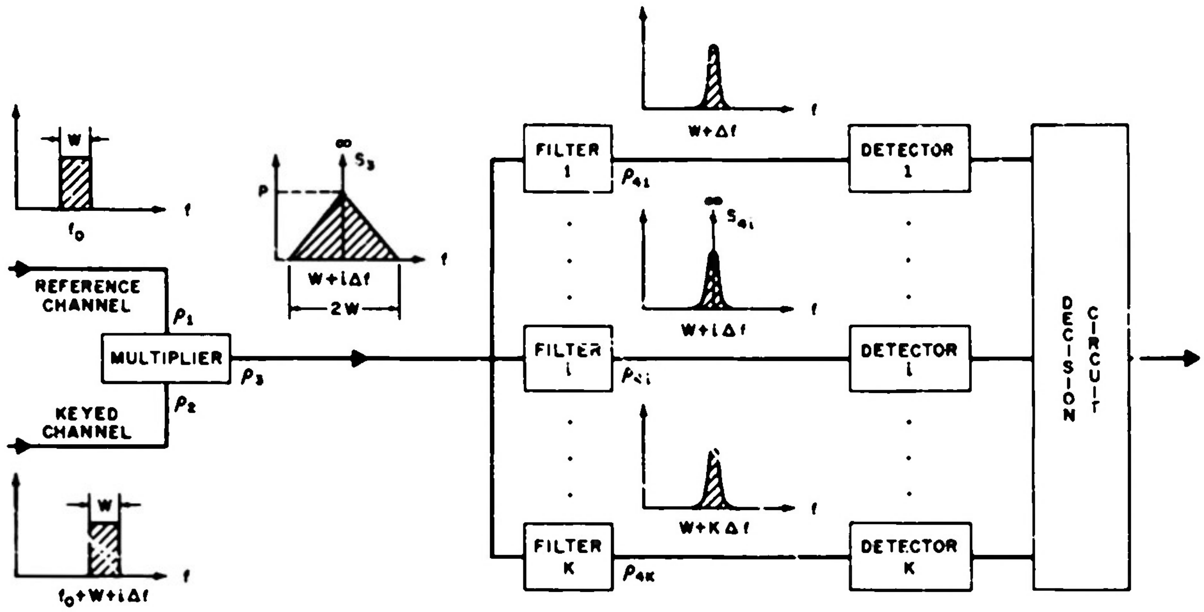

Fig. 34-8. Receiver of typical frequency-shift NOMAC system.[1]

points when the i^{th} symbol is transmitted; only the band centered at W + iΔf at the multiplier output has been shown, since it is the one of interest. The ρ's stand for the signal-to-noise ratios at the indicated points.

The following assumptions are made about this system:

(1) The detectors are either ideal synchronous or ideal envelope detectors.

(2) The decision circuit chooses the largest of the detector outputs as the correct symbol.

(3) The filters are narrow-band band-pass filters, all of which have the same shape transfer function but different center frequencies. Further, the bandwidth of these filters is assumed to be small compared to the frequency increment Δf, so that when the i^{th} symbol is transmitted there is a signal present only at the output of the i^{th} filter.

(4) The frequency increment Δf is small compared to the channel bandwidth W. This permits us to assume that the noise-power density at the input to the filters is essentially constant across the band occupied by the filters, and is equal to P, the noise-power density at the center frequency of the i^{th} filter. Then the noise-power outputs of all the filters are the same.

(5) Either ρ_1 or ρ_2 (or both) is very much smaller than unity. This assures us that the noises at the outputs of the filters essentially have Gaussian statistics.

(6) ρ_{4i} is very much greater than unity. This implies that the bandwidth of the filters is very much smaller than the channel bandwidth W, an assumption already implicit in (3) and (4) above.

(7) Each symbol is transmitted for a long enough time so that steady-state conditions have been reached before the decision circuit makes a decision. It can be shown that this

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assumption may be discarded without changing the results if the filter characteristics are suitably chosen and the decision circuit samples at the right instants.

The relationship between the input signal-to-noise ratios ρ_1 and ρ_2 and the i^{th} filter output signal-to-noise ratio ρ_{4i} is known.[12] The problem remains to find the probability of error at the decision circuit output as a function of ρ_{4i}. This is the probability of error if the i^{th} symbol is sent. However, from the symmetry of the system (and the assumptions made) this is the same as the total probability of error, which is independent of the symbol sent.

For synchronous detection, the probability density distribution at the detector outputs is

$$p_j(x_j) = \begin{cases} \frac{1}{\sqrt{2\pi}} \exp\left[-\frac{(x_j - a)^2}{2}\right] & j = i \ , \\ \frac{1}{\sqrt{2\pi}} \exp\left[-\frac{x_j^2}{2}\right] & j \neq i \ , \end{cases} \tag{34-36}$$

where the variables have been normalized to the common rms value of noise at the filter outputs, and where $a = 2\rho_{4i}$. (The 2 is introduced since the mean value of the detector output is equal to the peak value of the sine wave input.)

For envelope detection, the probability density distributions are[13]

$$p_j(x_j) = \begin{cases} \left.\begin{cases} x_j \exp\left[-\frac{x_j^2 + a^2}{2}\right] I_o(ax) & x_j \geq 0 \\ 0 & x_j < 0 \end{cases}\right\} j = i \ , \\ \left.\begin{cases} x_j \exp\left[-\frac{x_j^2}{2}\right] & x_j \geq 0 \\ 0 & x_j < 0 \end{cases}\right\} j \neq i \ , \end{cases} \tag{34-37}$$

where the variables are again normalized and a is the same as before. In Eq. (34-37), I_o is the zero-order modified Bessel function of the first kind.

The probability of error is

$$P_e = 1 - \int_{-\infty}^{\infty} p_i(x_i)\, dx_i \prod_{j \neq i} \int_{-\infty}^{x_i} p(x_j)\, dx_j \quad . \tag{34-38}$$

Using Eqs. (34-36), (34-37) and (34-38), and assumption (6) we get the following expressions for probability of error:

Synchronous detection:
$$P_e \approx \frac{(K-1)}{2} \frac{\exp\left[-\frac{\rho_4}{2}\right]}{\sqrt{\frac{\pi\rho_4}{2}}} \tag{34-39}$$

Envelope detection: $$p_e \approx \frac{(K-1)}{2} \exp[-\frac{p_4}{2}] \quad . \qquad (34\text{-}40)$$

From these results, we see that synchronous detection is superior to envelope detection for probability of error.

The expression for probability of error with synchronous detection does not agree with one derived by Basore.[14] His expression is

$$p_e \approx \frac{(K-1)}{2} \frac{\exp[-p_4]}{\sqrt{\pi p_4}} \quad . \qquad (34\text{-}41)$$

The discrepancy derives from an approximation made by Basore which has been shown to be invalid by numeral integration of the exact expression (34-38) for a particular value of K.

Because of the approximations made in deriving Eq. (34-39) and Eq. (34-40), the right-hand sides of these expressions are actually upper bounds of the true probabilities of error, and, indeed, for bounded K the probabilities of error approach these upper bounds as limits as $\rho_4 \to \infty$.

G. L. Turin

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TRANSISTORS AND SOLID STATE

GROUP 35

TRANSISTORS AND SOLID STATE*
GROUP 35

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I. CIRCUITS AND APPLICATIONS§

A. Receiver Study†

The development of transistor circuits for the Signal Corps receiver R-392(XC-1)/GRC-19 has continued during the past quarter.

Because of the power-supply problems reported previously, p-n-p junction transistors have been investigated for the IF amplifier. By using some conjugate matching, a grounded-base stage at 455 kcps may be designed to produce a power gain of 18 db over a 20-kcps band (between 3-db points). Such a stage consumes about 15 mw supply power. The bandwidth should be reduced to about 10 kcps by using special IF coils which have a Q of 200 at 455 kcps. These coils are available from the Sickles Company. If electromechanical filters (made by Collins Radio Company) become available in 2- and 4-kcps bandwidths, with characteristics similar to the one reported previously (3-kcps bandwidth), the possibility of meeting the electrical specifications of the entire IF amplifier looks promising. The 8-kcps bandwidth would be built into the interstage networks

*This section of the Quarterly Progress Report of Division 3 is now available in a separate unclassified jacket. The same is true of the corresponding section of the Quarterly Progress Report of Division 3, dated 15 January 1953. This policy will continue in the future.

§For additional transistor-circuit research, refer to the reports of: Division 3, Group 33; Division 6, Group 62; Division 2, Group 24; Quarterly Progress Reports of the M.I.T. Research Laboratory of Electronics.

†A dagger denotes work done in close cooperation with members of the M.I.T. Research Laboratory of Electronics. Such work is also reported in the Quarterly Progress Reports of that Laboratory.

of the IF strip, with the two electromechanical filters switched in to reduce the band. It must be pointed out, however, that the appropriate filters are not yet available, and that the only one on hand has 26-db insertion loss in the pass band. Even if the loss in the final filters is reduced to 10 to 15 db, as suggested by the manufacturer, it requires at least one extra stage in the IF strip. For these reasons, some theoretical study is being given to alternate ways of achieving this three-bandwidth IF amplifier design. Two methods are outlined below.

One calculation has been made for the design of an IF amplifier, using point-contact transistors, to meet specifications close to those required in the actual receiver.

(1) Center frequency: 455 kcps,

(2) Three bandwidths as follows:

2 kcps between 6-db points; 10 kcps between 70-db points,
4 kcps between 6-db points; 14 kcps between 70-db points,
8 kcps between 6-db points; 26 kcps between 70-db points,

(3) Power gain: 60 db (using point-contact transistors),

(4) Maximum allowed Q of coils: 150.

As is evident from Fig. 35-1, multiple-order poles are used in the design, staggered to give a flat-top frequency response. The amount of staggering is varied by changing only the capacitors to give the three bands. Seven stages are required; the values given for one stage in the figure are typical. The interstage transformers are all physically realizable and all are exactly alike. Observe that the dissipative complementary networks in the emitter leads also provide a margin for low-frequency short-circuit stability. It is possible that, by allowing the coil Q's to go up to 200, the skirt behavior could be brought more nearly to the 80-db receiver specifications. It is emphasized that this design is presented only for general comparison purposes. It will probably not be built for testing, unless unforeseen problems arise in the design that involves the Collins filters.

The second IF design study being undertaken centers about a survey of crystal filters. It is quite certain that quartz crystals will be unsatisfactory for the wide bands involved, but EDT, DKT, and ADP crystals offer more attractive electrical possibilities. These crystals might possibly yield the two narrow bands (2- and 4-kcps) in a single lattice design which employs

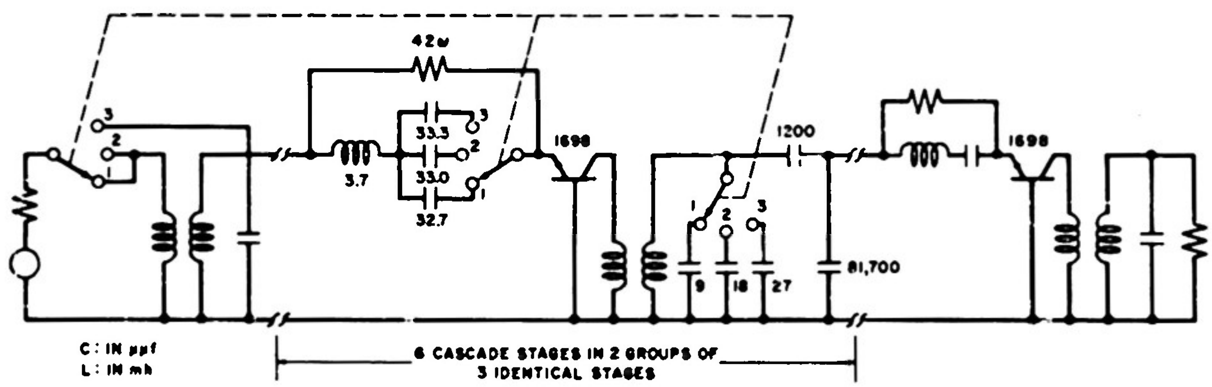

Fig. 35-1. Representative stages for three-band IF amplifier (power supply not shown).

only 4 crystals, with bandwidth changes accomplished by altering auxiliary capacitors. The passband insertion loss might be kept well below 6 db. Unfortunately, the mechanical, thermal and hygroscopic properties of these crystal materials may well render them unsuitable for this particular receiver application, but it is certainly worth while from a general transistor-circuitry point of view to continue the survey to a definite conclusion.

Further work has been pursued on the audio amplifier, using a single p-n-p junction transistor. When driven from a conventional 1N34 crystal-diode detector, the audio-stage power output was 16 mw. The efficiency was 20 per cent including the transformer loss. The power gain of the audio stage was 26 db. If p-n-p (alloy) junction transistors are to be used in the final design, this result indicates that the required 200 mw of power into the speaker can be obtained only by using special power transistors. The construction by this laboratory of a few such units is therefore being contemplated.

W. E. Morrow, Jr.
S. Schwartz
M. Boisvert
J. Craig

B. Small-Signal Parameter Measurements

Additional data have been taken on Raytheon and BTL junction transistors. Careful measurements of collector resistance (r_c) and base resistance (r_b) as functions of emitter bias current indicate the presence of a strong "Early effect."* These results call for caution in operating the units at extremely low bias currents. Measurements have been made also on a small number of RCA p-n-p junction transistors, which appear to have characteristics similar to those transistors made by Raytheon. The number of units measured in all cases has been much too small to obtain very accurate statistical results, but, for convenience, the data are being collected in an internal laboratory report.

W. E. Morrow, Jr.
S. Schwartz

II. SEMICONDUCTOR DEVICES

A. Diode Matrix

Previous experiments have indicated that the major fabrication problem in the matrix is the elimination of material at those cross-points where an open circuit is required. An alternate technique is now being studied as a means of applying the indium metal to the n-type germanium in only those regions where diode action is desired. After deposition of the indium, and subsequent alloying, the unit will be sawed or sandblasted as described in the previous report.

R. H. Kingston

B. "Channel" Studies

Brown and Shockley[1] have discussed the phenomenon of n-type conduction along the surface of the p-type base region in an n-p-n transistor. This effect, which produces a low-impedance path from emitter to collector, may be examined by use of the circuit in Fig. 35-2.

*J.M. Early, Proc. I.R.E. 40, 1401-6 (November 1952).

GROUP 35

If there is no n-type "skin" over the base region, a value of open-circuit emitter voltage given by

$$V_e = \frac{-kT}{q} \ln(1 - a)$$

may be predicted from the theory.[2] If, however, there exists a layer of n-type material at the surface, commonly called a "channel," the open-circuit emitter voltage will rise to a value near V_{cc}. The known sensitivity of the channel effect to water vapor suggested the following experiment. The transistor was placed in an atmosphere of dry nitrogen with vapor pressure controllable between 4.6 and 11 mm Hg. After the collector voltage was applied, the time required for the emitter voltage to reach a final equilibrium value was measured. This final equilibrium value was much greater than both the theoretical potential predicted for no channel and that observed for extremely low water-vapor pressure (less than 0.01 mm). Moreover, the mean time of buildup of the emitter voltage appeared to be an exponential function of vapor pressure. Specifically, the time of buildup varied from approximately 10 sec at 4.6 mm water-vapor pressure to approximately 1 ms at 11 mm. Further tentative results indicate a decrease in buildup time with decreasing transistor temperature.

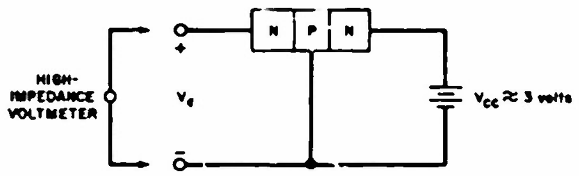

Fig. 35-2. Test circuit for "channel" effect.

R. H. Kingston

C. Diffusion of Copper in Germanium

Several workers in the field have suggested that the abnormally high diffusion coefficient of copper atoms in germanium[3] may be explained by assuming that the copper diffuses as a singly charged positive ion, through interstitial sites. To check this theory, a sample of germanium was saturated with copper, and then heated *in vacuo* to 825°C by the passage of a direct current through it. The resulting sample should show a constant resistivity over its length if the copper atoms are neutral, while, if they are positively charged, they should migrate to the negative end of the bar. There is no definitive result from this experiment yet, although the positive-ion picture does seem plausible from the first tentative data. If copper does behave as a positive ion at elevated temperatures, it may have a bearing upon the "forming" problem in point-contact transistors.

R. H. Kingston

D. N-P Junction Diodes Prepared by Fusion of Alloys into P-Type Germanium

Experiments have been conducted with the object of making an n-p junction diode with a back resistance of the order of 1 megohm. Alloys of Pb, Sn, and As (to provide the donors) were tried. The composition of these alloys was about 50:50 Pb and Sn with a few per cent (estimated between 1 and 10) of As. After the junctions were fused, they were etched electrolytically in 20 per cent NaOH to eliminate the conducting film (of the alloy) which shorts the junction. It was found that the diode could be improved by prolonged etching (up to about

1-1/2 hours). The best diode characteristics obtained were as follows: front resistance, 50 ohms; back resistance, 10K. These results were compared to those obtained with diodes using BTL n-type solder as the fusing alloy; the latter results showed a small improvement: front resistance, 50 ohms; back resistance, 25 to 30 K.

J.G. Mavroides
R.H. Kingston

E. Ultrasonic Excitation of Germanium Diodes

A back-biased germanium junction diode was subjected to an ultrasonic wave of frequency 400 kcps. It was hoped that there would be a noticeable variation of the bias current with pressure, and that this current variation could then be related to variations in the height of the barrier, or to changes in the width of the space-charge region, or both. Unfortunately, noise on one hand and magnetostriction of the nickel leads on the other have thus far masked all the possible (small) effects.

J.G. Mavroides

F. Forming

Efforts to obtain reliable correlation between the type of forming treatment, the size and shape of the formed region, and the collector characteristics of point-contact transistors continue to give negative results.

The forming circuit has been redesigned to provide better reproducibility of pulses and a wider range of forming parameters.

E. Rosenblum
J. E. Thomas, Jr.

G. Pilot-Plant Studies

Routine service of preparing germanium specimens, and general sample preparation, has continued.

A design of emitter and collector points using sheet beryllium copper and phosphor bronze, respectively, has been developed to give high mechanical strength and good control of relative point positions. Such a design may eliminate the need of a potting compound for many applications.

A number of the above units, having α greater than 5 and a frequency cutoff in the region of 5 Mc/sec, has been made. At present an attempt is being made to use very thin germanium samples to raise the frequency cutoff and to determine the effect of the germanium properties on the frequency response.

J. E. Thomas, Jr.
E. F. Dillaby
H. F. Priest

III. PHYSICS

A. Hall Effect and Resistivity Measurements

The new sample holder, mentioned in the previous Quarterly Progress Report, has been completed and seems to be operating satisfactorily. The outer chamber has been rebuilt to eliminate a 6-lead glass press which leaked at liquid-air temperatures.

A sample holder for room-temperature Hall effect and resistivity measurements has been built and operated successfully.

W. Krag

B. Measurement of Carrier Lifetime

1. Optical Modulation of the Conductivity (Optical-Injection) Method

In a rectangular sample 1 × 1 × 1.5 cm illuminated on one face, the decay of the excess conductivity was found to depart significantly from a simple exponential. This is the result of the presence of higher modes in the decay process. The general expression for the excess carrier density consists of a sum of terms, each representing one mode. The time-dependent term for each mode is

$$e^{-(\lambda + \nu_{l,m,n})t} \quad ,$$

where $\lambda = 1/\tau$ = reciprocal bulk lifetime; l, m, n = mode numbers for x-, y-, and z-directions, respectively.

The initial amplitudes of the various modes are determined by excess carrier density at the time the light is turned off. It can be shown that some of the higher modes have amplitudes comparable with the fundamental, and that they decay more rapidly than the fundamental. By a proper choice of sample dimensions, the higher modes can be made to decay reasonably rapidly, leaving only the fundamental; from the decay constant of the latter we can find λ. The initial amplitude of the fundamental can be enhanced by illumination of more than one face, and by making long the period of illumination.

R. J. Keyes
D. T. Stevenson

2. Pulse-Injection Method

A further increase in amplification is still being sought. A new non overloading amplifier and clipper designed in this laboratory will be built.

A phenomenon believed to be carrier extraction has been observed in an n-type sample with the present setup. After the large pulse has been applied to the sample and shut off, the resistance of the sample is higher than its thermal-equilibrium value and then decays back to its normal value. At a temperature slightly below room temperature, the signal disappears; and at still lower temperatures, carrier injection takes place. Extraction is observed only with one polarity of the extracting pulse. The extraction phenomenon is probably associated with the electrical contacts to, or with resistivity gradients in, the sample. It may provide a useful tool in the study of broad-area electrical contacts to germanium.

G. C. Hunt
G. I. Hitchcox
D. T. Stevenson

3. Measurement of Very Long Carrier Lifetimes

A modification of the Morton-Haynes Method[4] for measuring excess carrier lifetimes, described by W. H. Brattain,[5] has been used to measure a lifetime of the order of 1500μsec in a sample of germanium supplied by Bell Telephone Laboratories. A large flat area is etched and then sandblasted, with the exception of a small area where the collector point is to be placed. The collector is "formed," and the AC signal at the collector due to a chopped line-source of light is measured as a function of the distance from the collector to the light. The theoretical response is

$$I_{collector} \sim \frac{K_1(\frac{r}{L})}{(\frac{r}{L})} ,$$

where

K_1 = Hankel function of the second kind,

r = distance from collector to light,

$L = \sqrt{D\tau}$, diffusion length of excess carriers,

D = diffusion constant,

τ = lifetime.

For accurate results at these long lifetimes, the chopping frequency must be reduced from our present 90 cps.

Recombination at the sandblasted surface is high enough to make zero the excess carrier density at the surface. The effect of the surface is then always a known factor, which is not true for the etched surface used in the method described by Valdes.[4] The small etched region about the collector point allows, so to speak, a sampling of the density slightly under the surface, and also allows the "forming" of the collector point.

R. J. Keyes
D. T. Stevenson

4. Electrodeless Method

When a piece of germanium is inserted into the coil of an RF resonant circuit, it was reported previously that the Q of the circuit depends significantly on the conductivity of the germanium. A transient change in conductivity should therefore cause a transient change in the Q of the system. This phenomenon offers the possibility of providing a measurement of minority-carrier lifetimes.

It has been verified that this transient change can be detected in the output of a vacuum-tube oscillator when the germanium is inserted in the tank coil. For the present, the method appears to be limited to germanium of lifetime greater than 500μsec.

J. E. Thomas, Jr.
N. Janet

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C. Semiconductor Theory*

1. Transient Response

The decay of excess carriers in a bar of semiconducting material with linearly varying lifetime has been investigated theoretically, and the carrier concentration has been obtained in terms of Bessel functions. Further numerical work on the effective decay time is continuing.

S. F. Neustadter

2. Dielectric Relaxation and Carrier Motion in Semiconductors

An exact analysis, based on Laplace transform methods, has been made of the linearized equations describing the motion of excess carriers in semiconductors. The familiar formulae for the relaxation time, effective lifetime, mobility, and diffusion coefficient are again found, the latter two with small correction terms. It is shown that, for an infinite medium, excess densities acquire, within the relaxation time, a dipole-moment density

$$\delta P = K\left(\frac{\delta\sigma}{\sigma}\right)\vec{E} \quad ,$$

where K is the dielectric constant, σ the conductivity, and $\vec{E}$ the field intensity.

J. Keilson

D. Photoeffects in Germanium

The last period has been spent in installing and aligning the Perkin-Elmer monochromator and associated recording equipment. Components have been obtained for an accessory optical system to be used in conjunction with the monochromator and high-pressure apparatus..

D. M. Warschauer
F. Reif
R. L. Powell

E. Electroluminescence

A few preliminary observations have been made of the light output of electroluminescent phosphors when they are subjected to pulses of voltage. Methods of preparing the phosphors for test are being worked out. Measurements of the light output as a function of voltage, pulse duration, and wavelength will be undertaken.

S. Autler
W. F. Roat

*Considerable related theoretical work is being carried out by six members of the group: Drs. Kleiner, Koster, Meckler, Parmenter, Pratt and Schweinler. For the time being, they are continuing research along their original lines, which stem from the work of the M.I.T. Solid State and Molecular Theory Group (ONR contract N5ori-07856). To maintain continuity, and avoid duplication, they will continue to report their work as members of Lincoln Laboratory in the Quarterly Progress Reports of the aforementioned group. When convenient, their work will gradually be transferred to the Lincoln Laboratory reports.

F. Ferrites*

1. Physical Properties

Preliminary observations were made with a ferrite in a cylindrical cavity. Photographs were taken of the resonance split to show that the system was workable. Design difficulties with the cylindrical cavity necessitated a change to a square cavity, resulting in higher Q and greater separation of the undesired modes. The new cavity is being tested.

B. Lax
P. E. Tannenwald
J. O. Artman

2. Theory

Additional theoretical results were obtained for the impedance expressions associated with a cavity containing a ferrite and having either one or two inputs. An interesting result for the single-input cavity was obtained in terms of the natural rotating modes of a degenerate cylindrical cavity. As a consequence of this, an equivalent circuit of the system was derived. The results were presented at the 1953 I.R.E. Convention.

B. Lax
A. D. Berk

Calculations have been carried out for the problem of rotating the plane of polarization of an electromagnetic wave using two slabs of ferrite, and curves have been plotted exhibiting the dependence of angle of rotation on applied fields. A solution has been obtained to the equation for the magnetization of a ferromagnetic material which does not neglect contributions of second (and higher) order in small quantities. This work is being continued to determine possible experiments to test the implications of this solution.

M. Balser
B. Lax

IV. CHEMISTRY, PHYSICAL CHEMISTRY, METALLURGY

Electrochemical Production of Germanium Hydride

The apparatus for the determining of GeH_4-H_2 ratios is working satisfactorily. Further electrochemical studies are under way. A more accurate value for the refractive index of GeH_4 has been obtained.

P. H. Robinson
M. Green

Electrochemical Studies on Semiconductors

A preliminary Tafel curve has been obtained for n-type germanium.

M. Green
S. Zwerdling

Purification of Germanium Tetrachloride

The $GeCl_4$–$AsCl_3$ vapor-liquid equilibrium system has been fully investigated, and

*A considerable amount of work on ferrites for memory-storage purposes has been done by Drs. Gold and Kulin of Group 35, in cooperation with the program of Group 63. This work is reported by Division 6 (Group 63) in its section of this Quarterly Progress Report and is therefore not duplicated here.

is now complete. The relative volatility is constant at 2.5 from 5 mole per cent $AsCl_3$ (liquid) down to 10^{-6} mole per cent.

M. Green
J. Kafalas

Analytical Chemistry

Final analyses on arsenic unknowns are being made, using the light-scattering technique.

The chemical analysis of ferrites continues. Faster analytical procedures for iron, nickel, and zinc are being sought.

J. Lowen
R. Smythe
M. Green

Effect of Adsorbed Gases on Noise Characteristics

The gas-handling apparatus is being tested.

Our technique for cutting germanium filaments now appears to be satisfactory.

The modified Penning gauge is being calibrated, and seems to be operating properly.

T. G. Maple

Diffusion of Hydrogen in Germanium

The diffusion of hydrogen through a germanium membrane is being investigated. The hydrogen is generated electrolytically at the surface of the metal. The apparatus for this experiment has been completed.

Electrolyses have been performed with a germanium cathode in acid solution. The object of these experiments has been to determine whether such treatment causes hydrogen to enter the germanium and, if so, whether this would affect the lifetime of excess carriers. No conclusive results have been obtained.

M. Green
D. T. Stevenson
R. J. Keyes

Production of Germanium Metal

Production of germanium metal has continued, and some additional data on the reaction kinetics have been obtained. The present experiments are aimed at finding the critical minimum rate of flow of hydrogen. With the completion of these data, as well as some on the effect of pressure, this phase of the work will be terminated.

A production unit, designed only for producing metal, is under construction. It will be used to keep a supply of metal available for crystal pulling.

G. L. Priest

Purification Procedures

Additional data have been obtained on zone melting in high vacuum. Present indications are that oxygen might play a more important role in purification procedures and semiconductor properties than had been expected previously. General definitive experiments to confirm this indication are presently being started.

Vacuum distillation of metal has been delayed by vacuum problems, but this is now progressing satisfactorily.

In order to eliminate the use of graphite boats, the induction-heated, gas-atmosphere zone melter has been reconstructed, using carbon heating rings with a quartz boat. Operation has proved to be very satisfactory, giving narrow, reproducible zones. The system yields resistivities above 35 ohm-cm at room temperature, after 8 passes.

G. L. Priest
H. F. Priest
E. S. Candidus

Crystal Growing

The vacuum crystal furnace has been tested for leaks and found to be tight. It is presently in the shop for final mechanical assembly.

The gas-atmosphere crystal puller has been used regularly. The effects of a number of variables are being considered at present. The size of crystal that can be pulled is related to crucible geometry; various geometries are therefore being tried. The effects of rotation, oscillation, and seed orientation are being examined critically, although the necessary physical-characterization studies are so time-consuming that the over-all program progresses slowly.

Rather detailed studies on lineage grains in crystals are being made, using x-ray techniques.

An attempt is being made to meet device requirements for special doped materials by obtaining a number of tubes, crucibles, and pull mechanisms for the crystal puller. Separate parts are required to avoid cross-contamination.

H. F. Priest
G. L. Priest
E. S. Candidus

Surface Studies

The cathetometer for the gas-adsorption surface-area unit has not yet been received, but some additional work has been done on the gas-measuring system.

Preliminary gravimetric studies will be made using quartz spiral-spring balances, and final specifications for the springs are now being established.

The electrolytic surface-measurement equipment has been run for a sufficient length of time to give preliminary data of satisfactory reliability. A final reassembly of the electronic and electrical measuring equipment is being completed, and the detailed examination of surfaces prepared in a variety of ways is being undertaken. Present data indicate that a monolayer of oxide forms on germanium at room temperature at a very rapid rate.

S. Naiditch

Electron-Diffraction Studies

Preliminary low-speed electron-diffraction peaks have been found from germanium

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after outgassing at 700°C for some time. Interpretation of the results, which include some data on the work function of the crystal surface, will be undertaken as more information is accumulated.

The tube for studying the photoelectric work function of germanium is being changed slightly, because the glass blower was unable to produce the original design with sufficient precision. Most of the rest of the equipment for these studies is assembled, so that completion of the tube envelope will enable work to proceed rapidly.

Brown University
H. E. Farnsworth (supervisor)

Diffusion Studies

New samples of germanium have been centerless-ground to give accurate cylinders, and work has progressed on radiochemical procedures designed to give the best data from the copper-diffusion studies. Additional diffusion studies on the systems involved in fused junctions are being considered.

W. Z. Leavitt
J. F. Wenckus

Ferrites

The hydrothermal-synthesis studies have successfully produced crystals of about 0.5-mm length of crystal edge, but it has not yet been possible to control their composition.

Further studies on the production of ferrite powders by spray-drying appear to show considerable promise. A small, high-temperature experimental model is being designed to produce powders of spinel structure and of controlled composition.

H. F. Priest
G. C. Kennedy (consultant)
W. Z. Leavitt
J. F. Wenckus

Electroluminescent Materials

Additional samples of polycrystalline electroluminescent phosphor have been prepared, and the initial group having about 10 variations is nearing completion.

The reconstruction of the single-crystal furnace was delayed for two months by inability to obtain Kanthal wire, but the furnaces have now been rebuilt and are ready for use. A special Globar furnace for this purpose, which will allow for higher temperatures, has been designed and components have been ordered.

H. F. Priest
G. L. Priest

Metallurgical Studies

Considerable effort has been spent on ferrite samples submitted by Group 63. These were memory-cell ferrites, for which it was necessary to develop techniques for examination of microstructure. Preliminary conclusions show that microstructure is intimately associated both with magnetic properties and with preparation techniques.*

*Refer to footnote on p. 73.

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GROUP 35

Studies are under way on the phase relationships in the copper-germanium system.

Preliminary work has been started on the problem of ohmic contacts by alloy soldering and fused junctions in general. It is planned to expand this effort considerably in the next few months.

S. A. Kulin

X-Ray Studies

Work continued on setting up the x-ray laboratory. Studies have been carried out with the 180° spectrometer to establish its sensitivity. Some crystal-orientation work on lineage grains is under way (see section on Crystal Growing).

The double-crystal spectrometer is nearly ready for use, and studies on lineage and mosaic structures will begin in the near future.

S. A. Kulin
A. Kurtz

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SYSTEMS ENGINEERING

GROUP 36

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SYSTEMS ENGINEERING
GROUP 36

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I. INTRODUCTION

Group 36, Systems Engineering, was established in January 1953 to undertake a number of tasks involving development, procurement, and installation of special communications research facilities for Division 3. The group also has responsibility for the low-frequency phase-coherent array studies being carried on for the Department of State (DIC 6924).

Group 36 was assigned the responsibility for coordinating the Lincoln Laboratory activities and commitments on PROJECT COUNTER CHANGE with those of the Bell Telephone Laboratories and the Western Electric Company. These commitments included, among other things, the design and fabrication of two special teletype systems and transmitter conversions.

Much of this work was done by personnel assigned temporarily from Groups 33 and 34, but is reported here.

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II. VHF NOMAC TELETYPE SYSTEM

Equipment for six 2-channel VHF NOMAC teletype systems is presently being constructed by Group 34. The equipment being built is essentially a duplicate of the VHF NOMAC system described in the 15 January 1953 Quarterly Progress Report of Division 3, (pp. 100, 101). A block diagram of the equipment for each system is shown in Fig. 36-1. The major changes from the previously built VHF NOMAC system include a modification of the band-pass filters to a 6-pole Butterworth design, as described in the 15 January 1953 Quarterly Progress Report, and accommodation in the receiving terminal equipment for dual-diversity reception. As shown in the block diagram of Fig. 36-1, the dual-diversity signals in the receiving terminal equipment may be added either at the output of the adding circuit following the first set of averaging filters, or at the output of the envelope detectors following the second set of averaging filters.

The design of the equipment has allowed for operation in the frequency range from 31.00 to 38.00 Mc/sec, without changing any component with the exception of plug-in crystal units. However, changing the frequency of operation of any one system will involve a realignment of the VHF mixer, VHF power amplifier, and receiver front end.

It is planned to have equipment for four systems, as well as the instruction manual, completed by 1 May and the equipment for the remaining two systems completed by 1 June.

R.S. Berg

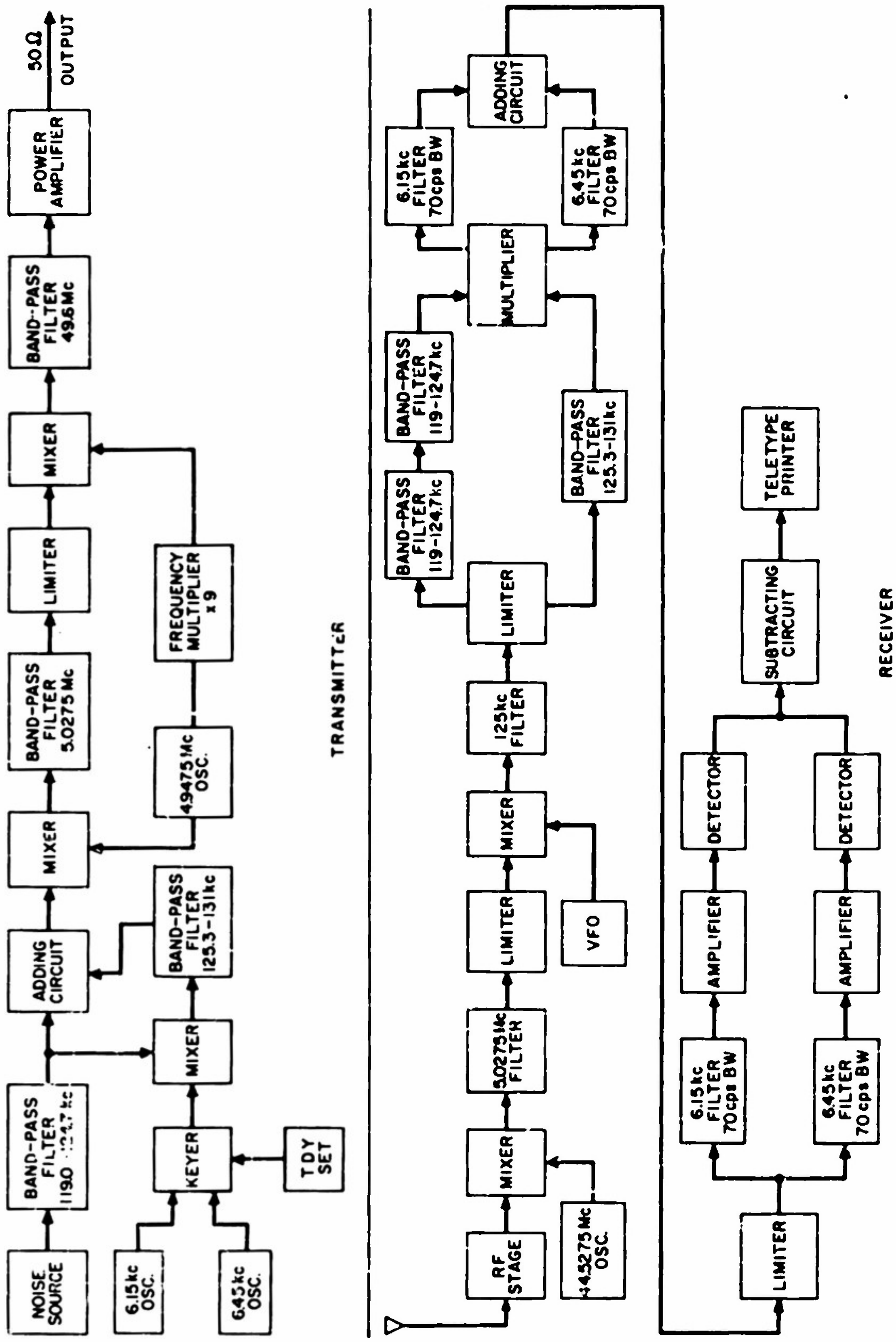

Fig. 36-1. Block diagram of VHF NOMAC teletype system.

III. ANTENNAS

In connection with the investigation of VHF-scatter circuits, studies have been made of various types of antennas and arrays of antennas. The following were considered: rhombics, arrays of rhombics, paraboloidal sections, arrays of corners, broadside arrays, and array of helices.

Work has gone forward through the design stage on an array of eighteen corners designed for a power level of 500 kw in the 25-Mc to 50-Mc frequency domain.

A. J. Poté

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IV. VHF FREQUENCY-SHIFT TELETYPE SYSTEM

As one of the communication systems for the PROJECT COUNTER CHANGE experiment, we have designed a high-precision frequency-shift teletype system to be used with the VHF scatter links. We are constructing six complete equipments consisting of one dual-diversity receiver and one transmitter exciter each. We are also constructing four test oscillators. The receiver consists of: a front end; two IF strips; a converter; a power supply; and a Sorenson regulator. The transmitter exciter consists of: a low-frequency unit; a high-frequency unit; and a power supply.

We have completed six receiver power supplies, twelve IF strips, six converters, and one front end. Subchassis for the remaining front ends are 95 per cent completed. Of the exciter, we have completed six power supplies and six low-frequency units. The low-frequency units are currently undergoing modification.

J. Granlund
D. Gray
W. Morrow

V. AN/FRT-6A TRANSMITTER MODIFICATIONS

Part of the group operations was the installation of an AN/FRT-6A transmitter in the Balloon Building at the Lexington Field Station. This transmitter is capable of 40 kilowatts of power output and is normally designed for use in the 4- to 26-Mc band. A Collins Radio Company modification was made upon this transmitter by personnel of the Lincoln Laboratory and E. C. Page Consulting Radio Engineers, converting the transmitter to use in the 30- to 38-Mc band. A folded dipole was installed on the Balloon Building. This was used as a load for operational tests.

Glenn Mellen

VI. HIGH-POWER PULSE TRANSMITTER SYSTEM

Although every effort was made to complete assembly of the transmitter by the end

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of March, delays in receiving components of the power amplifier cavities and water-cooling system, combined with difficulties in recruiting an adequate technician staff, have resulted in postponement of the proposed completion date until the end of April. The special water-cooled load being manufactured by Continental Electronics Manufacturing Company is long overdue and is not now expected until some time in April.

Work during this period has therefore consisted mainly of assembly of electrical components into the two-power amplifier cavities; assembly, wiring and preliminary testing of the power supply over-all control circuit; assembly and wiring of the entire 460-volt load-center and circuit-breaker system; intercabling connections among the various bays of the transmitter; assembly and wiring of part of the intermediate power amplifier cascade, and assembly of numerous small parts, hardware, locks, interlocks, and connections.

B. J. Cosman