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FINAL REPORT

UHF AIR-GROUND COMMUNICATION BEYOND HORIZON

Prepared For

AIR FORCE CAMBRIDGE RESEARCH CENTER
Air Research and Development Command
Laurence G. Hanscom Field
Bedford, Massachusetts

[redacted]

ABSTRACT

Studies were conducted of radio wave propagation phenomena with application to USAF air-ground communications, pursuant to a report "The Possibility of Extending Air-Ground UHF Voice Communications to Distances Far Beyond Radio Horizon," by L. A. Ames, E. J. Martin and T. F. Rogers, AFCRC Confidential Technical Report TR-56-111, ← [illegible] June 1956.

The results of the studies indicate that UHF tropospheric scatter may be employed successfully and with advantage to extend the range of communications with high altitude aircraft to beyond 400 miles with very high reliability even in the Arctic. A grid network of ground relay stations would be employed to extend communication coverage still further. Typical SAC and MATS missions are discussed using the proposed scheme.

A new set of UHF tropospheric scatter loss curves is derived from more recent data. Predicted performance is compared with very recent measurements of AFCRC in March 1957, using a B-47 aircraft flying at 40,000 feet, with good agreement.

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Results of system studies suggest the use of SSB, vertical polarization, and dual spaced diversity, even in the aircraft where the two spaced antennas would be near each wing-tip.

Recommendations for further work are given. Included is a recommendation for immediately implementing an experimental operational test of one typical grid setup centered near Boston, using recommended system parameters and components. Further measurements and studies are needed of propagation phenomena, particularly on fading characteristics in the Arctic and at high altitudes for very high reliability levels.

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I. INTRODUCTION

I. INTRODUCTION

Certain studies have been conducted which appear to make promising the application of UHF tropospheric scatter to USAF air-ground communications. These studies were conducted by AFCRC and the results were reported by L.A. Ames, E. J. Martin and T. F. Rogers in an AFCRC Technical Report of June 1956, entitled "The Possibility of Extending Air-Ground UHF Voice Communications to Distances Far Beyond the Radio Horizon". Further studies were indicated on propagational and system aspects and these have been undertaken under this program.

Results of propagation studies are given in Section II primarily those of tropospheric scatter. A new set of scatter loss curves is presented, based upon a wealth of more recent data and their evaluation which takes into account aperture-medium coupling losses and terminal obstructions and profiles. Extensions of the curves to the Arctic and Tropics are outlined. Included are results of studies of VLF, LF, HF regular layer propagation, VHF ionospheric and tropospheric scatter, and VHF meteoric scatter.

Section III contains results of system design studies, including propagational factors affecting system design, predicted performance and some details of the equipment (airborne and ground) pertaining to antennas, transmitters and receivers.

How the proposed system will improve USAF air-ground communications is given in Section IV. Outlined are the broad plan and typical SAC and MATS missions.

Section V is a summary and contains recommendations based on results to date.

Appendix A details methods used in obtaining line-of-sight from meteorological data.

Appendix B outlines scatter loss calculation methods.

In Appendix C is contained a detailed analysis of AFCRC flight data at 220 mc, 38,000 - 40,000 ft. altitudes from flights made in March 1957. Our curves are used to predict signal levels and height gain. Agreement within 3 db of predicted and measured values is shown.

Appendix D is an extensive bibliography used during the study.

II. PROPAGATION STUDIES

A. INTRODUCTION

A great deal more data on UHF tropospheric scatter has become available in the last year and a half. It is believed desirable to review these data and reassess previous transmission loss curves.

Since we believe the effective distance concept is applicable here, methods for deriving line-of-sight values from meteorological data are required for other than analytic profiles. These were outlined and data permitted obtaining line-of-sight contours for three climatic regions, Arctic, Midlatitudes and Tropics.

To properly assess tropospheric scatter for the application intended, other propagation modes were studied, emphasizing the Arctic auroral zone areas. These include studies of VLF, LF, HF, VHF ionospheric and meteoric scatter propagation.

These topics have been studied and the results of the study program are given below.

B. TROPOSPHERIC SCATTER PROPAGATION STUDIES

1. TRANSMISSION LOSS EQUATIONS

For tropospheric scatter circuits with elevated terminals such that transmission takes place well beyond radio horizon, it is assumed that the principal transmission losses are (a) that associated with free-space attenuation between antennas separated a total transmission distance d and (b) those associated with a scattering loss referred to free space evaluated for the "scatter distance" d_s beyond the radio horizon. The total loss has a steady term associated with (a) and a fading term associated with (b). The total path distance $d = d_o + d_s$ where $d_o = d_R + d_T$ is the line-of-sight distance to be subtracted from d to give the scatter distance beyond horizon d_s and d_R and d_T are the horizon distance at the receiving and transmitting terminals. (In what follows, the spherical geometry is such that great circle distances and ray distances are approximately equal).

At a given total transmission path length d and for a fixed antenna height at one terminal (say transmitter) this "effective distance" concept leads to a second concept of "height gain". This comes about in that as the aircraft height (receiver terminal) increases the line-of-sight term d_R increases which increases d_o and reduces d_s. We shall see that the scatter loss decreases with d_s and consequently the total loss, resulting in an increased received power, or "height gain".

To assess the characteristics of tropospheric scatter transmission[1], we shall use Norton's[2] transmission loss definition suitably modified. The total transmission loss L is then the ratio of transmitter power P_T to the received power available at the receiver

$$L = \frac{P_T}{P_R} \quad . \tag{1}$$

or in decibel notation

$$L_{db} = 10 \log \frac{P_T \text{ (watts)}}{P_R \text{ (watts)}} = P_T \text{ (dbw)} - P_R \text{ (dbw)} \tag{2}$$

The total transmission loss for a system may be considered due to several terms (in decibel notation) as follows:

$$L = L_{f_o} + L_s + L_c + L_A - G_T - G_R - G_D + F_T \tag{3}$$

in which

L_{f_o} = free space loss between isotropic antennas (db);

L_s = scatter loss in db below free space between broad beam antennas, a function of scatter distance d_s and frequency f;

L_c = aperture-medium coupling loss, important for very narrow beam antennas and large scatter distances (db);

L_A = ohmic losses of feeders, filters, diplexers, etc. (db);

G_T, G_R = plane wave antenna gains with respect to an isotropic radiator (db);

G_D = diversity gain, due to use of several orders of diversity, giving a shift in median signal (db);

F_T = fading factor, measured with respect to the median value (db); consists of a slow fading factor F_s (due to distribution of hourly median values) and a fast-fading factor F_f (due to short term fading within an hour).

These terms of equation (3) will be discussed further presently. The quantity $G_T + G_R - L_c$ would correspond to Norton's "path gain G_p". The quantity $L_{f_o} + L_s$ would correspond to his basic transmission loss between isotropic antennas L_b.

We shall give our estimated curves of L_s vs d_s below. These were obtained by analysis of data from several sources. The data for the most part give distributions of hourly median values of P_R for known values of P_T, d, L_A and G_T and G_R. In this case distribtuions of L_s hourly median values may be obtained from (3) with $F_T = F_s$, $G_D = 0$ db and calculations made for L_{f_o} and L_c. The appropriate values of d_s are determined from path profiles.

Assuming the estimated curves of L_s vs d_s thus derived, system performance of other circuits may be predicted by the use of (3). The performance may be depicted by curves of P_R vs d and, by comparing results with required received power, maximum range or margin for a given range may be determined.

We shall be interested in estimated performance in three general climatic areas, the Arctic, temperate Midlatitudes, and the Tropics. For these areas we shall give estimates of line-of-sight distances and the scatter losses, leading to overall system transmission losses.

2. REFRACTIVE INDEX PROFILES AND LINE-OF-SIGHT DISTANCES d_o

Refractive-index profiles may be used to estimate the radio line-of-sight distances d_T, d_R and d_o by methods given in Appendix A. The profiles may be those based upon weather balloon soundings or on other assumed profiles. Using actual Weather Bureau data of Ratner the profiles are used to obtain d_o by a piece-wise summation of incremental line-of-sight terms. The reference profile of Fannin and Jehn is analytic and the value of d_o may be calculated by an approximate integral.

The index profiles are shown in Appendix A. The resulting line-of-sight values of d_o are shown in Figure 1, for a ground antenna height of 100 ft. Also shown in Figure 1 is the curve of AFCRC; it is understood that this curve is based* upon d_T (or d_R) in miles $= \sqrt{kh_{ft}}$ where h_{ft} is the altitude in feet. The value of k was 2 at the surface (4/3 earth), 1.5 at 60,000 feet and approximately 1.75 at 30,000 feet and smoothly interpolated in between.

In final curves of predicted performance, the line-of-sight curves of Figure 1 labelled "Arctic", "Midlatitudes" and "Tropics" are used for values of d_o appropriate to those regions. To be noted is an important difference in d_o, especially at high altitudes, for the various regions.

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* E. J. Martin, AFCRC, private communication.

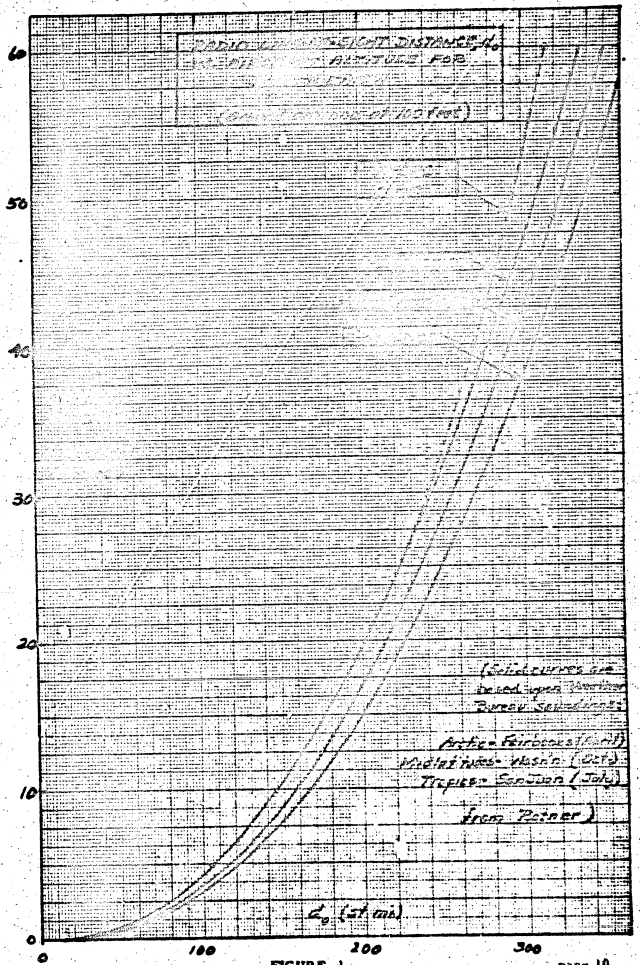


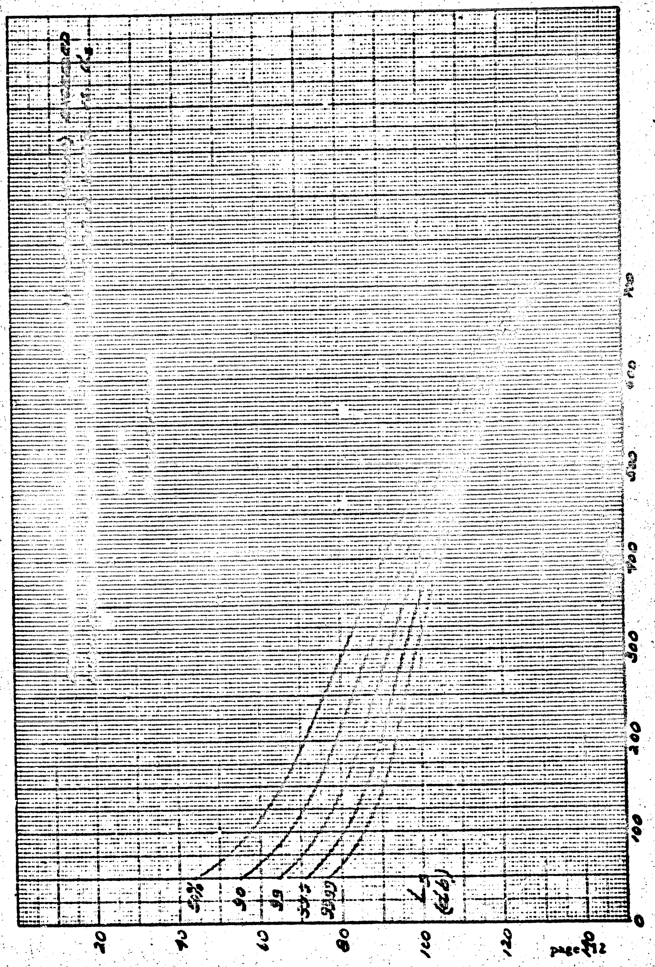
FIGURE 1

3. ESTIMATES OF L_s vs d_s

Our estimated curves of scatter loss for broad beam antennas L_s vs the scatter distance d_s well beyond horizon for antennas on the surface are given in Figure 2. The curves give the hourly median values of L_s exceeded for 50, 90, 99, 99.9 and 99.99% of the hours of a year. They are based largely upon Lincoln Laboratory 400 mc data[3] and thus are assumed applicable for Midlatitudes. The data were analyzed as suggested above and outlined in Appendix B, with due account given to terrain profile near the terminals and aperture-medium coupling losses. Other data from RCA, CRPL, BTL and Syracuse were employed in making our estimates.

In our analysis of the data, we conclude that there is a frequency dependence of L_s in that L_s increases according to the first power of frequency, so that L_s is 10 db higher at 4000 mc than our curves at 400 mc. This assumption was recently substantiated by Chisholm et al[4] of Lincoln Laboratory in comparing signal levels at 400 and 2000 mc on 188 and 350 mile paths. A frequency dependence was put forth by Bullington[5] (by reference to his Figure 2).

A comparison may be made between our estimated yearly median value of L_s vs d_s curves with those of others which have been published. As mentioned above, Bullington's curves[5] show a frequency dependence according to the first power of f, as do ours. However, his distance dependency varies with d_s^6 such that each doubling of distance increases L_s by 18 db. His values of L_s at longer distances are much less than ours. The curve of

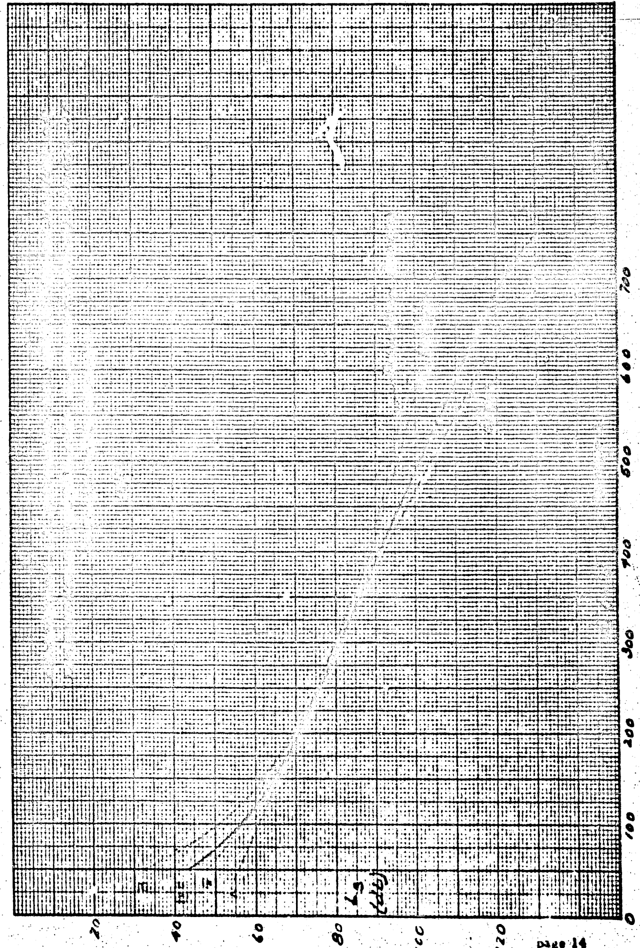


Gerks[6] assumes no frequency dependence and a slope of about 12 db per 100 miles. Norton's[7] empirical curve assumes no frequency dependence and at 600 miles his value is less than ours. The curves of Mellen et al[8], Morrow[9], and the winter values of Altman[10] show no frequency dependence. The curve of Davidson and Potè[11] is similar to that of Gerks but has a slope of 13 db per 100 miles; the values at 100 miles are identical at 57 db below free space. Davidson and Potè state the rate may be as low as 11 db/100 miles in semi-tropical, over-water paths and as high as 15 db/100 miles for dry arctic conditions, the value at 100 miles possibly varying with the attenuation. Their curve is frequency independent.

RCA's curve[12] uses values of L_s 6 db greater than Bullington's and corresponds to the highest expected monthly median losses. Their estimate is also frequency independent.

In the 300-400 mc region, all of our curves agree within three db in the 200-350 mile distance range.

A set of four estimates of L_s vs d_s is shown for comparison in Figure 3. Shown are Bullington's curve for 300 mc (derived from his Figure 2 of reference 5), that of Norton et al typical of afternoon hours (derived from their empirical curve of L_{b_m} - 10 log d_{mi} - 20 log f_{mc} vs d - d_R - d_T, Figure 7 of reference 7) and that of Gerks (derived from his Figure 2, of reference 6). Also shown unlabelled is our curve for 400 mc. These estimates apply to Midlatitudes.



In summary, our estimates of L_s vs d_s give values of L_s which apply:

a) for broad beam antennas,

b) for an hourly median value exceeded 50% of the hours of a year,

c) for temperate midlatitude regions,

d) for antennas on the surface ($d = d_s$),

e) for 400 mc (Figures 2 and 3); for other frequencies, say 400-4000 mc, add 10 log f_{mc} - 26 db,

f) for d_s exceeding about 50 miles.

4. LONG-TERM FADING FACTOR, F_s

Estimated hourly or long term distributions of the hourly median value of L_s are shown in Figure 4 for various scatter distances d_s. The distributions are db Gaussian.

The estimated hourly distributions of the slow fading factor or fading depth F_s i.e. the departure of L_s for a given percentage of the hours from the median value, are shown in Figure 5. The distributions of F_s are also db Gaussian. Note that the values of F_s decrease with increasing distance for a given percentage of the hours, as observed experimentally on the Lincoln Laboratory test circuits.

The scatter distance dependency of F_s for a given fading percentage (long term reliability) is shown in Figure 6. Beyond 150 miles, the values of F_s decrease with increasing d_s approximately according to log d_s.

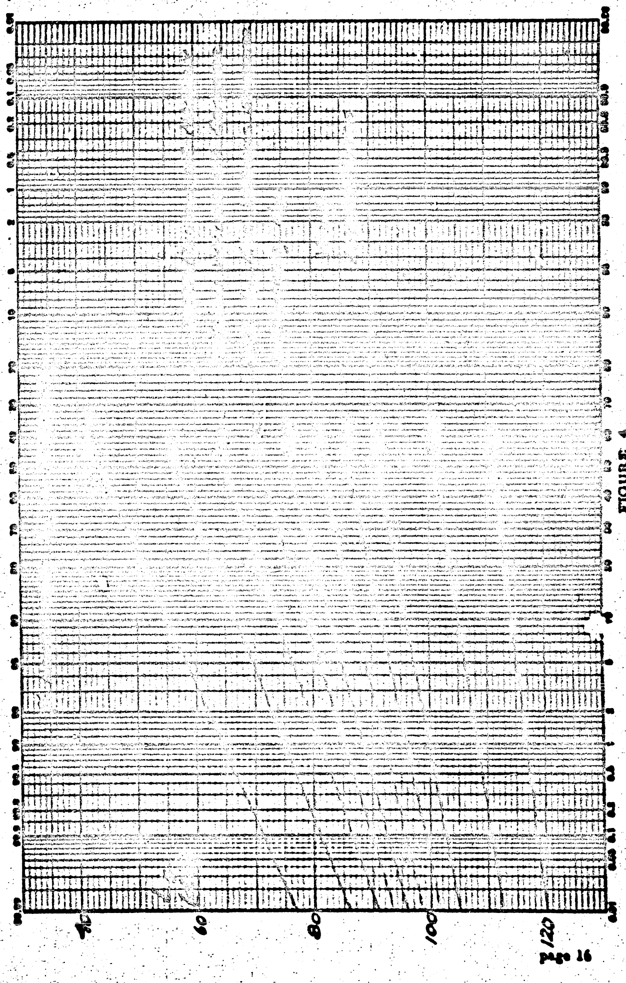


FIGURE 4

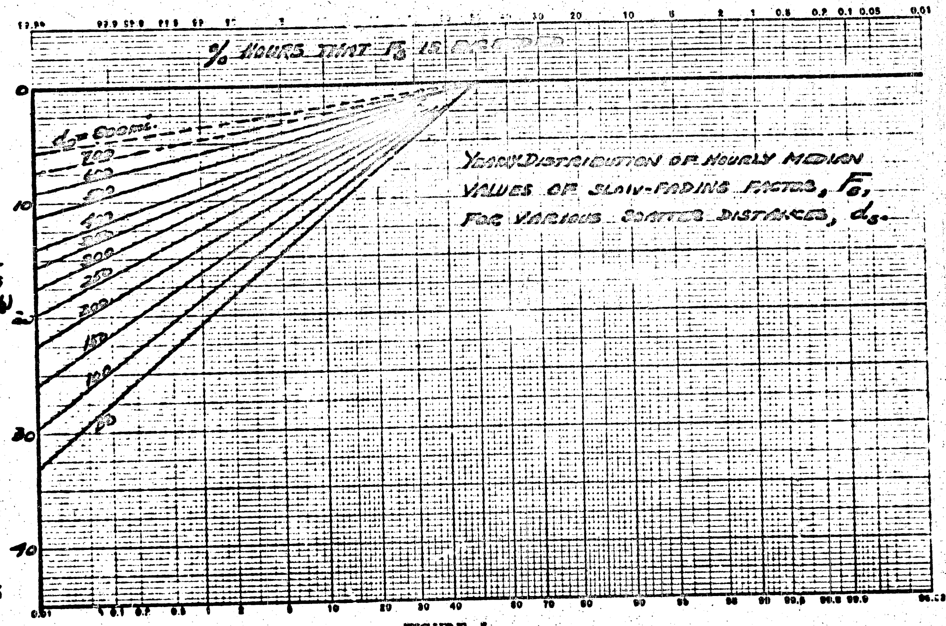


FIGURE 5

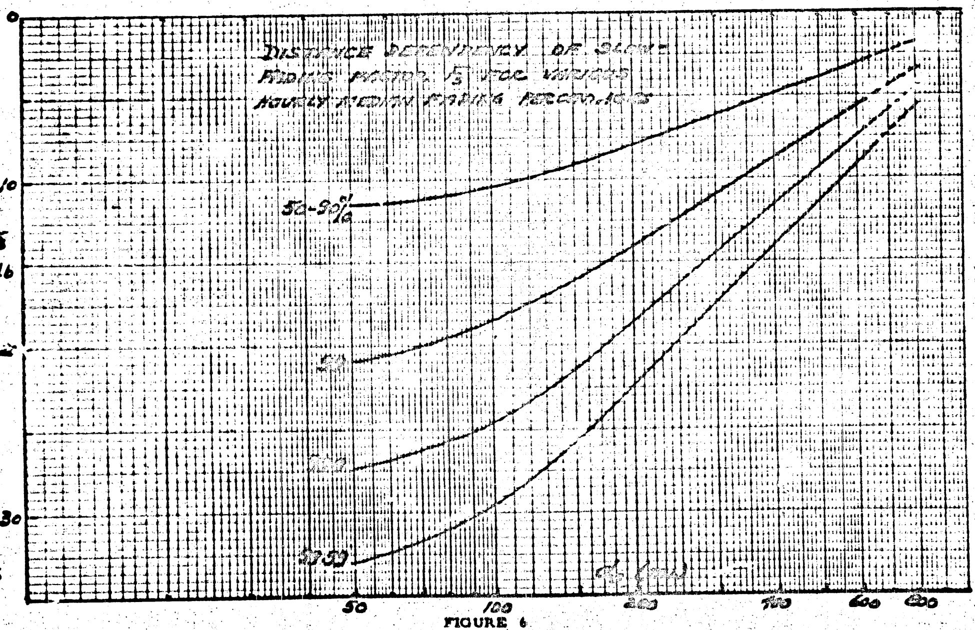


FIGURE 6

These distributions differ from those of Gerks (his Figure 4 of reference 6) in that his distributions are independent of distance. The curves of Mellen et al[8], Morrow[9] and Altman[10] show a decreasing slow-fading depth with distance.

5. SHORT-TERM OR FAST FADING FACTOR F_f

For single antennas, the fading for short periods of time (less than an hour) is Rayleigh distributed. When more than one stage of diversity is used the fast fading factor F_f is decreased and there is an increased shift in the median value, or diversity gain, G_D. For increasing order of diversity, the instantaneous distributions flatten out from the Rayleigh distribution and become more nearly db Gaussian.

For combiner diversity, we use the curves of Staras[13] which are slightly different from those of Mack[14]. The results are shown in Figure 7 for n = 1, 2 and 4 for distributions of signal-to-noise referred to the median value for n = 1. Shown are values of diversity gain G_D (4 db) and F_f (14.5 db) for n = 2, 99.9%.

6. TOTAL FADING RANGE F_T (RELIABILITY)

The distributions of signal-to-noise for fast fading become more nearly log normal or db Gaussian as the order n of diversity increases. Assuming the fast fading to be log normal, we obtain the total fading factor $F_T = \sqrt{F_s^2 + F_f^2}$ where each of the F factors are in db. The curves of F_T resulting from use

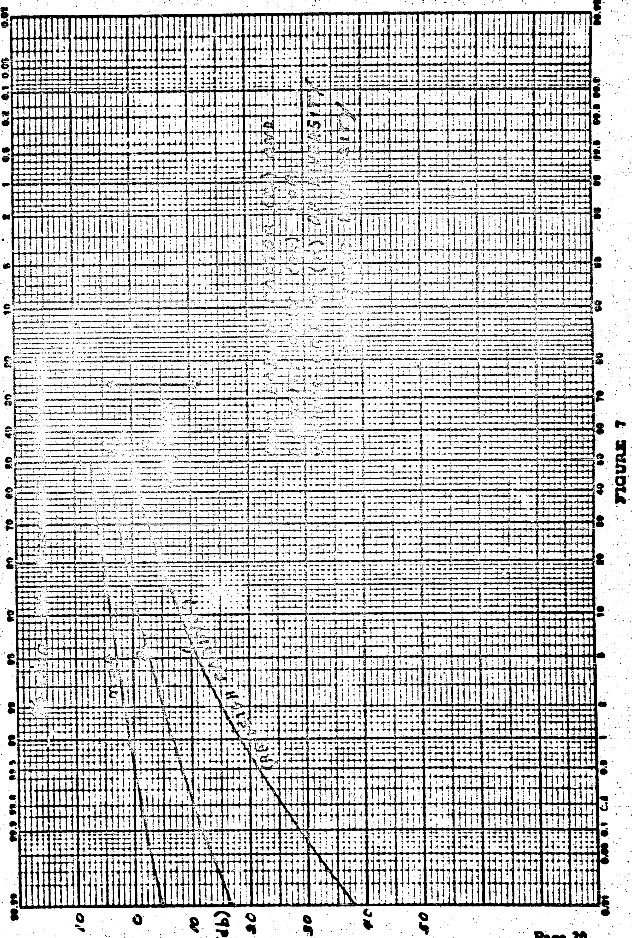


FIGURE 7

of F_s in Figure 5 and for 2-stage combiner diversity are shown in Figure 8. The diversity gain G_D is then 4 db.

7. APERTURE-MEDIUM COUPLING LOSS, L_c

The phenomenon and estimates of aperture-medium coupling loss L_c were pointed out by Booker and deBettencourt[15] for narrow beam antennas. This occurs when the antenna beams are narrower than the scattering polar diagram. The values of L_c are more important the greater the distance beyond horizon (d_s) and the narrower the beam angles (α) of the antennas. Their estimates assumed scattering parameter was constant with height, scattering was isotropic, and that antennas had identical conical beams. They guessed their estimate might result in too large a value for L_c. Since then, further approximate calculations were made by us assuming a scattering parameter varying with height h according to h^{-2}, with isotropic scatter. The results gave substantially lower values of L_c although the general trends with d_s and α were preserved. In discussions with Staras, he discussed extensions of the original Booker-deBettencourt work to include anisotropic turbulent scattering and scattering parameter varying inversely with square of height. The work was published recently[16]. He pointed out that his method allows estimates of L_c to be made for dissimilar antennas. Staras' basic curves are shown in Figure 9. Separate curves are given for the loss due to vertical beam dimension L_{cV} and horizontal dimension L_{cH}. The curve L_{cV} is a function of the quantity $(\theta_o/\alpha_o)_V$ where

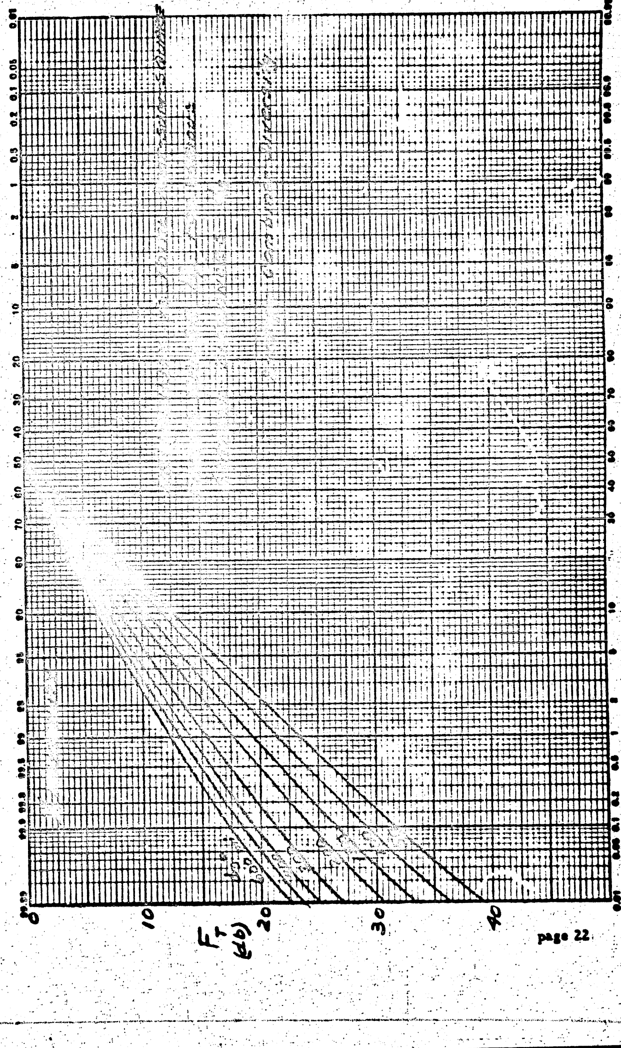


FIGURE 8

$\theta_o = d_s/a_e$ and a_e is taken as the effective earth radius (5280 miles for 4/3 earth). The quantity α_{oV} is related to the vertical beam widths α_{1V} and α_{2V} of antennas 1 and 2 according to

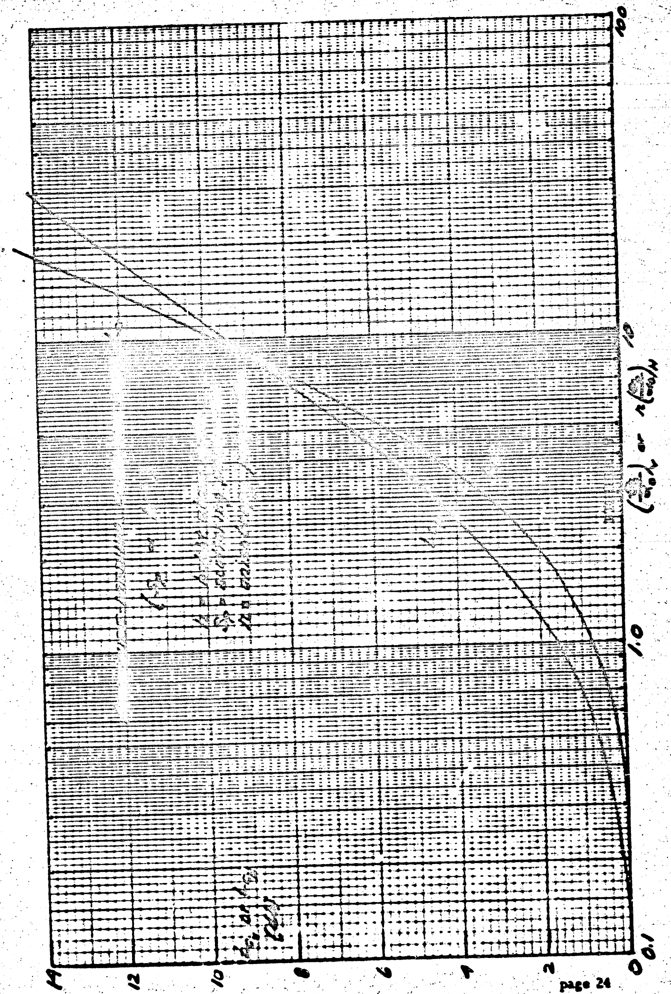
$$(\alpha_{oV})^{-2} = (\alpha_{1V})^{-2} + (\alpha_{2V})^{2}$$

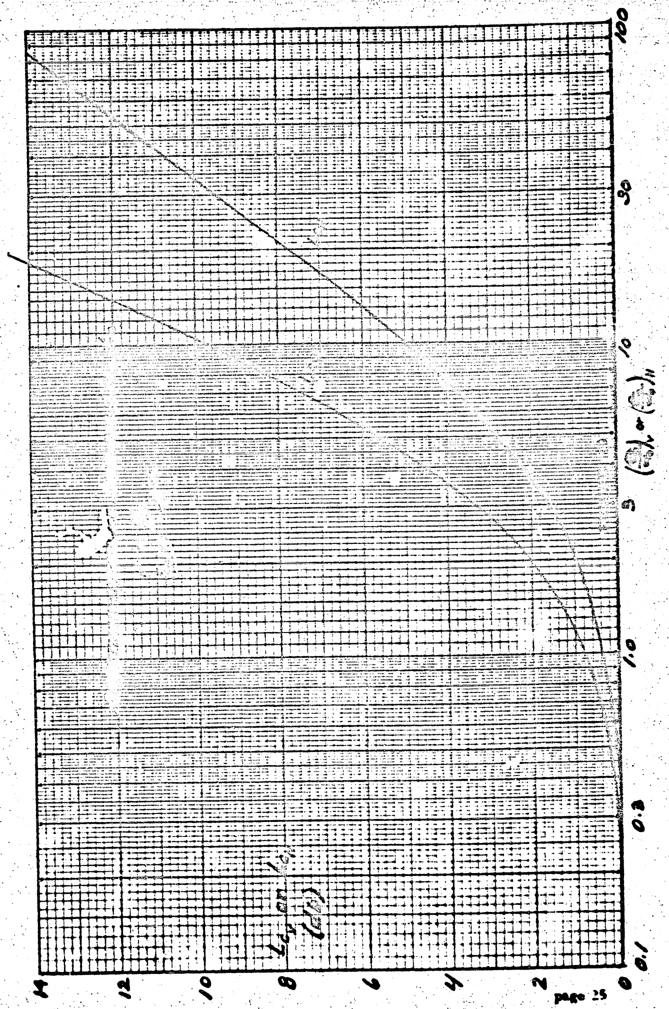
Similarly, the loss associated with the horizontal dimension L_{cH} is a function of the quantity $r(\theta_o/\alpha_o)_H$ where r is the anisotropy parameter. The total loss L_c in db is the sum of L_{cV} and L_{cH} each in db.

For $r = 1/3$, Figure 9 may be redrawn as shown in Figure 10, where L_{cV} is a function of $(\theta_o/\alpha_o)_V$ and L_{cH} is a function of $(\theta_o/\alpha_o)_H$.

Assuming identical paraboloidal beam antennas, curves for various estimates of L_c were drawn and are shown in Figure 11. Note the values of L_c are plotted against θ_o/α not θ_o/α_o, where α is the beam argle (here $\alpha_o = 0.707\alpha$). The dashed curve is the original Booker-deBettencourt estimate for isotropic turbulent scattering and scattering parameter constant with height. The solid curves apply for a scattering parameter varying inversely with height squared. The solid curves show the effect of anistropy. The experimental evidence seems to indicate that $r \leq 1$.

The curves indicate that the assumption of height dependence of scattering parameter strongly influences the resultant value of L_c, as mentioned before. This point was also noted recently by Waterman[17] in discussions.





For our purposes we shall assume now the estimates of L_c given by S_p varying with h^{-2} and $r = 1/3$ For identical beam antennas we would use the solid curve of Figure 11 labeled $r = 1/3$. For dissimilar antennas we use the curves of Figure 10.

8. ANTENNA FEEDER, FILTER AND DIPLEXER LOSSES, L_A

In ground-to-ground tropospheric scatter circuits, experience seems to indicate that ohmic losses in a typical installation (L_A) are given approximately by L_A (db) = 5 log f_{mc} -10 above 100 mc. At 400 mc the ground system is assumed to have a loss L_A = 3 db.

We do not have as much experience with airborne systems. We shall assume L_A here to be 3 db, however, for 400 mc.

9. REQUIRED RECEIVED POWER AND MAXIMUM TOTAL LOSS L

For an SSB system with 4 kc channel bandwidth the received power required may be estimated. An r.m.s. signal-to-r.m.s. noise power ratio of 10 db is assumed, whence noise peaks exceed r.m.s. noise 99.9% of the time. Then received power required is P_R (dbw) = $\overline{NF}$ (db) - 204 + 10 log B + $\frac{S_o}{N_o}$ where $\overline{NF}$ is receiver noise figure in db. For a noise figure of 5 db at 400 mc, the required power is 5 - 204 + 36 + 10 = -153 dbw. The received power will have to equal or exceed this value to obtain the required quality of voice reception.

Again, if a transmitter power P_T of 10 kw (40 dbw) is assumed, the maximum allowable total transmission loss is 40 -(-153) or 193 db. For a

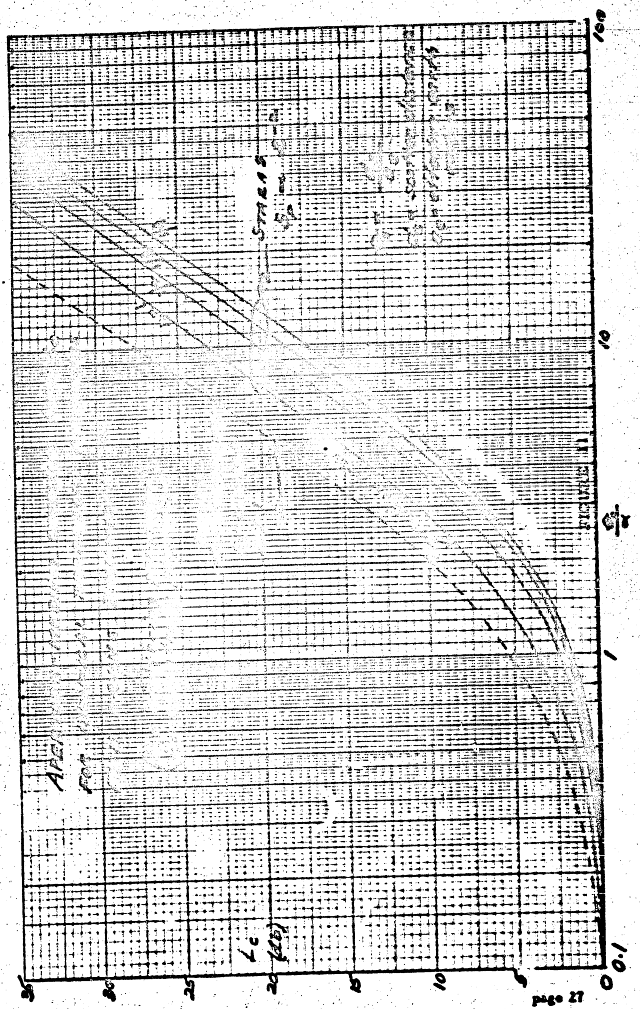


FIGURE 11

1 kw (30 dbw) transmitter power, the maximum allowable loss is 183 db.

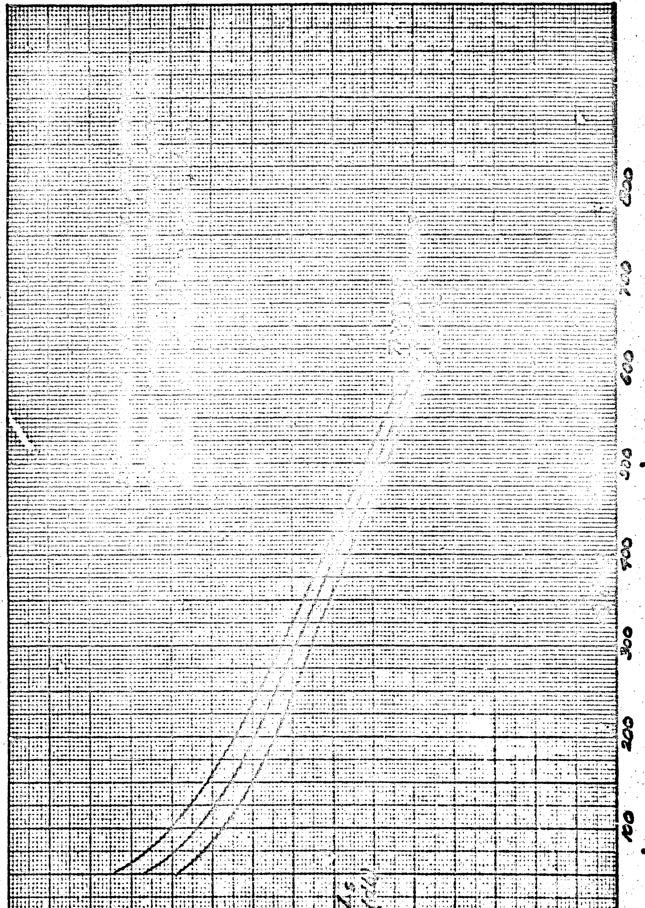
10. PREDICTED LOSSES IN OTHER CLIMATIC REGIONS

It is assumed that the hourly median value of L_s exceeded 50% of the hours in a year (yearly median) for the arctic corresponds to the winter median value of L_s for Midlatitudes. The winter median value for Midlatitudes, based on Lincoln Laboratory data, is a value exceeded approximately 80% of the hours of the year. Hence an estimate of the median value for a year for the Arctic may be obtained by adding to the Midlatitude values of L_s the values of F_s exceeded 80% of the hours.

In a similar fashion, it is assumed that the yearly median value of the hourly median value of L_s for the Tropics is obtained by subtracting F_s exceeded 80% of the hours from the Midlatitude values of L_s.

The results are plotted in the curves of L_s vs d_s in Figure 12, showing the comparison of L_s (hourly median) exceeded 50% of the hours of the year for the Tropics, Midlatitudes and Arctic regions.

We are aware of some of the weaknesses of our assumptions. However, we are emphasizing the regions of high reliability signals rather than levels of importance in estimating interference potential. This means levels exceeded 90, 99 and higher percentages of the time at which it is believed the influences of super refraction, ducting, etc. are less important relatively. A more exacting analysis would be that of correlating refractivity gradients



near the surface (say the gradient for first kilometer of height) with scatter loss following methods of Bean and Meaney[18] There are other correlations being studied the results of which are not yet available. Lacking correlative measurements with signal loss in other climatic areas we shall follow our initial assumptions.

The next important assumptions are those regarding slow fading factor F_s and fast fading factor F_f for the Arctic and Tropics. We assume them to be the same in all three regions. The assumption appears reasonable for the fast fading factor F_f. The validity of the assumption for F_s is open to the same sort of argument given above regarding L_s.

We proceed now to estimate the hourly median value of basic transmission los L_b between isotropic antennas exceeded for various percentages of the hours of a year for the three regions. The results are shown in the curves of Figures 14, 15 and 16 for the Arctic, Midlatitudes, and Tropics respectively. Figure 13 compares 50% reliability curves for these 3 curves.

It may be of interest to compare some very recent data with our predicted curves. On a truly Arctic path, measurements were made during the month of March 1957, with surface temperatures -20° to -30° F. One hundred hours of useful data* were obtained. The circuit constants were

d = 698 mi

d_s = 570 mi

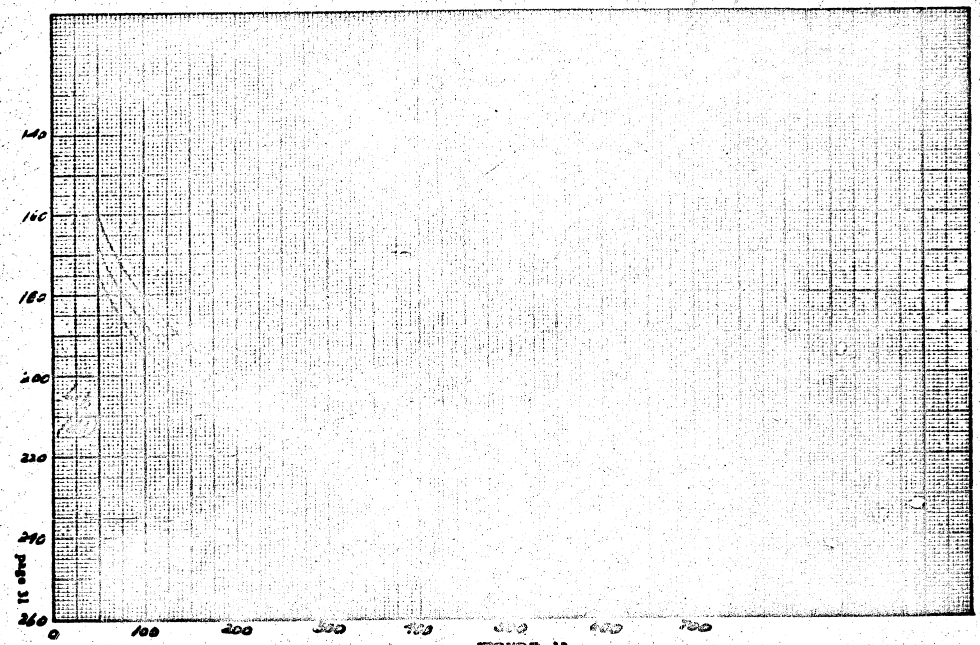
P_T = 10 KW (40 dbw)

f = 420 mc

$G_T = G_R$ = 28 db above isotropic (28 ft paraboloids)

L_A = 2.5 db

* E. J. Martin, AFCRC, private communication. The data were taken under the supervision of Lincoln Laboratory, M.I.T.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160
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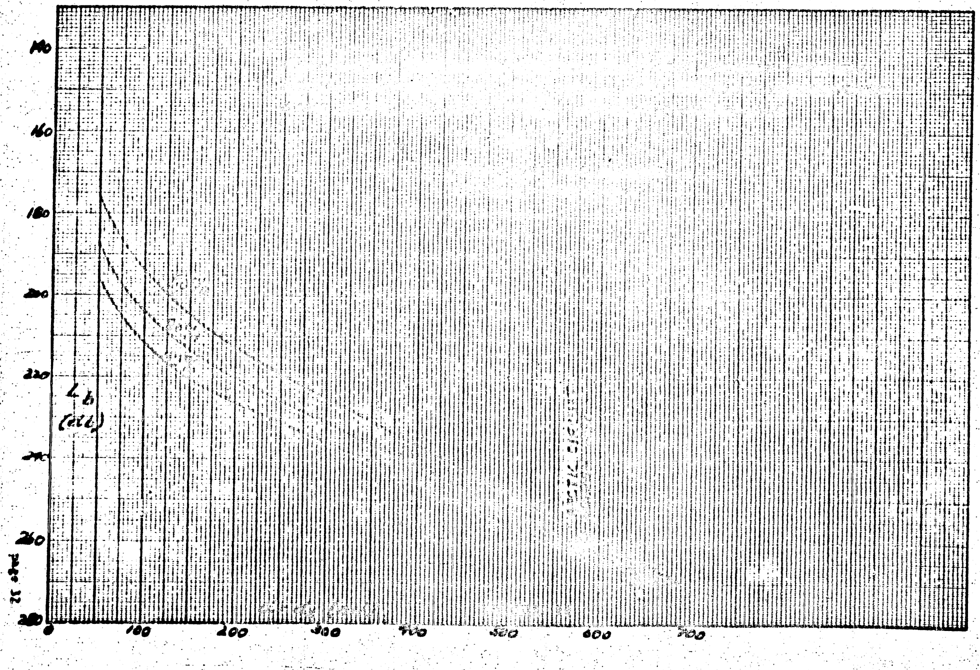
FIGURE 11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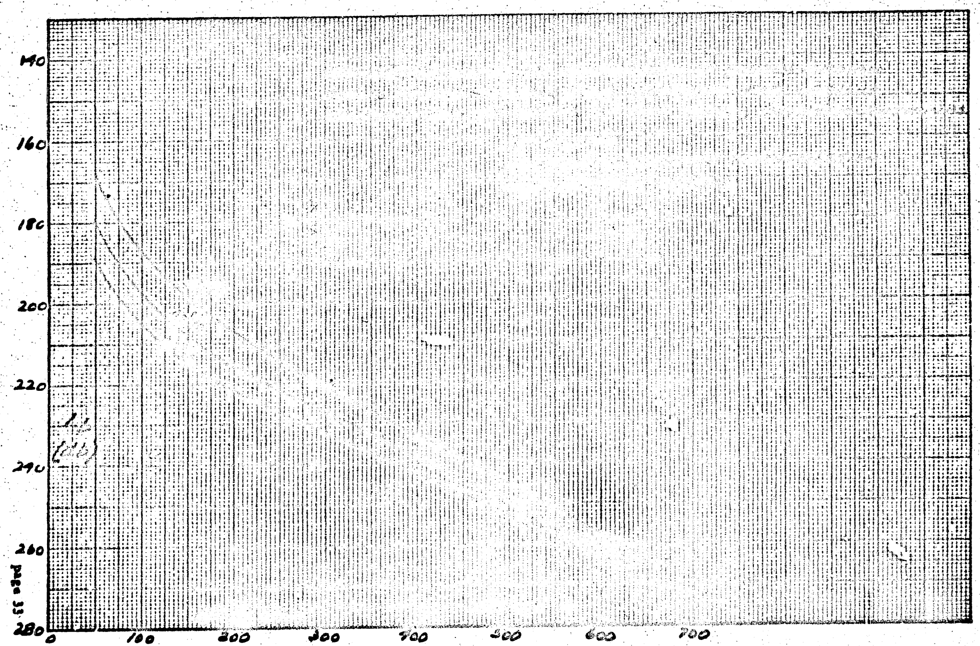
L_b
(db)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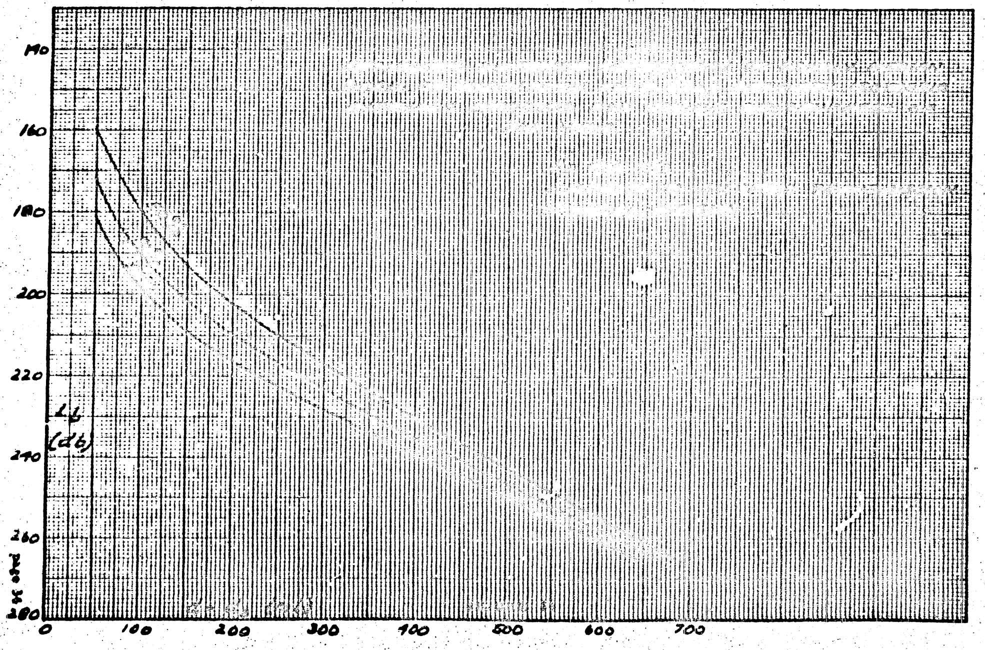
140
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L
(db)
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700

The hourly median received power was -136.5 dbm and -138.9 dbm exceeded 50% and 99% of the hours of the month, respectively. Thus the total transmission losses (L) were 204.0 and 206.4 db respectively. A coupling loss L_c (for $r = 1/3$) is calculated to be 2.0 db ($\alpha = 1.2\lambda/D = 0.1$ rad $= 5.7°$, $\theta_0 = 570/5280 = 0.108$ rad, $\theta\ /\alpha = 1.08$). Free space loss between isotropic antennas is $36.6 + 20 \log 420 + 20 \log 698 = 146.0$ db. Hence $L_s = L - L_{f_0} - L_c - L_A + G_T + G_R =$ 109.5 and 111.9 db respectively. The basic transmission loss L_b between isotropic antennas is then calculated to be 258.0 and 260.4 db exceeded 50% and 99% of the hours respectively. If one plots these values of L_b on the curves of <u>Figure 14</u>, the agreement may be considered good. We should remember that March is a weak month with lower monthly median than for the year and F_s depths are less than for the year (see e.g. Bell Telephone Labs. data for Newfoundland 505 mc circuits, October 1955, Proceedings IRE).

11. HEIGHT GAIN

Predicted values of height gain G_H may be made in that the transmission loss at very high altitude will be less than those on the surface, for a given <u>total</u> transmission distance d. The method would follow that given in <u>Appendix C</u>, and would give hourly median values for various percentages of hours of the year.

However, we prefer to perform such calculations after a discussion and decision on probable system parameters given in the next <u>Section</u> on System Design Studies. The results are given in Section III-B.

12. OBSTRUCTIONS, SITING AND OBSTACLE GAIN

In the choice of ground sites, the problem is made more difficult in some respects than point-to-point scatter communication, particularly in the Arctic because communication is desired with one terminal which is moving rapidly in range and azimuth. In some areas it will be a formidable task indeed to find a site free of foreground obstruction at all azimuths. However, every effort must be made to do so. It is easy to show that if there is an obstruction having an elevation e° above local horizon, then this is equivalent to increasing the scatter distance by 92 e° miles, for small elevation angles. If the obstruction is 1° high this amounts to a 92 mile increase, or approximately an additional loss of 10 db (depending on scatter loss attenuation rate).

In some cases, a true obstacle gain may be encountered unavoidably. This study has not included a detailed inquiry into the subject for direct application but techniques have been outlined for calculated "obstacle gain" [19, 20].

C. EXTERNAL FACTORS AFFECTING PERFORMANCE

1. GENERAL

The minimum detectable signal of any system is determined in part by the ratio of the available signal power to noise power at the input terminals of the receiver. In addition, practical receivers deteriorate the available received S/N due to the contribution of noise power in the receiver input stages. Since it is convenient to refer all rf noise power to a common point, the receiver input, the effect of external noise is combined with the receiver noise figure to provide an "Equivalent noise figure"(F_E), such that the noise power generated by the receiver is F_E KT Δf,

where

T = °Kelvin (288° K = room temperature)

K = Boltzman constant (1.38×10^{-23})

Δf = Bandwidth of receiver (cps)

Noise figures of high quality input circuits have been given as varying between 2 db and 8 db. Figure 17[a] shows what can be achieved using grounded grid planar triodes WE416B below 900 mc and special crystal diode mixers above 1000 mc. Figure 17[b] shows typical receiver noise figure in more detail.

2. EXTERNAL NOISE

The sources of external noise are:

1) man made such as diathermy, ignition, arc discharge, etc.

2) terrestrial

3) extra terrestrial

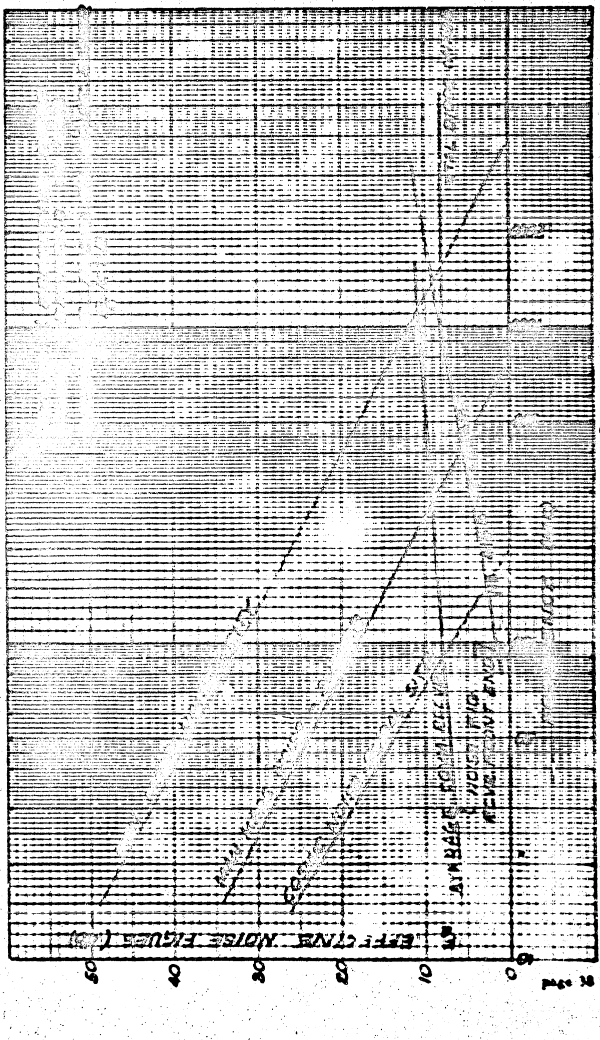

FIGURE 17a

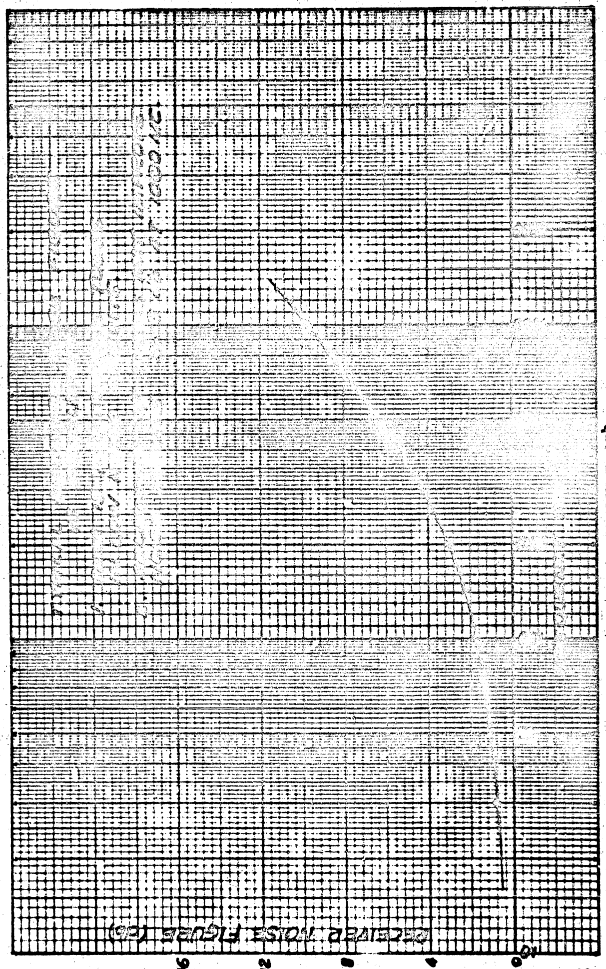

FIGURE 17b

MAN MADE NOISE

The intensity of man made noise is so variable that it is not possible to be specific. The noise from any single source is difficult to predict, but it is usually possible to plan the station so as to avoid the single noise source entirely.

In urban areas there are so many contributing noise sources that the received noise power has characteristics similar to thermal noise. Prediction of man made noise must be based on empirical data rather than theoretical considerations, and admit that such data is effected by antenna directivity, transmission line losses and other factors. The general magnitude and frequency distribution of this type of noise is shown in Figure 17a and is indicative of what may be expected of antennas of medium gain. It is best to attempt to eliminate the effect of this type of noise by locating ground receiving antennas in rural areas and by paying particular attention to other ground siting requirements.

4. TERRESTRIAL NOISE

This noise is usually generated by lightning discharges in thunderstorms, mainly in the tropics and is propagated by regular ionospheric transmission. The noise level is therefore dependent on frequency, diurnal, seasonal, geographic and meteorologic factors.

At frequencies above the MUF (30 - 40 mc) terrestrial radio noise disappears and radio noise emanating from extra terrestrial sources - so

called cosmic noise - becomes a factor in VHF communications up to 200 mc.

By convention the term cosmic noise includes both solar and galactic noise.

5. GALACTIC NOISE

Incomplete surveys of the continuous background radiation from interstellar space have been made and have established that the noise is not uniform over the celestial sphere.

It is weakest at the galactic poles, "brightest" at the center (Sagittarius) and 'bright' along the galactic equator (Milky way). Since the earth's rotation causes the antenna beam to sweep across the celestial sphere, the received cosmic noise power depends upon time, antenna azimuth and directivity. This cosmic noise has characteristics similar to thermal noise and is the predominant source of noise at isolated (rural) stations for VHF. The contribution to the effective noise figure is small, except for antenna high gain oriented to intercept bright portions of the galactic center.

6. RADIO STARS

The noise emanating from radio stars is not significant for frequencies above 300 mc even for the most intense source, Cassiopeia.

7. SOLAR NOISE

The sun is a serious source of radio noise. In northern latitudes, the sun is above the horizon and at low elevations for a considerable portion of the summer day. It is therefore important to estimate how serious this interference can be. Solar noise can be divided into steady components which are present all the time and erratic components of high intensity which occur during disturbed conditions.

The steady component (quiet sun) consists of a constant basic component and a slowly varying component. Noise from a disturbed sun is comprised of noise storms, outbursts and isolated bursts.

When the antenna beam is pointed directly at the sun the effective noise figure will be increased by a factor of 9-25 (6-14 db), even under quiet conditions. When the sun is disturbed its contribution to effective receiver noise figure increases and on occasion may be as much as 30 db. Even when the antenna beam is not pointed directly at the sun, noise entering the side lobes may be sufficient to cause trouble.

8. AURORA

There is a possibility that the aurora may emit noise in the UHF band. There has been little published material on the subject and until detailed evidence is available, evaluation is not possible. It is evident from work reported at lower frequencies that noise power is not significant and becomes less important as the frequency is raised.

D. LOW AND VERY-LOW FREQUENCY TRANSMISSION

Quantitative experimental information about air-to-ground and ground-to-air radio propagation in the low frequencies and very-low frequencies is rather limited. Unless the data are collected with well calibrated and standardized instruments, they are difficult to correlate and analyze, and controlled experiments, on a scale large enough to be of interest, are expensive. However, experience gained since World War II in the development of various navigation systems and in the siting and testing of high power transmitting stations has served to check the earlier empirical and theoretical relationships for ground-wave and sky-wave propagation.

The Beetle low frequency loran program[21] operating at 180 kc (and an input power of 100 kw) along the Alaskan and Canadian coast of the Arctic Ocean demonstrated the effects of geomagnetic and ionospheric activity and of the low conductivity of the frozen muskeg on sky-wave and ground-wave propagation. Field intensity and propagation time difference were measured on many flight tests, some to ranges of 1400 nautical miles, and at seven monitor stations at various distances out to nearly 1400 nautical miles from the transmitting station. Comparison between the observed variation of ground-wave field intensity with distance and the theoretical variation yields a value of 2 to 4 x 10^{-14} electro-magnetic units for the ground conductivity of the muskeg. This accounts for the unusually rapid attenuation of the ground-wave signal in the Arctic region. A summary of the ground-wave and sky-wave field intensities is given in Figure 18, in which the first-hop and second-hop sky-wave field intensities for cosine of the sun's zenith distance equal to

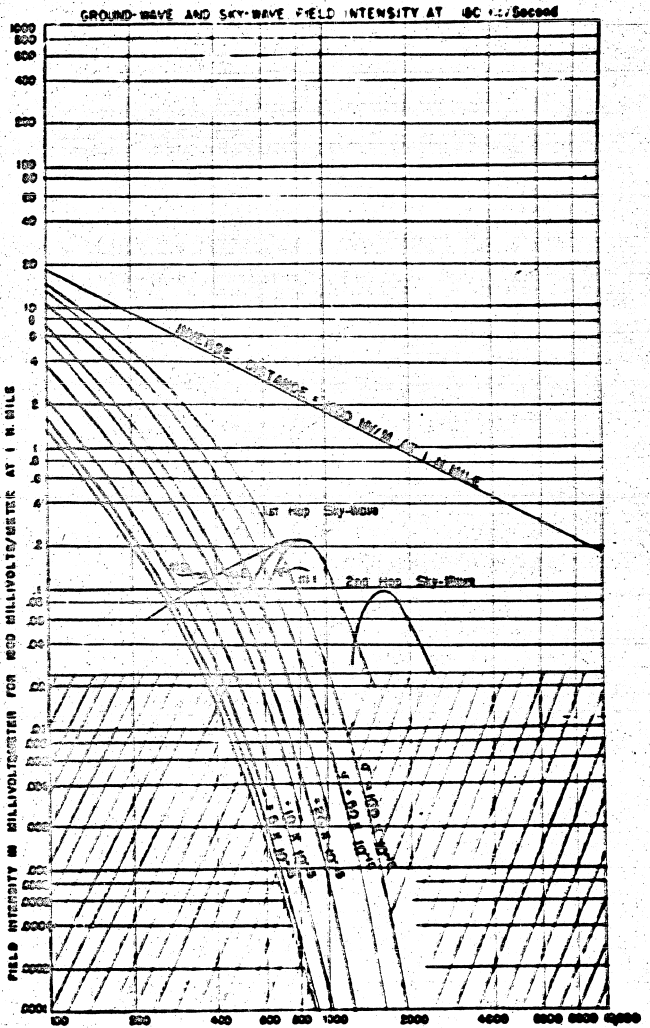

FIGURE 18

-0.5 were estimated for field intensities and time differences measured at the ground monitor stations and in flight. At ranges of predominant sky-wave propagation (more than 200 or 300 miles) the field intensity and sky-wave propagation time were found to correlate well with the cosine of the sun's zenith distance. Attempts to correlate auroral activity and geomagnetic disturbances with the observed field intensities indicate that they have little effect, and if anything, they tend to enhance sky-wave propagation at this frequency.

In 1950 the Air Force[22] set up a transmitting station at Allaire, New Jersey, operating at 100 kc and radiating 800 watts of unmodulated carrier power. Over a period of nearly a year, signal and noise field intensities were measured at six ground monitor stations with a maximum range of 2200 nautical miles and on 13 flight tests. From these measurements the reflection coefficient was determined as a function of the sun's zenith distance for various angles of incidence upon the ionosphere. For incident angles of more than 75° the average day and night reflection coefficients were found to be 0.15 and 0.6, respectively. From these values of reflection coefficient and an assumed ground conductivity the ground-wave and sky-wave field intensities of Figure 19 for daytime and Figure 20 for nighttime were derived. Attempts to correlate the observations with geomagnetic K character figures and with the I character figures of ionospheric storminess indicated that at this low frequency the absorption is not significantly changed by ionospheric or geomagnetic activity. There is an indication, however, that sudden ionospheric disturbances do appreciably increase the ionospheric reflection coefficient.

Some information on radio propagation at very low frequencies was obtained on a Navy-sponsored survey[23] of noise and field intensities of signals radiated

Figure 18. Variation of Winter Field Intensity with Cosine of Sun's Zenith Distance at Hastings College (1107 N. Miles)

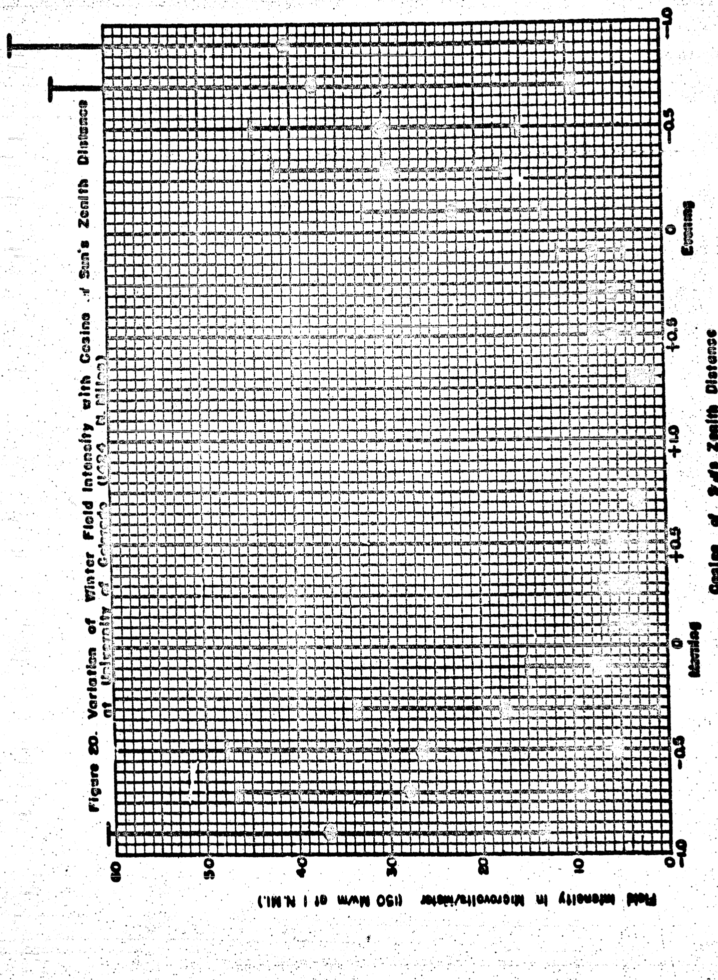

Figure 20. Variation of Winter Field Intensity with Cosine of Sun's Zenith Distance at University of Colorado (1494 N. Miles)

from the British station GBR at Rugby (16 kc), from the Annapolis, Maryland station NSS (19 kc) and from the Marion, Massachusetts station AFA2 (25.82 kc). Field intensities were measured in Newfoundland, Iceland, Ireland, Germany and French Morocco, and on a Naval aircraft. The results are in good agreement with a recent empirical study of low frequency propagation[24]. They indicate, however, that the ground conductivity at points of reflection from the earth strongly effect the received field intensity. Land masses of low conductivity such as Labrador and the ice caps of Greenland cause a loss of 6 or 8 decibels per reflection. The results of the flight measurements showed that the field intensity observed at any one time and place is largely dependent upon the vector addition of the ground-wave and sky-wave components.

Similar airborne measurements of the field intensities of signals from Jim Creek, Washington, at 19 kc and from Haiku in Hawaii at 40 kc have recently been reported[25]. These results were obtained mostly from daytime flights over the Pacific.

While the available information about low and very-low frequency propagation is limited to ground-to-ground and ground-to-air observations, air-to-ground communication may be feasible for some applications. Some of the advantages of this frequency range are that the antennas are essentially omnidirectional, ground-wave and sky-wave propagation characteristics are excellent, especially over water, and propagation is not adversely effected by auroral or geomagnetic activity. The disadvantages are that rather high power is required to overcome atmospheric noise, the transmitting antenna is awkward and the bandwidth is limited to a kilocycle or so.

E. HIGH FREQUENCIES

1. REGULAR LAYER HF PROPAGATION

Transmission loss computations for HF regular layer propagation can be made using standard prediction techniques outlined in NBS circular 462, Signal Corps Radio Propagation Unit Reports, and similar sources.

However, the low reliability of such propagation in the Arctic auroral zone is well known[26] and makes one conclude such propagation is insufficiently reliable for USAF air-ground use. It is useful some of the time and may be the only means available for long range communication, albeit unreliable in the Arctic.

2. OTHER POSSIBLE HF MECHANISMS - HF BACKSCATTER

If the communication requirement can accept a single 3 kc wide channel with operators at both terminals to check errors, there is a possibility of using high frequency circuits. Such a system will require at least one terminal to use more complex equipment than normal, if reliable communications are to be established at any range and time. However, many techniques have been established which have neither been widely adopted nor simultaneously applied. Two such HF techniques are reviewed here.

To reliably establish a circuit to a remote mobile point by high frequency techniques, it is a requirement that the correct operating frequency and best propagation mode be known to a much better degree than is usually true. Two factors influence any communications from the United States to more northern points: the erratic nature of sporadic E-layers, and the high absorption due to auroral zone effects.

Sporadic E-layers will either prevent successful transmission when F-layer modes are attempted, or greatly enhance transmission if E-layer reflections are intentionally utilized. The best way to insure that the correct propagation mode is used is to make continuous oblique incidence ionosphere soundings in the direction of interest and with sounding system parameters consistent with the communication system parameters. By using the COZI techniques, the best mode of propagation and the best of the available assigned frequencies can be continually known for any path not influenced by auroral zone effects.

Transmission to points in or, by means of multi-hop transmissions, beyond the auroral zone, can be accomplished if the angle of the down coming wave is such that it does not pass through the E and D layers in the disturbed auroral area. This places a severe limitation on the percentage of time that any fixed station with only a few assigned frequencies can communicate to given points in the Arctic by means of great circle routes. However, by taking advantage of ground reflected back or side scatter transmissions, non-great circle communication paths can be accomplished*. With the normal HF forward ground reflected wave, appreciable energy is also scattered at other than forward directions. The backscattered components are

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* Much of this data is due to a paper, "Some Ionosphere Scatter Techniques," by D.A. Hedlund, L.C. Edwards, and W.A. Whitcraft of Raytheon Mfg. Company, which was read at an IRE, PGAP Symposium held in Washington, D.C., 15 November 1955.

those normally used in oblique incidence sounding techniques. The loss over the one hop F-layer forward wave and the one hop returned scatter is not prohibitively high. A peak power of 400 watts at a one percent duty cycle, 20 kc bandwidth and 10 to 15 db antenna gains will give usable signal levels for sounding techniques. Experiments have demonstrated that a 400 watt, voice modulated signal is sufficient between two points along the Atlantic Coast by using two hop propagation paths via a surface reflection point well east by the two terminals. These experiments used a surface point which fell along the perpendicular bisector of the baseline between the two stations, and frequencies high enough that oblique incidence soundings disproved chances of normal propagation modes. Antenna rotation experiments also confirmed the mode of propagation.

The back or sidescatter mode would allow much greater freedom in communicating through the auroral regions. A sufficiently low angle of arrival can be chosen by utilizing frequencies near the MUF, and long two hop paths. It may well be possible to use more than one non-great circle scatter hop. If this be the case, any point in the arctic area could be reached. COZI soundings can be used to indicate the high absorption areas of the auroral zone for the first section of any particular transmission, because little or no backscatter is received from these azimuths.

The equipment requirements of the backscatter mode transmissions are not more complex than those of most other communication systems

proposed for reliable long range circuits. High power transmitters and high gain ground station antennas are required. Both are possible and at least as economical as tropospheric scatter equipment. Airborne equipment would be no less feasible than in other systems. It would be desirable that an omnidirection aircraft antenna be used, but some gain could be obtained by limiting the vertical plane radiation to low angles above the horizontal. A means of constantly monitoring several assigned frequencies would be required at the non-control terminal. The medium bandpass is sufficient for voice, but, because of the nature of the backscatter signal, care must be used in high speed automatic code transmissions. This latter requirement is due to the possibility of equal multipath signals or abrupt changes in propagation time at slowly repeated intervals.

The scatter mode of propagation can be used to advantage in any geographical area and provides a method of communication to a mobile station from zero range out to several thousand miles. It suffers the following disadvantages: it is not suitable for completely automatic coded transmissions unless much redundancy is used, it requires highly skilled control station operators, it utilizes frequencies which have both high noise levels and high interference from other signals, the reception at the control station could be readily jammed from far distant points, the reception at a mobile terminal could be readily jammed if its position were known, practical considerations create limited bandwidth and limited frequency assignments, sporadic E at ranges beyond detection by oblique incidence soundings can interfere with many of the intended paths, other ionosphere disturbances can prevent reliable transmissions, the required control station high gain steerable antenna array requires careful siting and large land areas.

F. VHF IONOSPHERIC AND TROPOSPHERIC SCATTER LOSSES

1. ESTIMATED LOSSES

a. Tropospheric Loss

The 300 mc data of Bullington[5] was adjusted to 50 mc by reducing the loss in ratio of 50/300 = approximately 8 db. This tropospheric scatter loss is compared with ionospheric scatter loss at the same frequency, 50 mc, in Figure 21.

b. Ionospheric Loss

Measured circuit losses were taken from Bailey, Bateman and Kirby[27]. (Geographic and antenna data was taken from Table I, page 1185 of this reference.) No attempt was made to adjust the maximum antenna gain values shown for gain at elevation angles other than the maximum. Although a check of actual elevation angles for an 85 kilometer layer height results in somewhat lower angles than the stated antenna lobes, this is a function of the actual layer height. The difference, in any case, will not amount to more than a few decibels and, in view of the large path loss variations, seems unimportant. The Cedar Rapids - Sterling, Anchorage - Barrow and Fargo - Churchill links were used as typical circuits. Received power values for each path were estimated from the seasonal data. However, comparison shows that the average seasonal data do not differ greatly from the average daily values excepting the hours between 1000 and 1600.

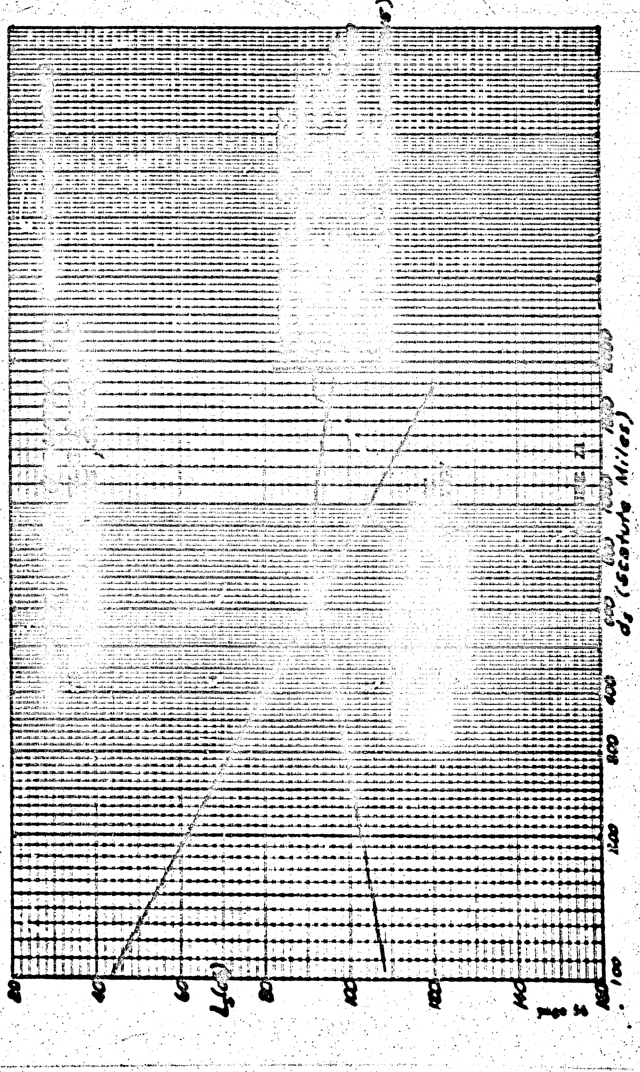
The three circuit scatter losses were found to be:

Sterling	97 db
Fargo	88 db
Anchorage	93 db

The curve was fitted to the 93 db level below free space.

The curve of Figure 21 for ionospheric scatter has a loss versus range dependency as determined by the Booker-Gordon theory. If Eckersley theory were used, and again fitted to 700 to 800 mile measured values, the range at which the tropospheric scatter curve crosses the ionospheric curve would be about fifty miles greater. This difference will be less than 100 miles for a wide variation in assumed scatter losses for the 700 to 800 mile circuits.

The daily variations in the ionospheric circuit will cause the range at which the mean signal equals the tropospheric signal to vary from perhaps 150 miles less than indicated (from 635 to 485 miles) at noon, to perhaps 50 miles greater. Seasonal changes will cause an additional ± 50 mile change. Thus the total variation of the equal mean signal point will vary between a range of 435 miles to 735 miles and the ionospheric scatter loss will vary between 82 db and 96 db for this condition.



G. VHF METEORIC SCATTER

1. INTRODUCTION

Besides LF, HF and weak background VHF ionospheric scatter modes there is another mode competing with UHF tropospheric scatter for air-ground communications which requires evaluation. This mode is that of VHF meteoric scatter.

There is considerable research being done in the field currently and some of the theoretical and experimental aspects are reviewed in what follows. The field is one in which the state of the art is being continually advanced, and results require further study before useful system designs could be advanced. Before this step is taken toward system design, the relative merits of this propagation mode vs UHF tropospheric scatter needs to be evaluated.

2. REVIEW OF THEORETICAL AND EXPERIMENTAL ASPECTS

Before evaluating the meteor burst transmission mode, let us outline the pertinent theoretical and experimental aspects of the field.

a. Nature of Meteor Burst Ionization

When signals are transmitted over a path approximately 600 to 1100 miles long at frequencies exceeding the MUF, the received signals are weak but omnipresent and possess characteristics of practical importance to reliable communication. The signals consist of a background component plus enhancements of several varieties. The transmission takes place via the E-region of the ionosphere, the weak background signal being explained by turbulence effects on the ionization present, whatever the cause, and the enhancements by this and other means. There are enhancements, where the total level is raised for long periods of time, attributed to reflection from sporadic E-clouds. Then there are the spiky, sharp rising enhancements due to specular reflection from ionized, rectilinear columns or trails caused by meteor impact. It is with these spiky enhancements, or meteoric scattering, that we shall be concerned for communication over an oblique-incidence ionospheric path.

The meteoric enhancements are principally of two types: (a) those which rise sharply and decay exponentially into background noise after a short time, in the order of couple of seconds or less, and (b) those which rise sharply and after a few seconds fade irregularly with amplitudes decreasing into the background noise. Most of the theory and experimentation has been concentrated on echoes of type (a). The theory has developed from that applicable to radar backscattering from the ionized trails to that for forward off-beam scattering from such trails. The major difference in these two cases of scattering is the echo duration, in which back-scattered echoes of type (a) last a few tenths of a second and are termed short-duration echoes; back-scattered echoes of type (b) are called long-duration echoes, the breaking up of the echo occurring after a few tenths of a second.

Generally, forward scattered echoes last much longer and are more intense than backscattered echoes.

b. The Radar (Backscattering) Case

The scattering of energy from a meteoric ionized trail at ionospheric heights has been shown to be due to electrons rather than heavier ions in the trail[1]. If there is a line density of q electrons/meter, a trail is said to be underdense when $q < 10^{14}$ and otherwise, overdense. Echoes of type (a) are explained by the theories of Lovell and Clegg[28, 29] and Herlofson[30]. For echoes of type (b), the theory of Kaiser and Closs[31] applies for the initial interval for over-dense trails. An explanation for the fading effects thereafter is not completely agreed upon, but a very recent theory of Booker and Cohen[32] with experimental data uses turbulence for the explanation as opposed to an earlier specular reflection theory by Eshleman and Manning[33].

The radar equation for received power P_R due to backscattering from an underdense meteoric ionized trail, having a density of q electrons/m, is

$$P_R = \frac{P_T G^2 \lambda^3}{32\pi^2 R^3} \; q^2 \; (r_e)^2 \text{ watts} \qquad (q < 10^{14}) \qquad (4)$$

where P_T = transmitter power, watts

G = identical transmitter and receiver antenna gains over isotropic

λ = wavelength, meters

R = slant range, meters

r_e = classical electron radius, meters

$= (\mu_o e^2/4\pi m) = 2.818 \times 10^{-15}$ meters

(with e = electronic charge = 1.602×10^{-19} coulombs, m = electronic mass $= 9.106 \times 10^{-31}$ kilograms, and μ_o = permeability of free space = $4\pi \times 10^{-7}$ amperes/meter).

Comparing (4) with the usual radar equation for a radar echoing area σ_T for the trail, one obtains

$$\sigma_T = 2\pi r_e^2 q^2 R\lambda, \text{ sq. m.} \qquad (5)$$

$$= \sigma_e q^2 R\lambda/2, \text{ sq. m.}$$

where σ_e is the radar echoing area of an electron $(4\pi r_e^2 = 10^{-28}$ sq. m.)

For a small meteor of 1 milligram mass the trail is barely visible (5th magnitude) and the line density is 10^{14} electrons/m. Thus, the radar cross section, σ_T, for an ionized trail due to such a meteor is about 10^6 sq. meters at 250-km slant range on a radar frequency of 40 mc. Such a large radar area for such a small meteor explains the ease with which meteor trails may be detected.

The above treatment applies to the case of underdense trails, which Kaiser and Closs[31] have shown to apply for trails where $q < 10^{14}$ electrons per meter. For $q > 10^{14}$ electrons/m., they and Greenhow[34] have shown that (4) becomes[35]

$$P_R = \frac{P_T G^2 \lambda^3}{54\pi^3 \; R^3} \; (q \, r_e)^{1/2} \qquad (q > 10^{14}) \qquad (6)$$

Equation (4) gives the peak received power because the ionized trail will immediately diffuse and the value of P_R will decrease with time exponentially as ϵ^{-t/T_B} where T_B is the decay time constant for backscattering.

For underdense trails T_B is given by

$$T_B = \lambda^2/32\pi^2 D \qquad (q < 10^{14}) \tag{7}$$

in which D is the diffusion coefficient (approx. 3 m²/sec.). Thus T_B varies as λ^2 for underdense trails, but does not depend on q.

For the high density trails $(q > 10^{14})$ the duration is[32]

$$T_B = \text{const.}\ q\lambda^2/D \qquad (q > 10^{14}) \tag{8}$$

and thus the duration depends on q as well as λ^2.

c. The Communication Case (Forward, off-beam Scattering)

Of interest in communications is the case of a separated transmitter and receiver where propagation takes place via an oblique path from transmitter, T, to meteor trail at M to receiver, R. The foregoing equations for backscatter must be modified by several obliquity factors, mostly geometric, to be applicable to the forward or off-beam meteoric scattering.

We may relate the power received, P_R, via oblique meteoric scattering to that which would be received in free space over the same distance P_F by the relation

$$P_R = P_F L\, \epsilon^{-t/T_F} \qquad (q < 10^{14}) \tag{9}$$

in which

L = loss factor of P_R with respect to free space

T_F = echo duration for forward scattering.

The free space received power, P_F, is

$$P_F = \frac{P_T\, G^2 \lambda^2}{(4\pi)^2 (R_1 + R_2)^2} \tag{10}$$

where R_1 is the distance from transmitter, T, to meteor trail, M, and R_2 is the distance from the trail at M to the receiver, R.

An expression for L may be derived from equations in the literature[35], for low density trails. The loss factor, L, is a complicated function of wavelength, path length, trail density and orientation. We may express L as the product of several factors

$$L = \lambda F_R F_q F_P F_G \tag{11}$$

where

λ = wavelength

F_R = range factor = $\dfrac{R_1 + R_2}{R_1 R_2}$

F_q = trail ionization factor = $(r_e q)^2$

F_P = polarization factor = $\sin^2 \alpha$

F_G = geometrical factor = $\dfrac{1}{1 - \cos^2 \beta \sin^2 \phi}$

in which α = polarization angle, i.e. the electric angle between incident vector and the line R_2 from trail to receiver

β = angle between axis of the trail and the plane containing T, R and the center of principal Fresnel Zone on the trail

2ϕ = angle between the lines R_1 and R_2, i.e. the forward scatter angle (this is 180° - 2θ where θ is Booker's off-beam angle of scattering)

In what follows we shall assume horizontal polarization such that the polarization factor, F_P, is unity. A typical example will suffice. Consider a meteor trail at the midpoint of a 1000 km path ($R_1 = R_2 = 500$ km) such that $\beta = 0$ (trail axis is in the plane containing T and R). Further assume a line density of $q = 10^{14}$ electrons/m. Then, $F_R = 2/R_1 = 4 \times 10^{-6} m^{-1}$; $F_q \cong 8 \times 10^{-2}$; $F_G = \sec^2\phi$ and

$$L = 3.2 \times 10^{-8} \lambda \sec^2\phi \qquad (12)$$

At 50 mc, and assuming plane earth geometry for a trail 100 km above the earth such that sec ϕ is about 5, then L becomes about 5×10^{-6} or about -53 db loss with respect to free space. (This is about 50 to 60 db stronger than a typical weak background ionospheric scatter signal[37]).

For a 1 KW, 50 mc transmitter with antenna gains of 10, then, the free space received power, P_F, is about 2.3×10^{-6} watts. Hence, for our example, P_R from the trail is about 1.1×10^{-13} w.

That such a signal is readily detectable comes from receiver input noise considerations. At VHF with a good receiver, the noise limitation is more probably that of cosmic origin rather than set noise. When pointed at the Milky Way the antenna captures cosmic noise power which may be 6 db or so above set noise at 50 mc[37]. The available noise power, P_N, would be in this case

$$P_N = (N - 4) kTB \qquad (13)$$

where N = receiver noise figure, $kT = 4 \times 10^{-21}$ joules, and B = receiver bandwidth (cps). For a 1 kc bandwidth and a noise figure of 3, then, P_N becomes 2.8×10^{-18} watts. The meteoric trail received power in our example is then about 46 db above background noise.

The system performance will depend upon the ultimate signal to noise ratio. In the VHF band, for broad beam antennas, Cottony and Johler[38] showed that the cosmic noise power varies as $\lambda^{2.3}$. If this power exceeds set noise, then we may express the available total noise power, P_N, in terms of an equivalent noise figure, N_e,

where

$$N_e = 0.25\lambda^{2.3} \qquad (\lambda \text{ in meters})$$

$$\therefore P_N = 0.25\lambda^{2.3} \; kTB \qquad (14)$$

Thus, with all other quantities constant over the VHF band (e.g. for constant gain antennas) the ultimate cosmic-noise-limited signal-to-noise ratio varies as $\lambda^{0.7}$, so that lower frequencies in the band are slightly favored, for individual echoes.

Comparing the strength of the forward scattered echo with that for the backscattered echo, we see the received power is $\sec^2\phi$ times as strong for the oblique path ($\beta = 0$). This feature is outstanding in the potential of meteor trails for use in communication.

When considering the durations, the duration T_F is $\sec^2\phi$ longer than T_B for the back-scatter echo, where

$$T_F = \lambda^2 \sec^2\phi \, / \, 32\pi^2 D \qquad (q < 10^{14}) \tag{15}$$

which is $\sec^2\phi$ times equation (4). Thus, for the 50 mc example, T_B works out to be .04 sec. and T_F about 1 second. (The values of T apply where P_R falls to 1/e times the values at t = 0, or maximum P_R. On a voltage basis multiply T by 2).

Other analyses may be performed on the above case for forward-scattered, underdense trails. Experiment agrees very well with theory. For forward-scattered long-duration echoes due presumably to overdense trails, there is not too satisfactory a theory. A back-scattered long duration echo theory has recently been put forth by Booker[32] with some experimental confirmation ($T_B > 0.4$ sec). These cases have importance for, while less frequent, they do give rise to a substantial fraction of the ultimate meteoric duty cycle.

2. PRACTICAL TOPICS ON METEORIC SCATTER COMMUNICATIONS

Let us now consider a few topics in the practical application to communication. One topic concerns antenna design. Another concerns the frequency power spectrum. Finally, the integrated communication capacity for many meteors as opposed to the single meteor theory above is presented.

a. Antenna Beam Orientation

Because of the geometry of meteor trail forward scattering the Stanford group early suggested that more power would be received over an oblique path if the antenna beams were swung mutually to one side of the great circle plane than if the antennas were beamed along that plane. The earlier work assumed that meteor radiants were uniformly distributed over the celestial hemisphere. Results of their more recent measurements and analyses together with those of other agencies such as CRPL (NBS) and RPL (Ottawa) have been described in several recent symposia[39, 40, 41, 42] on scatter propagation including meteoric scatter in particular at which representatives of the company participated. Considering communication path orientation in the Northern Hemisphere (N. of 23°) and the diurnal variation of received power the prediction was that

1) on an E-W path, received signals would be stronger at 0600 with beams mutually oriented North of the path and at 1800 with beams intersecting South of the path;

2) on a N-S path, similarly, at 1200 antenna beams should be directed to the East and at 0000 they should be oriented to the West.

These points were in part confirmed by the Cedar Rapids-Sterling CRPL measurements and those of Lincoln Laboratory[37]. More complete tests just reported[42] by CRPL group indeed do strongly indicate a beam swinging antenna would be very useful, resulting in about 6 db increases in median power on the Cedar Rapids-Sterling link with alternate sites at Shepherdstown and Careysbrook either side of Sterling. The antennas would be beamed at noon along the path, north in the morning and southerly in the evening. Similar experimental results by the Canadians were given for Ottawa-Greenwood (E-W) and for Suffield-Churchill (NS) and Suffield-Winnipeg (EW) circuits. Theoretical support was given by Hines et al[46]. The Stanford experiments were based on TV station signals received at Stanford from Spokane and from Phoenix, the results confirmed roughly the simple model prediction. Recent work of the DSIR (Slough, England) on the circuit to Gibraltar was interpreted by Eshleman[41] as confirmation of the beam shift desired (5° to the East in daytime, 5° to the West at night).

b. Frequency Power Spectrum

The Canadian RPL group made measurements of the frequency power spectrum of meteor trail oblique transmissions. Vogan reports[47] results of measurements of the signal received at Ottawa off beam from the Cedar Rapids 50 mc signals beamed toward Sterling with particular reference to long-duration as well as the short burst signals. The long duration echoes (4 to about 19 seconds) had a spectrum extending beyond 60 cps above the carrier, being about 12 db down at 50 cps. The short duration echoes (0.2 to 0.4 seconds duration) had a much narrower spectrum. Vogan attributed the significant broadening of the long duration echoes to the probable effects of diffusion and turbulence on the trails after their formation. (This feature was elaborated upon for back-scattered radar echoes in the theory of Booker and Cohen[32]. The theory should be extended to oblique propagation for communications and elaborated upon.)

c. Duty Cycle

Of importance in communications is the percentage of time the meteoric circuit will be "open", i.e. the duty cycle. If it is assumed that the times of occurrence of the bursts are random, the duty cycle d may be defined as

$$d = 1 - \epsilon^{-T} \tag{16}$$

where T is the integrated time that meteors exist above a reference level in a unit time interval. The integrated time T is found by summing the durations of each echo above this level. From (16), when $T = 0.693$, the reference level is the median (50% exceeded) level. If $T \geq 3$, the signal is said to be continuous (95% of time). If $T \leq 0.3$, then $T \cong d$, i.e. for small T, the value of T is approximately the duty cycle, d.

Eshleman[43] gives

$$T = 5.4 \times 10^{-17} \left[\frac{P_T \lambda^7}{P_r R^3} \right]^{1/2} \int_{A_h} (P' \sec^2 \phi)(G_R G_T)^{1/2} \, dA_h \qquad (17)$$

where most of the symbols have been given previously. Others are

P_r = reference received power level (watts) above which the signal exists for the time T, i.e. $(1 - \epsilon^T)$ fraction of the time.

2R = great circle path distance (meters)

A_h = area in the h plane (plane containing meteor trails at height h above the earth) common to T and R antenna beams.

P' = probability of detecting randomly oriented meteor trails occurring within a given h-plane area.

By making some idealized assumptions, the above equation may be simplified in order to assess effects of various system parameters. Thus for idealized pencil beam antennas at transmitter and receiver, $G = 16/\delta^2$ where δ is the beam width in radians. Then $A_h = \left[\pi R^2 \delta^2 \sec \phi\right] / 4$ and

$$T = 3.3 \times 10^{-3} \left[\frac{P_T}{BK} \right]^{1/2} \lambda^{2.35} (P' \sec^2 \phi) \sec^{3/2} \phi \qquad (18)$$

where B is receiver bandwidth and K is the ratio of reference level P_r to cosmic noise level (assumed to vary as $\lambda^{2.3}$), with cosmic noise assumed to be limiting background level. Because of some of the idealized assumptions, there is probably about a factor of 10 db between practical circuits and the theory expected of equation (18).

The equation (18) indicates that the duty cycle for burst communication, i.e. the fractional time the signal power exists above a reference level P_r which is a fixed ratio K above the cosmic noise background varies

1) with the 2.4 power of λ
2) with the 7/2 power of sec φ
3) with the square root of P_T/B
4) inversely with the square of the ratio K

In addition T is proportional to the number of trails suitably oriented to produce reflections in the common illuminated h plane area.

As an example, at λ = 10 m with idealized narrow antenna beams on a 1000 km path (φ = 77°), the computed value of required transmitter power P_T is 13 watts for a duty cycle of 10%. This assumes a probability P' of 0.02 and a desired signal-to-cosmic noise level of 20 db (K = 100). In practice several hundreds to a thousand watts of power may be needed with practical antennas and receivers to achieve this duty cycle. Thus for a 800-mile path on 50 mc, with 3-element Yagis, a transmitter power of 1 kw and a receiver bandwidth of 1 kc, bursts with an average duration of 1 second are received at a rate of 300 to 400 per hour. Transmission is

thus available 10 to 15% of the time[44].

d. Distance and Frequency Range

While meteor height ranges of 80 to 120 km indicate a maximum range of about 2500 km for a single meteoric reflection, for a useful communication system the practical range appears to be 200 to 1000 miles (300 to 1500 km). The long range limitation is due to finite minimum vertical angles for antennas and a useful common volume of antenna beams in the ionosphere. The minimum range is dictated by the fact that the shorter the range the shorter the echo duration of a meteor reflection.

Regarding frequency, the 25-75 mc range appears most practical. The lower limit should be raised somewhat at this time of the sunspot cycle since MUF's are very high and interference and jamming via regular F layer propagation is possible. Backscatter self interference is being experienced on ionospheric scatter circuits right now on 30 to 40 mc systems.

Considerable information is desired on long duration echoes and assessment of regular F-layer interference to meteoric circuits. The assessment depends upon type of intelligence to be transmitted and the modulation method.

e. Communication Channel Capacity

The communication capacity C of an ideal channel with additive Gaussian noise is[45]

$$C = B \log_2 \left[1 + \left(\frac{S}{N}\right) \right] \text{, bits/sec} \qquad (19)$$

in which B is the channel bandwidth (cps), S the average signal power (watts) and N the average background noise power (watts). Speech transmission bandwidths of 3 kc and a S/N ratio of 6 db result in relatively intelligible speech. For this case, C is 7000 bits/sec. It is assumed the channel characteristics change but slightly during the time needed to transmit several hundreds of bits, i.e. quasi-stationary.

Weaver[45] has analyzed the behavior of C applicable to under-dense trail meteoric forward scatter. There does not appear much advantage in using (S/N) ratios less than unity because of increased bandwidth requirements. On the other hand (S/N) ratios 10 to 20 db restrict the capacity C to low values.

The signal power S for underdense trails has the form

$$S = S_m \, \epsilon^{-t/T_F} \qquad (20)$$

where S_m is the maximum signal power at $t = 0$ and T_F the decay time constant for forward scattering aforementioned. (At 50 mc, T_F is about 1 second for a 1000 km path; S_m is $P_F L$ which was discussed previously, where P_F is the free-space received power and L the loss with respect to free-space). In this time varying case the instantaneous communication capacity C (t) is

$$C(t) = B \log_2 \left[1 + \frac{S_m \, \epsilon^{-t/T_F}}{N_e KTB} \right] \qquad (21)$$

where N_e is the equivalent noise figure for cosmic noise (assumed limiting noise level). The total information capacity of a burst I is

$$I = \int_0^T C(t)\,dt \qquad (22)$$

Weaver has analyzed this equation for three possible methods of using the meteoric channel:

1) operate at a constant rate set by a threshold $(S/N)_{min}$ as long as (S/N) exceeds this level;

2) vary B with time so as to keep (S/N) constant; and

3) for fixed B, adjust for optimum (S/N).

Scheme (2) has but a slight (about 1.7 times the capacity) advantage over (1) if the time of use is the decay time constant. If $(S/N)_{max}$ varies from one meteor reflection to the next, then it may be worthwhile to consider varying the transmission rate accordingly.

III. SYSTEM DESIGN STUDIES

A. PROPAGATION FACTORS AND ALLIED SUBJECTS AFFECTING SYSTEM DESIGN

1. GENERAL

In assessing propagation mode and frequency choices for the air-to-ground application we are led to conclude that UHF tropospheric scatter is the best choice today for ranges out to 400 or 500 miles in the Arctic. This is based upon a study of many factors. From the standpoint of range, however, say 1000 miles or more, the state of the art is such that other mechanisms of transmission must be employed. Thus VHF ionospheric scatter might be employed at 600 to 1100 miles, and the shorter range being tropospheric. But VHF ionospheric and meteoric modes would require large antennas as on the aircraft. Also these modes are more efficient at the lower VHF frequencies, and here a circuit will be susceptible to longer distance noise, backscatter and interference or jamming, particularly during high sunspot number periods.

The lack of reliability of regular layer HF in the Arctic auroral zone has already been mentioned and is ruled out for USAF use here. It would be easy to jam and intercept.

A very efficient mode is that of LF and VLF. However, huge ground structures must be employed. It is easy to intercept and jam and channel capacity is more limited than higher frequency modes.

Relative cost is difficult to assess for various modes. If a very high order of reliability is required with multichannel operation, and if the system must be difficult to intercept and jam, then relative cost is favorable at VHF and UHF using tropospheric scatter. Antenna size must be small for high speed jet aircraft and not exposed and this suggests the UHF spectrum rather than lower frequencies.

An attempt has been made to tabulate the relative performance of various modes for each of several characteristics important in air-to-ground USAF communications. The weighting of these is open to some criticism and an overall addition of relative values might be misleading. With these remarks we conclude that UHF tropospheric scatter, within the state of the art today, is the recommended mode of transmission for USAF air-to-ground communications for ranges beyond horizon up to 400 or 500 miles and particularly in the difficult communication region of the Arctic.

2. TROPOSPHERIC SCATTER FACTORS AFFECTING SYSTEM DESIGN

Many factors of tropospheric scatter propagation will affect the choice of parameters of an air-to-ground communication system. The parameters include:

1) frequency
2) power
3) system receiver sensitivity

4) antenna gain and beamwidth
5) polarization
6) modulation
7) diversity

In the UHF region there appears to be little choice between say 200 and 1000 mc. The free space loss increases with frequency squared. The scatter loss increases with frequency. If constant apertures are assumed (i.e., a certain area can be used at any frequency), the gain product increases as frequency to the fourth power. However, coupling losses increase with decreasing beam size towards higher frequencies. Noise figure deteriorates the higher the frequency. Thus for a constant transmitter power, aperture sizes, and required channel signal-to-noise ratio about the same performance would be expected. Other assumptions might be made about keeping beam shapes the same at all frequencies, in which case the higher frequencies are slightly preferred.

We recommend the 225-400 mc band, and prefer that allocations be made for this special service at the high end of the band. Here a pair of bands separated 20 or 25 mc would suffice. The bands would be sufficiently wide in allocation, say, for perhaps 36 or more channels.

Regarding antenna beamwidths, there are restrictions on minimum size. In the aircraft it would appear that 50° in azimuth, with proviso for forward and rearward switching would suffice. 50° in elevation beamwidth is about as narrow as one would care to go in view of restricted vertical aperture in high-speed jet aircraft. Beam maxima should be horizontal, or if feasible slightly (a few degrees) depressed.

The ground antenna beams need to be broad in azimuth to give good coverage with minimum difficulty of location and tracking the communication signal. The vertical beamwidth can be quite narrow consistent with vertical aperture size and coupling loss. The horizon should be illuminated.

The gains required of the antennas should be at least a total of 40 db because of transmitter power limitation (approximately 1 KW in the aircraft). Thus we settle on the following approximate antenna characteristics:

bandwidth		5%
ground antenna		
	gain	30 db over isotropic
	azimuth	17.5°
	vertical	2.5°
aircraft antenna		
	gain	12 db over isotropic
	azimuth	50°
	vertical	50°

Horizontal or vertical polarizations have the same losses, so other than ground reflection coefficient there is little to choose. We may elect either polarization consistent with siting. There is a preference for vertical polarization in the aircraft from a structural point of view (see System Design).

On diversity, higher orders than unity are desired to minimize fast fading depths. Use of two antennas, spaced, can be accomplished as will be seen on the aircraft with the design proposed. Spaced diversity on the ground is considerably easier and is well known. Polarization discrimination (not polarization diversity since the two components, vertical and horizontal, fade coherently) may be used, but aircraft structure would rule against this. Frequency diversity might be added and has been considered. However, this complicates equipment further and is not being considered at this phase of the study for recommendation in equipment design.

Numerous advantages appear to favor SSB over FM although the latter has been used more in the past. SSB affords less bandwidth, is less susceptible to multipath in the presence of other aircraft and requires less power, according to Morrow et al[46].

B. PREDICTED SYSTEM PERFORMANCE

1. SYSTEM PARAMETERS

We shall assume the following system parameters in order to predict performance:

frequency (f_{mc})	400 mc
dual spaced diversity, combiner type (aircraft and ground)	
transmitter power	
airborne	1 kw
ground	10 kw
channel bandwidth	4 kc
modulation	SSB
channel signal-to-noise	10 db
antennas	
ground (30 db)	2.5° vertical, 17.5° horizontal
aircraft (12 db)	50° vertical, 50° horizontal

2. TOTAL FADING FACTOR AND RELIABILITY

The total fading factor F_T for dual combiner diversity is given in the curves of Figure 8. It must be remembered that our reliability F_T, say of 90%, means a slow fading factor for 90% of the hours and a fast fading factor for 90% of the hours (root sum square). It is thus a larger fading factor than just the slow fading factor F_s alone and is representative of reliability.

3. APERTURE MEDIUM COUPLING LOSS, L_c

The estimated aperture-medium coupling loss L_C is plotted in Figure 22 as a function of d_s, for the antenna beams assumed for the system.

4. REQUIRED RECEIVED POWER

For a 10 db rms signal-to-rms noise ratio, a noise figure of 5 db and a 4 kc bandwidth, the required power received is 10-204+5+36 or -153 dbw. For a 10 kw transmitter power, this gives a maximum allowable transmission loss of 193 db (a 1 kw transmitter power gives 183 db).

5. TOTAL TRANSMISSION LOSS CURVES

Using the line-of-sight curves of Figure 1 and equation (3) with the above factors the total transmission loss of the system was computed for various altitudes and reliabilities.

For 90% reliability and altitudes of 10,000 and 50,000 feet, the resulting values of L are shown in Figure 23 for the three climatic zones. For a given maximum allowable total L, it is seen that the maximum ranges are less in the Arctic than elsewhere, and are smaller for lower altitudes.

The predicted values of L for the Arctic for various altitudes for a given reliability appear in the next three figures. Figures 24, 25, and 26 apply to values of L for 50%, 90% and 99% reliabilities, respectively.

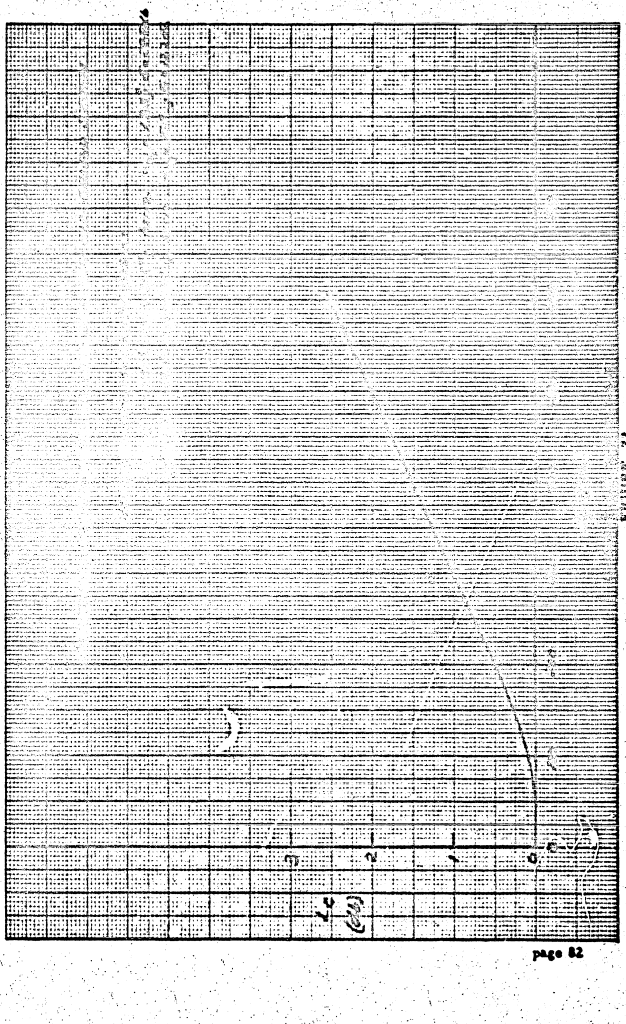


FIGURE 22

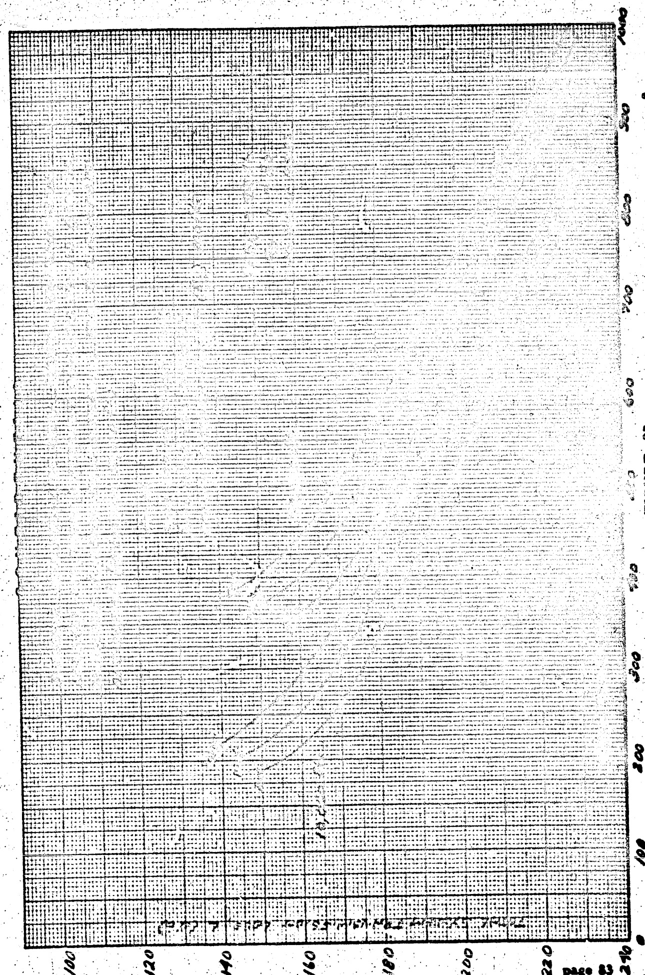

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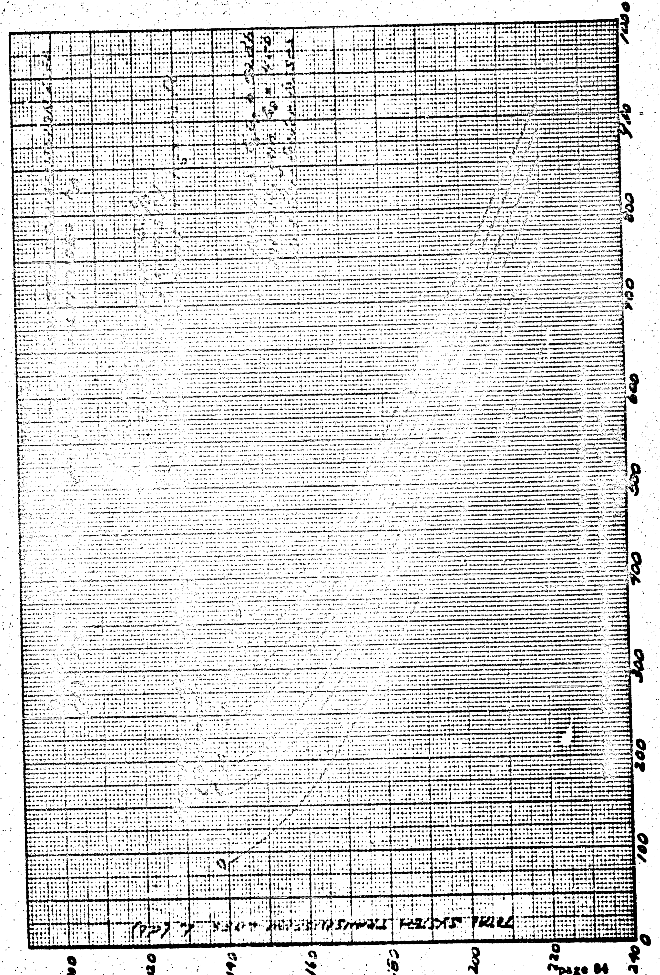
FIGURE 23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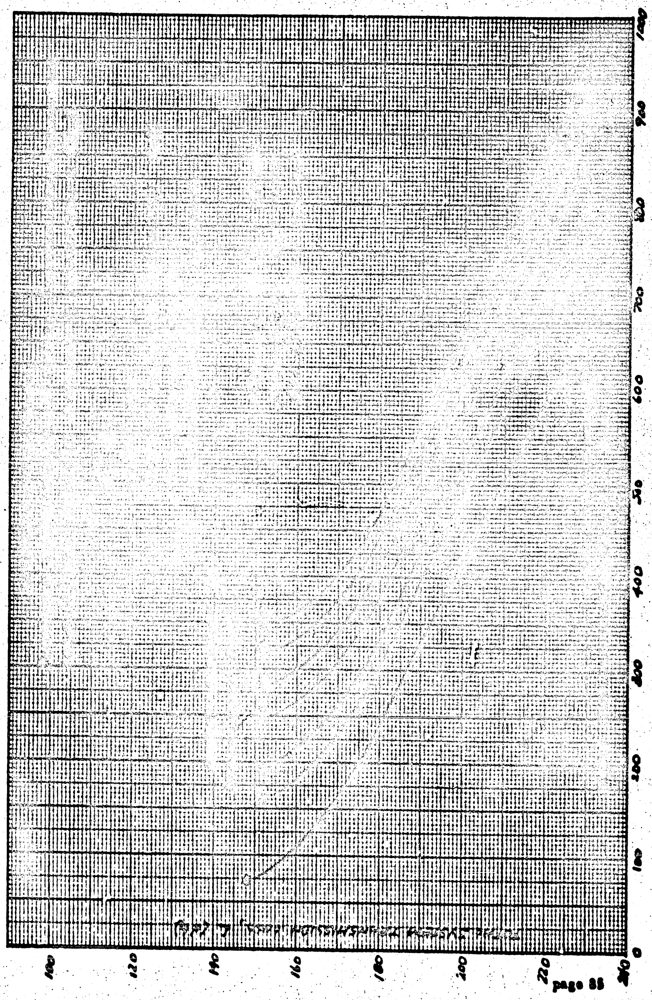

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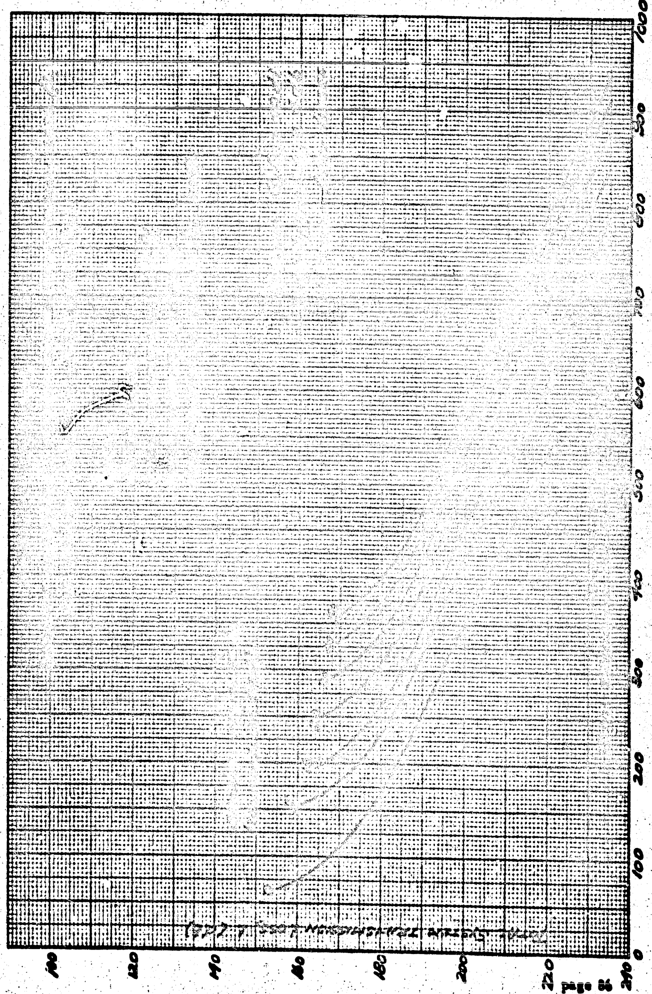

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6. ESTIMATED HEIGHT GAIN, G_H

Estimates were made for the height gain in the Arctic for various distances and altitudes for 90 and 99% reliabilities, based upon Figures 25 and 26. The results are plotted in Figure 27 for G_H for 90% and in Figure 28 for 99% reliabilities. In each of the figures for G_H, the curves show separately the dependence of G_H on height for a given total path distance d, and the dependence of G_H on d for a given altitude. For distances exceeding 350 miles, the curve of G_H vs h_A is practically the same for all distances. For a given height, G_H increases but slightly with distance beyond minimum G_H, unfortunately. For a given distance d, G_H increases with height as shown.

7. ESTIMATED MAXIMUM RANGE

The maximum range for given services is shown plotted for the Arctic in Figure 29 for various reliabilities and two transmitter powers (10 kw and 1 kw, typical of ground and airborne transmitters).

Shown in the figure is a curve, labelled 85% d_o, meaning an estimate of the line-of-sight communication range. The maximum range increases with height and transmitter power, decreasing for increasing reliability.

In the Midlatitudes, typical of some of the flight paths envisioned to Europe, the maximum ranges would be substantially larger than those shown in Figure 30. At 15,000 ft, the range is at least doubled by tropospheric scatter in the Arctic over the line-of-sight communication range, for 90% reliability.

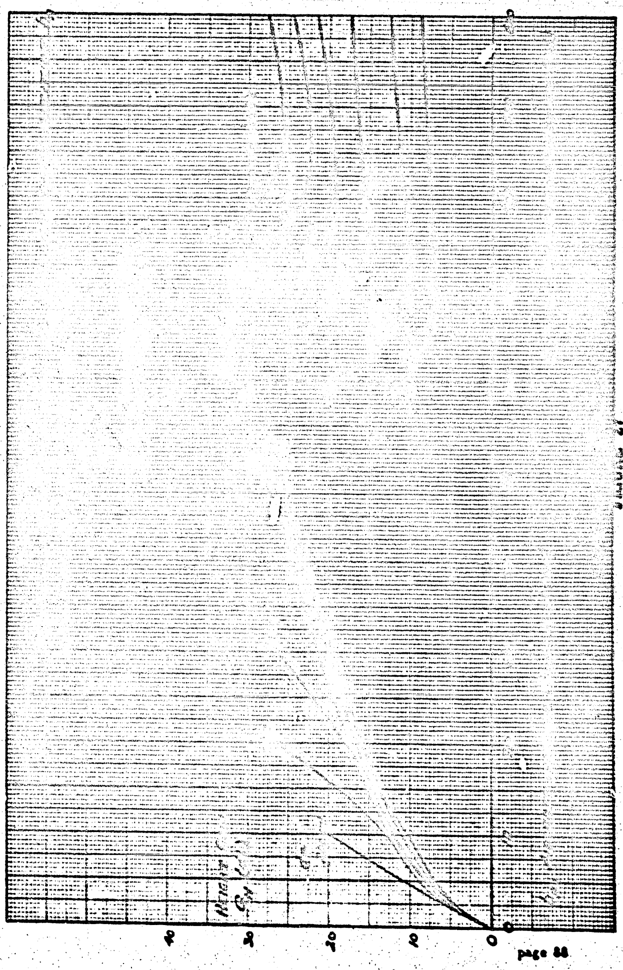
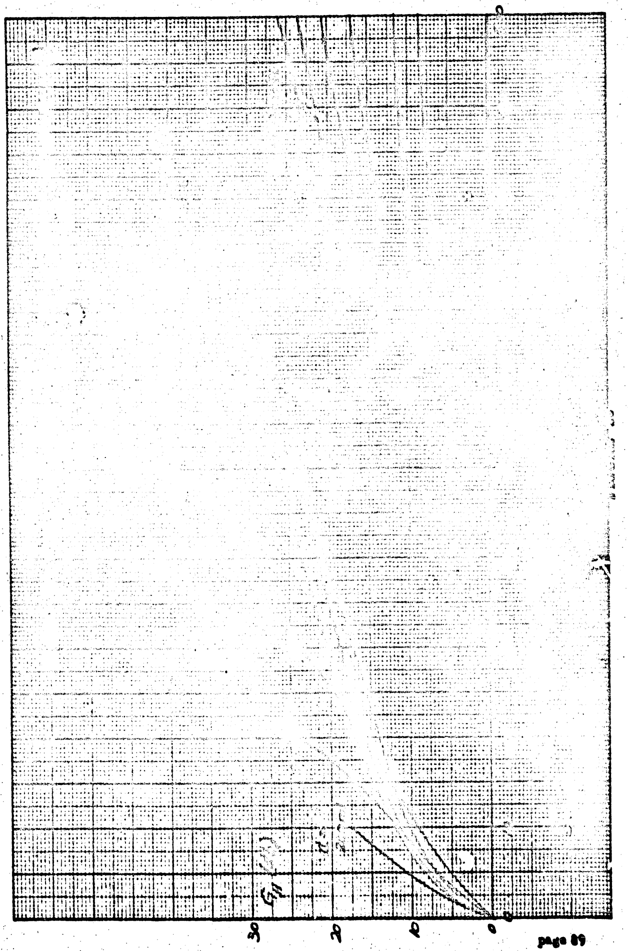
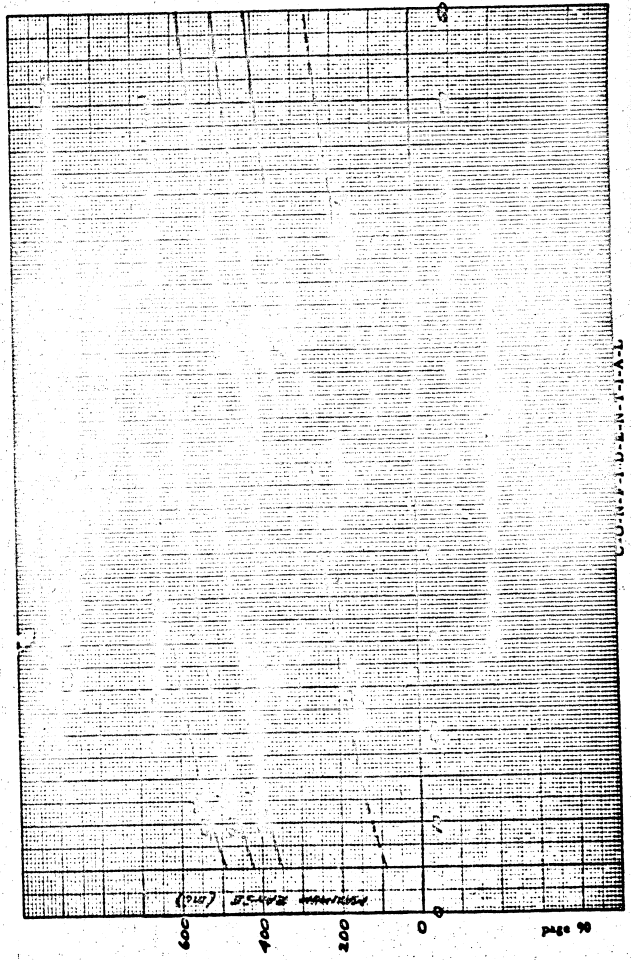


FIGURE 27



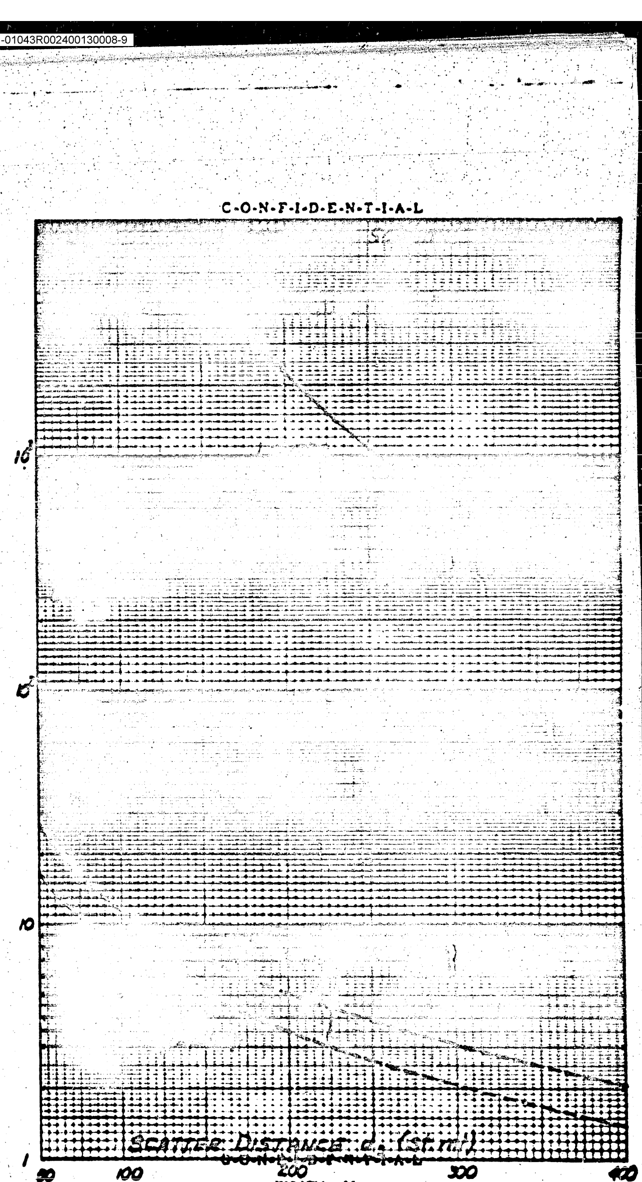


C-O-N-F-I-D-E-N-T-I-A-L

It is fair to mention that our curves have not considered the gains available from the use of companding, which may be of the order of 8 db. Referring to Figure 26 at 20,000 feet, 99% reliability, this would increase the maximum range for a 10 kw transmitter from 420 to 520 miles, i.e. by 100 miles.

8. BANDWIDTH OF TRANSMISSION

The influence of other aircraft on reducing bandwidth is shown in Figure 30. The medium itself allows a very large bandwidth to be employed. However, with aircraft crossing the beams the bandwidth allowed deteriorates an amount depending on the modulation employed. SSB with one pilot is worse than PCM, in turn worse than FM. However SSB with two pilots will allow 10 times the bandwidth shown for one pilot[46]. (These curves for aircraft multipath are based upon severe multipath assumptions but do show that aircraft in the scattering beams will deteriorate bandwidth).



FIGUR 30

C. EQUIPMENT DESIGN STUDIES

1. ANTENNA DESIGN STUDIES

a. Introduction

In order to accomplish effective air-to-ground communications using tropospheric scatter techniques, the antennas incorporated in the overall design must be such that the maximum operational efficiency is derived from the system. The requirements for this for both air and ground systems are the same; however, the aircraft system poses some problems due to size, weight, positioning on the aircraft, and the effect of the aerodynamics of the ship by the use of various types of antennas. On the other hand, the ground system is not limited in design except by required coverages, cost, and ease of operation. In both systems the highest practical directivity to afford some protection from interference, jamming, or undesirable monitoring of the circuit is desirable. The beams must not be too narrow in azimuth to ease tracking. Vertical polarization is preferred* although other polarizations are acceptable and are proposed in some design considerations. It is also desirable to use two antennas at either end of the circuit so spaced as to offer two distinct transmission paths, thus providing space diversity operation.

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*From the propagation standpoint, either horizontal or vertical polarization is acceptable. However, in considering transmission from airborne, flush-mounted antennas, vertical polarization offers the most advantages in attaining the desired radiation characteristics.

For the 225-400 mc range of operation for tropospheric scatter air-to-ground communications, several antenna types have been studied. These types can be specified under two general classifications:

1) ground station antennas

2) airborne antennas.

Typical ground station antennas studied are sectional parabolas, corner reflectors, and broadside arrays. A wide variety of each of these types have extensive use in other communications systems and a review of the state of the art indicates that many possibilities exist for usage in this particular application.

Installation of antennas on present-day aircraft should be accomplished without introducing additional drag. This suggests flush mounting for any antenna system. A study was made of all possible types of flush mounted antennas that could be used which have approximately 10 db gain. Antennas such as end-fire slots, cavity-type, traveling wave antennas, semi-flush yagi arrays, stubs, and helix types were surveyed as potential radiators to fulfill the requirements of the system.

b. Requirements

(1) Ground Antennas

To operate effectively, the ground station antennas should have a configuration such that the radiated beam would have a gain of 30 db with respect to an isotropic source, minimum sidelobes, and half power beamwidth angles of about three degrees and seventeen degrees in the elevation and azimuth respectively.

The arrays may have provisions to allow azimuth rotation for alignment purposes. At each ground station there will be two antenna arrays spaced approximately one hundred and fifty wavelengths (600 feet) apart so that two distinct paths of transmissions may be utilized.

The high gain, beam shaping, and space diversity are essential requirements for the ground station.

(2) Aircraft Antennas

It would be desirable to obtain the same characteristics in the aircraft antennas as are obtainable in the ground station arrays. However, the limitation on size, weight, and available space for mounting immediately limit the acceptable design.

The antennas should be directional with a gain of 10 db with respect to an isotropic source and have half-power beamwidths in both planes of about 50°.

c. Analysis

(1) Ground Station Antennas

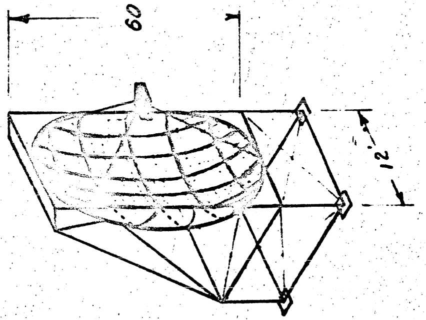
Selection of one type of a variety of different existing antenna systems is based on many factors, among which consideration should be devoted principally to:

1) Performance
2) Reliability
3) Cost
4) Ease of Maintenance
5) Simplicity.

The investigation of possible antenna designs for this requirement evolved around the need of such large structures for antennas to develop 30 db gain. Structures such as that shown in Figure 31, corner reflectors, and "vee" antenna arrays were analyzed. Broadside arrays, although having narrow band characteristics, were also included in the study.

(a) Corner Reflectors

A common type of directional antenna that has enjoyed widespread use is the corner reflector. Basically, this device is made up of two flat reflecting sheets that intersect at an angle, usually 90°, which are excited by a primary source such as dipole feed. To obtain the desired radiation characteristics, it is necessary to array colinear radiators in the focus of the reflector. By proper choice of the number of primary sources and reflector parameters, an antenna may be developed that satisfies the requirements of the system.



SECTIONAL PARABOLA

FIGURE 31

1. Structure Size

An analysis of the design data of corner reflectors, as outlined in Kraus[48], indicates that a 60° corner reflector for a given antenna-to-corner spacing (in wavelengths) possesses a higher gain than others. With an attempt to develop an antenna of the highest gain possible for a fixed size, we have designed an antenna that appears to be a realistic approach to this problem.

The size of the reflector can be determined after the line source parameters have been established. Since the beamwidth in elevation is required to be 2.5°, the length of the array can be approximated by:[49]

$$\theta = \frac{51^\circ}{D}\lambda$$

where θ = beamwidth in degrees at the half power point.

λ = free-space wavelength at the operating frequency.

D = aperture dimension of the reflector in the plane of polarization.

For an antenna operating at 400 mc, where $\lambda = 2.46$ feet, the array length is derived:

$$\theta = \frac{51^\circ\lambda}{D}$$

$$D = \frac{51^\circ\lambda}{2.5^\circ} = \frac{(51)(2.46)}{2.5} \sim 50 \text{ feet}$$

To obtain maximum efficiency from the array in the elevation plane, the reflector length should be greater than the array length. The literature indicates that a corner reflector height of 60 feet is appropriate for a line source of 50 feet in length.

In the azimuthal plane where it is desired to have a 17.5° beamwidth and the intensity across the aperture is of constant phase and tapers smoothly from the center to the edges of the reflector, the beamwidth is given by:[49]

$$\theta^\circ = \frac{70^\circ}{D}$$

Therefore,

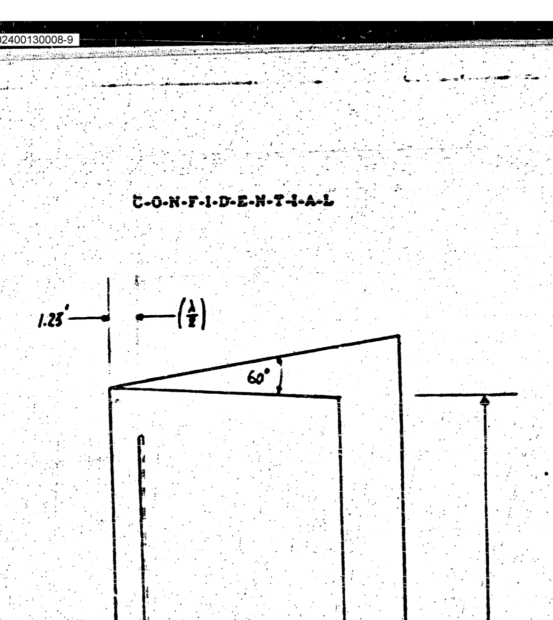
$$D = \frac{(70)\ (2.46)}{17.5} \approx 10 \text{ feet}$$

If one chooses an antenna-to-corner spacing of .5λ, the system would appear as illustrated in Figure 32.

2. Corner Reflector Illuminators

An interesting problem arises in attempting to illuminate a reflector of the size shown in Figure 32, and it is one of determining a suitable colinear array for the primary source. A unique design is presented that affords maximum simplicity of construction at minimum expense.

Common colinear arrays for corner reflector sources consist of .4λ dipoles appropriately spaced and phased to obtain the necessary illumination for the reflector. For a 50 foot array, where .4λ ≈ 1 foot, the number of elements (n) required is $n = \frac{50}{1} = 50$ radiators. Obviously such an



CORNER REFLECTOR
FIGURE 32

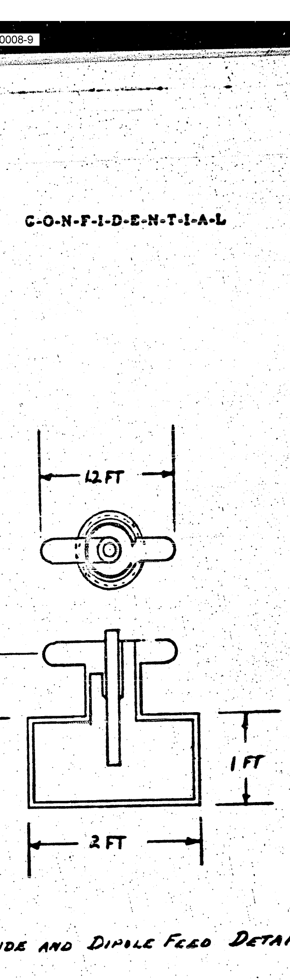
array requires a fairly complex feeder system at 400 mc and could present serious structural problems. Such a design was ruled out in favor of a waveguide transmission line that is used to excite a number of dipole feeds that are properly positioned on the broad face of the waveguide as shown in cross-section in Figure 33[50].

A preliminary design of a waveguide linear array feed has been completed that should satisfy the operational requirements for the purposes considered herein. The waveguide cross-section is 24" x 12" and at 400 mc, the guide wavelength is approximately 3.2 feet, where $\lambda_g / \lambda = 1.3$. For a 50-foot resonant[51] array and a $.5 \lambda_g$ spacing of the radiating elements, the number of elements required becomes $n = \frac{50 \text{ ft.}}{1.6 \text{ ft.}} \approx 31$.

Thus, if a 1 5/8" diameter coax line from the high power transmitter (10 kw) terminates in a coax-to-waveguide transition which in turn transfers energy down the waveguide and with suitable coupling to the 31 radiators it is possible to excite the 60 feet x 10 feet corner reflector and develop a 2.5° x 17.5° beam at 400 mc. This feed structure is shown in Figure 34. For purposes of pattern symmetry an even number of radiators are required in the array and therefore 32 dipoles are illustrated.

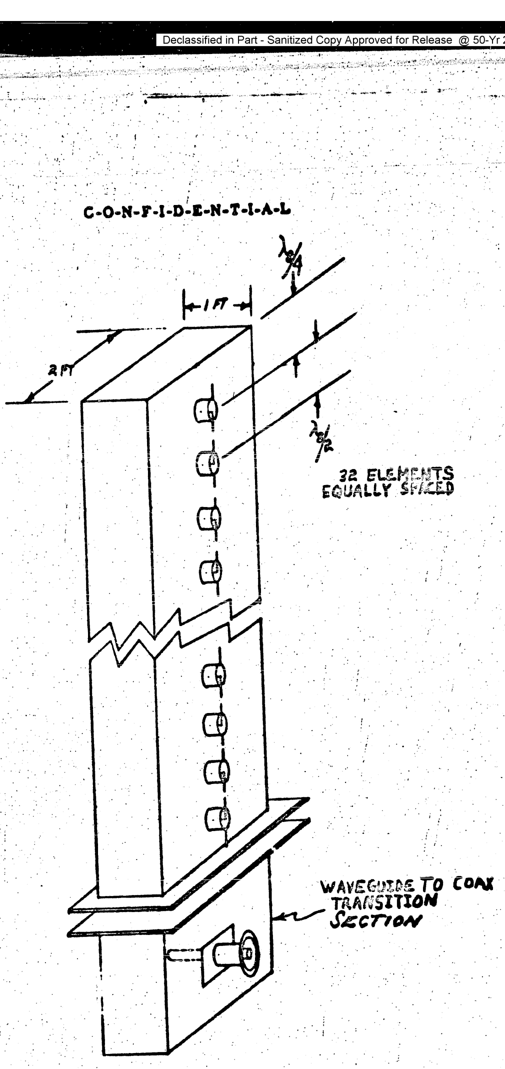
(b) Antenna Array Configuration

Based on the preceeding analysis of the reflector size and the line source, a design is proposed for a practical antenna system. The system would consist of two 60 foot x 10 foot flat reflectors arranged to form



WAVEGUIDE AND DIPOLE FEED DETAILS

FIGURE 33



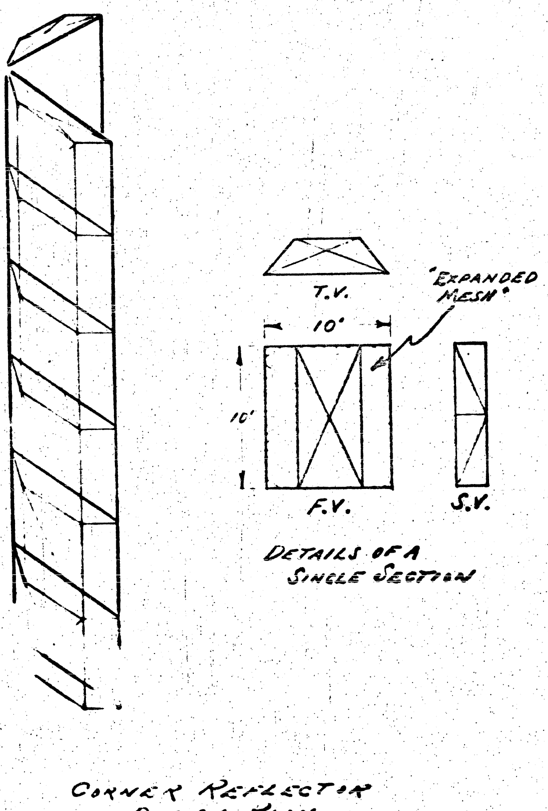
Linear Array of Dipoles

FIGURE 34

a corner reflector as shown in Figure 32. The reflector sides are made up of six (6) 10 foot x 10 foot sections of 2" expanded aluminum mesh fastened to suitable framing that provides the necessary structure for allowing erection. The panel sections being identical can be fabricated in an economical manner and produced so that pre-erection is not necessary. One scheme is shown in Figure 35. The array design can be such that the waveguide feed is positioned at the apex of the corner and be used as a structural member in the system. The radiating elements will protrude from the broad face into the focus of the reflector as shown in Figure 35. It is conceived that the reflector sides can be fastened to the edges of the broad faces of the waveguide using common 1/2" diameter bolts and nuts. If, in a more complete mechanical analysis it is found desirable to use a waveguide of smaller cross-section then various techniques such as those employed in broad-banding waveguides may be employed to accomplish this end.

(c) A Resonant-Slotted Waveguide Array Design (Broadside Array)

Another practical antenna design is considered based on the characteristics of radiation from shunt-loaded slots[52, 53, 54] cut in the narrow face of a waveguide. Using the same array analysis as previously described an antenna design has been established that is electrically and mechanically simpler than the corner reflector antenna. This is shown in Figure 37, and except for the feed all other characteristics would be the same for the corner reflector design.



DETAILS OF A SINGLE SECTION

CORNER REFLECTOR DESIGN PLAN

FIGURE 35

LINEAR ARRAY IN CORNER REFLECTOR

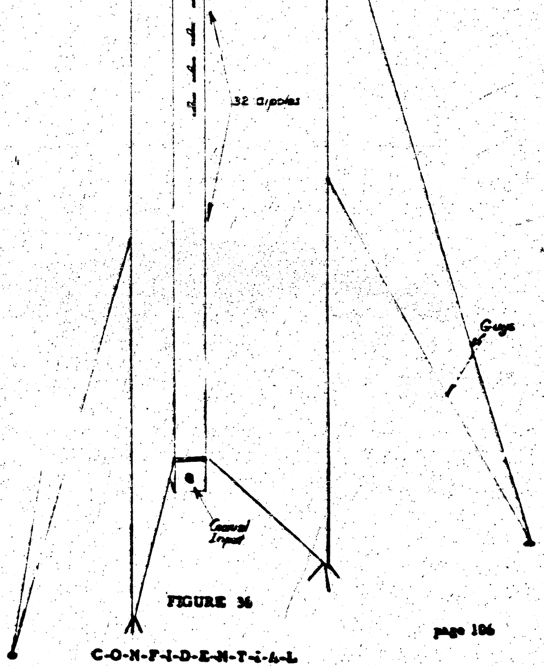
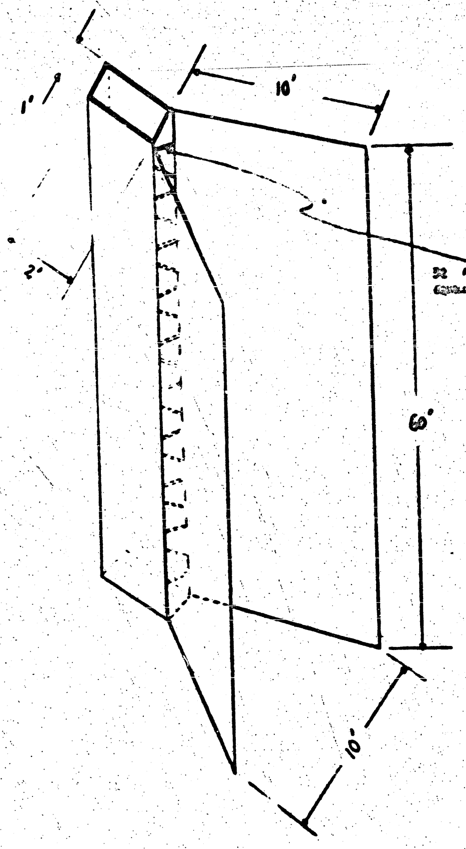


FIGURE 36



WAVEGUIDE LINEAR ARRAY

FIGURE 37

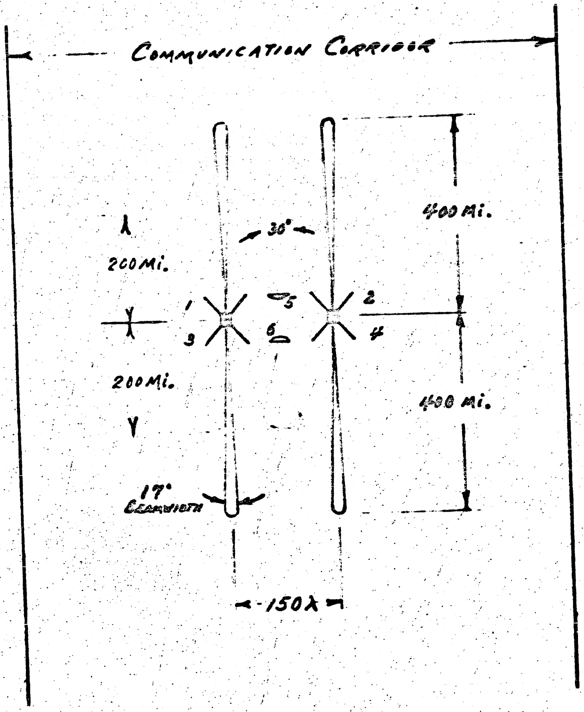
(d) **Sectional-Parabola Antennas**[55]

Many common scatter communication systems employ 60 foot diameter parabolic reflectors, with the horn radiators for the primary source. Although these antennas are large and expensive, practical designs exist that have facilitated production of large numbers of these systems.

In view of the fact that large parabolic type reflectors can be developed on a practical basis, it was considered desirable to design a section of a paraboloid as a reflector. One such design having the proportions shown in Figure 31, and incorporating a horn feed should satisfy the requirements. It is obvious that a feed system for such a reflector can be a simple horn and the simplicity of feed design is procured at the expense of requiring larger structures than previously described using arrays as sources.

(e) Tentative Design For A Typical Ground Station

For the proposed system of tropospheric scatter air-to-ground communications, one terminal may be required to have four large arrays and two smaller arrays in operation. The operational aspects of the system will be described elsewhere in this report, however, the station antenna installation plan is described below, consistent with the operational aspects. If corner reflectors and waveguide arrays are to comprise an antenna then the plan shown in Figure 38 is considered an efficient installation. It is to be noted that the two high gain arrays, (No. 1 and No. 2) pointing in the same direction provide coverage on a space diversity basis



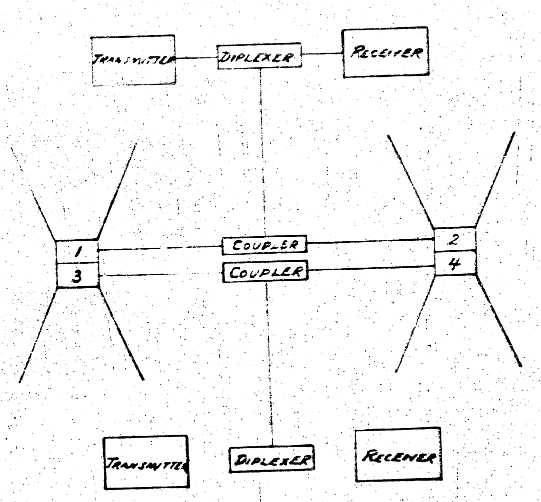
Plan View

Schematic of Ground Station Installation
FIGURE 38

up to 400 miles. Two more identical arrays (No. 3 and No. 4) are shown pointing 180° away from the first two arrays. The "paired" arrays serve the same purpose but differ in the direction of communication. Figure 38 is a schematic of the high gain array arrangement. Also illustrated in Figure 38 are smaller antennas designated No. 5 and No. 6. These antennas could be 6 foot diameter parabolic reflectors, having approximately a 30° beam at 400 mc. They serve to provide short range coverage and allow flight paths to be off the high gain array beam axis and still maintain suitable communications at shorter range.

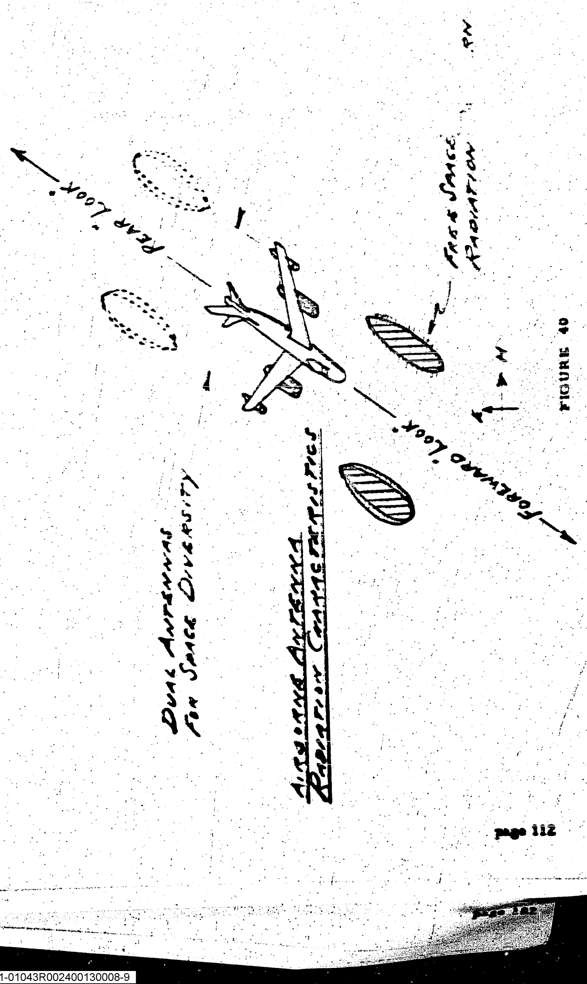
(2) Airborne Antennas

A major problem involved in the aircraft antennas is the effect of the airframe on the radiation pattern. Below 1000 mc the radiation characteristics of an antenna of practical size are, to a large measure, dependent upon the aircraft structure. Hence the antennas themselves may or may not have free-space or ground plane patterns which display the desired characteristics. The achievement of a good specialized system depends upon finding locations such that the performance of the antenna in the presence of airframe is satisfactory. Ideally, the radiation characteristics of such a system for aircraft installation to satisfy the requirements should have approximately a 50° E-plane and a 50° H-plane beamwidth. The axis of the radiated beam should coincide with the longitudinal axis of the fuselage as illustrated in Figure 40. In addition, the antenna system should be capable of radiating either foreward or aft of the craft, not simultaneously.



Schematic High Gain Array Arrangement

FIGURE 39



Airborne Antenna Radiation Characteristics

FIGURE 40

and the forward or rear "look" would be dependent upon whether the craft is going toward or away from the ground station being contacted. The study was mainly concerned with determining the existing state of the art of flush mounted aircraft communication antennas. Also considered was exploring the possibility of utilizing new techniques to design a system suitable for aircraft installation.

The final selection of an aircraft antenna was based primarily on the operational efficiency that can be obtained with the various proven designs. Many antennas that would provide a 10 db gain over an isotropic source such as arrays consisting of flush mounted slots, vertical stubs, yagi arrays, and helices were investigated.

It was determined that in order to provide some space diversity transmission the antennas should be mounted outboard on the wings, and designed so that forward or rear directivity could be accomplished by some method of switching.

The following is an analysis of the antenna types investigated for use on the aircraft:

a. Resonant 1/2λ Slots

An array of flush mounted, 1/2λ, resonant slots was model studied. Physically such an array could be mounted on the extremities of the wings and offer no aerodynamic problems. The studies indicated that due to the nature

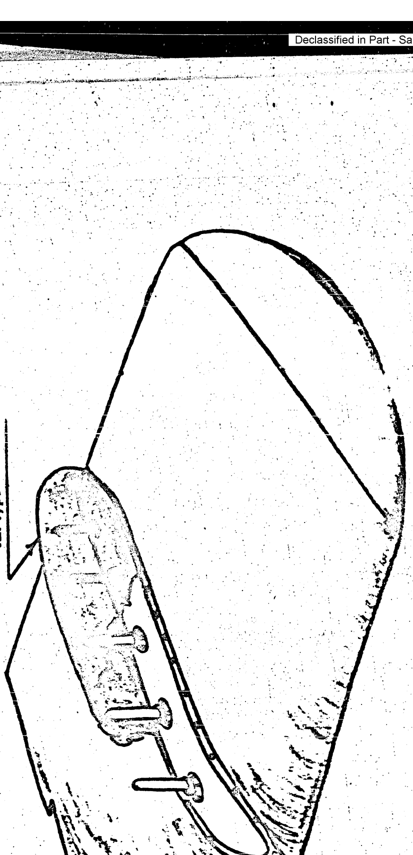
of flush mounted slots on a finite ground plane the major lobe of the radiated energy could not be directed forward or aft efficiently. In all cases the axis of the radiated pattern was 25° - 35° above the axis of the ground plane. Investigation into existing antennas used in aircraft indicated that the same problem is encountered[56].

b. Traveling Wave Slots

Flush mounted traveling wave slots have desirable pattern and impedance characteristics for use in this application[56]. However, the usual length is about 7λ (17 ft) long by 4λ (10 ft) wide. They fire in only one direction and cannot be reversed except by reversing the array. This necessitates the use of two antennas in each wing to accomplish the operational requirements. The problems involved in mounting such large structures into an aircraft wing, if the wing should be large enough, are evident.

c. Vertical Stubs

The use of a half yagi (1/4λ vertical stub parasitic array) or an array of driven 1/4λ vertical stubs has been considered as a practical antenna for use in this application. The stubs would be placed on the wing as shown in Figure 41. Positioning above or below the wing would depend on the effect received from the airframe on the radiated pattern. Such an array can be directed forward or aft by changing the driver element in the case of the parasitic array or reversing the phase in the case of the driven array. Both arrays would be physically about the same size; the stubs 7 1/2 inches long



END FIRE QUARTER WAVE DIPOLE ARRAY

FIGURE 41

and spaced 7 1/2 inches in the driven array and about the same distance in the parasitic array for a frequency of 400 mc. The array would be covered with a fin type radome which would be about four feet long by one foot high and about four inches wide overall. The stub array appears to be the simplest method of accomplishing the desired results.

d. Helices[48]

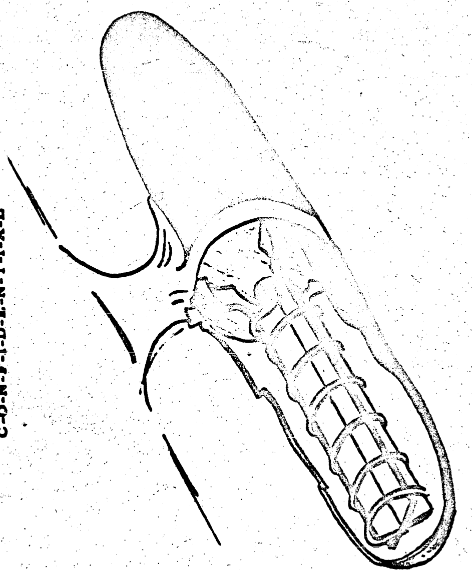
An antenna that could be pod mounted and is quite adaptable for use in this communications problem is the circularly polarised helix. A standard six turn, 12 to 15 degrees pitch angles, helix mounted on a circular ground plane has overall dimensions of about three feet long by about 1λ in circumference (ten inches in diameter). The ground plane would be about 22 inches in diameter. A sketch of the proposed design is shown in Figure 42.

Two such antennas mounted back to back with a common ground plane would provide the proper pattern coverage. The helix is a broad band antenna with good impedance characteristics of a 1.75 to 1 band.

Half power beam width of the helix may be obtained by the quasi-empirical relations:

$$\text{Beam Width} = \frac{52}{C_\lambda \sqrt{n\, S_\lambda}}$$

Where C_λ is the circumference of the helix, n is the number of turns, and S_λ is the spacing between turns (center to center). Beam widths in the order



POD MOUNTED HELICES

FIGURE 42

of 50° in both planes, and overall gain in the order of 10 to 12 db over an isotropic source are easily obtainable.

The helix appears to be the most satisfactory externally or radome mounted antenna available with the proper electrical characteristics for use in the scatter communication circuit. It is felt that if it is pod mounted on the wing tips the airframe would have a little effect on the radiation pattern.

2. TRANSMITTING AND RECEIVING EQUIPMENT

a. General Considerations

The following system parameters are the basis for the purposes of this equipment design proposal:

Frequency	150 mc to 400 mc
Transmitted Power:	
Airborne Unit	1 kw average
Ground Station	10 kw average

Modulation:
Single voice channel using single sideband modulation with a low level pilot signal.

Receiver:
150 mc - 400 mc fixed frequency, noise figure less than 5 db.

Provisions for diversity operation.

b. Airborne Equipment

Specific equipment must be tailored to the type of aircraft and the external facilities available. In general, it would seem more desirable to use liquid cooling to minimize the size of the actual electronic packages. While the combined weights of the liquid to air heat exchanger and associated coolant lines may result in the total weight of the equipment exceeding that of an air cooled package there is a distinct advantage in using several individual small sealed packages. Air cooling requires either a maximum altitude limitation or a necessity for the installation to be in a pressurized compartment with probable failure if the pressure is lost.

The transmitter power stages proposed use a single type 4W 300A driving a pair of GL6019 tetrodes. At the frequencies being considered here and with Class AB or B linear amplifiers, a very high gain will be realized. Other stages will require only miniature or sub-miniature tubes. All low level stages should be constructed using the techniques of current telemeter equipment design; making maximum use of printed circuitry, heat dissipating wrap-around tube shields, and individual component brackets. Wherever possible, the use of shock and vibration mounts should be avoided and instead, component parts and circuits should be designed to withstand the environment. All power supplies proposed are 400 cycle, three phase designs, utilizing silicon diode rectifier stacks, high temperature, oil or evaporation cooled, transformers and assembled as liquid cooled, sealed components.

The single-sideband type of modulation requires accurate carrier reinsertion. In addition, due to the radial velocity of the aircraft, corrections

must be made for the doppler shift of the received signals. It is proposed here that all reference frequencies be generated on the aircraft and all correction for reference errors and doppler shift be made at the ground station. Sufficiently stable crystal-controlled oscillators are available for aircraft use and, by utilizing these components, the more complicated error correcting devices can be ground equipment.

Figure 43 represents the airborne equipment. Current techniques will allow the 12.5 mc to 25 mc stable oscillator source for the high frequency injection to have a 24 hour stability in the order of 1 parts in 10^{6}. Existing airborne navigational 2.5 mc or 5 mc standard oscillators approach a 24 hour stability of several parts in 10^{8}*. Thus, in the VHF-UHF region, the aircraft velocity can introduce the major frequency error. Although a frequency discriminator circuit similar to that proposed for the ground station could also be included as part of the aircraft equipment, the addition of this feature will lower the system reliability by requiring servo-controlled motor driven equipment, or an extremely stable electronic tuning device. In addition, the pilot signal would need be well above the noise level and the airborne equipment could be jammed by relatively weak interfering CW signals. Instead, as will be described below, a relatively simple calling procedure will avoid the necessity of airborne pilot detection equipment. One exception to this simplification might be in the case of utilization of

* See for example, Borg Equipment Division, G. W. Borg Corporation, Model 150b Airborne Frequency Standard, WADC Exhibit WCLN-1135.

AIRBORNE EQUIPMENT SYSTEM

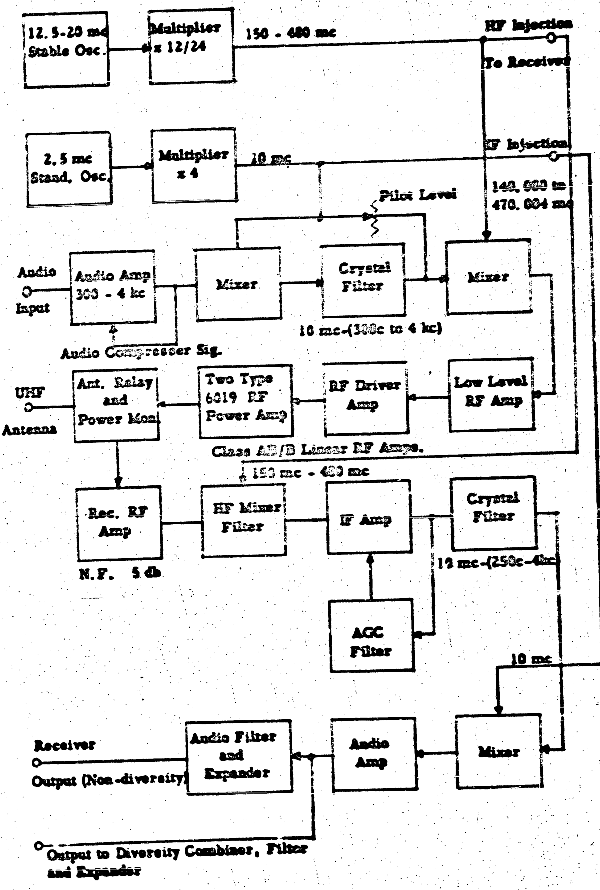


FIGURE 43

several narrow band pulse code signals such as teletype, instead of the single voice channel. In this case, a simple reference pilot may be required at both terminals to provide precise decoding data. However, the servo-frequency correction equipment would not be necessary in the aircraft.

The use of 10 mc crystal filters in the airborne transmitter and receiver may require special care in vibration and shock isolation and some form of crude temperature control. However, this filter design allows considerable simplification by eliminating the need for additional heterodyning.

The rest of the proposed airborne system is conventional. Diversity reception can be obtained by combining, at the audio frequency level, using parallel connected cathode followers in a "signal squaring" type circuit. Level clipping and audio compression - expansion is required to limit the dynamic range of the transmitter power stages. However, if only one voice signal is to be used, the amplitude control and intermodulation distortion specifications need not be severe.

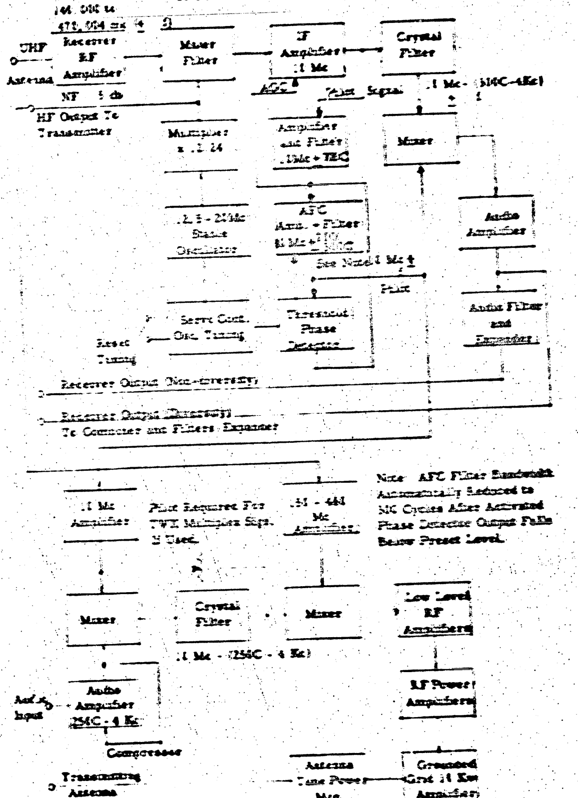
c. Ground Equipment

A ten kilowatt VHF-UHF transmitter is not a particularly difficult design problem. For frequencies below about 250 mc, there are many choices of acceptable power tubes. In the spectrum between 250 mc and 400 mc the choice of reliable power amplifier tubes is more limited, but not impractical. Possible tube choices at the higher frequencies are the Eimac 4W 20,000 A tetrode, the GL-6019 triode, a ring circuit of several GL-6182 tetrodes, the RCA 6448 tetrode, the RCA 6806, the GL-6251, and certain power klystrons such as the Eimac X509. Although the high power gain of the klystrons is very attractive, much of the advantage of reduced drive level is lost by the added requirement of many extra power supplies, each requiring careful adjustment and regulation. The klystron circuitry is particularly difficult with single sideband modulation. All of these possible tube choices are expensive, limited production, designs and some have serious shortcomings in performance. Thus, a careful study of the actual performance records of existing applications of each type is highly recommended. In the spectrum below 300 mc, the final choice would probably be a conventional grounded grid triode or tetrode design.

With the exception of the A.F.C. circuits, inspection of Figure 44 will show the overall proposed ground station equipment to be conventional in design. The receiver requires a high frequency injection oscillator with a short term stability of perhaps several parts in 10^8. The 10 mc pilot signal, transmitted by the airborne equipment, is filtered from the IF amplifier output, amplified and used to transpose the speech signals directly to the audio range. Thus, as long as the high frequency injection signal is within 10 mc + 50 cycles of the incoming signal, a usable output is obtained.

If, due to doppler shift or airborne equipment errors, the received signal is as great as 700 cycles off the assigned frequency, the phase shift

GROUND STATION SYSTEM



Note: AFC Filter Bandwidth Automatically Reduced to 50 Cycles After Activated Phase Detector Output Falls Below Preset Level.

Frequency = 156 to 446 Mc
Less 11 Mc - (250 to 4 Kc)
Plus Corrections and [illegible]
Frequency Errors and Doppler Shift

Ground Station System Block Diagram

FIGURE 44

across the second pilot signal amplifier filter in the ground receiver will actuate the phase detector and cause the servo-motor drive to vary the local high frequency injection frequency and bring the pilot frequency back to 10 mc. If the received signal is low, the frequency of the injection oscillator will be raised.

The receiver HF injection signal is also utilized as the ground transmitter high frequency signal. Thus a doppler frequency shift downwards in the received signal will cause the ground station transmitted signal to be equally displaced upwards, and the signal received at the aircraft will be the desired frequency.

The wide range of the possible combined doppler shifts and equipment errors requires that the ground receiver pilot amplifier filters have a passband extending into the desired audio range (depending of course on the radio frequency chosen and the aircraft relative velocity). To prevent speech frequencies interfering with the AFC action, several procedures may be used: offset the pilot frequency further below the speech signals, which would require additional equipment in both the airborne and ground stations; establish initial contact in either direction by means of a 1000 cycle audio code plus the pilot channel, thus providing sufficient time for synchronizing the ground station equipment; or establish initial contact using an audio filter in the aircraft modulator to eliminate speech frequencies below 700 cycles until the ground station operator can be clearly understood by the aircraft operator. The last

two methods result in simpler equipment than the first method and either should be satisfactory with trained operators. The use of a spectrum between 700 cycles and 4 kc for speech during the initial contact only, will not result in an unintelligible signal to the ground operator and, if initial contact is made by the aircraft operator, no unintelligible transmissions will result. Should the ground station initiate the contact, a speech frequency translation as large as 7000 cycles might prove unintelligible, however it will be clearly heard by the aircraft operator who can correct the signal by merely transmitting his pilot tone for a few seconds before responding. Thus, the third suggested method is the simplest and most reliable from the design standpoint and should prove satisfactory operationally.

After each contact the ground station operator should reset his injection frequency oscillator to its normal frequency by operating a position reference circuit included as part of the servo control system.

If it is desired to have the ground station simultaneously monitor and contact several aircraft, the error correcting system proposed here could be modified to cause each received signal and pilot tone to vary the center frequency of the associated sub-carrier in the ground station transmitted signal. A guard band of 1400 cycles would be used between the nominal sideband extremes of each sub-carrier.

IV. IMPROVEMENT IN MILITARY COMMUNICATIONS

A. IMPROVEMENTS TO MILITARY COMMUNICATIONS RESULTING FROM APPLICATIONS OF THE BEYOND-LINE-OF-SIGHT COMMUNICATORS TECHNIQUE

The implementation of beyond-line-of-sight techniques at UHF extends the communication range of military UHF command radio equipment from about 170 miles to more than 400 or 500 miles a altitudes of 20,000 feet in the Arctic. The ranges are greater in Midlatitudes. This has been demonstrated in Figure 29. This doubling or trebling of the ground-air-range at first look does not appear to assure a reliable (90%) communication facility meeting long range requirements. Yet when this new capability is placed in the context of the USAF World-Wide Ground Point-To-Point Communication System (System 456L), highly reliable performance characteristics become apparent, as compared with the present use of high frequency (HF) transmissions for long range communication.

First, let us review the existing SAC reporting system as an example of today's operation. As shown in Figure 45, the air to ground link is High Frequency Radio with its erratic reliability and ease of interception. Because of the uncertainty of the high frequency link, air to ground reporting and ground-to-air command transmissions usually require a considerable amount of time plus a number of test transmissions at several frequencies to find a channel that will be useful.

All active USAF bases will be netted, by several different propagation modes, th form the World Wide Communications System. It is apparent that a

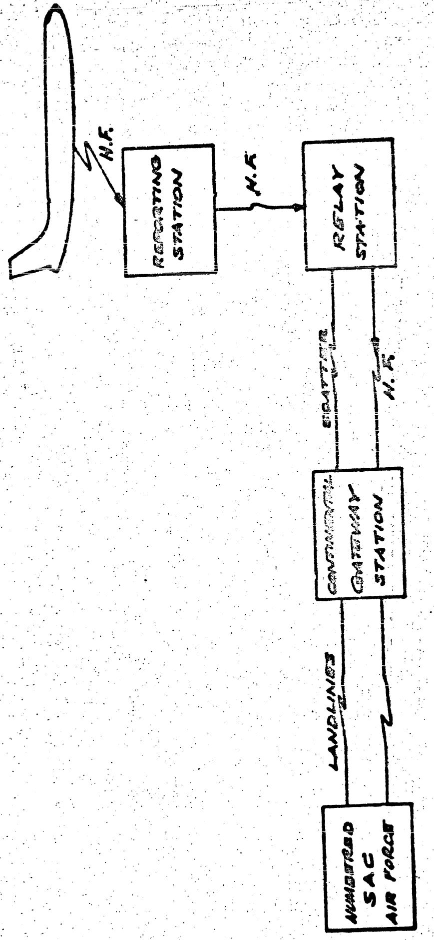


FIGURE 45 - EXAMPLE OF PRESENT COMM. REPORTING - S.O.C.S. AND SAC COMM.

reliable point-to-point communication circuit exists to any USAF base, once a message is entered into the ground net. This network is typified in Figure 46. With this philosophy in mind, it now becomes important to have a highly reliable air-ground circuit, which at all times can communicate with a station in the ground communication system. The increase in range by UHF scatter makes it possible to find enough geographical sites to implement a reliable communications circuit for long distance flights.

Let us now look at the new communication building block. The air-ground coverages supplied is shown in Figure 47. This coverage is in two parts, a circular coverage of 200 mile radius (line-of-sight at 20,000 feet) plus a scatter coverage extended by at least 200 miles more. This service is generated by a dual ground installation; an omnidirectional service of the conventional type of UHF command communication, plus a dual UHF scatter ground-air link. One scatter circuit covers the left half of the service area, the other the right half. This is due to the necessary use of high gain antennas which have a 17.5 degree beamwidth. The available bandwidths as shown in Figure 30 in the scatter circuit are more than ample to provide the same 100 kc channels as in the UHF command allocations.

Let us now see how this new concept can result in a reliable long range communications channel for:

1) A SAC Bomber on a mission from Presque Isle, Maine to Moscow

2) A MATS Flight from New York to London, England.

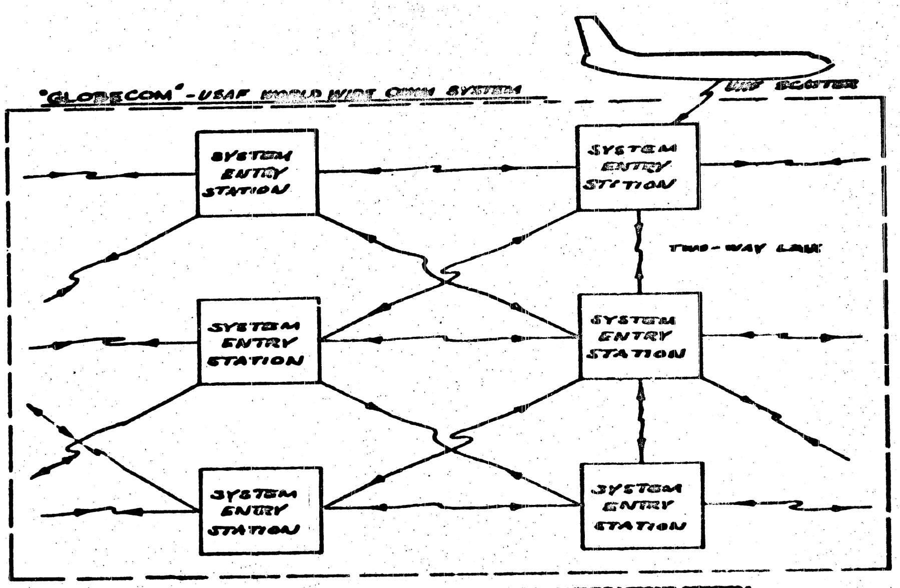
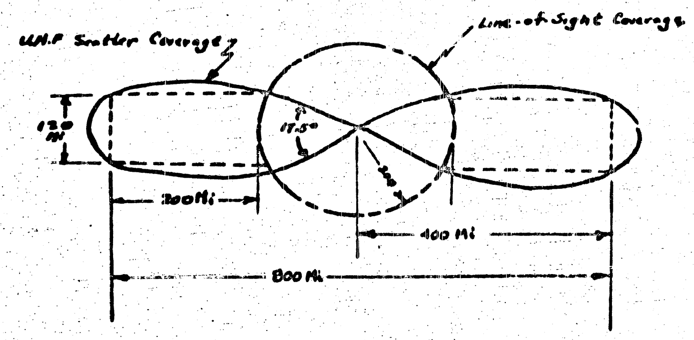


FIGURE 46 - CONCEPT OF IMPROVED USAF COMMUNICATIONS SYSTEM (SYSTEM 486-L)



Combined Line-of-Sight & Scatter Coverage for A Ground Station

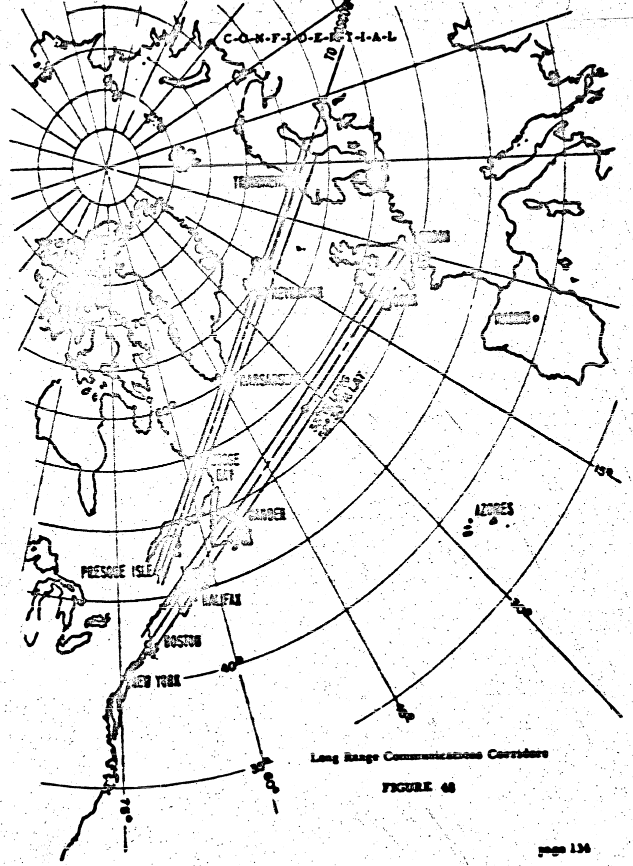
FIGURE 47

1. TYPICAL SAC MISSION

Should hostilities ever break out, a typical SAC mission might be staged out of Presque Isle, Maine to strike in the area of Moscow. In Figure 48, is shown such a flight path. If UHF scatter ground to air stations were installed near Presque Isle, Maine; Goose Bay, Laborador; Narsarssuak, Greenland; Reykjavik, Iceland and Trodheim, Norway communication with 90% or better reliability would be provided for the entire mission. These bases are separated by less than 800 miles so continuous coverage is provided. The bases selected for this example are already in use by the USAF or in friendly countries. During this entire mission, a SAC aircraft commander could be in direct contact with his home base by using air-ground UHF command-scatter communications plus the USAF point-to-point ground net. Of particularly importance is the fact that the aircraft uses only the UHF band during the entire flight. The particular military advantages of using this frequency sions are described later.

2. TYPICAL MATS MISSION

In MATS operation, an important run is the New York to London mission. For this corridor, the UHF ground stations could be at Boston, Massachusetts, Gander, Newfoundland; a picket ship station near 33° west longitude and 52°33' north latitude and Cork, Ireland. Here again none of the ground sites is more than 800 miles apart. There is another interesting by-product in the use of this concept. Since these communication links operate at UHF, it is entirely



Long Range Communications Corridors

FIGURE 48

possible to transmit from the ground precise navigation data of the VORTAC (rho-theta) type, and so provide fixing data also. This facility would increase the allowable traffic density for transatlantic greatly.

In summary, the implementation of air-ground UHF tropospheric scatter for extension of range will increase the operational communication capability by:

1) Providing at least 90% reliability of communications for important long distance flight paths.

2) Requiring the use of UHF command equipment only during an entire mission.

3) Providing effective long range communications to aircraft with the very limited possibility of detection of the scattered energy by unfriendly listening posts.

4) Providing precise navigation data during the entire mission, if desired.

5) Providing a long range communication capability which is quite impervious to enemy jamming.

The anticipated improvements which will accrue by the use of this concept are sufficiently great that every effort should be made to justify the theoretical data with experimental results. 10 kilowatt UHF transmitters for ground operation are already well within the state of the art. The ground directional antennas can be readily designed and built. The

airborne antenna will require some but modest aircraft modifications for its installation. Suggested airborne and ground based designs for antennas, transmitters and receivers are given in <u>Section III-C.</u>

B. AEW APPLICATION

Studies have been made previously of application of tropospheric scatter to AEW aircraft communication. It was thought desirable to make dual usage, radar and communication, of the equipment existing in such aircraft for the radar function. During the radar antenna slow scan, perhaps for a sector of about 30° or less, the radar would switch to the low information rate pulse coding at about the same duty cycle as for radar, the sector being generally centered on the direction of the ground terminal.

Preliminary calculations indicate high reliability for 600 miles range under such assumptions. Studies of modulation techniques are continuing.

V. SUMMARY AND RECOMMENDATIONS

V. SUMMARY AND RECOMMENDATIONS

Studies were conducted of propagation phenomena and applications to USAF air ground communications, using as a basis the report of Ames. Martin and Rogers*. The results therein have been extended under the study program.

The studies indicate the superiority of UHF tropospheric scatter for communications reliably in the Arctic for ranges of up to 400 to 500 miles at high altitudes and thus far beyond horizon.

A new set of scatter loss curves and distributions of hourly median values (slow fading) were derived. The curves are extensions to ranges of 500 miles beyond horizon, covering the total distance ranges contemplated. A wealth of data was used as a basis.

Methods were employed to obtain the radio line-of-sight (d_o) vs aircraft altitude for conditions typical of the Arctic. Midlatitudes and Tropics, and were based upon Weather Bureau meteorological data.

A method was used for extending the Midlatitude curves of scatter loss to other climatic regions of the Arctic and Tropics. The losses are greater in Arctic and less in the Tropics. Some experimental confirmation was found for the Arctic but considerably more data are needed.

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* L. A. Ames, E. J. Martin, and T.F. Rogers, "The possibility of extending Air-Ground UHF Voice Communications to Distances Far Beyond Radio Horizon," CONFIDENTIAL, AFCRC Technical Report TR-56-111, June 1956.

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The scatter loss in db below free space is considered to be due to a term L_s (db) appropriate for broad beam antennas and the aperture-to-medium coupling loss L_c (db) for narrow beams. Recent revisions in estimates of L_c were used for dissimilar antennas, anisotropic turbulent scattering ($r = 1/3$) and a scattering parameter varying inversely with height squared. This estimate of L_c was employed in analyzing experimental data to obtain our estimates of L_s vs d_s.

A reliability or total fading range factor F_T is employed when specifying performance reliability expected. It is larger than the slow fading factor F_s and is smaller than the sum (in db) of F_s and fast fading factor F_f, usually discussed. F_T is taken as the root-sum-square of F_s and F_f, for multi-stage diversity. Thus 99% reliability means 99% slow fading and 99% fast fading.

Using the effective distance concept the height gains were derived for high altitude aircraft. Predictions are given for the Arctic. The method was used to predict values to be compared with recent AFCRC data at 40,000 feet. Agreement within an error of 3 db was obtained between predicted and measured values (Appendix C).

A set of system parameters is proposed for a system design, using SSB. Dual spaced diversity, ground and airborne, is to be employed with vertical polarization. Airborne antennas would be "pod-mounted" under and out on the

wings, switching for forward or rearward directivity (50° x 50° beams, 12 db gain). Several ground antenna designs are suggested (2 1/2° vertical by 17.5° horizontal, 30 db gains). Dual diversity would be used here with 150 wavelength spacing. Pairs of such diversity antennas would be used, one pair for a forward direction and the other for the rearward direction to permit required coverage. Supplemental broader beam, (30°) short range, non-diversity antennas (paraboloids) would be used. Ground transmitter power is 10 kw and aircraft transmitter power is 1 kw.

Predicted performance of such a system is in excess of 400 miles at high altitudes for high reliability in the Arctic.

The operational system studies indicate "corridors" or "elongated cells" 800 miles by 120-150 miles may be used in a network to solve air-ground communications with high speed SAC and MATS aircraft. Typical examples of missions were laid out, using some existing ground terminals for relay.

A large bibliography was assembled of material used in the studies and is attached as Appendix D.

In order to check our predictions, particularly in the Arctic, it is recommended:

1) more data be obtained, ground-to-ground and particularly air-to-ground at high altitudes. The data pertain to signal loss and fading characteristics, and are lacking particularly in the Arctic.

2) detailed studies of the short-term fast fading factor should be made at higher altitudes at various distances and with different airspeeds to assess effects of diversity, doppler, and modulation.

3) further studies be made of rough terrain effects in the Arctic, particularly of positive horizon angles and "obstacle gain" phenomena.

4) before finally determining detailed antenna characteristics, model studies, perhaps 10 to 1 modelling scale, should be conducted for the antenna designs suggested for both airborne and ground arrays suggested.

5) an experimental operational setup should be implemented immediately. It is suggested that our 800 by 150 mile range coverage contour be planned, using the suggested antenna arrangement. One possible location for the forward and rearward coverage ground station would be at Plum Island, with flight paths from New York-New Jersey area to the Newfoundland direction. Communications potential and reliability plus propagational data would be obtained.

6) further studies be conducted on the feasibility of "simultaneous" radar and communications for AEW aircraft using UHF tropospheric scatter and essentially the AEW radar equipment with modest modification.

VI. REFERENCES

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VII. ACKNOWLEDGEMENTS

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VIII. APPENDICES

APPENDIX A

RADIO LINE OF SIGHT COMPUTATIONS FROM REFRACTIVE INDEX PROFILES

1. INTRODUCTION

It is often required to compute the radio line of sight between elevated antennas given a knowledge of the refractive index profile, i.e. the variation of radio refractive index with height. If the profile is analytic, then often it is possible to calculate the line-of-sight directly. More often, the profiles are obtained from Rawin Sonde balloon or other measurements. In such case a piece-wise summation can be used as outlined below.

2. SNELL'S LAW OF REFRACTION - SPHERICAL ATMOSPHERE

Consider a spherical earth and a spherically uniform atmosphere. From geometrical optics, it can be shown that[1] for a ray leaving the earth:

$$\frac{d}{ds}\left(n\frac{d\rho}{ds}\right) = \text{grad } n \tag{1}$$

where n is the refractive index, ds is the scalar arc length along the ray and $d\rho$ is the differential of the position vector for points on the ray. A point on the ray has a position h and $\text{grad } n = dn/dh$, a vertical vector. Let β_o be the angle of departure (or arrival) measured above the horizontal at the earth's surface where the refractive index is n_o. Let β_1 be a similar angle at a point h_1 above the earth where the index is n_1 (the angle is measured above the normal to the radius vector from the center of the earth, at $h = h_1$). Then it can be shown that (1) becomes

$$n_o \cos \beta_o = n_1 (1 + \frac{h_1}{a}) \cos \beta_1 = n_2 (1 + \frac{h_2}{a}) \cos \beta_2 \text{-----}$$

$$= \text{constant}$$

$$= n (1 + \frac{h}{a}) \cos \beta \qquad (2)$$

where β and n are appropriate to height h. Equation (2) is Snell's Law for a spherical earth and atmosphere, where a is the earth radius[2].

Denoting by N the modified refractive index at height h for a spherical earth, i.e. $N = n(1 + \frac{h}{a}) = n + h/a$, and by M the excess modified refractive index such that $M = N - 1$, we may obtain several useful relations.

In most of the work the angles β, β_o, β_1, etc. are sufficiently small that

$$\cos \beta \cong 1 - \beta^2/2$$

$$\cos \beta_o \cong 1 - \beta_o^2/2$$

etc.

Hence (2) becomes

$$N (1 - \frac{\beta^2}{2}) = N_o (1 - \frac{\beta_o^2}{2})$$

$$\beta^2 - \beta_o^2 = 2 (N - N_o) = 2 \Delta N \qquad (3)$$

whence

$$\beta = \sqrt{\beta_o^2 + 2 \Delta N} \qquad (4)$$

where

$$\Delta N = N - N_o = (N - 1) - (N_o - 1)$$

$$= \Delta M$$

so that (4) becomes

$$\beta = \sqrt{\beta_o^2 + 2 \Delta M} \qquad (5)$$

3. LINE-OF-SIGHT DISTANCE

Let us choose a coordinate system such that x is the great circle distance measured along the surface of the earth corresponding to a point on the ray at (x, h) where h is the height of the point measured radially from the earth's surface.

Then

$$dx = \frac{dh}{\tan \beta \, (1 + \frac{h}{a})} \qquad (6)$$

Let $\Delta x = x_2 - x_1$ be the increment in x corresponding to two points at (x_2, h_2) and (x_1, h_1). Then

$$\Delta x = \int_{x_1}^{x_2} dx = \int_{h_1}^{h_2} \frac{dh}{(1 - \frac{h}{a}) \tan \beta} \qquad (7)$$

or approximately

$$\Delta x = \int_{h_1}^{h_2} \cot \beta \, dh = \overline{(\cot \beta)}_{1,2} \, \Delta h \qquad (8)$$

where $\overline{(\cot \beta)}_{1,2}$ is the average value of $\cot \beta$ over the height interval $\Delta h = h_2 - h_1$.

For the case under consideration Schulkin[3] shows (see also reference 4)

$$\overline{(\cot \beta)}_{1,2} \cong (\beta_m)^{-1} = (\frac{\beta_1 + \beta_2}{2})^{-1} \qquad (9)$$

where

$$\beta_1 = \sqrt{\beta_o^2 + 2(M_1 - M_o)}$$

$$\beta_2 = \sqrt{\beta_o^2 + 2(M_2 - M_o)} \qquad (10)$$

Hence (8) becomes

$$x_2 - x_1 = \frac{h_2 - h_1}{\beta_m} \qquad (11)$$

If the profile is analytic and the resulting equation (7) can be integrated, then the line of sight is determined from (7) with β_o set equal to zero to give the tangent ray.

Otherwise the line-of-sight is determined from summations of incremental Δx values from (11). If $\beta_o = 0$, then

$$\beta_1^2 = 2(M_1 - M_o)$$

$$\beta_2^2 = 2(M_2 - M_o)$$

etc.

$$\beta_i^2 = 2(M_i - M_o) \qquad (12)$$

and

$$\beta_m = \frac{\beta_i + \beta_{i-1}}{2} \qquad (13)$$

and

$$\Delta d_i = \frac{h_i - h_{i-1}}{\beta_m} = \frac{\Delta h}{\beta_m} \qquad (14)$$

whence the line-of-sight d is given by

$$d = \sum_i \Delta d_i \qquad (15)$$

An example of the calculations is given below for the U.S. Standard Atmosphere in Reed and Russell[5].

h (ft)	0	2000	4000
$(n - 1) \times 10^6$	324.3	298.8	275.7
$\frac{h}{a} \times 10^6$	0	95.7	191.4
$M \times 10^6$	324.3	394.5	467.1
$\frac{\beta_i^2}{2} = M - M_o \; (10^6)$	0	70.2	142.8
$\beta_i^2 \times 10^6$	0	140.4	285.6
β_i (mr.)	0	11.85	16.90
β_m (mr.)	-	5.925	14.375
Δh ft.	-	2000	2000
Δd mi.	-	63.9	26.4
d mi.	0	63.9	90.3

4. OTHER PROFILES AND LINE-OF-SIGHT

Estimates of the line-of-sight for various bilinear profiles including the 4/3 earth radius case. These were:

Zone	$(n_o - 1) \times 10^6$	$\frac{dn}{dh}\ 10^9\ ft^{-1}$
Humid Tropics	400	-20.32
Temperate	365	-13.91
Dry Polar	300	- 6.10

The results were useful in a preliminary fashion for propagation studies for high altitude air-to-ground transmission. However, data from actual Weather Bureau soundings were actually employed which were published by Ratner[6]. The refractive index was computed from

$$(n - 1) \times 10^6 = 79 \frac{p}{T} (1 + 4800e/pT) \tag{16}$$

where

n = refractive index

T = temperature in °K

p = total pressure in millibars (p_e + e)

p_e = partial pressure of dry air

e = water vapor pressure

Profiles were then drawn for the Arctic (Fairbanks, July), Midlatitudes (Washington, October) and Tropics (San Juan, July). The results are shown in Figure 1 of the body of this Report.

Another model profile used was the reference profile of Fannin and Jehn[7] given by

$$(n - 1) \times 10^6 = 542.9 + 4.854\, h_1 - 100.86 \sqrt{h_1 + 3.919} \tag{17}$$

where h_1 is the height in thousands of feet. When used in equation (7) the result can be integrated and the line of sight calculated directly. The calculation was repeated using segments and equation (15), the results agreed within 1 to 2 miles. The resulting curve of d_o vs h_A for h_G = 100 ft agreed very closely with our curve for Midlatitudes shown in Figure 1 of the body of this Report.

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APPENDIX B

TRANSMISSION LOSS CALCULATIONS AND DATA ANALYSIS

1. TRANSMISSION LOSS EQUATIONS

In analyzing available data in order to come up with predicted transmission and scatter loss curves, one finds the bulk of the data is given in terms of distributions of hourly median values of received power and results from use of single antennas, usually identical. In such case, equation (3) of the body of the Report may be rewritten as

$$L = L_{f_o} + L_s + L_c + L_A - G_R - G_T + F_s \qquad (1)$$

the fading factor F_T being set equal to F_s and the diversity gain being set equal to zero.

The antenna gains G_R and G_T were mostly measured values. The value of L is the ratio of known transmitter power P_T and the known, median received power P_R. Values of slow fading factor were obtained from distributions of P_R for the most part using totalizers. Occasionally values of L_A are given or lumped in with measured G_T and G_R.

The value of L_c was computed using the methods outlined in the body of the Report. Accordingly, values of L_s were obtained from measured data (plus calculated L_{f_o} and L_c) according to

$$L_s(\text{db}) = P_T\ (\text{dbw}) - P_R\ (\text{dbw}) - L_{f_o}\ (\text{db}) - L_c\ (\text{db})$$
$$- L_A\ (\text{db}) + G_R\ (\text{db}) + G_T\ (\text{db}) - F_s\ (\text{db}) \qquad (2)$$

and the result gives the distribution of L_s according to F_s.

Where possible, the terrain profiles were taken into account in order to obtain d_s from $d_s = d - d_R - d_T$. This was true in the Lincoln Laboratory data[1, 2, 3, 4]. This was also true in Syracuse University data[5] on the Lexington-Syracuse 915 mc link studies for AFCRC as well as the Arctic 420 mc link data furnished by E J Martin. The profile information for BTL's data in Newfoundland[6] was not available in detail.

The Lincoln Laboratory data were analyzed for summer 50% and 90% values, complete winter distributions and yearly for the various receivers at Riverhead (winter only), Crawfords Hill, Alpha, and Winston-Salem. Data at Dewy Rose (October 1956, 800 miles, 412 mc) were also employed

The results led to the curves of Figure 2 in the body of the Report.

2. REFERENCES

1. J. H. Chisholm, P. A. Portmann, J. T. deBettencourt, and J. F. Roche, Proceedings IRE, Volume 43, pages 1317-1335, October 1955.
2. Same authors as (1), Lincoln Laboratory Technical Report 84, 26 September 1955.
3. See Reference 3 in body of Report.
4. See Reference 4 in body of Report.
5. J. H. Dienst, F Farner, R. Wyrick, S Goldman, Syracuse University Interim Technical Report, November 1956.
6. K. Bullington, W. J. Inkster, and A. L. Durkee, Proceedings IRE, Volume 43, pages 1306-1316, October 1955.

APPENDIX C

DATA AND ANALYSIS OF AFCRC FLIGHT MEASUREMENTS (MARCH 8-APRIL 1, 1957)

1. INTRODUCTION

Additional flights were made by AFCRC Communication Laboratory in March 1957 at altitudes of 38,000 to 40,000 feet. The flight and balloon sounding profile data were made available.

The flight data were in the form of graphs and tabulations of 10 and 20 mile median values of transmission loss vs distance. The balloon data were in the form of graphs of refractive index and tables of meteorological elements and computed refractive index vs height at Portland and Sable Island.

The refractive index profile data served to obtain line-of-sight d_o for a ground antenna at 80 ft, following the methods of Appendix A. They also served to obtain the appropriate values of k for use in computing the diffracted field component, outlined below.

In the scatter region, for the month of March 1957, we used our values of scatter loss L_s (hourly median) exceeded 90% of the year to typify the month values, from Figure 2 of the body of the Report. With appropriate allowance for other losses, with $d_s = d - d_o$ our predicted value of the loss at 38,000-40,000 feet was obtained and compared with the data.

2. REFRACTIVE INDEX DATA AND PROFILES

Meteorological data for each day of flights (except March 25) were furnished, from balloon flights at Portland and Sable Island. A sample calculation of modified index $N = n - 1$ is given in Table C-I for March 8, 1957.

The values of N were plotted and an average of the two curves was computed. A sample calculation of average N profile to typify the day's flight is given in Table C-II, for the same date.

The line-of-sight, for a ground antenna height of 80 feet ($d_T = 12.7$ mi) vs aircraft altitude h_R was then computed using the method outlined in Appendix A. A sample calculation is given in Table C-III for March 8, 1957.

The computed values of line-of-sight d_o for an 80-foot transmitter height are shown plotted in Figures C-1 through C-6, for March 8, 18, 22, 27, 29 and April 1, respectively.

3. CALCULATION OF DIFFRACTED FIELDS

The calculation of field strength and transmission loss just beyond horizon follows the methods outlined by Norton[1] and given in such texts as Terman[2]. However, some of the values required extensions of available tables or curves, and these were computed from the initial theory outlined by Norton. Also it was necessary to take into account a non linear value of

SAMPLE CALCULATIONS FOR MARCH 8, 1957

To Determine Modified Index (N) of Refraction From Meteorological Data Obtained at Portland and Sable Island on March 8, 1957, 1000 Est.

$$N(10^6) = (n - 1)\,10^6 = 79\,\frac{P}{T} + 379200\,\frac{e}{T^2}$$

where $N = (n - 1)$ = modified refractive index

n = refractive index
P = total pressure in millibars (mb) and is equal to $P_d + e$
P_d = partial pressure of dry air in millibars (mb)
e = water vapor pressure in millibars (mb)
T = Temperature in degrees Kelvin (°K)

At Portland:

H (ft)	$P_{(mb)}$	T(°K)	e(mb)	$N \times 10^6$
Sea Level	1014	276	5.16	314
430	1000	274	4.73	311
1,270	970	274	4.40	301
1,780	952	273	1.61	283
2,270	934	271	1.55	280
3,110	904	270	3.28	281
3,990	876	271	4.33	276
4,610	850	270	4.16	269
8,290	744	265	3.14	238
9,390	724	267	3.60	232
9,670	700	265	2.90	223
18,000	500	251	0.41	160

At Sable Island:

H (ft)	$P_{(mb)}$	T (°K)	e (mb)	$N \times 10^6$
Sea Level	1021	275	6.44	324
560	1000	274	6.13	317
1,430	969	273	5.89	309
2,140	944	276	7.47	305
3,750	888	276	6.87	286
4,870	850	276	2.05	253
9,910	700	265	—	—
12,260	644	262	0.84	199
16,190	551	255	1.12	177
16,550	543	256	1.13	174
18,290	500	253	0.44	159
20,770	458	247	0.27	148
23,560	400	239	0.21	134

TABLE C-I

TABLE C-II

Average Modified Index (N) of Refraction For Portland and Sable Island on March 8, 1957, 1000 Est.

Altitude (Thousands of Feet)	Index (N) at Portland	Index (N) at Sable Island	Average Index (N)
Sea Level	314	324	319
.5	310	318	314
1.0	304	312	308
1.5	292	309	300
2.0	282	306	294
2.5	281	301	291
3.0	280	294	287
3.5	278	289	284
4.0	276	278	277
5.0	266	252	259
6.0	257	245	251
7.0	248	238	243
8.0	240	230	235
9.0	233	222	228
10.0	220	214	217
12.0	205	200	202
14.0	190	188	189
16.0	175	177	176

Altitude (Thousands of Feet)	Index (N) at Portland	Index (N) at Sable Island	Average Index (N)
18.0	160	160	160
20.0	151	151	151
22.0	142	142	142
24.0	*	*	130
26.0	*	*	119
28.0	*	*	110
30.0	*	*	100
32.0	*	*	93
34.0	*	*	86
36.0	*	*	79
38.0	*	*	73
40.0	*	*	67

* Average Modified Index (N) of Refraction for These Altitudes Were Determined by Extrapolation on a Lin-Log Basis.

TABLE C-III

To Determine The Line-of-Sight Curve For March 8

From Calculated Values of Index of Refraction For That Day

Transmitter Height (h_t) at 80 Ft.

$$d(mi) = d_x(mi) + d_t(mi)$$

$$d_t(mi) \simeq \sqrt{2h_t(ft)} = 12.65\,(mi)$$

$$d_x(mi) = \sum_{o}^{x} \Delta d_x(mi)$$

$$\Delta d_x(ft) \cong \frac{\Delta h\ (ft)}{B_m\ (rad)}$$

$$\Delta h(ft) = h_2\ (ft) - h_1\ (ft)$$

$$M_x = (N_x + \frac{h_x\ (ft)}{a\ (ft)})$$

$$N_x = n_x - 1$$

$$a = \text{radius of earth} = 20.9 \times 10^6\ (ft)$$

$$B_x = \sqrt{2(M_x - M_o)}\ (rad)$$

$$B_m = \frac{B_{x1} + B_{x2}}{2}\ (rad)$$

$$M_o = N_o = 318 \times 10^{-6}$$

TABLE C-III (cont'd)

h (ft)	$N_x(10^{-6})$	$M_x(10^{-6})$	$B_x(10^{-3})$	$B_m(10^{-3})$	Δd_x(mi)	d_x(mi)	d(mi)
0	319	319	0			0	0
				3.1	30.52		
500	314	338	6.3			30.52	43.17
				7.5	12.62		
1000	308	356	8.7			43.14	55.79
				9.5	9.95		
1500	300	372	10.4			53.09	65.74
				11.2	8.45		
2000	294	390	12.0			61.54	74.19
				12.8	7.40		
2500	290	410	13.6			68.94	81.59
				14.3	6.62		
3000	287	430	15.0			75.56	88.21
				15.6	6.17		
3500	282	449	16.2			81.73	94.38
				16.7	5.67		
4000	276	467	17.3			87.40	100.05
				17.7	5.35		
4500	267	482	18.1			92.75	105.40
				18.5	5.11		
5000	259	498	19.0			97.86	110.51
				19.5	4.85		
5500	254	517	20.0			102.71	115.36
				20.5	4.62		
6000	251	538	21.0			107.33	119.98
				21.5	4.40		
6500	247	558	21.9			111.73	124.38
				22.4	4.22		
7000	243	578	22.8			115.95	128.60
				23.3	4.06		
7500	239	597	23.7			120.01	132.66
				24.1	3.93		
8000	235	617	24.5			123.94	136.59
				24.9	3.80		
8500	231	637	25.3			127.74	140.39
				25.7	3.68		

TABLE C-III (cont'd)

h (ft)	$N_x(10^{-6})$	$M_x(10^{-6})$	$B_x(10^{-3})$	$B_m(10^{-3})$	Δd_x(mi)	d_x(mi)	d(mi)
9000	227	657	26.1			131.42	144.07
				26.4	3.59		
9500	223	677	26.8			135.01	147.66
				27.2	3.48		
10,000	217	695	27.5			138.49	151.14
				27.9	3.39		
10,500	213	715	28.2			141.88	154.53
				28.6	3.31		
11,000	209	735	28.9			145.19	157.84
				29.3	3.23		
11,500	206	756	29.6			148.42	161.07
				30.0	3.17		
12,000	202	776	30.3			151.59	164.24
				30.7	3.08		
12,500	199	797	31.0			154.67	167.32
				31.3	3.02		
13,000	196	817	31.6			157.69	170.34
				31.9	2.97		
13,500	193	838	32.3			160.66	173.31
				32.6	2.90		
14,000	189	858	32.9			163.56	176.21
				33.2	2.85		
14,500	186	879	33.5			166.41	179.06
				33.3	2.80		
15,000	183	900	34.2			169.21	181.86
				34.5	2.74		
15,500	180	921	34.7			171.95	184.60
				35.0	2.70		
16,000	176	941	35.3			174.65	187.30
				35.6	2.66		
16,500	173	962	35.9			177.31	189.96
				36.2	2.61		
17,000	169	982	36.4			179.92	192.57
				36.7	2.58		
17,500	165	1001	37.0			182.50	195.15
				37.3	2.54		

TABLE C-III (cont'd)

h (ft)	$N_x(10^{-6})$	$M_x(10^{-6})$	$B_x(10^{-3})$	$B_m(10^{-3})$	Δd_x(mi)	d_x(mi)	d (mi)
18,000	160	1020	37.5			185.04	197.69
				37.8	2.50		
18,500	157	1041	38.1			187.54	200.19
				38.4	2.47		
19,000	155	1063	38.6			190.01	202.66
				38.9	2.43		
19,500	153	1085	39.2			192.44	205.09
				39.5	2.40		
20,000	151	1107	39.7			194.84	207.49
				40.0	2.37		
20,500	149	1129	40.3			197.21	209.86
				40.6	2.33		
21,000	147	1151	40.9			199.54	212.19
				41.1	2.30		
21,500	145	1173	41.4			201.84	214.49
				41.6	2.27		
22,000	142	1194	41.9			204.11	216.76
				42.1	2.25		
22,500	139	1215	42.4			206.36	219.01
				42.6	2.22		
23,000	135	1234	42.8			208.58	221.23
				43.3	4.34		
24,000	130	1277	43.8			212.92	225.57
				44.7	8.47		
26,000	119	1362	45.7			221.39	234.04
				46.6	8.12		
28,000	110	1448	47.6			229.51	242.16
				48.4	7.82		
30,000	100	1534	49.3			237.33	249.98
				50.2	7.55		
32,000	93	1623	51.1			244.88	257.53
				51.9	7.29		
34,000	86	1711	52.8			252.17	264.82
				53.6	7.07		
36,000	79	1800	54.4			259.24	271.89
				55.3	6.85		
38,000	73	1889	56.2			266.09	278.74
				56.9	6.65		
40,000	67	1979	57.6			272.74	285.39

$d_o = d_t + d_r = 278.74$ mi $= 279$ mi

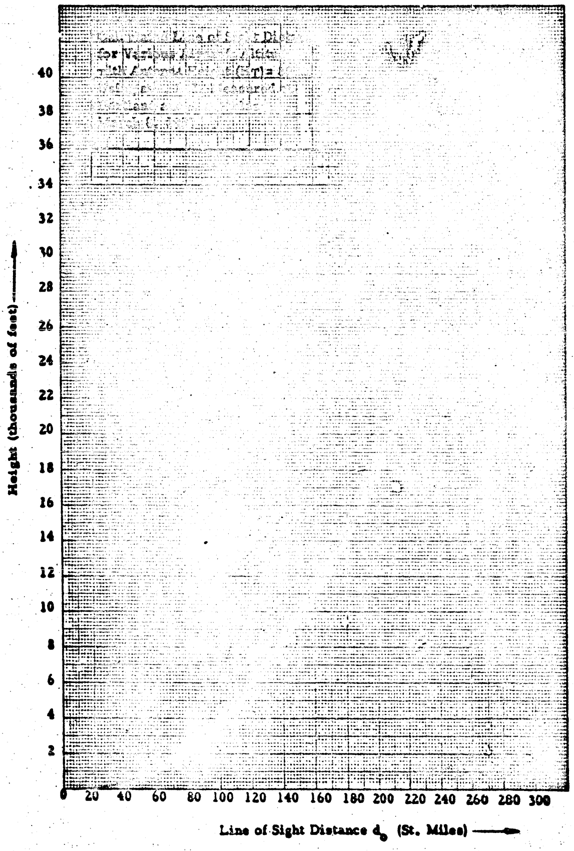


FIGURE C-1

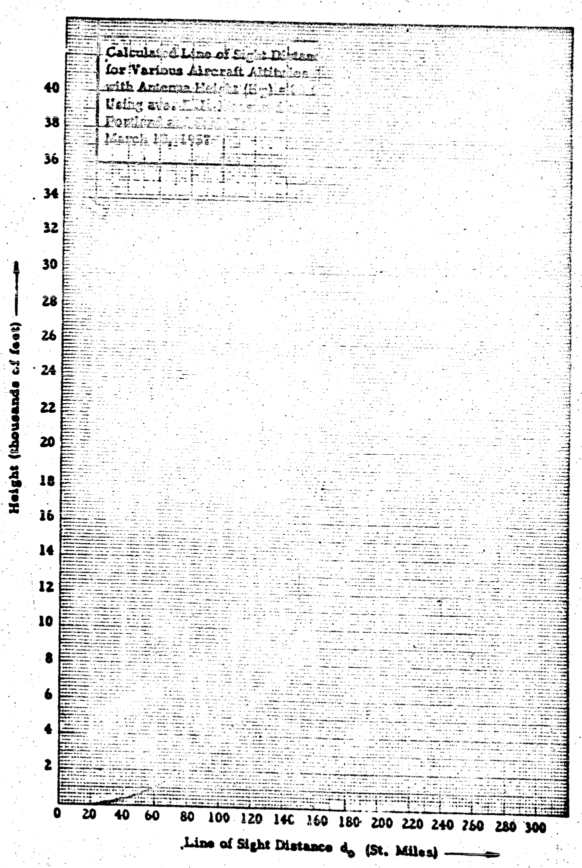


FIGURE C-2

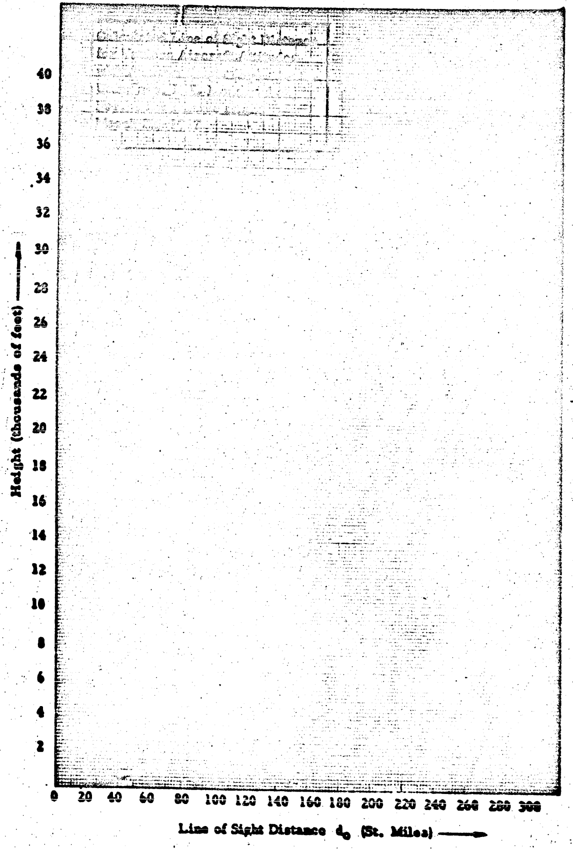


FIGURE C-3

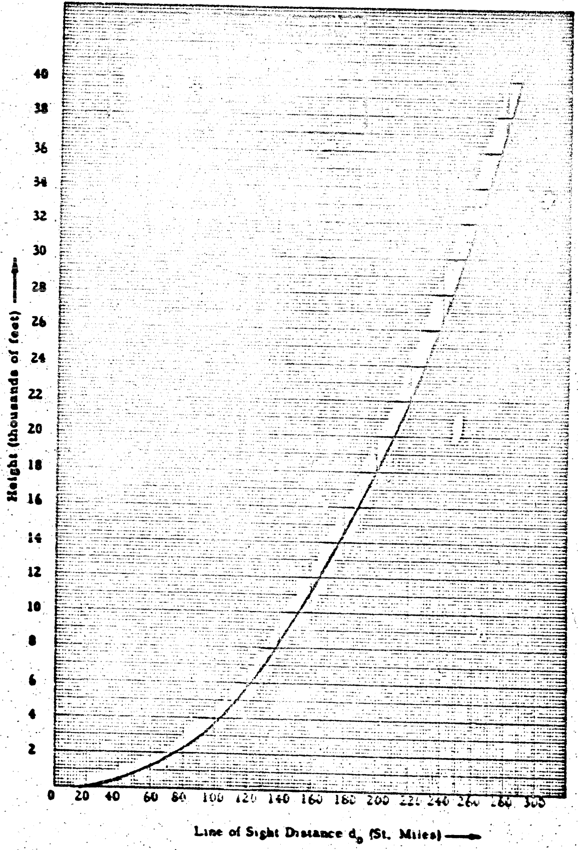


FIGURE C-4

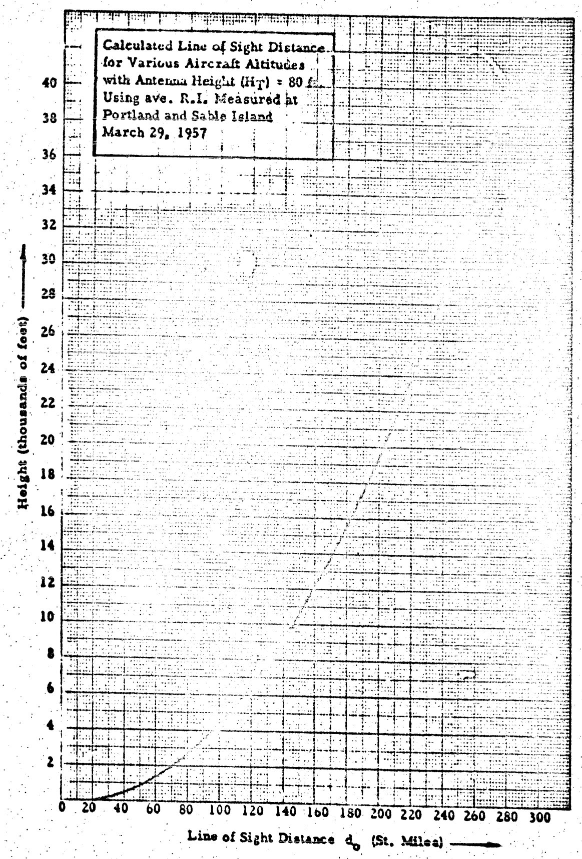


FIGURE C-5

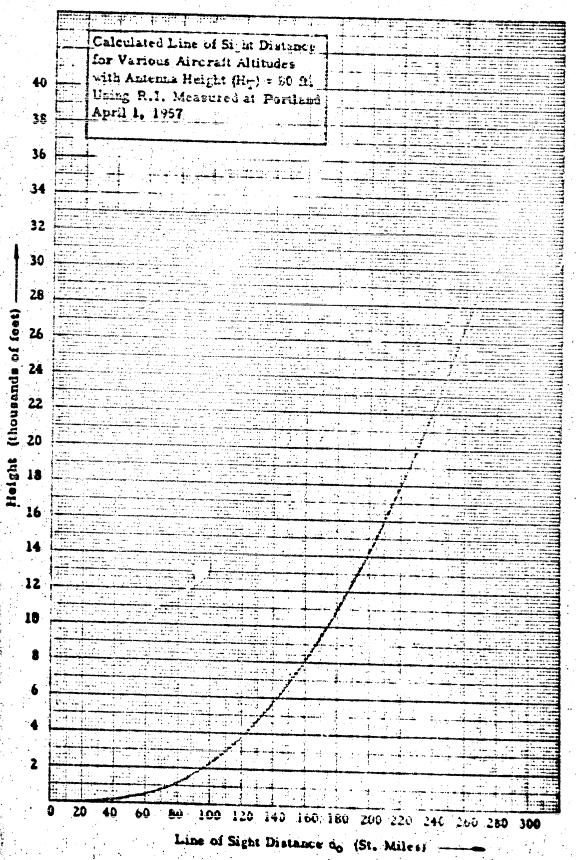


FIGURE C-6

index gradient. An approximation was used here. At low heights the appropriate value of k near the surface was used. For the height-gain factor for high heights an average value of k was used from the relation $d_R^2 = 1.5 k h_R$ where d_R was the value appropriate to h_R and obtained from the profile data, as given above.

To assist calculations, several additional tables were compiled and these are given in Tables C-IV through C-IX, for various quantities and variables, particularly k, used in the work.

For loss computations, a table of free space loss between isotropic antennas is given in Table C-X, for f = 220 mc.

Finally, a sample calculation is given in Table C-XI for computing the transmission loss between isotropic antennas for a ground height of 80 feet, an aircraft antenna altitude of 38,000 feet, f = 220 mc for March 8 using the index profile and line-of-sight data for that date.

The computed diffraction losses are shown in the curves of the next section.

4. TROPOSPHERIC SCATTER LOSSES AND HEIGHT GAIN G_H IN THE REGION FAR BEYOND RADIO HORIZON

The basic transmission losses were computed for the scatter mode using the methods of Appendix B. It was assumed that coupling loss L_c was small (actually the loss L_c was computed for the assumed ground and airborne antenna

TABLE C-IV

$$\left.\frac{E_{\eta'=2}}{E_o}\right]_{db} = -195.9892 + 20 \log k^{-4/3}$$

k	log k	4/3 log k	20 log $k^{-4/3}$	$E_{\eta'=2}/E_o$ db
1.0	0	0	0	-195.99
1.1	0.041393	.05520	-1.1040	-197.09
1.2	.079181	.10558	-2.1116	-198.10
1.25	.096910	.12922	-2.5844	-198.57
1.3	.113943	.15192	-3.0384	-199.03
1.333	.124612	.16614	-3.3228	-199.31
1.4	.146128	.19484	-3.8968	-199.89
1.5	.176091	.23478	-4.6956	-200.69
1.6	.204120	.27216	-5.4432	-201.43

TABLE C-V

$M(\eta')$ db $= 27.7 - 16.31\eta'$ $(\eta' \geq 9)$

η'	$M(\eta')$	$M(\eta')$ db
9	1.18×10^{-6}	-118.56
10	1.70×10^{-7}	-135.40
11	2.60×10^{-8}	-151.68
12	4.20×10^{-9}	-167.52
13	5.70×10^{-10}	-184.88
14	9.40×10^{-11}	-200.52
15	1.35×10^{-11}	-217.40
16	2.20×10^{-12}	-233.16
17	3.20×10^{-13}	-249.88
18	4.60×10^{-14}	-266.72
19	8.10×10^{-15}	-281.84
20	1.20×10^{-15}	-298.42

TABLE C-VI

$$E_{o,o}/E_o \text{ (db)} = E_{\eta'=2}/E_o \text{ (db)} + M(\eta') \text{ (db)}$$

k		1.0	1.1	1.2	1.3	4/3	1.4	1.5	1.6
$\frac{E_{\eta'=2}}{E_o}$ (db)		-196.0	-197.1	-198.1	-199.0	-199.3	-199.9	-200.7	-201.4
η'	$M(\eta')_{db}$	Values of $E_{o,o}/E_o$ (db)							
9	-118.6	-314.6	-315.7	-316.7	-317.6	-317.9	-318.5	-319.3	-320.0
10	-135.4	-331.4	-332.5	-333.5	-334.4	-334.7	-335.3	-336.1	-326.8
11	-151.7	-347.7	-348.8	-349.8	-350.7	-351.0	-351.6	-352.4	-353.1
12	-167.5	-363.5	-364.6	-365.6	-366.5	-366.8	-367.4	-368.2	-368.9
13	-184.9	-380.9	-382.0	-383.0	-383.9	-384.2	-384.8	-385.6	-386.3
14	-200.5	-396.5	-397.6	-398.6	-399.5	-399.8	-400.4	-401.2	-401.9
15	-217.4	-413.4	-414.5	-415.5	-416.4	-416.7	-417.3	-418.1	-418.8
16	-233.2	-429.2	-430.3	-431.3	-432.2	-432.5	-433.1	-433.9	-434.6
17	-249.9	-445.9	-447.0	-448.0	-448.9	-449.2	-449.8	-450.6	-451.3
18	-274.4	-470.4	-471.5	-472.5	-473.4	-473.7	-474.3	-475.1	-475.8
19	-281.8	-477.8	-478.9	-479.9	-480.8	-481.1	-481.7	-482.5	-483.2
20	-298.4	-494.4	-495.5	-496.5	-497.4	-497.7	-498.3	-499.1	-499.8

TABLE C-VII

$$\log C_1 = 1.17844 + 2/3 \log k$$

$$C_1 = \left(\frac{1}{6.6307 \times 10^{-2}}\right) k^{2/3}$$

k	log k	2/3 log k	log C_1	C_1
1.0	0	0	1.17844	15.081
1.1	.041393	.02760	1.20604	16.071
1.2	.079181	.05279	1.23123	17.031
1.25	.096910	.06461	1.24305	17.500
1.3	.113943	.07596	1.25440	17.964
1.333	.124612	.08307	1.26151	18.260
1.4	.146128	.09742	1.27586	18.874
1.5	.176091	.11739	1.29583	19.762
1.6	.204120	.13608	1.31452	20.631

$$C_2 = \frac{80}{3} k$$

$$C_3 = 1.08147 k^{-2/3}$$

$$C_4 = -168.3 - \frac{80}{3} k$$

$$E_{o,o}/E_o \text{ (db)} = C_4 - C_3 \text{ d(mi)}$$

k	1.0	1.1	1.2	1.25	1.3	1.333	1.4	1.5	1.6
C_2	0.00000	1.10384	2.11152	2.58424	3.03848	3.32296	3.89672	4.69576	5.44320
C_3	1.08147	1.01490	0.95770	0.93199	0.90795	0.89320	0.86417	0.82533	0.79057
C_4	-168.3	-169.4	-170.4	-170.9	-171.3	-171.6	-172.2	-173.0	-173.7
d (mi)	Values of $E_{o,o}/E_o$ (db)								
100	-276.5	-270.9	-266.2	-264.1	-262.1	-260.9	-258.6	-255.6	-252.8
120	-298.1	-291.2	-285.3	-282.7	-280.3	-278.8	-275.9	-272.0	-268.6
140	-319.7	-311.5	-304.5	-301.4	-298.4	-296.7	-293.2	-288.5	-284.4
160	-341.3	-331.8	-323.6	-320.0	-316.6	-314.6	-310.5	-305.1	-300.2
180	-362.9	-352.1	-342.8	-338.7	-334.7	-332.5	-327.8	-321.6	-316.0
200	-384.6	-372.4	-361.9	-357.3	-352.9	-350.2	-345.0	-338.1	-331.8
220	-405.2	-392.7	-381.1	-375.9	-371.0	-368.1	-362.3	-354.6	-347.6
240	-427.9	-413.0	-400.2	-394.6	-389.2	-386.0	-379.6	-371.1	-363.4
260	-449.5	-433.3	-419.4	-413.2	-407.4	-403.9	-396.9	-387.6	-379.2
280	-471.1	-453.6	-438.6	-431.9	-425.5	-421.8	-414.2	-404.1	-395.0
300	-492.8	-473.9	-457.7	-450.5	-443.7	-439.6	-431.5	-420.6	-410.8
320	-514.3	-494.2	-476.9	-469.1	-461.8	-457.6	-448.7	-437.1	-426.6
340	-535.9	-514.5	-496.0	-487.8	-480.0	-475.5	-466.0	-453.6	-442.4

TABLE C-VIII

$C_5 = 3.4552 \times 10^{-3} k^{-1/3}$

$H = C_5 \, h \text{ (ft)}$

Value of f (H) Obtained from Curve in Terman's Handbook
Page 689, Figure 18

k	$C_5 \times 10^3$	h = 38,000 ft. H	log f (H)	20 log f (H)	h = 80 ft. H	f (H)	log f (H)	20 log f (H)	δ (db)
1	3.4552	131.30	12.35	247.0	.2764	1.03	+.0125	+ .25	66.3
1.1	3.3471	127.19	12.15	243.0	.2678	.99	-.005	- .10	66.6
1.2	3.2515	123.56	11.97	239.4	.2601	.96	-.017	- .34	66.8
1.25	3.2075	121.88	11.89	237.8	.2566	.94	-.027	- .54	66.9
1.3	3.1658	120.30	11.81	236.2	.2533	.93	-.032	- .64	67.1
1.333	3.1400	119.32	11.75	235.0	.2512	.92	-.036	- .72	67.1
1.4	3.0887	117.37	11.65	233.0	.2471	.91	-.041	- .82	67.3
1.5	3.0184	114.70	11.51	230.2	.2415	.89	-.050	-1.00	67.5
1.6	2.9541	112.26	11.38	227.6	.2363	.87	-.060	-1.20	67.7

TABLE C-IX

TABLE C-X

Free Space Loss Vs Distance at 220 Mc.

$L_{f_0} = -83.43 - 20 \log d \text{ (mi)}$

$G_1 = G_2 = 1.00 \text{ (0 db)}$

d (mi)	L_{f_0} (db)
20	109.45
40	115.47
60	118.99
80	121.49
100	123.43
120	125.01
140	126.35
160	127.51
180	128.53
200	129.45
220	130.29
240	131.03
260	131.73
280	132.37
300	132.97
320	133.53
40	134.07
[illegible]	134.55
380	135.03
400	135.47

d (mi)	L_{f_0} (db)
420	135.89
440	136.31
460	136.69
480	137.05
500	137.41
520	137.75
540	138.07
560	138.39
580	138.69
600	138.99
620	139.27
640	139.55
660	139.83
680	140.09
700	140.33
720	140.57
740	140.81
760	141.05
780	141.27
800	141.49

TABLE C-XI

To Calculate Transmission Losses (Diffraction) Near Radio Horizon on March 8, 1957, 1000 Est. (f = 220 Mc.)

$$\bar{k}_1 = \frac{1}{1 + a\left(\frac{dn}{dh}\right)_{\substack{\text{near}\\\text{surface}}}} = \frac{1}{1 + 20.9\left(\frac{-71}{4870}\right)} \simeq 1.5$$

$$\overline{k_2} = \frac{{d_t}^2}{1.5 h_r} = \frac{(266.1)^2}{(1.5)\ (38,000)} \simeq 1.25$$

$\bar{k}_1 = 1.5$, $h_t = 80$ (ft), $f(H_1)$ db $= -1.0$ (db)

δ_1 db $= 67.5$ (db)

$\bar{k}_2 = 1.25$, $h_r = 38,000$ $f(H_2)$ db $= 237.8$ (db)

δ_2 db $= 66.9$ (db)

$$E_{1,2}/E_o = E_{o,o}/E_o\ (\bar{k}_1) + 371.2\ ;\ \Big[F(H_1) \cdot F(H_2)\Big]_{db} = 371.2\ \text{(db)}$$

d(mi)	$\frac{E_{o,o}}{E_o}(\bar{k}_1)$	$\frac{E_{1,2}}{E_o}$	L_{f_o} (db)	L (db)
240	-371.1	+0.1	-131.0	-130.9
260	-387.6	-16.4	-131.7	-148.1
280	-404.1	-32.9	-132.4	-165.3
300	-420.6	-49.4	-133.0	-182.4
320	-437.1	-65.9	-133.5	-199.4
340	-453.6	-82.4	-134.1	-216.5

beams vs d_s, the results showed L_c to be less than 1 db in the range of d_s (not d) encountered, $d_s < 350$ mi)

The data from the flights were given in terms of the 10 mile medians (median of several observations in 10-mile intervals) of basic transmission presumed computed from measured received power transmitter power antenna gains line losses and free space loss These points are indicated as circles in the graphs which follow

Table C-XII contains the tabulated transmission loss data of AFCRC for the flights

Table C-XIII contains a sample calculation of predicted transmission losses L_b and height gain G_H for h_R = 38,000 feet for March 8.

The calculated and measured basic transmission losses for each of the flights are shown in Figures C-7 through C-13 for March 8 18, 22, 25, 27, 29 and April 1, 1957 The calculated free space curve is shown, but not the interference pattern within horizon The calculated diffraction curve applies from roughly 30 to 35 miles beyond horizon The calculated scatter loss applies roughly for distances greater than 50 miles beyond horizon where it greatly exceeds the diffraction field.

As aforementioned, the scatter loss L_s curve we used for prediction was that exceeded 90% of the hours in a year as being appropriate to March. The value at the surface (h_R = 0) assumes that h_T is 80 feet of course.

TABLE C-XII
B-47 10 MILE MEDIANS - BASIC TRANSMISSION LOSS
AFCRC

Distance	8 Mar.	18 Mar.	22 Mar.	25 Mar.	27 Mar.	29 Mar.	1 Apr.
105	138.5		145	125.5		135.5	127
15	126		131	130	135	139.5	130
25	125	128	130	127.5	127.5	130	131
35	129	(133)	127	126.5	129	127.5	131
45	136	135	132.5	129	137.5	129.5	152
55	146	135	143.5	129	151.5	132.5	139.5
65	136	137.5	144	141	143	146	130.5
75	126	132.5	136	127.5	135.5	140	131
85	126	127	142.5	125.5	142.5	130.5	130
95	124	125.5	133	124	134	127	127.5
205	126	125	139.5	124.5	142	126	133.5
15	126	132.5	131	125.5	133	129	134.5
25	128	131	130.5	126.5	127	132	134.5
35	128.5	136	133.5	129.5	125.5	133.5	144.5
45	133	154	134	139		138	151
55	139	147	143.5	(146)		145.5	150.5
65	149	(159)	149	145	(141)	154	150
75	157.5	168	152.5	153.5	(155)	159	159
85	166	174.5	160.5	160	157	167.5	168
95	173	188.5	165.5	161	161.5	172	170
305	181.5	189	171	170	168	182.5	174.5
15	185		176	181	166.5	189	176
25	190	191	178.5	182	177	194.5	189.5
35	192	201.5	183	191.5	171.5	195.5	192
45	193.5	199	190	204.5	172	195	187
55	198	205	194	199	173	201	198
65	195.5	204	203.5	201.5	188.5	203	197.5
75	200	204	207	201	191	200.5	194
85	202.5	208	208.5	206	199.5	202.5	201
95	(199)	211.5	207.5	206	200	208	196.5

TABLE C-XII (cont'd)
B-47 10 MILE MEDIANS
AFCRC

Distance	8 Mar.	18 Mar.	22 Mar.	25 Mar.	27 Mar.	29 Mar.	1 Apr.
405	200.5	211	210	206	201	209	207.5
15	204	212	208	(203)	206.5	214	203.5
25	204	214.5	215.5	205	198.5	211	202
35	205.5	210.5	218	218	208	210	202
45	201	216	218.5	219	203	217	208.5
55	204	217	220.5	(210)	208.5	217.5	212.5
65	207	218	220	213.5	213	222	209.5
75	208	217	218.5	224.5	215	221	216.5
85	210	220.5		220	216	208	213
95	210.5	223.5	225	221.5	218.5	218	(220)
505	221	217.5	228	214	216		227
15	216	231.5	230.5	207.5	217		224.5
25	217	231	234.5	212	223		212
35	217		234		221.5		211
45	211		232		(214)		220
55	215.5		227.5		220		216
65	231		228		227.5		216.5
75	232				230		222
85	228.5				228.5		227.5
95	230				233		229.5
605					232.5		234.5
15					234		234.5
25							231
35							228.5
45							235.5
55							236
65							228
75							235
85							236.5

SAMPLE CALCULATIONS FOR MARCH 8, 1957

To Determine Basic Transmission Loss (L_b) and Height Gain (G_H) vs Distance (d)
(Measured vs Predicted Values)

d_s = scatter distance (mi)
d_r = receiver line-of-sight distance (mi)
d_t = transmitter line-of-sight distance (mi)
G_T = transmitter antenna gain (db)
G_R = receiver antenna gain (db)
F_s = slow fading factor (db)

L_{f_o} = free space loss (db) for isotropic antennas
L_c = antenna coupling loss (db)
L_A = transmission line loss (db)
$L_s(50)$ = hourly median scatter loss (db) for 50% of the hours
L_b = hourly median basic transmission loss (db)
L = total transmission loss (db)

(1) $L_b = L_{f_o} + L_s(50) + F_s + L_c$

(2) $L = L_b + L_A - G_T - G_R$

(3) $L_s(90) = L_s(50) + F_s(90)$

(4) $d_s = d - d_t - d_r$

(5) $d_t = \sqrt{2h_t} = 12.65$ mi (h_t = 80 ft)

(6) $d_r = 266$ mi (h_r = 38,000 ft) from line-of-sight curve

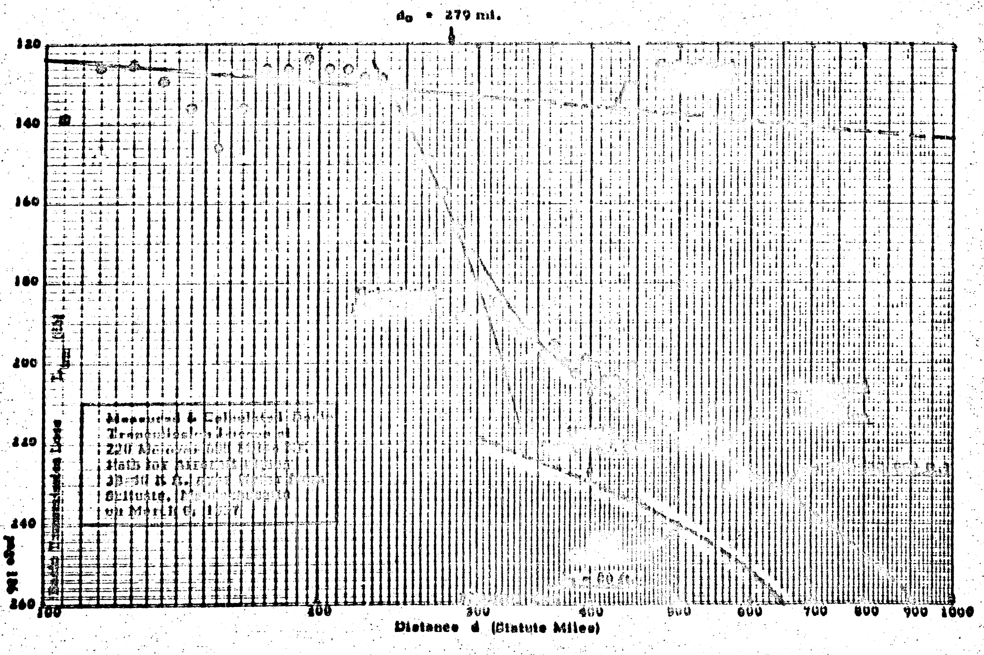
TABLE C-XIII

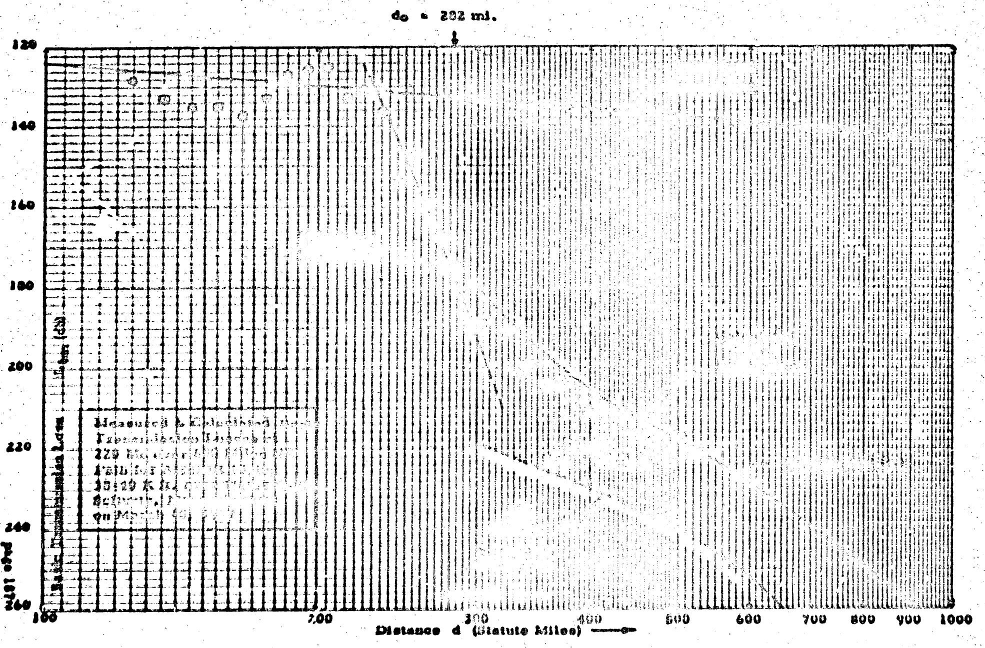
d (mi)	d_s(mi) (h_R = 38 K)	d_s (mi) (h_R = 0)	$\left[L_s + L_c + F_s(90)\right]$ db (h_R = 38 K)	$\left[L_s + L_c + F_s(90)\right]$ db (h_R = 0)	L_{f_o} (db)	(Calc.) L_b (db) (h_R = 38 K)	(Calc.) L_b (db) (h_R = 0)	(Meas.) L_b (db) (h_R = 38-40 K)	(Calc.) G_H (db) (h_R = 38 K)	(Meas.) G_H (db) (h_R = 38-40K)
305	26	292		86.7	133.1		219.8	181.5		
315	36	302		87.5	133.4		220.9	185.0		
325	46	312		88.3	133.7		222.0	190.0		
335	56	322	56.5	89.0	133.9	190.4	222.9	192.	32.5	30.9
345	66	332	59.3	89.9	134.2	193.5	224.1	193.5	30.6	30.6
355	76	342	61.7	90.8	134.4	196.1	225.2	198.0	29.1	27.2
365	86	352	64.0	91.6	134.7	198.7	226.3	195.5	27.6	30.8
375	96	362	66.0	92.4	134.9	200.9	227.3	200.	26.4	27.3
385	106	372	67.8	93.3	135.1	202.9	228.4	202.5	25.5	25.9
395	116	382	69.5	94.1	135.4	204.9	229.5	(199)	24.6	30.5
405	126	392	70.8	94.9	135.6	206.[illegible]	230.5	200.5	24.1	30.0
415	136	402	72.0	95.9	135.8	207.8	231.7	204.	23.9	27.7
425	146	412	73.3	96.7	136.0	209.3	232.7	204	23.4	28.7
435	156	422	74.5	97.6	136.2	210.7	233.8	205.5	23.1	28.3
445	166	432	75.5	98.6	136.4	211.9	235.0	201.	23.1	34.0
455	176	442	76.6	99.5	136.6	213.2	236.1	204.	22.9	32.1
465	186	452	77.4	100.3	136.8	214.2	237.1	207.	22.9	30.1

TABLE C-XIII (cont'd)

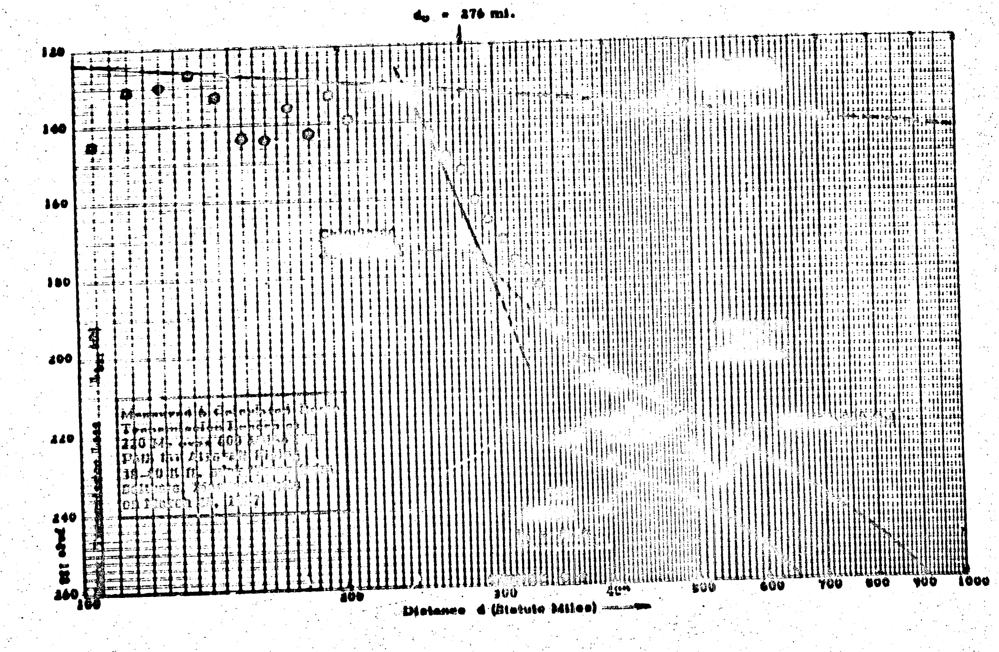
d (mi)	d_s(mi) (h_R = 38 K)	d_s(mi) (h_R = 0)	$L_s + L_c + F_s (90)\Big]$ db (h_R = 38 K)	$L_s + L_c + F_s (90)\Big]$ db (h_R = 0)	L_{f_o} (db)	(Calc.) L_b (db) (h_R = 38K)	(Calc.) L_b (db) (h_R = 0)	(Meas.) L_b (db) (h_R = 38-40K)	(Calc.) G_H (db) (h_R = 38 K)	(Meas.) G_H (db) (h_R = 38-40K)
475	196	462	78.3	101.2	137.0	215.3	238.2	208.	22.9	30.2
485	206	472	79.2	102.2	137.2	216.4	239.4	210.	23.0	29.4
495	216	482	80.2	103.1	137.3	217.5	240.4	210.5	22.9	29.9
505	226	492	81.1	103.9	137.5	218.6	241.4	221.	22.8	20.4
515	236	502	81.9	104.8	137.7	219.6	242.5	216	22.9	26.5
525	246	512	82.7	105.8	137.8	220.5	243.6	217.	23.1	26.6
535	256	522	83.5	106.8	138.0	221.5	244.8	217.	23.3	27.8
545	266	532	84.4	107.8	138.2	222.6	246.0	211.	23.4	35.0
555	276	547	85.3	108.7	138.3	223.6	247.0	215.5	23.4	31.5
565	286	552	86.2	109.7	138.5	224.7	248.2	231.	23.5	17.2
575	296	562	86.9	110.7	138.6	225.5	249.3	232.	23.8	17.3
585	306	572	87.8	111.7	138.8	226.6	250.5	228.5	23.9	22.0
595	316	582	88.8	112.7	138.9	227.4	251.6	230.	24.2	21.6

TABLE C-XIII (cont'd)

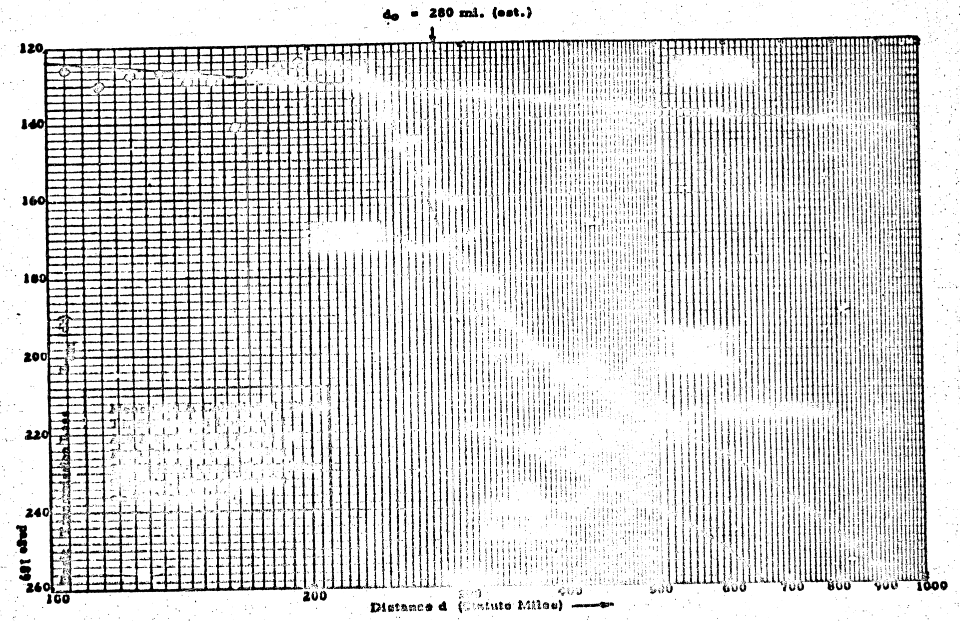


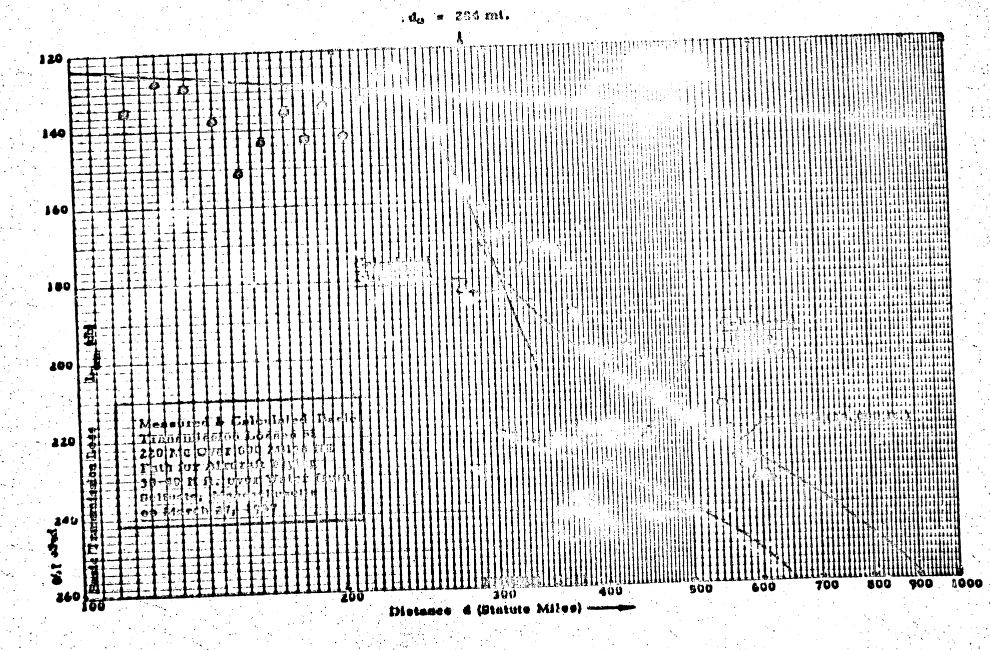


d_o = 282 mi.
120
140
160
180
200
220
240
260
Basic Transmission Loss — L_{bm} (db)
Measured & Calculated
Transmission Loss
220
100
200
300
400
500
600
700
800
900
1000
Distance d (Statute Miles)



d_o = 276 mi.
120
140
160
180
200
220
240
260
Transmission Loss — L_{bm} (db)
Measured & Calculated
Transmission Loss
220 Mc
600
Path
30-40
on
100
200
300
400
500
600
700
800
900
1000
Distance d (Statute Miles)





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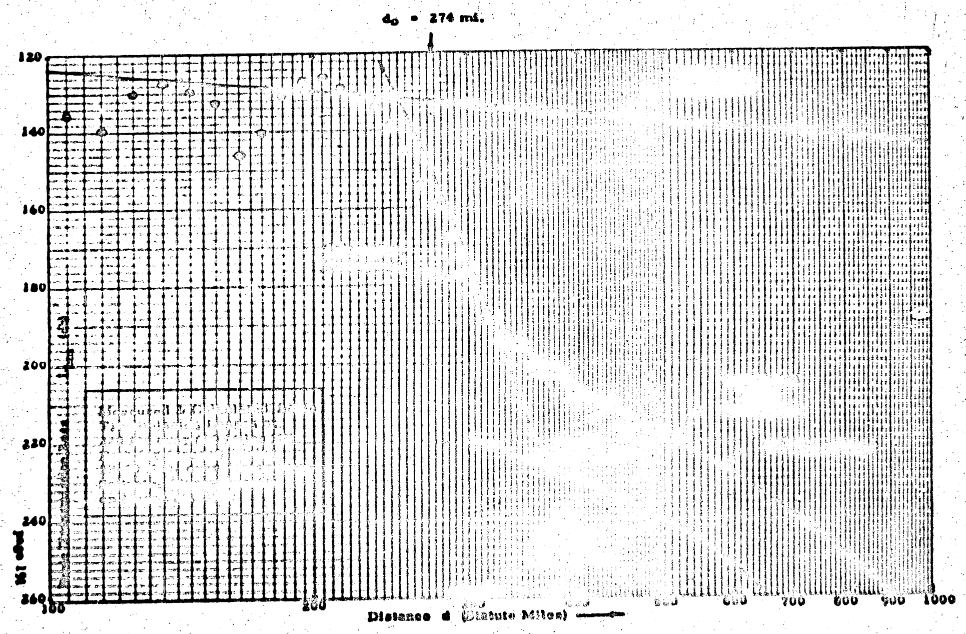
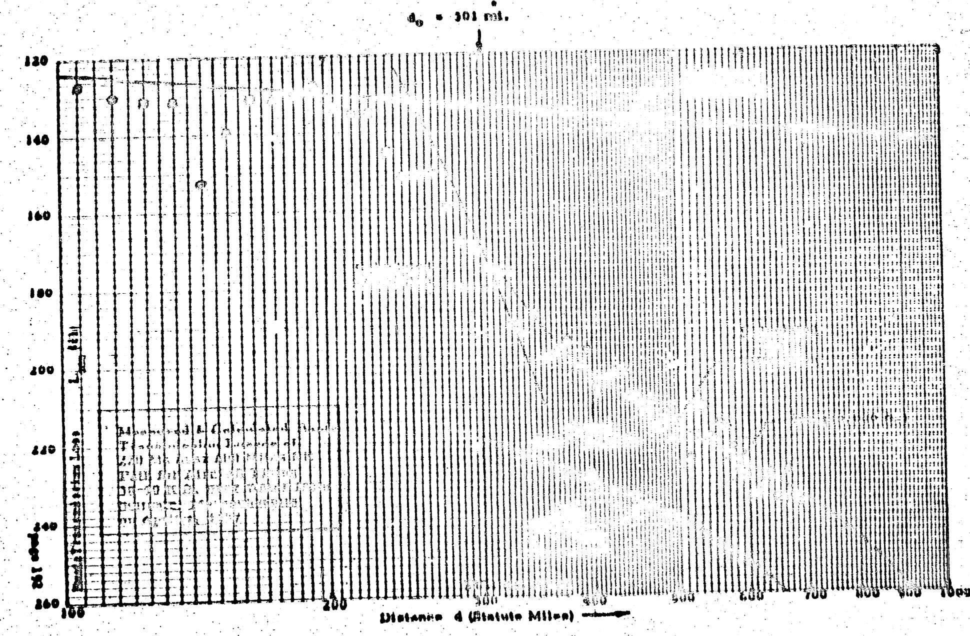

d_0 = 274 mi.
120
140
160
180
200
220
240
260
Basic Transmission Loss L_bm (db)
page 192
100
200
300
400
500
600
700
800
900
1000
Distance d (Statute Miles)


d_0 = 301 mi.
120
140
160
180
200
220
240
260
Basic Transmission Loss L_bm (db)
page 192
100
200
300
400
500
600
700
800
900
1000
Distance d (Statute Miles)

The predicted or calculated height gain in db is the difference in db between the value calculated for the surface (db) and that for the altitude 35,000 feet. The "measured" height gain was the difference in db between the calculated value of L_b at the surface and the measured values.

In Figures C-14 through C-20 are predicted and measured height gains for each of the flights. The data for G_H from these seven flights may be averaged and drawn to represent the month period covered by the seven flights. The results are shown in Figure C-21. A calculated curve is drawn based on a median d_o of 280 miles for the period. A median of the seven 10-mile medians for a given distance is obtained and plotted as a circle in Figure C-21. The standard deviation of the measured and predicted curve was 2.7 db.

There seems to be evidence in Figure C-21 of some peaking of measured gain with distance. One peak occurs at about 425 miles, another at 550 miles and a third at 665 miles (this is questionable as a trend for the period due to the sparsity of data at the extreme range). This is believed explainable on the basis of partial reflection from elevated layers[3].

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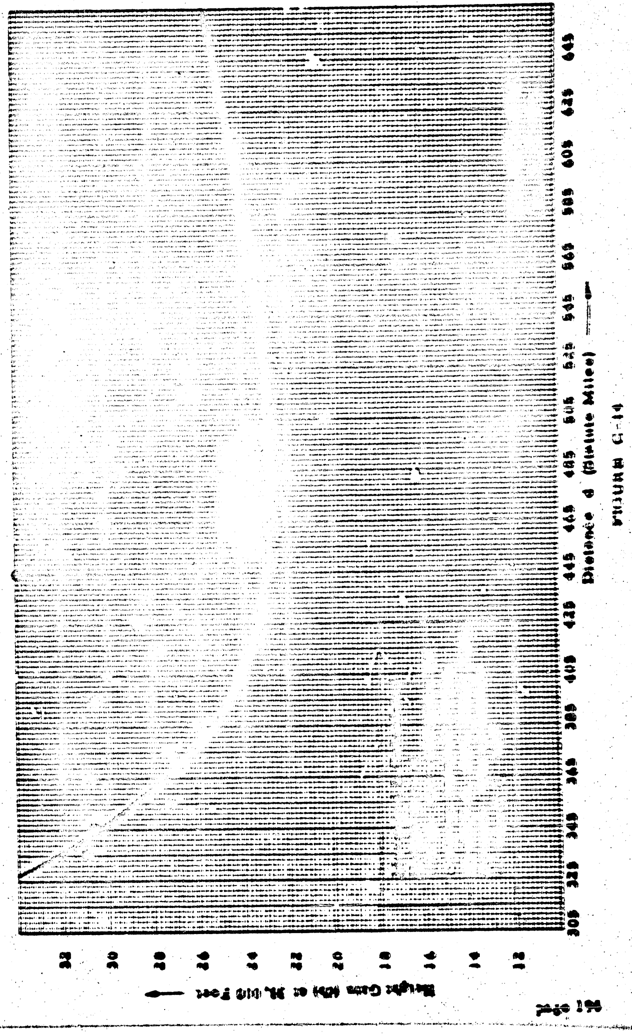


FIGURE C-14

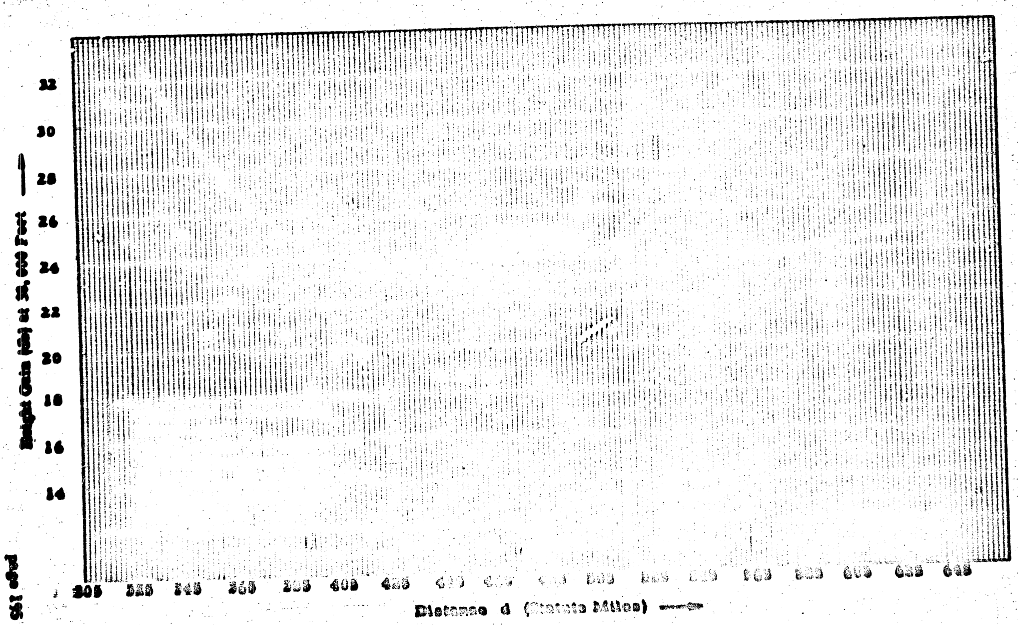


FIGURE C-15

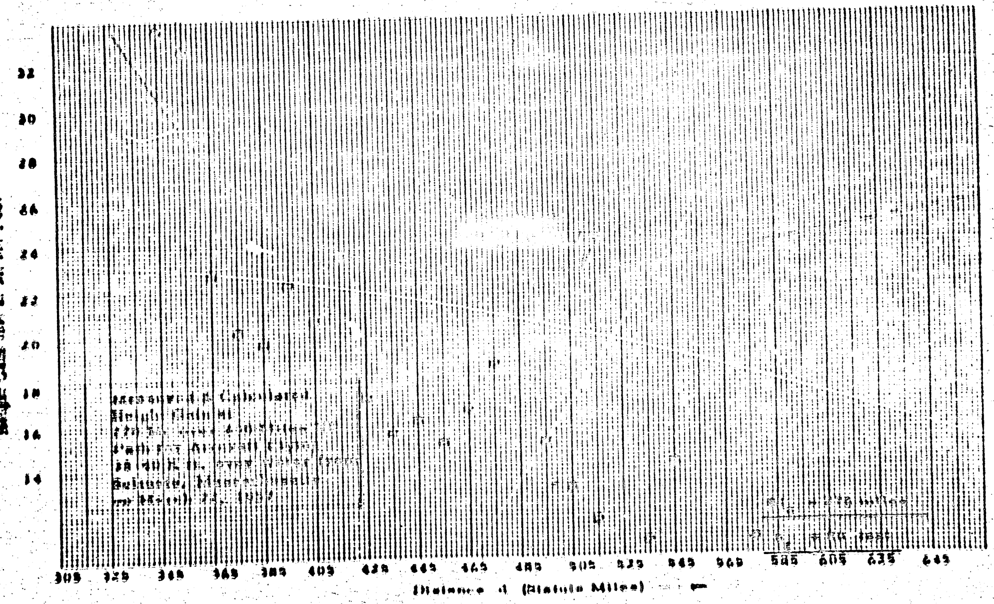


FIGURE C-16

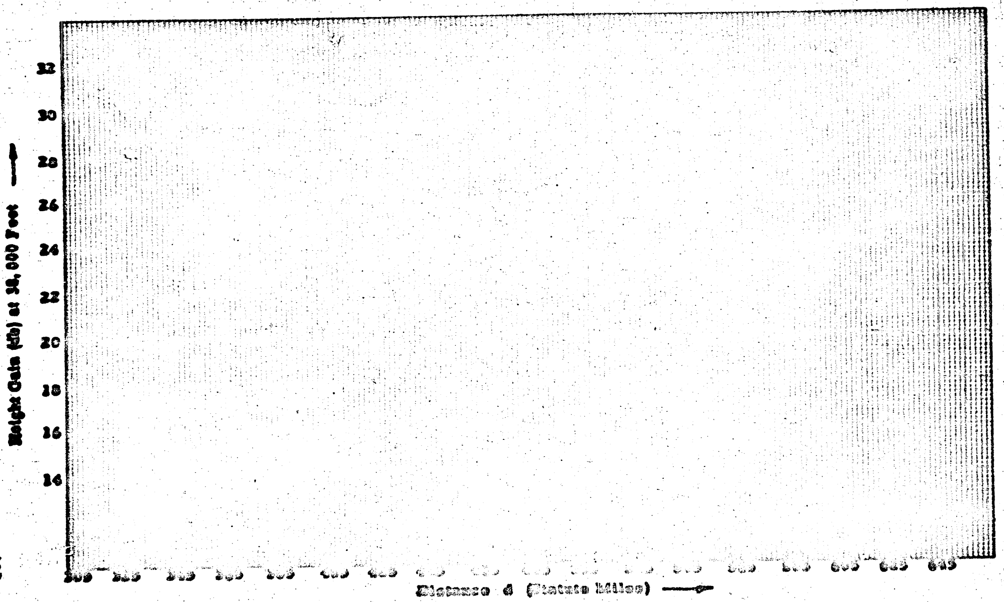


FIGURE C-17

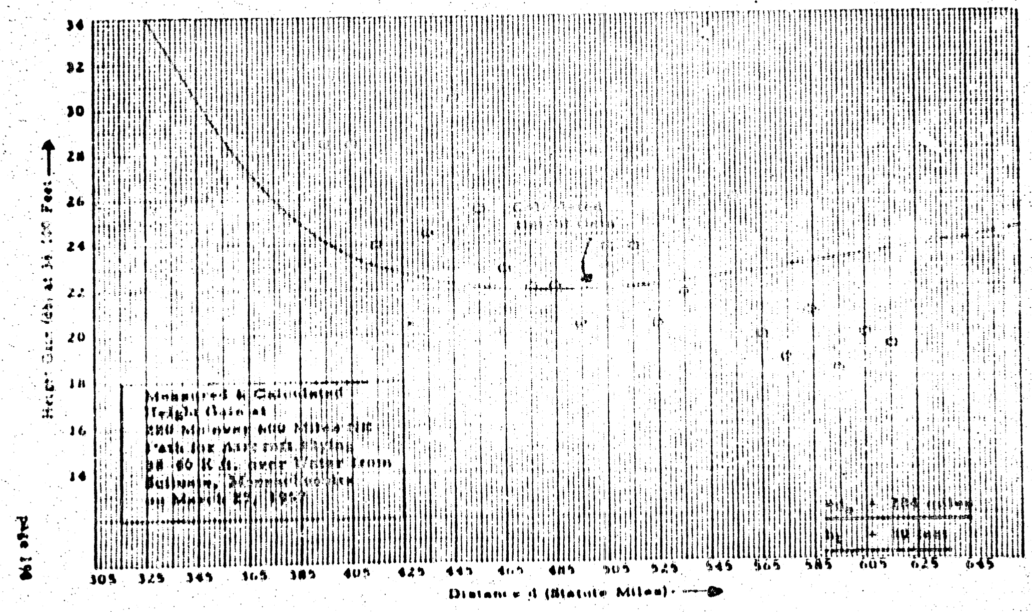


FIGURE C-18

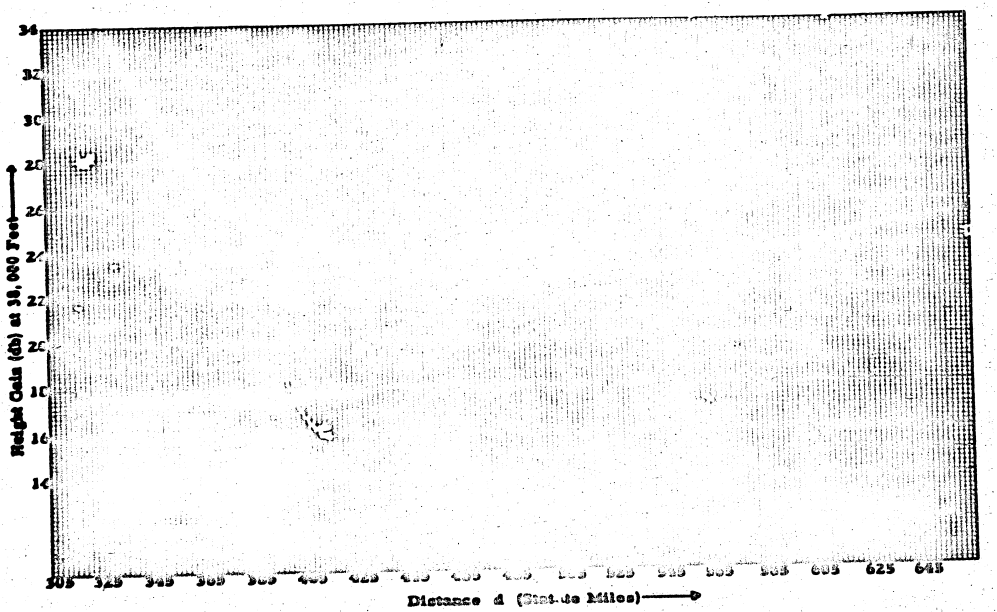


FIGURE C-19

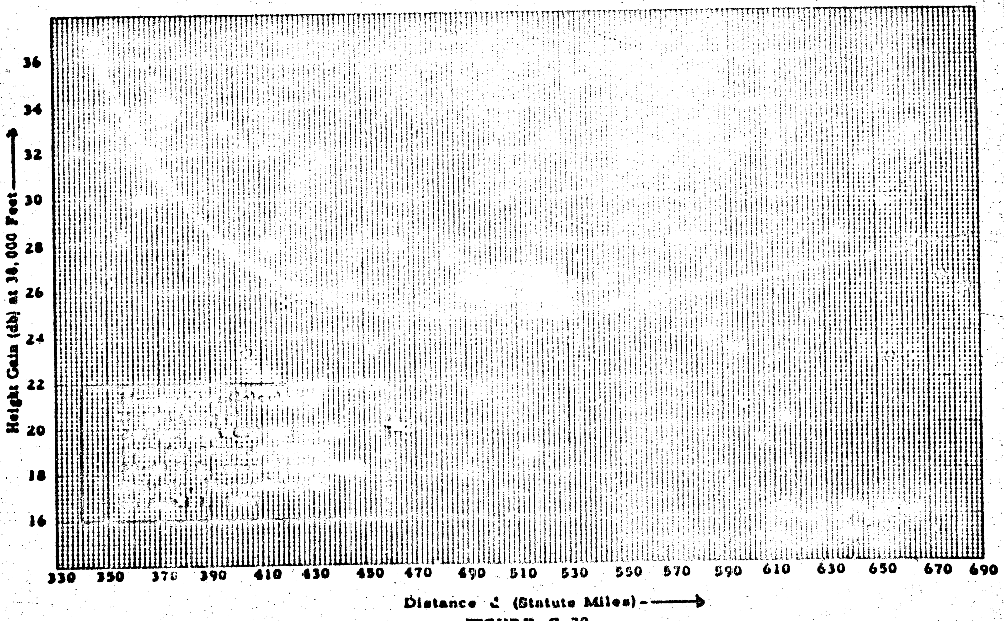
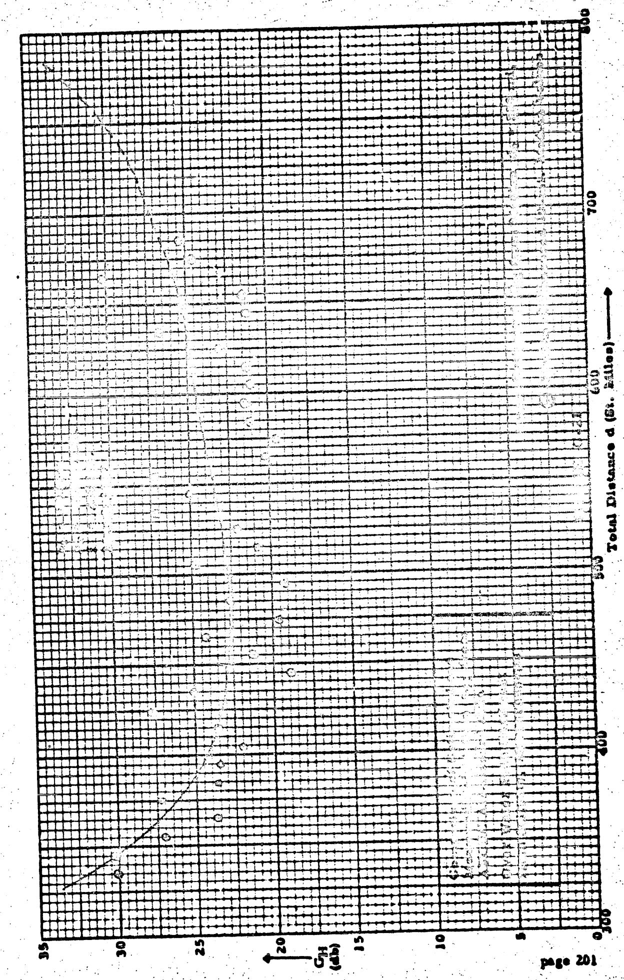


FIGURE C-20



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