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Ocean Wave Profiling Radar System

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Tracking Branch
Radar Division

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Ocean Wave Profiling Radar System

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An airborne oceanographic sensor, using a precise radar-ranging technique, is used for high-speed profile mapping of the ocean wave structure. The instantaneous distance to a spot on the ocean surface contour directly beneath a low-altitude level-flying aircraft is continuously recorded on a roll chart. This record portrays a laterally compressed cross section of the ocean surface along the flight track. Frequency modulation of a cw microwave source is used to obtain a measure of transmission delay by phase comparison of the demodulated return with the modulating reference signal. The use of microwaves is necessary to develop a narrow vertical beam, from a convenient size aircraft antenna, in order to short-range spot-illuminate and resolve the fine structure in each ocean wave as the profile is followed. Wave heights may be resolved to less than 6 inches, over a dynamic range up to 50 feet, by making the radar-illuminated spot sufficiently small and the transmission bandwidth sufficiently large. Multi-path interference between reflecting wave facets within the beam may seriously distort the phase coherence of the side frequencies in an unsmoothed FM ranging system unless this distortion is minimized by the transmission of a diverse spectral bandwidth. The equipment thus uses a sinusoidal FM output over a 4-percent spectral bandwidth centered at 9400 Mc to overcome this effect and to obtain the resolution necessary to prove the principle. Flight tests over various sea states experimentally verified the feasibility of the system and helped to establish design parameters for future equipment.

BACKGROUND

Successful oceanographic predictions and surveillance over the vast ocean areas of interest requires a current knowledge of sea conditions for the optimization of numerous operations. The desirability of obtaining accurate sea-state data, at a high rate, from many ocean wave samples is well recognized. In the system described, an airborne radar provides precision measurements of the differential path length between the airplane and a small spot on the ocean surface contour directly below a low-flying aircraft and plots a profile or relief map of the sea. Since the aircraft's horizontal velocity provides an accelerated scanning rate, compared to the real-time ocean wave progression, a long track of many waves may be quickly sampled. A series of tracks made at various headings, or in a "star" flight pattern, may be used to effectively evaluate the sea state from a statistical point of view.

To provide a radar system with an absolute range, or at least a consistent range reference point, it is necessary to "tag" the transmitted signal in some manner in order to measure the time delay of transmission to and from the target. For this reason, the subject radar utilizes the principle of frequency-modulated continuous-wave transmission to maintain the time reference. If a continuously transmitting microwave carrier is frequency-modulated and if the receiver is provided with sufficient isolation to avoid direct coupling from the transmitter, the reflected signal will enter the receiver, and the phase of the recovered modulation may be compared with the phase of the modulation of the outgoing signal. This comparison provides a continuous measure of the propagation delay and is read out in terms of changes in differential range as the ocean contour is scanned.

Even though the range-sampling spot is necessarily small relative to the ocean wavelength, the diverse nature of the surface reflectivity within the illuminating beam may still cause a form of target noise distortion due to multipath interference. This distortion will seriously affect the fidelity of the recorded profile. The interference effect is largely due to the granular structure (at microwave frequencies) of the major points of surface reflectivity. These points may be displaced in radial range by increments of a quarter or a half rf wavelength, which results in destructive or constructive interference, respectively. Both kinds of interference, of course, may be present. The interfering multiple reflections are vectorially

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combined at the receiver's input, with a consequent disruption to the coherence of the FM side frequencies. This distorts the phase of the demodulated envelope from a true phase delay reference, as determined by the range to the illuminated wave's center of reflection. The range error from the distortion is most severe during a cancellation period when the echoes from a pair of dominant surface reflectors of maximum vertical separation are in phase opposition. This develops a phase quadrature component that may be of sufficient magnitude to appreciably alter the indicated range of the system.

A high relative phase rate of the interfering rf signals, developed from a wide frequency deviation, will reduce the statistical importance of the cancellation period as it affects the mean range to target. The multiplicity of possible interfering paths in the practical case also reduces the net phase error as derived from a simplified two-path model.

It was shown, both theoretically and experimentally, that a narrowband FM/cw radar suffered from this multiple-interference anomaly to the extent of completely masking the true profile.* Peak errors in range, due to multipath distortion, may be several times the actual wave height of the surface wavelets within the illuminating radar beam.

A simplified three-path analysis of multiple interference showed that the quadrature component developed by a pair of cancelling paths may cause the resultant modulation phase to shift considerably in excess of the amount expected from the simple alternation between any pair of single transmission paths. The combined received power is amplitude-modulated by the multipath cancellations and reinforcements. However, these power changes are well within the limiting and phase-stable range of the practical receiver. Therefore, there may be no correlation between the range noise and the received power, a fact which appears to explain the effects observed in flight over the ocean using the narrowband radar.

The effects were also verified in the study by means of a waveguide multipath simulator.* The narrowband FM/cw transmitted signal is split into three paths, differentially controlled in amplitude and delay over these paths, and recombined at the input of the receiver. When the recombined signal power is held within the levels typical of in-flight reception, an error of 10 feet in indicated range may result due to signal interference from travel over a waveguide pair having a 2-ft differential delay. The third waveguide path is shifted in rf phase to show the extremes of the interference effect. The phase shift of the modulation is shown to be directly related to the differential range between the interfering paths and the relative amplitudes of the signals over these paths.

The study further showed that by making the transmitted frequency deviation large, the multipath distorting effects may be minimized. This fact provides a premise upon which the radar system design is based.

BASIC DESIGN CONSIDERATIONS

Introduction

The problems of precise radar ranging to an extensive-area target from a moving aircraft will be discussed, with the recording of ocean wave profiles as a specific objective. Since the ocean surface within the radar beam has a random character that subjects the reflections to a multipath environment, the radar system design requires an awareness of the extent that the interfering returns influence the fidelity of the differential range measurements. Spot illumination of the ocean surface with a wide transmitted spectrum localizes the range readout to within the wave height range of the surface wavelets. As the bomber-type aircraft moves over the surface at a speed of about 300 ft/sec, the capability of high speed translation of the wave height and period with a precision commensurate to other less-mobile methods is desired. The design philosophy was to construct as simple an operating system as possible, consistent with good short-range response, using experimental equipment to prove the feasibility of the technique. Recommendations toward prototyping an operational oceanographic survey tool will be given later in the report.

Ocean Wave Resolution

The ability of the recording radar to resolve the fine structure in the ocean surface profile is largely dependent upon the surface area sampled. If a narrow microwave beam is projected vertically downward from an aircraft in low-level flight so that the circle of surface illumination covers

*C.M. Morrow, "Radar Altimeter—Ocean Wave Profiler. Part 1," Radar Return Symposium Transactions, pp. 169-197, NOTS Report TP-2338, DDC Document AD-244937, 1959.

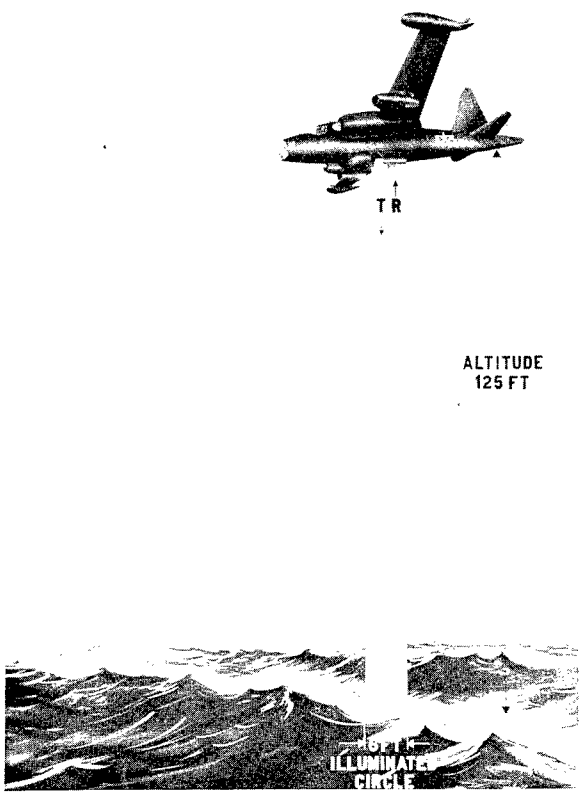

Fig. 1 - Airborne ocean wave profiling radar system

only a small portion of the significant length of the ocean wave, then continuous differential range profile measurements may be made as the beam is swept over the waves. The size of the circle of surface illumination determines, to a large extent, how much of the fine structure that rides on top of the major wave shape will be resolved.

Since the fixed vertical beam is swept over the surface at the aircraft velocity, an approximate surface distance of 100 yards is scanned horizontally along the flight track during each second. Figure 1 is an illustration of the mode of operation and shows an imaginary visual projection of the microwave beam pattern from the transmitting antenna (T) and the portion of the scattered return (R) accepted by the receiving antenna. The high scanning rate of 100 yd/sec requires that the electronic system, shown in block diagram form in Fig. 2, have the necessary response to follow the incremental range changes as the 6-ft-diam microwave beam progresses over the surface undulations. (For altitudes greater or less than 123 feet, the beam diameter will increase or decrease, respectively.) Since the diameter of the projected spot from the antenna is finite, the indicated range will yield only a mean height of the combined capillary waves and short wavelets that ride on the major wave structure. These two minor wavelets may, therefore, not be recorded in their true perspective due to this beam averaging.

The resolution of the flight tested system, using the beam parameters shown in Fig. 1, is adequate for recording waves that may be as small as 6 inches in height and as short as 15 feet in length. System linearity and stability appear adequate to reproduce maximum wave profiles up to 50 feet high from crest to trough. The ability of the aircraft to maintain altitude, or the deflection sensitivity of an aircraft motion corrector, or a combination of the two, determines the maximum length of the ocean waves that may be accurately recorded.

The wave profile measurements are recorded visually on a roll chart. The pen deflection is calibrated in wave height, and the paper speed is scaled to the ground speed of the aircraft to yield wavelength. Typical chart recordings of radar profiles obtained from two different sea states are shown in Fig. 3. The strips each represent a horizontal distance along the flight track of approximately 0.7 mile that was recorded in a 12-sec period. The vertical distance is calibrated to indicate a full-scale chart deflection of 10 feet. A lateral compression ratio of 55:1 permits the display of many wave samples within a reasonable chart length. The long-period drift, shown more clearly on the calm water chart, is that of the aircraft's vertical motion. The wave heights that are resolved correlate well with visual estimates and are indicative of the background and resolution capability of the system.

Ocean Wave Radar Reflectivity

The slope of the wave affects the reflection coefficient of the surface, as well as the degree of range averaging within the radar beam. A high wave slope would be expected to cause extensive signal fading because the beam impinges on the side of the waves at a high angle of incidence. This should divert more of the reflected signal away from the receiver if it were not for the mechanics associated with wave generation. Since the wave slope is related to the wind speed or

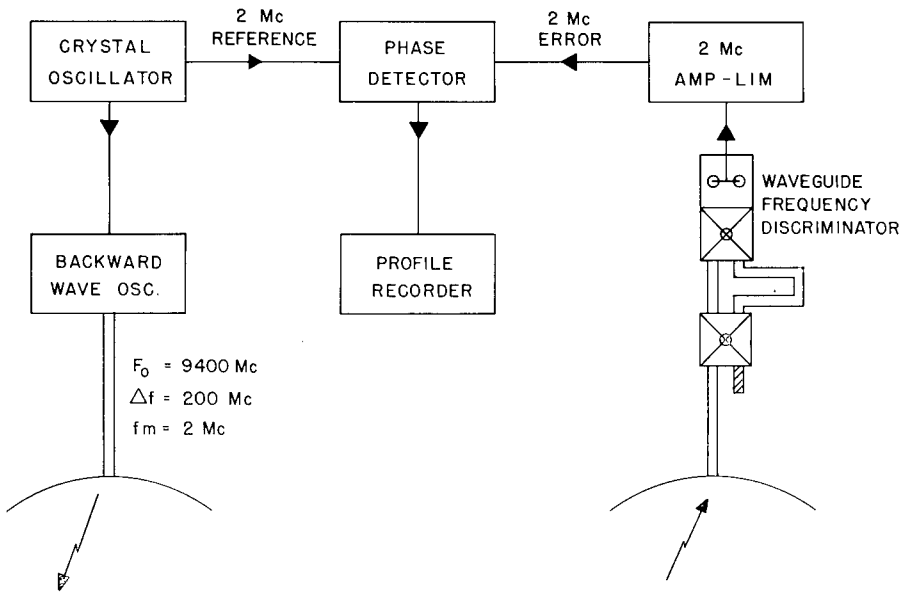

Fig. 2 - NRL wideband FM/cw radar system

force of generation, the duration of build up, and the fetch or distance traveled after generation, the higher slope waves are generally roughened by the wind to an extent that compensates for much loss in reflection.

A correlation between radar backscatter and wind conditions is particularly possible at transmission wavelengths of three centimeters or less if an angle of incidence near grazing is utilized. It would appear that the return should maximize at an operating wavelength that matches the dominant physical wavelength of the capillary waves. However, to relate radar reflectivity to wind, and thus to sea state, is not valid due to the short time effect of the wind on the backscatter, which may or may not be related to the height of the waves.

Surface reflectivity appears to have the greatest variability over high waves of long fetch, a synoptic condition usually associated with sea state after the passage of a storm. The complexity of the wave structure therefore makes the direct range measuring technique utilized for recording wave height most desirable. Both the changes in wave height and the roughening of the water surface by local wind may independently affect the amplitude of the return. It is therefore necessary to suppress those amplitude changes that influence the accurate determination of the range or wave height. Accuracy in reproducing the wave profile will be obtained when the range changes are derived solely from the changes in phase delay of the modulation envelope of the FM signal. This instantaneous phase must be truly indicative of the propagation time required to transmit the signal to and from the target's center of reflectivity within the main beam lobe of the antennas. The wideband deviation of an FM transmitter, combined with phase-stable amplitude limiting over a large dynamic input range at the receiver, will satisfy these requirements.

Ranging Fidelity

A radar range measurement can only be as precise as the measurement of the end points of the path, *i.e.*, the point of reflection and the position of the radar. Since the round-trip propagation time of a radar signal is 0.25 μsec when the radar is 123 feet from a reflecting surface, the motion of a 200-knot aircraft flying parallel to this surface and carrying the radar is only one-thousandth of an inch during this interval. Therefore, a single measurement or phase comparison of the distance to the reflecting surface should not be influenced by aircraft motion during the transit time of the radar echo.

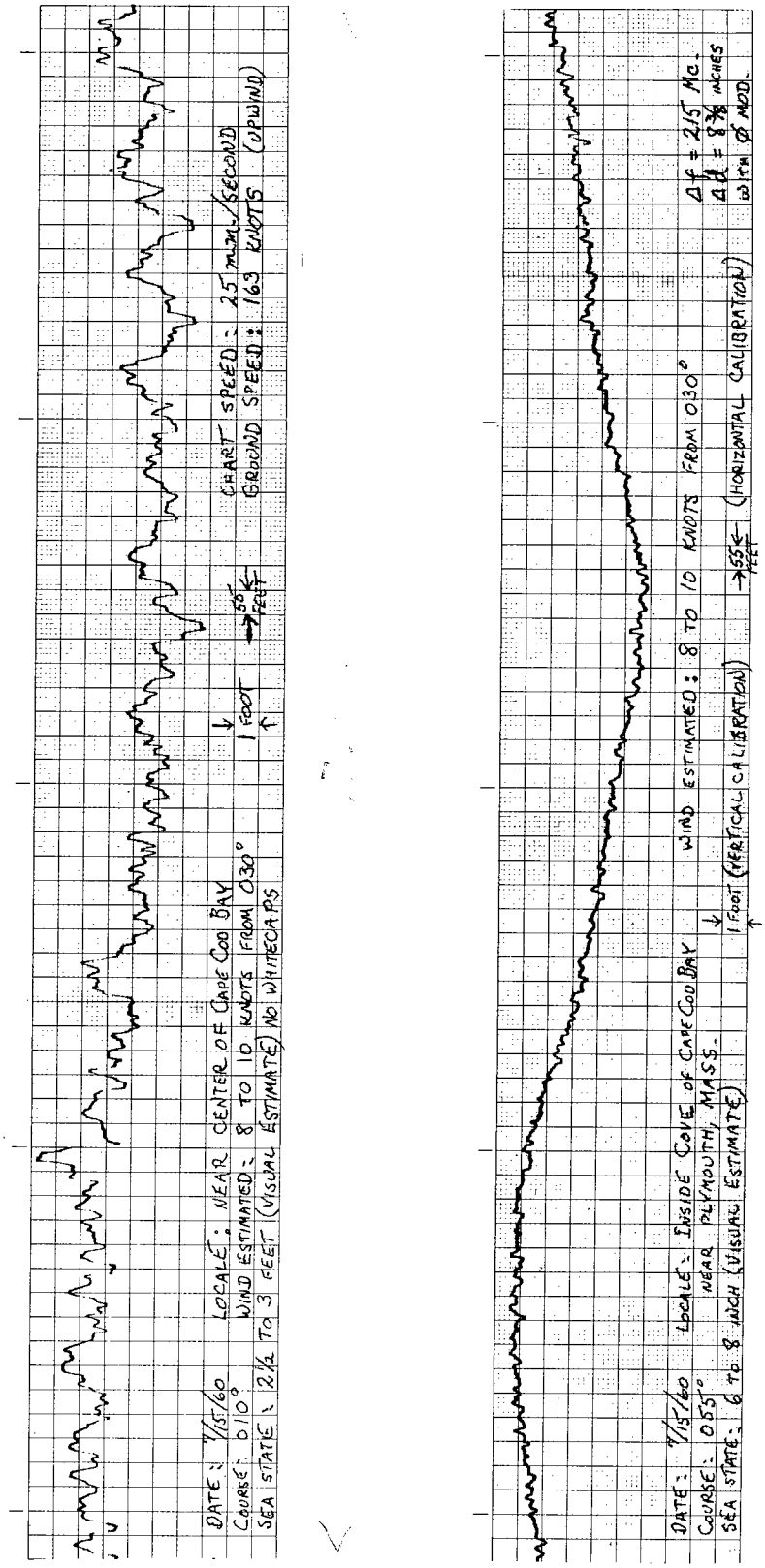

Fig. 3 - Recorded wave profile from two different sea states. Each profile spans a horizontal distance of approximately 0.7 mile and was recorded over a 12-sec period of time. The full vertical range on each chart represents 10 feet.

The FM transmitter operates in a cw mode with range information being continuously obtained, and therefore no range fidelity loss is experienced by positional changes during interpulse periods. This does not infer that a short-pulse radar of sufficient repetition rate and bandwidth would be degraded in ranging fidelity. In either case, ranging fidelity is largely dependent on how closely the apparent centers of reflection from the numerous wave facets within the illuminated spot conform to the shape of the wave. Some of these glittering points of reflection may be displaced more in radial range than others. The instantaneous range measured will be the result of the vectorial combination of the returned signals from the many reflectors that are illuminated by the divergent antenna beam. This indicated range will vary in an ambiguous manner within the spatial depth of the reflecting area due to changes in reflection characteristics of the individual reflectors. So, the technique of illuminating a small area restricts the spatial depth of ambiguity to an acceptable range error provided that a sufficient transmission bandwidth is utilized.

If it were possible to focus the beam to a point of reflection that would instantaneously trace the surface, the range ambiguity should be reduced to zero. However, the practical limit to the reduction of range ambiguity by beam narrowing appears to be within the peaks of the short wavelets on the surface waves. The illumination area is an acceptable compromise between the size of the antenna, the operational height, and the microwave operating frequency suitable for all-weather reliability. (These factors are more fully dealt with in the appendix.) Optimum ranging fidelity is obtained by transmitting a narrow beam of wideband energy to confine the reflection to a homogenous return from within the mean continuum of the wave slope. Foam and wash from breaking waves must necessarily be smoothed over, together with the minor undulations that ride on the major wave structure.

Selection of Modulation Frequency

A sinusoidal modulation with a frequency of 2 Mc is used to deviate the microwave FM carrier over a maximum bandwidth of ±200 Mc. The selection of these parameters permits an accurate measurement of the propagation time to and from the target provided that the sensitivity, linearity, and stability limits of the phase-sensing detector are not exceeded. Changes in aircraft altitude, as well as the range changes due to the ocean wave fronts, affect the delay and, hence, the relative phase between the modulation in the transmitted signal and in the echo.

Delay changes are continuously compared with the modulation reference in a quadrature phase bridge to produce an output proportional to the changes in range. The polarity and amplitude of the bridge output follow the changes in phase of the demodulated echo, which is balanced at a mean 90-degree phase difference between the 2-Mc reference and the echo. A phase correction control in the reference channel is available to set up this balance at the start of each operational run at the desired altitude. This assures that the phase bridge will be operated over the most linear portion of its range, which is essentially uniform over a ±50-ft change in altitude or profile range. The output is applied to a zero-center balanced chart recorder. Mean sea level will be indicated by the middle of the chart, and the wave crests and troughs will be traced out above and below this central portion.

The extent to which a given phase shift affects the indicated range is inversely proportional to the modulation frequency. For example, a 1-1/2-degree phase shift to a 2-Mc modulation frequency represents about a 1-ft shift in indicated wave height, while if a 1-Mc modulation frequency were used and shifted in phase by 1-1/2-degree, the wave would appear to shift by 2 feet in height.

The modulation frequency of two megacycles (2 Mc) was determined by the accuracy requirement of the system, the problem of redundancy in readout from the phase-sensing circuitry, and the inherent crosstalk rejection when operating at an altitude of 123 feet. The propagation transit time over a round-trip path of 246 feet is 0.25 μsec, which also represents a 180-degree phase shift or delay for a 2-Mc frequency. This out-of-phase relationship between the modulation components of the return and the local leakage results in immunization of the beat spectral energy in the cross product of the two. The contribution to the fundamental is of the second order since the energy is spread over many spectral lines when the time delay of the primary target results in π radians of the modulation frequency. This produces a high phase rate of simultaneous side

frequencies, resulting in a high order of harmonics that are rejected by the tuned circuitry of the 2-Mc amplifiers. In addition, any crosstalk or direct coupling from the crystal oscillator or modulator into the 2-Mc amplifiers or limiters is expected to be either in phase or 180 degrees out of phase. This coupled signal will vectorially add to the demodulated return with less phase distortion than would the same signal if added in quadrature.

The readout redundancy or identical dc response from the phase-sensing circuits will be obtained at odd altitude multiples of 123 feet (*i.e.,* 123, 369, 615, etc.), and an opposite sense to the readout will be indicated at the even multiples (246, 492, 738, etc.). This has the effect of recording the wave crests with an upward deflection on the chart when operating at the odd intervals of altitude, and with a downward deflection at the even intervals. The periodic modulation causes a repetitive zero readout, or "altitude hole," at multiples of 180 degrees of phase delay. These altitude holes are undesirable in an altimeter because the repetition confuses the altitude information, and generally a lower modulation frequency is used to avoid this ambiguity. However, the profiling application is not hampered by this type of response since level flight, within a ±50-ft variation, is used during a run. Also, the accuracy of the system is retained by the use of the relatively high modulation frequency of 2 Mc and permits the use of a simplified type of phase detector.

It may be shown that if the modulation frequency were to be decreased, the phase distortion due to the multipath interference will also decrease. However, the range error or noise will not be proportionately decreased since a phase error at the lowered frequency now represents a greater range error, and the other desirable factors of using a modulation frequency near 2 Mc will be sacrificed. An increase to the frequency deviation, on the other hand, decreases the phase distortion, and when a sufficiently diverse spectral distribution is transmitted, the multipath distortion is minimized. The parameters indicated by the physical requirements of the flight-tested system show that a frequency deviation Δf up to 200 Mc is necessary.

OPERATIONAL DESIGN CONSIDERATIONS

A Navy patrol bomber was utilized at the Naval Air Development Unit, South Weymouth, Massachusetts, for field measurements of the wideband FM/cw radar developed for ocean wave profiling. The conical patterns of illumination for transmission and reception, shown in Fig. 1, are developed from two X-band antennas mounted in the bomb bay section of the P2V-5 aircraft. Low-level operation permits the narrow beams, which are 2.75 degrees in half-power beamwidth, to illuminate and observe a common 6-ft spot directly beneath the aircraft. This circular field is small relative to the length of most ocean waves. Steep wave slopes, associated with high-amplitude waves or wind-driven wavelets, are resolved if the area sampled is small relative to the surface area of the ocean wave.

The compatibility of an antenna size, suitable for mounting on the aircraft, with the chosen microwave frequency range of radiation establishes the sharpness of the beam. The beams are derived from two 30-in. paraboloids operating near 9400 Mc. The paraboloids have a depth of 5.6 inches and are illuminated with 15-db, tapered, Question Mark feed horns at a 10-in. focal length. Separate antennas (shown in Fig. 7) are used for the transmitting and receiving modes. This provides an isolation in excess of 50 db to the local leakage. When the transmission losses approach the decoupling losses or isolation between the two antennas, the result is excessive noise in the profile record. This effect appears to be most severe over glassy-like waves when surface winds are essentially absent. Beam tilt due to aircraft pitch or roll causes large changes in the echo amplitude from the smooth sea.

An improvement to this noise level is possible by a nulling technique. It is possible to cancel or balance out some of this antenna feedthrough by the injection of an out-of-phase, equal-amplitude signal over a secondary path. This path length should be equivalent to the feedthrough or local leakage path to offset the modulation phase crosstalk and should be controllable to balance out the phase and amplitude of the microwave leakage. The technique of nulling the rf leakage on an aircraft has an inherent problem of stability (temperature and microphonism), as well as a narrow bandwidth of control over only one primary leakage path.

Since it was desired to record waves as small as 6 inches in height, operation was restricted to low altitudes because of the size of the illuminating spot on the surface necessary to resolve such waves. This spot size appeared to be more important in obtaining good records than the possible noise contributed to the profile from excessive transmission loss. However, noise due to the local leakage was minimized through high-altitude nulling. Improved techniques of decoupling or desensitization to feedthrough are available which may be used to duplex the cw signal and allow single antenna operation at a higher altitude, but the added sophistication was avoided in this evaluation program.

No antenna stabilization or alignment correction, other than aircraft trim experiments, was used during the flight tests. The trim experiments showed a noncritical response to a change of ±3 degrees in trim over a scattering or normally wind-roughened sea. However, the signal was adversely affected by similar trim changes over a specular target, such as a calm lake or airport runway. Thermal air currents over the land at low altitude will cause severe pointing errors due to pitch and roll of the aircraft. The homogeneity of the air over the open sea improves the flight stability and is more favorable for operation with a nonstabilized antenna mount.

Since the radar responds to range changes between the aircraft and the surface spot, vertical motion of the radar platform must be separated from the rise and fall of the wave profile beneath it. The use of a high-pass filter in the radar's profile output was partially effective during the flight tests since the wave data generally covers a higher frequency band than the vertical drifts in the airframe. However, this filter is not successful during crosswind flights or while operating in unstable air. On a crosswind flight, the flight track is more parallel to the major wave pattern and this lowers the recorded frequency components in the waves until they may well approach those of the aircraft motion, and thus the filter removes some of the desired wave data. On the other hand, a weather front of unstable air may cause the airframe to bounce at a period similar to that of the waves and add undesired information to the profile record. This discrepancy might be minimized on most ocean flights by the addition of an aircraft motion canceller that is capable of accurately responding to vertical deflections of ±50 feet of the aircraft. A technique useful for obtaining a space displacement from some reference position is to doubly integrate the output of a sensitive accelerometer. The forces of vertical acceleration of a P2-V in level flight might typically equal the pull of gravity in turbulent air, and in smooth air rise and fall with a thousandth of this force. Obviously, the minimum sensitivity and calibration of such a unit may require blocking of the extreme low-frequency components from the wave record.

A pilot's meter, which indicated slow changes from the desired operational altitude, was found to be of considerable help in accounting for aircraft vertical motion during a profiling run. This meter obtained a driving signal from the wave profile output of the radar. Violent meter flutter from the wave data was avoided by dividing the profile output into two channels and separately filtering the signal for the pilot's meter in a low-pass filter. This limits the response to the acceleration components that are normally in control of the pilot and proved to be a reliable operational aid.

The typical procedure used to obtain a useful wave recording requires level flight at a low altitude on a given course for a period up to approximately five minutes. The relative modulation phase of the echo and the 2-Mc reference is set near a quadrature relationship at the selected operational altitude. A reference phase corrector is available for this purpose. A balance produces a zero mean dc voltage from the phase bridge, and any subsequent range change causes a plus or minus deviation from the zero or centerline position on the chart recording. The deflection is proportional to the profile changes and to the mean altitude of the aircraft. A calibration delay step equivalent to 5 feet may be inserted in the reference channel at will to establish the desired recorder gain settings.

SYSTEM OPERATION

Modulation Phase Indicates Wave Height

The block diagram of Fig. 2 illustrates the fundamental components of the wideband FM/cw radar. The propagation delay is proportional to the phase delay of the 2-Mc frequency modulation (FM) that is deviating the

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continuous-wave 9400-Mc carrier (F0) over a spectral bandwidth of ±200 Mc (Δf).

A change in the propagation delay takes place when the contour of the reflecting surface is scanned. If the range to the water surface changes either by vertical displacement of the aircraft or by movement of the radar beam to another vertically displaced position on the surface contour, then the propagation delay is similarly changed. For example, if level flight moves the vertical beam from a wave crest to its trough, the round-trip distance between the radar and the reflecting water surface is increased. This proportionately increases the delay of the 2-Mc modulation recovered from the received microwave echo. When this modulation is phase-compared with the original 2-Mc reference signal in the phase detector, an output is obtained which is increasing negatively in proportion to the depth of the wave trough and is applied to the chart recorder. The amplitude of the downward chart deflection may be scaled from the calibrated grid lines on the chart paper to determine the depth of the wave trough in absolute units.

These grid lines may be calibrated to represent a given number of feet by momentarily adding a fixed delay to the system. This steps or offsets by a known amount the relative phase of the two 2-Mc signals applied to the phase detector. The recorder sensitivity control is then adjusted to cause the chart pointer to deflect by an amount that matches the equivalent range delay step. For example, a 10-nsec delay inserted in the reference channel will step the chart deflection upward by an amount equivalent to a change of 5 feet of wave height. Recorder gain is then set to produce a deflection offset of five grid lines from its normal position.

Information Handling Requirement

The information rate to be recovered from the radar echo is determined by the ocean wavelength, the fine structure in the wave, its aspect ratio or slope, and the scanning velocity of the aircraft. High-speed flight over short-crested waves on a course normal to the major wavefronts produces the highest wave data rate. If we pass over an assumed sinusoidal ocean wave with a crest-to-crest distance of 20 feet, at a ground speed of 300 ft/sec (178 knots) we will trace out one wave in 1/15 second. The radar receiver should then apply a sinusoidal 15-cps waveform to the recorder with an amplitude proportional to wave height. Waves customarily have a greater distance than 20 feet between crests, and most wave data of interest is confined within a 0 to 5 cps bandwidth at this scanning velocity.

On the basis of required information rates, the system appears to be extremely wasteful of spectral bandwidth. However, this is not true where the phase of the information must be accurately preserved during transmission in a multipath environment.

The phase of the modulation envelope of a narrowband FM or an AM system is markedly disturbed by interference distortion components developed whenever a signal travelling over a supplementary rf path passes through a 180-degree rf phase difference with respect to that of the primary path. The extent of this distortion depends on the relative amplitudes of the two signals, their differential path lengths, and any modification due to additional supplementary paths. The envelope phase is governed by the coherent relationship of the side frequencies recovered by the receiver. When only a few closely spaced side frequencies exist with any appreciable signal strength, as in the case of narrowband FM, then they may be simultaneously and similarly affected by the interference to cause a phase pulling of the modulation. However, when a wideband spectrum is transmitted, the modulation envelope phase is now governed by the coherent relationship of many side frequencies, which are spread out over a bandwidth 50 to 100 times broader than the narrowband system. Although these supplementary rf phase cancellations still exist, the side frequencies are randomly affected, and the interference is averaged with little or no effect to the gross coherence of the envelope.

A study of these multipath effects, when an idealized predetection limiting case was assumed,* indicated the need for a deviation of 200 Mc in order to reduce an anticipated range distortion error to less than 6 inches. An analysis of post-detection limiting also indicated a similar requirement for a broad transmitted bandwidth. The latter study assumed a high signal-to-noise ratio with negligible quadrature interference due to

*Ibid.

leakage. The broad transmitted bandwidth is primarily necessary because the illuminated water surface is not a point source type of reflector. The frequency deviation covers a bandwidth that is many times that required simply for the reproduction of relative target motion. However, the required bandwidth is comparable to that of a short-pulse radar that would be capable of range discrimination to less than 1 foot.

Information Retrieval and Detection

Wideband signal processing in a microwave receiver generally requires the use of travelling-wave-tube amplifiers in the rf link and distributed line amplifiers for the video components of the signal. Since the information rates are relatively low and the signal-to-noise ratio high, it is practical to discard the wide spectrum as soon as it has served its usefulness, *i.e.*, in combating the phase distortion to the modulation over the transmission path. This is done at the receiver input through a waveguide frequency discriminator which extracts the 2-Mc modulation directly from the FM microwave signal recovered by the receiving antenna. A frequency-sensitive, balanced, crystal-video type of detection is used. The response to changes in frequency is accomplished in a closed waveguide interferometer that is primarily sensitive to phase only. This frequency-discriminating demodulator, shown in Fig. 4, is more completely described in a following section.

Since the system is operated at short range, an adequate signal-to-noise, or quieting, ratio is obtained. This permits the use of low transmitter power and direct crystal conversion at the receiver input.

Superheterodyne conversion and predetection limiting is normally required in FM communications where the signal levels are low and the information content is high. In the wave profiler these factors are reversed, and, therefore, the use of postdetection limiting following the demodulation in the waveguide discriminator is practical. This avoids the sophistication of feedback techniques or wideband amplification and limiting at the receiver.

System Sensitivity Determination

In order to simplify the bandwidth requirements of the receiver by the use of crystal-video techniques, we have sacrificed the high receiver sensitivity normally available with superheterodyne reception. It is therefore practical to determine the feasibility of the simplified system using the postdetection limiters.

A series of calculations obtained from (a) the measured system parameters, (b) the radar reflectivity measurements, and (c) the operational use of the equipment over various sea states are described in the appendix. This serves to clarify what limitations are imposed on the operation of the system and also why recommendations for improvement are offered.

Microwave Frequency Discriminator

A frequency selective or responsive microwave detector is desired for demodulation of the broadband frequency modulation where a 1 to 2 percent bandwidth is utilized. Most FM/cw radar altimeters use a small frequency deviation, typically less than 5 Mc, at a low modulation rate permitting direct mixing of the transmitted signal to produce beat frequency audio conversion. In this case, where a frequency deviation of 200 Mc at X-band is used, it is desirable to introduce a frequency selective element that will detect or recover the modulating signal directly without the use of heterodyne, homodyne, or wideband i-f conversion techniques.

Microwave cavities are typically constructed having Q values in the region of 2000 to 4000 and are of the absorption or bandpass type. Such elements are particularly useful where it is desired to accept or reject an extremely narrow bandwidth and may functionally be used in tandem as a sharply selective discriminator. The physical movement of the cavity tuning adjustment is quite critical and may be detuned a full bandwidth by a movement of only 1/Q of the wavelength. Therefore an alternate method for obtaining selectivity is desirable.

The discriminator (Fig. 4) attains its selectivity by means of a microwave interferometer technique.* As indicated by the block diagram of Fig. 2, the incoming signal from the receiving antenna may be split into two paths by means of a sidewall hybrid power splitter. One of these paths exceeds the other by 10 to 20 guide wavelengths.

*H. P. Raabe, "Measurement of Instantaneous Frequency with a Microwave Interferometer," *Proc. IRE* **45**(No. 1):30 (1957).

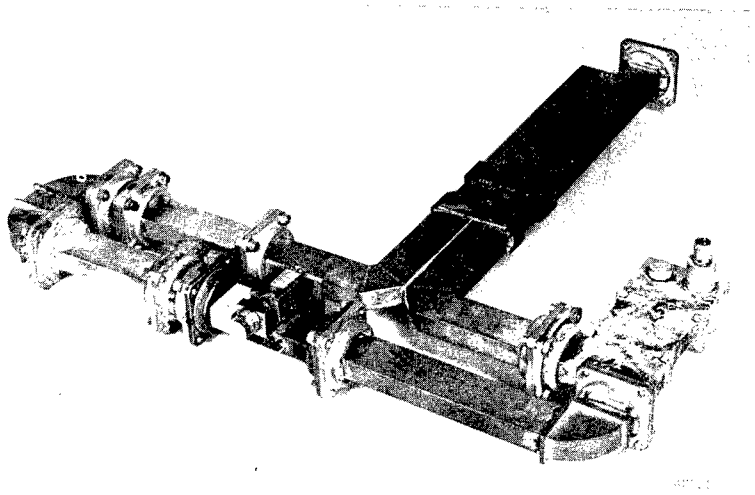

Fig. 4 - Waveguide microwave frequency discriminator

When these two paths are recombined in the dual balanced crystal mixer (to the right in Fig. 4) there will be a repetitive series of center frequencies where the phase and amplitude of the combined signals at the crystals will drive the series-connected diodes by an equally balanced amount. This will produce a zero dc output at the bridged-T coaxial jack because an equal-amplitude and opposite-polarity rectified current flows in the crystals. The phase interference of the signals over the differential paths likewise produces an unbalanced drive to the crystals because the incoming frequency is deviated $\pm\Delta f$ from a center frequency. This results in detection of the sinusoidal frequency modulation of 2 Mc by the alternate cancellations and reinforcements of current flow in the individual crystals.

Since the Q of the discriminator is approximately equal to 4n, where n is the number of differential guide wavelengths, a linear broadband response from the crystals over a 1-percent bandwidth (Q = 100) is conveniently attained. Physical adjustment of the differential paths by $\lambda/2$ is required to offset the discriminator tuning by its full bandwidth, and thus it is considerably less critical to adjustment and temperature stability than a discriminator of the cavity type.

Since it is desired to make the FM detector immune to AM, a dual balanced mixer detector of the quadrature injection type is employed to reject the common-mode modulation-frequency amplitude variations. The use of a double probe detector, described in a previous paper,* functions similarly. This requires that the phase shifters be broad-banded, the power recombination be uniform, and the crystal detectors be matched in sensitivity and voltage-current characteristics.

Functionally, the discriminator unit divides the input signals into two equal-amplitude and quadrature phase paths by means of a short-slot sidewall hybrid junction. One arm of the four-port device is terminated in a matched load to absorb any reflected energy due to possible mismatch. The length of the two paths are suitably adjusted to obtain the desired Q for optimum reception of the transmitted bandwidth.

Since the unit is not frequency limited by any element other than the bandwidth of the guide, hybrid, and crystal holder, signals outside of the transmitted bandwidth may cause interference by nonsymmetrical biasing of the detectors. It is to be noted on Fig. 4 that one path is provided with a dielectric phase shifter for adjusting the discriminator response symmetrically about a selected center frequency. This may likewise be accomplished by suitably shimming the guide length of the differential or trombone section, or conversely, by selecting the center frequency at the zero crossover of the detector output. To maintain symmetry of output with minimum harmonic distortion, it is important to be able to set and maintain a stable center frequency. Symmetry of

*Ibid.

the frequency modulation in producing balanced sidebands from the transmitter likewise will affect the amount of output distortion. Redundant and out-of-phase output responses from the discriminator are obtained at odd multiples of the discriminator bandwidth. To avoid this, as well as undesirable interference from off-frequency reception, it should be possible to band-limit the input signals. This may be accomplished by means of suitable bandpass filters of the (a) stagger-tuned cavity or (b) waveguide-beyond-cutoff types.

The video output is taken at the junction of the two crystals by means of the bridged-T coaxial connection in the dual balanced mixer, and thus the video signals from the two diodes are added and divided by two in this bridging connection. The balance of the video sensitivity between the two crystal detectors minimizes the even-order harmonics in the detection process. Therefore, it is desirable that the crystal diodes be matched in characteristic impedance and rectification efficiency. The power sensitivity of the detector bridge is increased approximately 9 db when the crystals are dc-biased to cause a crystal current flow of 30 to 50 μamps in the absence of an rf input signal.

The video-crystal detectors are a matched pair of the reversible polarity type MA418AM which, when connected in a balanced configuration, effectively suppress 2-Mc phase distortion due to amplitude variations of the input wave. Ideally, the phase of the demodulated error signal would be unaffected by the variations in its amplitude if a balanced linear demodulator were used. However, the most practical broadband detection device for microwaves is a crystal-video detector whose characteristics at low signal level operate in the square law region with a high figure of merit. The dc bias optimizes this figure of merit by the large increase in current sensitivity relative to the decrease in the video resistance. A balanced power split and phasing of the signals in the discriminator arms is necessary to minimize phase distortion of the detected output.

A balanced detection of the 2-Mc modulation, therefore, produces an output with a phase delay proportional to the range and an amplitude that is dependent on (a) the power radiated, (b) the absorption by the transmission medium, (c) the multipath reinforcements and cancellations, and (d) the changes in the reflection coefficient of the surface. The amplifiers and limiters which follow in the receiver should effectively remove the amplitude flutter from the detected signal to provide a truly phase-sensitive ranging system.

Component specifications require that the match of the hybrid junctions in the power splitter, the recombiner, and the crystal holder be good over the operating bandwidth. This avoids phase pulling of the demodulated video output due to reflections and commensurate amplitude unbalances or nonlinearities as the input frequency is deviated.

If the frequency discriminator is not preceded by rf predetection limiting, then an rf power fade of 10-db results in a video amplitude fade of 20-db (square law detection). Therefore, to suitably recover the phase of the demodulated envelope in the presence of fading, the receiver must accomplish phase-stable amplification and limiting over the dynamic range of the video amplitude fades.

Typical quadrature-type phase bridge detectors are both amplitude and phase sensitive to the error signal. It is therefore important that the demodulated signal applied to the receiver's phase detector input be of constant amplitude when the rf power input to the discriminator changes over at least a 30-db range. This is an equivalent amplitude change of 60 db to the detected 2-Mc signal taken from the junction of the two crystal detectors, assuming they are functional in a truly square law manner.

Amplifier-Limiter

Noise reduction from amplitude fades is accomplished in the amplifier-limiter. Since the phase of the 2-Mc modulation determines the target range, it is important that changes in the amplitude of the demodulated signal, due to fading, do not distort its phase. These amplitude flutters and extraneous noise are effectively removed to produce a constant output amplitude. A bandwidth of 4 kc is used in the amplifiers and limiters in order to maintain the phase stability during the limiting process.

When information sidebands no higher than 15 cps are contained in the 2-Mc envelope, the use of very narrowband amplification, with consequent noise reduction, would be permitted if it were not for the phase distortion accompanying limiting. This occurs when the input and output

impedances of the limiting stages vary with input signal amplitude and detune the bandpass circuits. Stray coupling, excess recovery time, and nonlinear amplification also seriously affects the phase.

Since the range information is carried by the signal phase, it is vital that little or no phase error is self-generated in the receiver. To maintain a phase stability of less than 1 degree over a 60-db limiting range, a series of gated-beam deflection limiters are utilized. They are isolated from 2-Mc tuned amplifiers with neutralized cathode-follower decouplers. Seven of these rather unorthodox stages, each containing an amplifier-follower-limiter, are cascaded to provide constant amplitude with phase stability over the required range. The gated-beam mode of limiting, as suggested in a paper by Wingrove,* helps to minimize impedance shifts with level. This effect is further decreased by swamping with the low-impedance cathode-follower decouplers. Gain is primarily provided by high-transconductance pentode amplifiers.

A sufficient gain is built into the amplifiers to produce self-limiting on system noise, and thus the phase-stable locking range is the primary design criteria. A cascode type of input amplifier is used to minimize the input noise level, maximize the system sensitivity, and, thereby, extend the range of satisfactory operation under minimal signal conditions.

Transmitter

The transmitter consists of a microwave generator that is sinusoidally frequency-modulated from a master quartz crystal reference oscillator and modulator unit. The deviation rate is phase-stabilized by the crystal oscillator, which also provides a 2-Mc phase reference to the phase detector as a basis for determination of propagation delay.

A backward wave oscillator (BWO) (type VA-161), which is capable of being frequency-modulated over the broad bandwidth required, is the microwave generator. This BWO radiates approximately 63 mw of power (+18 dbm) at an X-band frequency of operation. The microwave power is coupled to the transmitting antenna through a ferrite isolator which adds to the stability of the radiation. This particular BWO is frequency sensitive to changes in load impedance, thereby making the antenna decoupling a desirable feature.

The dual balanced detectors of the receiver's discriminator are connected in reversed polarity. The advantage of this scheme over a single-ended detector is that it rejects the phase error caused by a 2-Mc amplitude modulation of the transmitter as its frequency is deviated. In addition, the simultaneous detection of the standing wave pattern at two positions spaced one-quarter wavelength apart subtracts out the AM crystal bias effects due to signal fades and leaves the desired signal with doubled amplitude.

The transmitting antenna is a 30-in. parabola having a forward gain of 32.8 db relative to an isotropic radiator. The parabolic dish has a depth of 5.6 in. and is illuminated by a 15-db, tapered, wideband, Question Mark feed horn at a 10-in. focal length. A similar antenna for reception is mounted nearby, as shown in Fig. 7.

The modulator provides a maximum driving signal of ±20V to the frequency-sensitive elements of the BWO through a pulse-type modulation transformer. The cathode and heater voltages are swept from −230V to −270V at a sinusoidal 2-Mc rate, thereby frequency-modulating the output microwave signal over the deviation bandwidth.

The spectrum of the radiated power contains modulation side frequencies in the form of spectral lines spaced at 2-Mc intervals over a 400-Mc bandwidth from 9200 to 9600 Mc. The frequency deviation ($\Delta f = \pm 200$ Mc), which is maximized at this value, was decreased during the flight evaluation program to determine the bandwidth requirements for satisfactory operation.

A tuning meter, not shown in the block diagram of Fig. 2, is coupled to the transmitter's output through a 20-db directional coupler, a resonant cavity, and a crystal detector. This provides a monitor of relative power output when the BWO is unmodulated and also a means of setting the center frequency of operation to match the waveguide discriminator tuning.

A nulling network, also not shown in Fig. 2, is likewise coupled to the transmitter output through a directional coupler for the purpose of injecting an amplitude and phase-controlled portion of the transmitted output into the receiver's input. This effectively cancels out a portion of the local

*E.R. Wingrove, Jr., "Phase Stable Limiting I-F Amplifiers Using Beam Deflection Tubes," PGMIL-IRE Conf., 4th National Convention on Military Electronics, p. 21, Washington, D.C., June 1960.

leakage within the system to improve the operation under minimal signal return. A remotely tuned 20-db attenuator, used in conjunction with two 20-db directional couplers at each terminus of the network, provides a control of the injection over a −40 to −60 db portion of the BWO power into the receiver's discriminator. Remote phase control over 360 degrees is provided in a dielectric microwave phase shifter. A delay line equivalent to a leakage path of 6 feet of free space connects these elements of the network together.

Phase Detector

A quadrature-type of balanced phase detector is used to compare the phase delay of the 2-Mc modulation obtained from the receiver's amplifier-limiter output with the 2-Mc reference from the crystal oscillator. The phase-detector unit includes a variable phase shifter, a gated-beam phase discriminator, and a dc amplifier, cathode-follower driver.

The gated-beam phase detector effectively functions as a limiting coincident-gating circuit. The error signal from the receiver is applied to one of the quadrature grids of the beam tube and the reference signal is applied to the other, each having sufficient amplitude to produce sharp grid limiting. When the phase of one of the grid signals is shifted relative to the other, the duty cycle of the detector's output is altered. This waveform is integrated in a low-pass filter to produce a dc level that follows the phase changes as the target range is varied. A low-impedance output is obtained from the cathode follower, where it is dc balanced to zero when a quadrature or 90-degree relationship exists between the two input signals.

The detector's output is essentially linear for an equivalent 50-ft wave height variation centered about the dc-balance zero point. A sensitivity of greater than 0.6V/ft is available, which may be attenuated to match the calibration sensitivity of the recording equipment. Amplification of the input signals in the detector unit, as well as harmonic filtering of the error signal, minimizes extraneous output not due to phase shift. Adequate phase stability of the modulation is accomplished by the use of a quartz crystal for 2-Mc frequency control.

A continuously variable phase shifter is provided in the reference channel to balance the bridge output at the desired operational aircraft height, thus giving linear response about this range level. Since the phase delay of the modulation increases 180 degrees for every 123 feet of altitude increase, the phase detector will have ambiguous balance points or zero crossings at these intervals. A profile recorder requires only precise differential range information, and, therefore, this redundancy does not negate its usefulness, although it would complicate its use as an absolute altimeter. The use of the balancing phase shifter permits some choice in the operational height without degrading the profile linearity, although wave resolution does suffer at the higher altitudes due to the divergence of the antenna beam.

The vertical motion of the aircraft causes a slow drift in the detector's output, and it is desirable to minimize this effect. Changes in wave height are not separable from changes in aircraft altitude, except for the rate of change. This is evident on the wave record shown in Fig. 3. The use of a bandpass filter in the profile output reduces this effect and helps to maintain the recording within the limits of the chart. This filter passes range rates that have periods between 0.05 and 10 seconds with a skirt selectivity of 6 db/octave.

Pilot's Meter

An operational aid in the form of a Hi-Lo type of ILS meter is provided near the pilot's instrument panel. It is actuated by the voltage output from the phase detector and indicates corrective altitude information to the pilot during a profiling run. The meter is balanced at its midpoint at the start of a profiling track, when the selected course and altitude of the aircraft has been attained. The variable phase shifter permits a zeroing of the meter at some reference altitude. The radar then supplies a plus or minus voltage to the meter that is proportional to the aircraft's vertical departure from this reference level.

It is necessary to integrate or smooth the wave record from the metering circuit with a low-pass filter to avoid confusing meter flutter. The filter time constant is selected to slow the meter response sufficiently to accommodate the response time of the pilot and aircraft controls. An RC product of 0.25, coupled with the inherent damping of the meter movement, smooths out the sudden altitude

changes or wave swell response from the radar. The sensitivity of the meter was adjusted to indicate over a full-scale altitude range of approximately ±25 feet. The proper use of the pilot's meter enhances altitude stability during a run, which in turn improves the fidelity of the wave record.

Recorder

A Sanborn-type pen recorder having a chart speed of 25 mm/sec was used to trace out the wave profile. The input signal to the recorder was not integrated in the same manner as the pilot's meter but was adjusted to pass all frequencies in their true perspective from 0.1 to 17 cps. The lower corner frequency was provided to filter out some of the aircraft motion and help maintain the record on the chart without readjustment. The upper corner was provided to filter out system noise and interference that may adversely affect linearity of readout.

FIELD TESTS

A series of flight tests of the radar system was made during the development program to determine background level, resolution, bandwidth requirements, and accuracy of reproduction as compared to an instrumented wavestaff.

Specular Reflectivity

Aircraft runs over specular reflecting surfaces such as calm lakes or runways were highly unsatisfactory due to the large dynamic fading range as the aircraft pitched and rolled. Most of these targets, observed for short periods, were subject to thermal air currents that prevented aircraft stability and caused small altitude changes which were inseparable from the wave data in the reflective return. It must be recognized that extreme fading that exceeds the signal limiting capabilities of the system will result in noisy readout. This condition was experienced over mirror-like surfaces when the trim of the aircraft was difficult to maintain due to thermal updrafts. Slight changes in trim tilted the unstabilized antenna from normal, causing the beam axis to move from directly beneath the aircraft. The incident and reflected beams were no longer in coincidence, causing the specular reflector to divert most of the return away from the receiving antenna and resulting in a noise contaminated output from the receiver.

Background Level

Since the ocean air was more stable and generally had enough motion to at least slightly disturb the capillary wavelets of the surface, signal fades were less of a problem. Figure 3 (bottom graph) illustrates a run over a reasonably calm surface inside a cove, immediately followed by a run (upper graph) in a bay where the waves were of higher amplitude. The calm water background shows less than a 6-in. disturbance while the bay shows waves up to 3 feet. The drift in the aircraft motion is quite evident by the 10-sec cyclic period across the chart length on the calm water run. Although not quantitative, these runs do indicate the low background level resolvable.

Bandwidth Versus Range Noise Tests

Subjective flight tests to determine the effect on the fidelity of the profile with changes in deviation bandwidth were performed. To avoid the reduction of receiver input with a decrease in deviation, it is necessary to increase the Q of the discriminator. This is done to optimize the sensitivity of the system by matching the acceptance bandwidth of the receiver to the transmitted bandwidth.

When the length of the differential arm of the frequency discriminator is increased by an integral number of wavelengths, the Q increases and the deviation must be reduced to bring the response within its linear range. This maintains a more constant amplitude of the demodulated signal for a given input power, even though the deviation is altered.

The several settings of frequency deviation which were selected (30, 60, 100, 190, and 250 Mc) were individually used during separate and repeated runs over a given track of ocean. Two mechanical trombone sections, differing by 34 guide wavelengths, were used in the differential arm of the waveguide discriminator. It would

be more desirable to match a particular trombone length to each deviation step used in order to produce a consistent average signal amplitude in the return. However, the practicability of physically changing the differential path length of the discriminator within a short period of time while in flight limited this change to the two lengths.

The discriminator with the longer path length in the trombone section has a narrower linear acceptance range of frequency, which was exceeded when the 190- and 250-Mc deviation was used. The longer section was most useful when the three lower deviation steps were used, and the shorter section was more applicable at the three higher steps. Use of the short differential path with the 30- and 60-Mc deviation was not successful because the average output amplitude faded consistently below the minimum acceptable level of the receiver. However, the overlap between the settings of deviation with respect to the Q of the two detectors assisted in isolating the amplitude effects from the deviation effects to the profile record.

It was shown that with the high-Q detector and the 30-Mc deviation a degradation of the wave record due to noise spikes was evident, even though the recovered signal amplitude was adequately high. The deviation setting of 100 Mc with this detector appeared to optimize the quality of the wave record. When the deviation of 250 Mc was used with the low-Q detector, the record again developed noise even though the signal level and the discriminator acceptance bandwidth was adequate. An explanation of this behavior indicates that, in the 30-Mc deviation case, an insufficient bandwidth failed to suppress the multipath interference effects. In the 250-Mc case, the rise in spurious response may be attributed to a combination of FM and AM effects. Although the increased frequency deviation should have worked as well as the 100-Mc deviation from a pure FM analytical point of view, a superposition of quadrature AM may seriously affect the phase of the detected waveform. The power output from the BWO changes by approximately 15 percent when the frequency is slowly moved over the 250-Mc bandwidth. Unless an amplitude hysteresis or output delay is involved at the 2-Mc frequency sweep rate to produce quadrature AM, this AM should not account for the increased noise. In a like manner, the frequency-selective fading characteristics of the illuminated target area may also produce an amplitude modulation that increases with frequency deviation. If the reflectors differentially modify the phase and amplitude of the side frequencies, then quadrature AM can develop in the return to modify the demodulated phase.

The random character of the sea surface prohibited a definitive evaluation of the optimum frequency deviation. However, sharply defined waves were obtained during these tests and in subsequent flights over ship wakes using the 100-Mc deviation. A lesser deviation did not appear to resolve the waves as clearly, a fact that correlates well with the analytical study.

Profiling Resolution

Additional runs were made over the wakes or disturbances left in the ocean surface by the motion of surface ships. Such a run is illustrated in Fig. 5 where the aircraft course was set at a right angle to the ship's course with the intercept to pass approximately 200 yards aft of the ship's fantail. Timing marks are indicated when the aircraft visually approached the edges of the bow wake

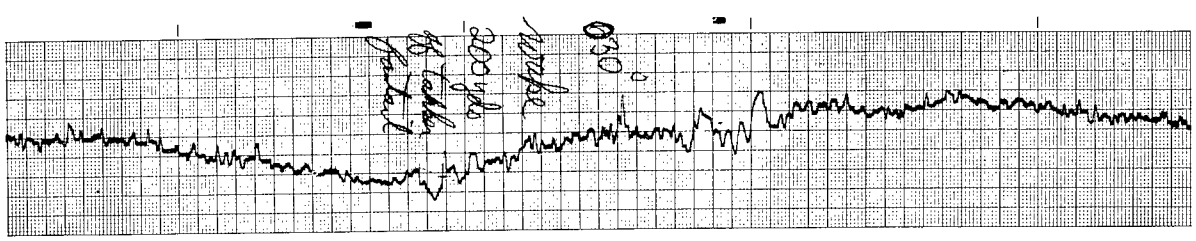

Fig. 5 - Profiling run perpendicular to a ship's wake. The run crossed the wake at 200 yards off the ship's fantail.

left in the water. The sea was quite calm, and a definite delineation of the wake against the background was obtained.

Wavestaff Comparison

Further runs were made over the ocean in the vicinity of Texas Tower No. 4 where an instrumented wavestaff simultaneously recorded the wave motion beneath the tower. Although point-to-point correlation is not possible, spectral distributions of the wave frequencies obtained by the two methods were compared. Figure 6 indicates several aircraft runs on various headings, in addition to the wavestaff frequency distribution obtained from the simultaneous wave measurements.

The frequencies recorded at the top of each chart in Fig. 6 are in real time. The upwind and downwind runs show a major portion of the wave spectral distribution between 1.0 and 2.4 cps. This means that if our aircraft ground speed was 300 ft/sec, the horizontal ocean wavelengths were dominantly between 300 feet and 125 feet. The deep-water gravity wave formula relates the length of the ocean wave to its horizontal velocity in passing a fixed point, such as a wavestaff. Waves of these respective lengths should show a computed frequency distribution between 0.13 cps and 0.2 cps, which correlates reasonably well with the wavestaff analysis. The diagonal and crosswind runs show a blending of the large amplitude spike due to the aircraft motion with the wave motion. To properly account for the aircraft motion, even after filtering, it was necessary to reduce the amplitude of the wave spectra. Therefore, a better method for the removal of the aircraft motion, such as a doubly-integrated accelerometer, is recommended.

INSTALLATION

The equipment installed aboard the aircraft was typical laboratory equipment and developmental units. The system was designed primarily to prove feasibility, and no major attempt was made to compress the volume of the functional radar. Additional instrumentation was also used to evaluate the operation, thus making the equipment quite cumbersome. The antenna rack illustrated in Fig. 7 shows two dishes mounted within the bomb-bay section of the aircraft. Reflections within the bomb bay from the open doors and bulkheads undoubtedly add to the feedthrough problem, making nulling necessary in this case. Greater spacing between antennas, or operation as a higher microwave frequency, should diminish this problem.

CONCLUSION

Low-level operation of a microwave system utilizing pencil beam illumination can obtain sufficient resolution of the ocean contour to plot the profile of the wave structure in excess of a 6-in. wave amplitude. Quantitative measurements are extremely difficult due to the lack of control over comparison measurements. The diverse nature of the surface reflectivity requires estimates of system parameters, based on random scatterers, which for different sea states are not exact. However, qualitative comparisons with visual observations of the sea on numerous runs show good correlation.

The fidelity of the system is somewhat degraded in the presence of high wave slopes that are not resolved by the spot size of illumination. Also, specular or mirror-like reflecting surfaces make it difficult to accommodate the dynamic fades due to tilting of the antenna beam.

The effectiveness of an airborne system to rapidly survey the existing sea state is realized. Acceptable precision in the reproduction of significant wave shapes by these radar techniques appears practical, and further refinements to an operating prototype unit are under way.

ACKNOWLEDGMENTS

Acknowledgment for contributions to the progress of this program is extended to my colleagues in the Radar Division, with specific reference to Arno M. King and Chauncey G. Myers who materially assisted in the developmental and flight test phase. Gratitude is also extended to the operation personnel of the Naval Air Development Unit of South Weymouth,

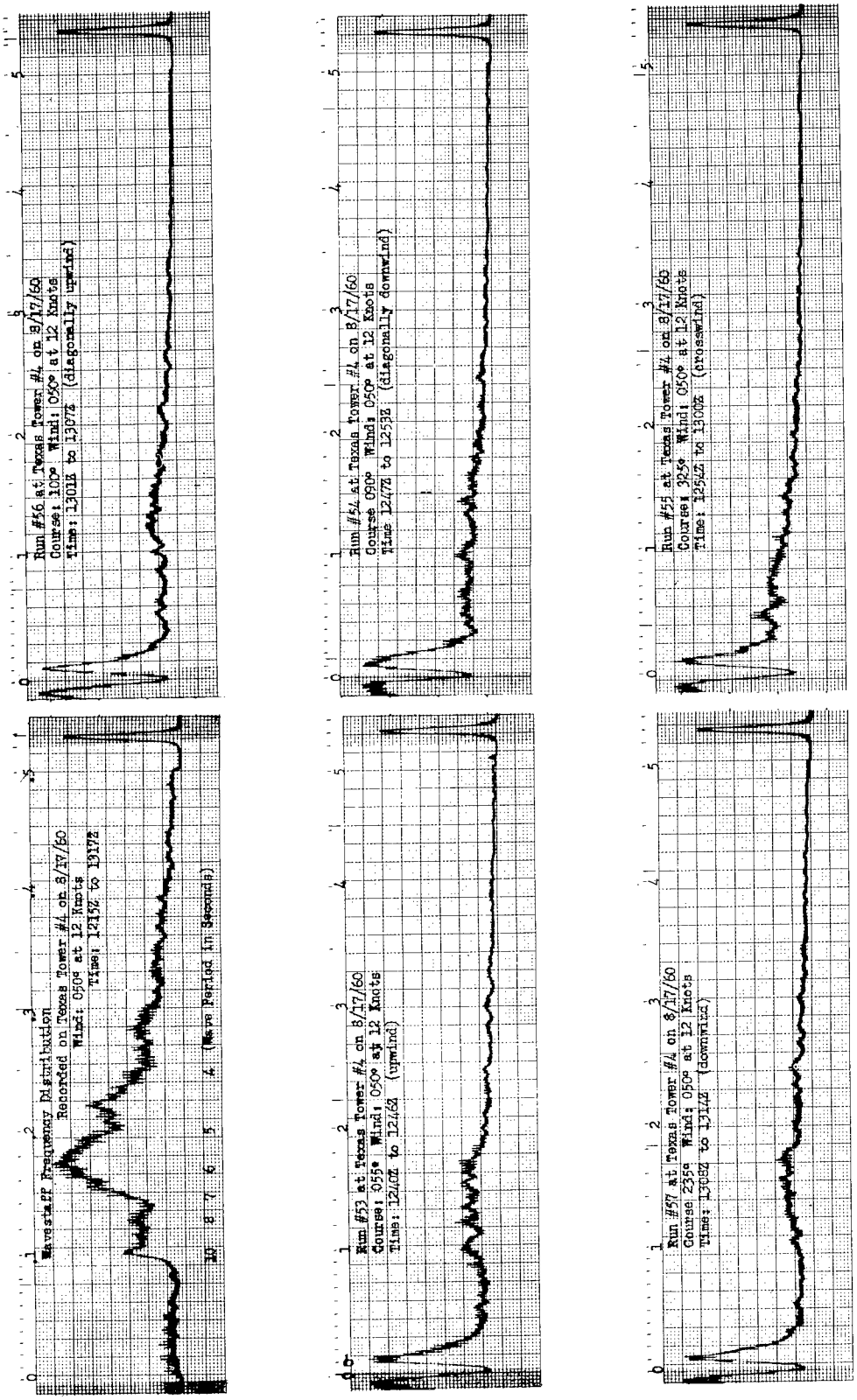

Fig. 6 - Comparison of wavestaff measurement of ocean-wave frequency distribution with simultaneous aircraft measurements of the same. The aircraft runs were made over a period of 34 minutes, with a 1-minute interval between each 6-minute run.

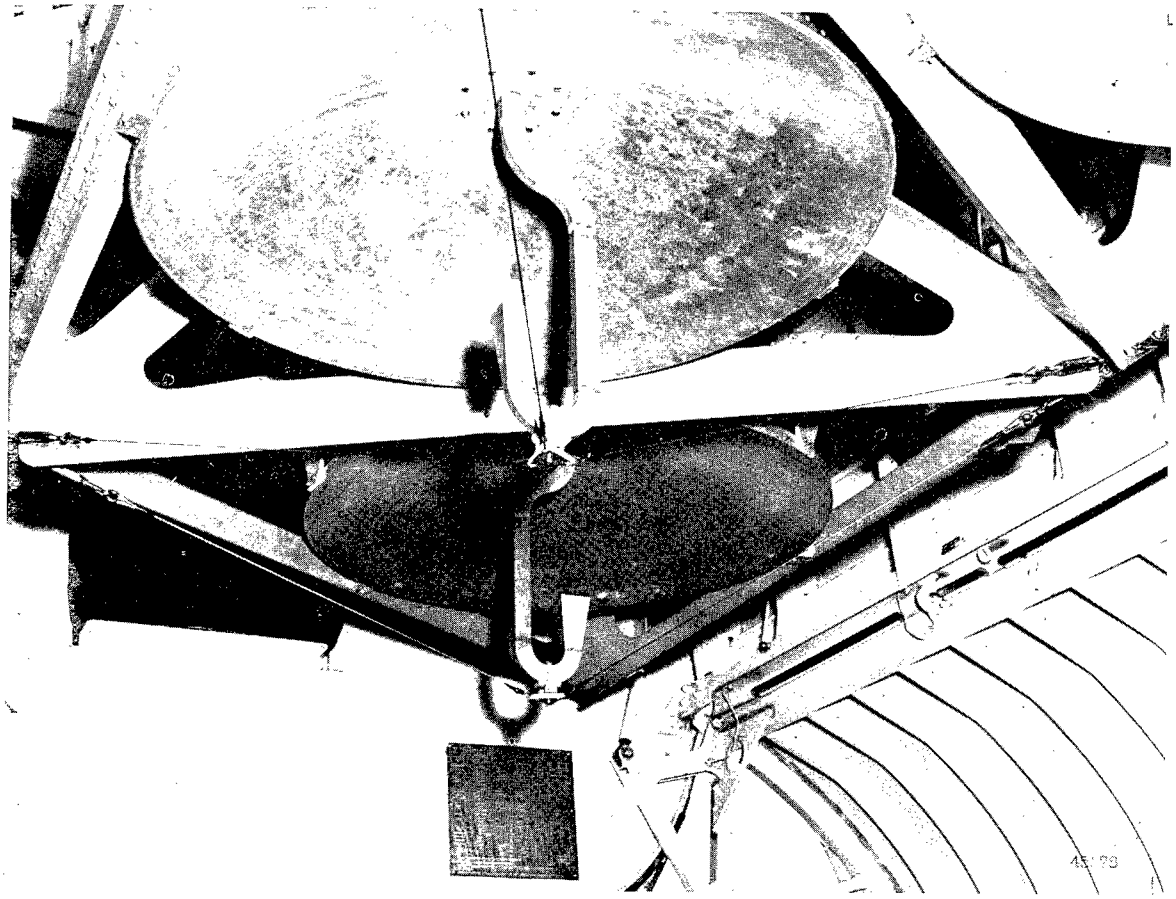

Fig. 7 - Transmitting and receiving radars mounted in bomb-bay section of aircraft

Massachusetts, who provided the flight test facilities and assistance. Valuable assistance from the Oceanography Prediction Division of the U.S. Naval Oceanographic Office was provided to man the Texas Tower facility and assist in data analysis.

APPENDIX
OPERATIONAL LIMITATIONS AS DETERMINED FROM SYSTEM PARAMETERS

A phase-sensitive frequency modulation system determines range either by directly heterodyning or beating the modulated radio frequency of the return with the outgoing signal, or by demodulating the return prior to phase comparison. The first system is not seriously affected by local leakage but, unless a rather complicated processing technique is used, it suffers from appreciable error. Phase detection of the demodulated signal, as used in the subject radar system, requires a high isolation of the local signal to avoid seriously limiting the operating altitude.

Decoupling of Leakage

In order to determine the extent to which the leakage of the local signal, or feedthrough, influences the maximum useful range of the system,

it is necessary to compute losses from a knowledge of the circuit parameters. The major sources of feedthrough are antenna spillover and side-lobe radiation which feed directly from the transmitting antenna into the receiving antenna. Additional feedthrough occurs when some of this sidelobe energy, which would normally not leak into the receiver over a direct path, is bounced off a portion of the aircraft and then enters the receiver. Waveguide and coaxial leakage, as well as crosstalk of the modulation signal in the post-detection portion of the receiver, may also be lumped into the local leakage problem.

Typical decoupling ratios for adjacent mounted antennas is between 50 and 60 db at X-band. At K-band this decoupling may be as high as 70 db; at L-band it may be as low as 20-db for a similar type radiation structure. When the round-trip transmission loss approaches the decoupling ratio, range information is contaminated. This fact occurs regardless of the amount of power radiated by the transmitter, provided of course that a sufficient amount of energy is returned to the receiver to exceed the minimum sensitivity threshold of the detector.

If a single antenna is used for both transmission and reception, a directional selective circulator must be used to accommodate the continuous operating mode. Typical state-of-the-art ferrite circulators have an isolation of less than 40 db and, therefore, single antenna operation with the subject system would accentuate the decoupling problem.

Antennas

The isolation that is accomplished with dual antennas is dependent on the distance separating the radiators, their directivity or beamwidth, the illumination taper of the feed horns, reflections from adjacent parts of the aircraft, local shielding by intervening objects, and the F/D ratio of the parabolic dishes. This F/D ratio of the focal distance with respect to the dish diameter determines the curvature of the parabolic section used as a reflecting aperture. This influences the sidelobe structure, the beamwidth, and the spillover or back radiation. Most of these factors are interrelated, and a compromise must be made to attain the optimum design. The choice of mounting position of the antennas within the bomb-bay section of the aircraft restricts both the antenna size and the distance separating the dishes.

The desired wave resolution and minimum altitude restrictions require a narrow beam of approximately 3 degrees to be selected. The 30-in. dishes chosen to produce this beam at X-band, have an F/D ratio of one-third, and the feed horn illumination pattern was tapered by 15 db across the aperture in an effort to further reduce spillover. Antenna beamwidth is essentially linearly related to the radiation wavelength; thus a smaller dish will produce the same beam if the operating frequency is proportionally increased.

In a system not sensitive to local coupling, or where a common antenna is employed, a 10-db taper of the feed pattern is typically used for optimization of forward gain with respect to spillover and sidelobe loss. However, in the present case, a sacrifice in forward gain and a slightly broader beam produced greater decoupling, with lowered sidelobes, in an effort to improve the short-range operational stability. This concentrates more of the available power in the main forward lobe of the beam. Use of a paraboloid with a deeper cross section or a smaller F/D ratio accomplishes this same net result in an almost analogous manner.

Increasing the feed horn taper from 10 to 15 db changes the respective horn beamwidth from 72.5 to 68.5 degrees. Therefore, less of the energy is reflected by the area near the outer edge of the dishes and so the effective area of the parabola is reduced.

To compute the efficiency of forward radiation for the antenna used, the antenna pattern measured in the far field was projected to an annular ring power-density plot. Incremental area summation of this plot shows a power distribution between the main lobe and side lobes of 1 to 0.11. This indicates a forward efficiency of 89 percent.

Radiated and Reflected Power

The average power output of the backward wave oscillator (BWO) used for the transmitter is 63 mw, of which 89 percent or 56 mw is radiated forward in the main lobe. If this 56 mw (+17.5 dbm) is now isotropically reradiated from the reflecting surface over 2π steradians, then the power captured on reception is diminished by the ratio of the effective receptor's area to the area

over which the reflected power is hemispherically distributed at the operational altitude. A hemisphere with a radius of 125 feet has a surface area of 98,175 square feet. The measured effective area of the receiving antenna is 34 percent of its projected area. Illumination taper is the major factor affecting this parameter. This value is obtained from the absolute measured gain (32.8 db) of the antenna as it relates to the theoretical area gain (37.5 db) from $4\pi/\lambda^2$, which assumes a square illumination of the projected area. The −4.7-db difference (above isotropic) between these two gain ratios is equivalent to a ratio of 0.34:1 – hence the 34 percent effectiveness. Therefore, the 30-in. dish, with a projected area of 4.91 square feet, has an effective capture area of 1.67 square feet. A ratio of 1.67 to 98,175 indicates that the captured power will be 47 db below the reradiated or reflected power.

Propagation and Reflectivity Losses

The theory that the ocean surface is a "perfectly rough" scatterer is not wholly substantiated by field measurements* which show a repeated fixed degree of specularity, combined in the surface return, that is independent of the sea state. The observations indicate that the microwave echo from a surface illuminated at normal incidence is 5 db greater than the echo which would be obtained from a perfect hemispheric scatterer. On this basis the transmission loss will therefore be 5 db less than the computed −47 db, resulting in a captured power that will be −42 db less than the incident power. A high degree of specularity in the return is substantiated in an independent study.†

Radar transmission losses due to absorption by gases (primarily oxygen and water vapor) and scattering by suspended particles (precipitation and dust) are neglected in these computations. Such losses are inconsequential at X-band frequencies over a short path length where visual contact is established. Longer range operation at shorter wavelengths or in an atmosphere of heavy precipitation may require further consideration of these factors.

Signal amplitude changes, due to changes in the contour of the ocean surface, must also be considered. "The range of fading (defined here as the range between the level below which only 5 percent of the return power is found and the level below which 95 per cent is found) varied from about 3 db, for some smooth water surfaces exhibiting principally specular returns, to about 19 db for certain nonhomogeneous target areas...the theoretical range of fading‡ for a randomly scattering ground for which the signal amplitudes are Rayleigh-distributed is 17.5 db." This means that interference patterns from the random scatterers will cause 5.5-db increases and 12-db decreases from the average returned power based on a 5-percent distribution criteria. Since we are interested in determining the receiver's sensitivity requirement at minimum signal, adding this −12 db to our computed loss of −42 db gives a total path loss of −54 db that will only be exceeded less than 5 percent of the time. This loss should be independent of the sea state, assuming that the scattering criteria at 125 feet is modified by 5 db due to specularity. Therefore a computed minimum power level of −36.5 dbm is received at the input of the waveguide discriminator during most fades. This is obtained by subtracting the total path loss of 54 db from the +17.5 dbm of power (56 mw) reflected from the surface. This is further reduced by 3 db in the power splitting hybrid of the discriminator to give a level of −39.5 dbm at the crystal detectors. On this basis the system sensitivity is not limited by the minimum sensitivity threshold of the crystals, which is computed to be at a level of −68 dbm.

Detector Minimum Sensitivity

The selected MA-418AM crystal detectors are rated at a tangential minimum sensitivity of −51 dbm, based on a 10-Mc bandwidth. This amount of power will be detected with a signal-to-noise ratio of 8 db above unity. The tangential rating is based on a given noise bandwidth and is inversely proportional to the square root of the bandwidth for a given S/N ratio.

*G.G. MacFarlane and H. Davies, "Radar Echoes from the Sea Surface at Centimetre Wave-Lengths," *Proc. Phys. Soc. (London)* **58**:717 (1946).

†M.L. Meeks, C.H. Wilcox, and N.A. Logan, "Reflection Properties of Radar Targets," Georgia Institute of Technology, Engineering Experiment Station, p. 37, Section 6.3.5 of Supplement to report, 1954 (Conf. Report and Conf. Suppl.).

‡A.R. Edison, R.K. Moore, and B.D. Warner, "Radar Terrain Return Measured at Near-Vertical Incidence," IRE Trans. Antennas Propagation, AP-8(No. 3):246 (1960).

If the crystals are connected to amplifiers or detectors that change the noise bandwidth, the sensitivity threshold is thus modified. The waveguide frequency discriminator of the system is coupled to the amplifier-limiters, which have an overall bandwidth of 4 kc. This 2500:1 bandwidth reduction from 10 Mc to 4 kc provides a 50:1 sensitivity improvement to the threshold. This 50:1 ratio extends the −51 dbm minimum sensitivity of the crystals by 17 db, that is, to −68 dbm.

The computed received power level of −39.5 dbm at the crystals appears to make the above sensitivity improvement inconsequential. However, if the aircraft altitude ware quadrupled (1:4) to a 500-ft altitude, an additional range-squared-dependent loss (16:1, or 12 db) would exceed the crystal rating at the 10-Mc bandwidth.

Phase Distortion Due to Leakage

The recording system is responsive to the phase of the strongest coherent signal appearing at the output of the amplifier-limiters. This means that when the propagation loss becomes comparable to the decoupling loss, the vector addition of the modulation components of the leakage with the echo may cause the limited signal applied to the phase bridge to shift in phase. The extent of this recorded range distortion depends on the relative amplitude and phase of the combining signals. For example, a fading interval which occurs up to 5 percent of the time results in a computed path loss that equals or exceeds 54 db. During this interval the leakage power may be as much as 2 db higher than the echo power. (This result is based on a local leakage measurement of the system, under flight test conditions, which showed an antenna decoupling loss of 52 db.) The momentary control of the error phase by the leakage modulation during the deeper fades may seriously affect the noise level and fidelity of the profile. Minimization of the leakage to a level that is more consistently exceeded by the echo is therefore desirable. It may also be shown that if an operational altitude is chosen whereby the mean modulation phase of the echo is either in-phase or out-of-phase with the leakage, the range distortion to the profile during severe fades will be less severe. Operation at an altitude of 123 feet (or multiples thereof) should introduce this desired transmission delay.

Leakage Reduction by Nulling

The distorting effect due to leakage was minimized during the flight test phase by the installation of a secondary microwave path over which an equal amplitude, out-of-phase signal could be coupled into the receiver to cancel at least a portion of the local leakage. The parameters of this path were adjusted for optimum cancellation by observing a minimum tuning point in the receiver's detected output when the echo signal signal was essentially absent. Such a condition was obtained by operation of the aircraft in a steep bank at a high altitude, where the strength of the signal reflected from the water is minimized. It was found that 52 db of attenuation inserted in the out-of-phase secondary nulling path, between the transmitters output (+18 dbm) and the receiver's input due to leakage (−34 dbm), reduced the detected local signal by at least 5:1, or 14 db, when properly phased for cancellation. Therefore, this nulling will reduce the received minimum power limit from −34 dbm to −48 dbm. Since the computed minimum of −36.5 dbm of echo power is about 12 db (actually, 11.5 db) above this new leakage threshold, an ample margin of satisfactory operation should exist. In fact, this 12-db margin should permit a quadrupling of the operational altitude and still meet the 5-percent fading criteria. This local leakage does not suddenly cause the system to fail when the path loss exceeds a given value of attenuation, but instead it vectorially adds the leakage to the reflected signal. This causes a gradual deterioration of the response to range changes, and flight tests show an adverse effect to the fidelity of the profile at altitudes in excess of 400 feet. Firm predetection limiting should substantiate the computed operable altitude threshold of 500 feet.

Several flight tests were made using two cascaded travelling wave tubes as predetection amplifier-limiters. This consisted of a type HA-20W driving a type HA-20J tube whose output was linked to the waveguide discriminator. An isolating pad of 13 db is used between the tubes to stabilize the limiting characteristics. Measurements on this combination show a stable output of +6.5 dbm over a power input range from 0 dbm to −25 dbm. No dramatic improvement to the profile fidelity was observed when operating at an altitude of 125 feet. The additional micro-

wave amplification assisted the high-altitude nulling procedure, but since the minimum signal sensitivity of the crystal detectors was not involved, no noise reduction was expected. Signal reduction to a level less than the −25 dbm required for firm limiting appeared to occur consistently and, therefore, an evaluation of predetection versus postdetection was not practical during the flights without a major system change.

The optimization by nulling is not absolute since it is not practical to null out the effect of many possible local feedthrough paths with a single, secondary, closed, cancellation loop. However, a broad minimum in the null was attained when a secondary path was used whose length was equivalent to about 6-ft of free space. The assumption may therefore be made that the majority of the local leakage between antennas travels over a 6 ft path between the points where the secondary path is coupled into the waveguide structure. The equaling of the two paths (plus or minus a half rf wavelength) is important in maintaining phase coherence of the FM sidebands for optimization of cancellation over the full rf bandwidth. Vibration of airframe reflectors in the free path of local leakage likewise reduces the cancellation effectiveness.

Conclusions and Recommendations

Antenna decoupling seriously affects the noise level of the return under marginal operating conditions. If the local leakage problem could be reduced by improvement of the decoupling ratio between the antennas to approximately 70 db, nulling would not be necessary. This might be accomplished in a combination of ways, such as increasing the operating frequency, increasing the spatial separation between the antennas, use of an isolating septum between antennas, or by shading the side-lobe pattern of radiation. This, of course, does not exhaust all means for improvement of hardware; the major function of this project is to investigate a concept rather than to refine operational equipment.