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MASSACHUSETTS INSTITUTE OF TECHNOLOGY

LINCOLN LABORATORY

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A POLARIZATION-DIVERSITY SIMULTANEOUS-LOBING ANGLE-TRACKING RECEIVER

W. B. Renhult

31 August 1961

The work reported in this document was performed at Lincoln Laboratory, a center for research operated by Massachusetts Institute of Technology; this work was supported by the U. S. National Aeronautics and Space Administration under Contract NAS1-229.

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CONTENTS

ABSTRACT

This report describes a simultaneous-lobing angle-tracking receiver operating in the 225-260 Mcps telemetry band and employing polarization diversity. Its operation is considered primarily in the context of the Mercury range and tracking of the Mercury capsule.

Several methods of providing diversity are briefly considered, and a number of ways of implementing the phase shifts required at one polarization for coherent signal addition are discussed.

A prototype receiver is briefly described although circuitry which may be somewhat novel is covered in greater detail. No attempt has been made to include all of the sophistication one might expect in a receiver of this type; circuits have been simplified in some areas where, for example, a manual control can replace an automatic function and reduce complexity.

Some conclusions are drawn as to how this receiver might perform in the Mercury environment.

Introduction

The use of diversity reception to combat fading and improve signal-to-noise ratio in radio communication is nearly as old as radio itself. It may take many forms. One of the most obvious is time diversity, which in its simplest form consists of sending the same message twice. Space diversity takes advantage of the lack of correlation between fades over separate paths and might take the form of two receivers at separated locations receiving the same message. More elaborate is frequency diversity, in which two transmitters operating at different frequencies transmit the same message. Still another method is polarization diversity, which uses the lack of correlation between fades for orthogonally polarized antennas.

For some applications it is not essential that a complex signal-combining scheme be employed. For communication systems using space diversity, one might merely compare the messages received at the two sites and fill in any blanks or correct any obvious errors. However, for a tracking device with a high information rate and a necessarily rapid response time, it is clear that any signal selection or combining must be automatic and fast. In general, there are three approaches*. Selection diversity automatically chooses the best signal of several channels, rejecting all others. Maximal-ratio diversity achieves maximum output ratio when the gain in each channel is proportional to the RMS signal and inversely proportional to the mean square noise.

Equal-gain diversity, as the name implies, is one in which all channels maintain equal gain with their outputs simply added. In practice, the

D. G. Brennan, Linear Diversity Combining Techniques, Proc. of the IRE, pp. 1075-1101, June 1959 equal gain performs better than selective diversity and almost as well as maximal-ratio systems . Since the equal-gain system is much more simple to implement than maximal-ratio, this was the type chosen.

Since the system we are discussing is a tracking device, in which the phases of signals in the azimuth and elevation error channels relative to the phase of the signal in the sum channel determine the "direction" of the error voltages and hence the direction of antenna correction, it is clear that predetection summing of signals at the two polarizations is required and that the summed signals must be phase coherent. It is also implicit that the relative phases of the signals in the three channels from the antenna to their ultimate destination at a phase detector must be held within close tolerances. Conceivably in some devices the differential phase shift between channels would be constant and could be initially adjusted for. In general, however, some automatic phase-adjusting circuitry is required to insure that coherence is maintained.

One type is a form of phase-locked loop, presumably that used by Altman and Sichak. It employs two local oscillators, one for each of the two channels to be summed. The phase of one is controlled by the output of a phase discriminator connected between the channels; this output varies the bias and hence the oscillator-tube capacity via the Miller effect.

A variation on this theme has been used at Lincoln Laboratory, in which a single local oscillator is used but with a voltage-variable delay line inserted between the oscillator and mixer in one channel. This delay line is then varied by the output of a phase detector connected

*F.S. Altman and W. Sichak, A Simplified Communication System for Beyond-the-Horizon Links, IRE Trans. on Communication Systems, Vol. CS-4, pp 50-55, March 1956.

between the two channels. This somewhat simplifies the equipment and removes the danger that the two oscillators might phase lock from mutual coupling.

However, in the system that we are considering, there are not two but six channels that must be combined in pairs. These are the sum and the azimuth and elevation error channels for both polarizations. If all signals were brought separately to some intermediate frequency and there phase-corrected and summed, six complete receivers would be required. The phase correction at the intermediate frequency could be by either of the two schemes just mentioned. It must be remembered that the four error channels during track contain signals that are some 20-40 db down relative to the sum, and hence, in all likelihood, would not be useful signals to compare in a phase detector; that is, phase comparison and correction could not be done separately for each pair of channels. The probable approach would be similar to that shown in Fig. 1, in which the two sum signals are compared and an error voltage developed to delay one of them. This same error voltage must then be applied to the other two delay lines, which implies that they must have the same delay versus voltage characteristics. This may be difficult, though clearly not impossible, to do within usable tolerances.

An alternate approach is to do the phase correcting at the carrier frequency. In general, as shown in Fig. 2, this permits the use of only four channels since the azimuth and elevation signals may immediately be combined in hybrids after phase correction. In addition to reducing the amount of required equipment, it also eases somewhat the problem of getting a number of channels to track in phase as input frequency varies or as AGC is applied.

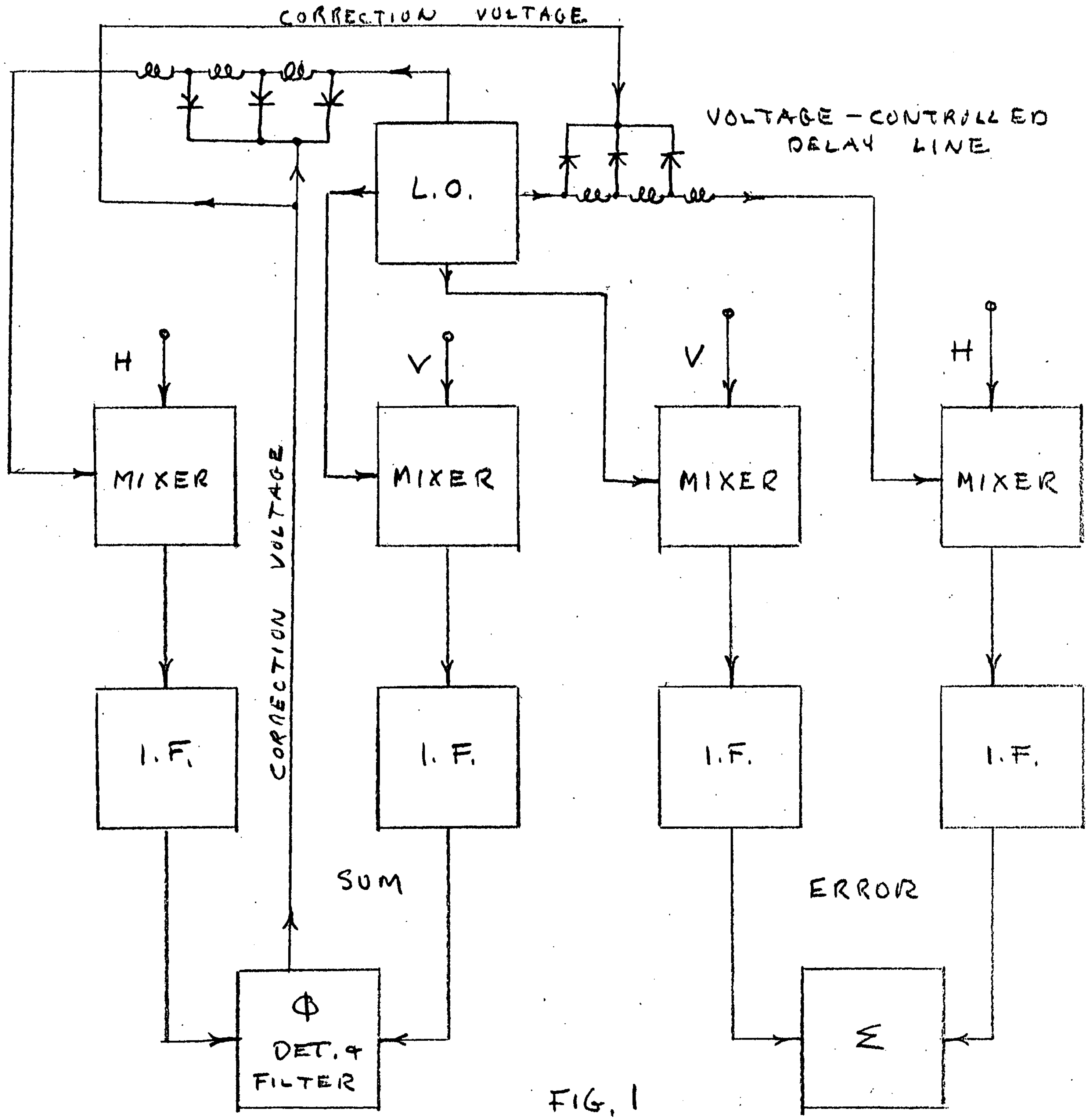


FIG. 1

PHASE CORRECTION
AT I.F.
(SINGLE ERROR CHANNEL SHOWN)

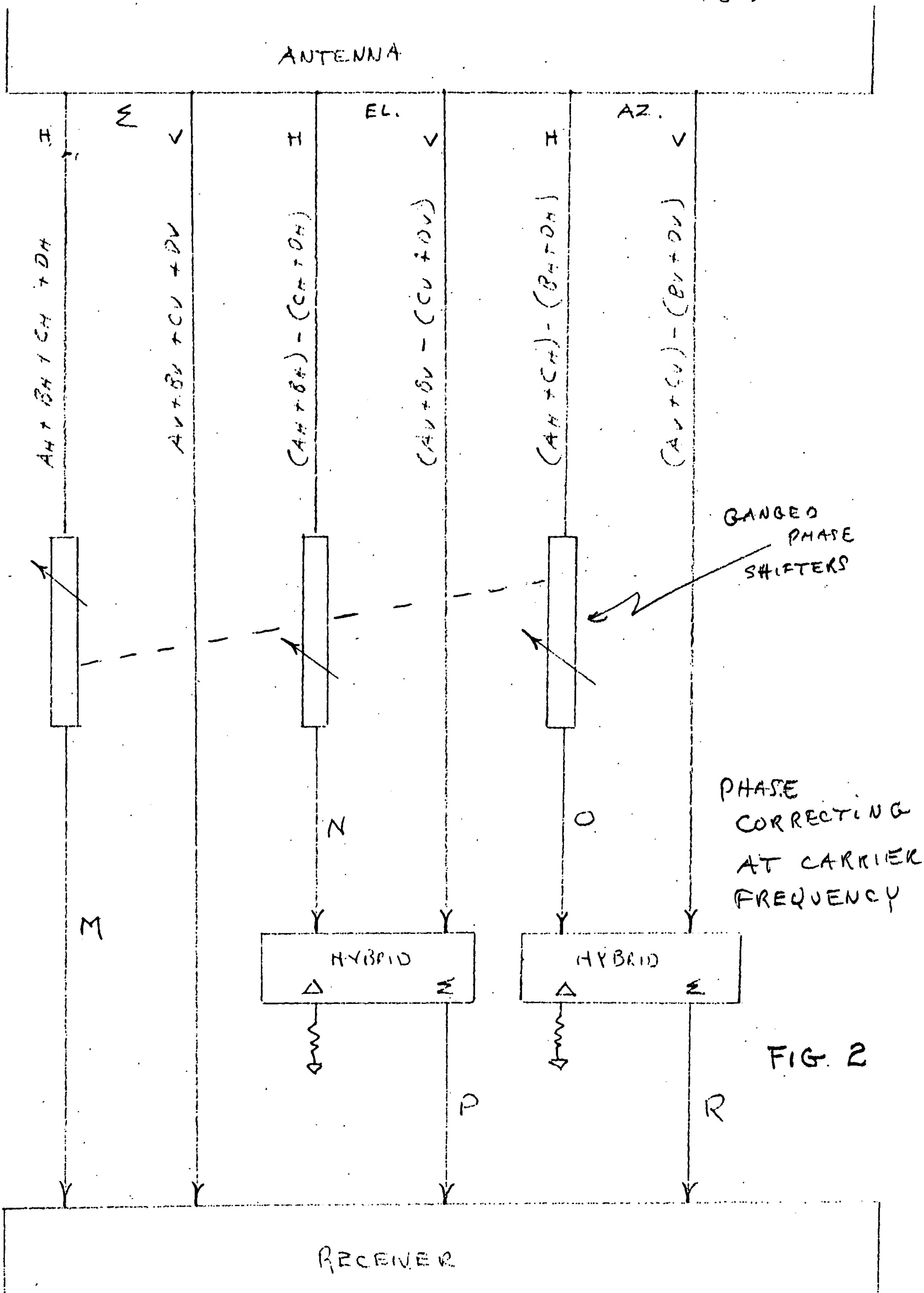

ANTENNA
Σ
EL.
AZ.
H
V
H
V
H
V
$A_H + B_H + C_H + D_H$
$A_V + B_V + C_V + D_V$
$(A_H + B_H) - (C_H + D_H)$
$(A_V + B_V) - (C_V + D_V)$
$(A_H + C_H) - (B_H + D_H)$
$(A_V + C_V) - (B_V + D_V)$
GANGED PHASE SHIFTERS
PHASE CORRECTING AT CARRIER FREQUENCY
M
N
O
HYBRID
Δ
Σ
HYBRID
Δ
Σ
P
R
RECEIVER

FIG. 2

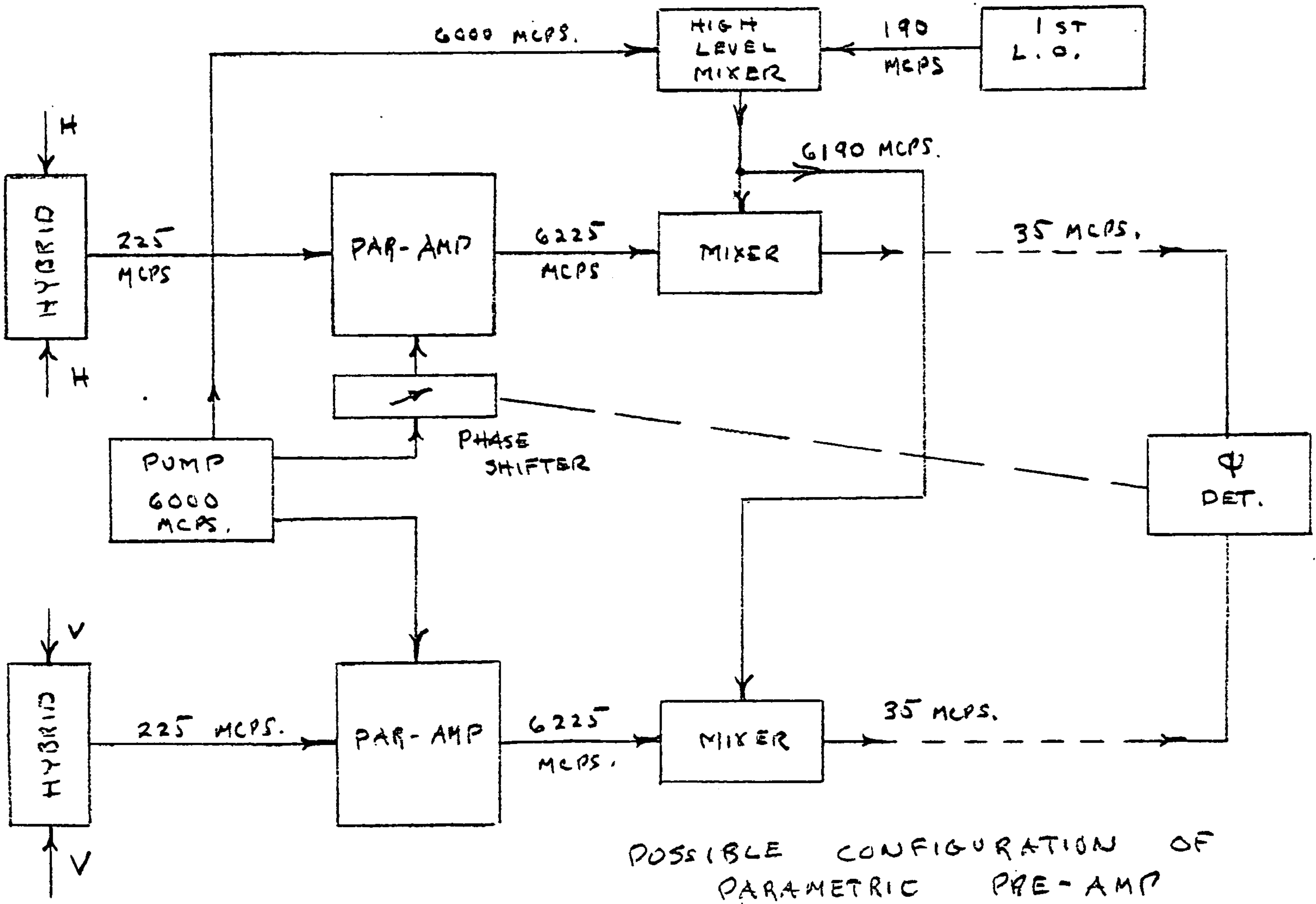
By putting the phase shifters ahead of the RF amplifiers in the signal channels, one is immediately constrained to employ a low-insertion-loss device in order not to degrade the over-all noise figure of the system. Then, too, at this wavelength almost any scheme gets bulky, and attractive methods of phase shifting in waveguide are not feasible.

A Discussion of Methods of Phase Shifting

Several approaches were considered. One suggestion was to use voltage-variable delay lines such as were employed at the lower frequencies. After several experimental models were constructed, it was decided that the practical difficulties of constructing a low-loss line, in getting three to track, and in keeping the characteristic impedance within reasonable bounds as the capacitors were varied to give delay made this method unattractive.

A second solution that was investigated was to use ferrite phase shifters. Telephone conversations with several vendors indicated that such a device was not a shelf item, that it would require a developmental program, but that it was not beyond the state of the art. Guarded replies as to insertion loss made it appear that it might be high enough to negate any improvement in signal-to-noise ratio that one might obtain through diversity. It also appeared that the driving power supplies might become prohibitively large if rapid response time were demanded. This approach was somewhat reluctantly abandoned.

Another method, also based on the use of ferrites but predicated on parametrics as the preamplifiers is shown in simplified form in Fig. 3. By the use of a parametric amplifier, it is possible to put the phase shifters in the pump arms. Two advantages accrue from this configuration.



POSSIBLE CONFIGURATION OF PARAMETRIC PRE-AMP SHOWING SUM CHANNEL WITH PHASE SHIFTER IN ONE PUMP ARM

FIG. 3

By putting the phase shifter in what is analogous to a local-oscillator arm, any insertion loss it may have is kept out of the signal channels; and since the pump operates at a very short wavelength, it is possible to use less cumbersome equipment. A ferrite is attractive because of its response time at this frequency, but mechanical types in guide are also feasible at this frequency. As in the case of correction at the intermediate frequency, the two phase shifters in the error channels would have to be ganged to the sum-channel phase shifter or operate on the output of the sum-channel phase detector.

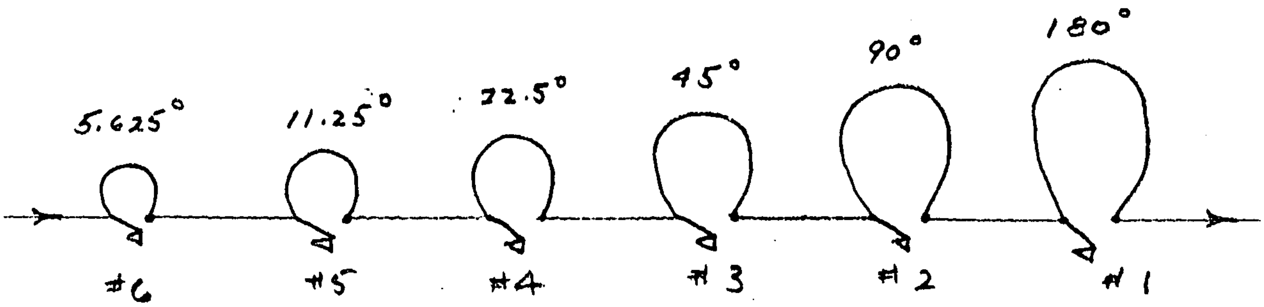
However, here again, six preamplifiers are required. Because of this fact, because of availability and time of delivery, and because of some forebodings as to instability and differential phase shifts that might occur between the six preamplifiers, this idea was discarded, at least for the present*.

The most direct way to produce a low-loss shift in phase at these frequencies is by a simple coaxial line stretcher. Mechanically, it lends itself to ganging, and close tolerances in phase shift among three such lines are easily obtained and maintained. Although large in size they are not impossibly so when folded and may be had in a reasonable package. They do suffer from the fault that they are mechanical devices and move in a finite length of time.

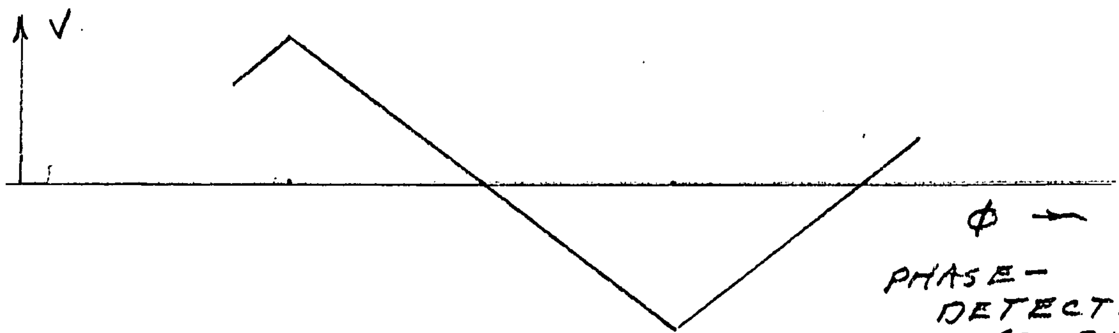
Still another possibility is the use of a digital phase shifter. In its simplest form, this might consist of six lengths of cable arranged as shown in Fig 4a with facilities for switching any cable in or out of the circuit. If the length of the first cable corresponds to a phase lag of 5.625 degrees, the second, 2 x 5.625 degrees, and so forth, the

*Such an amplifier has recently appeared in the literature. ELECTRONICS of 10 March 1961 pg. 170 describes a preamplifier operating in this frequency band.

FIG. 4-A

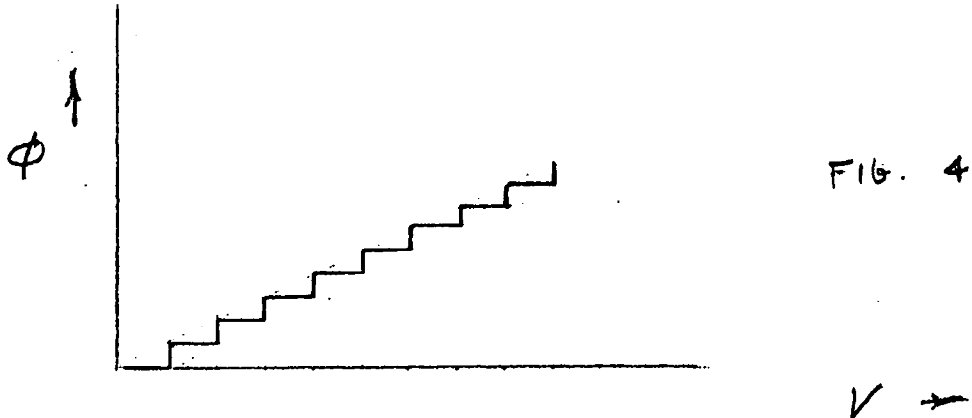


ARRANGEMENT OF COAXIAL CABLES AND SWITCHES TO PRODUCE 360° OF DELAY IN 5.625° STEPS



PHASE-DETECTOR RESPONSE

FIG. 4-B



PHASE-SHIFTER RESPONSE

resultant is a six-binary-bit with 64 available phase shifts to 360 degrees as below:

6	5	4	3	2	1	Phase Shift (degrees)
0	0	0	0	0	1	5.625
0	0	0	0	1	0	11.25
0	0	0	0	1	1	16.875
0	0	0	1	0	0	22.5
0	0	0	1	0	1	28.125

and so forth

While such a line may easily be programmed, there are certain problems which are encountered in making it part of a closed loop. The gain of such a closed loop is determined by the volts per radian which the phase detector produces and the radians per volt which the phase-shifting device produces. Somewhere in the system the analog output of the phase detector must be converted to digital information. We can imagine a box with unspecified contents which accepts a voltage from the phase detector and delivers on six wires some 64 arrangements of 1's and 0's. If the entire swing of the phase detector output is used to produce these 64 discrete phase shifts as would seem reasonable then the converter does not enter into the gain calculation. Since we have postulated 360 degrees of total shift the overall gain (although it varies from zero to infinity) is 360 degrees out for 180 degrees in.

The effect of such low gain is, of course, to permit large phase errors to exist. This may be seen intuitively by reference to Fig. 4b and the following discussion. The characteristic linear-phase-detector transfer characteristic is shown; a certain phase difference produces a certain voltage. This voltage is then translated into a binary code corresponding to the phase correction required. Imagine a step function input in phase error producing some output voltage which immediately produces the proper phase correction. At this point the phase-detector voltage returns to its original value which is promptly translated into a new phase correction and so forth. In practice we would expect the loop to sequentially shift phase to some compromise value which is characteristic of a low-gain loop. In contrasting this device with the voltage-controlled delay line we find that a delay line can be built with many radians per volt of gain so keeping the residual error small. In order to make an equivalent gain increase the total length of the line must be made much greater leading to a many-binary-bit which is undesirable. Thus additional d.c. amplification must be provided after the phase detector.

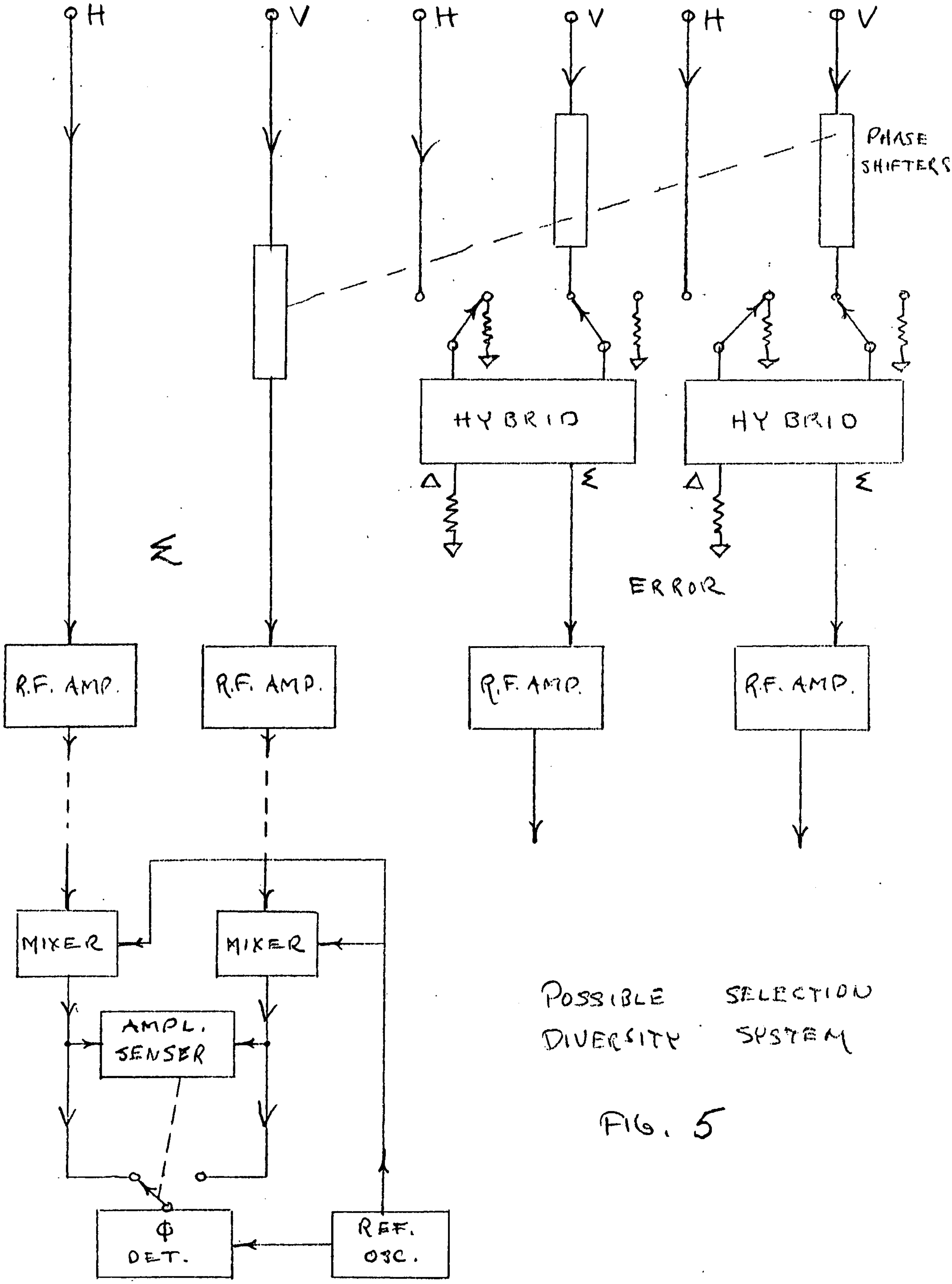
A convenient way to provide both the required additional gain and a form of analog-to-digital conversion would be to use a servo motor driven by the phase-detector output which in turn operated a 64-segment commutator. (In the limit, if the steps were made small enough, such a system approaches the motor-driven linear phase shifter.) The system is only quasi-stable since in general an error voltage always exists and the motor would oscillate between adjoining segments unless a threshold were provided.

If the motor-commutator were replaced by all-electronic circuitry with faster response time the switches shown on each cable could be replaced by diodes. The principal objection to this configuration is that it degrades the overall noise figure. The factors include the attenuation of the cable, the ohmic losses in the diodes and the excess noise generated by the conducting diodes. This material is covered in greater detail in Lincoln Laboratory Technical Report No. 228, Phased Array Radar Studies in which similar devices are discussed starting on pg. 77.

Although we have been discussing the various approaches within the context of equal-gain signal combining it is possible that in some field situations selection diversity might prove to be the most desirable solution. While it is sometimes convenient to think of a plane wave as made up of two circularly polarized signals with some phase relationship and rotating in opposite directions, for the receiving antenna we are using it is simpler to consider that the received signal consists of two orthogonal plane waves whose amplitudes and relative phase vary in some fashion since this is the information that the antenna extracts. If we assume that the telemetry transmitting antenna is a simple dipole with its E field along the line of capsule travel then a straight overhead pass would appear to the receiver as if only vertically polarized signals were being transmitted. Conversely, a pass low on the horizon would result mainly in horizontally polarized signals.

In these two admittedly idealized situations the contribution of one of the channels to the total signal is mainly noise. Then, too, as the signal in one of the input channels of the phase-sensing element goes to noise, erratic wandering of the phase shifter would result unless an amplitude-sensing shut-off were provided. Thus it would seem useful to at least examine what would be involved in a selection-diversity system or, better still, one whose mode of operation can be switched.

Figure 5 illustrates a possible configuration of such a system while still maintaining equal-gain combining if desired. An amplitude-



POSSIBLE SELECTION DIVERSITY SYSTEM

FIG. 5

sensing circuit is required to determine which of the two sum signals is the larger, this signal then being used to lock to the reference oscillator. Electrically ganged switches prior to the hybrids then connect to the appropriate polarization in the error channels, the other input to the hybrid at that point being terminated. This unfortunately adds additional noise. This would not be true in a six-channel configuration since the switching could be done at the intermediate frequency.

Several Sources of Signal Phase Shifts

In considering a polarization-diversity system combining horizontally and vertically polarized signals (as defined by the launching dipoles of the receiving antenna) in a coherent manner it behooves the designer to examine the possible sources of differential phase shift, their magnitudes, and the rates at which they occur.

One major contributor is from multipath-propagation effects. For horizontally polarized transmissions over land the reflection coefficient may be close to unity while in general the phase shift at reflection is close to 180 degrees. This is approximately correct for all vertical angles.

At very low grazing angles this is also true for vertically polarized radiation. However, as the vertical angle increases the reflection coefficient drops abruptly and the phase shift at reflection changes radically until a vertical angle of about 30 degrees is reached. Here the phase shift at reflection has reached nearly 90 degrees and the coefficient then increases rather slowly to its value at 90 degrees*.

One can see in a qualitative way that such behavior on the part of the input signals can result in deep fades for horizontally polarized signals, less deep for the vertically polarized and differential phase

* Reference Data for Radio Engineers, page 698, Fig. 31, 4th edition.

shifts between the two signals at the antenna (or decorrelation in amplitude fades since the differential phase shift occurs mainly in the reflected ray.)

Figure 6 is a plot of predicted amplitudes of vertically and horizontally polarized signals for the second pass at Southern California*. This is a fairly low pass rising to only about 20 degrees above the horizon. The horizontally polarized signals do indeed show very deep nulls and the vertically less so. The fades are fairly well decorrelated.

With the signal source on the horizon both reflected signals behave in about the same manner in both amplitude and phase. However, at about 5 degrees above the horizon the reflectivity for vertically polarized signals has dropped almost to 0.3 and the phase lag at reflection is now only about 90 degrees.

We can get some idea of the magnitudes and rates of differential phase shifts involved by constructing somewhat simple-minded vector diagrams. Let A_h and A_v represent the direct ray at the two polarizations. (See Fig. 7) Let B_h and B_v represent the reflected ray. Because of the motion of the capsule relative to the receiving antenna the B vectors will rotate but since they rotate at the same rate and we have assumed equal amplitudes there will be no differential phase shift between the two vector sums and the phase shifter will not be called upon to make a correction.

At some later time we can imagine a situation somewhat as illustrated in the lower vector diagrams. At some instant B_v is shown lagging 90 degrees behind B_h and reduced in amplitude. Since, for an average pass, it requires a minute or more for the capsule to reach an elevation of 5 degrees

* All plots of this type furnished courtesy of H. C. Peterson, Grp.28,L.L.

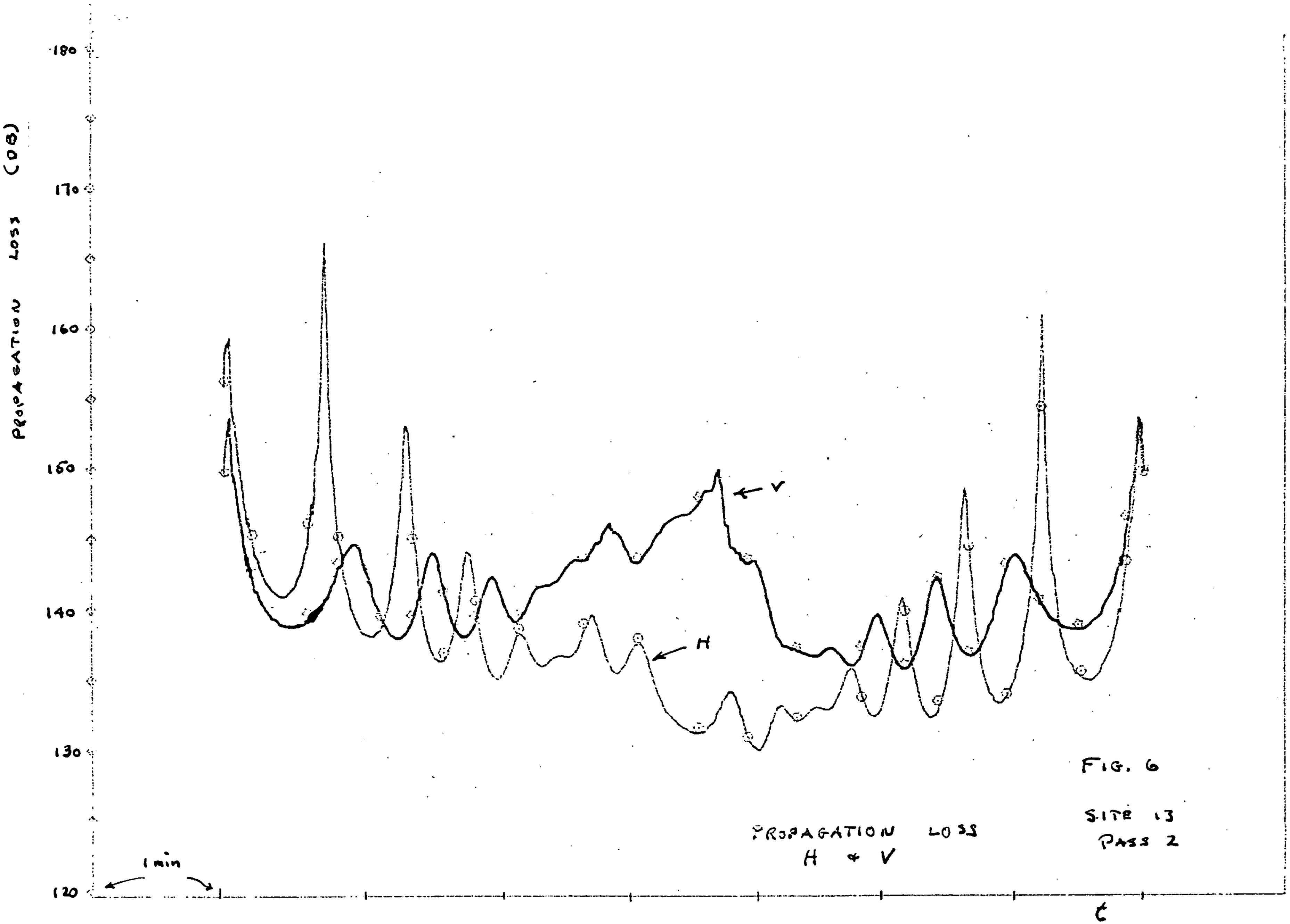


FIG. 6

PROPAGATION LOSS
H & V

SITE 13
PASS 2

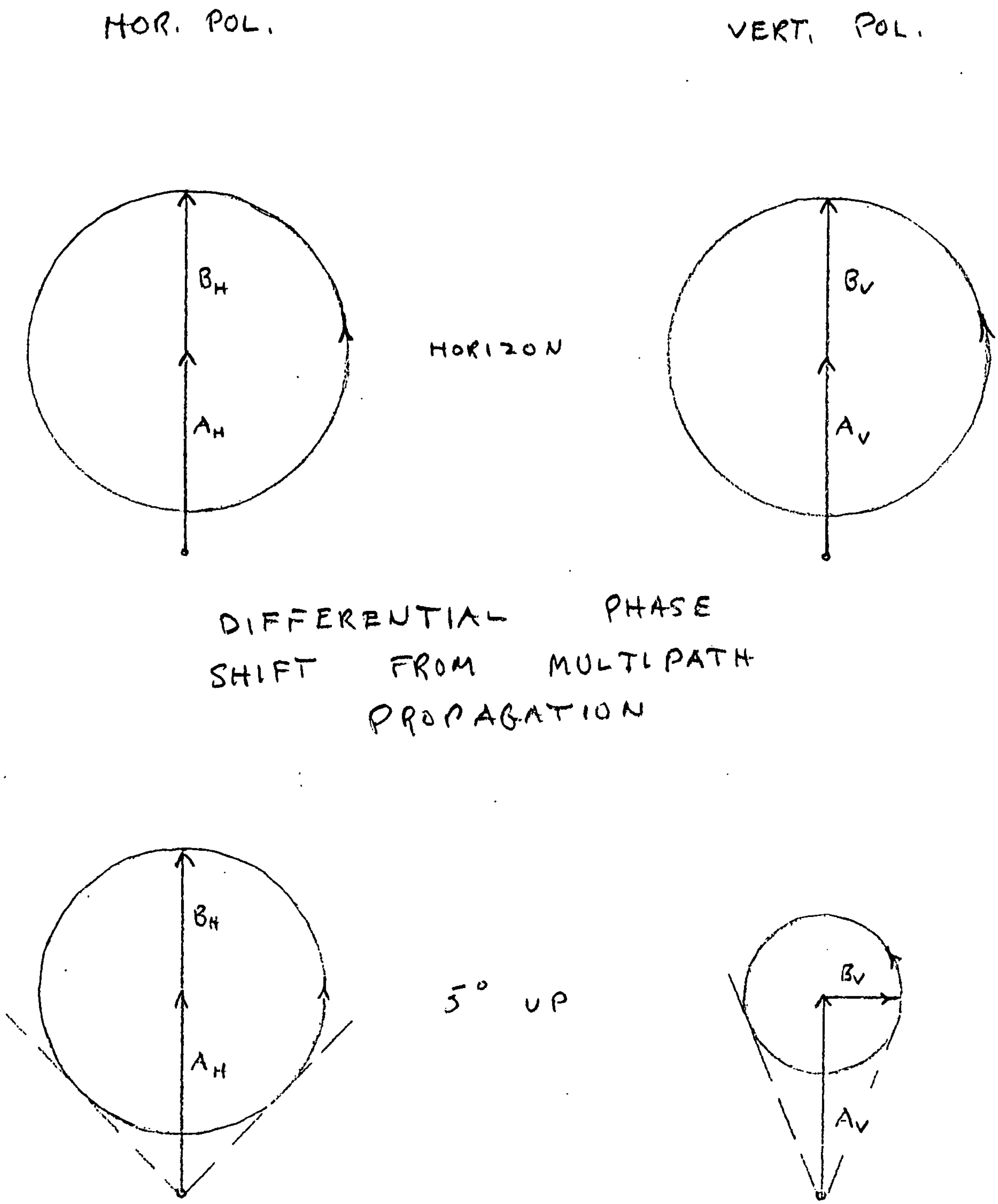

HOR. POL.
VERT. POL.
B_H
A_H
HORIZON
B_V
A_V
DIFFERENTIAL PHASE SHIFT FROM MULTIPATH PROPAGATION
B_H
A_H
5° UP
B_V
A_V

FIG. 7

no particular demands are made of any phase shifter to make this slow correction.

A second and more rapid source of differential phase shift arises because B_v is smaller than B_h in general. As B_h rotates, the vector sum of the direct and reflected ray swing, in our illustration, through about 90 degrees while the corresponding vertically polarized vector swings through only 45 degrees. The rate at which this occurs varies with the station, depending on its altitude above the reflecting surface and the pass. For a 50' altitude a typical number might be two to three revolutions per minute so that the phase shifter might have to correct 45 degrees in 20 seconds to pick a ball park figure. This is a most reasonable rate for almost any device. For higher receiving antennas this rate goes up but even an increase by a factor of five represents a correcting rate of only about 10 degrees per second.

For high passes (Fig. 18) the interference rate also goes up at angles about 20 degrees above the horizon and continuing to about 30 degrees. The direct ray is of course still in the main beam of the receiving antenna but the reflected must now enter the antenna on a side lobe and is much attenuated. Our vector diagrams, if we extended them to this situation, would consist of a very small vector adding to and subtracting from a very large signal. The total phase shift of the combined signal relative to the direct ray is thus very small and the differential between the two combined signals even less.

It also seems plausible that amplitude variations in either the vertical or horizontal components do not cause additional relative phase shifts. If A_h becomes smaller then we would expect B_h to become correspondingly smaller so that the total phase gyrations of the horizontal signal would not be greatly affected.

Admittedly the preceding analysis is a superficial one. The problem is infinitely more complex, involving as it does the patterns of the capsule antenna at various aspects, the pattern of the receiving antenna and the geography and surrounding terrain of the different sites. However, it does seem sufficiently valid to give some idea of the order of magnitude of the differential phase shifts and the probable rates.

Another possible source of differential phase shifts between the orthogonally polarized signals is that occasioned by Faraday rotation. While the total polarization rotation which may be experienced depends on a number of factors, some of which are time variable and some not well known, 70 degrees for one-way transmission at 200 Mcps has been given as a reasonable figure.*

When a plane wave enters the ionosphere, it breaks up into two counter-rotating circular (or elliptical) polarized signals which then proceed independently through the medium at different group velocities, the relative phase between them constantly changing. Upon emerging, we can think of the waves again combining with some arbitrary phase relationship depending on the medium and the path length through it. The only change we can observe is that the plane of polarization has shifted.

It is again convenient to think of the signal from the capsule as two orthogonal plane waves. This signal we can imagine as breaking up into four circularly polarized signals, two of each sense. Each pair (CW and CCW) then experiences the same phase shift and upon emerging we again have two orthogonal plane waves with the same phase relationship but oriented

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* Michael S. Macrakis, Faraday Rotation for the Mercury Capsule at 200 Mcps., Memo No. 20-0026, 26 Feb. 1960.

in some different frame of reference.

Thus at the receiving antenna--which establishes its own frame of reference as far as polarization is concerned--the horizontal and vertical components will be extracted with no differential phase shift due to Faraday rotation.

There is one special case that should be considered. Imagine that the capsule at some instant and attitude effectively is transmitting two orthogonal signals of equal amplitudes whose planes of polarization are aligned with the receiving antenna as in Fig. 8a.

If then a 45-degree rotation of the transmitted planes of polarization occurs, the situation is as in Fig. 8b. The receiving antenna extracts the horizontal component but the vertical components are out of phase and cancel. Further rotation will again restore a vertical component but 180 degrees out of phase, and the phase shifter must make a correction.

Thus far, in discussing effects due to Faraday rotation, only the direct rays have been considered. What additional differential phase shift can we expect, contributed by the reflected signal, as a result of rotation?

If the preceding analysis is substantially correct, we can expect changes in relative amplitude of the direct rays as the plane of polarization is rotated depending on the antenna pattern and attitude of the capsule but no relative phase changes except at a null. Hence the reflected signals will in turn exhibit the amplitude changes in the two polarizations but, as observed earlier, if the amplitude ratio of the direct and reflected rays stays approximately the same, no additional phase shift results. While there has been effectively a change in path length, the change has been the same for both rays. Hence we should expect only minor differential phase shifts from any Faraday effects.

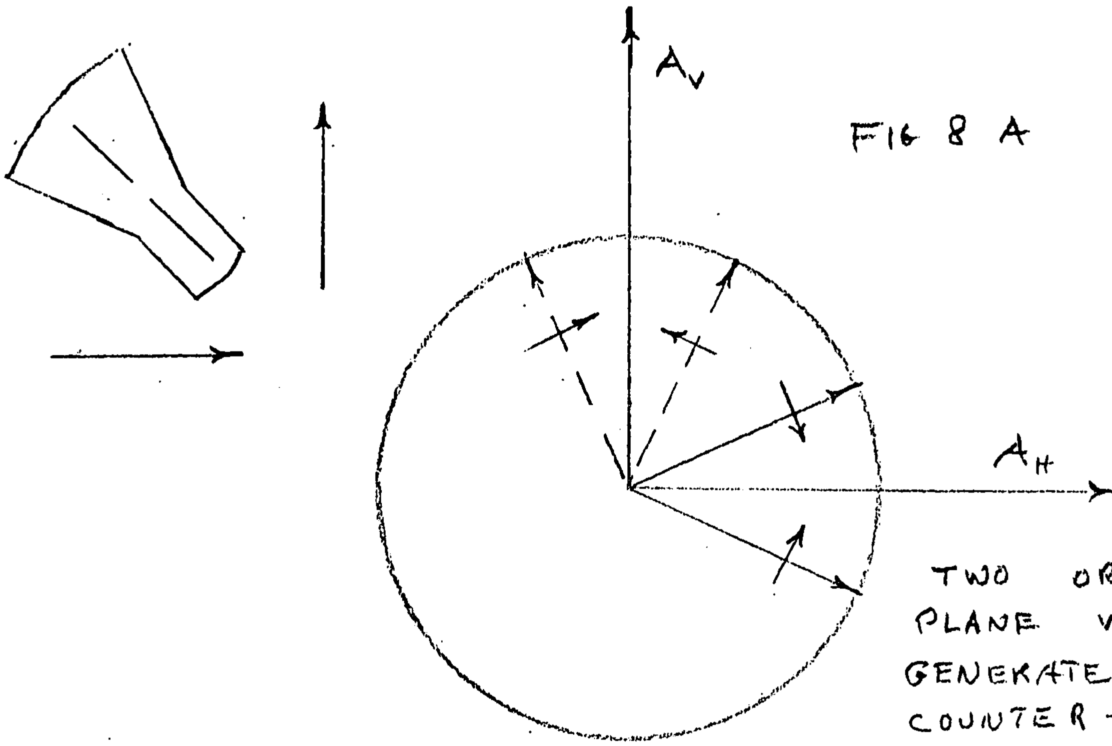

FIG 8 A

TWO ORTHOGONAL PLANE WAVES GENERATED BY COUNTER-ROTATING CIRCULAR WAVES

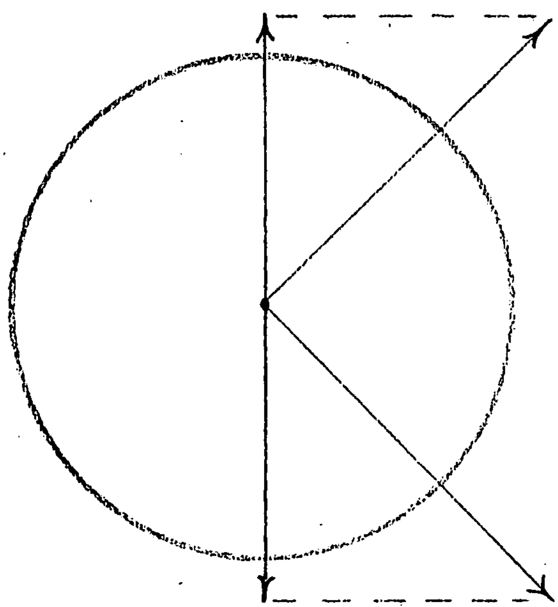

FIG. 8 B

45° ROTATION OF PLANE OF POLARIZATION VERTICAL COMPONENTS CANCEL AT RECEIVER

Any antenna that one can build is not an isotropic radiator through its entire solid angle, but rather exhibits gain in some directions and nulls in other directions. With each null is associated a change in phase of 180 degrees. If one passes through a null at one polarization, the phase between the two received signals reverses and a correction is required. Again, however, predicted signal strengths do not show these nulls.

To summarize, it would appear that at angles close to the horizon rather rapid differential phase shifts occur from multipath effects, that these diminish above 30 degrees, but that there is the possibility during the pass that nulls will occur with resultant phase reversal.

It does not appear, except for the sudden reversals, that phase rates will exceed those which can be handled by a mechanical phase shifter. For this reason, because of its quick availability and the greater complexity required by other devices, the motor-driven line-stretcher approach was chosen for this experimental receiver.

A Brief Description of the Equipment

The antenna employed is a General Bronze four-element bi-polarized "Swept Volume Efficiency" type. The electrical characteristics have been previously reported*. It was mounted on a SCR-584 pedestal and driven by lineal descendents of the servo system of that venerable machine. As can be seen in the accompanying photograph, long weighted booms were required to balance the antenna whose center of gravity is far forward of its mounting pad, so that it would perform adequately in elevation.

A second mechanical problem which was encountered is that of wind effects. A strong breeze blowing other than parallel to the circular

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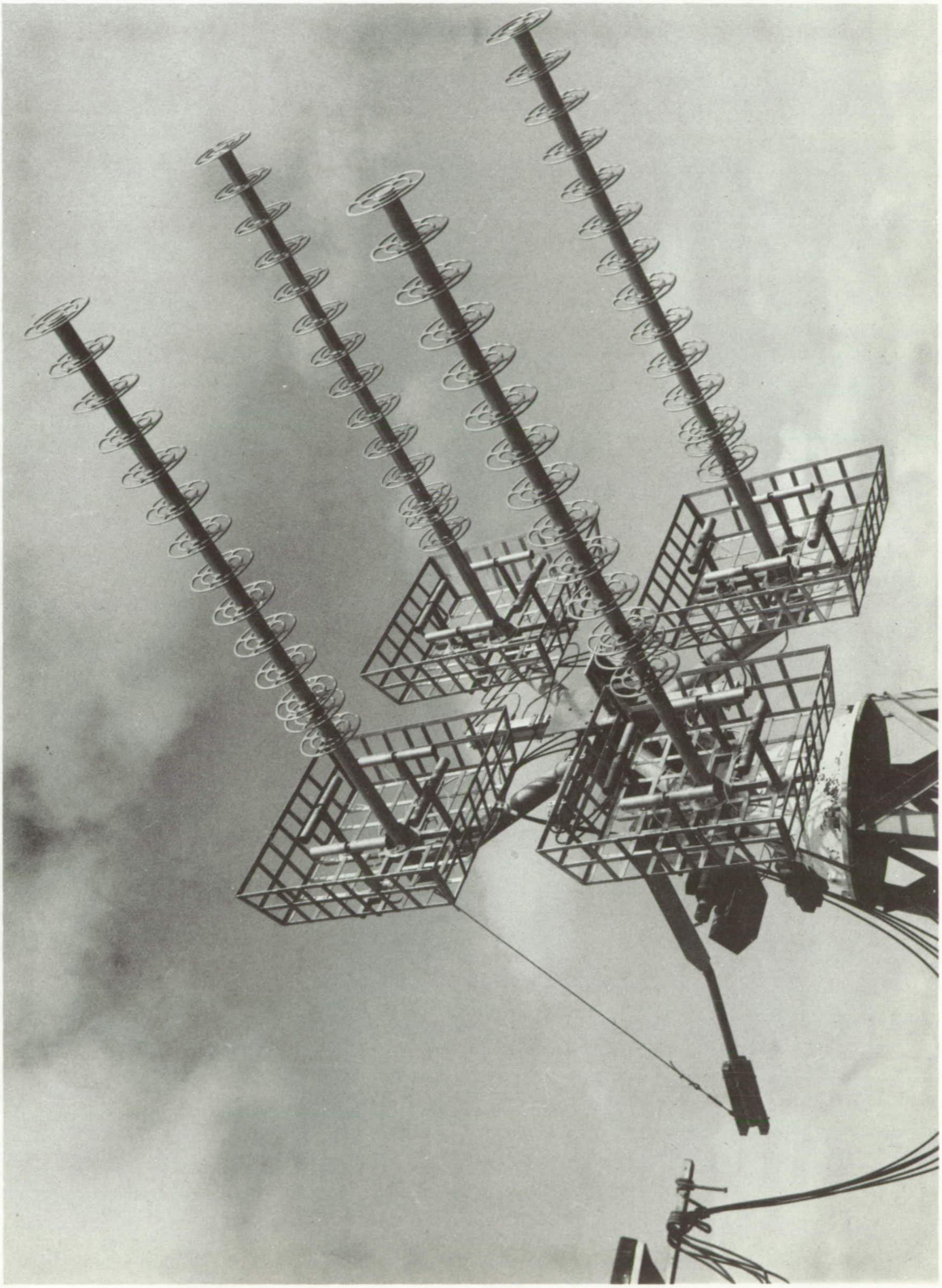
* F. Sheftman, Electrical Test of General Bronze Bi-polarized SVE Antenna, Lincoln Laboratory Report 20-0057, October 27, 1960.



beam-forming elements tends to rotate the structure and to heavily load the servo when the antenna is trying to beat up-wind. Gusty wind conditions tend to set up strong oscillations in the four beam-forming element supports.

The hybrids used were Alford 1027 types. Figure 19 shows the complete inter-cable diagram through the phase shifters and up to the r.f. amplifier inputs.

For mechanical convenience and to avoid extensive weather-proofing, the phase shifters were mounted on the receiver racks and the signals from the antenna were brought down on six cables. From a noise-temperature point of view, this is not the most desirable arrangement. Considerations involved in mounting the r.f. components on the antenna point up the desirability of light-weight, low-volume phase shifters.

The phase shifters were designed and built by the Radar Design Corp. They consist of three trombone line stretchers 360° long which are mechanically ganged together and driven by a long lead screw. Electrically they consist of two 50-ohm lines joined by a 40-ohm line, the discontinuities being a multiple half wavelength apart. As can be seen from Figs.(9) and (10) this frequency dependence results in a change in input impedance and hence VSWR from the frequency for which they were designed, 225 Mcps to the other end of the band, 259 Mcps. However, designs are available which can be made flat over a much wider frequency range. For this particular application where operation was always at 225.7 Mcps, this feature was not important.

Mechanically, the breakaway torque of the line stretches was measured as 15-inch ounces with the average running torque about 12-inch ounces.

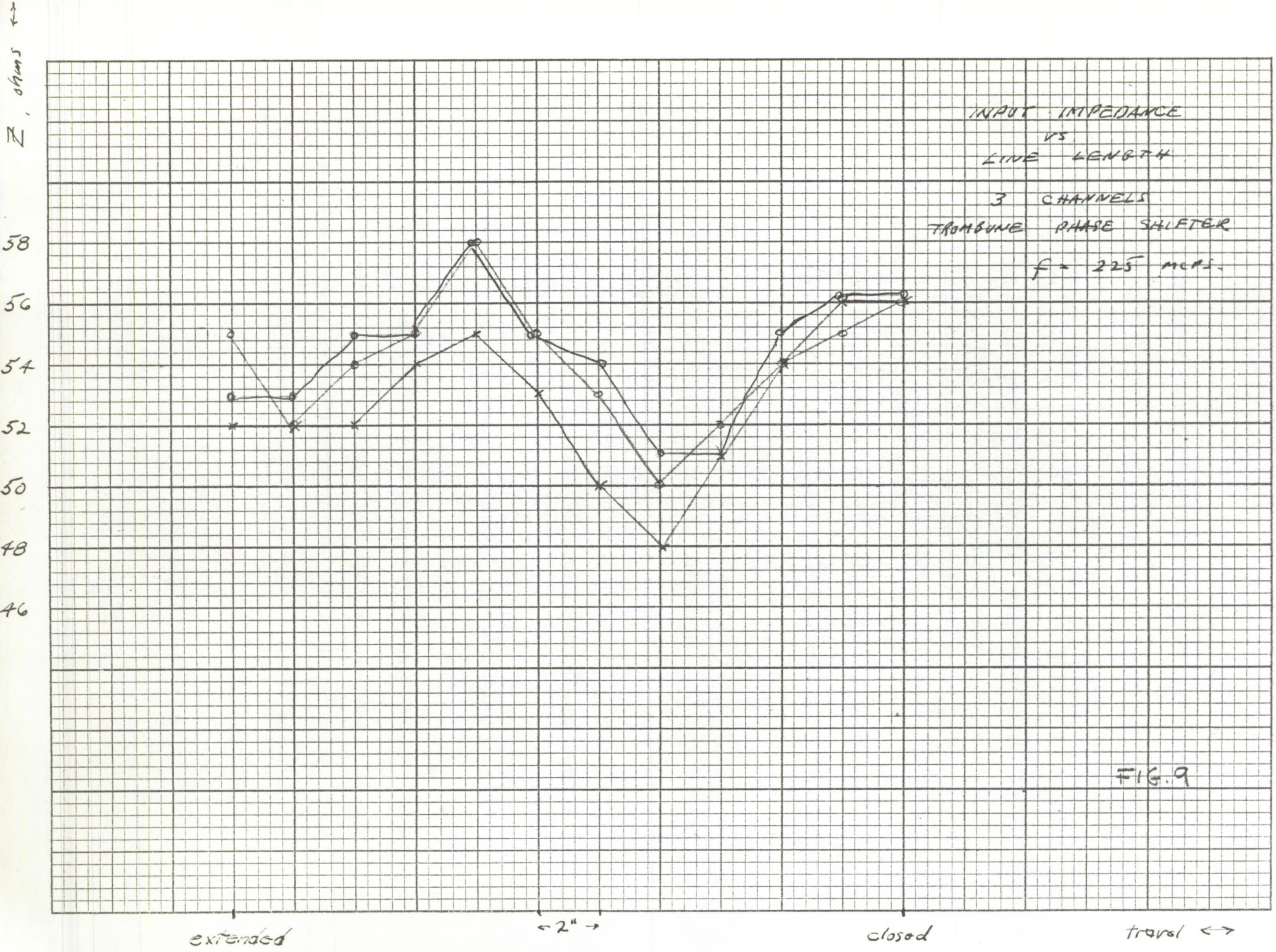

Z, ohms ↔
58
56
54
52
50
48
46
INPUT IMPEDANCE
VS
LINE LENGTH
3 CHANNELS
TROMBONE PHASE SHIFTER
f = 225 MCPS.
extended
← 2" →
closed
travel ↔

FIG. 9

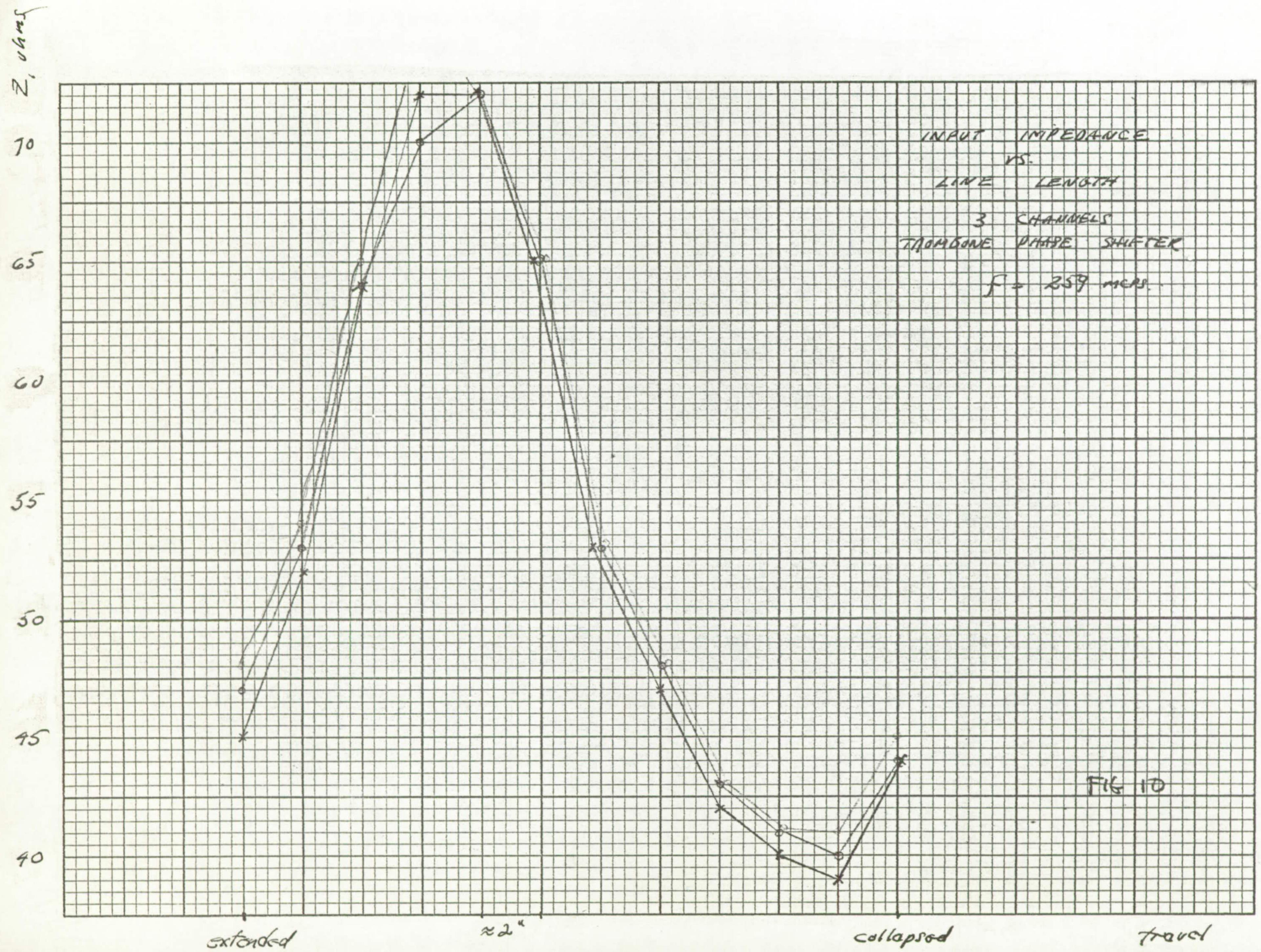

Z, ohms
70
65
60
55
50
45
40
INPUT IMPEDANCE
VS.
LINE LENGTH
3 CHANNELS
TROMBONE PHASE SHIFTER
f = 259 mcps.
extended
≈2"
collapsed
travel

FIG 10

Using a 10-watt motor as the driving source and a gear ratio to give sufficient torque, the phase correction rate is approximately 10 degrees per second. The average phase error to overcome starting friction is about 2 degrees with a phase lag of approximately 10 degrees at the maximum phase correcting rate. It is felt that with a better mechanical design of the line stretchers, the tracking rate could be improved by a factor of about four. This corresponds to a linear velocity of 5 inches per second or 360 degrees of phase shift in 10 seconds. This then becomes competitive with 225 Mcps ferrite phase shifters for which figures of "several seconds" were quoted.

The r.f. amplifiers used are commercially available shelf items. The pass band is 5 Mcps wide at 225 Mcps, gain is of the order of 30 db and the advertised- and measured-noise figure is 4 db. The following mixers are Empire Products type CM-107A.

The first local oscillator consists of (for this receiving frequency) a 47.675 Mcps crystal oscillator followed by four broadband quadruplers. The i.f. amplifiers are conventional in-house built, double-conversion units. The first intermediate frequency is 35 Mcps which is then mixed with the output of the voltage tuned oscillator at 32.785 Mcps to the second intermediate frequency of 2.215 Mcps. The over-all bandwidth to this point is 50 kcps which insures adequate capture and lock range in the phase-locked loop for the kind of tests planned.

In the dual sum channels the outputs of the two i.f. amplifiers are compared in phase in a standard configuration 6BN6-type phase detector. The resultant d.c. signal is chopped and amplified and used to drive a motor which in turn drives the triple-ganged phase shifters.

Since the phase shifters are linear devices in the sense of being non-rotating devices and since they have finite travel, safety micro-switches are mounted at each limit. These actuate a quick-return circuit returning the shifters 180 degrees to mid-position. An optional approach would be to cause these microswitches to switch in (or out) additional lengths of line. However, it is not reasonable to expect a continuous phase shift in one direction on the part of one of the polarizations since this would amount to a frequency change for one polarization; that is, differential phase shifts, integrated over sufficient time must equal zero.

Still in the sum channel, it is here that the vertically and horizontally polarized signals are added together to produce the composite sum signal which locks the voltage-tuned oscillator and generates the AGC voltage. These two loops will be considered in more detail in a later section.

Since the signals to the phase detector are held at 90 degrees out of phase, the normal operating point for a correcting phase detector, an additional phase shift is required in one leg to the summing circuit. Note also that compensating cable lengths must be added prior to the summing hybrids in the two error channels.

The two error channels, circuit-wise, are similar to the sum channels through to the second i.f. amplifier. Here bandwidth-limiting filters are employed to reduce the noise bandwidth. After lock one can also reduce the bandwidth in the sum channel, but for simplicity this feature was not included.

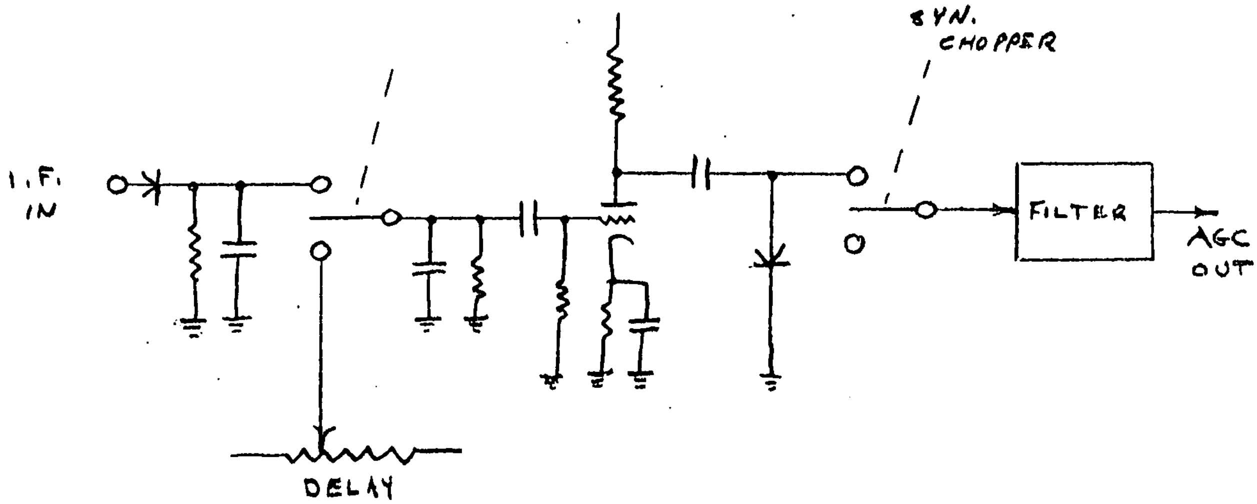
As can be seen from the block diagram, the crystal-generated reference signal which is compared with the error signals in each channel passes through additional phase shift. This circuit too is discussed in greater detail in a later section.

The AGC circuit accepts the rectified composite sum signal and develops a bias which is applied to all i.f. amplifiers. Thus, if the signal at one polarization is very large and a large AGC bias is developed, the noise contribition of the channel with the weak signal is hence reduced.

While the circuitry is simple, it is discussed since the method of applying delay may be somewhat novel. As can be seen in Fig. (11), the summed i.f. signal is rectified and filtered and applied to one contact of a synchronous chopper. The delay voltage, variable as desired, is applied to the other contact, the chopper arm alternately sampling each. The action is as follows: when the rectified i.f. signal is less than the delay voltage, a square wave of some amplitude and phase is fed to the grid and amplified. The output of the amplifier is diode clamped to ground. The amplitude of the output signal represents the difference between delay bias and signal amplitude. However, the other contact of the synchronous chopper samples only during the period indicated by the arrows and no AGC signal results.

When the rectified signal becomes larger than the delay bias, a chopped signal of the opposite phase results. Again the diode clamps to ground but now during the period of time corresponding to the delay voltage and the chopper is now sampling the signal voltage and a negative bias results.

One of the simplifications made in the circuitry was to use an LC oscillator as the voltage-tuned oscillator in the phase-locked loop.


I.F. IN
SYN. CHOPPER
FILTER
AGC OUT
DELAY

AGC

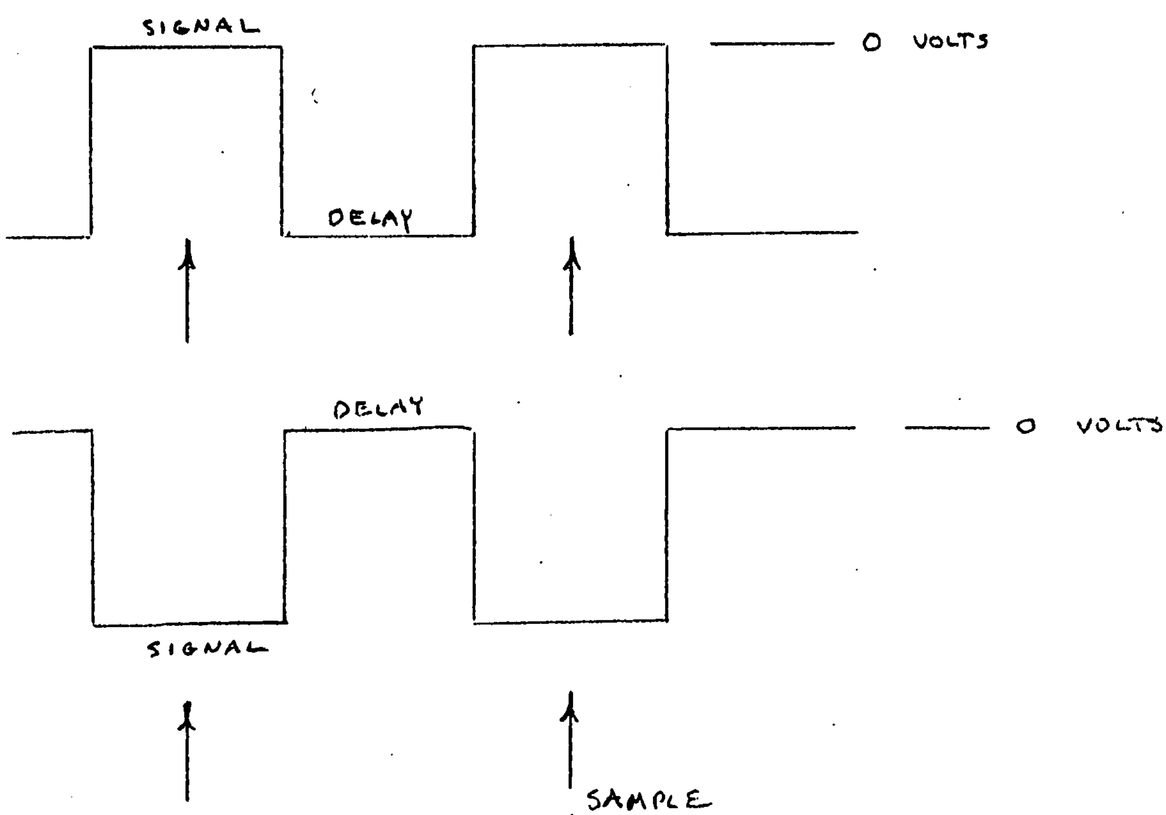

SIGNAL
DELAY
0 VOLTS
DELAY
0 VOLTS
SIGNAL
SAMPLE

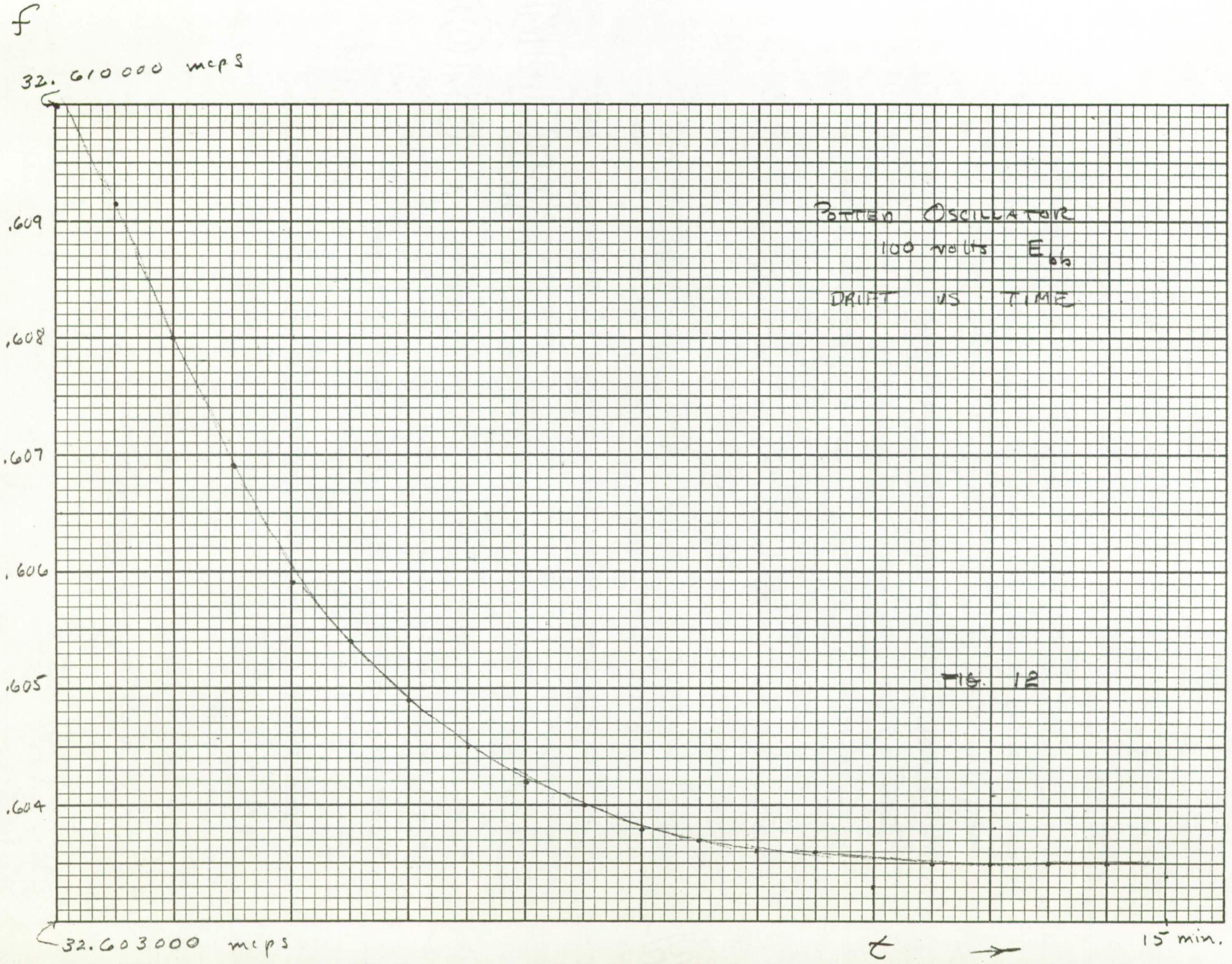
FIG. 11

This made it relatively easy to achieve the loop gain required to capture and track over a fairly wide frequency range. The voltage-tuned oscillator has a gain of 2π x 1000 radians per second per volt while the phase detector has a gain of 20 volts for π radians. The lock range thus becomes 2π x 20,000 radians per second or 20 kcps. Since no bandwidth limiting is used in this application, the capture range is the same.

Of course, given an oscillator that is easy to "swing," there exists the possibility that it is prone to drift. While a manual control is available for re-tuning, constant re-tuning can be annoying and so some effort was made to stabilize the frequency so that it stayed within the lock range of the loop.

Thermal drift was minimized by choice of components, by using very light chassis construction so that thermal equilibrium was quickly reached and by operating at low plate voltages. The entire assembly was then potted in a high-temperature wax except for one trimmer capacitor. Because of the dielectric properties of the wax some cutting and trying was necessary to insure that once the oscillator was potted, one still had an oscillator and that it covered the correct frequency range.

Through these relatively minor steps, the improved oscillator reached a first thermal equilibrium in a little over ten minutes after a drift of 6500 cycles per second. (See Fig. 12) After 2-1/2 hours of running time the frequency had drifted back almost back to the starting point. This frequency drift is well within the capture and lock range. The short-term stability, with the oscillator tube operating with a.c. on the filaments, varied from 3 to 30 cycles per second, the sampling being done at 3-second intervals. However, once the oscillator is locked, it maintains itself at exactly the difference between the crystal


f
32.610000 mcps
.609
.608
.607
.606
.605
.604
32.603000 mcps
t
15 min.
POTTED OSCILLATOR
100 volts E_{bb}
DRIFT VS TIME
FIG. 12

reference oscillator and the incoming signal frequency. The oscillator circuit was a simple Colpitts using a conventional triode.

However, for best operation of a tracking system using a phase-locked loop, it is not sufficient that the voltage-corrected oscillator's unlocked or free-running frequency stay somewhere within the lock range but rather that it stay within fairly close limits. A better way to state this is that the phase detector controlling its frequency should operate around a limited portion of its transfer characteristics where the two input signals are close to 90 degrees out of phase and should not be required to make corrections to the limits of its lock range. These remarks apply only when the crystal-oscillator signal is used as the reference in the azimuth and elevation error channels.

Let us assume an ideal system in which the sum signal and the oscillator reference signal to the frequency-correcting phase detector are exactly 90 degrees out of phase and that further, through appropriate phase shift, the two error signals are in phase with the reference on one side of boresight and 180 degrees out of phase on the other side of boresight. If we now experience a frequency shift, either of the transmitter or in the natural frequency of the voltage-tuned oscillator, a correction is required. The voltage to make this correction is derived from the phase detector in the phase-locked loop by allowing the phase of the sum signal and the locally generated reference to "slide by" each other. If we now examine the signals at the two error-channel phase detectors, we find that they are no longer in or out of phase with the reference on either side of boresight but rather have some arbitrary phase relationship depending on the magnitude of the frequency correction that was required.

This can have two deleterious effects on a system's tracking performance. One is that the over-all gain of the servo controlling the antenna position goes down, can result in sluggish operation, and, in the limit, the gain can go to zero. The other is a boresight error, the amount of which also depends on the amplitude balance or depth of null from the antenna-hybrid complex.

Data on this effect were taken on the polarization-diversity receiver using a boresight tower with a dually-polarized antenna as the signal source. First, the azimuth antenna difference pattern was measured rather roughly by swinging the antenna across the boresight tower and recording the peak-to-peak output of the azimuth error-channel i.f. amplifier. The voltage null of 0.12 volt was estimated to be twice noise and this point was then taken as 0 db. The null depth obtained as shown in Fig. (13) is about 29 db which agrees fairly well with other data taken previously on this antenna.

The first test made was to manually point the receiving antenna at the boresight tower and then tune the voltage-controlled oscillator over most of its lock range while recording the correcting voltage developed as well as the error-channel phase-detector output. Knowing the V.C. O characteristics the correcting voltage may be correlated with frequency. As shown in Fig. (14) a change of 14 kcps, roughly 0.7 of the total lock range, resulted in phase-detector outputs of plus and minus 2-1/2 volts. Since this was done in an open-loop condition, the data are of interest only in a qualitative sense.

Of more interest is what one might call the restoring force; the phase-detector signal generated as the antenna is swung past boresight.

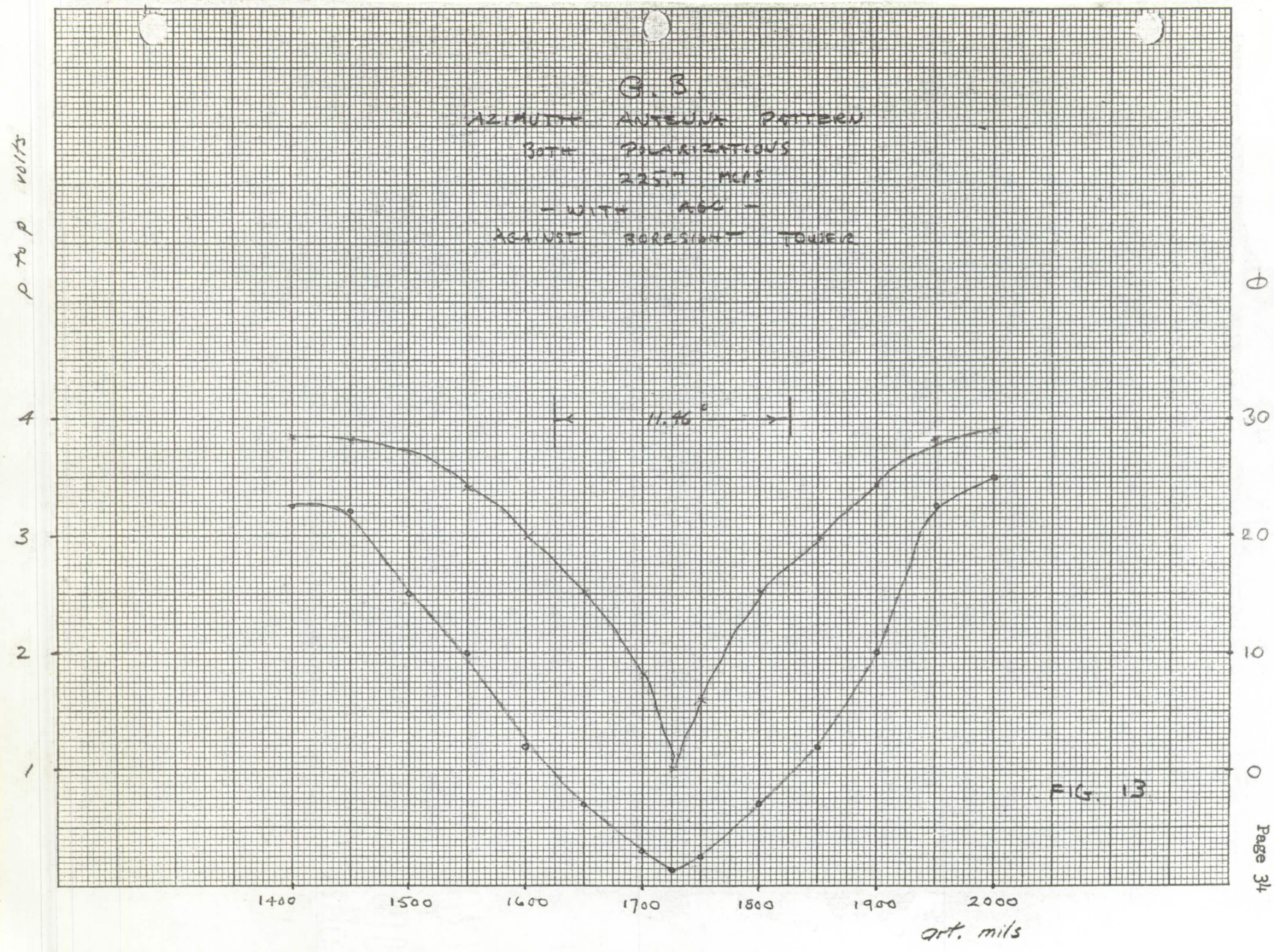

G.3.
AZIMUTH ANTENNA PATTERN
BOTH POLARIZATIONS
225.7 MCPS
- WITH AGC -
AGAINST BORESIGHT TOWER
11.46°
p to p volts
1
2
3
4
0
10
20
30
1400
1500
1600
1700
1800
1900
2000
art. mils

FIG. 13

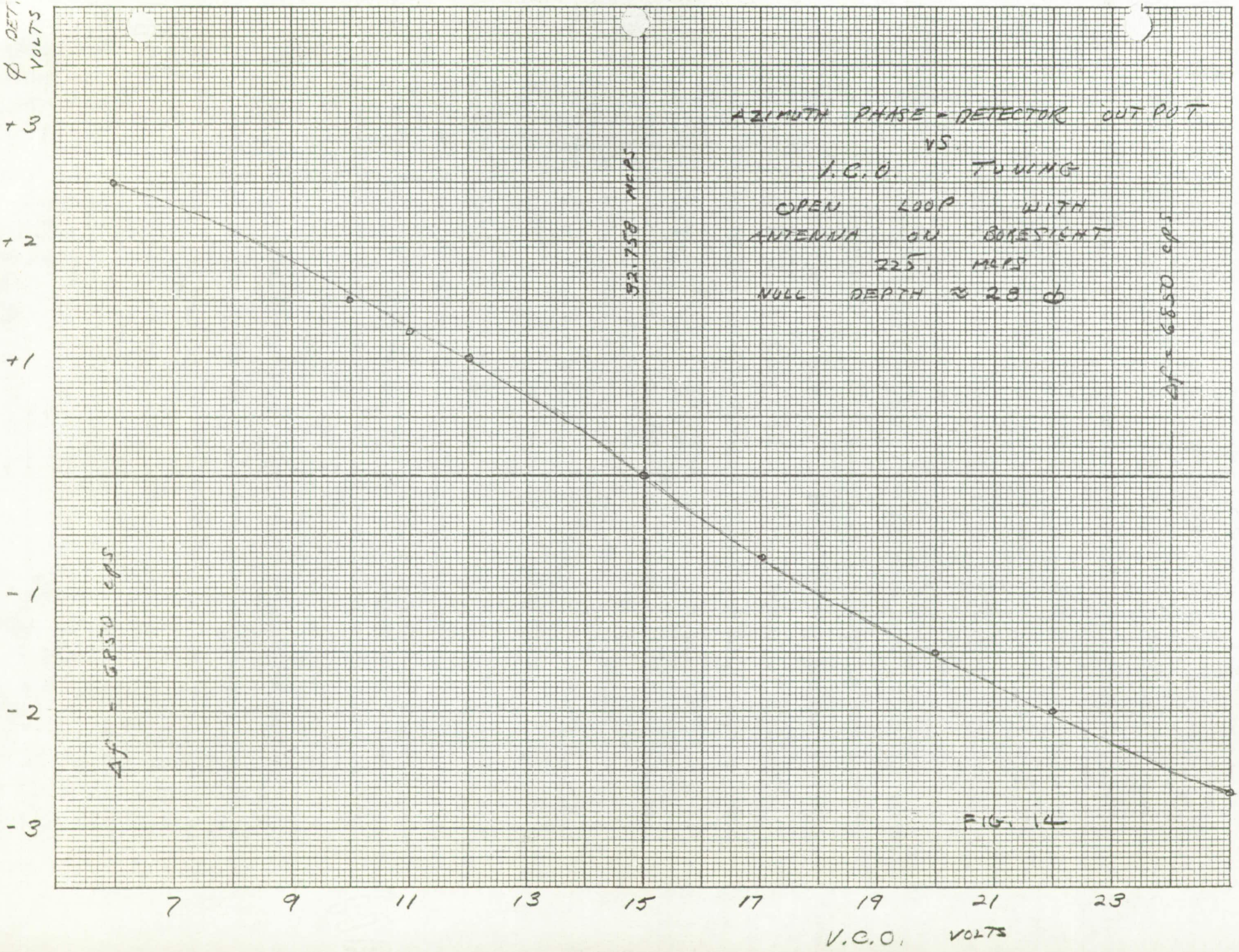

AZIMUTH PHASE - DETECTOR OUTPUT
VS
V.C.O. TUNING
OPEN LOOP WITH
ANTENNA ON BORESIGHT
225. MCPS
NULL DEPTH ≈ 28 db
32.758 MCPS
Δf = 6850 cps
Δf = 6850 cps
φ DET. VOLTS
+3
+2
+1
-1
-2
-3
7
9
11
13
15
17
19
21
23
V.C.O. VOLTS

FIG. 14

Several of these plots are shown in Fig. (15). Again, the volts listed are the operating points of the phase detector and correspond to some inserted frequency error. These data were taken as follows: The system was aligned for some V.C.O. frequency, in this case the one corresponding to 28 volts. The technique used for this alignment is to swing the antenna to one side of boresight, look at the reference signal and the error signal on a dual scope and adjust the manual phase shifter in the reference channel until the two signals are in- or out-of-phase, depending on the sense of the correcting voltage required.

Once this was accomplished the antenna was swung across the boresight tower and the phase-detector output plotted as a function of angle. As can be seen from the curve labeled 28 volts a symmetrical curve resulted passing steeply through zero at boresight.

The V.C.O. was then tuned to a different frequency--or what would have been a different frequency if a correcting voltage were not applied--and the curve labeled 20 volts resulted. This change in voltage corresponds to a 5-kcps shift. As can be seen, the curve is less steep and passes through zero at a point some 15 artillery mils removed from the first curve. The curve labeled 15 volts corresponds to an 8-kcps shift, shows still a further reduction in gain and the error at boresight has increased to 45 mils or some 2.5 degrees.

It should be emphasized that the data which show rather large errors or changes in gain were taken at close to the extreme of the lock range. It is also true that the error-channel phase detectors are linear and that the problem is somewhat mitigated with a cosinusoidal transfer characteristic.

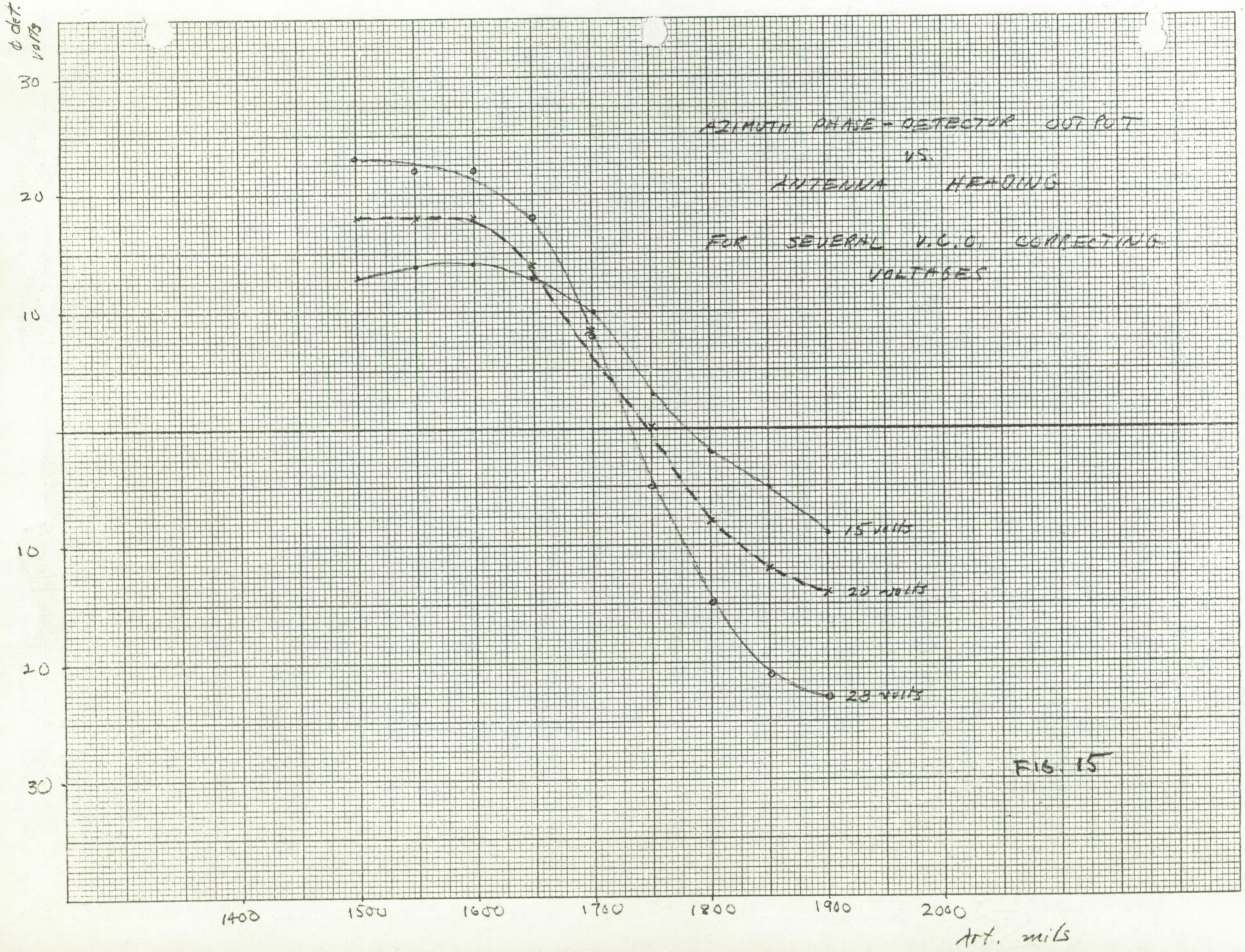
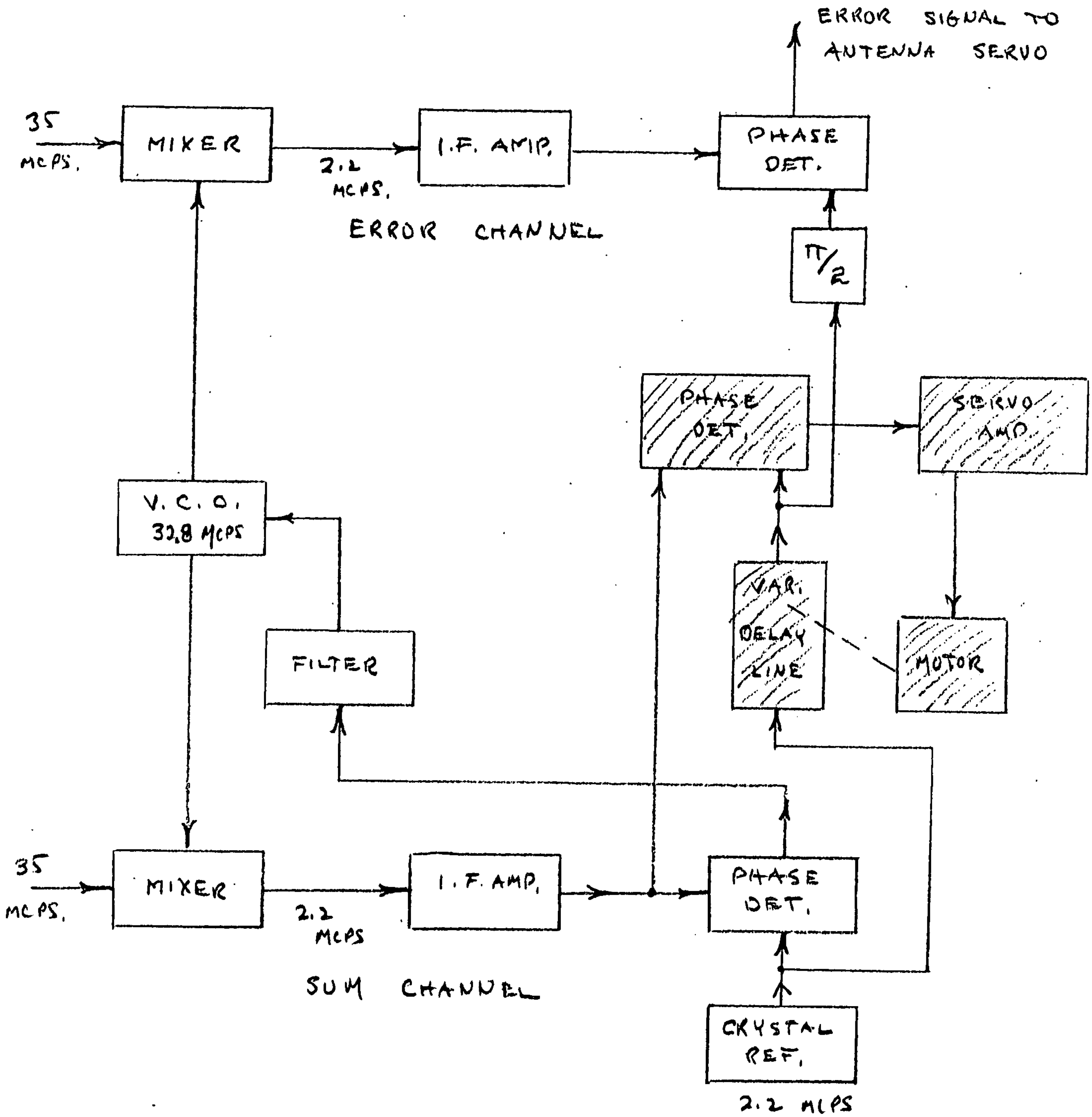

AZIMUTH PHASE-DETECTOR OUTPUT
vs.
ANTENNA HEADING
FOR SEVERAL V.C.O. CORRECTING VOLTAGES
φ det. volts
30
20
10
10
20
30
15 volts
20 volts
28 volts
1400
1500
1600
1700
1800
1900
2000
Ant. mils

FIG. 15

There are several solutions if such a problem does exist. The most obvious is to design the oscillators to be sufficiently stable and the phase-locked-loop gain high enough so that only a small section of the possible lock range is used. These two situations are mutually contradictory; a stable, "stiff" oscillator has low gain in a phase-locked loop and rather complex schemes of hetrodying and multiplying are sometimes employed. A second solution is simply to use the sum signal as the reference for the error channels. The difficulty is that the sum signal can be noisy while the locally generated reference is "clean."

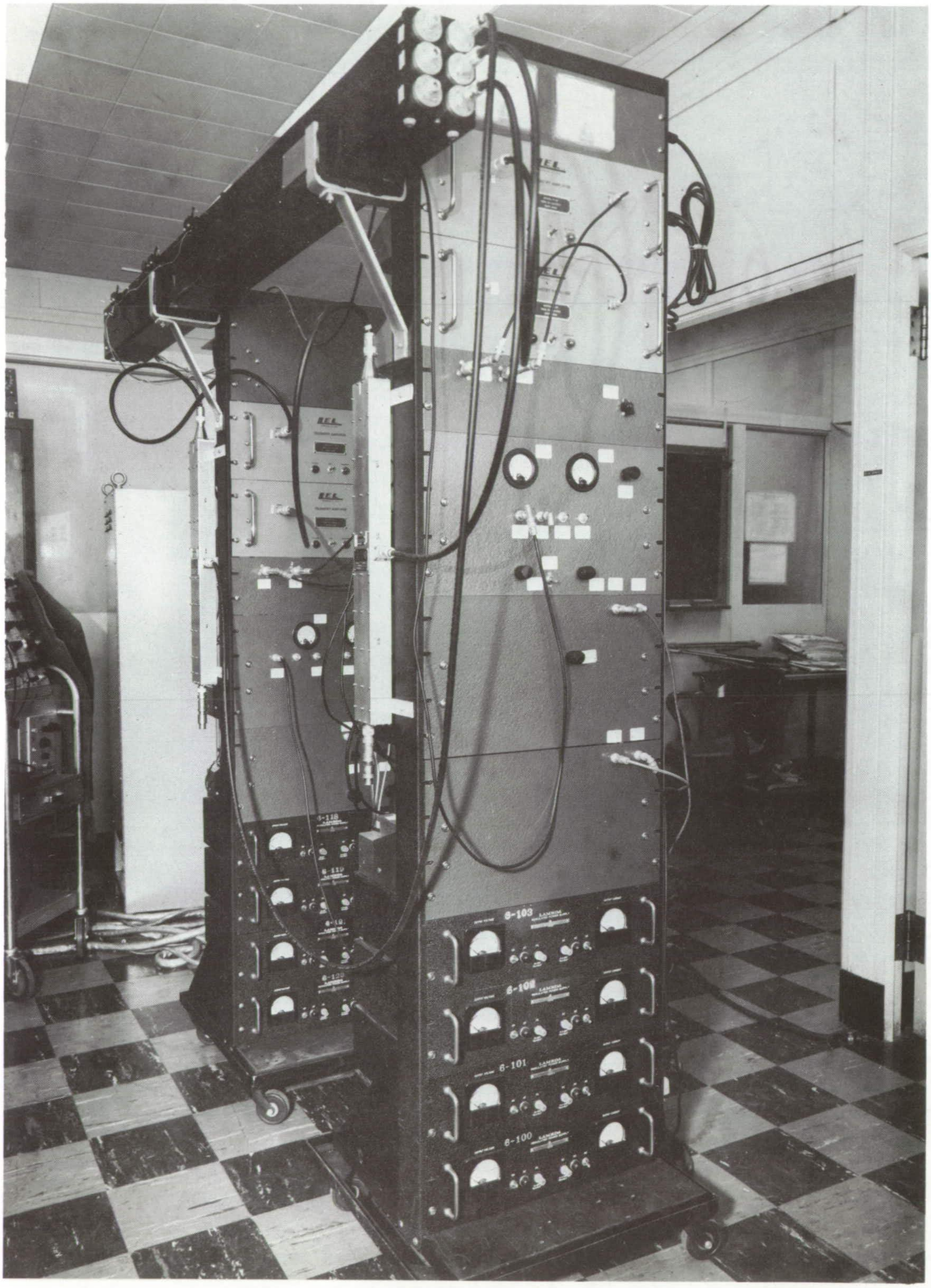
There is a compromise which was used in this receiver in order to enjoy the luxury of a high-gain, easily tuned V.C.O. without paying any penalty for its penchant for drifting. This is shown in the block diagram in Fig. (16), the added components shown shaded. These function to keep a constant phase relationship between the sum signal and the local reference after the motor-driven delay line and hence to the two error channels, thus permitting the designer to use the full lock range of his phase locked loop. Thus the reference signal is a clean signal but is of the same phase as the sum signal at all times.

The accompanying photograph shows the complete receiving equipment. The front rack principally contains the sum-channel components including the voltage-tuned oscillator and associated circuitry, the AGC, first local oscillator, etc. From top to bottom may be seen the meters indicating the magnitude and direction of the pointing error in azimuth and elevation, the two r.f. amplifiers followed by crystal-current meters and operating controls. Power supplies are located in the bottom. Much the same arrangement makes up the error-channel circuitry in the rear rack. The



USE OF ADDITIONAL LOOP TO CONTINUOUSLY CORRECT REFERENCE PHASE

FIG. 16


6-103
6-101
6-100

three-channel phase shifter is mounted on brackets and is supported between the two racks. Two hybrids may be seen at the left of each rack.

WBR/mhd

CONCLUSIONS

In designing an air-to-ground communication link the designer has some latitude in the choice of antennas and in the polarization employed both at the vehicle and at the ground site. One choice is circular to circular. However circular on the bird has no real meaning since evidence indicates that the received signal can vary from LHC to RHC with resultant very deep amplitude fades*.

Linear to linear would also prove disastrous as the plane of polarization varied with look angle and from Faraday effects. Many systems use as linear a polarization as one can obtain with complex vehicle shapes for the transmitted signal and circular polarization at the receiving antenna with a resultant 3 db. loss in received signal. Clearly polarization diversity in the receiving equipment is desirable.

The flight-test program which was set up to demonstrate the extent of possible improved performance did not yield the hoped-for data. This is not to say that successful tracks were not established and maintained on transmitter-carrying aircraft and free and tethered balloons, but rather that it was not obvious from the results that better performance was obtained using polarization diversity than could have been obtained using a single polarization or, more realistically, circular.

There are two general areas in which one would like to improve the performance of a tracking device of this kind. One is what we might call an average improvement in the signal-to-noise ratio over the time that the bird is in sight. If deep fades were encountered in receiving

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* R. C. Baker, A Circularly Polarized Feed for an Automatic Tracking Telemetry Antenna, IRE Trans. on Space Electronics and Telemetry, p. 103; Sept. 1959

a circularly polarized signal which could be "averaged out" by the use of diversity, then we have gained in the improved reliability in the transmission of intelligence. The other broad area is that of improved tracking ability for which we might establish a figure of merit as the percentage of time during a pass when the antenna is pointing at the target within some angular limits. Clearly these two areas are not independent of each other. No system will track well on a noisy signal, and in general any improvement in S/N will improve track. However there are situations where an improvement in this parameter may actually degrade the figure of merit.

Two principal difficulties were encountered in trying to simulate signals that would be anticipated during an actual capsule pass. The first was geographical. Interference-pattern predictions are most simply made if one considers a smooth conducting earth or sea. As the real estate gets more intricate the patterns grow more complex. The area surveyed by the polarization-diversity antenna consists of low rolling hills, many nearby structures, moving and parked automobiles, assorted aircraft, a number of antennas, etc. The multipath propagation, even on high passes, as observed by monitoring error signals, AGC voltage or simply by observing the antics of the phase shifter were so great as to mask any differences one might have observed between diversified and non-diversified reception.

The second problem involved the transmitting antenna. As has been observed, all antennas have amplitude nulls where the phase reverses. If this happens rapidly enough the phase shifter simply never catches up. This effect was probably more severe with balloon-borne antennas, but it is difficult to separate the symptoms of antenna nulls from the multipath-propagation effect.

Fig. (17) is a recording of the AGC voltage with the system tracking a balloon. Since the AGC voltage is derived from the sum of the two polarizations added coherently, any difference in the amplitude of either one or any error in their phase relationship shows up as a change in AGC voltage. As can be seen from the record, with the phase shifter operating it is making valiant efforts to correct and maintain maximum signal but not quite making it. The variations, with the phase shifter off, are of the order of 0.3 cps.

It is apparent from the above that a mechanical phase shifter with a response time of some ten degrees per second is not adequate for the rate of differential phase shifts encountered at this set-up. It is not quite as evident that these rates will be encountered at a site where some effort has been made to insure less complex interference patterns and where the attitude of the capsule and hence its transmitting antenna changes much more slowly with time. On the other hand, the advantages of rapid electrical as opposed to mechanical phase shifting are so attractive that if a second-generation polarization-diversity receiver were to be built this would be included if the state of the art permitted.

It is possible to make some intelligent guesses about the performance of a polarization-diversity receiver in the Mercury environment at a given site as well as certain suggestions for improved performance by studying the predicted signal strength for representative passes. The predicted propagation loss for horizontally and vertically polarized signals as shown in Fig. (18) for the third pass at the Southern California site illustrates some of the problems as well as the advantages of using polarization diversity.

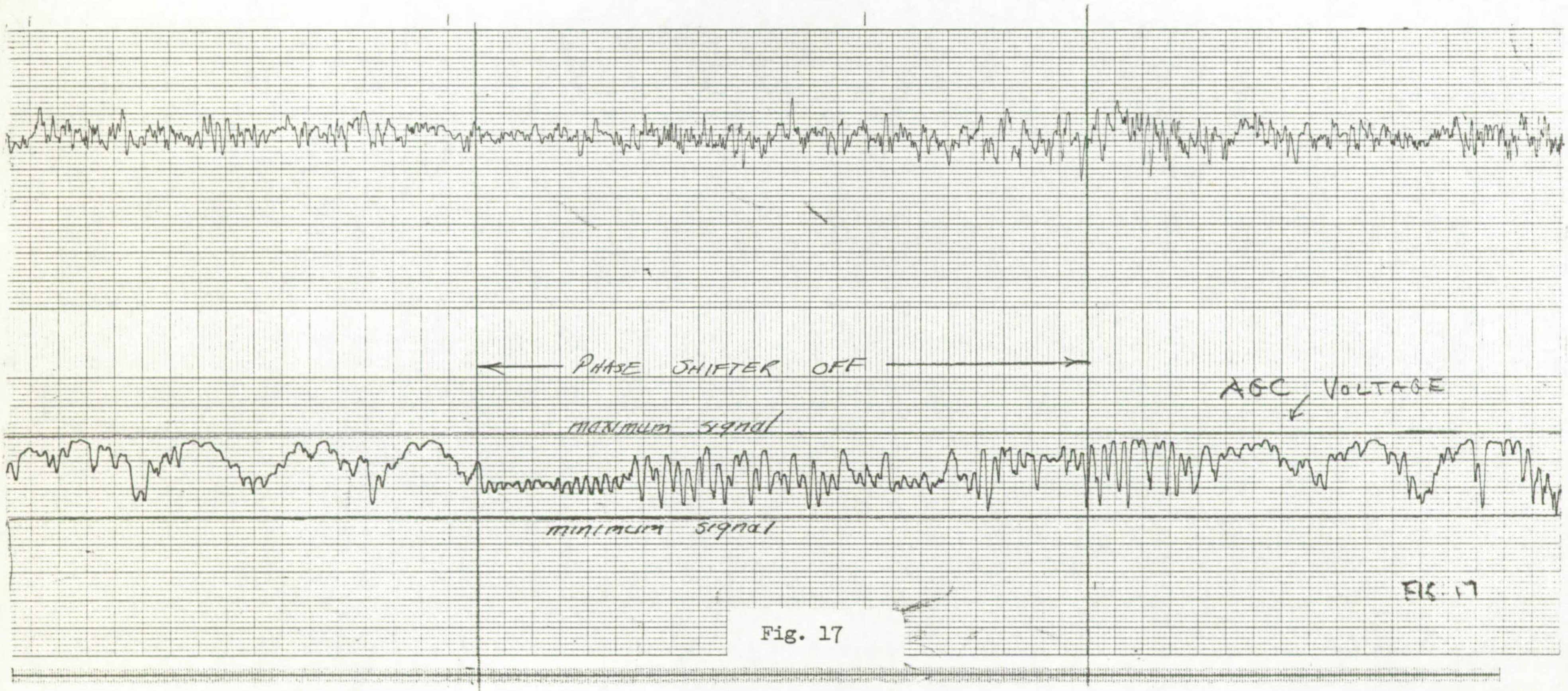
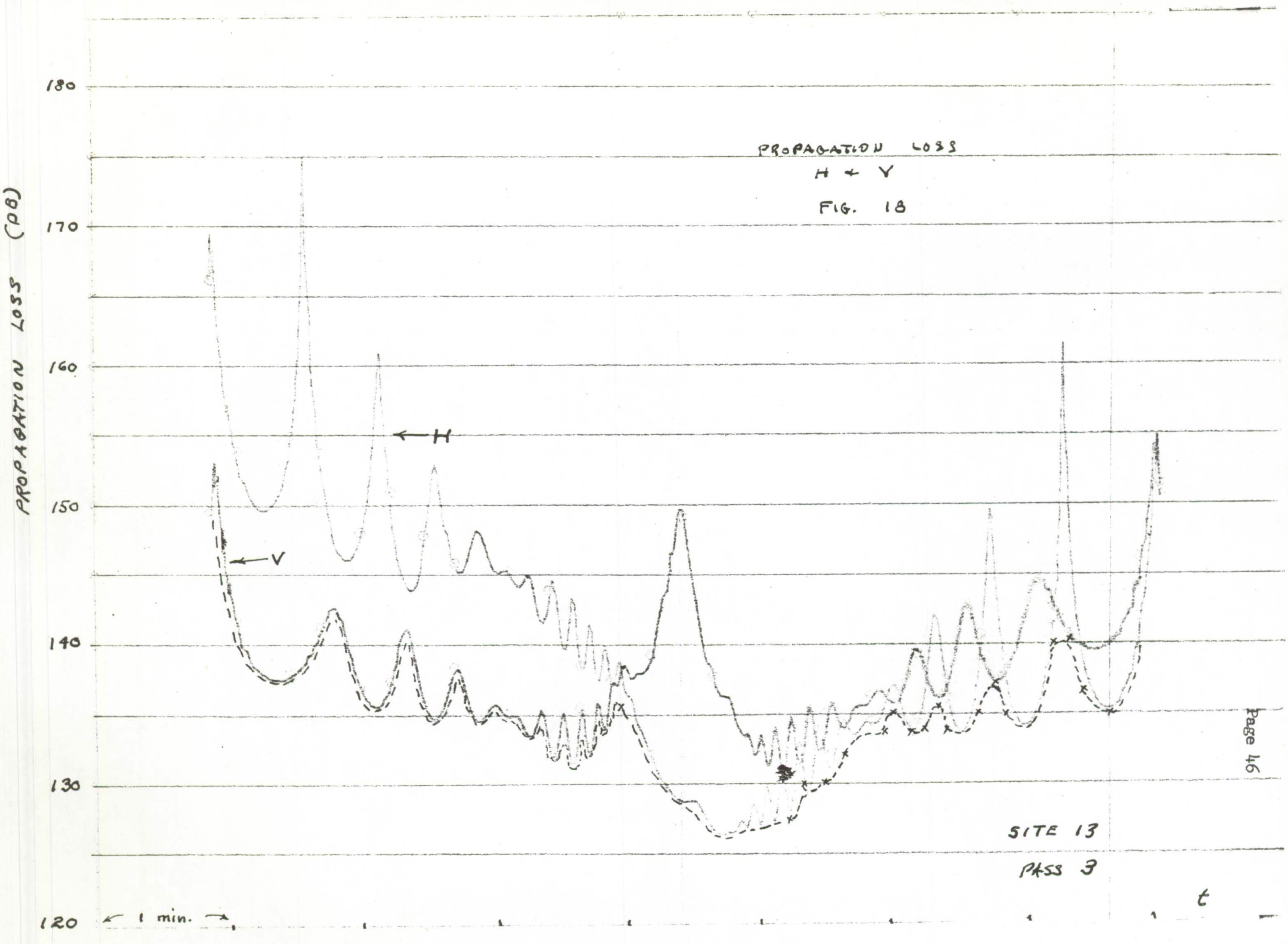

PHASE SHIFTER OFF
AGC VOLTAGE
maximum signal
minimum signal
FIG. 17

Fig. 17


PROPAGATION LOSS
H + V
FIG. 18
PROPAGATION LOSS (DB)
180
170
160
150
140
130
120
← H
← V
SITE 13
PASS 3
t
← 1 min. →

If one assumes perfect coherent addition of the two signals, one may quickly sketch in the approximate resultant space loss, shown dotted, by assuming 3 db. less propagation loss when the signals are equal in strength, that the received signal is equal to the stronger when the weaker is 10 db. down and a rough sliding scale for points between.

A space loss of 149 db has been estimated to be the maximum allowable for good lock and track as applied to the Cubic machine a commercial version of a similar device, and that is the figure that will be used here*.

It is clear that if one were restricted only to the horizontally polarized signal, trouble would be experienced during the first 1 1/2 minutes of track. However, then the vertically polarized signal drops to the minimum allowable level. Reference to the RHC and LHC predicted signal strengths for this pass (see Fig. 13-3, Memo No. 20-0019, Signal Strength Calculations for Mercury Capsule Communications) reveals that each drops to or below the minimum level during the last two minutes of the pass. A casual glance through the rest of the RHC and LHC predicted strengths show this situation to be more common than rare.

However, the sum of the horizontally and vertically polarized signals are below 149 db only briefly at the start of the run and again at the end of the pass. Thus from the point of view of signal strength only, diversity has a clear advantage over circular based on predicted strengths and performance.

What improvement might one expect in improving the figure of merit of tracking? The greatest source of pointing error external to the equipment is that resulting from multipath-propagation effects. Stated quite

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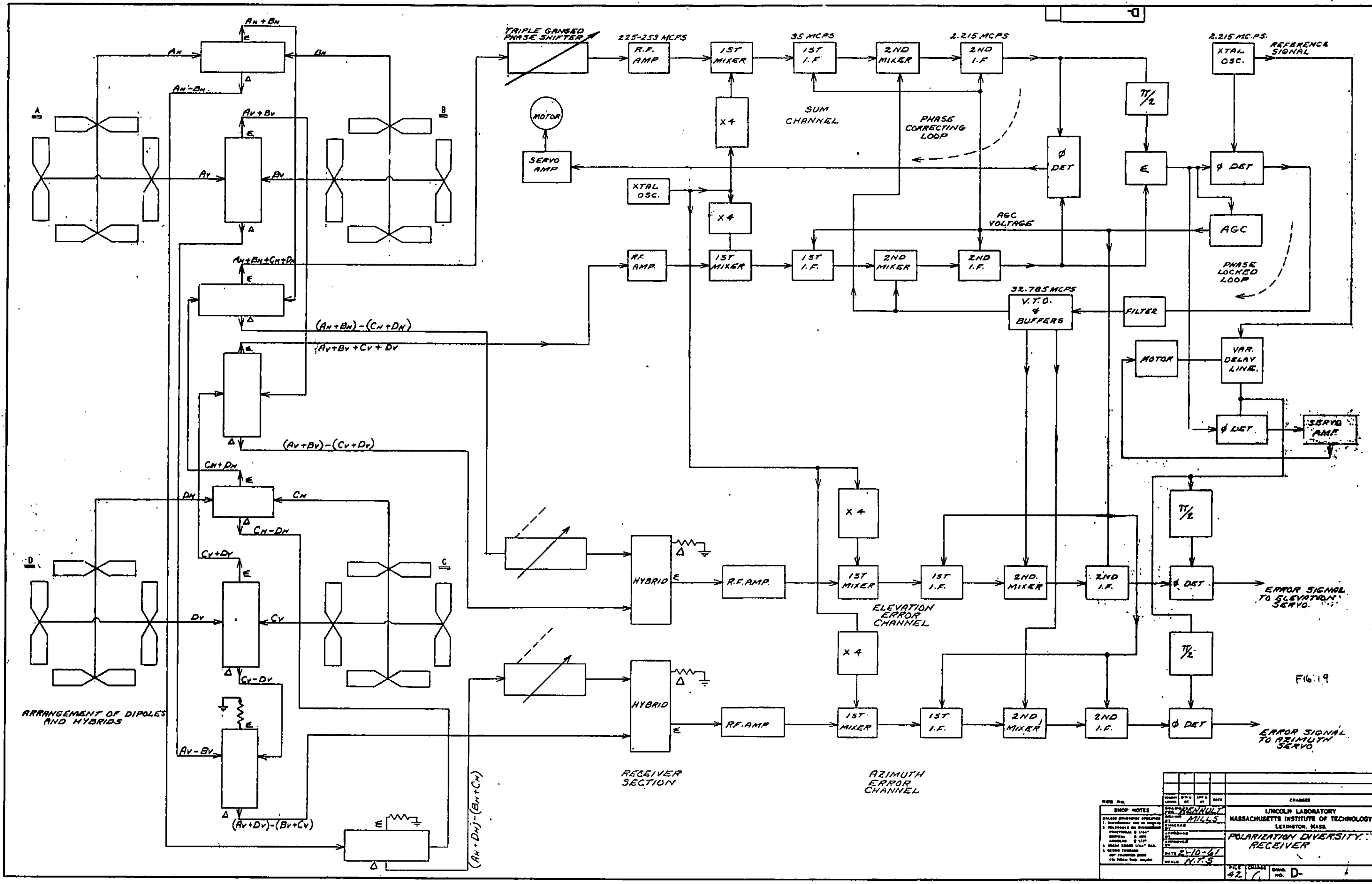
* R. Vacca, Allowable Space Loss for Acquisition Aid Receiver, 13 May 1960, 20-0039

simply the ray which enters the antenna by reflection from the sea appears to come from a different point in space than the direct ray. Obviously diversity is no panacea in this situation, for any system whose function it is to combine orthogonally polarized signals will happily combine both the direct and the reflected and perhaps aggravate the tracking problem. One possible area of improvement does suggest itself. Since the coefficient of reflection for vertically polarized waves drops rapidly for small angles above grazing the receiver could be programmed to utilize only vertically polarized signals for angles up to 10 to 15 degrees above the horizon at which point diversity combining could commence.

WBR/mhd


TRIPLE GANGED PHASE SHIFTER
225-259 MCPS
R.F. AMP
1ST MIXER
35 MCPS
1ST I.F.
2ND MIXER
2.215 MCPS
2ND I.F.
2.215 MC.P.S.
XTAL OSC.
REFERENCE SIGNAL
MOTOR
SERVO AMP
X4
SUM CHANNEL
PHASE CORRECTING LOOP
π/2
φ DET
ε
XTAL OSC.
AGC VOLTAGE
AGC
PHASE LOCKED LOOP
32.785 MCPS
V.T.O. & BUFFERS
FILTER
VAR. DELAY LINE
SERVO AMP
HYBRID
ERROR SIGNAL TO ELEVATION SERVO
ELEVATION ERROR CHANNEL
ERROR SIGNAL TO AZIMUTH SERVO
RECEIVER SECTION
AZIMUTH ERROR CHANNEL
ARRANGEMENT OF DIPOLES AND HYBRIDS
AH+BH
AH
BH
AH-BH
AV+BV
AV
BV
AH+BH+CH+DH
(AH+BH)-(CH+DH)
AV+BV+CV+DV
(AV+BV)-(CV+DV)
CH+DH
DH
CH
CH-DH
CV+DV
DV
CV
CV-DV
AV-BV
(AV+DV)-(BV+CV)
(AH+DH)-(BH+CH)
LINCOLN LABORATORY
MASSACHUSETTS INSTITUTE OF TECHNOLOGY
LEXINGTON, MASS.
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RENNHULT
MILLS
2-10-61
N.T.S
FILE 42
DWG. NO. D-

FIG. 19

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