PHYSICAL SCIENCE LABORATORY

NEW MEXICO STATE UNIVERSITY

University Park, New Mexico

A NON-AMBIGUOUS INTERFEROMETER SYSTEM

Wesley Joosten

Contract No. NAS 5-3318
3 September 1963

A NON-AMBIGUOUS INTERFEROMETER SYSTEM

1.0 HISTORY

The possibility of utilizing signals from three antennas instead of the conventional two for an SSD coarse interferometer was discussed as early as 1961 at Wallops. At that time, it was concluded that the idea was impractical due to the instrumentation problems and the receivers which were then in use.

Since then, the increasing demand for real time data from the SSD has spurred system developments in an effort to meet the real time data demand.

Since then, it has been found in the use of narrow band audio filters in the present SSD interferometer that quite good phase stability can be obtained. Also, a solid state VLF - IF receiver has been found to be feasible. Other VLF mixer and multiplier circuits have since been found to be practical and are now in use in the SSD.

Therefore, the three antenna interferometer is now reconsidered in the light of these recent developments.

2.0 RESULTS

Unambiguous coarse phase can be derived from a center offset antenna.

Both coarse and fine interferometer data are simultaneously available from the same set of instrumentation. No degradation of fine data is seen.

The accuracy of antenna location has been found to have much less effect on system error than the instrumentation electrical phase error.

Improved stability in the fine phase is expected because of the characteristic of the filters used.

The individual electronic circuits considered in the system have been investigated and found to be acceptable for this application.

Phase zero can be accomplished independently and fairly easily for the coarse and fine systems. Calibration cycles can be easily obtained.

The system sensitivity would be expected to approach -135 dbm on both coarse and fine. This should equal or exceed the sensitivity of the present system.

The system cost is not expected to exceed that of the present interferometer, and would quite probably cost less.

The system could conceivably be so constructed as to fit on the wall space presently used by the trombone board. No large amount of rack space would be needed.

Coarse angular measurement accuracy is shown to be sufficient for quick look data, and could conceivably be used for prime data.

Tape back-up of the interferometer could be accomplished if desired and trajectories could be computed on playback, as is now done with Doppler.

It would be necessary to move the SSD van to outside the antenna field.

It would be very desirable to use a straightforward AFC on the one local oscillator to eliminate Doppler phase error in all coarse and fine systems.

3.0 ANTENNA FIELD

One additional antenna, (M), would be located in about the middle of the field as shown in Fig. 1. The four other antennas would assume the nonsymmetrical position as shown, which is highly distorted to show the change in position as compared to a square array.

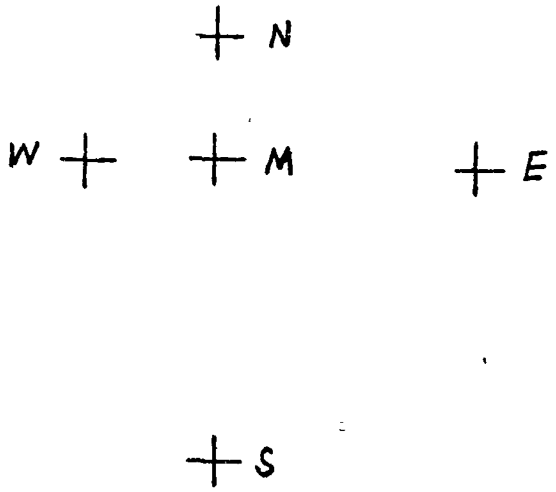

FIG. 1 - 5 ANTENNA ARRAY

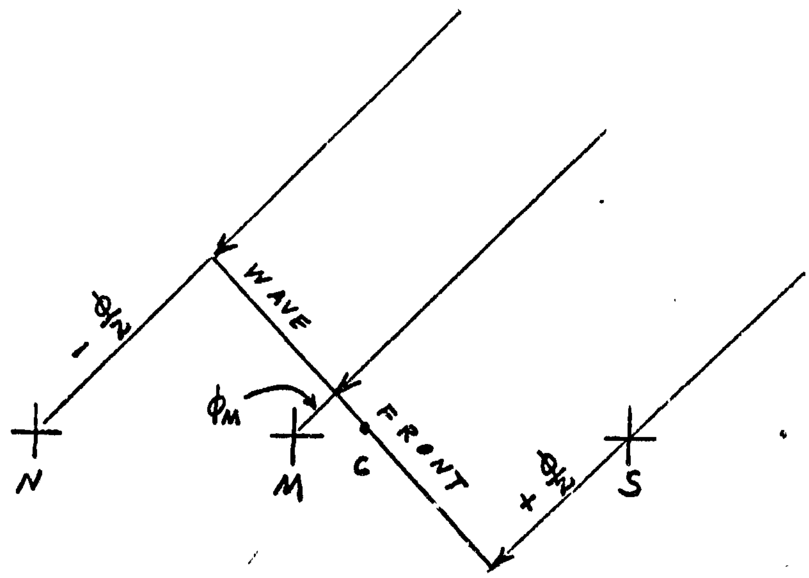

FIG. 2 - NMS AXIS

Fig. 2 shows one axis of the interferometer. The antennas are N (north), M (middle), and S (south). The array center is C. These three identical antennas drive three identical receiver systems, except for phase zero control to be discussed later. These receivers would be the 30 Kc IF type, driven from a common local oscillator.

The N to S spacing is $d = n\lambda$, the N to C spacing is the same as the C to S spacing and M to C spacing is $b = p\lambda$.

Using the array center C as reference, the phase is:

$$\phi_N = -\phi/2 \qquad (1$$

$$\phi_M = -\phi p/n \qquad (2$$

$$\phi_S = \phi/2 \qquad (3$$

4.0 THE GENERAL EQUATION

For any interferometer, the phase shift between the two antennas is, from Fig. 2:

$$\phi_{NS} = K\frac{\phi}{2} + L\frac{\phi}{2} = (K + L)\frac{\phi}{2}$$

where K is a multiplier in the south instrumentation and L is a multiplier in the north instrumentation.

In general, $L = K$, in which case:

$$\phi_{NS} = K\phi \qquad (4$$

When K is less than unity, then phase division results.

The total phase shift of an interferometer from horizon to horizon is known to be:

$$\phi_T = (2n)\ 360 \text{ electrical degrees} \qquad (5a$$

or in general:

$$\phi_T = 720\ Kn \text{ electrical degrees} \qquad (5b$$

From equations 1 and 2, the difference in phase of N minus M is, in general form:

$$\phi_{NM} = \phi\,(-K\tfrac{1}{2} + G\,\tfrac{p}{n}) \qquad (6$$

where G is phase multiplication associated with the M instrumentation.

From equations 2 and 3, the difference in phase of M minus S is, in general form:

$$\phi_{MS} = \phi\,(-G\,\tfrac{p}{n} - K\tfrac{1}{2}) \qquad (7$$

Adding equations 6 and 7 produces the fine phase:

$\phi_{SN} = K\phi$ the same as equation 4.

The difference of equation 6 minus 7 gives the coarse phase:

$$\Delta = \phi\, G\,(2\,\tfrac{p}{n}) \text{ electrical degrees} \qquad (8$$

Combining equation 5a and 8, we find the total coarse phase variation:

$$\Delta_T = 720\ G\ (2p) \text{ electrical degrees} \qquad (9$$

which has the same form as the fine phase in equation 5b.

Examination of equation 9 shows two interesting facts.

If any multiplying of phases is done, the total coarse phase variation is multiplied by the same amount. This implies that for a non-ambiguous output, the distance b must be decreased to compensate for any instrumentation frequency multiplying which produces a net phase multiplication.

The total phase variation is twice (compare equation 9 to equation 5) the variation from the standard interferometer. This implies that the distance b must be half that of a regular interferometer.

There are basically two types of phase detectors. Type A produces a saw tooth (flip-flop type) which essentially has no ambiguous points in 360 degrees. Type B produces a triangular wave (coincidence type) which has two ambiguous points in 360 degrees.

Solving for b ($b = p\lambda$) in equation 9,

$$b = \frac{\Delta_T}{1440\ G}\lambda \qquad (10$$

5.0 DATA ERRORS

Letting G = 1,

for type A, $\Delta_T = 360$, $b = \lambda/4$, and

for type B, $\Delta_T = 180$, $b = \lambda/8$.

If a system can be used so that G = 1, then for type A, allowing one ambiguity (two cycles of output), $\Delta_T = 720$ and $b = \lambda/2$.

For the two most probably cases, the spacings are $b = \lambda/4$ and $b = \lambda/2$. For the type A phase detector, the errors in measurement angle can be calculated for an assumed survey distance measuring error of 0.01 feet (0.1 inch).

b/λ	1/4	1/2	wave length
b	3.34	6.68	feet

For an error of 0.1 inch, the error in angle measurement would be:

At Horizon	4.5	2.2	deg. space arc
At 45° Elev.	6	3	min. space arc
At 70° Elev.	2.5	1.5	min. space arc

The error near zenith per electrical degree phase error for $b/\lambda = 1/2$ would be 15 minutes of space arc and for $b/\lambda = 1/4$, 30 minutes of space arc. The accuracy of phase zero setting by electrical phase adjustment would be determined by the type of recorder or other output. It is felt that the time stability of the instrumentation could be held to one degree or less using the solid state equipment.

The use of two cycles of type A phase detector output, $b = 1/2\lambda$, would mean that data would be correct only above 60° elevation angles in the slant plane.

A review of past missile trajectories shows that the elevation is greater than 60° from about 0.5 seconds to about 200 seconds flight time.

This is roughly equivalent to about 5,000 to 7,000 feet altitude ascending. The altitude at 60° descending is dependent on the impact distance. Past trajectories indicate the altitude to be between 230 K ft and 290 K ft.

Fig. 3 shows the electrical output of a type A phase detector for $b/\lambda = 1/2$ (two cycle) and $b/\lambda = 1/4$ (one cycle).

This does not appear to be an objectionable price for real time data.

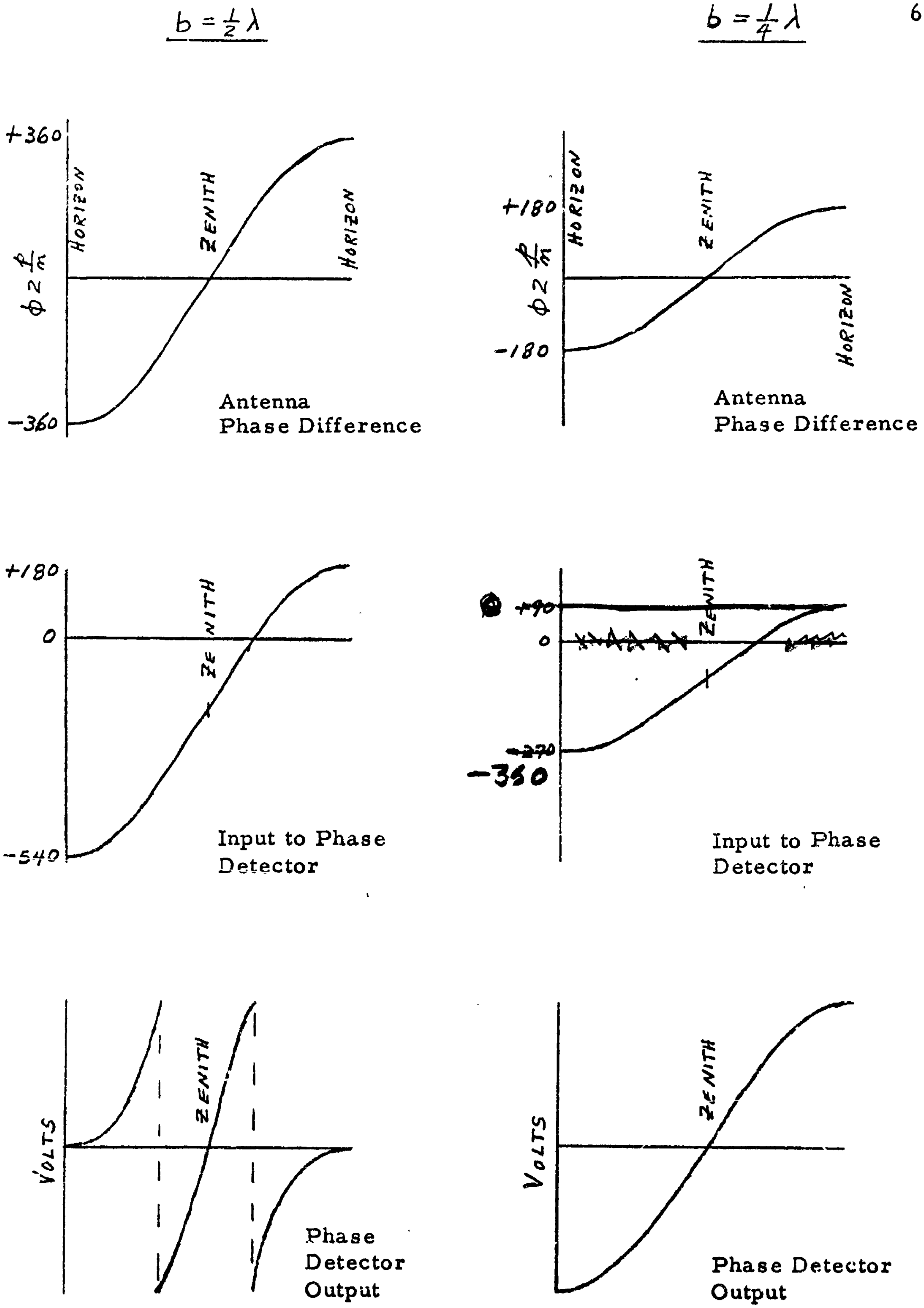

FIG. 3 - PHASE DETECTOR OUTPUT

5.0 ELECTRONIC SYSTEM

In order to convert the phases into the non-ambiguous electrical output the problem begins; that is, how to obtain the data without the additional phase shift expected from the instrumentation.

Since the mixing (additive and subtractive), and multiplying would be done at audio frequencies, a bread board was set up and various mixing frequencies were used. The leakage and distortion was measured in each case.

F_1 (Kc)	F_2 (Kc)	$F_1 - F_2$ (Kc)	Leakage
90	60	30	-34 db
45	30	15	-35
60	45	15	-42
75	60	15	-44
45	15	30	-20 (not useable)

A bread board multiplier was also set up and tested for leakage.

X2	-50 db
X3	-42 db
X5	-36 db

This data (plus the equations in Part 4) lays the ground rules for designing the instrumentation.

These ground rules are:

1) No multiplying of phases is desirable.
2) Electrical phase errors of all types must be kept to a minimum.
3) Only certain frequencies can be mixed.
4) Only difference mixers are allowed.

After seemingly endless attempts to arrive at an instrumentation which obeys the rules and yields the desired results, the system was arrived at as shown in Fig. 4.

The frequencies for this diagram are as follows:

D = Doppler plus roll
N = $73.6 - D + \phi_N$
M = $73.6 - D + \phi_M$
S = $73.6 - D + \phi_S$
NIF = $73.6 - LO - D + \phi N + n$, where IF = $73.6 - LO$
MIF = $73.6 - LO - D + \phi M + m$
SIF = $73.6 - LO - D + \phi S + s$

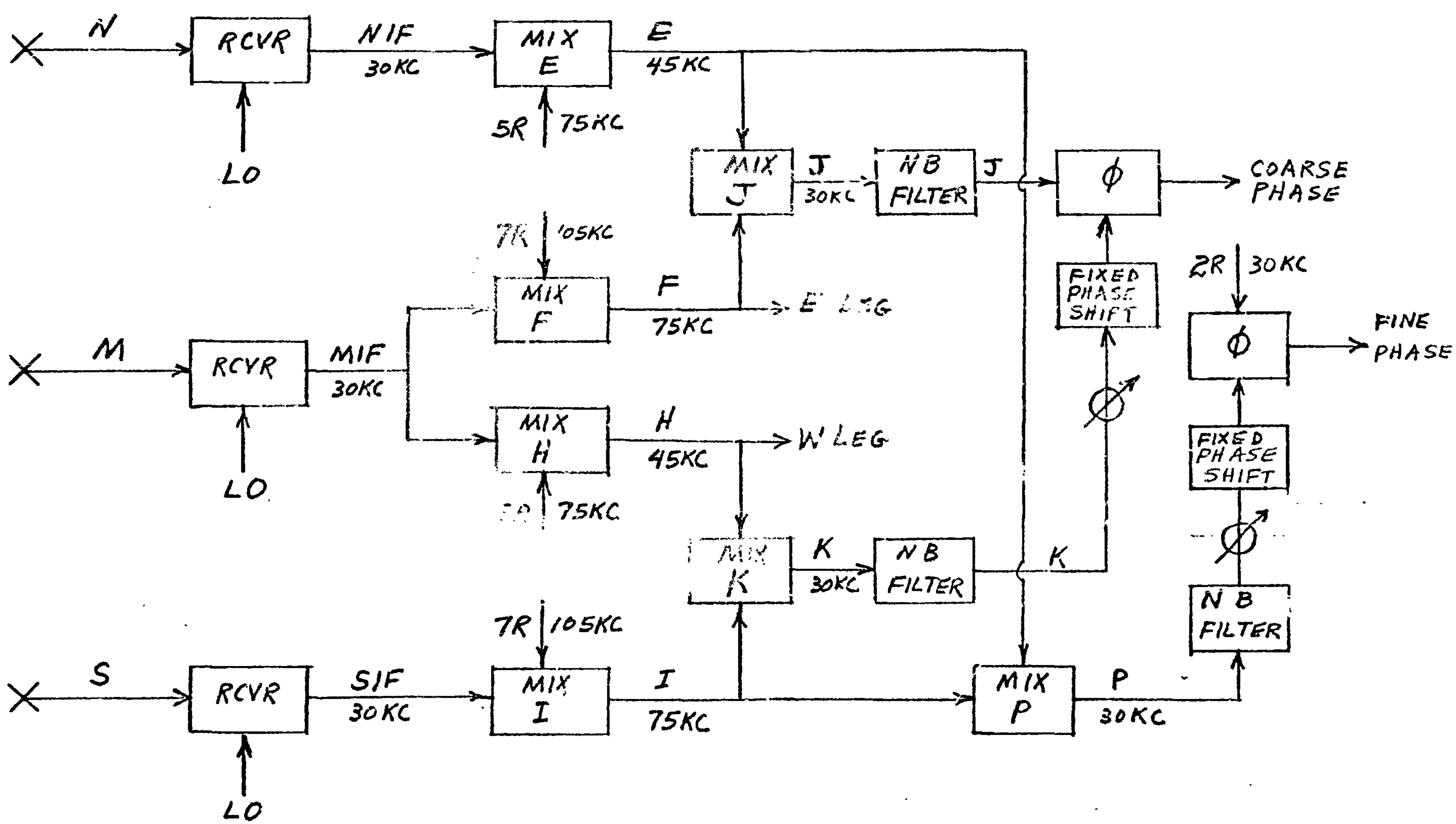

FIG. 4 - SYSTEM BLOCK DIAGRAM

where n, m and s are receiver phase shifts.

$$
\begin{aligned}
R &= 15 \text{ Kc}, \; 5 R = 75 \text{ Kc}, \; 7 R = 105 \text{ Kc}, \; 2 R = 30 \text{ Kc}. \\
E &= 5 R + e - IF + D - \phi_N - n \\
F &= 7 R + f - IF + D - \phi_M - m \\
H &= 5 R + h - IF + D - \phi_M - m \\
I &= 7 R + i - IF + D - \phi_S - s
\end{aligned}
$$

where e through i are the mixer filter phase shifts.

$$
\begin{aligned}
J &= F - E = 2 R + \phi_N - \phi_M + f - e + n - m + j \\
K &= I - H = 2 R + \phi_M - \phi_S + i - h + m - s + k
\end{aligned}
$$

where j and k are the narrow band filter phase shifts.

$$\text{Coarse } \phi = K - J = 2 \phi_M - \phi_S - \phi_N + (e - h) + (i - f) + (2 m - s - n) + (k - j)$$

The differential phase shifts in parenthesis have been grouped by identical circuits. Ideally, each one of these parentheses would be zero. Actually, this never is realizable. The Doppler shift effects (n, m, s, e, f, h, i) could be expected to produce undesirable errors. Therefore, it would seem most practical to utilize a crystal controlled local oscillator driven by a very slow AFC loop such that in the absence of a correcting signal, the local oscillator would be in the middle of the pass band. Discriminator circuits are now available which could accomplish this.

Assuming zero differential phase shift:

$$\text{Coarse } \phi = 2 \phi_M = -2 \phi \frac{p}{n} \qquad (11$$

because

$$-\phi_S - \phi_N = -(\frac{\phi}{2} - \frac{\phi}{2}) = 0$$

Thus, equation 11 is the same as equation 8 for the case G = 1.

The fine phase is also derived.

$$P = I - E = 2 R + i - e + \phi_N - \phi_S + n - s + p$$

The fine phase is:

$$\text{Fine } \phi = P - 2 R = \phi_N - \phi_S + (n - s) + (i - e) + p$$

The differential phase shifts would again be ideally zero. The phase shift p would be caused by the frequency of the interferometer phase and would not cause appreciable error after 10 seconds of flight. It could possibly amount to a few minutes of arc during powered flight. Because of the fixed component audio filters, and the use of AFC, the phase stability of this system will be markedly superior to the present SSD.

The sensitivity of the system would be determined by the narrow filters which could be made as narrow or even more narrow than the present SSD system of 100 to 200 cycles.

6.0 PHASE ZERO ADJUSTMENT

Since the system would be used to furnish coarse and fine phase simultaneously, a method of zero phase adjustment is necessary which will allow entirely independent phase zero adjustment of any coarse or fine phase channel.

Instead of the present two-output phase matched power divider, a five output phase matched power divider (three two-outputs would suffice) would be used to simultaneously feed five signals to the five receivers N, S, E, W and M.

Instead of the costly trombones now in use, audio (or video) RC phase shifters would be used. These are quite applicable here.

Examining the phases present at P, it is seen that at phase zero, $\phi_N - \phi_S$ ideally is zero. If this is not true because of a residual phase in the system, a phase shift introduced here would correct this. The same would hold for the E - W leg.

Examination of the phases at J, K and coarse phase show that a shifter placed in the K (or J lead) would accomplish the same result. A shifter in the same place in the W leg would allow the E - W coarse phase to be zeroed.

Note that these are all independent adjust nents!

7.0 CYCLE CALIBRATION

The cycle calibration for phase could be easily accomplished by injecting a 30 Kc signal into the IF of M and S and a 30,001 signal into the IF of N. The relative stabilities of these two signals would be quite easy to obtain from audio crystal oscillators. This ease of obtaining calibrating signals would be a marked system improvement over the present SSD.

8.0 TAPE BACK-UP

If for any reason, it was desired to record the phase data for post firing playbacks, a suitable recorder could be used to record the data.

If all fine and coarse data would be desired, five channels would be necessary, that is the IF outputs of the five receivers (phase zero would be accomplished on playback). If only coarse data were required, signals J and K (from both the NS and EW systems) would be recorded, a total of four channels. Likewise, four channels would be needed if only fine data were required. No recorded reference would be necessary since this could be locally generated for playback operation.

9.0 INSTALLATION

The equipment would take the form of small modular units. It would be not too useful to place these small units on standard chassis in standard racks. These modular units could much easier be attached to a plate on the wall, such as the present trombone board. Rack space would be utilized only for items such as the local oscillator, recorder, etc. This would represent a significant saving in space, not to mention a savings in rack cost and installation.