(5) 627 400 (1)

NRL Report 5055

(6)

PROJECT VANGUARD REPORT NO. 23
MINITRACK REPORT NO. 3, RECEIVER SYSTEM

(9) INTERIM REPT.

(10)

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Project Vanguard

419878

December 6, 1957

NAVAL RESEARCH LABORATORY
Washington. D.C.

$3.60

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PREVIOUS REPORTS IN THE MINITRACK REPORT SERIES

NRL Report 4995, "Project Vanguard Report No. 18, Minitrack Report No. 1; Phase Measurement," by C. A. Schroeder, C. H. Looney, Jr., and H. E. Carpenter, Jr., July 26, 1957

NRL Report 5035, "Project Vanguard Report No. 21, Minitrack Report No. 2; The Mark II System," by R. L. Easton, September 12, 1957

CONTENTS

ABSTRACT

The Minitrack system is designed to acquire and track artificial earth satellites which Project Vanguard will attempt to place in orbits as part of the United States' effort in the International Geophysical Year. The satellites will be provided with 108-Mc signal sources to illuminate the tracking stations. These sources will radiate sufficient power to permit the desired tracking accuracy. The Minitrack system functions as a radio interferometer, providing the direction cosines of a satellite's position vector with respect to the tracking stations at any given instant during a transit.

The receiving equipment must amplify the signals obtained from the satellites via interferometer antenna pairs, with the introduction of a minimum of noise and differential phase shift, and by means of a unique double-local-oscillator principle translate this intelligence to a frequency readily adaptable to phase measurement.

This report describes, with schematics, the five units comprising a single rf channel, i.e., a pair of front ends, a signal adder, a combined i-f amplifier, a special local oscillator, and a calibration source.

PROBLEM STATUS

This is an interim report on one phase of the problem; work is continuing.

AUTHORIZATION

NRL Problem A02-18
Projects NR 579-000 and NR 579-001

Manuscript submitted October 18, 1957

PROJECT VANGUARD REPORT NO. 23
MINITRACK REPORT NO 3, RECEIVER SYSTEM

GENERAL DESCRIPTION

The Minitrack system of radio angle tracking operates on the interferometer principle. The radio path lengths from a signal source to each of a pair of antennas are compared, and the difference between these path lengths is determined by measuring the phase difference between the signals at the antennas and subsequently at the output of the receiver.* The function of the receiver is to amplify these signals and translate them to a frequency (500 cps) readily adaptable to phase measurement, without causing undue signal deterioration by the addition of receiver noise or differential phase shift.

The antenna baselines are oriented north-south and east-west, forming a cross; the distance from the center of the array to the outer antennas is 250 feet. Coaxial 7/8-inch-diameter transmission lines connect the antennas to the receivers (Fig. 1) located in a trailer near the center of the array. The loss in these lines, 0.4 db per 100 feet at the 108-Mc signal frequency, is not considered sufficiently signal-degrading to outweigh the convenience of placing the preamplifiers at the trailer instead of at each antenna. The lines are buried in trenches to minimize differential phase shift between antennas of a pair due to temperature changes altering the electrical line lengths.

The 7/8-inch lines terminate in a transmission line terminal box located near the trailer, where a transition is made from 7/8-inch line to RG9/U solid dielectric cable. These cables enter the trailer and connect directly to the receiver system. The transmission line terminal box contains pressurizing equipment including valves and gauges for each line and a tank of compressed nitrogen to pressurize the lines in order to prevent moisture from altering the line characteristics. The inputs to the receiver front ends are connected to coaxial switches which select either the antenna, via the transmission lines, or a calibrating signal produced by a calibration source.

The front end employs two tubes as grounded-grid amplifier stages, a GL 6299 planar triode and a 6AN4. The circuits associated with these tubes, together with a balanced mixer incorporating two 1N82A crystals and a signal-dividing hybrid, comprise the front end. The signal from one of a pair of antennas is amplified with a minimum of added noise, after which it is translated in the balanced mixer to the lower of the two i-f frequencies, 11.2950 Mc. The translated signal is then connected to two type-N connectors through a signal-dividing hybrid junction. This signal division permits a single antenna channel to be used for two phase measurements.

The other signal comprising a phase comparison channel is likewise amplified in a front end but is translated in frequency to the higher of the two first i-f frequencies, 11.2955 Mc, by means of the special local oscillator. This local oscillator provides signals separated in frequency by 500 cps, this reference 500-cps signal being locked in phase with the 500-cps standard signal obtained from the time standard equipment.

*A description of the phase measurement portion of the Minitrack System is given in "Project Vanguard Report No. 18, Minitrack Report No. 1; Phase Measure," NRL Report 4995, July 26, 1957

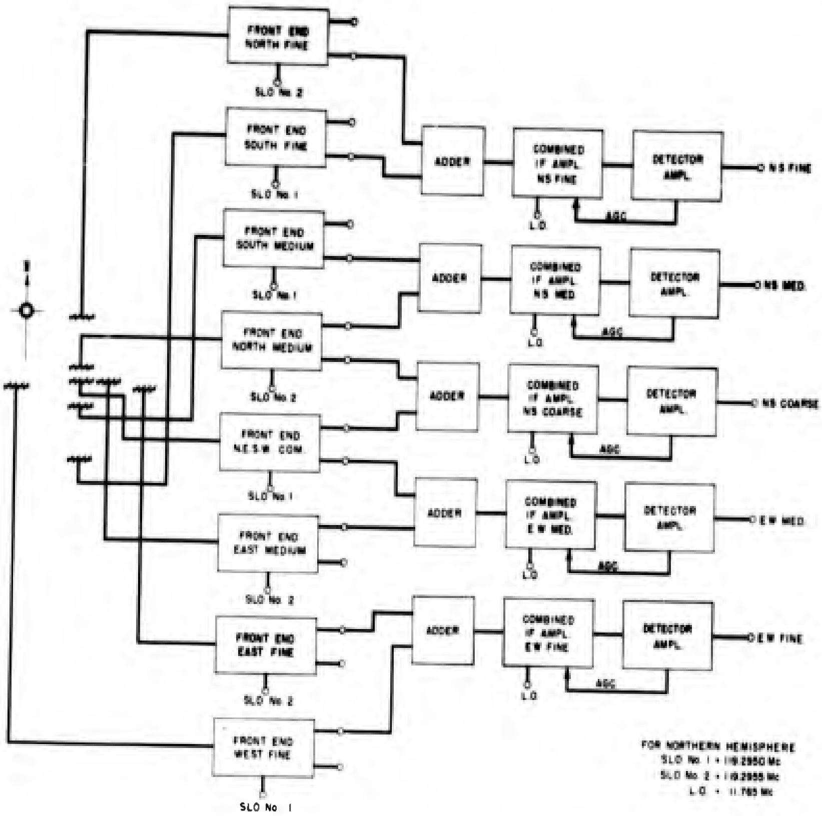

Fig. 1 - Minitrack receiver block diagram

The output signals from the two front ends, separated in frequency by the 500-cps difference in their respective local oscillator frequencies, are combined by means of a signal-adder hybrid junction, the output of which is connected to the combined i-f amplifier unit. The major portion of the gain required to increase the level of the exceedingly weak antenna signals to the 12 volts peap-to-peak necessary for adequate phase measurement is obtained in the i-f amplifier; only the minimum gain necessary for a good noise figure is permitted in the front end units. This is because of the desirability of a minimum of signal processing prior to combination and, therefore, a minimum of differential phase shift between the signals. The chance of differential phase shift occurring after signal combination is greatly reduced in that the phase characteristics of the i-f amplifier affect both signals almost identically, owing to their close proximity in frequency.

Approximately half the required receiver gain is supplied by the first i-f amplifier, after which the combined signals are translated to the second i-f frequencies of 470 and 470.5 Kc by means of the second local oscillator signal. The second i-f stages provide the required pre-detection bandwidth of about 12 Kc, along with the necessary amplification.

The combined signals, which at this point have frequencies of 470 and 470.5 Kc, respectively, are applied to a detector which reconstitutes the 500-cps signal. The phase relationship existing between this receiver output 500-cps signal and the 500-cps reference signal is identical to the phase relationship that exists originally between the signals received at the two antennas. A cathode follower provides a low-impedance output for this

500-cps signal. It is then fed by means of coaxial cable to the phase measurement equipment, wherein the phase difference between it and the 500-cps reference signal is determined and presented in both analog and digital form.*

Although the usable input signals applied to the system range between -135 dbm and -60 dbm, the phase measurement equipment requires that the maximum amplitude deviation of the receiver output signal voltage be 8 db. In order to meet these requirements the receivers are provided with automatic-gain-control (AGC) circuitry by which a negative dc voltage proportional to the output signal level is applied to the grids of two of the i-f stages, varying the transconductance and hence the gain of these stages. The AGC servo-loop is provided with sufficient gain to meet the specifications described above.

Internal calibration of the receiving system is provided by the calibration source which consists of a 108.000-Mc crystal-controlled oscillator, a variable attenuator, and coaxial switches which connect the signal directly to the front ends. Periodic checks may be made with this unit to determine any change in the phase characteristics of the receiving equipment.

RECEIVER FRONT END ANALYSIS

The relatively weak signal to be radiated by the satellite makes it necessary to provide the Minitrack receiver with a good noise figure. At a satellite radiating level of 10 milliwatts, for the maximum anticipated range of 1500 miles, the power received by a ground antenna with a gain of 40 is approximately 10^{-15} watts, or -120 dbm.

The available power of the thermal noise present at the input of a receiver is equal to kTB, where k is Boltzmann's constant, T is the temperature in degrees Kelvin, and B is the bandwidth in cycles per second. Numerically $kTB = 4 \times 10^{-17}$ watts (-134 dbm) for a 10-Kc bandwidth – a reasonable value for the pre-detection bandwidth of the receiving equipment associated with the Minitrack system. Thus, the signal-to-noise ratio, S/N, at the input terminals of the receiver is 134 - 120 = 14 db.

The noise figure is defined as

$$F = \frac{kTB + N}{kTB},$$

where N is the noise contribution of the receiver referred to the input terminals.

The total noise power, referred to the input terminals, is

$$W_n = kTB + N = F \times kTB.$$

Thus, the signal-to-noise ratio at the second detector of the receiver described above is (14 - F) db. A pre-detection signal-to-noise ratio of 10 or 12 db is desirable for phase measurements requiring accuracies of less than one electrical degree; correspondingly, a noise figure of 3 db or less is equally desirable.

The preamplifier section of the front end described here consists of two grounded-grid stages and is capable of producing a noise figure of less than 2.5 db with good bandwidth, stability, and gain characteristics. The tube chosen for the first stage is the

*See footnote on page 1

GL 6299 planar triode, which features a transconductance of 12,000 micromhos and a transit-time conductance of 30 micromhos at 108 Mc. The high power gain available in the first stage permits the use of the less expensive, commercially available 6AN4 grounded-grid tube for the second stage without sacrifice of performance

A schematic of the front end is given in Fig. 2. The operation of the circuit is as follows. The 50-ohm signal source is connected to the input terminals and for noise figure considerations is transformed to about 420 ohms by C1 and L1. The plate circuit of the first stage is resonant at the signal frequency by means of C6 and L4. The auto-transformer, L4, transforms the output impedance of the first stage to a value which provides a good second-stage noise figure and the desired bandwidth. Numerically this value is in the region of 500 ohms. L10 resonates with the stray capacitance existing at the plate of the 6AN4 and, through its turns ratio, transforms the high output impedance of V2 to a value which is compatible with the input impedance of the mixer hybrid composed of L12 through L15 and C14 through C16. The 108-Mc signal is mixed with the 119.2950- or 119.2955-Mc local oscillator signal and by means of the two 1N82A mixer crystals is translated in frequency to 11.2950 or 11.2955 megacycles, the first i-f frequency.

The identical outputs from each side of the balanced mixer are added directly by means of C17 which, together with C18, serves to block the dc crystal currents. These currents are monitored with milliammeters mounted on the control panel.

The signal is divided at this point by the dividing hybrid composed of inductors L20 through L22 and capacitors C19 through C21. This enables each front end and its accompanying antenna to be used in two phase-measuring channels, and also matches the mixer output impedance (about 150 ohms, depending upon the local-oscillator insertion level) to the 50-ohm line.

This front end 50-ohm output impedance is transformed upward at the input to the i-f amplifier by a factor of about 10, resulting in noise-free gain and reducing the contribution, if any, of the first i-f stage to the total receiver output noise.

The expression for the noise figure of a single grounded-grid stage is

$$F = 1 + \frac{G_{1\,eq}}{G_s} + \frac{\beta G_t}{G_s} + \frac{R_{eq}(Y_s + G_t)^2}{G_s}$$

where

$G_{1\,eq} = G_1 + G_D$,

G_1 = the input circuit losses,

G_D = the output circuit losses,

G_t = the transit time conductance,

G_s = the transformed source conductance,

β = a constant with a value of approximately 5,

R_{eq} = the equivalent noise resistance of the tube, approximately $2.5/g_m$ for most triodes,

$Y_s = G_s + G_1 + jY_1$, and

jY_1 = the susceptance presented by the input network to the input terminals of the tube.

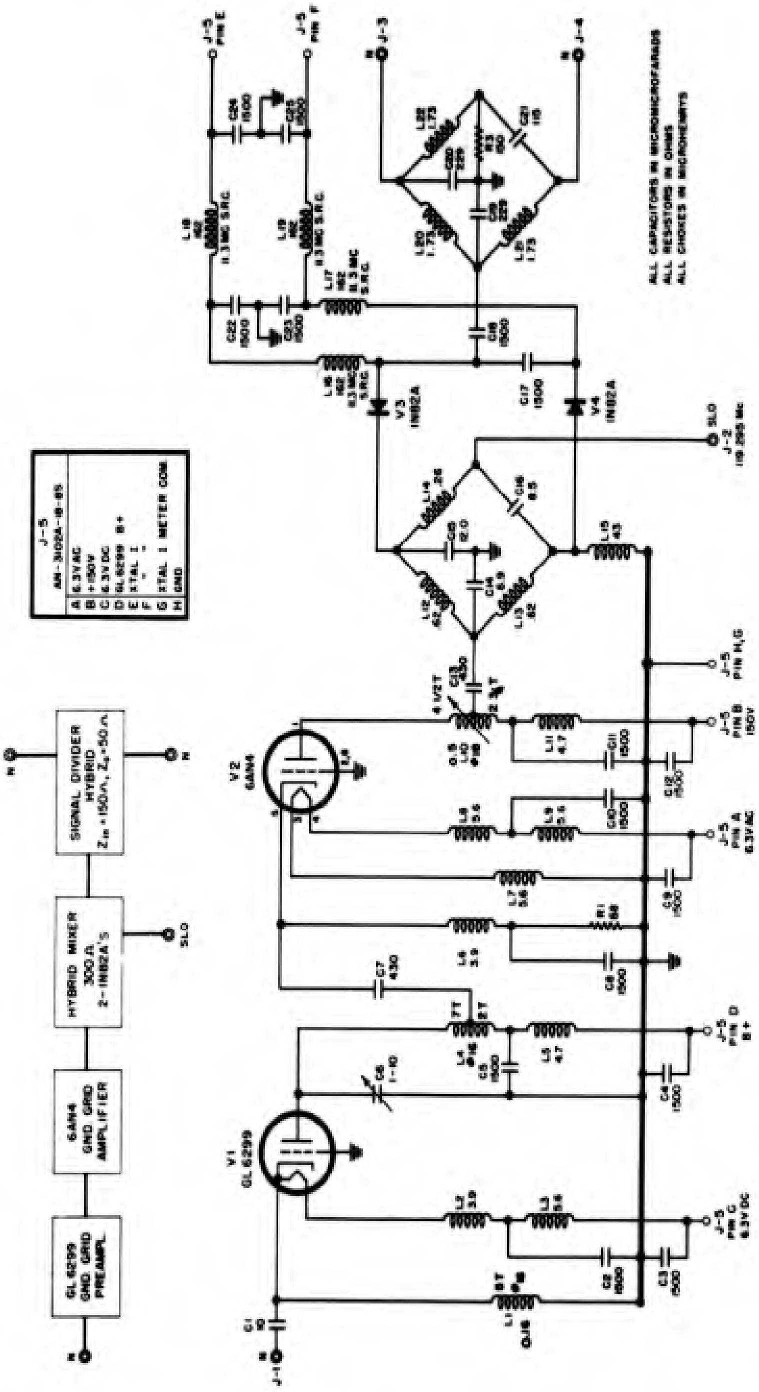

Fig. 2 - Minitrack front end

Several conclusions may be made immediately concerning this expression:

1. The noise figure is a minimum when $jY_1 = 0$; this occurs when the admittance presented to the input terminals of the tube is purely conductive.

2. All coil losses or other losses represented by $G_{1\ eq}$ increase the noise figure.

3. G_t and R_{eq} play a prominent part in the value of the noise figure, making the choice of tube type highly important.

4. G_s appears in both the numerator and the denominator, indicating that an optimum value exists. Of course, this is usually the prime consideration for optimum noise figure design and cannot be overemphasized.

The optimum value of the transformed source conductance, $G_{s\ opt}$, may be determined by differentiating the noise figure expression with respect to G_s and setting the first derivative equal to zero. The resulting expression, after dropping the insignificant terms, is:

$$G_{s\ opt} = \sqrt{\frac{G_{1\ eq} + \beta G_t}{R_{eq}}}.$$

The corresponding value of the optimum attainable noise figure is

$$F_{1\ opt} = 1 + 2\sqrt{(G_{1\ eq} + \beta G_t) R_{eq}}.$$

The overall noise figure of the amplifier is expressed as

$$F = F_1 + \frac{F_2 - 1}{W_1} + \frac{F_3 - 1}{W_2} + \ldots + \frac{F_n - 1}{W_{n-1}},$$

where W is the stage power gain. This series is rapidly convergent for all practical cases, so usually only two terms are significant, and sometimes only one.

These equations may be applied to the amplifier described above. Its circuit parameters are:

	GL 6299	6AN4
G_t	30 μmhos	160 μmhos
R_{eq}	125 ohms	250 ohms
g_m	12,000 μmhos	10,000 μmhos
r_p	9,600 ohms	7,000 ohms

$$f_o = 108\ \text{Mc}: \quad \omega_o = 6.78 \times 10^8\ \text{radians/sec.}$$

In obtaining the first-stage noise figure, it will be assumed that the Q of input and output coils is 200. Then the loss conductance of the input coils is

$$G_1 = \frac{1}{Q\omega L1} = 40 \times 10^{-6}\ \text{mhos}.$$

The loss conductance of the output coil plus the losses from second-stage loading is

$$G_D = \frac{1}{Q\omega L4} + N^2 G_{in\ 2} = 250 \times 10^{-6} \text{ mhos},$$

where N is the turns ratio of L4, and $G_{in\ 2}$ is the input conductance of the 6AN4 stage. Also,

$$G_{1\ eq} = G_1 + G_D = 290 \times 10^{-6} \text{ mhos}.$$

Then

$$G_{s\ opt} = \sqrt{\frac{G_{1\ eq} + \beta G_t}{R_{eq}}} = 1900 \times 10^{-6} \text{ mhos},$$

and

$$F_{1\ opt} = 1 + 2\sqrt{(G_{1\ eq} + \beta G_t) R_{eq}} = 1.5 = 1.8 \text{ db}.$$

The values for the optimum transformed source conductance and the resulting noise figure of the second stage obtained by the same method as applied to the first-stage calculations are

$$G_{s\ opt\ 2} = 2200 \times 10^{-6} \text{ mhos};$$

$$F_{opt\ 2} = 2.1 = 3.2 \text{ db}.$$

The tap point on L10 is positioned so that the input impedance of the mixer loads the plate tank to a Q which provides the desired bandwidth. If the tap point were adjusted to provide a match between the output impedance of the stage and the input impedance of the mixer, the resulting gain would be maximized, but at the cost of a severe restriction on the bandwidth. A good compromise for the bandwidth of this stage is 7 Mc (Fig. 3). This bandwidth requires a Q of approximately 15.

The available power gain, W_1, of the first stage is

$$W_1 = \frac{G_s}{G_D + \dfrac{G_s}{r_p\, g_m}} = 9.$$

The overall noise figure, F, is

$$F = F_1 + \frac{(F_2 - 1)}{W_1} = 2.1 \text{ db},$$

which amounts to a second-stage contribution of only 0.3 db to the noise figure.

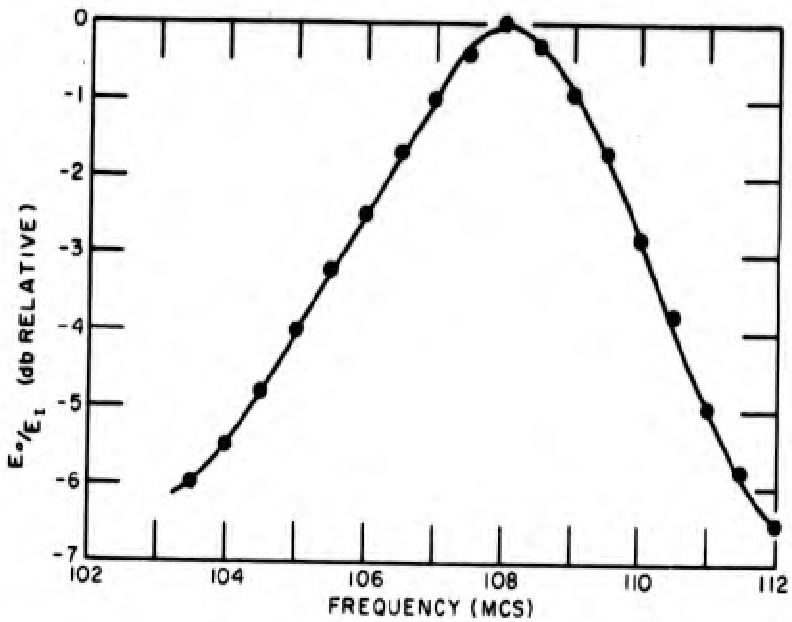

Fig. 3 - Front end bandpass characteristics

The bandwidth of the coupling network between the two grounded-grid stages is quite narrow when the optimum G_s is presented to the input terminals of the 6AN4. However, if the tap point on L4 is raised from the optimum point the bandwidth will be increased with only a relatively small resulting increase in the noise figure. The input resistance of the 6AN4, which serves to load the tank circuit through the turns ratio of L4, thus controlling the bandwidth, has the value

$$R_{in} = \frac{G_L + g_p}{g_m G_1} = 175 \text{ ohms},$$

where G_L is the load conductance at the plate circuit of the 6AN4, with a value of about 200×10^{-6} mhos. This value, when reflected through the turns ratio of L4, provides a a damping resistance, R_q, equal to $N_4^2 \times R_{in}$, where N_4, the turns ratio of L4, is about 5:1. Thus $R_q = N_4^2 \times R_{in} = 4400$ ohms,

$$G_q = \frac{1}{R_q} = 230 \times 10^{-6} \text{ mhos}.$$

This conductance and the output conductance of the stage, 30×10^{-6} mhos, provide the total damping. The Q of this stage, considering the total conductance and capacitance associated with this stage, is about 15. This Q provides a bandwidth of about 7.2 Mc.

The input transformer, consisting of C1 and L1, is simple but extremely effective at the impedances involved. The impedance R_o presented by the source to the input terminals of the front end is 50 ohms and must be transformed to about 520 ohms as determined by the preceding calculations for $G_{s\ opt}$. R_o, L1 and C1 form a series resonant circuit, R_o being the series damping resistor. The relationship between the series damping resistor and the parallel damping resistor is $R_s R_p = (1/\omega C)^2 = (\omega L)^2$

where

R_s = the series resistance = R_o, and

R_p = the parallel resistance = $1/G_{s\ opt}$

Thus, the source conductance $1/R_o$ is transformed to $G_{s\ opt}$.

The values of C1 and L1 which provide the desired transformation are easily found:

$$\frac{1}{\omega C1} = \sqrt{R_s R_p} = \sqrt{50 \times 520} = 160,$$

$$C1 = \frac{1}{6.78 \times 10^8 \times 160} = 9.2\ \mu\mu f.$$

The optimum value determined experimentally is 10 $\mu\mu f$.

It is generally desirable to utilize as high an L/C ratio in the plate circuits as possible in order to obtain the maximum bandwidth. It is never wise to supply additional damping by means of resistive loading, because a value of resistance capable of supplying this damping causes deterioration of the noise figure. Any additional damping appears as an increase in the circuit losses represented by $G_{1\ eq}$ and is to be avoided.

Several beneficial results are obtained by using the balanced mixer for the heterodyne frequency conversion:

1. The signal and local oscillator are combined with negligible insertion loss.

2. A high degree of isolation is obtainable between the signal and local oscillator input; this is necessary for reducing crosstalk between receivers employing the same local oscillator.

3. Noise from the local oscillator is cancelled.

The hybrid form of balanced mixer is the lumped-constant counter part of the familiar coaxial-line junction commonly called the "rat race," or of the "magic-T" formed from a three-dimensional waveguide junction. It is composed of four reactive matching networks with the required image resistances, three having positive 90-degree phase shifts and one having a negative 90-degree phase shift. The networks are connected in the form of a ring, the input impedance of one serving to satisfy the image impedance requirements of its neighbor. The relationship between the image resistances and reactances is:

$$\frac{1}{\omega C} = \omega L = \sqrt{2\ R1\ R2},$$

where R1 and R2 are the input and output image impedances for each network.

When the individual networks are connected, the parallel reactances at the junctions combine to form a single equivalent component or, in many cases, disappear entirely owing to an antiresonant condition. Although most applications call for equal image resistances, hybrids using transforming sections can be made, i.e., $R1 \neq R2$, in which case the impedances at the terminals of the hybrid are not equal. The hybrid used in this particular mixer has three 300-ohm terminals and one 50-ohm terminal. The crystals, which at this frequency have an impedance of approximately 300 ohms, terminate two of the junctions while the 50-ohm terminal properly loads the local oscillator line. The fourth terminal, at 300 ohms, is the load on the second-stage output through the turns ratio of the auto-transformer, L10.

Because two of the antennas in the Minitrack system are each employed in two separate phase measuring channels, at least two of the front ends must have provisions for two signal outputs each. In the interest of equipment uniformity it was decided to provide all front ends with two outputs. The hybrid divider, L20 through L22 and C19 through C21, provides this dividing function admirably, the usual hybrid junction attributes of low insertion loss and high isolation making this network ideal for this function. It also serves to match the input impedance of the mixer, about 150 ohms, to the 50-ohm line.

The experimentally measured characteristics of the front end unit, which conform very closely to the computed values, are as follows:

Voltage gain . . . . 17 db

Bandwidth . . . . 4.5 Mc

Noise figure . . . . 2.5 db (nominal)

The 17-db gain is sufficient to override the noise contribution of the i-f amplifier following. The bandwidth is a compromise between (1) phase stability requirements and (2) reduction of the chances of intermodulation products due to received signals spaced at frequency intervals equal to the first i-f frequency, and heterodyning at the first mixer.

SIGNAL ADDER ANALYSIS

The signal adder unit (Fig. 4) is a device for adding or combining the outputs of two front ends for subsequent channeling to the combined i-f amplifier. Classical receiver operation requires precisely controlled signal levels prior to combining in order that the two combined signals interfere completely as indicated by their apparent 100-percent modulation. The two level-set adjustments per channel are rough approximations to constant-impedance variable attenuators (Fig. 4). With the variable resistor set at its midpoint, the input and output impedances are close to 50 ohms. At other settings the impedance remains fairly constant, and the resulting 5- or 6-db variable attenuation is more than adequate to insure balance of the signals from any two front ends.

The hybrid itself is an ordinary 50-ohm symmetrical unit calculated at 11.295 Mc for both isolation and power division. In general, if components are selected somewhat carefully it is not necessary to adjust this hybrid, the isolation and power division being usually better than 35 db and 1 db, respectively.

The following signal additions are made in this hybrid to provide five phase-comparison output channels from the eight input antennas:

1. East fine plus west fine to form the east-west fine channel.

2. North fine plus south fine to form the north-south fine channel.

3. East medium plus common to form the east-west medium channel.

4. North medium plus south medium to form the north-south medium channel.

5. North medium plus common to form the north-south course channel.

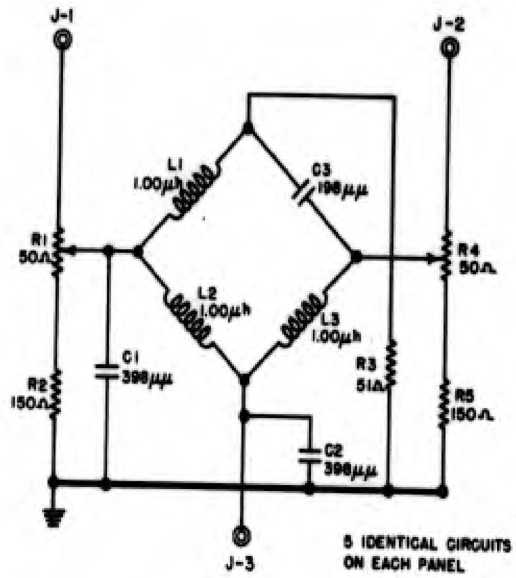

Fig. 4 - Signal adder unit

This nomenclature is correct for stations in the northern hemisphere. For stations in the southern hemisphere the entire antenna field is rotated 180° with the following results: For the five systems the antennas are connected for their true nomenclature. East medium becomes west medium and the west medium combined with common forms east-west medium. South medium and north medium have interchanged in such a way that north-south course must be obtained from the common and the new south medium antennas because of the antenna spacing. This change results in the center of the station being north of the common antenna.

COMBINED I-F AMPLIFIER ANALYSIS

The output of each adder in the adder unit is connected through an RG 9/U cable to the receiver input terminal where its impedance is stepped up by means of the tuned input transformer. The input impedance looking into the receiver unit from the input terminal is roughly 50 ohms in spite of the apparent termination of 150 ohms presented by the resistor R1. The input impedance to the transformer, with its Q of about 40, is sufficient to provide the 50 ohms input impedance when paralleled by the 150-ohm resistance.

The following two stages (Figs. 5 and 6) are conventional i-f amplifiers, with gains of about 30 db each, and bandwidths of about 200 Kc. The second stage has AGC voltage applied to its grid through a 47-kilohm resistor which is bypassed by capacitors. It should be noted that the cathode and screen voltage of each stage controlled by the AGC voltage is independent of the bias conditions. Thus the bias, hence the transconductance and gain of each of these stages depends upon the known relationship between grid voltage and transductance; this relationship is logarithmic, with the result that a plot of input signal level in db versus AGC voltage is fairly linear. In addition the expression $\ln(\Delta E_g/E_g)$ which appears in the servo-loop gain expression is fairly constant, relaxing the requirements on the filtering and phase shifting networks in the AGC loop.

The high output impedance is stepped down by the autotransformer in the plate circuit of V2 in order to provide a better match between the 300-ohm input impedance and the second mixer hybrid. The input impedance of the hybrid loads the tank circuit through the 8:1 turns ratio of the autotransformer. The second mixer-converter is identical in operation

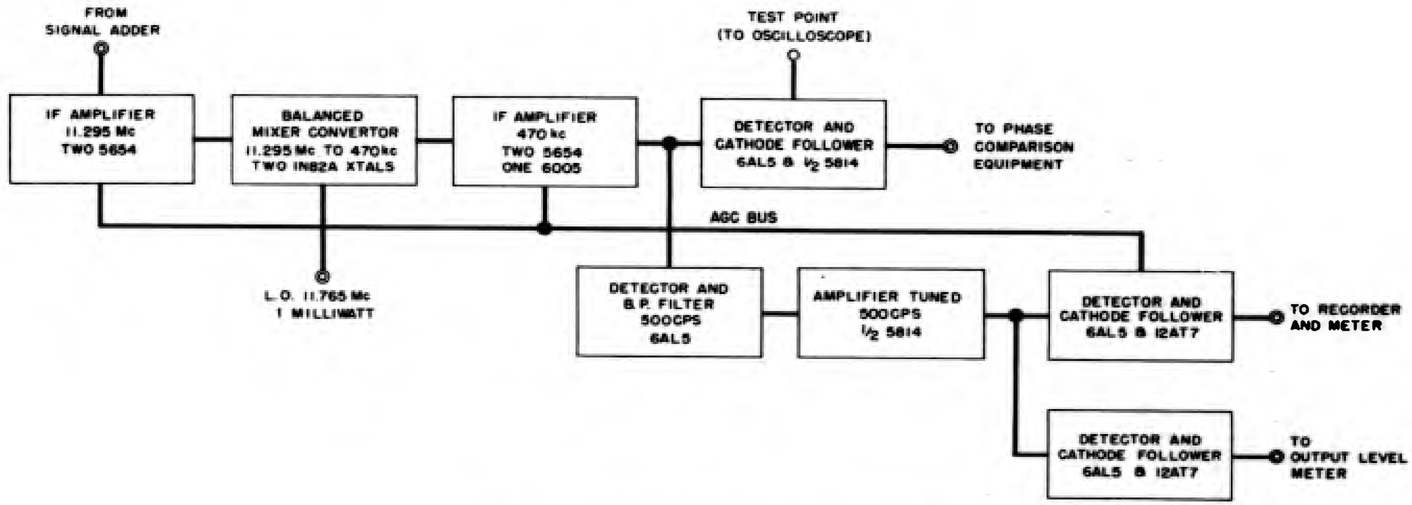

Fig. 5 - I-F amplifiers, block diagram

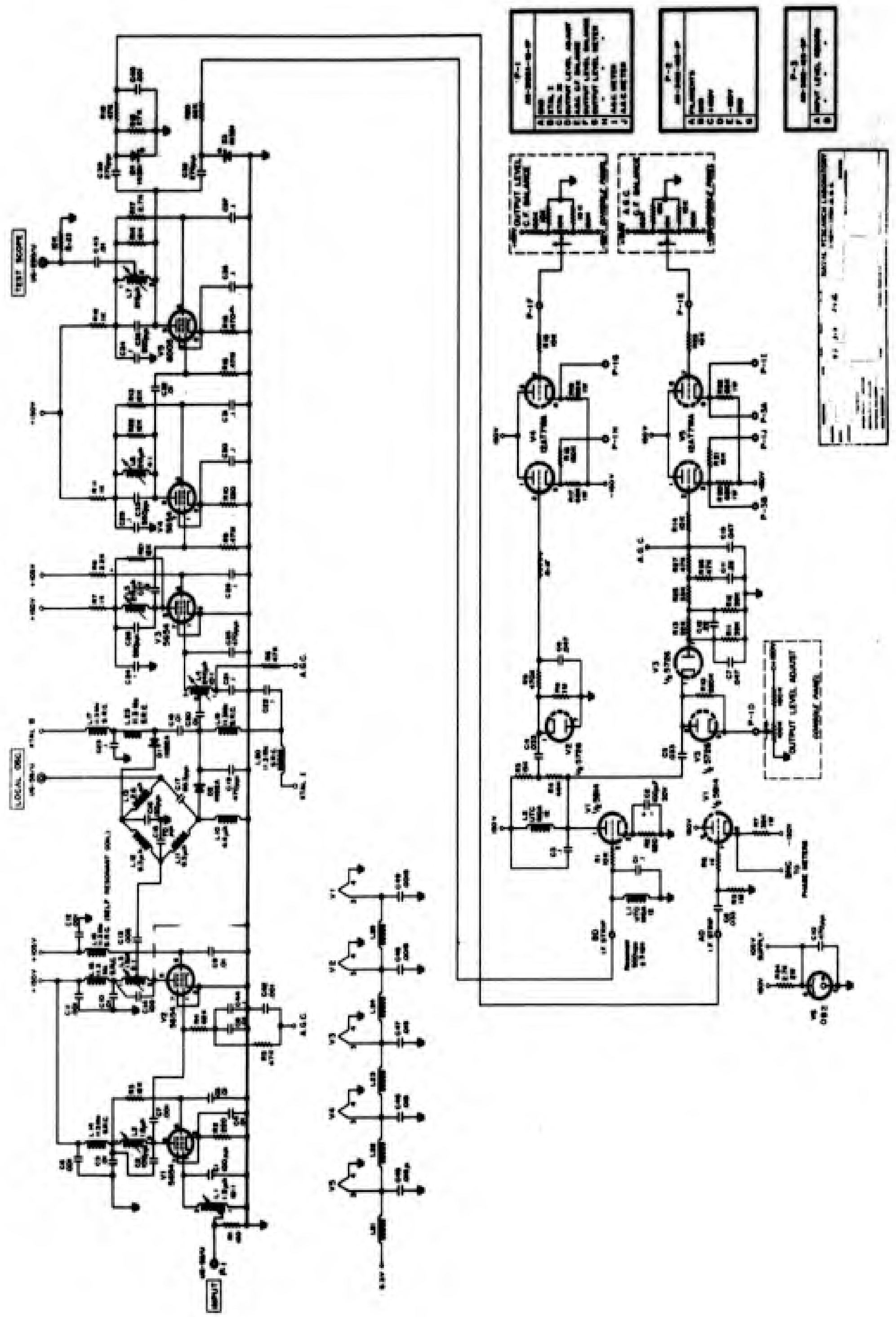

Fig. 6 - I-F amplifiers, schematic diagram

to its facsimile in the front end except that the frequency conversion is from 11.295 Mc to 470 Kc. The input terminal and the two crystal terminals each have an impedance of 300 ohms while the local oscillator terminal loads the local oscillator line with its characteristic impedance of 50 ohms.

The output of the second mixer converter is connected to the grid of the first 470-Kc amplifier through the autotransformer, L4, considerable gain being obtained by this simple expedient. The relatively low output impedance of the converter, approximately 150 ohms, thus loads the tuned autotransformer through the 10:1 turns ratio.

The tuned autotransformer, and the tank circuits in the plate circuits of the following three stages, are essentially the sole pre-detection bandwidth-limiting networks. These stages are synchronous single-tuned circuits affecting the bandwidth according to the equation

$$\% \text{ bandwidth} = \sqrt{2^{1/n} - 1}$$

where n is the number of tuned circuits. When n equals 4 this factor equals 0.44. The bandwidth of each stage is about 27 Kc; therefore, the overall pre-detection bandwidth is slightly less than 12 Kc.

The other stage of the pair of stages to which AGC voltage is applied is V3. The following stage, V4, is a conventional i-f amplifier. The stage following V4 is an i-f power amplifier utilizing the 6005 tube, which closely matches the required 15-kilohm load resistance, thus providing a high output voltage swing with good linearity. The normal peak-to-peak output signal level at this point is only about 20 volts. However, when a high-level signal is applied to the input terminals this stage must be capable of providing far more than the normal swing, with good linearity. In the presence of a weak signal the accompanying noise may be many time the normal output voltage. Although the noise constitutes the majority of the output voltage, it must be amplified without distortion if the signal, which later has the noise removed by selective filtering, is to remain undistorted. Amplitude distortion, even of the order of a few percent, will cause differential phase shifts amounting to several degrees; this indicates the importance of distortionless signal processing and the need for a power amplifier.

The output i-f signal is detected by two crystals, one operating on the positive portion of the wave, the other on the negative portion. One detected signal is fed through a cathode follower to the phase measurement rack where its phase is compared to that of the 500-cps reference signal obtained from the special local oscillator unit. The other 500-cps detected signal is amplified by a 500-cps bandpass amplifying stage with a 500-cps resonant tank in the grid circuit. The resulting bandwidth is about 50 cps centered at 500 cps; this bandwidth is comparable to that obtained by subsequent filtering on the 500-cps signal fed to the phase measurement rack. The final bandwidth is about 10 cps. Thus the voltage which is fed to the AGC detector has somewhat the same signal-to-noise ratio as has the signal in the phase measurement racks after filtering. Therefore the AGC circuit, which cannot differentiate between signal and noise, will maintain the signal plus its noise, in a 50-cps bandwidth, approximately constant so that the phase measurement signal will likewise remain nearly constant in amplitude regardless of signal level.

The AGC detector utilizes both halves of a double diode tube in a voltage doubler circuit. The shunt detector member of this team has its cathode returned to a variable positive voltage which constitutes the output level adjust control, otherwise known as the AGC delay voltage. For an explanation of the operation of this circuit, let it be assumed that the control is shifted more positive. Then the signal voltage presented to its plate circuit is insufficient to cause the first and the second diode to conduct. The AGC voltage will then rise toward zero, which will increase the tranconductance of the two stages controlled by the AGC voltage. Hence, the gain of these stages will increase, resulting in a sufficient increase in signal voltage applied to the AGC detector circuit to cause diode conduction and the resulting gain control. The network in the plate circuit of the series AGC detector is a filter and a phase-correcting network which is necessary to prevent the AGC loop from oscillating. This tendency of the AGC loop to oscillate became quite a problem in developing the system; the severe phase shifts encountered at various points in the system, coupled with the extremely rigid filtering requirements on the AGC bus, posed many difficult questions. In addition, the change in the slope of the transconductance with respect to grid bias or AGC voltage of those tubes controlled by the AGC voltage became yet another factor in the AGC loop oscillations. At very high input signal levels, and hence large negative values of AGC voltage applied to the grids of the AGC controlled amplifiers, some 5654 tubes showed a tendency to cause oscillations around the loop. Probably the transconductance changes vary rapidly with AGC voltage in this region for some tubes.

A portion of the 500-cps signal voltage at the output of the 500-cps amplifier is detected with respect to ground and is therefore proportional to the output signal level. For monitoring purposes the AGC voltage and output signal level voltage activates meters. Cathode followers are needed in order to lower the impedance of the detected signal source for this purpose. The duo-triode cathode follower used for this application has several benefits over a single triode unit. The second unit provides a balancing voltage from a low-impedance source. Changes in plate voltage or heater voltage affect both units in the same manner so that the meter reading is relatively independent of such fluctuations. Also, in the event that either the positive 150 volts or negative 150 volts should fail or not be applied concurrently, the meter is not damaged, there being no great potential difference across it as there would be if the meter return were grounded as in a single-stage cathode follower. In addition to the meter circuit, the AGC cathode follower also supplies AGC level information to the recorder.

The noise figure of this unit with a typical 5654 tube in the first stage is about 9 db. Receivers with noise figures in this area make a very small, perhaps 0.1 db, contribution to the overall system noise figure. There are type 5654 tubes, however, which are extremely noisy, and if by chance one of these is put in the first stage of the receiver system, noise figures may deteriorate from a normal 2.5 db to as much as 10 db. Therefore some selection of the receiver first-stage tube is advisable. The receiver characteristics are shown in Figs. 7 through 12.

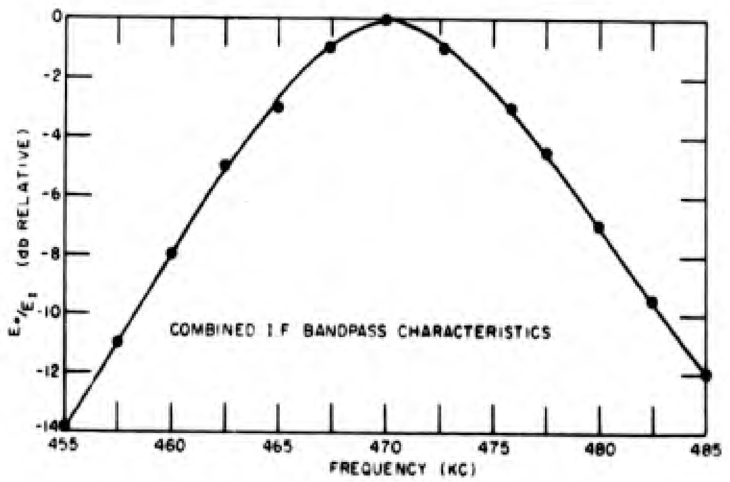

Fig. 7 - Combined i-f bandpass characteristics

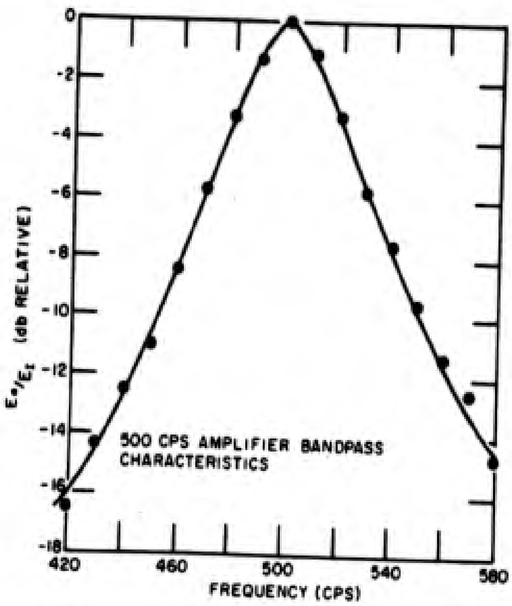

Fig. 8 - 500-cps amplifier bandpass characteristics

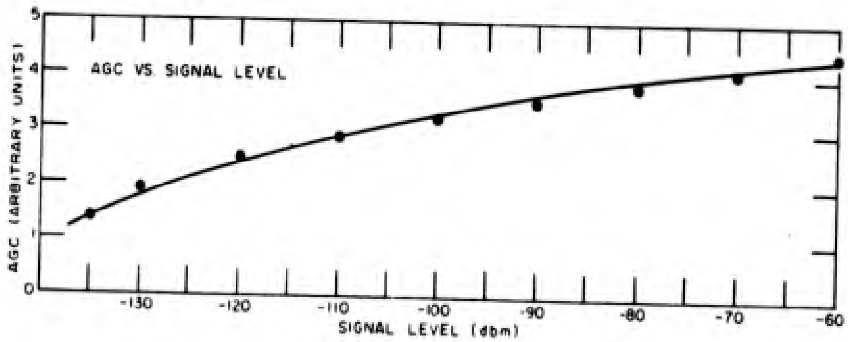

Fig. 9 - AGC vs signal level

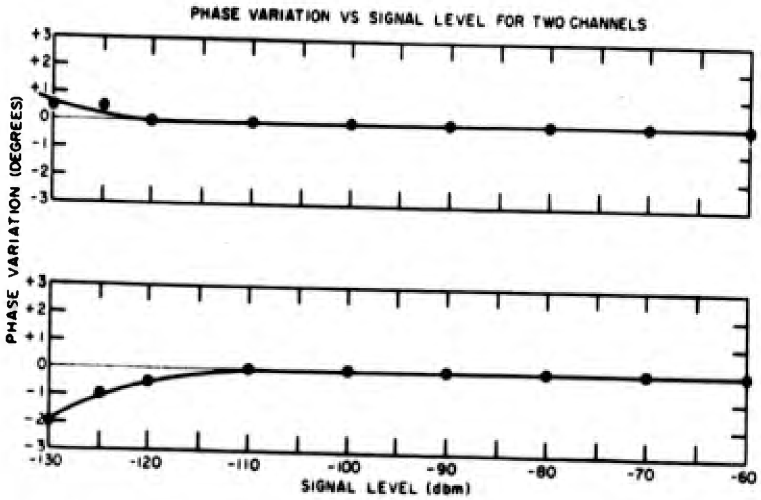

Fig. 10 - Phase variation vs signal level for two channels

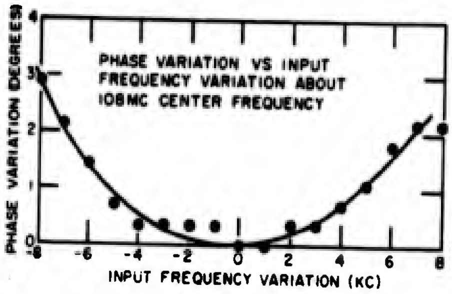

Fig. 11 - Phase variation vs input frequency variation about 108-Mc center frequency

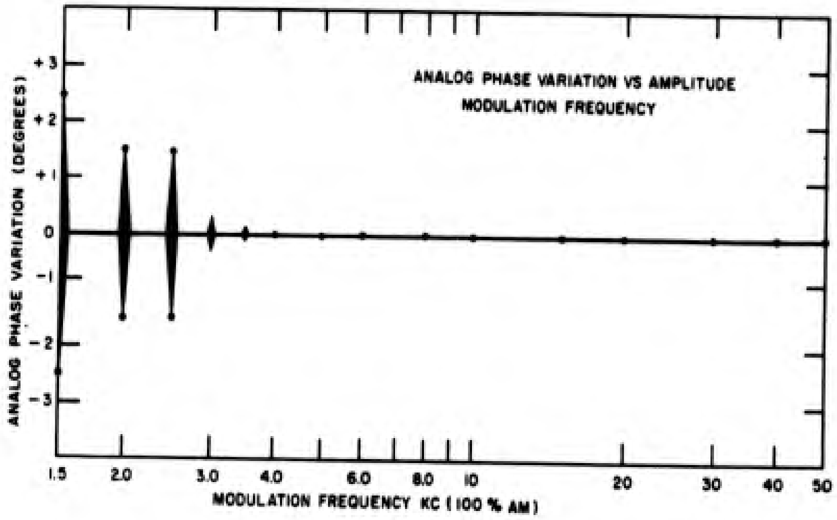

Fig. 12 - Analog phase variation vs amplitude modulation frequency

SPECIAL LOCAL OSCILLATOR ANALYSIS

The special local oscillator unit (Figs. 13 and 14) provides four types of signals. They are:

1. 119.2950 Mc at a power level of 1 milliwatt for local oscillator insertion to four of the 8 front ends.

2. 119.2955 Mc at a power level of 1 milliwatt for local oscillator insertion to the remaining four front ends.

3. 11.7655 Mc at a power level of 1 milliwatt for local oscillator insertion to the five second mixers in the i-f amplifiers.

4. The 500-cps reference signal is produced by beating the 119.2950-Mc and 119.2955-Mc oscillator outputs and amplifying the result, which is subsequently connected to the phase measurement equipment and serves as the reference signal for the actual phase measurement.

The only input signal is the 500-cps standard which is provided by the time standard unit and to which the 500-cps reference is locked in phase.

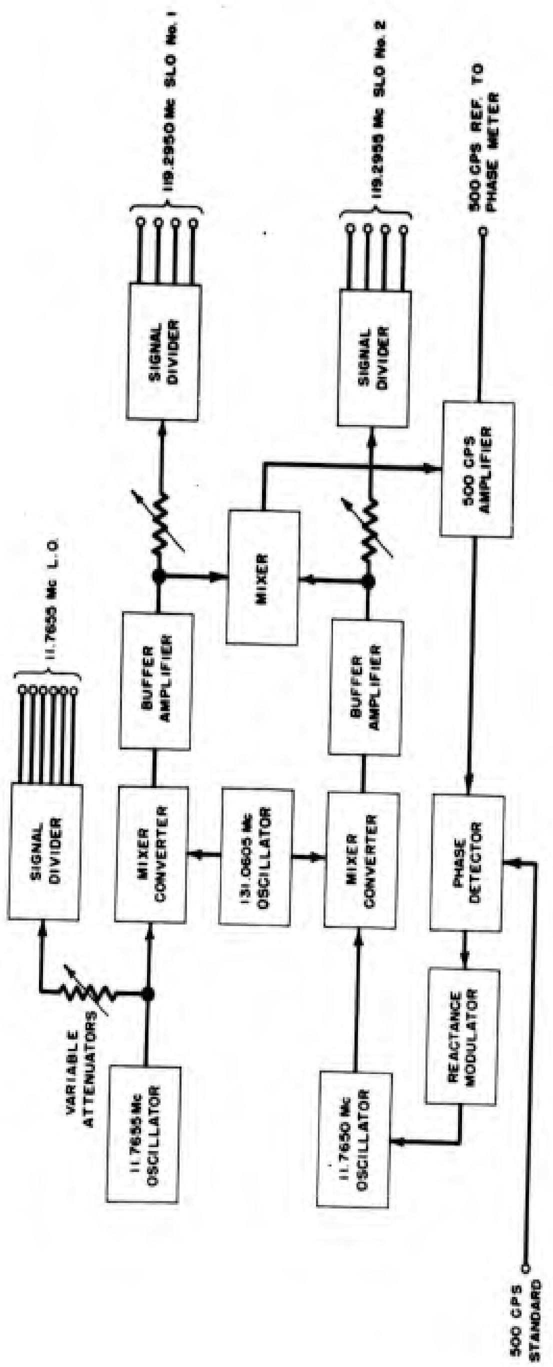

Fig. 13 - Special local oscillator unit, block diagram

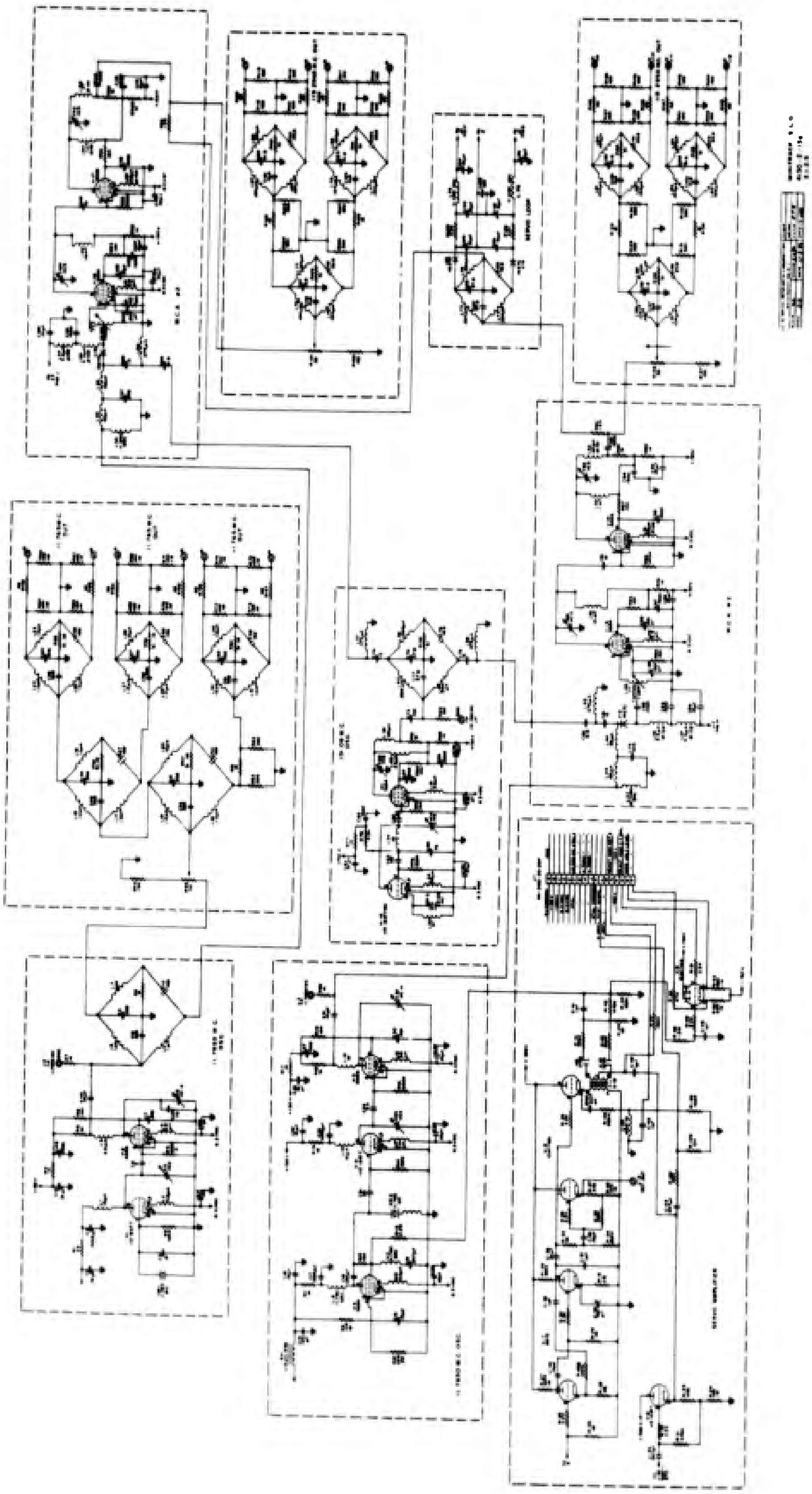

Fig. 14 - Special local oscillator unit, schematic diagram

The two phase-locked signals are first produced at 11.7650 and 11.7655 Mc after which they are translated in frequency to 119.2950 and 119.2955 Mc by means of the 131.0605-Mc oscillator common to both channels. Incidentally, one of these 11.7655-Mc oscillators provides the five (and one spare) outputs at that frequency for the second local oscillator function.

The phase locking is accomplished in the following manner. Assuming that the 119.2950-Mc and 119.2955-Mc outputs are functioning, the two signals are mixed in a balanced mixer hybrid and, by means of the two 1N270 diodes, the 500-cps difference frequency is produced; it is then amplified and fed to a phase discriminator. The phase of this signal is compared to the phase of the 500-cps standard, an error voltage being produced proportional to their deviation from quadrature relationship. This error signal is connected to the grid of a reactance tube, thus controlling the gain of that stage and hence the frequency of the 11.7650-Mc oscillator associated with the reactance tube. A shift in frequency of the uncontrolled oscillator results in the same shift at the translated frequency (119.295 Mc), so that the beat frequency between 119.2950 and 119.2945 Mc tends to shift. This tendency to shift, however, results in an error signal, produced by the phase discriminator, which pulls the frequency of the controlled oscillator, through the reactance tube, causing it to parallel the drift in the first oscillator. Thus, the phase relationship between the 500-cps standard and 500-cps reference changes only slightly because the high servo-loop gain is sufficient to reduce relatively large frequency swings to only a few degrees, i.e., a fraction of a cycle per second. Generally, the phase deviation between the standard and reference signals, after sufficient warm-up time has elapsed, is only 2 or 3 degrees, which puts the phase stability of the network in the region of one part in 10^{10} . However, even this small error is intolerable with regard to the accuracy desired in the phase measurement system. This error is prevented by the simple expedient of comparing the receiver output signal to the reference signal instead of to the standard signal.

The 11.7655-Mc oscillator is the simplest type of tuned-plate tuned-grid oscillator; the crystal serves as a parallel-resonant tank in the grid circuit and the oscillations are sustained by feedback through the interelectrode capacitance of the tube from the tuned plate circuit. This oscillator is extremely stable, its warm-up drift being only a few hundred cycles (Fig. 15), and it has a random short-term variation of only a few cycles. The output of this stage, V1, is coupled to a 6005 power amplifier, V2, which furnishes approximately 100 milliwatts to a divider hybrid, L5 through L6 and C11 through C13. This hybrid in turn supplies power to the second mixer local oscillator hybrid dividers, and to the translator mixer where the frequency is translated to 119.2950 Mc. The five hybrid dividers in the 11.7655-Mc section, and the output hybrid just discussed, are identical and symmetrical 50-ohm hybrids calculated at the second i-f frequency of 11.295 Mc. It should be noted that, although the power division takes place at 11.765 Mc, they are not calculated at that frequency because their primary function is to provide isolation between outputs in order to decrease the deleterious effects of crosstalk between receivers which might occur at the i-f frequency. However, in spite of the fact that the power division occurs at a frequency other than the optimum, the frequency response with regards to power division is sufficiently broad to permit power divisions within 1.5 db at 11.765 Mc and isolation greater than 40 db at 11.295 Mc. A 3-db pad, R61 through R63, compensates for the insertion loss in the associated hybrid, thus providing equal power levels to each 11.7655-Mc output. A rough approximation of a constant-impedance variable pad is provided by R54, R55. It is, of course, impossible to obtain constant impedance with only one variable element; nevertheless the match is not too far off, and near the center of the variable range it is quite good. The range in attenuation is approximately 6 db. The pi networks at each output constitute a 6-db attenuator. This provides an increase in isolation of 12 db through the power divider and also terminates the output hybrids independently of the relatively poor match of the second mixer. Thus the necessary isolation is obtained in the output hybrids which might otherwise decay because of the possibility of a poor match.

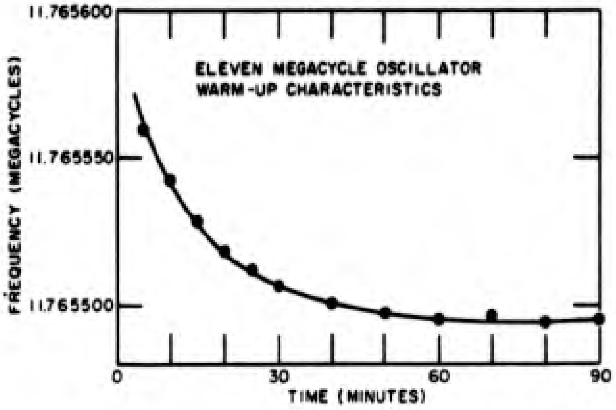

Fig. 15 - 11.7655-Mc oscillator warm-up characteristics

The 131.0605-Mc oscillator and buffer unit, V10, V11, serves to translate the 11-Mc phase-locked oscillator outputs to the 119.295-Mc first local oscillator frequency. This 131.0605-Mc bridged-T crystal oscillator relies on the negative transfer characteristics of the LC pi network (C145, L34, and C69 together with the plate-to-grid capacitance of the tube) to establish proper feedback conditions. These values also include the stray capacitance that exists at the grid and plate of the tube. The oscillator frequency is controlled by the crystal because of the high cathode degeneration existing at frequencies other than the low series-resonant frequency of the crystal, together with the steep phase characteristic in the vicinity of resonance. The warm-up characteristics of this oscillator are due to the temperature characteristics of the crystal. The warm-up characteristics of a typical oscillator are shown in Fig. 16. The network in the plate circuit of the buffer amplifier transforms the high output impedance of the stage to provide an output impedance of 50 ohms. The power is divided at this point by a hybrid junction which is designed for maximum isolation at 119.295 Mc. Much as in the case of the dividing type hybrid in the 11.7655-Mc oscillator unit, the power division at 131.060 Mc remains sufficiently close.

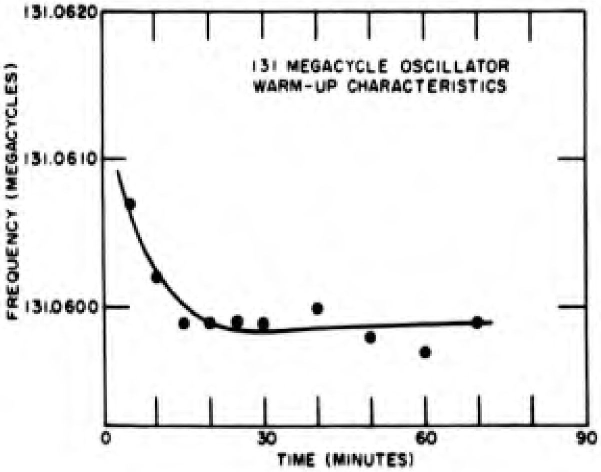

Fig. 16 - 131-Mc oscillator warm-up characteristics

Because a high signal level is desirable at the input to the mixer converter, resistance mixing is not practical; instead the great difference in frequency between the two signals is exploited by using high-pass and low-pass filters which accomplish the mixing with negligible insertion loss while providing considerable isolation between the oscillators. The conversion is obtained by means of V15, the 1N270 crystal diode. This crystal, in addition to being capable of withstanding considerable input power, has an extremely low forward resistance which provides a fair impedance match for the 50-ohm filters which it terminates. The resulting 119.295-Mc local oscillator signal is amplified by two bandpass amplifier stages which remove to a great extent the two mixer input signals. If either of these two frequencies are allowed to enter the receiver, the result is a differential phase shift between the combined received signals which is a direct system error, changing with signal level. This may be explained by taking the signals one at a time and investigating the result of their presence on the receiver output signals.

If some 11.7655-Mc signal were to enter the first local oscillator line, the insertion loss between the local oscillator input and the first i-f amplifier would reduce the signal's amplitude, but a small portion of it would nevertheless arrive at the second mixer with an amplitude depending on the gain of the first i-f amplifier at this frequency. The first i-f amplifier has a bandwidth of several hundred kilocycles, so at this frequency, only 470 Kc removed from the center frequency, the stage gain may well result in a level at the second mixer which is comparable to conventional i-f signal levels. This level, of course, depends upon the gain, which is controlled by the signal level through the AGC network. Thus the resultant local oscillator signal at the second mixer is dependent upon the input signal level. Obviously one such extraneous signal would not be sufficient to cause a direct phase error; however, if an 11.7650-Mc signal correspondingly leaks through the other combined i-f input and appears at the second mixer, the resulting beat between the authentic signal, if present, and these extraneous signals results in the correct second i-f frequency. When detected, this spurious second i-f frequency produces a 500-cps component which, combined with the authentic 500-cps signal, causes a phase error depending upon their relative phases and amplitudes.

Because of the large difference between 119.295 Mc and 11.765 Mc, the chance of the latter leaking through is relatively small. On the other hand, 131.0605 Mc is much closer to 119.295 Mc and is therefore considerably more difficult to remove by selective filtering. Let us examine the effect of this frequency leaking into the front end along with the authentic input frequency. The first item which confronts these two frequencies is the first mixer. The result of the two signals entering a mixer is the production of beats including, among other frequencies, their difference, 11.7655 Mc or 11.7650 Mc, depending upon the particular front end. The result, then, is identical to that which was just discussed with reference to 11.765 Mc leaking in directly.

The output of the mixer converter amplifiers (M.C.A. in Fig. 14), is coupled through a pad to the 119.295-Mc mixer converter. The outputs from both mixer converter amplifiers are mixed and the 500-cps beat difference is obtained. The mixer is a customary hybrid balanced mixer of a type which has been discussed elsewhere in this report.

The 119.295-Mc hybrid dividers are adjusted for maximum isolation at 108 Mc, thereby causing a power division error of about 1.5 db, which can be tolerated. By adjusting the coils, i.e., spreading or compressing the turns, it is possible to adjust these units to provide over 40 db of isolation if the capacitors and the load resistor are sufficiently precise.

The 500-cps reference obtained from the servo loop unit (Fig. 14) is amplified by a two-stage feedback-stabilized amplifier to a level of about 30 volts peak-to-peak. The output impedance of this amplifier, is transformed by a cathode follower, after which it is connected to a conventional phase detector. When the phase difference between the reference and standard signals is 90 degrees, the resultant vectors at the inputs to the detectors are equal, and since the diodes are reversed and the resultant detected voltages are added, the output is zero. A deviation in differential phase in one direction provides a positive voltage output while a deviation in the other direction causes a negative voltage output. This dc error voltage is fed to the grid of a conventional reactance tube which controls the frequency of its associated oscillator by causing varying amounts of reactive currents to flow in the oscillator tank circuit. Although employing a crystal, this tank circuit can be pulled slightly in frequency owing to its being neutralized to some extent by a series inductance which makes the characteristics of the crystal somewhat similar to those of an L/C antiresonant circuit.

The reactance tube modulator reference voltage is merely a variable controlled negative dc voltage which is fed to the reactance tube grid through, for practical reasons, the phase detector. This control provides a remote manual method of adjusting the frequency of the controlled oscillator in order to bring it within locking range. The remainder of this oscillator unit is identical to its 11.7655-Mc counterpart except for the elimination of the hybrid divider at the output, which is, of course, unnecessary, there being no second local oscillator output requirement from this unit.

CALIBRATION SOURCE ANALYSIS

The 108-Mc oscillator and buffer amplifier (Fig. 17) is identical in operation to the crystal-controlled 131.0605-Mc oscillator in the special local oscillator unit. The output of the buffer amplifier is matched to 51 ohms and supplies power to the variable attenuator. This output is also connected through appropriate resistors to the test point where frequency monitoring is accomplished, and to the level-monitoring detector circuit. The 1N82A crystal detects the output signal, which is then filtered and fed to a meter through a variable calibrating resistance. When this meter is set to the proper point by adjusting the screen voltage of the buffer amplifier, the attenuator dial reads directly in dbm.

The oscillator unit is very well shielded by two boxes, one inside the other. All leads entering these boxes are well decoupled, with the result that no leakage from the unit is discernible in the output of the extremely sensitive receivers mounted in the same rack and tuned to the oscillator frequency. Separate B+ switches are provided for the buffer and the oscillator in order to permit freedom of operation. For example, it is usually desirable to keep the oscillator operating in order that it remain stable in frequency. Whether the buffer remains on is less important, but it has some effect on the crystal environment and hence its frequency of oscillation. On the other hand, it is often suspected that some extraneous signal is leaking into the receiver, and in such cases these on-off switches provide a quick and simple method of eliminating the calibration unit as a source of interference. It is of course desirable to turn off the calibration unit during an actual operating run.

The attenuation of the 8-way output signal divider is approximately 30 db, while the insertion loss of the variable attenuator is about 40 db. Thus, since the desired output signal level is -60 dbm with the variable attenuator set close to minimum attenuation, the output power of the buffer amplifier at the junction of C8, R8 and R9 is about 10 dbm, i.e, 10 milliwatts.

SUMMARY AND CONCLUSIONS

The prototype model of this equipment was designed and constructed during the summer of 1956. It is the result of an accelerated program in which a concentrated effort was expended for a relatively short period. In spite of the severe restrictions on preliminary investigations with regard to optimizing circuitry and packaging details, the equipment which has been manufactured by the Bendix Radio Division of the Bendix Aviation Corporation remains virtually identical to the prototype furnished them.

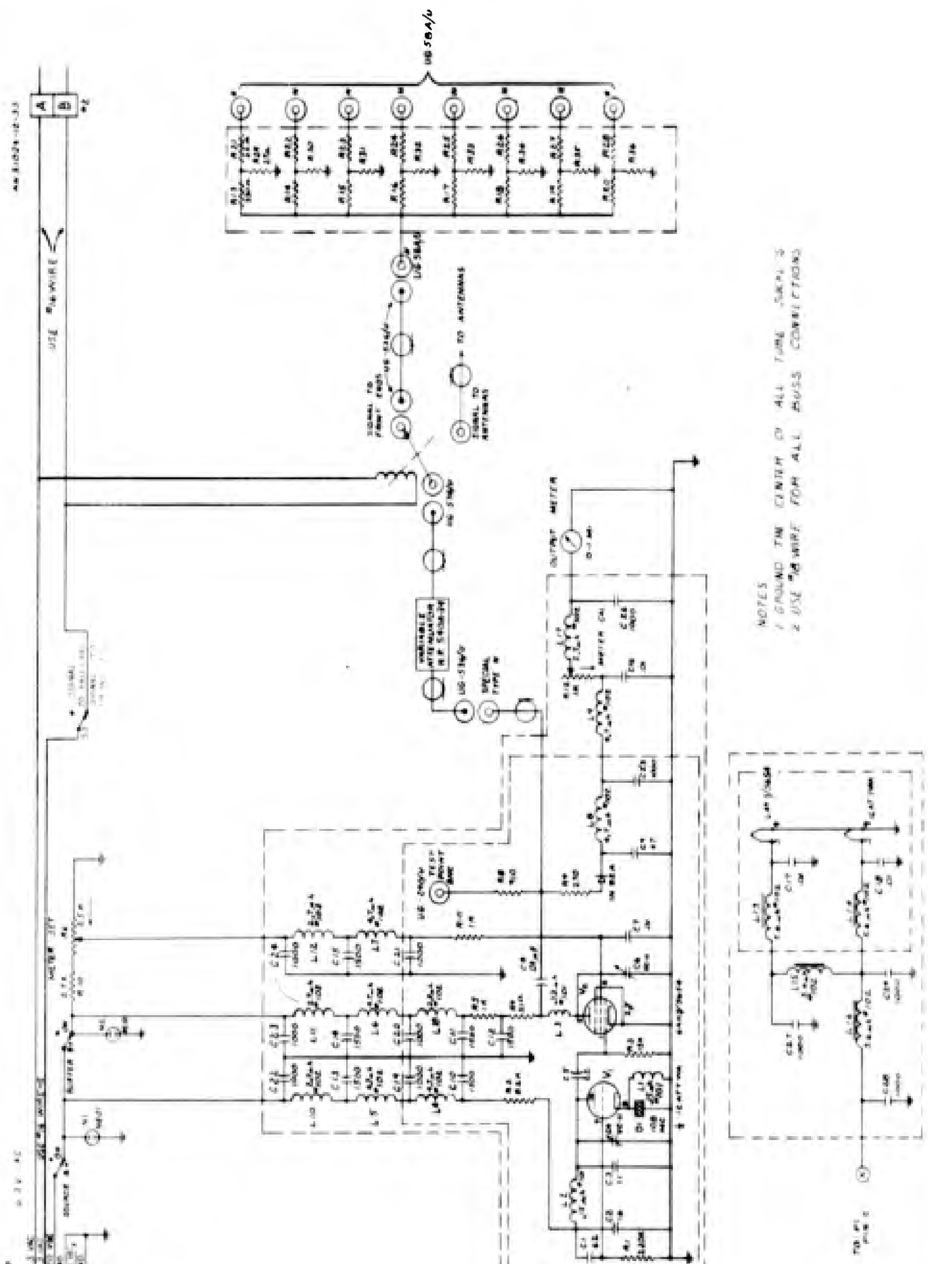

Fig. 17 - Calibration source

At the inception of this program the principal requirements confronting the development of this equipment were a noise figure less than 3 db at the 108-Mc transmitting frequency, local oscillator signals with a difference frequency of 500 cps locked precisely in phase with the time standard signal, and practically infinitesimal differential phase errors resulting from input level changes, frequency changes, and long-term drift. The prototype package, consisting of five phase-measuring channels, meets these requirements and in some cases was sufficiently better to permit tightening of the specifications. Preliminary tests on the Bendix-built production models indicate that their operation is essentially identical to that of the prototype model; thus it appears that the undertaking has been successful.

ACKNOWLEDGMENTS

The services of J. B. Martin, G. C. Kronmiller, Ens., USN, W. B. Moriarty, E. T. Hinkley, and D. W. Phipps who together with the authors constitute the Receivers Group of the Tracking and Guidance Branch of Project Vanguard, are gratefully acknowledged.

* * *

APPENDIX
Minitrack Receiver System Specifications

Front Ends

Bandwidth:* 4.5 Mc ±0.75 Mc.

Gain: 17 db ±3 db.

Noise figure: Less than 3 db.

Combined I-F Amplifier

Output signal amplitude variation: Less than 8 db for input signal variations between -60 dbm and -120 dbm.

Gain: Sufficient to amplify receiver noise, in the absence of signal, to a level which will excite the AGC circuit.

Differential phase shift vs input signal level: The phase difference between the 500-cps reference from the special local oscillator chassis and the 500-cps output from the combined i-f chassis shall not vary by more than 1 degree for changes in input signal level between -60 dbm and -120 dbm. The average must not change by more than 3.6 degrees for changes between -60 dbm and -130 dbm.

Differential phase shift versus input frequency: The phase difference between the 500-cps reference from the special local oscillator chassis and the 500-cps output from the combined i-f chassis shall not vary by more than 3 degrees for input frequency variations of ±5 Kc, and 1.5 degrees for input frequency variations of ±3 Kc about the 108.000-Mc center frequency at any input signal level between -120 dbm and -60 dbm.

I-F pre-detection bandwidth:* 12 Kc ±2 Kc.

AGC speed: The 500-cps signal from the i-f amplifier shall not fluctuate in amplitude by more than 50 percent for input signal level modulations of 20 db at modulating frequencies between 0 and 6 cps.

Receiver interchannel crosstalk: For this test, a 360-degree phase shifter is inserted in the line connecting the calibration source to a particular front end with all other channels connected directly. A 360-degree phase shift introduced in the designated front end shall not affect the phase meter readings of any disassociated channel by more than 0.75 electrical degrees at any signal level between -60 dbm and -120 dbm.

AGC 500-cps bandwidth and tuning accuracy: The AGC 500-cps amplifier shall resonate at 500-cps ±3 cps and have a bandwidth* of 75 cps ±10 cps.

---

* Between 3-db points

Special Local Oscillator

Reactance modulator servo loop gain: The frequency of the 11.7655-Mc oscillator should be pulled slightly by detuning the oscillator tuning control, meanwhile noting the resulting change in the phase difference between the 500-cps standard and 500-cps reference voltages. The gain is expressed as cps per electrical degree. This ratio shall be less than 5 cps per degree.

Oscillator warm-up stability: The frequency of the three oscillators in this unit shall not drift more than is consistent with the temperature specifications of their respective crystals. The temperature coefficient of frequency for all crystals shall be less than 1 part per million per degree centigrade.

Oscillator stability in the presence of varying plate supply voltage: The 131.060-Mc oscillator shall not deviate from its operating frequency by more than 1 Kc for plate supply voltage changes of 20 volts (i.e., from 150 to 170 volts). The 11.7655-Mc oscillator frequency shall deviate less than 100 cps for the same change.

Local oscillator output amplitudes: All local oscillator output amplitudes shall be sufficient to excite the mixer diodes (1N82A) to a level between 0.5 and 1.0 milliamperes.

Calibration Source

Oscillator warm-up stability: The frequency of the 108.000-Mc oscillator shall not drift more than is consistent with the temperature specifications of the crystal. The temperature coefficient of frequency for the crystal shall be less than 1 part per million per degree centigrade.

Oscillator stability in the presence of varying plate supply voltage: The 108.000-Mc oscillator shall not deviate from its operating frequency by more than 600 cps for plate supply voltage changes of 20 volts (i.e., from 150 to 170 volts).

Accuracy of attenuator dial readings: The attenuator dial readings shall be capable of being adjusted, at settings between -90 and -120 dbm, to within ±1 db as compared with a standard signal generator.

Output power divider accuracy: The interterminal variation in output power from the eight terminal power dividers shall be less than 2 db.

Signal leakage: There shall be no discernible leakage as observed by monitoring the AGC voltage.

* * *

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Naval Research Laboratory. Report 5055. **PROJECT VANGUARD REPORT NO. 23, MINITRACK REPORT NO. 3, RECEIVER SYSTEM**, by V. R. Simas and C. A. Bartholomew. 27 pp. & figs., December 6, 1957.

The Minitrack system is designed to acquire and track artificial earth satellites which Project Vanguard will attempt to place in orbits as part of the United States' effort in the International Geophysical Year. The satellites will be provided with 108-Mc signal sources to illuminate the tracking stations. These sources will radiate sufficient power to permit the desired tracking accuracy. The Minitrack system functions as a radio interferometer, providing the direction cosines of a satellite's position vector with respect to the tracking stations at any given instant during a transit.

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1. Radio tracking systems - Design
I. Simas, V. R.
II. Bartholomew, C. A.
III. Project Vanguard Report 23
IV. Minitrack Report 3
V. Minitrack

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The receiving equipment must amplify the signals obtained from the satellites via interferometer antenna pairs, with the introduction of a minimum of noise and differential phase shift, and by means of a unique double-local-oscillator principle translate this intelligence to a frequency readily adaptable to phase measurement.

This report describes, with schematics, the five units comprising a single rf channel, i.e., a pair of front ends, a signal adder, a combined i-f amplifier, a special local oscillator, and a calibration source.

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