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PROJECT VANGUARD REPORT NO. 24

MINITRACK REPORT NO. 4
THE SATELLITE TELEMETRY RECEIVER SYSTEM

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Project Vanguard

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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

NRL Report 5055, "Project Vanguard Report No. 23, Minitrack Report No. 3; Receiver System," by V. R. Simas and C. A. Bartholomew, December 6, 1957

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ABSTRACT

Scientific information from the artificial earth satellites will be transmitted to the telemetry systems located at each of the Minitrack stations. The telemetry receiver will amplify the received signals with the addition of as little receiver noise as possible and convert the information to a form which permits demodulation and subsequent analysis. The receiver is of the double-conversion type with crystal-controlled local oscillators. Three pre-detection bandwidths are available which provide a means of increasing the output signal-to-noise ratio for those experiments in which the information bandwidth is less than the maximum. Grounded-grid preamplifier circuitry is employed to achieve stability and a low noise figure.

PROBLEM STATUS

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

AUTHORIZATION

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INTRODUCTION

In addition to the Minitrack transmitter, the Vanguard satellites will contain scientific measuring equipment to obtain information pertaining to enviromental conditions, solar Lyman-alpha intensity, cosmic-ray intensity, and many other experiments. Telemetering equipment in the satellites will code this information and transmit the data to ground receiving equipment located at each of the Minitrack tracking stations.

The information obtained by the scientific instrumentation within the satellite is in the form of voltages which are applied to the telemetry coder. These voltages are sampled or commutated in a predetermined time sequence by the coder unit. Each sample is then translated into either a frequency or a time duration. The output of the coder unit is applied as double-sideband amplitude modulation on the 108-Mc carrier of the Minitrack transmitter. It was early established that within certain frequency limitations, severe modulation of the Minitrack oscillator caused practically no deterioration of the tracking function. Therefore, this oscillator not only provides the tracking signal but the telemetry carrier as well, thereby saving the additional weight and power consumption of a separate telemetry oscillator.

The modulation waveform used to convey information about the satellite's instruments is shown in Fig. 1. This waveform is a series of tone bursts (f_1, f_2, etc.) and blanks (T_2, T_4, etc.). By applying a known conversion factor, the reading of instrument A is given by frequency f_1, the reading of instrument B is given by T_1, and so on. Times T_1, T_2, etc., range from 15 to 30 milliseconds; frequencies f_1, f_2, etc., lie between 2.5 and 15 kc.

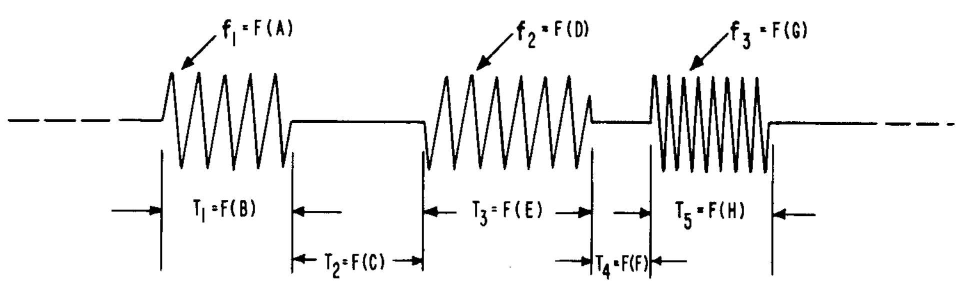

Fig. 1 - Telemetry modulation waveform

GENERAL DESCRIPTION

The function of the telemetering receiver is to amplify the incoming signals to a level which permits demodulation and subsequent analysis. It is necessary to obtain this amplification with a minimum of added receiver noise. The requirements also include specifications for the pre- and post-detection bandwidths and the phase characteristics of the band-pass filter networks. In addition, the system is supplied with automatic gain control (AGC) which must maintain an approximately constant output for wide input signal level fluctuations.

The following discussion on the circuit operation is facilitated by close reference to the block diagram, Fig. 2. The circuit diagram is shown in Fig. 3. The input signal from the antenna is first amplified by the GL 6299, a low-noise grounded-grid planar triode, where a noise figure of less than 3.0 db is achieved. The second stage, utilizing a 6AN4 tube, is also connected as a grounded-grid stage in order to obtain adequate bandwidth and to maintain the first-stage noise figure as the overall system noise figure.

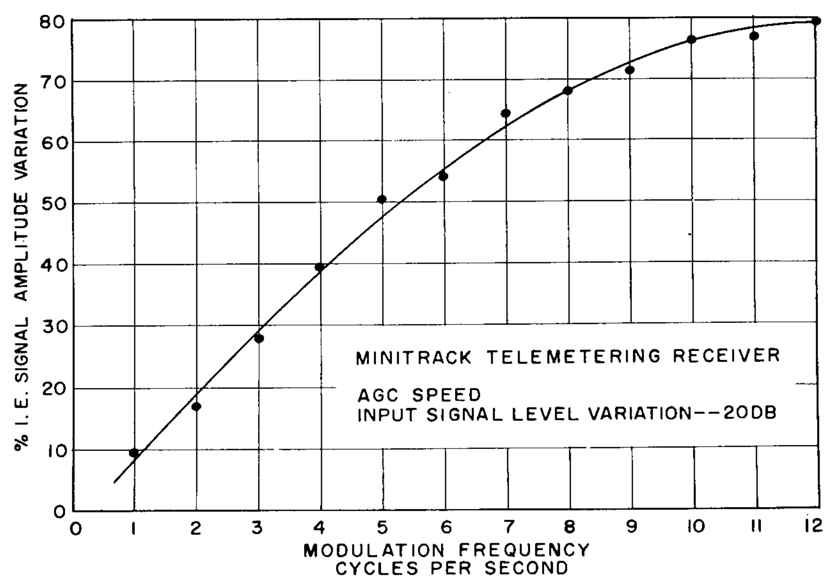

Fig. 2 - Block diagram, Minitrack telemetering receiver

At this point, the 108-Mc signal is mixed with the 119.295-Mc first local oscillator signal and converted to 11.295 Mc, the first i-f frequency. The first i-f amplifier is composed of a conventional pair of amplifying stages utilizing type 5654 tubes. The gain of each of these stages is set at about 30 db by resistive loading in parallel with the equivalent resistance of the plate tank circuits. Reasonable bandwidths are obtained by selecting particular values for the L/C ratio which when damped with the load resistance provide the required circuit Q.

The 11.2950-Mc signal at the output of the first i-f amplifier is mixed with the second local oscillator output of 11.7650 Mc and converted to 470 kc, the second i-f frequency. The conversion is performed by a single 1N82A silicon diode.

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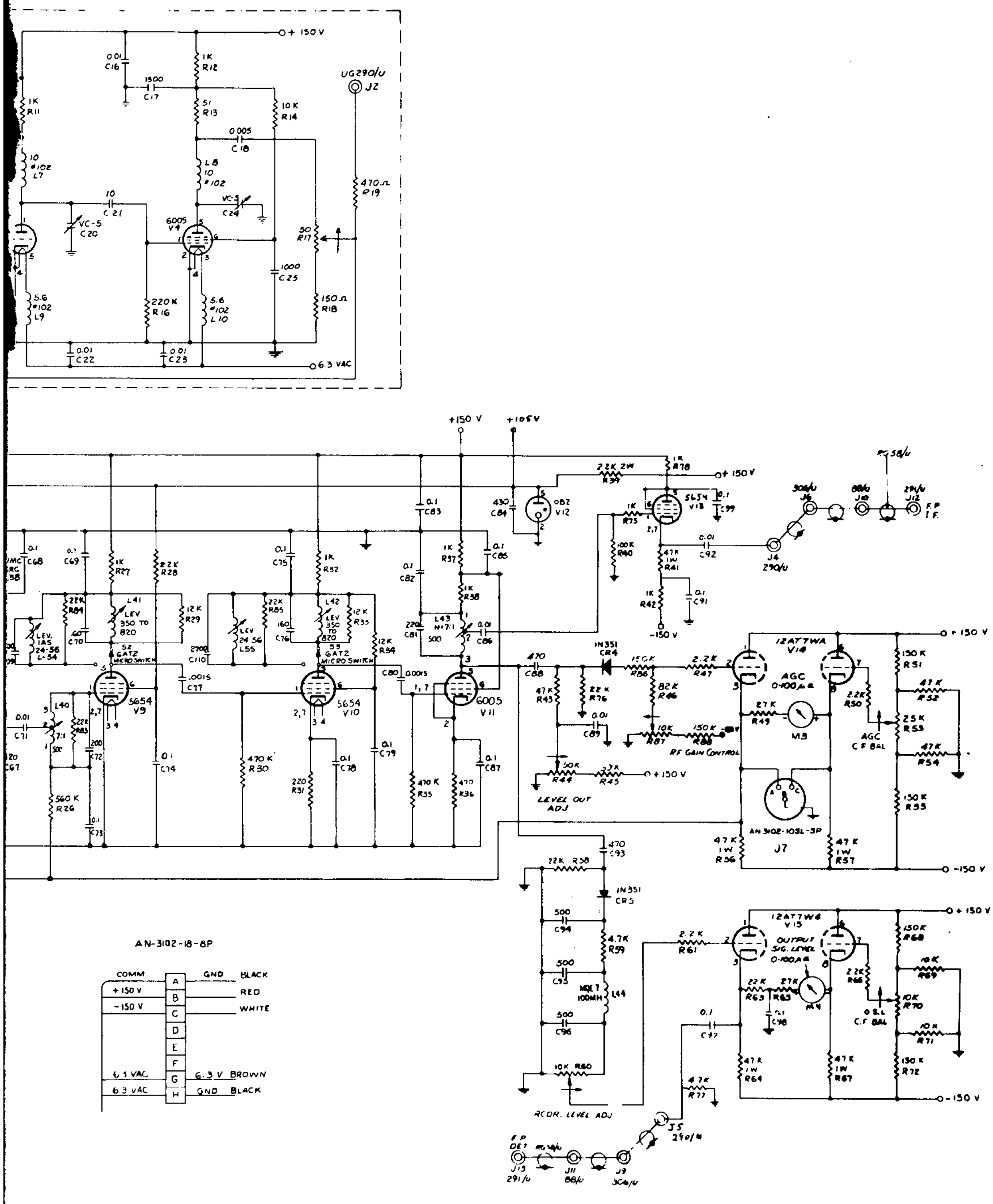

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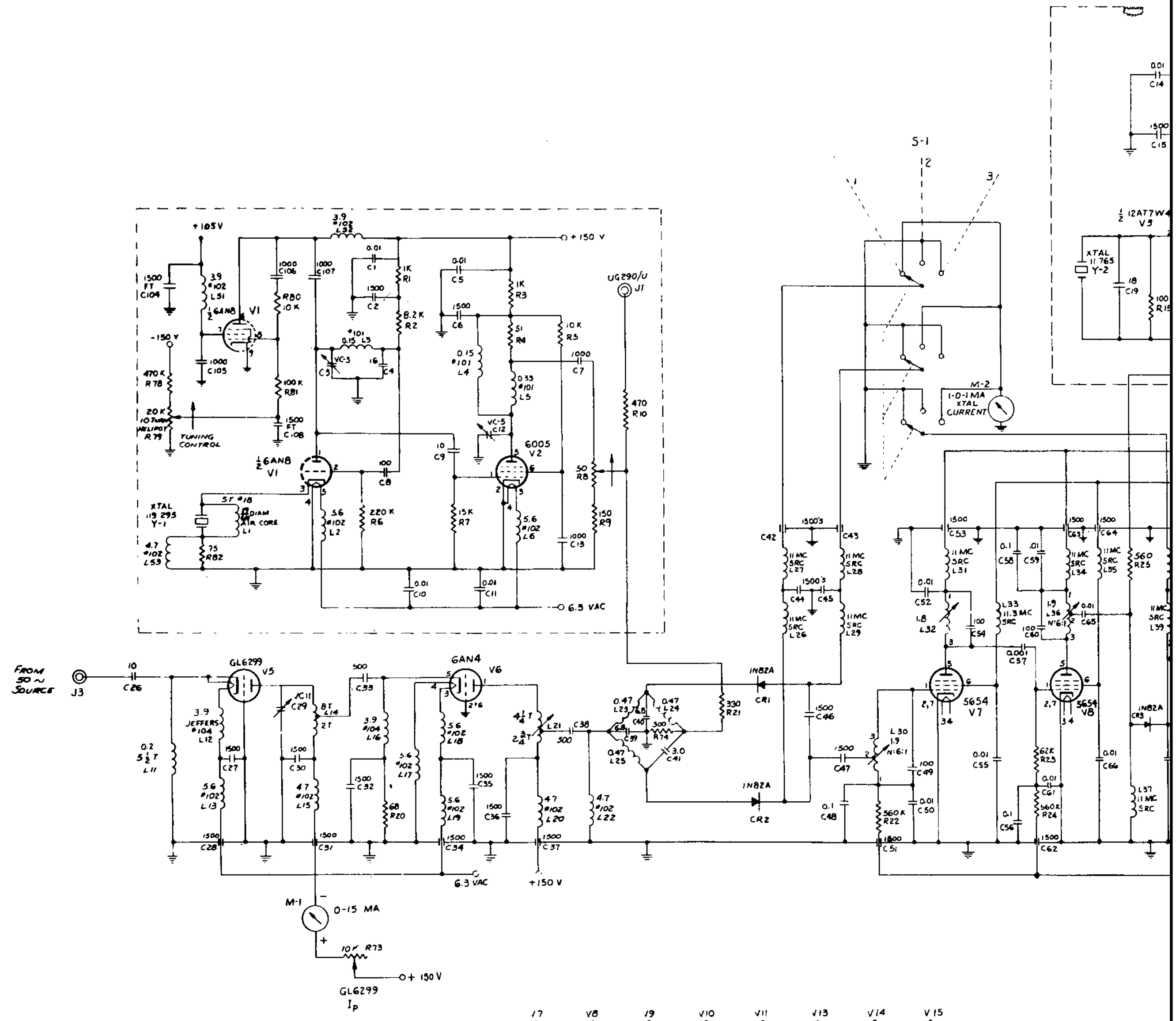

Notes:

1. All Capacitance Values In Micromicrofarads If One Or Higher And In Microfarads When Less Than Unity.
2. All Inductance Values In Micro Henries Except Where Otherwise Specified.

Fig. 3 - Circuit diagram, Minitrack t

The second i-f amplifier provides the remaining gain as well as three pre-detection bandwidths, any one of which can be selected by appropriate settings of the bandwidth switches. The three stages that comprise this unit utilize two type 5654 tubes and a type 6005 tube, each of which has approximately 30-db gain. The proper bandwidths for this unit are again provided by the selection of the proper L/C ratios in the plate tank circuit with respect to the equivalent plate load resistance of each stage. The plate circuits of two of the stages in this strip are provided with switches which select one of two tank circuits. With both switches thrown to select tank circuits with the highest L/C ratio, the bandpass is about 31 kc. With both switches thrown to select tanks with the lowest L/C ratio, the bandpass is about 3.4 kc. With one switch on low and the other on high, the bandwidth is about 5 kc. The ability to switch bandwidths enables the operator to optimize the signal-to-noise ratio for various modulation frequency spectrums. For example, the regular modulation channel requires a width on each side of the carrier of about 15 kc. However, it is anticipated that there will be times when the modulating information encompasses only about 1.5 or 2 kc. In such a case, it is advantageous to restrict the noise bandwidth along with the information bandwidth, thereby increasing the signal-to-noise ratio at the output of the receiver.

The 470-kc signal at the output of the amplifier is connected to a cathode follower which supplies signal voltages at a low impedance via a BNC connector to a linear detector unit or to various test instruments used to monitor the output signal. The 470-kc signal at the output of the 6005 tube is also detected by a signal detector and the AGC detector. The AGC voltage is converted to a low impedance by means of a cathode follower, the output of which supplies the AGC bus as well as input-level information to the recorder and monitoring meters.

The detected signal is filtered by a low-pass filter with a cutoff frequency sufficiently high to be almost perfectly flat out to 15 kc. The overall system bandwidth out to the 3-db points is therefore limited only by the pre-detection bandwidth. The impedance of the detecting circuit is transformed by a cathode follower and is available for recording. With the controls set at their approximate midpoints, the detected output to the recorder is 1 volt rms.

Both local oscillators are crystal controlled. In addition, the first L.O. is adjustable over an 8-kc range. A conventional reactance tube pulls the frequency of the oscillator. The frequency is varied by adjusting a potentiometer located on the front panel which controls the dc operating point of the reactance tube, which in turn controls the oscillator frequency. The reactance tube and oscillator functions are both performed by the triode-pentode tube, 6AN8. The 6005 buffer amplifier isolates the load from the generator and supplies power gain as well.

The second local oscillator is a conventional crystal-controlled TPTG oscillator in which the crystal operates in its parallel-resonant condition. One-half of a 12AT7WA is the oscillator tube, while the buffer is again the 6005 tube. The amount of signal insertion to the receiver is determined by the value of the series resistor connecting the two units. This resistance is chosen to provide a crystal current of approximately 0.7 ma.

RECEIVER TELEMETERING SPECIFICATIONS

Sensitivity: the gain shall be sufficient, with the rf gain control and the AGC level control fully clockwise, to amplify the receiver noise in the absence of any input signal to a level 10 db greater than that required to provide 6 volts p-p at the receiver i-f jack.

Noise figure: the noise figure shall be less than 3 db.

Maximum output level: with the input signal 100-percent modulated at about 1 kc and at a level of about -100 dbm, the receiver i-f pattern shall not show limiting at any output level below 10 volts p-p measured at the receiver i-f jack.

Output signal level: with the rf gain control and the recorder level adjust control approximately centered, the 470-kc output at the test point shall be 6 volts peak to peak and the detected signal level shall be 1 volt rms at input signal levels of -100 dbm.

Pre-detection bandwidths: the wide pre-detection bandwidth shall be 31 kc ±2 kc at the 3-db points. The medium pre-detection bandwidth shall be 5.0 kc ±0.5 kc and the narrow pre-detection bandwidth shall be 3.4 kc ±0.5 kc.

Post-detection bandwidth: the post-detection bandwidth shall be essentially flat to or beyond the pre-detection limitations; that is, it shall be flat (±1.0 db) from 0 cps to 15 kc.

AGC speed: the output signals shall not fluctuate in amplitude by more than 50 percent for input signal level variations of 20 db at modulating frequencies between 0 and 3 cps.

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

Crystal drive: the power dissipated within the 119.2950-Mc crystal shall be less than 5 milliwatts.

Crystal accuracy: the accuracy of the 119.2950-Mc and 11.7650-Mc crystals shall be within plus or minus 500 cps. There shall be no spurious resonances within 20 kc of the prescribed frequency.

CIRCUIT ANALYSIS

Preamplifier

The function of the preamplifier is to generate a good noise figure and to provide the gain necessary to establish the preamplifier noise figure as the system noise figure. It should be stable yet easily constructed and aligned.

A preamplifier consisting of two grounded-grid cascoded stages has characteristics ideally suited to the preceding requirements. In the first place, there are many tubes on the market which function well as grounded-grid amplifiers at 108 Mc. In addition, the grounded-grid configuration is the most stable of the three possible circuit arrangements. The tuning arrangement for this circuit is the most elementary possible, consisting of peaking two plate tank circuits. The difficulty sometimes encountered with this circuit is that of utilizing the high-output impedance of these stages. However, with the fairly low fractional bandwidths required for this system, this is no obstacle.

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

$$F_1 = 1 + \frac{G_{1\,eq}}{G_s} + \frac{B\,G_t}{G_s} + \frac{R_{eq}\,(Y_s + G_t)^2}{G_s},$$

where $G_{1\ eq} = G_1 + G_D$.

G_1 = input circuit losses.

G_D = the output circuit losses.

G_t = transit time conductance.

G_s = transformed source conductance.

B is a constant with a value of approximately 5.

R_{eq} is the equivalent noise resistance of the tube, approximately equal to 2.5/gm for most triodes.

$Y_s = G_s + jY_1 + G_1$, where jY_1 is the susceptance presented by the input network to the input terminals of the tube.

A cursory examination of this expression allows several immediate conclusions.

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

2. All coil losses or other losses represented by $G_{1\ eq}$ increases 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 both in the numerator and denominator, indicating that an optimum value exists. Indeed, 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, with respect to G_s, the noise figure expression and setting the first derivative equal to zero. The resulting expression, after dropping the insignificant terms, is:

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

The corresponding value of the optimum attainable noise figure is:

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

These two expressions are the key to computations concerning the noise figure of grounded-grid preamplifiers, hence anyone desirous of obtaining an understanding of the subject should analyze and study them.

The overall noise figure of the amplifier is expressed as:

$$F = F_1 + \frac{F_2 - 1}{W_1} + \frac{F_3 - 1}{W_2} \text{ ---------},$$

where W = power gain.

This series is rapidly convergent, for all practical cases, so that usually just two terms are significant and sometimes only one. The following is the application of these equations to the telemetering amplifier.

The circuit parameters are:

	GL 6299	6AN4
G_t	30 μmhos	160 μmhos
R_{eq}	125 ohms	250 ohms
g_m	12000 μmhos	10000 μmhos
r_p	9600 ohms	7000 ohms

$$f_o = 108 \text{ megacycles}; \quad \omega_0 = 6.78 \times 10^8$$

First-stage noise figure: assume the Q of input and output coils is 200.

$$G_1 = 1/Q\omega L_1 = 36.8 \times 10^{-6} \text{ mhos}$$

$$G_D = 1/Q\omega L_4 + N^2 G_{in} = 250 \times 10^{-6} \text{ mhos}$$

$$G_{1\ eq} = G_1 + G_D = 49.1 \times 10^{-6} .$$

Then,

$$G_{s\ opt} = \sqrt{\frac{G_1 + BG_t}{R_{eq}}} = 2370 \times 10^{-6} \text{ mhos}$$

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

$$= 1.8 \text{ db} .$$

The values of $G_{s\ opt}$ and $F_{1\ opt}$ for the second stage are:

$$G_{s\ opt} = 2200\ \mu\text{mhos}$$

$$F_{1\ opt} = 3.2 \text{ db} .$$

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 is equal to:

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

Thus the second stage contributes only 0.3 db to the noise figure.

The circuitry of the preamplifier can be divided into three main parts: (1) The input transformer, (2) the coupling transformer, and (3) the output transformer. The function of the input transformer, for noise matching conditions, is to transform the source impedance to the value which provides the lowest noise figure. This is the $G_{s\ opt}$ discussed in the previous paragraphs. There are many types of transformers which will perform this function adequately; however, the one used in this unit is possibly one of the simplest to construct and uses the fewest parts.

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 420 ohms, as determined by the preceding calculations for $G_{s\ opt}$. R_o, L11, and C26 form a series-resonant circuit, R_o being the series damping resistor. The relationship between the series damping resistor and its equivalent parallel damping resistor is:

$$R_s R_p = (1/\omega C)^2 = (\omega L)^2,$$

where

$$R_s = \text{series resistance} = R_o$$

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

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

The values of C_1 and L_1, which provide the desired transformation, are easily found:

$$1/\omega C_1 = \sqrt{50 \times 520 = 160}$$

$$C_1 = 1/6.78 \times 10^8 \times 160 = 9.2 \text{ mmf}.$$

This is exceedingly close to the optimum value, 10 mmf, determined experimentally.

The coupling transformer can be considered in exactly the same manner as the input transformer. Here the 6AN4 is the equivalent first stage and the transformer transforms the source conductance, in this case the output conductance of the GL 6299, to the optimum transformed, conductance $G_{s\ opt}$, of the 6AN4. The optimum transformed impedance, $1/G_{s\ opt}$, is about 520 ohms, determined in the same manner as for the GL 6299. The output impedance of the first stage, including the coil losses, is about 25 kohms. Therefore, the transformer must step this impedance down by a factor of about 50, which leads to a turns ratio of about seven to one.

This tank is loaded by the input conductance of the 6AN4 through the turns ratio of L14. This loading ability accounts in large measure for the usefulness of the grounded-grid configuration as the second stage of most preamplifiers. If a grounded cathode stage is used as the second stage, the high-input impedance provides little loading on the preceding tank, with the result that the bandwidth of this circuit is exceedingly narrow. If resistive damping is resorted to, the noise figure suffers. The usual way to combat this situation is to employ feedback, which lowers the input impedance while avoiding the introduction of losses into the circuit.

The calculations show that for optimum noise considerations, the second-stage contribution to the system noise figure is very small. Therefore, if bandwidth is an important consideration, a small amount of noise figure deterioration can be allowed in favor of obtaining a wider bandpass. The practical method of adjusting this circuit for particular conditions is to decrease the turns ratio of L14 by raising the tap position until the desired bandwidth is obtained or until the noise figure increases to a point beyond which it would not be acceptable.

The output transformer merely transforms the high-output impedance of the 6AN4 to a value compatible with the input impedance of the mixer hybrid junction. Matched conditions here are desirable if bandwidth considerations permit. However, if the bandwidth is insufficient, the tap position must be raised, thereby reflecting a greater load across L21 and increasing the frequency response.

First Mixer Converter

The mixer in this unit is in the form of a hybrid junction. It is the lumped-constant counterpart of the familiar coaxial-line junction, commonly called the "rat race," or the magic-T formed from a three-dimensional waveguide junction. It is composed of four reactive matching networks, each having the required image resistances, three of which have positive 90-degree phase shift and one having 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 terminal resistances and the circuit reactances is:

$$1/\omega C = \omega L = \sqrt{2R_1R_2},$$

where R_1 and R_2 are the terminal impedances of each network. The image resistance for each terminal of this junction is 300 ohms, which is the approximate impedance of the 1N82A crystals at this frequency and local oscillator drive. The local-oscillator insertion terminal is located opposite the signal-input terminal. The relatively high isolation between these terminals prevents crosstalk and loss in signal. The signal and L.O. combine at the adjacent terminals in such a manner that the beat between them is 180 degrees out of phase. The 1N82A crystals which serve as loads for the hybrid are reversed in order that the heterodyned output signals may be combined in the proper phase for direct addition. The blocking condensors and pi filters allow the current from each diode to be monitored separately by the crystal current meter.

I-F Amplifiers

The signal at this point has been converted from 108 Mc to 11.2950 Mc by the 119.2950-Mc translating oscillator. The output impedance of the mixer converter is about 150 ohms, and since we are still not out of the woods with regard to noise figure, it is advisable to transform the output impedance of the converter to a value which produces an i-f noise figure low enough to prevent undue contribution to the system noise figure. The input transformer, L30, is a tuned autotransformer with a step-up ratio of about 6:1. Therefore the input impedance is transformed to about 6000 ohms, a figure compatible with low-noise figures of this stage. It is not necessary, in spite of the noisiness of pentodes, to use a circuit configuration other than the standard pentode arrangement for the first stage in this i-f amplifier.

The gain of each of these stages, V7 and V8, is about 30 db, a practical value at this frequency. The second i-f amplifier determines the system pre-detection bandwidth, so the bandwidth of the first i-f is set at a value which provides convenient tank components. As a result, the Q's of the tank circuits in the first i-f amplifier are about 30, a convenient value.

Both stages, comprising the first i-f amplifier and the first stage of the second i-f amplifier, have their operating point, hence their gain, controlled by AGC voltage. The transconductance, hence the gain, of 5654 tubes varies markedly with grid bias voltage so that a dc voltage proportional to the output signal level connected to these grids provides control of the gain of these stages. Because the dc operating point is the quantity that is being varied, it is unwise to depend upon the tube current to provide the operating conditions. In other words, there should be no cathode bias resistor or screen dropping resistor. Therefore, the cathodes of these stages are grounded and the screens are tied to a regulated 105-volt source. Thus the AGC voltage does not first have to overcome these voltages in order to control the gain.

The output of the second stage, V8, is stepped down to a value compatible with the input impedance of the second mixer. The mixing of the signal with the local-oscillator voltage is accomplished by a form of resistive addition using the output impedance of the tube, V8, and the 1-kohm resistor, R25, in series with the output of the local oscillator. The combined signals are rectified and converted by CR3, the 1N82A silicon crystal. The input autotransformer, L40, is tuned to 470 kc, the difference frequency between the first i-f, 11.2950 Mc, and the second local oscillator frequency, 11.7650 Mc. The output impedance of the converter is fairly low, about 300 ohms, so a step-up transformer is employed in order to obtain a good bit of passive gain.

The tuned autotransformer, together with the tank circuits in the plate circuits of the following three stages, are essentially the sole pre-detection bandwidth-limiting networks. For the widest bandwidth, these tanks are all arranged to have very nearly the same Q, and since they are synchronous single-tuned circuits, the overall bandwidth is determined from the following expression:

$$\frac{\text{System bandwidth}}{\text{Single-stage bandwidth}} = \sqrt{2^{1/n} - 1},$$

where n equals the number of tuned circuits. In our case, n equals 4, so the bandwidth shrinking factor equals 0.44. The single-stage bandwidth for the widest bandwidth position is about 70 kc, so the overall bandwidth is about 31 kc. Two other pre-detection bandwidths can be obtained by switching other tank circuits into the plate circuits of V9 and V10. These narrow-band tanks both have Q's of approximately 90, so that when both are switched into the circuit, the bandwidth is about 3.4 kc. With one wide and one narrow, the overall bandwidth is about 4.9 kc (Fig. 4).

In spite of the variation in bandwidth, the gain of these stages remains virtually constant. This results from the fact that the equivalent plate load resistor for all the tuned circuits remains constant, the change in the bandwidth being effected by varying the L/C ratio and not the load resistor.

The signal at the output of the second i-f strip, plate of V11, is connected to three circuits: (1) A 470-kc output via a cathode follower for monitoring and connection of the linear detector, (2) the AGC circuit, which detects the signal and feeds back a dc voltage proportional to the signal level, thereby maintaining a somewhat constant output level, and (3) the signal detector circuit, which demodulates the signal and connects the output via a cathode follower, V15, to a BNC connector for connection to an oscilloscope for visual monitoring and to a meter which indicates signal level.

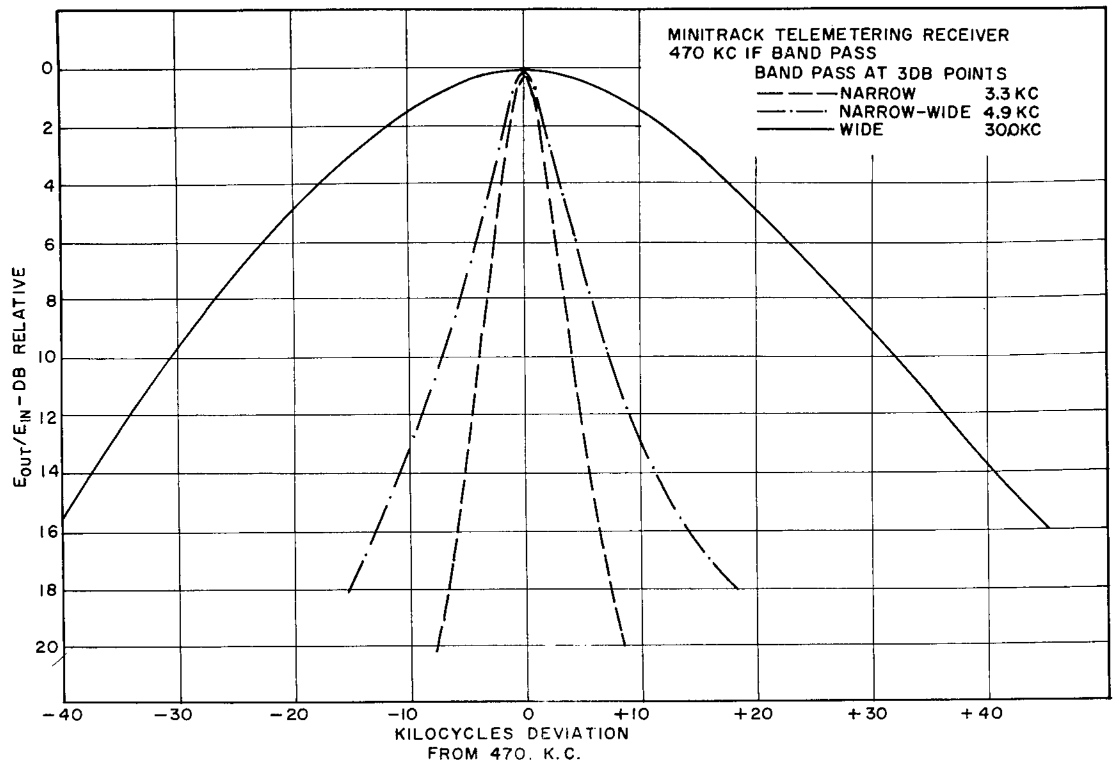

Fig. 4 - Minitrack telemetering receiver 470-kc i-f bandpass

The constant-K filter in the detector circuit, C95, C96, and L44, is flat beyond the bandwidth limiting capabilities of the pre-detection circuits. Thus, the resulting overall bandwidth is essentially that of the pre-detection circuits out to the 3-db points, after which it falls more rapidly due to the post-detection bandwidth circuitry.

The AGC detector utilizes a 1N351 rectifier which has its cathode tied to a variable positive voltage furnished by the output level adjust control. This voltage is the familiar AGC delay voltage, which prevents AGC action until the desired output level is attained. For an explanation of the operation of this circuit, assume that the control is shifted more positive. The signal voltage presented to its plate circuit is then insufficient to cause the 1N351 diode to conduct. The AGC voltage will then rise toward zero voltage, which will increase the transconductance of the three stages controlled by the AGC voltage. Hence, the gain of these stages will increase, resulting is a sufficient increase in signal voltage applied to the AGC detector circuit to cause diode conduction and the resulting gain limitation.

The rf gain control varies the negative voltage on the AGC bus. This limits and controls system gain until the input signal is sufficiently high to overcome this manual gain control. Additional operating flexibility is thus given the system.

Cathode followers are used in the AGC and detected signal output circuits to isolate and reduce the impedance of the output signals. The duo-triode cathode follower used in this application exhibits several beneficial effects which a single triode unit does not possess. The second unit provides a low-impedance balancing voltage. 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. In the event that either the positive 150 volts or negative 150 volts should fail or should not be applied concurrently, the meter is not damaged, there being not as great a potential difference across it as there would be if the meter return were grounded as in a single-stage cathode follower.

Crystal-Controlled Local Oscillators

The 11.7650-Mc oscillator, V3, is the simplest type of TPTG oscillator, the crystal serving as a parallel-resonant tank in the grid circuit and oscillation being 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 and having a random short-term variation of only a few cycles. The output of this stage is coupled to V4, a 6005 tube, which amplifies, isolates, and transforms the output to an impedance suitable for insertion into the mixer circuit. The transformer in this circuit is the same type that is used for transforming the source impedance to the optimum value for minimum noise figures in the preamplifier. However, instead of transforming upward, this application requires that the high-output impedance of the 6005 tube be transformed downward in order to provide isolation and power to the variable attenuator, R17, and load, R25.

In order to accommodate better the narrow bandpass of the receiver with respect to the variations in frequency of the incoming signal, the first local oscillator is made slightly tunable. It is still controlled by the crystal, however, and has excellent stability in spite of this tunable feature. The oscillator and reactance sections are enclosed within the single envelope of the 6AN8 tube, the pentode unit being the reactance tube and the oscillator being provided by the triode unit.

The bridged-T oscillator relies on the negative transfer characteristics of the LC pi network (C3, L3, and C4 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 is controlled in frequency by the crystal by virtue of the high cathode degeneration that exists at frequencies other than the series-resonant frequency of the crystal together with the rapidly changing phase shift in the vicinity of resonance. The crystal drive, i.e., power dissipated in the crystal, is maintained by the proper relationship between C3 and C4. In practice, C4 is varied, which adjusts the drive, and the circuit is kept resonant by compensating for the change in C4 by altering the value of C3. L1 is the neutralizing coil for the crystal which resonates with the crystal-holder capacitance, thus making the crystal-circuit impedance high in the vicinity of resonance.

The buffer amplifier associated with this unit is similar to the one in the 11.7650-Mc oscillator, except for the plate circuit values. Its operation is identical, and therefore the reader is referred to the previous paragraphs for information concerning this stage.

The reactance tube alters the frequency by introducing a reactance across the oscillator plate circuit, the value of which depends upon the reactance tube gain, which in turn is manually controlled by a potentiometer located on the front panel. The operation of the reactance tube is conventional and straightforward, even though the application with respect to its operation on a crystal-controlled oscillator is somewhat at odds with the usual practice. The oscillator signal on the plate is shifted in phase by means of an RC integrating circuit, after which it is connected to the grid of the reactance tube. The voltage on the grid is converted to current fluctuations which feed back into the oscillator plate circuit. These current fluctuations are shifted, due to the integrating circuit, by about 90 degrees, with the result that the oscillator plate views them as though thay originate from an inductance. The effect is as though there were an inductance introduced into the plate circuit which, of course, detunes the circuit slightly and pulls the frequency of oscillations.

PLAN OF ACTIVITIES

This equipment was designed and the prototype constructed at the Naval Research Laboratory by the Receiver Section of the Tracking and Guidance Branch of Project Vanguard. Ten of these units were manufactured by the Bendix Corporation. The equipment has been checked and is at this writing being installed at the ten Minitrack stations. There it will receive further checks and prolonged operation during station calibration runs using a Minitrack oscillator flown over the station. It will be checked and in operating condition for the first Satellite launching attempt occurring during the latter part of 1957 or the early part of 1958.

ACKNOWLEDGMENTS

The Author is pleased to acknowledge the excellent work and dedicated efforts of Mr. W. B. Moriarty who worked toward the successful completion of this equipment.

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