T

The Ground Transmitter and Receiver

By A. J. GIGER, S. PARDEE, JR., and P. R. WICKLIFFE, JR.

(Manuscript received March 22, 1963)

10880

Ground station equipment for the Telstar experiment includes a 6-kmc transmitter and a 4-kmc receiver used for television and telephone channels. This paper describes the over-all transmitting and receiving arrangements and presents detailed accounts of the RF and power-supply subsystems. Information is also presented on protection, control and equipment features, and on receiver noise performance.

AUTHOR

I. TRANSMITTER

1.1 *General*

The Telstar ground transmitter is designed to produce a 2-kw frequency-modulated signal in the 6-kmc band, suitable for the transmission of television or multiplex telephony. Fig. 1 is a simplified block diagram which shows the basic features of the four sections of the transmitter: FM deviator, modulator-amplifier, transmitter carrier supply, and power amplifier. The Telstar transmission objectives[1,2] are similar in many respects to those established for domestic microwave radio relay systems. As a result, it has been possible to use in the Telstar transmitter many pieces of equipment designed for that service. The FM deviator, modulator-amplifier and transmitter carrier supply are direct adaptations of units currently being manufactured for the 6-kmc TH radio relay system.[3,4,5] The power amplifier is designed around a 2-kw traveling-wave tube[6] developed at Bell Telephone Laboratories.

1.2 *Design Considerations*

Table I summarizes the transmission objectives established for the ground transmitter. These are derived from system objectives based, primarily, on one-way transmission of monochrome television and 600-channel multiplex telephone. Simultaneous two-way transmission through the satellite, together with its restricted AGC range, deter-

In its Telstar 1, Vol. 1 Jun. 1963
p 1063-1107 refs (See N64-10868 02-01)

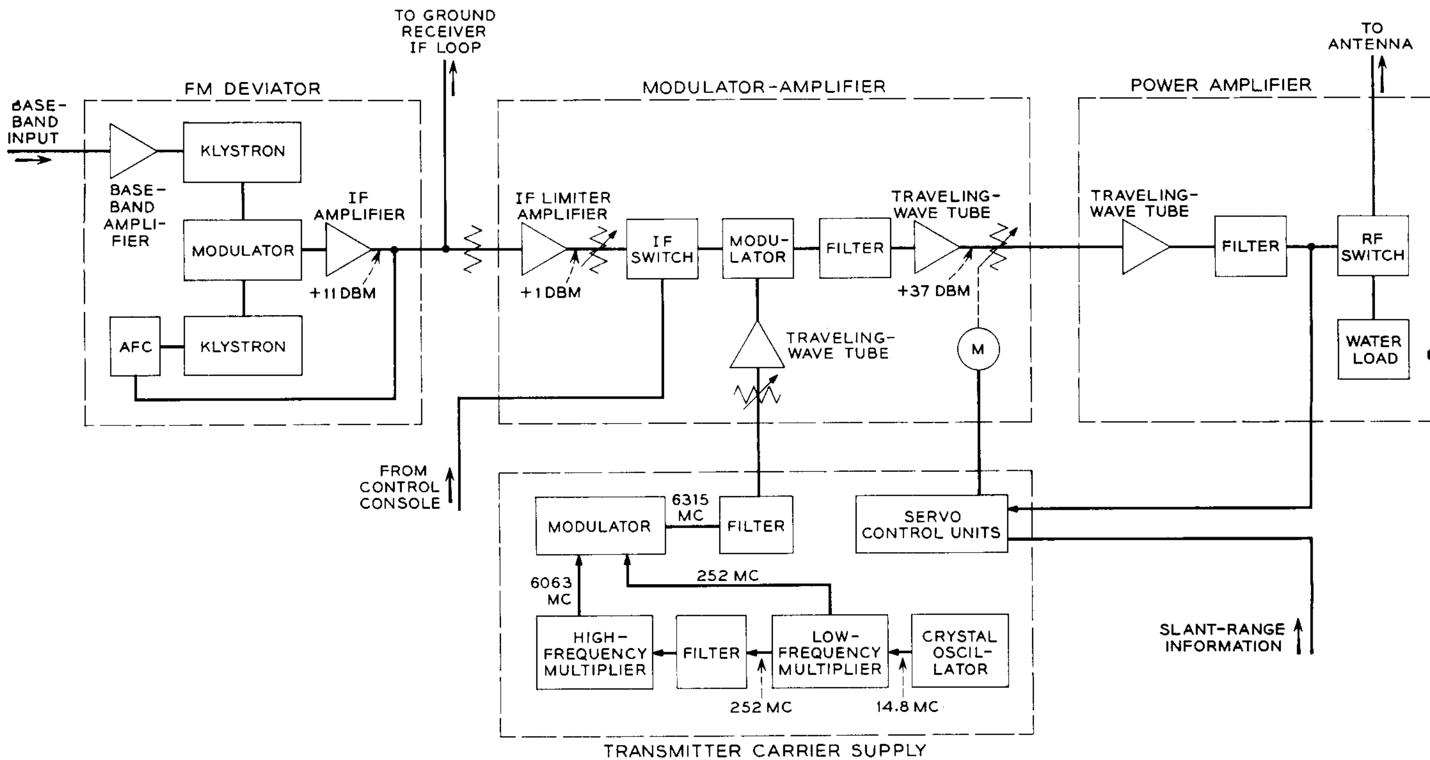

Fig. 1 — Block diagram of the ground transmitter.

TABLE I — TRANSMISSION OBJECTIVES, GROUND TRANSMITTER

Baseband response	2 cps to 5 mc
Peak frequency deviation	10 mc
RF and IF bandwidth	32 mc at 1-db points
Output power	2 kw, continuously variable from 2 kw to 0.02 kw
Center frequency:	
TV and 600-channel telephone	6389.58 mc
12-channel two-way telephone	6384.58 or 6394.58 mc
Frequency stability:	
FM deviator	
without modulation	±50 kc
with modulation	± [50 kc + 0.1 (rms frequency deviation)]
Transmitter carrier supply	±60 kc

mined the need to control the transmitter output. For 12-channel two-way telephony,* it is particularly desirable that the 6-kmc signals at the satellite be independent of slant range.

Operating practices are patterned after those established for domestic radio relay systems. Maintenance procedures, based on the use of standard Bell System test equipment,[7] enable all testing to be performed with the transmitter disconnected from the antenna. In addition, extensive alarm and control circuits have been incorporated in the transmitter to permit continuous, unattended operation.

1.3 *Description*

1.3.1 *General*

In this section the signal is traced through the transmitter; the following sections discuss the individual units of the transmitter in greater detail. Referring to Fig. 1, frequency modulation is generated by applying the television or multiplex telephone signal to the repeller of a reflex klystron. A 1-volt peak-to-peak signal at the 124-ohm balanced input to the baseband amplifier produces a peak deviation of 10 mc. The resulting frequency-modulated microwave signal is translated to the required intermediate frequency by combining it with the output of a second reflex klystron in a modulator. The output of the modulator, corresponding to the frequency difference of the two inputs, is the desired frequency-modulated IF signal. For television or 600-channel multiplex the center frequency is 74.13 mc; for 12-channel two-way multiplex it is 69.13 or 79.13 mc. Shifting the center frequency of the

* Twelve two-way channels is the number selected for this experiment and does not necessarily represent the capacity of the system.

FM deviator plus or minus 5 mc is a convenient means of obtaining either of the two transmitter frequencies required for two-way telephony. At the output of the wideband IF amplifier that follows the modulator, the signal power is +11 dbm. The automatic frequency control circuit samples the output and adjusts the repeller voltage of the second klystron to maintain a constant average output frequency.

In the modulator-amplifier the FM signal initially passes through an IF limiter, which removes any amplitude modulation, and then through an IF switch to the transmitter modulator. The controls of this high-speed switch are connected to the protective circuits in the power amplifier as well as to local and remote operating positions. In the modulator the IF signal is translated to the required RF frequency. The filter following the modulator permits only the signal corresponding to the sum of the frequencies of the IF signal and the 6315-mc output of the transmitter carrier supply to be amplified in the medium-power traveling-wave tube (TWT). At the output of this TWT the signal power is +37 dbm. The motor-driven waveguide attenuator at the output of the modulator-amplifier determines the input to the 2-kw traveling-wave tube in the power amplifier. The attenuator is part of the servo loop that varies the output of the ground transmitter to compensate for the changes in free-space path loss with satellite range. The filter following the 2-kw traveling-wave tube offers approximately 50 db suppression to second harmonic signals in the output. The waveguide switch and water load are part of the test facilities built into the power amplifier.

1.3.2 *FM Deviator*

The FM deviator is an adaptation of the FM terminal transmitter[5] designed for the TH radio relay system. The TH terminal transmitter is characterized by a 4-mc peak deviation capability, a 10-mc baseband response, excellent linearity, and an IF bandwidth of 32 mc. Increasing the gain of the original baseband amplifier was the only change required in the signal path to extend the peak deviation capability to 10 mc. Even at this increased deviation, the linearity of the Western Electric 450A reflex klystron, designed especially as a deviation oscillator for the TH FM terminal transmitter, is more than adequate to meet Telstar objectives. With a required baseband response of only 5 mc, it was possible to realize the additional baseband gain by simply increasing the interstage impedances in that amplifier.

While the basic method of automatic frequency control used in the TH equipment has been retained, increasing the peak deviation to 10 mc has necessitated extensive modifications in the AFC circuits. Requiring the FM deviator to operate satisfactorily with television signals for 405-, 525-, and 625-line systems further complicated the problem.

Referring to Fig. 2, the average frequency of the deviator output is compared with the frequency of a crystal-controlled oscillator. The AFC IF amplifier and the reference oscillator are alternately gated off-and-on at an $18\frac{1}{3}$ cps rate. (The reason for selecting this particular frequency will be discussed later in this section.) During one half of the gating period only the deviator output is effectively connected to the AFC discriminator; the dc output of the discriminator is proportional to the difference between the average frequency of the deviator and the center frequency of the discriminator. During the other half the reference oscillator is connected to the discriminator, and the dc output is then proportional

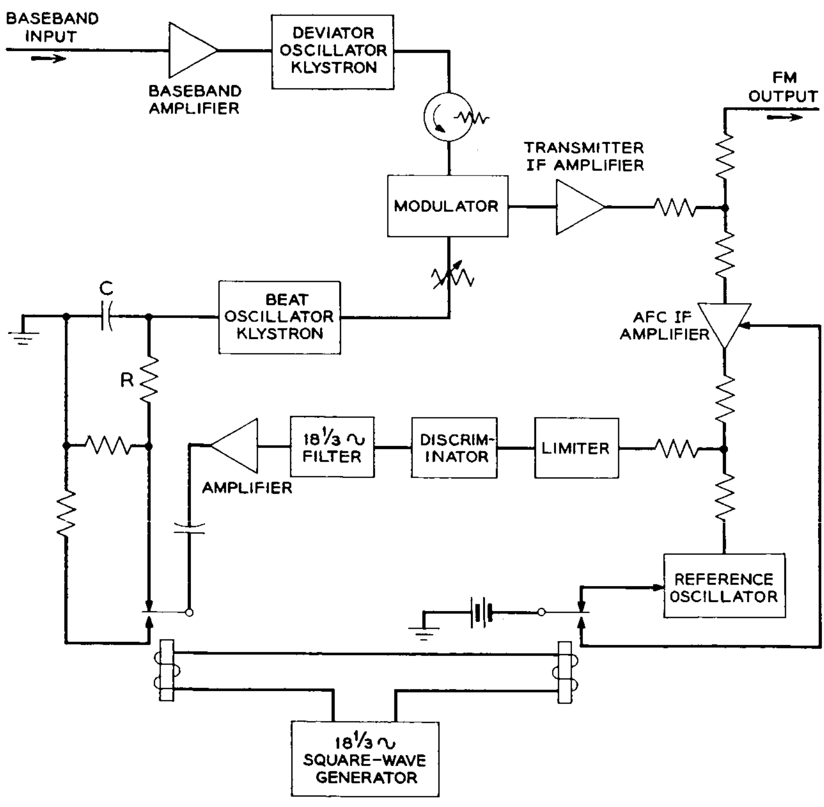

Fig. 2 — Block diagram of the FM deviator.

to the difference between the oscillator frequency and the discriminator center frequency. As a result, the AFC discriminator output contains an $18\frac{1}{3}$ cps square wave whose peak-to-peak average amplitude is proportional to the frequency difference between the reference oscillator and the average frequency of the deviator output. The function of the AFC limiter is to equalize signal amplitudes at the input to the discriminator. The bandpass filter following the discriminator effectively allows only the fundamental component of the $18\frac{1}{3}$ cps square wave to be amplified. At the output of the audio amplifier, the $18\frac{1}{3}$ cps sine wave is rectified in a synchronous detector and filtered. The dc signal is applied to the repeller of the beat-oscillator klystron with the proper polarity to reduce the frequency difference between the reference oscillator and the deviator output. Gating of the AFC IF amplifier and reference oscillator and the use of synchronous rectification eliminate the need for dc amplification and make the AFC loop insensitive to small drifts in the center frequency of the AFC discriminator. The open-loop gain of the AFC system is, typically, 36 db.

In the absence of modulation, frequency errors are in general the result of drifts in the klystron center frequencies. With modulation, the instantaneous frequency of the FM signal is continuously changing, and nonlinearities in the AFC discriminator characteristic will contribute terms to the average discriminator output that are not linearly related to the average frequency of the FM signal. To achieve comparable linearity at the increased peak deviation, the original AFC limiter-discriminator was replaced with the wideband limiter and discriminator of the TH FM terminal receiver.[5] Between 64 and 84 mc the first-order nonlinearity of the wideband discriminator is less than 2 per cent.

In gated AFC systems, objectionable "flicker" interference often arises as the result of beats between the harmonics of the gating frequency and low-frequency components of the television signal or power supply ripple. To make the FM deviator compatible with the 50-cps field rate of European television and the 60-cps U.S. standard, a gating frequency of $18\frac{1}{3}$ cps was chosen. Five-cps beats, developed between the 55-cps third harmonic of the gating frequency and either field rate, are effectively suppressed by the RC filter at the output of the synchronous detector. In addition, the bandpass filter at the output of the discriminator provides more than 20 db of attenuation to 50- and 60-cps signals before rectification At maximum deviation the 5-cps interference is more than 60 db below the 50- or 60-cps modulating frequency.

The wideband characteristics of the IF amplifier, limiter, and discriminator in the AFC loop make possible a simple method of shifting the deviator frequency for two-way telephone transmission. It is only necessary to substitute a 69.13- or 79.13-mc crystal in the reference oscillator, retune the oscillator, and adjust the beat-oscillator repeller supply so that the output of the synchronous detector is zero. The rest frequency of the beat-oscillator klystron is then 69 or 79 mc different from that of the deviation oscillator klystron.

1.3.3 *Modulator-Amplifier and Transmitter Carrier Supply*

The TH radio relay system provides the basic elements of the modulator-amplifier and the transmitter carrier supply. Detailed descriptions of this equipment and its performance are published elsewhere.[3,4] Modifications that affect operations are limited to the addition of the IF switch and the output power control system.

1.3.4 *Power Amplifier*

The power amplifier, including the power supplies and the TWT itself, is contained in five cabinets. Two of these are on the lower level of the antenna structure and contain the high-voltage rectifier and its controls. The other three are on the upper level and contain the high-voltage regulator, intermediate- and low-voltage supplies, and the TWT and RF equipment. This arrangement imposes the minimum space and weight requirements on the upper level of the antenna structure. In addition to the TWT heater and accelerator supplies, the intermediate- and low-voltage units include supplies for the solenoid focusing magnet, ion pumps, arc detector, and control and protective circuits. Water is supplied for cooling the TWT and also the dummy load. Air under pressure is brought into the cabinets for general cooling, supplemented where necessary by auxiliary blowers.

1.3.4.1 *The RF Circuit.* The M4040 TWT is described elsewhere in this issue.[6] Fig. 3 shows additional elements required in the RF path. Interposed between the modulator-amplifier and the TWT is the input circulator, the principal function of which is to provide isolation between the two for better impedance matching over the frequency band. In particular, as the input signal to the TWT passes through the tube, some small amount will be reflected at the sever (which is not a perfect match over the whole band). This power will reappear at the TWT input port with about 14 db amplification due to its passage through this first section of the TWT. If it undergoes a further reflection due to a mismatch at the modulator-amplifier output, a buildup will occur leading to oscillation and tube damage. A termination at the input circulator absorbs this reflected power, preventing regeneration.

At the output window of the TWT, a photomultiplier tube is mounted

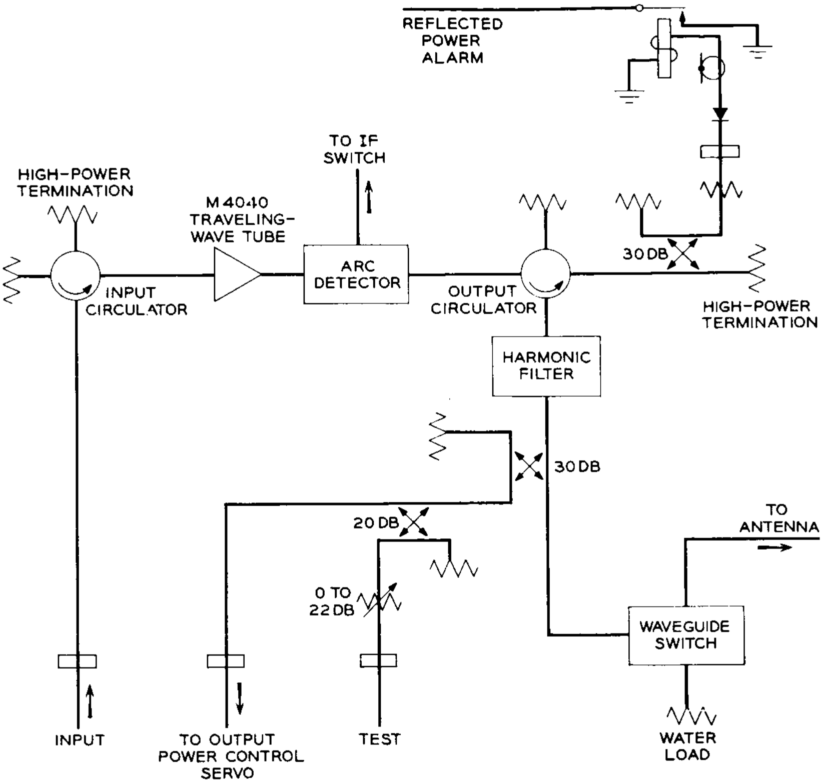

Fig. 3 — Block diagram of the power amplifier.

next to a small hole in the broad wall of the waveguide to detect any arcs that may develop in this region. Associated circuitry operates the IF switch, which removes drive from the tube, thus interrupting the arc.

A high-powered circulator is used in the output waveguide, again providing a minimum return loss of 14 db over the band to prevent oscillations in the output section due to excessive reflected power. In addition, one of the branches is connected through a 30-db directional coupler and an RF attenuator to a detector which monitors reflected power. If this exceeds 50 watts, the detector output operates a meter relay which opens the IF switch and removes the drive.

Following the output circulator, a harmonic filter reduces the second harmonic output of the transmitter, which may be only 25 db below the fundamental. To meet interference objectives, the filter is designed for 50-db minimum harmonic discrimination. To meet transmission objectives, the filter has 0.5-db maximum insertion loss and 30-db minimum return loss over the band of 5925 to 6425 mc.

Following the filter is a directional coupler and a waveguide switch and water load. The coupler is part of the output power control system, described in Section 1.3.5. The switch and water load permit all transmitter tests to be performed with the output disconnected from the antenna.

1.3.4.2 *DC Power Supplies*

(i) TWT Heater Supply. AC heater operation tends to cause small but measurable amounts of amplitude and phase modulation of the M4040 RF output. To eliminate these effects, a simple dc heater supply, consisting of a transformer, bridge rectifier and filter capacitor, is provided. A variable autotransformer on the power amplifier control panel determines the ac input to the power supply, and hence, the heater voltage. Normally, the heater supply is operated continuously to extend tube life and promote efficient outgassing. Meter relays connected to the primary of the heater supply cause high voltages to be turned off if the TWT heater current falls below a preset value.

(ii) High-Voltage Beam Supply. This supply, shown in Fig. 4, furnishes a regulated output of 0 to 20 kv at up to 1.5 amperes from 3-phase ac power at 50 or 60 cps. Its regulation for load changes from 1 to 1.5 amps, up or down, and/or line changes of ± 15 per cent, is better than 0.1 per cent peak-to-peak. Ripple and noise at rated load is less than 0.1 per cent peak-to-peak. The supply was designed with depressed collector operation of the TWT in mind for the future, so both plus and minus leads may be isolated from ground. Because of the requirement for regulated operation over the whole range of voltage from 0 to 20 kv, the supply was built with a motor-driven variable autotransformer to control the high-voltage applied to the rectifier, the output of which is further regulated by an electronically controlled series regulator tube. A limited degree of tracking between the autotransformer and the series tube control voltage is provided automatically.

The negative lead of this supply is fed to the TWT cathode at about 17 kv. The positive side, or ground end of the supply, is fed over separate leads to the collector and through the body-current sensing circuit to the body of the TWT. Regulated output voltage and collector and body current are monitored by meters on the control panel.

There are certain trouble conditions affecting the traveling-wave tube which might cause its destruction unless the high voltage is removed

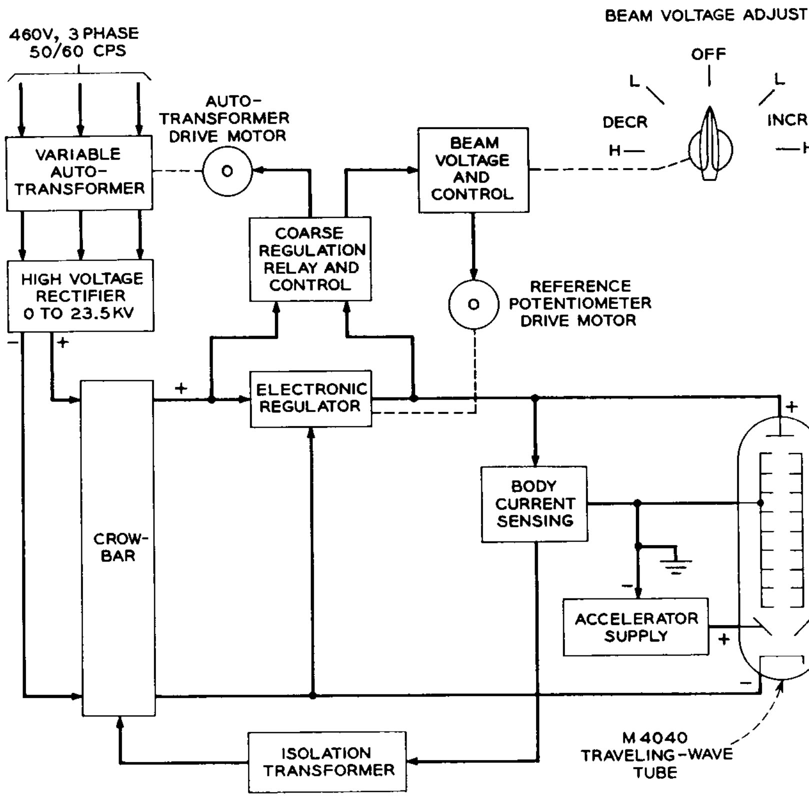

Fig. 4 — Block diagram of the high-voltage beam supply.

within a fraction of a millisecond. Protection of this sort is provided by the "crowbar" circuit, which can be operated by excessive TWT cathode current or body current. The crowbar tube is a type 5563A mercury vapor thyratron. This particular tube can be controlled by a grid voltage change of only 100 volts; the more usual hydrogen thyratrons or ignitrons require control voltages of several thousand volts. Thus, a simple and fast-firing circuit can be used. When fired, the crowbar puts a short circuit across the high-voltage rectifier, reducing the output voltage of the regulator to a few hundred volts in about 20 microseconds. After a few milliseconds delay due to relay operate time, the excessive current drain on the high-voltage rectifier causes its own protective circuits to shut it down.

(*iii*) *The Accelerator Supply.* The accelerator supply consists of a solid-state bridge rectifier and two stages of regulation. The first stage is a ferroresonant type of circuit, and the second stage is an electronic regulator. Line and load regulation and ripple measure less than 0.03 per cent. Variation of output voltage is controlled by a potentiometer across a voltage regulator tube reference supply. The ferroresonant first stage has a voltage control on the same shaft so that the two stages can be made to track. Under normal operating conditions the output voltage is adjusted to 600 volts. The output current monitor in the supply is part of the protective circuitry in the power amplifier. If the accelerator current becomes excessive, the beam supply and all intermediate-voltage supplies, except those for the ion pumps, are disabled.

1.3.4.3 *Cooling.* Deionized and filtered water is supplied to cool the M4040 TWT and the dummy load. Flow is set at about 18 gallons per minute and valved to the various using areas. Pressure switches are used as part of the protective system as an indication of proper flow in the TWT, the high-powered circulators, and the solenoid focusing magnet. If one or more of these flows are below a minimum, interlocks prevent the application of all but TWT heater voltage and the ion pump supplies.

In addition to the normal flow of filtered air through the cabinets, a spot blower is used to keep the TWT waveguide flanges cool, and another blower is used for the series tube in the beam regulator. A vane switch in the air stream of this latter blower is in the interlock circuit, again preventing the application of all but the heater, ion pump and solenoid voltages.

1.3.4.4 *Operating Interlocks, Controls, and Monitoring Devices.* Because of the experimental nature of the Telstar program, a greater ability to control voltages is necessary than would be expected of a commercial system. The manner of applying dc power to the TWT requires that the heater voltage be raised and lowered by carefully controlled amounts and that accelerator voltage and high voltage be applied in steps. Furthermore, the tube must be protected against arcs in the waveguide; excessive feedback or return power in the tube itself; overheating of the body, sever or collector; inadequate water flow to the body or the solenoid; and various over- or under-voltage or current conditions. The following sections describe some of these features.

(*i*) *Control Panel.* Operating controls for the power amplifier portion of the ground transmitter are located on a panel of the control bay. This control panel contains the power switches, the visual monitoring meters, status indicating lamps, alarm lamps and adjustment controls.

The status lamps indicate upper-level and lower-level standby power, heater on, plate ready, plate on, beam ready, beam on, beam adjust disable, and antenna/dummy load switch positions. The alarm lamps indicate such conditions as open circuit breakers, open interlock circuits, TWT water on or off, TWT over-temperature, over-current and under-current and over-voltage in certain circuits, high reflected RF power and RF drive disable. An audible alarm rings if the RF drive to the TWT is removed during normal operation. In addition to overload relays, additional protection is provided for the TWT by meter relays which will operate on out-of-limit readings to open the interlock protection circuit. Operation of the interlock protection circuit will remove power from the TWT.

(*ii*) *Circuit Protection and Operations Monitoring.* Individual circuits and functions are protected by 32 circuit breakers on a panel in the power bay in the upper level (there are additional circuit breakers in the beam rectifier control bay in the lower level). Elapsed time meters and operations counters are provided for heater, plate, and high-voltage monitoring.

1.3.4.5 *Mounting Arrangements and Maintenance Philosophy.* The equipment layout was planned with two particular considerations in mind: maintainability and personnel safety.

As much preventive maintenance and trouble analysis as possible is to be done in place, since bench maintenance facilities in the rotating structure are limited. To accommodate the objective of in-cabinet maintenance, sub-units in the control bay (20-kv regulator) are designed with test points of interest accessible from one side of the chassis, which is then mounted so that this side faces the front (or back) of the cabinet. Metal doors provide access to front and back of this cabinet and of the power bay (miscellaneous rectifiers), and inner plexi-glass panels prevent accidental contact with dangerous voltages. Small holes in these panels permit screwdriver adjustment to be made, or the insertion of a probe to some test points. In the case of the 20-kv regulator cabinet, opening these plexi-glass doors will operate interlocks, dropping the system back to the heaters-only condition, and will ground all high-voltage points with gravity-operated shorting bars. Most of the circuitry in this cabinet can be tested without high voltage being turned on, and in almost every case, troubles can be localized to the component requiring replacement without exposing maintenance personnel to undue risk.

The cabinets have been designed with RF shielding at the front and rear doors and some other openings, in order to keep the possibility of interference to the lowest possible level. Consideration of corona from high-voltage leads resulted in rounded corners and large-radius bends for potential corona sources. The cabinet containing the traveling-wave tube and its associated water-cooling, waveguide and RF measuring and monitoring gear was designed to isolate that part of the equipment which might "leak" RF. The TWT and the leads going to it which carry much of the high voltage, and the arc-detector and photomultiplier, are all in a metal-enclosed box, with the customary shorting bar for the 20-kv supply. Water valves and gauges are in a compartment to the rear of this, and temperature-monitoring pyrometers and RF-monitoring equipment in a compartment next to it. The doors to these two compartments are not interlocked. The traveling-wave tube bay, the control bay (20-kv regulator) and the power bay (intermediate voltage supplies) are of aluminum and are mounted in the upper structure near the end of the horn; the beam rectifier control bay and the beam rectifier bay are of steel, and are mounted below on the rotating platform.

1.3.5 *Transmitter Control*

Off-on control of the ground transmitter output centers around the IF switch. The switch consists of an IF diode gate[8] and a switch control circuit. In the off state the insertion loss of the switch is more than 85 db; there is essentially no input to the power amplifier. In the on state the insertion loss is 1.5 db, and the motor-driven attenuator in the modulator-amplifier effectively determines the output of the transmitter. In order to protect both equipment and personnel from hazardous operating conditions, the IF switch control circuit is inhibited by an extensive chain of relay and switch contacts. The chain encompasses the local and remote operating positions, the waveguide switch and protective circuitry associated with the power amplifier, and the motor-driven attenuator in the modulator-amplifier, as well as the antenna elevation indicators. An auxiliary connection exists between the arc detector in the power amplifier and the IF switch control circuit. It enables the switch control circuit to open the IF gate in less than 5 microseconds after an arc in the 2-kw TWT.

Fig. 5 shows a diagram of the output power control system. The arc detector, output circulator and harmonic filter have been omitted for simplicity. In loop 1 the input power to the RF detector remains essentially constant. Any change at this point unbalances the input to the difference amplifier, operates the polar relay and causes the shaded-pole servo motor to increase or decrease the attenuation of the AGC attenuator until the output of the M4040 TWT changes sufficiently to restore the RF detector input to its nominal value. As a result, the loop automatically compensates for compression in the input-output characteristic of the M4040 TWT and for changes in the RF output of the 5-watt TWT. The output of the transmitter is directly related to the attenuation of the reference attenuator and the manual attenuator.

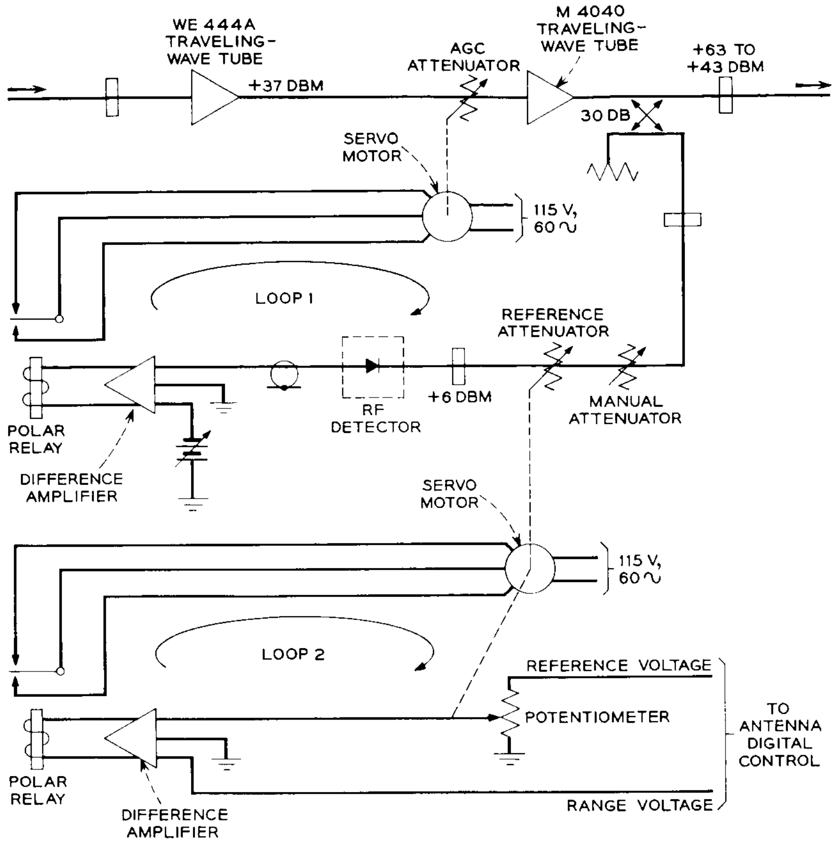

Fig. 5 — Output power control servo.

The control limits of loop 1 are 2 kw and 0.02 kw, equivalent to a 20-db variation in free-space path loss or a 10:1 variation of slant range.

The reference attenuator and hence the transmitter output are controlled by loop 2. The method of generating the two analog inputs to loop 2 is described elsewhere.[9] One is a fixed reference voltage; the second is proportional to the logarithm of the slant range or, in other words, free-space path loss. Expressed mathematically,

$$V_{\text{RANGE}} = 0, \qquad R < 0.1R_{\text{MAX}} \tag{1}$$

$$= V_{\text{REF}} \log_{10}\left(\frac{R}{0.1R_{\text{MAX}}}\right), \qquad 0.1R_{\text{MAX}} < R < R_{\text{MAX}} \tag{2}$$

$$= V_{\text{REF}}, \qquad R_{\text{MAX}} < R, \tag{3}$$

where R is the actual slant range and R_{MAX} is the maximum slant range at which it is possible to achieve a particular input power to the satellite. Since the transmitter is limited to 2 kw, the maximum slant range corresponding to a particular satellite input is determined by free-space path loss and transmitting and receiving antenna gains. For slant ranges equal to or greater than the maximum defined above, the voltage representing range is equal to the reference voltage; the reference attenuator in loop 1 and the potentiometer in loop 2 are set for maximum attenuation and resistance, respectively. As the slant range decreases, the voltage representing slant range also decreases. The inputs to the difference amplifier in loop 2 are no longer equal, and the servo motor adjusts the potentiometer to restore the equality. In so doing the attenuation in loop 1 and, consequently, the transmitter output are reduced. Since the loss in db of the reference attenuator and the resistance of the potentiometer are essentially linear functions of mechanical rotation, the decrease in transmitter output just equals the reduction in free-space path loss. As a result, the power incident on the satellite remains constant. In operation, a small second-order term is added to (2) to compensate for a slight nonlinearity in the attenuation vs mechanical rotation characteristic of the reference attenuator. With this addition, it is possible to maintain the signal incident on the satellite constant to within ± 0.5 db. The major portion of this error can be attributed to the dead space associated with the polar relays.

The IF switch and the output power control servo are interconnected with the ground station control console (GSCC) to permit remote operation of the transmitter from that location. Between passes, the transmitter output is connected to the water load and an out-of-service indication is displayed at the GSCC. Upon completion of pre-pass checks, the transmitter output is switched to the antenna and control is transferred from the local operating position to the GSCC. By activating appropriate switches, the console operator may choose any one of three operating conditions, provided the antenna elevation exceeds 5 degrees. These conditions are carrier off, standby, and carrier on. As long as the antenna is pointing below the horizon, however, the IF switch cannot be closed. The AGC attenuator is automatically positioned for maximum attenuation, independent of the output power control servo, and a carrier off indication is displayed on the GSCC. For antenna elevations above the horizon but less than 5 degrees, the choice is restricted to carrier off or standby. In the latter case, the IF switch is closed. However, the AGC attenuator is still positioned for maximum attenuation so that the radiated power is limited to approximately 0.2 watt. The standby condition is normally used during the pre-pass system checks with the satellite simulator at the test tower. For antenna elevations above 5 degrees, no restrictions exist. In the carrier on condition, the IF switch is closed and the AGC attenuator is under control of the servo system.

1.4 *Equipment Features*

As stated earlier, the high-voltage beam rectifier and associated controls are located on the lower level of the antenna structure. Fig. 6 shows the arrangement of the units of the transmitter mounted on the upper level. Cabinet doors have been temporarily removed. The transmitter carrier supply and modulator amplifier are located in the two pull-out bays on the left. To the right of the modulator-amplifier are the three remaining units of the power amplifier. The first cabinet contains the M4040 TWT and RF circuitry; the second cabinet contains the series tube regulator and crowbar circuits and the power amplifier control panel; and the third contains the circuit breaker panel and intermediate- and low-voltage supplies. The FM deviator is mounted in the cabinet on the extreme right. Each of the cabinets is 6 feet high and 3 feet deep. The combined width of the four cabinets and the two pull-out bays is $18\frac{1}{2}$ feet. The weight of the six units is approximately 6000 pounds.

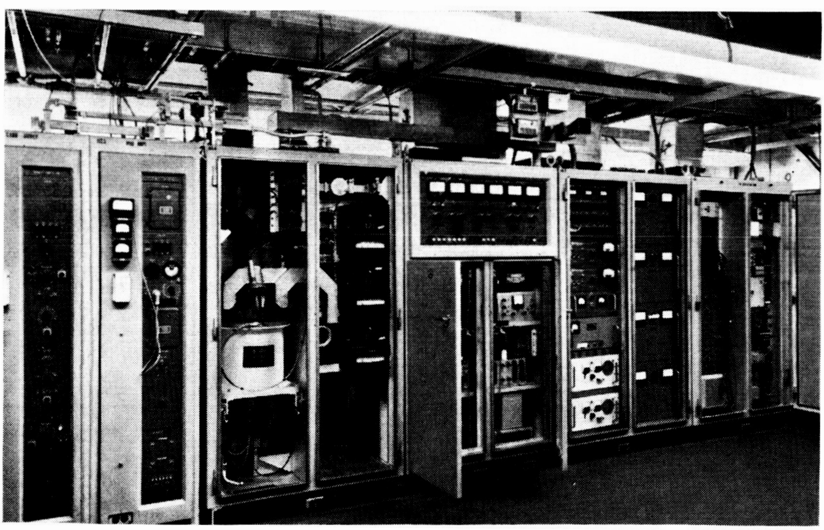

Fig. 6 — The ground transmitter.

II. THE RECEIVING SYSTEM

2.1 *General*

One of the major differences between a conventional microwave receiver — as used, for instance, in such ground-based microwave systems as TD2 and TH — and the ground receiver of the Telstar system, lies in the tremendously improved noise behavior of the latter. By the use of low-noise antennas and masers, the total noise temperature of satellite ground receivers can be made 100 to 200 times lower than for conventional receivers. The noise characteristics of the Telstar receiver are superior in still another respect. Present thinking calls for the use of wideband transmission methods, such as high-index FM, in order to keep the transmitter power in the satellite at a realistically low level. Such high-index systems necessarily work close to the threshold of detection. By the use of the FM with feedback (FMFB) principle, it is possible to extend the threshold of noise improvement in the Telstar system by about 4 to 5 db compared with the standard method of FM demodulation. This allows operation of the transmitter in the satellite with only one third the power otherwise required.

This part of the paper emphasizes the low-noise features of the receiver and describes the circuits responsible for such operation. An overall description of the receiver is given in Section 2.2. The pertinent operational characteristics of the maser, the parametric amplifier, and the FMFB receiver are described in Section 2.3. A description of the diplexer and other waveguide equipment required to connect the transmitter and receiver to the same antenna is found in Section 2.4. Various system noise measurements are presented in Section 2.5. Circuits of a more conventional type and those taken from the TH or TD2 microwave system, such as converters, IF amplifiers, and carrier supplies, are not described.

2.2 *Receiver Description*

2.2.1 *The Transmission Circuits*

A block diagram of the ground receiver is shown in Fig. 7. This equipment is located in the upper equipment room on the horn antenna structure. The received 4-kmc, left-hand circularly polarized signal from the satellite, after being picked up by the horn-reflector antenna, is guided toward the maser by a low-loss waveguide diplexer. The diplexer

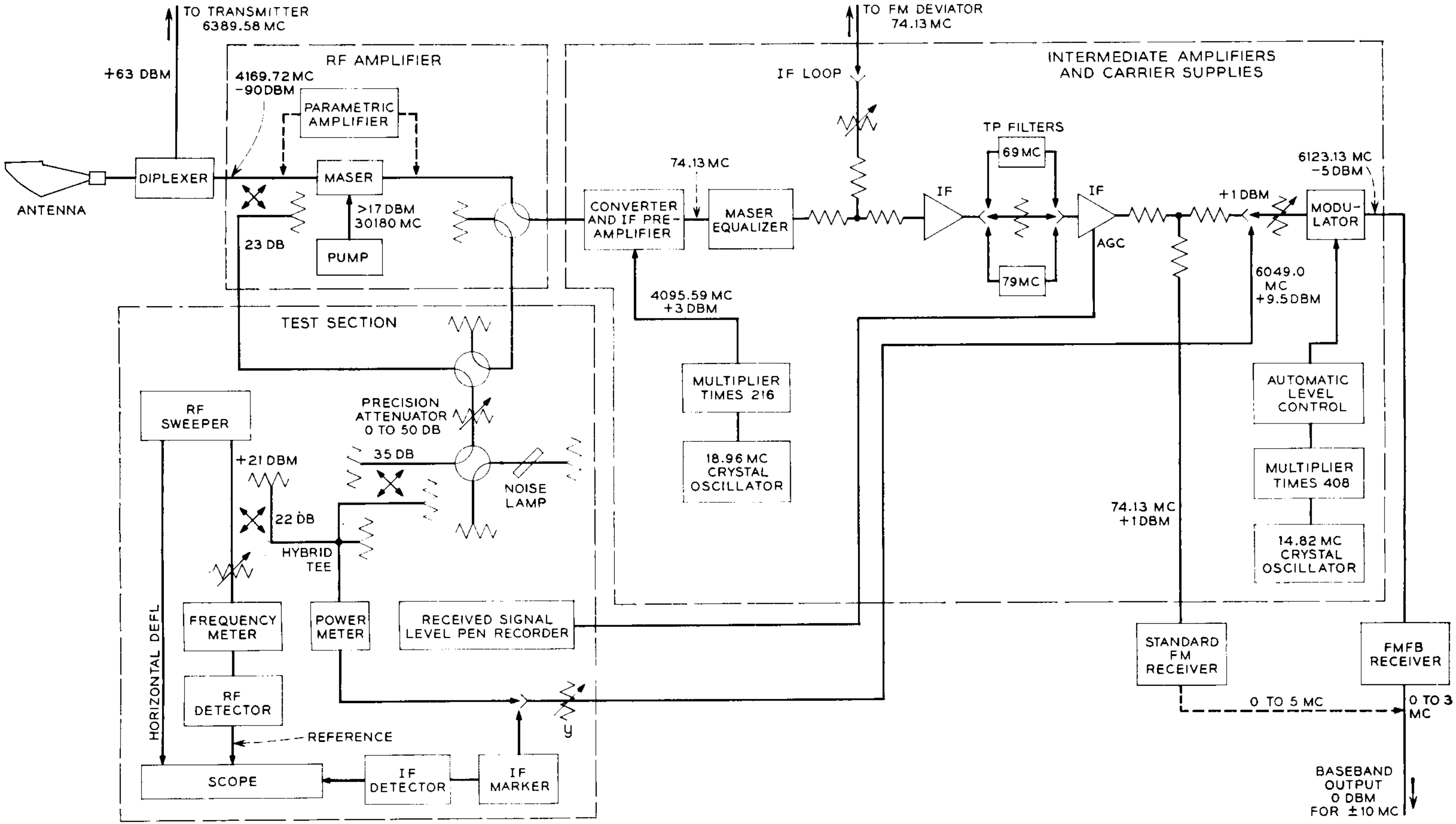

Fig. 7 — Block diagram of the ground receiver.

also accepts the high-power 6-kmc signal from the transmitter and feeds it to the same horn-reflector antenna, from where it is launched right-hand circularly polarized. The maser, operating at the temperature of liquid helium (4.2° Kelvin), has a gain of 42 db and a 3-db bandwidth of 16 mc. The receiver bandwidth is widened to the necessary 25 mc by the insertion of a maser equalizer at IF directly after the frequency converter. The 42-db gain of the maser is high enough to make the noise contribution of the following frequency converter negligible. Total noise temperature of the receiver measured at the input of the maser with the antenna pointed toward zenith is 32°K on a clear day. The maser requires for its operation a "pump" which supplies a frequency-stabilized microwave carrier of 30,180 mc at a power greater than 17 dbm.

It is also possible to use a broadband (60-mc) low-noise parametric amplifier instead of the maser for evaluation in the system.

The signal from the maser is changed to an intermediate frequency of 74.13 mc in a frequency converter. A crystal-derived carrier of 4095.59 mc serves as the local oscillator signal. A low-noise preamplifier and two more IF amplifiers bring the signal power up to a level of +1 dbm, which is held constant by an automatic gain control circuit in the second amplifier.

The transmission characteristic of the converter and IF circuits combined is flat and drops only 0.25 db at the edges of the 25-mc band. Except for the case where the parametric amplifier is used, the over-all transmission is limited by the combination of maser and maser equalizer. For TV transmission this bandwidth is 25 mc at the half-power points, and for two-way telephony it is reduced to 3 mc by the insertion of a special filter ahead of the last IF amplifier. This telephone filter is centered either 5 mc above or below the normal IF frequency of 74.13 mc, depending on the transmitting frequencies chosen in the two participating satellite ground stations. The insertion of the telephone filter is the only step necessary for changing the receiver from television to two-way telephone operation.

To the output of the last IF amplifier can be connected either a standard (conventional) FM receiver or an FMFB receiver after suitable frequency translation. The input frequency of the FMFB receiver is at 6123.13 mc for reasons explained under Section 2.3.3. A balanced diode modulator, together with a crystal-derived carrier of 6049.0 mc, translate the 74.13 mc signal to the required 6123.13 mc.

2.2.2 *The Test Equipment*

In addition to the basic transmission circuits just described, the Telstar ground receiver contains special test equipment for fast and precise measurements of noise temperature, transmission characteristics, gain and received signal level. This equipment is shown in the block called test-section of Fig. 7. It includes an RF sweeper which can be swept over 60 mc anywhere in the 3700- to 4200-mc range, a power meter for 4 kmc and 74 mc with an accuracy of a few tenths of a db, a waveguide noise lamp, a precision waveguide attenuator, an oscilloscope, and a pen recorder. A number of waveguide switches allow the distribution of either the 4-kmc carrier or the noise signal to either the input of the maser or the input of the converter. The output, for test purposes, is taken at the +1 dbm IF point.

The AGC voltage of the IF amplifier is a function of the received 4-kmc signal level and is recorded on a pen recorder. A wealth of information can be obtained from such recordings. Several examples of received level recordings are given in a companion paper.[10] The recorder is calibrated before each pass with a 4170-mc signal from the RF sweeper, which is applied to the input of the maser at the levels of −80 and −90 dbm. Absolute accuracy of the recording is about ±1 db. A second recorder of received level is located on the ground station control console in the control building. Its expanded range, going as low as −110 dbm, allows the personnel tracking the satellite to check whether the autotrack system has locked on the main beam and not on a sidelobe of the horn-reflector antenna. The recording also helps an experienced operator to easily discover troubles in the communications part of the system.

Whereas it is possible to test the entire receiver at Andover by means of an "RF loop" which includes the transmitter and the transponder on the test tower, it is often simpler to test important parts of the receiver in the "IF loop" (Fig. 7) which interconnects transmitter and receiver at intermediate frequencies. In this case, random noise from the input of the receiver can be injected into the communications channel for noise and threshold tests. For such tests the antenna should be pointed at the zenith to ensure stable noise power.

In addition to the integrated test equipment shown in Fig. 7, other specialized test sets are used to maintain all the circuits of the receiver on a routine basis. Tests to be performed include the measurement of IF-to-IF and baseband-to-baseband transmission characteristics, discriminator sensitivity and linearity, and IF and RF return loss. The receiver can also be left in continuous and unattended operation due to the provision of alarm circuits with remote indication.

2.3 *Circuits for Low-Noise Amplification and Demodulation*

2.3.1 *The Maser*

The heart of the receiver is the low-noise maser amplifier which is described in detail in a companion paper.[11] The single-tuned (or equalized) maser, as described in this reference, was used in the receiver rather than the low-gain stagger-tuned version. This choice resulted in a substantial saving of equipment by eliminating a low-noise second-stage RF amplifier. The single-tuned maser has a gain of 42 db and a half-power bandwidth of 16 mc. At a bandwidth of 25 mc, the maser gain is still 35 db. The noise contribution of the converter with its 12-db noise figure or 4300°K noise temperature, referred to the input of the maser, amounts to a negligible 0.27°K at midband and to 1.36°K at the band edges. The stagger-tuned maser, on the other hand, has a flat gain of about 27 db, and the frequency converter would contribute 8.6°K unless the maser were followed by a low-noise second-stage RF amplifier, such as a traveling-wave tube or a parametric amplifier. The terminal noise temperature of the maser alone is 3.5°K.

The relatively low level at which a maser begins to saturate fortunately does not pose a problem in a satellite communications system, as can be seen from the following. The ruby maser shows a one-db drop from its small signal gain at an output level of about −40 dbm. The single-tuned maser, therefore, will begin to saturate at a single-carrier input level of −82 dbm. The highest expected received signal level is about −75 dbm, at which the gain of this maser is 3.5 db below the maximum. If, instead of a single carrier, a signal with the same power but a flat spectrum of 25 mc, resembling an actual communications signal, were applied to the maser, the saturation effects would be smaller. The automatic gain control circuit of the IF amplifier compensates not only for the variations in received signal level but also for the changes of maser gain due to saturation. Saturation also causes a widening of the maser transmission characteristic. This effect changes the transmission by only a few tenths of a db up to input levels of −75 dbm and can therefore be neglected for all practical purposes.

The saturated maser, unlike an electron tube or transistor amplifier, does not show signs of instantaneous nonlinearity, which would give rise to intermodulation or signal clipping. The maser acts much like an extremely linear amplifier followed by a voltage divider containing a thermistor whose resistance is changed by the power in the amplified signal. As in this thermistor circuit, the maser gain reacts to a change in level very slowly, with a time constant of about 0.1 second.

The extremely low instantaneous nonlinearity predicted by maser theory was also confirmed experimentally. For this purpose, two strong signals of equal power at frequencies f_1 and f_2 were applied within the passband of the maser, and the third-order intermodulation products $2f_1 - f_2$ and $2f_2 - f_1$, falling again into the maser passband, were measured. If both signals f_1 and f_2 were at the level of -85 dbm, which is possible during two-way telephony operation, the intermodulation terms would be about 180 db lower. It is clear that such low distortion levels are completely negligible. One is also led to the conclusion that the maser is the most linear amplifier in existence.

For the design of the waveguide diplexer, it was necessary to determine how much of the 6390-mc transmitting signal could reach the input of the maser without causing saturation or other effects. Whereas saturation is evident around -82 dbm for signals in the passband (4170 mc), signal levels as high as 2.5 watts (34 dbm) can be applied to the (impedance-matched) maser input at 6390 mc without causing intolerable distortions of the transmission characteristic. In the latter situation, however, the helium boil-off rate will be increased considerably. Because we have no control over the impedance match of the maser at 6 kmc, it is not advisable to apply interfering signal levels in excess of a few hundred milliwatts.

It is possible to saturate the maser from the output side as well. Due to balancing and filtering in the converter, the leakage of the 4095.59-mc local oscillator signal is low enough not to cause saturation of the maser.

The over-all receiver bandwidth is widened to the required 25 mc at the half-power points by inserting an equalizing filter at IF. A simple bridged T single-pole filter is used with a loss of 7 db at midband and 3 db at the 25-mc bandwidth. The resulting transmission characteristic has two peaks of about 0.3 db, 5 mc away from the center frequency.

The magnetic field of 3300 gauss which tunes the maser to a frequency of 4170 mc is provided by a permanent magnet. Because of the high tuning sensitivity of 2.4 mc per gauss, it is necessary to avoid large temperature variations of the magnet, which will change the magnetic field at a rate of 0.4 gauss per degree centigrade and will therefore lead to a temperature sensitivity of 0.96 mc/°C. It is very helpful in this respect that the temperature of the upper equipment room in which the receiver is located is kept within a few degrees. It has further been found important that magnetic material like metal chairs should not be brought close to the maser to avoid detuning. It is even possible to notice the effect of the earth magnetic field, which amounts to about 0.16 gauss at Andover, Maine. When the horn antenna is turned in azimuth, the earth field will add or subtract from the maser field, and a slight tilting of the transmission characteristic can be observed. (See Fig. 19 of Ref. 10.)

The operation of the maser requires a supply of liquid helium and nitrogen. The handling of these cryogenic fluids has been simplified considerably over standard laboratory procedures by the availability of new 100-liter containers and transfer tubes. The maser accepts 10 liters each of liquid helium and nitrogen, and boil-off occurs in about 20 and 50 hours, respectively. No attempts were made to recover the expended helium gas because this was considered to be an interim situation which would be changed in time by the installation of a closed-cycle helium liquifier.

A ruby maser operating at a frequency of 4170 mc must be pumped with a signal at 30,180 mc. The pump power at the maser terminals should be greater than 50 milliwatts if the maser gain is not to drop more than about 2 db from its maximum. The pump frequency should remain within a few mc of the correct value to ensure maximum pumping efficiency and to avoid loss of gain. A reflex klystron serves as the source of pump power. Good long-term frequency stability is provided by the AFC circuit shown in Fig. 8, which has the following features: (*i*) the reference cavity is made low Q in order to produce a peak-to-peak separation of 30 mc in the Foster-Sealey type waveguide discriminator, (*ii*) a wideband (100 kc) dc amplifier stabilized by a mechanical chopper is used, and (*iii*) the repeller of the klystron is operated at about ground potential. The so-often unreliable voltage translation from ground potential to the high dc potential of the repeller is therefore avoided. This requires operation of the klystron shell at a high potential and makes the use of a high-voltage waveguide choke necessary. The RC circuit between the dc amplifier and the klystron repeller determines the stability of the feedback loop.

2.3.2 *The Parametric Amplifier*

A parametric amplifier[12] can be inserted into the system instead of the maser for evaluation of its performance in a satellite system or as a back-up for the maser. The amplifier consists of two cascaded stages of

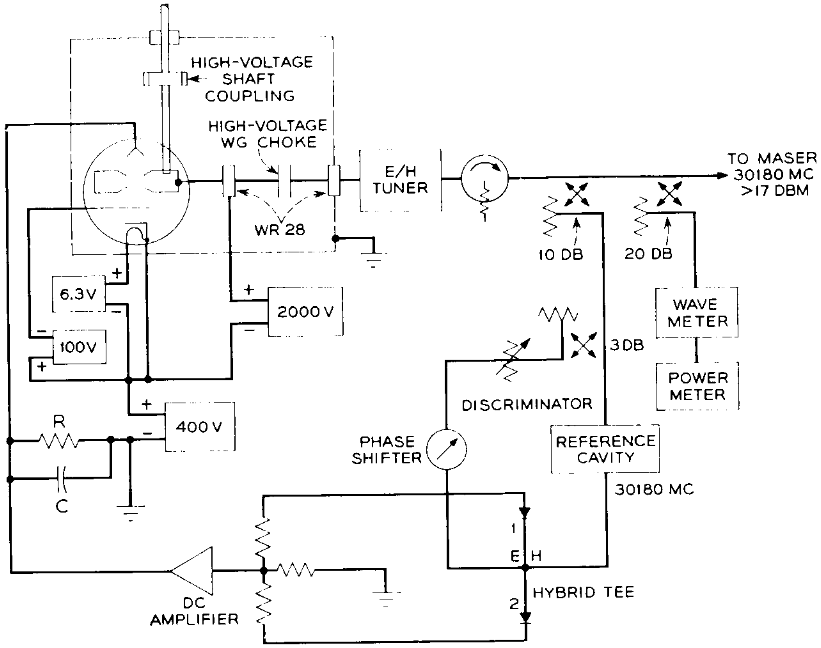

Fig. 8 — Circuit diagram of the maser pump.

similar design. The first amplifier is operated at liquid nitrogen temperature (77°K) and has a gain of 20 db, while the second works at room temperature with a gain of 18 db. The over-all gain of 38 db is therefore close to the 42 db of the maser. The room temperature parametric amplifier with its noise temperature of about 130°K serves as a buffer against the 4300°K noise of the converter circuit. The total noise temperature of the parametric amplifier, including converter, is about 50°K.

The bandwidth of the amplifier is 60 mc at the half-power points. It is obvious that no equalizing filter is needed at IF as in the case of the narrower maser. The over-all 3 db bandwidth of the ground receiver is then 32 mc as determined by the narrower waveguide filters and IF circuits. Saturation in this parametric amplifier, as expressed by a one-db drop from maximum gain, sets in at an input level of −50 dbm. It has been found that intermodulation is no problem for the typical satellite signals of −85 dbm. If a 6390-mc signal is applied to the input of the amplifier at a level greater than 0 dbm, then its 4-kmc transmission will be affected. A band rejection filter can be built into the first parametric amplifier, which raises the tolerable 6-kmc level to about 17 dbm without causing an increase in the noise temperature of the amplifier.

The good over-all stability of this parametric amplifier is due mainly to the stable design of the 23-kmc pump source. The same source is used to drive both stages of the amplifier. Refilling of the dewar with liquid nitrogen is required only every 10 days.

2.3.3 *The FMFB Receiver*

An FMFB receiver is part of the ground receiver. It serves to extend the threshold of noise improvement by 4 to 5 db over that obtained with a standard FM receiver. The feedback circuit is capable of demodulating high-index FM signals with peak frequency deviations of ± 10 mc. The circuit can be used for the reception of television, telephone multiplex and other types of signals.

An integral part of any FMFB receiver is a voltage-controlled oscillator. Such an oscillator should have good modulation linearity, high modulation sensitivity, small signal delay, and good long-term frequency stability. All these features were found in a 6-kmc reflex klystron. This made it necessary to enter the circuit at a frequency of 6123 mc, as indicated in Fig. 7. No further circuit details will be given here; the reader is referred to a companion paper[13] for a detailed description of the FMFB receiver.

2.4 *Diplexer and Associated Waveguide Circuits*

2.4.1 *The 4 kmc - 6 kmc Diplexer*

The waveguide circuit which allows the connection of the 6-kmc transmitter and the 4-kmc receiver to the same horn-reflector antenna is shown in Fig. 9. Although at present this diplexer is only required to work properly over a 25-mc band around the transmitting frequency of 6390 mc and the receiving frequency of 4170 mc, it was designed to handle a much wider band of frequencies. Several components are capable of operating over the full 4-kmc and 6-kmc common carrier bands, extending from 3700 to 4200 mc and from 5925 to 6425 mc, respectively. Other parts are more narrow-band, but studies are in progress to make the circuit work over the full 500 mc of the two common carrier bands.

Returning to Fig. 9, the 2-kw (63-dbm) signal from the 6-kmc transmitter arrives at the diplexer from the left in a rectangular waveguide and with horizontal polarization. The signal travels essentially unchanged over a waveguide transition and through the polarization coupler. The 6-kmc signal which leaks through the coupler into the arm towards the maser is at least 25 db lower. The polarizer then transforms

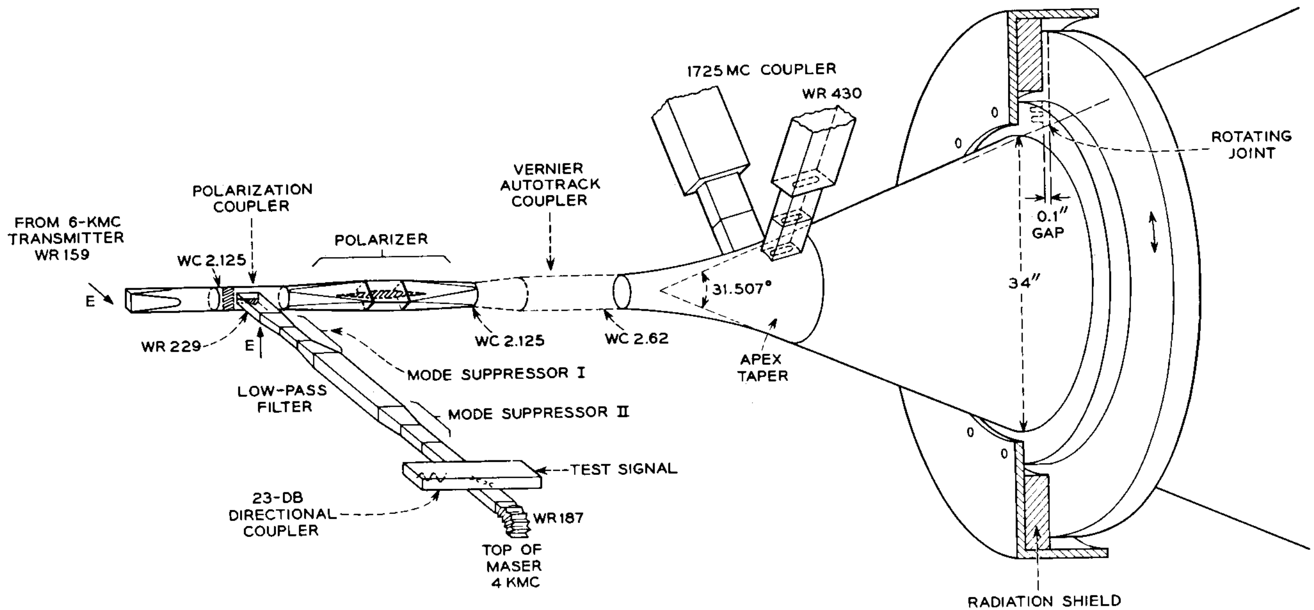

Fig. 9 — Sketch of diplexer and associated waveguide circuits.

the linearly polarized 6-kmc signal into circular polarization for launching by the antenna.

The 4-kmc signal from the antenna, after going through the polarizer, appears as a vertically polarized wave which is guided through the polarization coupler with little loss into the rectangular waveguide leading to the maser. The polarization coupler is of the same type as used in Project Echo, but is designed to work over both the 4- and 6-kmc common carrier bands.[14]

The diplexer was designed under the assumption that the 6-kmc signal level seen at the 4-kmc output terminal shall never exceed 0 dbm. This level is considerably below the value which is acceptable for the maser. It was chosen to allow the use of different types of parametric amplifiers. (See Section 2.3.2.) The diplexer therefore has to provide a total of 63 db of isolation for the 6-kmc signal. With a minimum guaranteed loss of only 25 db in the polarization coupler, additional loss is introduced by a waveguide low-pass filter. This filter was originally designed for the TD2 microwave system, working in the 4-kmc common carrier band, to suppress leakage from the TH system (6-kmc band). The filter has more than 60 db of attenuation for the fundamental mode, but negligible loss for the TE_{20} mode at 6 kmc. It also has an insertion loss of less than 0.09 db over the 4-kmc band. Generators of TE_{20} modes are the polarization coupler and the directional coupler on the maser side. Two mode suppressors were therefore inserted as shown in Fig. 9, consisting of a short piece of narrow-width (1.79-inch) rectangular guide and associated transitions. Each suppressor has little loss at 4170 mc but more than 40 db of attenuation for the TE_{20} mode at 6390 mc. With these precautions, the isolation of the diplexer is found to meet the 63 db isolation with a substantial margin.

The 23.35-db directional coupler at the maser is used for measuring the receiving system noise temperature and for gain and transmission measurements. A short piece of flexible waveguide connects the waveguide circuit to the top flange of the maser.

Because it is necessary to transmit a right-hand circularly polarized wave at 6 kmc and to receive a left-hand circular wave at 4 kmc, a polarizer is inserted between the polarization coupler and the antenna. The polarizer, suitable for operation at 4 and 6 kmc, is shown in Fig. 10 together with a plot of its performance in terms of axial ratio. For pure circular polarization to exist at the output of the polarizer, it is necessary for the field components parallel and normal to the dielectric plate to be equal in magnitude and 90° out of phase. If a differential loss or a deviation from 90° exists, the wave will be elliptically polarized, and

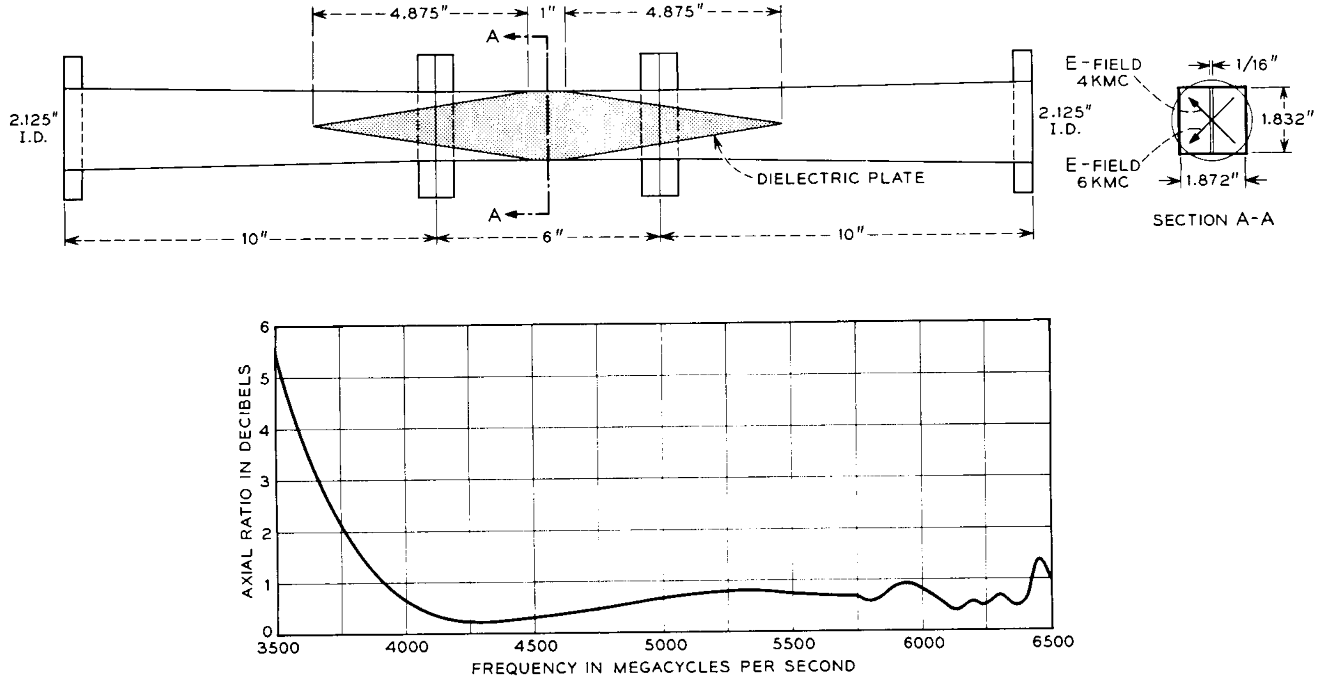

Fig. 10 — The polarizer.

the ratio of major to minor axis, the axial ratio, will be greater than 0 db. Fig. 11 shows how the axial ratio depends on the phase difference and on the differential loss between the two field components. The differential loss was kept below 0.01 db in the polarizer of Fig. 10 by using an extremely low-loss dielectric plate. A phase difference close to 90° is maintained from 4000 to 6500 mc in the polarizer shown. This was made possible by making use of the fact that a dielectric plate in a square waveguide produces a differential phase versus frequency characteristic which is opposite to the one obtained in a nonsquare guide. By locating the dielectric plate in a slightly nonsquare guide, a compensation over a wide frequency range is possible. As a consequence, the axial ratios plotted in Fig. 10 are found to be quite low over a frequency range of 2500 mc.

If imperfect polarizers are used in a system, they will cause a transmission loss. Fortunately, these losses are quite small and amount to only a few tenths of a decibel for the polarizer of Fig. 10 when working with a satellite whose axial ratio is lower than 4 db.

The diplexer is well matched over wide frequency bands at 6390 and

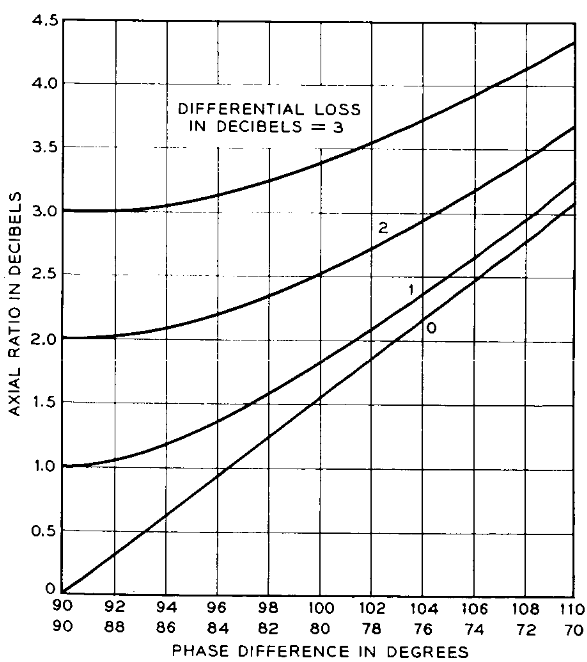

Fig. 11 — Dependence of the axial ratio of a polarizer on differential loss and phase difference.

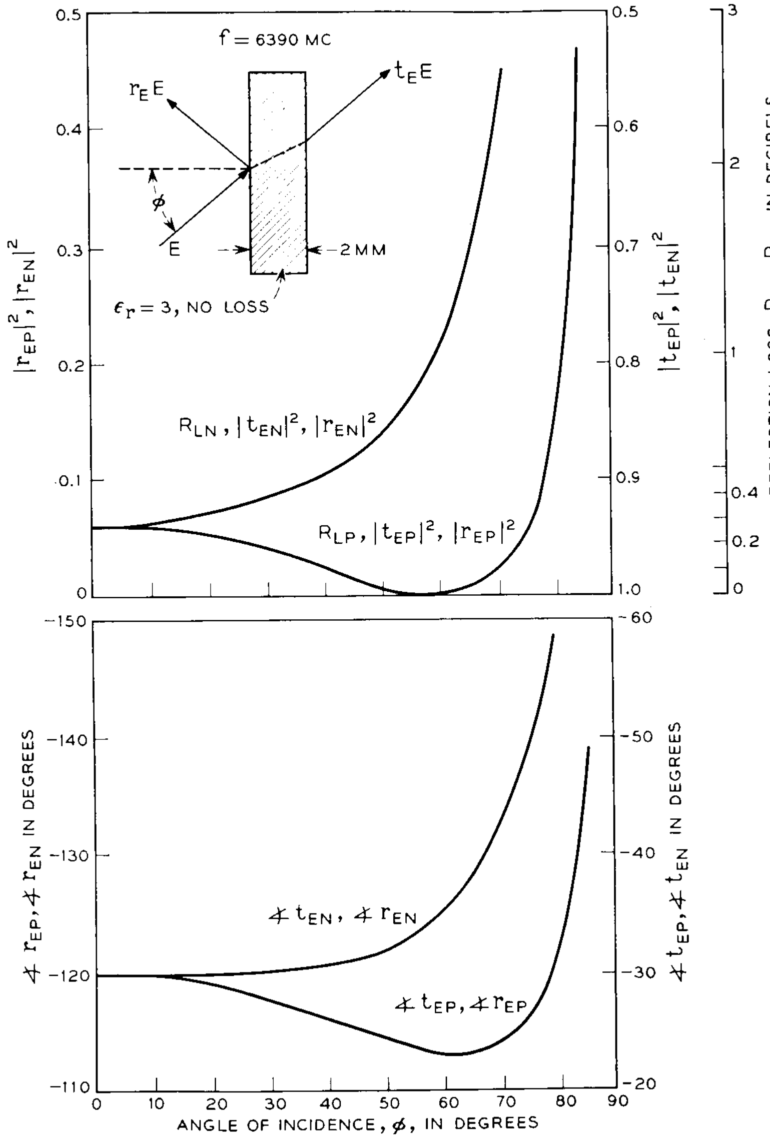

Fig. 12 — Reflection and transmission coefficients of the (lossless) radome at 6390 mc.

4170 mc, and the return losses at the three diplexer terminals are 30 db or higher.

2.4.2 *Effect of the Radome on the Operating Characteristics of the Diplexer*

The 4- and 6-kmc signals are affected in a number of ways by the radome which surrounds the horn-reflector antenna. The possible effects on the operating characteristics of the diplexer will be investigated here. The radome, for instance, will reflect the right-hand circularly polarized transmitted wave as a left-hand circular wave which, after being picked up by the horn antenna, will be directed towards the maser by the polarization coupler. The isolation of the polarization coupler, and in turn the isolation of the diplexer, could be reduced by this reflection. Measurements were made in the laboratory to determine the dielectric constant of the radome material. The power reflection $|r_E|^2$ and transmission $|t_E|^2$ coefficients and their phase angles were then calculated as a function of the angle of incidence Φ. Electric fields parallel (index P) and normal (index N) to the plane of incidence were considered. The results are shown in Figs. 12 and 13 for 6390 and 4170 mc, respectively. According to Fig. 12, the power reflection coefficient of the radome at 6390 mc, averaged over the possible angles of incidence from 8° to 52° and parallel and normal polarization, amounts to about 0.065, which corresponds to a return loss of only 12 db.

Only if all the reflected components from the radome appeared in-phase over the aperture of the antenna could a return loss of 12 db be seen at the apex of the horn. This obviously is impossible. The response of the antenna to reflections from the radome is very similar to its far-field response far away from the main beam. Measurements have indicated that the return loss of the radome as seen from the apex is much greater than 46 db. This figure is considerably higher than the 25 db of isolation provided by the polarization coupler. It can therefore be concluded that the isolation of the diplexer is not affected by the radome.

The radome material can also change the axial ratio of the circularly polarized signals and, therefore, affect the characteristics of the polarizer in the diplexer. The circularly polarized signal arriving at the radome can be decomposed into two linearly polarized waves which can be assumed to be parallel and normal to the plane of incidence. The curves of Figs. 12 and 13 through the voltage transmission coefficients t_{EP} and t_{EN} then tell how much differential loss and phase error is introduced by the radome. For angles of incidence up to 20° for 4170 mc and up to 12° for 6390 mc, no differential effects can be found. At 52°, the highest

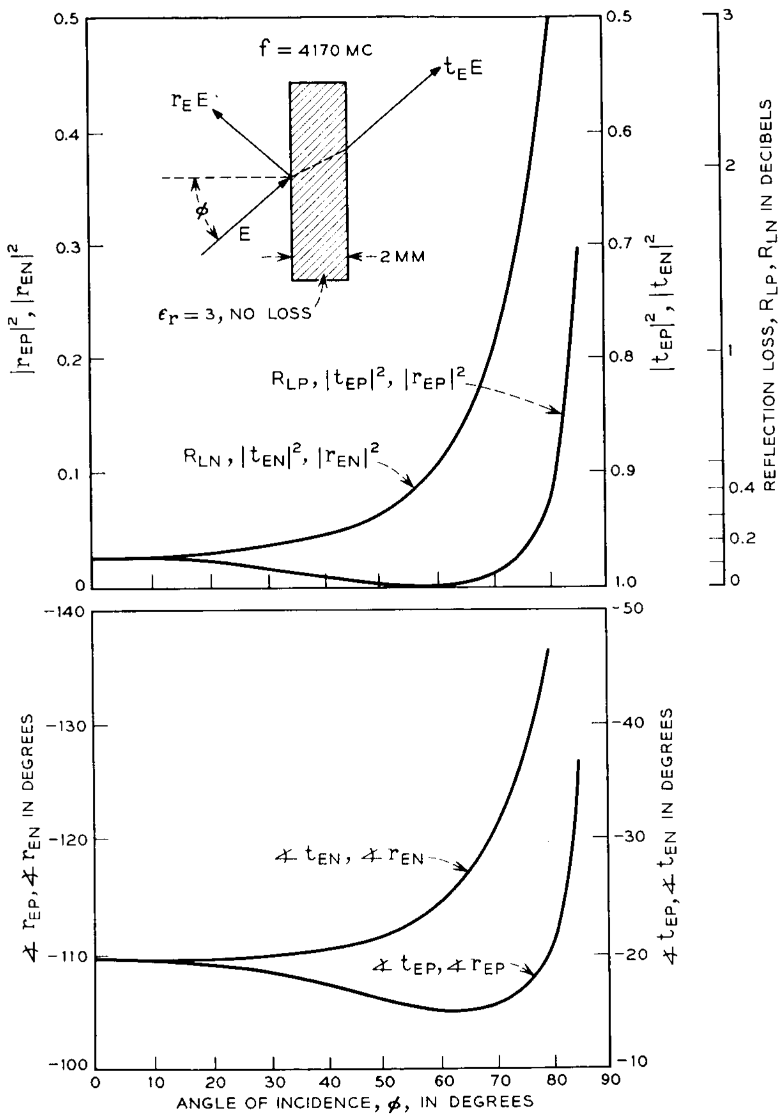

Fig. 13 — Reflection and transmission coefficients of the (lossless) radome at 4170 mc.

possible angle for the Andover horn, the differential losses amount to 0.30 and 0.65 db and the phase errors to 6.50 and 9° for the 4170- and 6390-mc frequencies, respectively. Using Fig. 11, this results in axial ratios of 1.08 db and 1.52 db for the two frequencies. Fortunately, only a very small part of the energy is radiated (or received) at angles as high as 52°, and no significant loss of signal will result from the degradation of axial ratio by the radome.

2.4.3 *Other Waveguide Circuits*

Following the polarizer shown in Fig. 9, an autotrack coupler is inserted into the line to the antenna. Its purpose is to extract, over narrow-band coupling slots, the TE_{11} and the TM_{01} modes which are generated in the horn antenna by the 4080-mc beacon signal from the satellite. The TM_{01} mode, which propagates freely in the 2.62-inch circular waveguide of the coupler, is completely reflected in the taper leading to the smaller sized polarizer guide. The autotrack coupling section has about 0.01 db of loss for the 4170-mc signal and does not affect transmission of the 6-kmc signal. A complete description of this coupler is contained in a companion paper.[15]

A taper whose impedance varies exponentially with length connects the 2.62-inch circular waveguide to the 31.507° conical apex section of the antenna. The 12.25-inch long taper has a calculated return loss of more than 40 db for the TE_{11} mode and more than 30 db for the TM_{01} mode over the frequency range of 3700 to 6500 mc.[16]

The 1725-mc transmitting signal used in NASA's Project Relay is coupled to the antenna in the conical section directly following the apex taper. Two 1725-mc signals of equal amplitude and 90° phase difference are coupled at right angles into the horn. The phase difference is such as to give a signal right-hand circularly polarized in space. The hybrid Tee and the associated waveguide used to produce the two driving signals are not shown in Fig. 9. The coupling slots are longitudinal and are located at a point where the circumferential wall current at 1725 mc has the first maximum. This maximizes the coupling for the TE_{11} mode and keeps excitation of the TM_{01} mode at a low level. The coupling slots are backed by two-cavity bandpass filters, giving a maximally flat transmission with a 60-mc bandwidth. The filters are built in a small-sized waveguide (4.0 by 0.9 inch) to reduce the number of undesired modes which might be excited at 4 and 6 kmc. Insertion loss at 1725 mc is less than 0.1 db. The filters also act as transformers into the conical section of the antenna and into the larger-sized waveguide WR430, which is used for the rest of the installation. Return loss of the coupler, including the hybrid Tee splitting arrangement, is better than 16 db over a 14-mc band centered at 1725 mc. The coupler has no measurable effect on the transmission of 6390- and 4170-mc signals. The noise introduced into the system by the coupler at 4 kmc is dependent on the losses in the iris located in the apex taper. By proper mechanical design of this iris, it was possible to reduce this noise contribution to less than 1°K.

A rotating joint is located at a cone diameter of 34 inches. This rather large diameter and a gap width of 0.1 inch simplify the electrical design of a joint which has to work at 6390, 4170, and 1725 mc. A double choke is used and optimized at 4170 mc. Measurements have indicated that for gap widths up to 0.5 inch, and with absorbing material covering the outside of the joint, the 4-kmc noise contribution was still below 1°K. A radiation shield of the form shown in Fig. 9 was put around the outside of the joint in order to keep the radiated energy at 6390 or 1725 mc below the Bell System safety limit for continuous exposure of 1 milliwatt per cm² at all points outside the gap.

2.4.4 *The 4-kmc Waveguide Losses and Noise*

The importance of using low-loss waveguide components in a low-noise system will now be explained. It can be shown that the noise introduced by a matched circuit with a power transmission coefficient a amounts to

$$T = T_R(1 - a) = T_R(1 - 10^{-A/10}),$$

where $A = -10 \log a$, the insertion loss in decibels, and T_R is the temperature of the circuit in degrees Kelvin. With T_R taken to be 290°K, the formula leads to the helpful approximation $T = 66.8$ A, if $A < 0.5$ db. For each tenth of a decibel loss, the noise will therefore increase by about 6.7°K. For a total loss of 1 db, the exact formula gives a temperature of 59.7°K. If one considers that the noise from all the nonwaveguide sources together amounts to about 20 to 30°K in a station like the one at Andover, it becomes clear that the waveguide losses should not exceed a very small fraction of a decibel.

It is enlightening to determine the effect of an additional 0.1 db of loss on the signal-to-noise ratio of a receiver with 26°K over-all noise temperature. We find that the added 6.7°K of noise is equivalent to an increase in the receiver noise power of one db. With the signal down only by 0.1 db the signal-to-noise ratio is degraded by 1.1 db. The effect of the loss is therefore 10 times greater on noise than on signal level. The situation becomes reversed, however, in cases where A is higher than a few db (very lossy waveguide circuits or high rain attenuation). The noise temperature then reaches 290°K asymptotically.

The waveguide components of Fig. 9 were measured with a dual-

TABLE II — LOSS AND NOISE OF THE WAVEGUIDE CIRCUIT AT 4170 MC

Waveguide Unit	Loss (db)	Noise Temperature °Kelvin
Autotrack coupler	0.010	0.67
Polarizer	0.021	1.40
Polarization coupler	0.007	0.47
Mode suppressor I	0.021	1.40
Low-pass filter	0.085	5.67
Mode suppressor II	0.016	1.07
23-db directional coupler, including coupling loss	0.030	2.00
Flexible waveguide	0.023	1.54
Total	0.213	14.22

channel insertion loss test set which gave an accuracy of a few thousandth of a db. Because noise is generated only by the absorptive part of the insertion loss, the loss due to reflections should normally be taken into account. However, the very high return losses (30–40 db) of the components made it possible to neglect the reflection losses for all practical purposes. Table II lists all the waveguide parts which contribute to the noise at 4170 mc.

No contributions from parts to the right of the autotrack coupler in Fig. 9 are shown because they are negligible. The noise temperatures shown in the table were calculated using the above formula. A more direct determination of the over-all waveguide noise temperature was made by using the maser as a low-temperature reference. The result was within 1°K of the above given figure of 14.22°K.

2.5 *System Noise Temperature*

2.5.1 *Method for Measuring Noise*

Those parts of the receiver which are important for the determination of the receiving system noise temperature are shown in Fig. 14. The noise energy coming from the sky, the radome, the antenna, and the waveguide circuits is lumped into a single term $kT_{\mathrm{IN}}B$. It is assumed here that B is the receiver noise bandwidth of about 29 mc. Noise from a noise lamp can be injected through a 23.35-db directional coupler into the signal path ahead of the maser. With the noise lamp turned off, we can write for the noise power at the output of the IF amplifier:

$$N_{\mathrm{OFF}} = kBg_M g_{\mathrm{IF}}\,(T_{\mathrm{IN}} + cT_R + T_M) + kBg_{\mathrm{IF}}\,(f - 1)T_R\,.$$

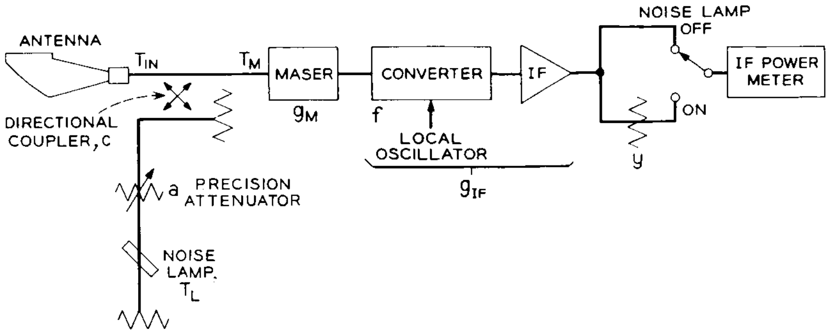

Fig. 14 — Circuit essential for the measurement of the noise temperature of the receiving system.

The g's are power gains, T_R is the room temperature, taken as 290°K, c is the coupling factor of the directional coupler $(=1/216)$, and f is the noise factor of the converter $(=15.85$ or 12 db). TM is the temperature of the maser alone, amounting to 3.5°K. Effects like imperfect impedance matches, finite directivity of the directional coupler, and differences in noise bandwidth between the maser and the IF amplifier were neglected because they are very small. If we introduce the new term

$$T_{\text{SYS}} = T_{\text{IN}} + cT_R + \frac{f-1}{g_M} T_R + T_M,$$

meaning the receiving system noise temperature measured at the input terminals of the maser, the noise power can be written as

$$N_{\text{OFF}} = kBg_M g_{\text{IF}} T_{\text{SYS}}.$$

With the noise lamp turned on, the noise power at the output of the IF amplifier, after going through an attenuator with power loss coefficient y, is found to be

$$N_{\text{ON}} = kB \frac{g_M g_{\text{IF}}}{y} [T_{\text{SYS}} + ac(T_L - T_R)]$$

where T_L is the temperature of the noise lamp, amounting to about 9500°K, and a the power transmission coefficient of the precision attenuator. The noise due to the insertion of attenuator y is neglected because its loss is small and the product $g_M g_{\text{IF}}$ is high. In the course of a noise measurement, the two noise powers N_{OFF} and N_{ON} are made equal by adjusting the precision attenuator. The setting of the precision attenuator will depend on, among other variables, the value of attenuator y, which can be changed in one-db steps. We then find for the system temperature:

$$T_{SYS} = \frac{ac}{y - 1}(T_L - T_R).$$

The maximum error of a noise measurement amounts to about ± 0.35 db or ± 8.5 per cent, with T_L contributing 0.2 db; $y - 1$, which includes short-term amplifier gain variations, 0.1 db; and the combination of a and c, 0.05 db. The improved accuracy of the noise lamp temperature over the 0.5 db of the commercial units was obtained by directly measuring the temperature of the lamp with the maser as a low-noise amplifier.

2.5.2 *Results of Noise Measurements*

Measurements of the system noise temperature made at Andover for different elevation angles of the antenna are shown in Fig. 15. The noise at zenith on a clear day is 32°K for the system containing a maser and is about 85°K with the parametric amplifier replacing the maser. The noise increases towards the lower elevation angles due to the higher noise originating in the atmosphere. When the antenna beam hits the ground, the noise increases abruptly to a value of about 235°K. This temperature depends on the actual temperature and the reflection coef-

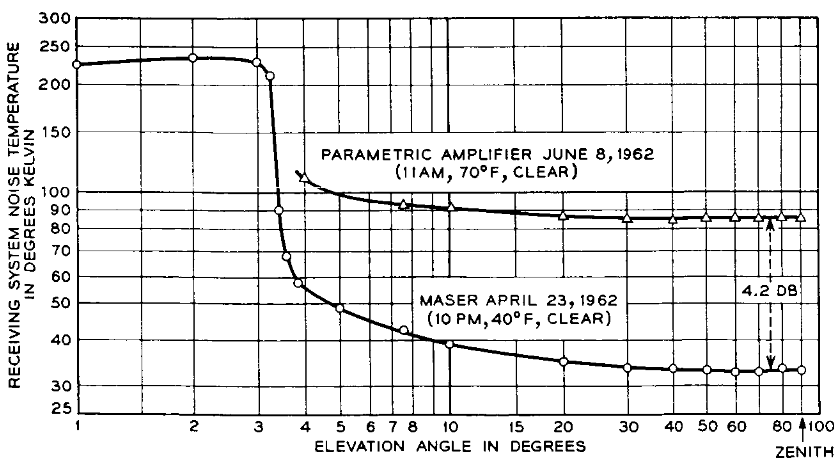

Fig. 15 — Receiving system noise temperature as a function of antenna elevation — clear day data. Azimuth = 142°.

ficient of the ground and on the losses of the signal through the radome and the waveguides. Knowing all these parameters, except the reflection coefficient of the ground, we find for the latter a value of 0.165 at 4170 mc. The sharp increase of the noise at the horizon was used to plot the radio horizon, which then was compared with the far optical horizon. The two coincided within about 0.1° for all regions where the far horizon was not hidden behind nearby obstacles.

The noise distribution shown in Fig. 15 was measured many times and was never found to vary more than one or two degrees on clear or cloudy days. During periods of rain or snow, vastly different noise temperatures were observed. Fig. 16 shows three typical curves. During a heavy rain (curve number 1), high noise values were found, with fast variations occurring in the region from 50° to 90° elevation. Variations at zenith over a period of about 15 minutes ranged from 68° to 126°K. Zenith temperatures between 70°K and 100°K are quite common during periods of moderate rain.

Curve 2 shows a case of light rain, and curve 3 was taken during a light and wet snowfall. The high noise temperatures found near the zenith (curve 3) are clearly due to the presence of the radome on which more snow has accumulated near the top than on the sides. Even curves

CURVE	DATE	LOCAL TIME	WEATHER
1	MAY 24, 1962	10:30 TO 11:30	HEAVY RAIN
2	SEPT 5, 1962	15:00 TO 15:20	LIGHT RAIN
3	OCT 26, 1962	07:10 TO 07:30	LIGHT WET SNOW

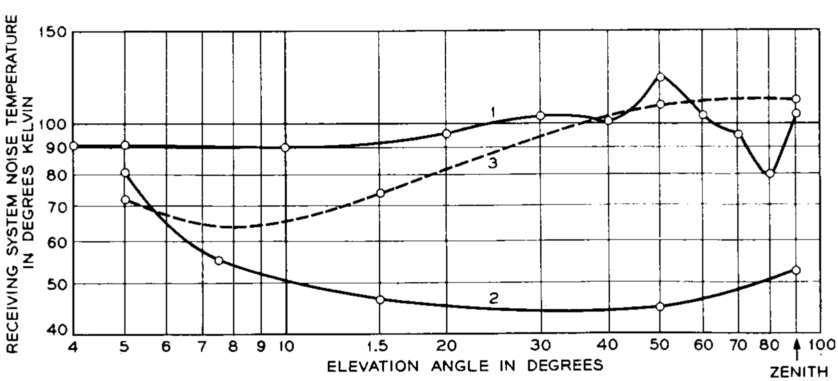

Fig. 16 — Receiving system noise temperature as a function of antenna elevation — rain and snow data. Azimuth = 142°.

1 and 2 show more noise at the higher angles than at, for instance, 15° elevation. This may again be due to some "accumulation" of rain on the top of the radome, but is also due to increased reflection of ground noise into the antenna aperture by the higher portions of the inner radome surface. The latter will become clearer in a moment when the electrical characteristics of the radome are described. The highest zenith system temperature observed to date in Andover due to rain is 135°K and 160°K due to snow. Within about half an hour after the end of a rainfall, noise temperatures close to the dry values of Fig. 15 can be found.

Similar measurements of noise during rain and snow, made at the Holmdel and Whippany, N. J., locations of Bell Laboratories, but without a radome, show a completely different character. Noise temperatures tend to be considerably lower, especially during snow, and they always seem to be lowest at zenith.

2.5.3 *Contributors to System Noise*

Table III gives a breakdown of the system noise into the different contributors. The values for the dry atmosphere were taken from measurements made at Holmdel.[17] The zenith temperature of 2.6°K of this reference appears reduced to 2.4°K by the total loss of 0.36 db in the radome and the waveguides. The contributions of antenna sidelobes, waveguides, maser and second stage are constant and amount to 19.2°K. They were discussed above with the exception of the antenna sidelobes. The horn-reflector antenna is probably the antenna with the lowest possible noise. If operated without a radome, all sidelobes and backlobes

TABLE III — CONTRIBUTORS TO SYSTEM NOISE

	Elevation			
	90°	30°	15°	7.5°
Dry atmosphere	2.4°K	4.8°K	9.2°K	18.4°K
Antenna sidelobes Waveguide circuits Maser Second stage		1.0 14.2 3.5 0.5	19.2°K	
Dry (wet) radome absorption	3 (12)	3 (12)	3 (12)	3 (12)
Dry (wet) radome scattering	7.4 (18.5)	6.0 (15.0)	4.1 (10.2)	1.4 (3.5)
Total (dry, measured)	32.0 °K	33.0 °K	35.5 °K	42.0 °K

will not contribute more than about 1°K if the antenna is pointed at least one degree above the horizon.

The effect of the radome on system noise is two-fold. First, the absorption of 4170-mc energy by the radome material is a measure of the noise it spontaneously emits at the same frequency; and second, the inner surface of the radome will reflect (scatter) ground noise into the antenna aperture. In order to investigate the first effect, it was necessary to know the loss characteristics of the radome. Laboratory measurements on dry and wet radome samples first gave values for the dielectric constant and the loss tangents. Calculations were then made to determine the power reflection $| r_E |^2$ and power transmission $| t_E |^2$ coefficients and the absorption transmission coefficient $a = | r_E |^2 + | t_E |^2$. The absorption loss in decibels is $A = -10 \log a$ and would amount to 0 db for the lossless material shown in Figs. 12 and 13. The results of the present investigation are shown in Figs. 17 and 18. If we consider again a maximum variation of angle of incidence from 8° to 52°, we find 3° and 12°K as rough averages for the absorptive noise in the dry and the wet case. It should be kept in mind that the data for the wet radome are based on tests made in the laboratory and that they cannot be completely representative of the conditions encountered in the field where the amount of wetness will vary with the rate of rain and possibly the position on the radome. The tests also showed that the dielectric constant and the loss tangent of the wet material decayed rapidly within minutes and later more slowly towards the "dry" values.

The scattering effect of the radome is difficult to determine by calculation. The noise values shown for dry radome scattering in Table III are therefore equal to the amount of noise which could not otherwise be accounted for. The higher scattering at high elevation angles is due to the particular antenna-radome geometry,* and to a certain extent is also dependent on the area surrounding the radome. The ground effects were determined by rotating the antenna 360° in azimuth and observing the change in zenith temperature. Values ranging from 31° to 35°K were found. The data presented in Figs. 15 and 16 and in Table III were taken at an azimuth of 142° (boresight tower).

The scattering effect of the wet radome was found by multiplying the dry values with the ratio of wet-to-dry power reflection coefficients as given in Figs. 17 and 18. This ratio is about 2.5 over a large range of

* Experiments with antennas whose apertures remain always vertical to the diameter of a spherical radome indicate that scatter-noise is minimum at zenith. Such a geometry is impractical for the horn-reflector antenna, however.

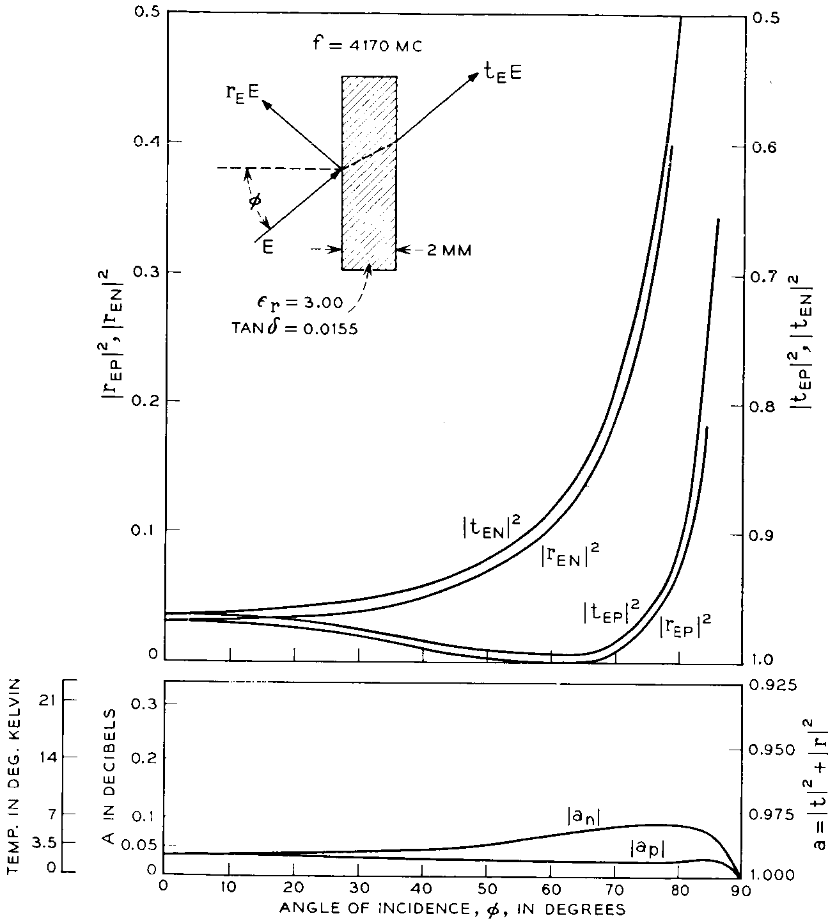

Fig. 17 — Reflection, transmission and absorption transmission coefficients of the dry radome at 4170 mc.

angles of incidence. Wet radome scattering is considerably larger at zenith than at lower angles. This explains the higher noise readings at zenith than at 15° elevation.

Recently, the diplexer of Fig. 9 was operated without the low-pass filter. It was found that one of the polarization couplers built had an isolation peak of 47 db in a narrow band around 6390 mc. This value of isolation allows unperturbed operation of the maser. The system temperature could thereby be reduced by 5°K, or to 27°K at zenith.

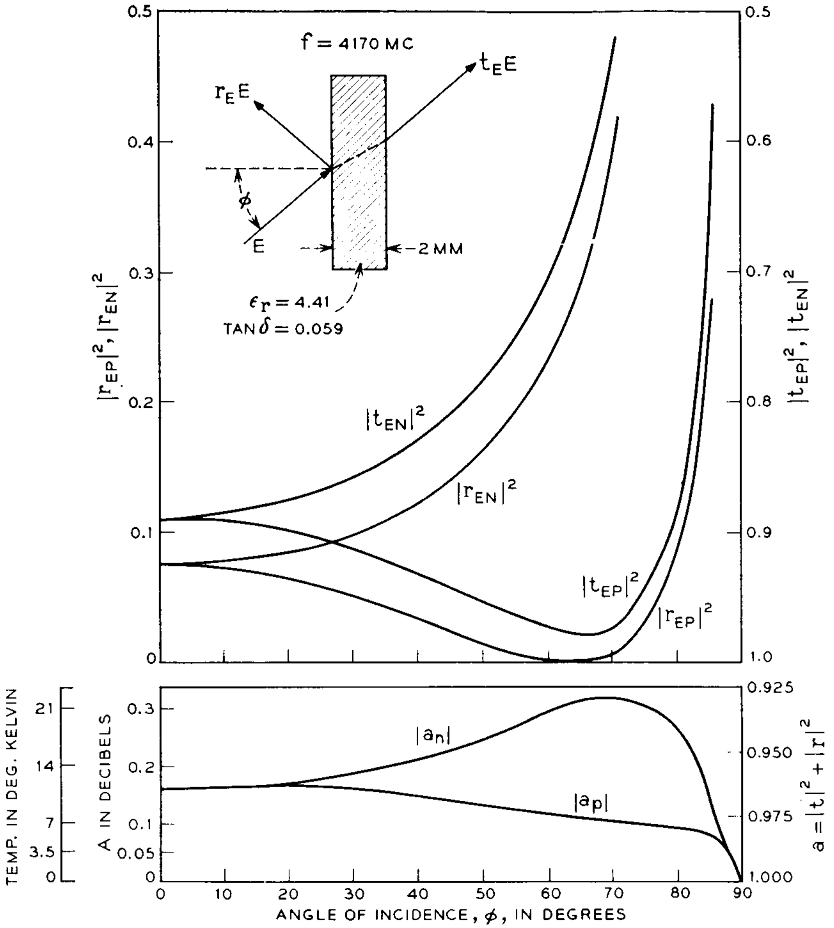

Fig. 18 — Reflection, transmission and absorption transmission coefficients of the wet radome at 4170 mc.

2.5.4 *Other Possible Sources of Noise*

A satellite communications system can be affected by extra-terrestrial noise sources, which is the case on such rare occasions when the satellite passes in front of one of these sources. The strongest source of noise at 4 kmc is the sun, followed by the moon and a number of radio stars. When the antenna is pointed at the center of the galaxy, a few extra degrees of noise will be introduced.

A measurement was made of the receiving system noise at and near the sun. The antenna was pointed in the path of the sun, which then drifted through the antenna pattern. The result is plotted in Fig. 19. Over an angle of about 0.4°, which is a little less than the 0.5° diameter of the sun, a maximum temperature of 17,000°K is maintained, representing an interfering noise signal of −81 dbm. If we compare this signal with the signal from the satellite, which is in the range of −75 to −100 dbm, it is clear that communications would be disrupted during this time. The drop in noise shown in Fig. 19 is as expected from the radiation diagram of the antenna. Only one degree away from the center of the sun, the noise has dropped to an operationally acceptable 65°K.

The actual temperature of the sun at 4170 mc is higher than the measured 17,000°K by about 1.6 db, or equal to 24,500°K. The 1.6 db is made up of 0.04 db atmospheric loss, 0.15 db radome insertion loss, 0.21 db waveguide loss, and 1.2 db loss due to the finite width of the antenna pattern.

The temperature of the moon was measured on May 10, 1962, 10:00

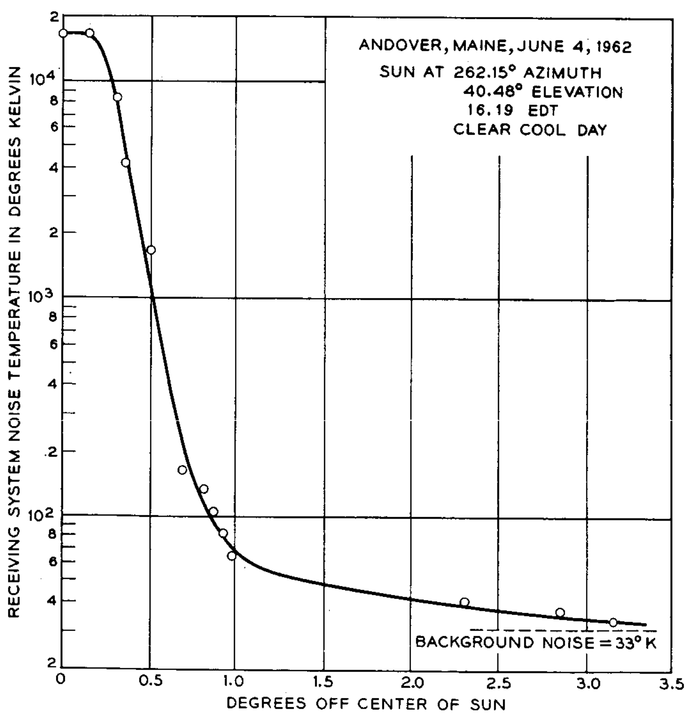

Fig. 19 — Receiving system noise temperature near the sun.

p.m. EDT. The maximum system temperature was found to be 195°K. If we subtract from this figure 34°K of system background noise and add 1.66 db due to the same type of losses mentioned before, we find for the temperature of the moon 236°K. The moon will produce a noise power in the communications channel of −100.5 dbm and therefore can affect the satellite link if the received signal level is lower than −95 dbm.

A search was also made for man-made sources of noise and interference. Such signals would be strongest close to the horizon. The measuring equipment included the receiver followed by a narrow-band IF amplifier which allowed the detection of signals down to −130 dbm. This is about 20 db lower than the normal broadband noise level of the receiver. No noise or other interfering signals could be detected. This is clearly a result of the topography of the Andover site.

III. ACKNOWLEDGMENTS

A number of people have contributed to the development of the transmitter and receiver for the Telstar ground station. The efforts of G. S. Axeling in the design of the receiver waveguide circuits are particularly acknowledged. W. Gabel, G. D. Hohmann and W. S. Irvine contributed to the power amplifier design, and J. Schill was largely responsible for the FM deviator.

REFERENCES

1. Hoth, D. F., O'Neill, E. F., and Welber, I., The *Telstar* Satellite System, B.S.T.J., this issue, p. 765.
2. Shennum, R. H., and Haury, P. T., A General Description of the *Telstar* Spacecraft, B.S.T.J., this issue, p. 801.
3. Sproul, P. T., and Griffiths, H. G., The TH Broadband Radio Transmitter and Receiver, B.S.T.J., **40,** November, 1961, pp. 1521–1568.
4. Susen, C. P., and Degan, J. J., The TH Radio Microwave Carrier Supply System, B.S.T.J., **40,** November, 1961, pp. 1569–1586.
5. Houghton, E. W., and Hatch, R. W., FM Terminal Transmitter and Receiver for the TH Radio System, B.S.T.J., **40,** November, 1961, pp. 1587–1626.
6. Collier, R. J., Laico, J. P., Helm, G. D., and Striny, K. M., The Ground-Station Traveling-Wave Tube, B.S.T.J., this issue, Part 3.
7. Houghton, E. W., and Drazy, E. J., Test Equipment for the TH Radio System, B.S.T.J., **40,** November, 1961, pp. 1717–1743.
8. Giger, A. J., and Low, F. K., The Automatic Protection Switching System of TH Radio, B.S.T.J., **40,** November, 1961, pp. 1665–1715.
9. Githens, J. A., and Peters, T. R., Digital Equipment for the Antenna Pointing System, B.S.T.J., this issue, Part 2.
10. Hatch, R. W., Bennett, S. B., and Kinzer, J. P., Results of the *Telstar* System Communication Tests, B.S.T.J., this issue, Part 2.
11. Tabor, W. J., and Sibilia, J. T., Masers for the *Telstar* Satellite Communications Experiment, B.S.T.J., this issue, Part 3.

12. Uenohara, M., Chruney, M., Eisele, K. M., Hanson, D. C., and Stillwell, A. L., A 4-gc Parametric Amplifier for the *Telstar* Satellite Communication Experiment, B.S.T.J., this issue, Part 3.
13. Giger, A. J., and Chaffee, J. G., The FM Demodulator with Negative Feedback, B.S.T.J., this issue, p. 1109.
14. Ohm, E. A., Project Echo Receiving System, B.S.T.J., **40**, July, 1961, pp. 1065–1094.
15. Cook, J. S., and Lowell, R., The Autotrack System, B.S.T.J., this issue, Part 2.
16. Tang, C., unpublished work.
17. De Grasse, R. W., Hogg, D. C., Ohm, E. A., and Scovil, H. E. D., Ultra-Low-Noise Antenna and Receiver Combination for Satellite or Space Communication, Proc. National Electronics Conference, **15**, October, 1959, pp. 370–379.