The Spacecraft Communications Repeater

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This paper describes the electrical characteristics of the waveguide components, the solid-state circuits and the traveling-wave tube used in the microwave communications repeater. The reflex-circuit method of obtaining local oscillator signals for the modulators and certain circuit stability problems are discussed.

I. INTRODUCTION

This paper describes the communications repeater portion of the Telstar satellite. This repeater receives the weak, −60 dbm nominal, FM signal from the earth at a center frequency of 6389.58 mc, shifts the frequency to 90 mc for amplification by transistors, shifts the frequency to 4169.72 mc for further amplification by the traveling-wave amplifier, and reradiates the signal at a minimum power of 33 dbm. The signals are received and transmitted through separate circularly polarized antennas which are nearly isotropic.[1] The satellite also radiates, for tracking purposes, a very stable microwave beacon signal at a frequency of 4079.73 mc and at a power of greater than 13 dbm. The bandwidth of the repeater is 50 mc, although to date only 25 mc has been used in the experiments because of bandwidth limitations of the maser in the ground receiver.

II. CIRCUIT OPERATION

The block diagram of the communications repeater, excluding power supply, is shown in Fig. 1; the numbers circled at various points in the circuit show the power levels in dbm. Fig. 1 shows exact frequencies, but for simplicity approximate frequencies will be used in the text. The signal path through the repeater is as follows.

The signal from the ground station is received by the nearly isotropic 6-gc antenna and is applied to the down converter at a nominal level of −60 dbm. A down converter shifts the center frequency to 90 mc where

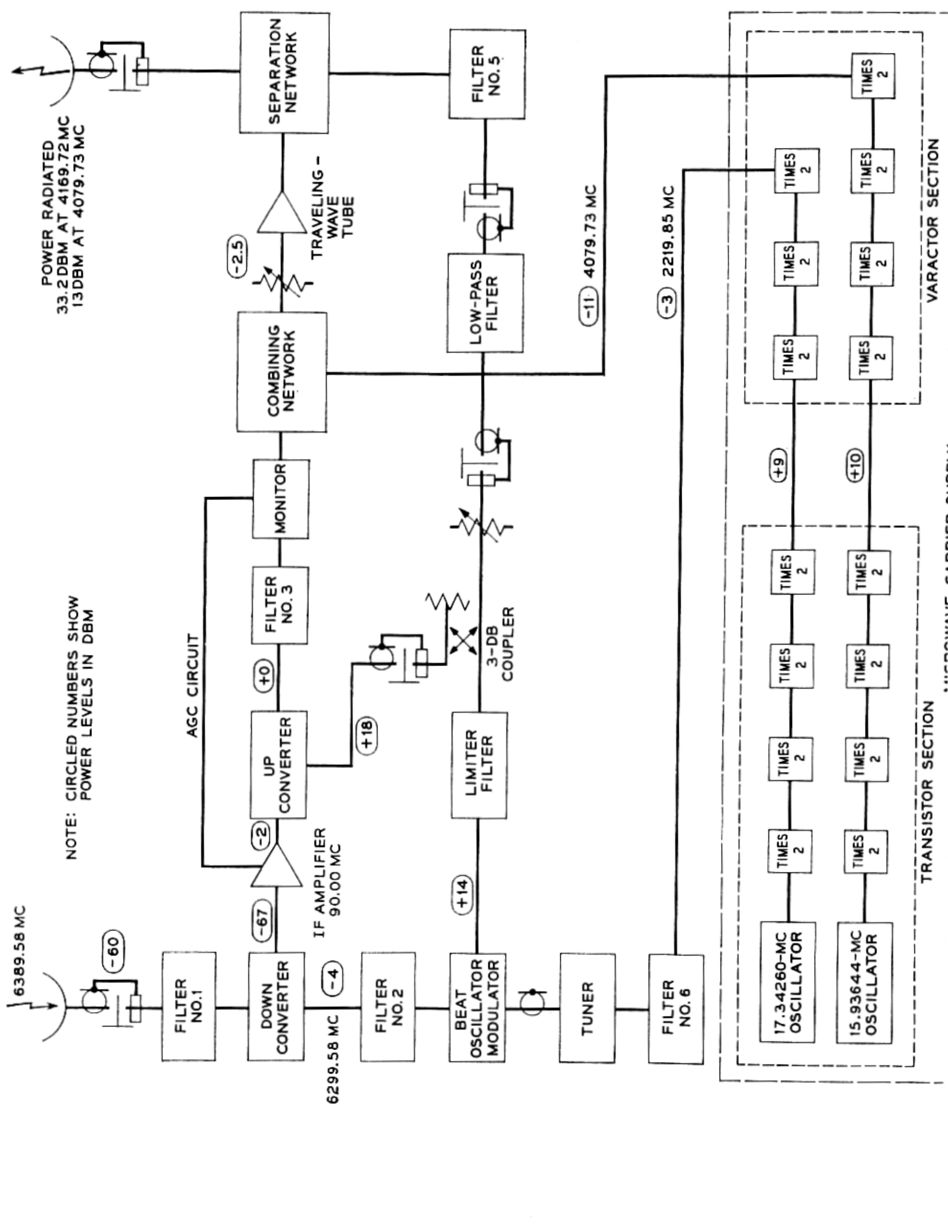

Fig. 1 — Communications repeater.

the signal undergoes gain of about 65 db in a 14-stage transistor amplifier. The signal is then applied to a balanced up converter which shifts the center frequency to 4170 mc. Varactor diodes are used in the up converter, so it provides some conversion gain. Filter 3 in the output of the up converter allows only the sum frequency (4080 + 90) mc from the up converter to pass into the monitor section. Silicon diodes in the monitor provide a dc output voltage which is a monotonic function of the power input to the monitor. This dc voltage, after amplification, controls the gain of the IF amplifier by changing the current through variolosser diodes in the IF amplifier. Because this automatic gain control (AGC) detector, the monitor, operates at 4 gc, the AGC loop includes the up converter and keeps the input power to the traveling-wave tube essentially constant as the input signal at 6390 mc varies from −55 to −72 dbm. Variations in the input signal are due to changes in satellite slant range and lack of isotropy in the receiving antenna. These variations are usually much less than 17 db when the transmitted power at 6390 mc is programmed. Ripple in the antenna pattern combined with spinning of the satellite causes the received signal to be amplitude modulated at frequencies of several hundred cycles per second. The frequency response of the AGC circuit is fast enough to smooth the amplitude modulation caused by the ripple in antenna gain.

After the signal passes through the monitor section, it is applied through a combining network to the traveling-wave tube (TWT). A card attenuator preceding the TWT is used so that small adjustments in tube drive can be made after the AGC adjustments have been completed. At the operating point selected for the tube, the gain for the 4170-mc signal is 37.5 db and the resulting power output is 35 dbm. The 4170-mc output of the TWT is applied to the transmitting antenna through a separation network which has an insertion loss of only 0.2 db at this frequency. However, the insertion loss of the transmitting antenna and connecting cables is 1.6 db, so the radiated power is 33.2 dbm.

The 4080-mc pump for the up converter is derived from a crystal oscillator at approximately 15.9 mc, followed by transistor and varactor frequency doublers. In order to obtain sufficient power for this up-converter pump, a special reflex circuit[2] using the TWT amplifier with a combining and a separation network is used, as shown in Fig. 1. In addition to furnishing the pump for the up converter, the 4080-mc amplified output of the TWT provides the microwave beacon needed for precision tracking and provides the pump for the beat oscillator (BO) modulator. The BO modulator furnishes the 6300-mc local oscillator for the down converter by combining the pump signal at 4080 mc with a signal at

2220 mc. The latter is also derived from a crystal oscillator and a series of frequency doublers. The yttrium-iron-garnet (YIG) limiter is used to ensure stability of the BO modulator under conditions of increased power at 4080 mc, conditions which exist when no 6-gc signal is transmitted to the satellite.

The gain of the TWT for the beacon signal is less than that for the communications signal, because the latter signal drives the tube into partial saturation: see Fig. 2. This graph shows the relationship of the power outputs at 4080 and 4170 mc when the 4080-mc input is constant and the 4170-mc input is varied. Three sets of curves also show the sensitivity of the tube to changes in the main regulator supply voltage; -16 volts is normal.* A circuit in the AGC amplifier changes the drive at 4170 mc to reduce the output variations if the supply voltage changes. Because two signals are amplified by the TWT, it is necessary to operate the tube about 1.5 db below saturation to reduce intermodulation to an acceptable level.

The reflex circuit is an electrically efficient way of obtaining sufficient power levels needed at 4080 and 6300 mc, but it introduces additional closed paths in the circuit, with the resulting possibility of instability. These closed paths are hereafter called feedback paths or loops, although the names are misleading because they are not used to improve circuit linearity or to realize the usual advantages of negative feedback. In this circuit, the feedback paths are undesirable by-products of the reflex method of obtaining the local oscillator signals for the converters.

The AGC circuit controls the gain of the IF amplifier so that the input signal power to the satellite is amplified by the proper amount to hold the power into the TWT constant. An increase in signal level of 3 db means a decrease in IF gain of 3 db with a corresponding 3-db decrease in noise drive to the TWT. The noise level to the TWT is low enough that noise amplification is linear, and that output noise increases almost linearly with a decrease in input signal so long as the signal falls within the AGC range.

When the input signal is removed, the gain of the IF amplifier is maximum and the noise drive to the TWT is enough to give a radiated noise power of 31 dbm in a 50-mc bandwidth. This noise will not cause trouble in any communications system unless the system uses the same frequency and has a very high-gain antenna pointed almost at the satellite. Sup-

* The variations in power output predicted from Fig. 2 are very pessimistic, because these data are shown for variations of ± 3 per cent in supply voltage. The predicted variation at the end of 2 years is less than ± 1 per cent, unless radiation damage is extensive.

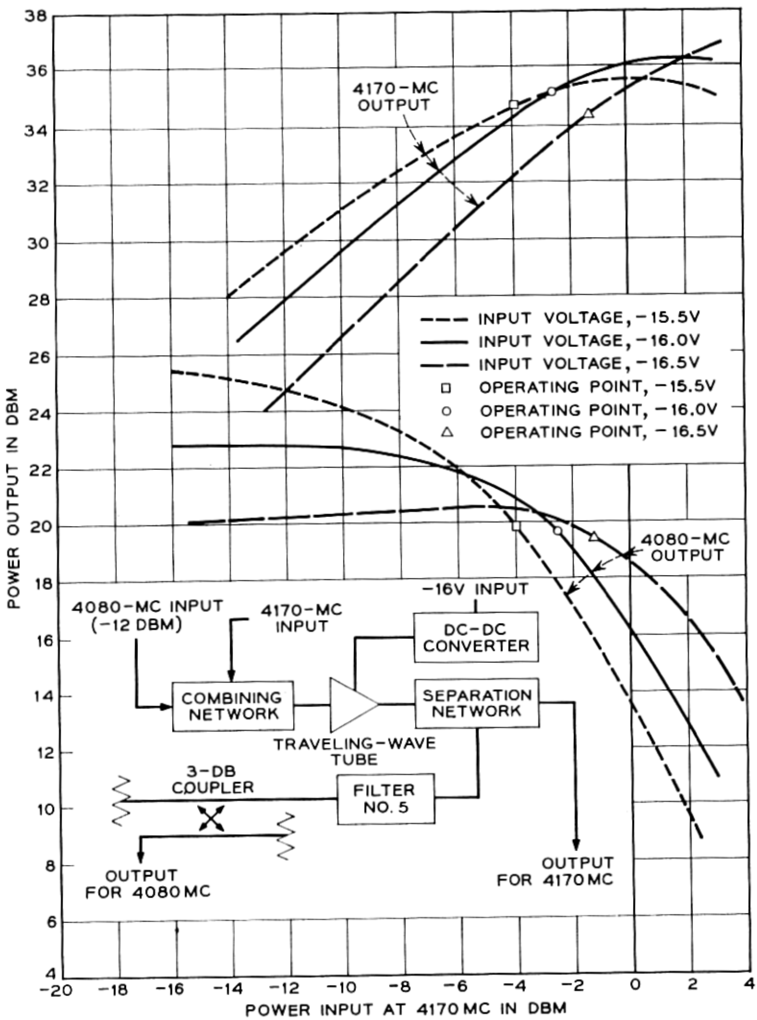

Fig. 2 — Operating characteristics for the traveling-wave amplifier.

pose, for example, there were two Telstar satellites in view, one turned on and in use, and the other turned on but having no input signal. The satellite not in use can increase the noise temperature of the Andover ground receiver by a maximum of only 2.2°K while the horn is pointing more than 3° away from the "noisy" satellite, and this would occur only when the slant range to the satellite is at its minimum possible distance of

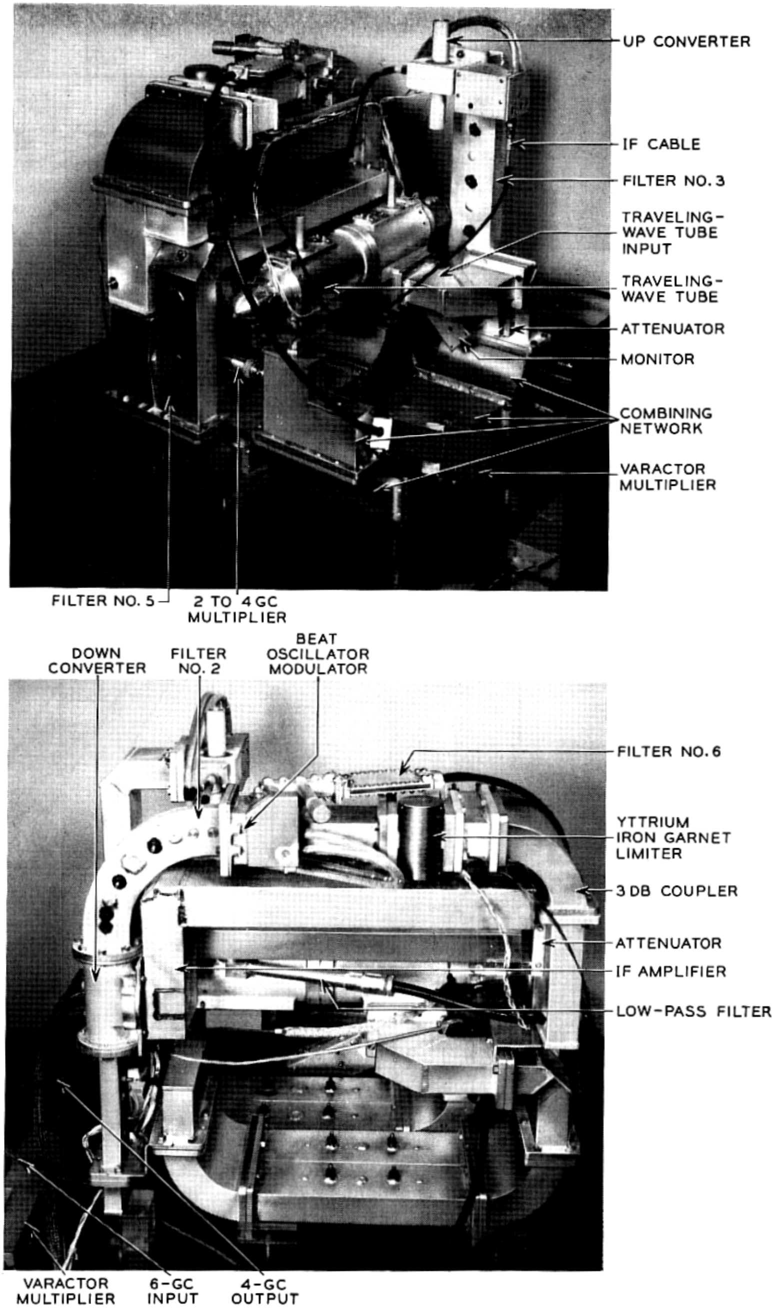

Fig. 3 — (a, b) Pictorial views of the repeater.

600 miles.[3] For a "normal" range of 2500 miles, the noise contribution from the "noisy" satellite when it is more than 3° from the beam of the ground antenna is less than 0.12°K.

When the spacecraft has no input signal, the TWT is not driven as hard as when a normal signal is used, so the microwave beacon power increases 2 db over its normal level. This extra power is advantageous, since the no-signal-input condition usually occurs when the ground tracking antennas are trying to acquire the satellite.

III. SUBASSEMBLY DESCRIPTIONS

In this section the various subassemblies of the repeater are described. The characteristics and unique features of the subassemblies are covered, but the descriptions are not intended to be design sections. Two views of the communications repeater, Figs. 3(a) and 3(b), show all the important subassemblies. All waveguide parts are made of magnesium; the 4-gc parts use WR187 guide, and the 6-gc parts use reduced-height WR137 guide. All subassemblies were tested before and after they were subjected to vibrational forces in excess of these expected in launch. The repeater must operate over a temperature range of 0° to 50°C, so all units were tested over at least this range.

3.1 *Down Converter*

The down converter shifts the 6390-mc broadband signal to an intermediate frequency centered at 90 mc. Fig. 4 shows the internal configuration of the down converter. The received signal passes through a waveguide filter into a circular cylindrical cavity in which two diodes are mounted. Inside the cavity, the signal power divides equally between the two diodes. The 6300-mc local oscillator signal is coupled into the opposite end of the cavity from a second waveguide filter; the principal axes of the two waveguides are at right angles to ensure isolation between the two frequencies present in the cavity. Input matching is provided at the signal input by the three screws which penetrate into the cavity as shown in the illustration. Two screws mounted in the waveguide close to the cavity are used to tune the local oscillator input.

The IF connection to each diode is decoupled to microwave signals by means of an RF choke which consists of two quarter-wavelength radial transmission lines. The resonant frequencies of the two lines are staggered to obtain high insertion loss over a wide bandwidth. The position of each choke along the IF output line is so chosen that an RF short circuit appears at the waveguide wall. The two IF output lines are parallel connected through coupling capacitors to the input of the IF ampli-

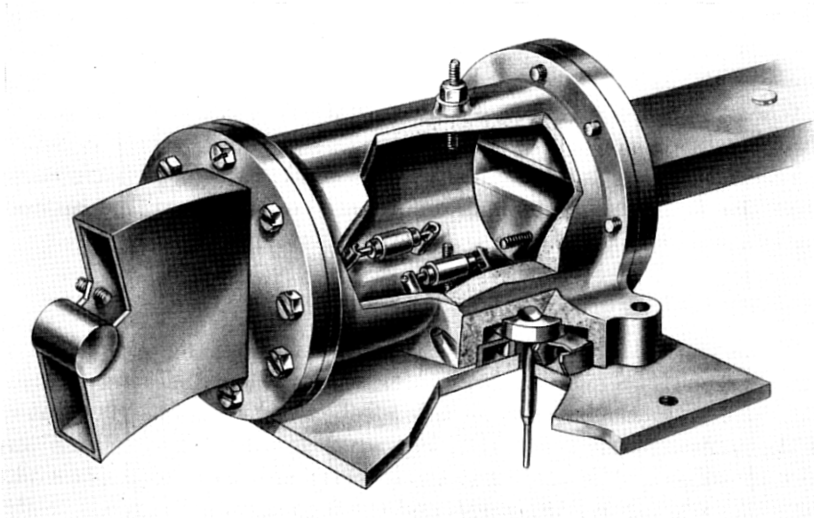

Fig. 4 — Internal structure of the down converter.

fier which is mounted directly onto the down converter. However, separate dc connections are made to the two IF output lines so that each diode may be forward biased for optimum performance.

It was experimentally determined that a good noise figure could be obtained most readily when the down converter was tuned to have a relatively narrow-band input match. The down converter is narrow banded by the presence of the resonant tuning screw placed between the two diodes, which reduces the bandwidth to about 120 mc. Narrow banding the down converter lessens the importance of the combination-frequency products generated by the modulator. With the down converter working into an IF amplifier with a noise figure of 4.5 db, the receiver noise figure is 12.5 db. Some improvement in noise figure would be obtained by increasing the level of the local oscillator signal; however, this would decrease the long-term stability and cause the properties of the down converter to be noticeably dependent on the exact level of the local oscillator signal. In the final design, the down-converter conversion loss is less than 7 db.

3.2 *IF Amplifier and AGC Circuit*

The IF amplifier provides the bulk of the amplification in the repeater, helps determine the repeater bandwidth,* and in conjunction with the

* One end of the receiver passband is determined by a high-pass filter at the input to the up converter.

microwave level monitor and the direct-coupled AGC amplifier provides automatic gain control.

Fig. 5 shows the levels encountered throughout the IF circuit. Note that the gain of the IF amplifier is adjusted by two variolossers whose losses are controlled by the output current of the AGC amplifier. The monitor for the AGC system is placed at the output of the up converter; thus the AGC system controls the input level to the traveling-wave tube.

3.2.1 *IF Amplifier*

Fig. 6 shows diagrammatically the basic design of the, IF amplifier, which uses diffused-base germanium pnp transistors throughout. Three different transistor IF amplifier configurations are employed.

The low-noise input stage consists of a common-emitter — common-emitter "doublet."[4] This configuration utilizes a large portion of the available power gain of the input transistor and thereby minimizes the effect on the over-all noise figure of the noise generated by the other transistors in the IF amplifier. A simple equalizer network is used at the output of the second transistor to compensate for the rolloff in gain of the doublet in the IF band. As one can see from Fig. 3(b), the input of the IF amplifier is physically located next to the down converter. A transformer with a 1:2 turns ratio is used between the down converter and the doublet to provide the proper mismatching for optimum noise figure. Averaged over the 65- to 115-mc band, the IF amplifier noise figure is 4.5 db, and the input impedance is about 30 ohms. The power gain of the doublet plus equalizer network is about 13 db.

To achieve the desired IF gain with a minimum number of stages while still realizing satisfactory aging and temperature performance characteristics, the common-emitter circuit with frequency-sensitive shunt feedback[4,5,6] is used for the large majority of IF amplifier stages. Frequency shaping is achieved by means of an RL network connected between base and collector. Consider the third transistor stage of Fig. 6. Resistor R_1 determines the low-frequency gain, and inductor L effectively removes the feedback at high frequencies and thereby plays the role of a broadbanding element. Resistor R_2 damps the resonance which occurs between the inductor L and the capacitive reactance presented by the transistor. Through adjustment of R_1, R_2 and L, gain and bandwidth can be exchanged; and the gain-bandwidth product is given approximately by f_T, the frequency at which the common-emitter short-circuit current gain is unity. The gain for each common-emitter shunt-feedback stage is approximately 8 db.

The principal requirement which dictates the selection of an IF out-

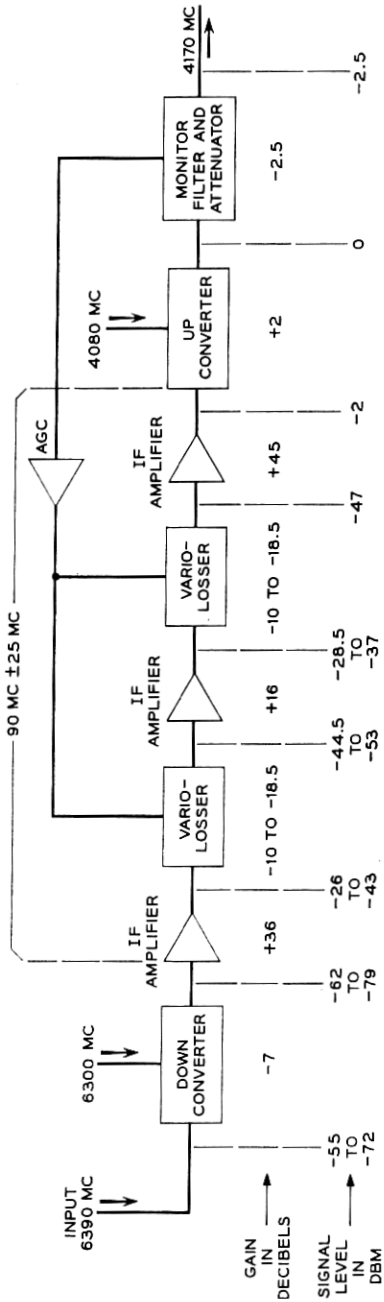

Fig. 5 — Level diagram of the IF amplifier.

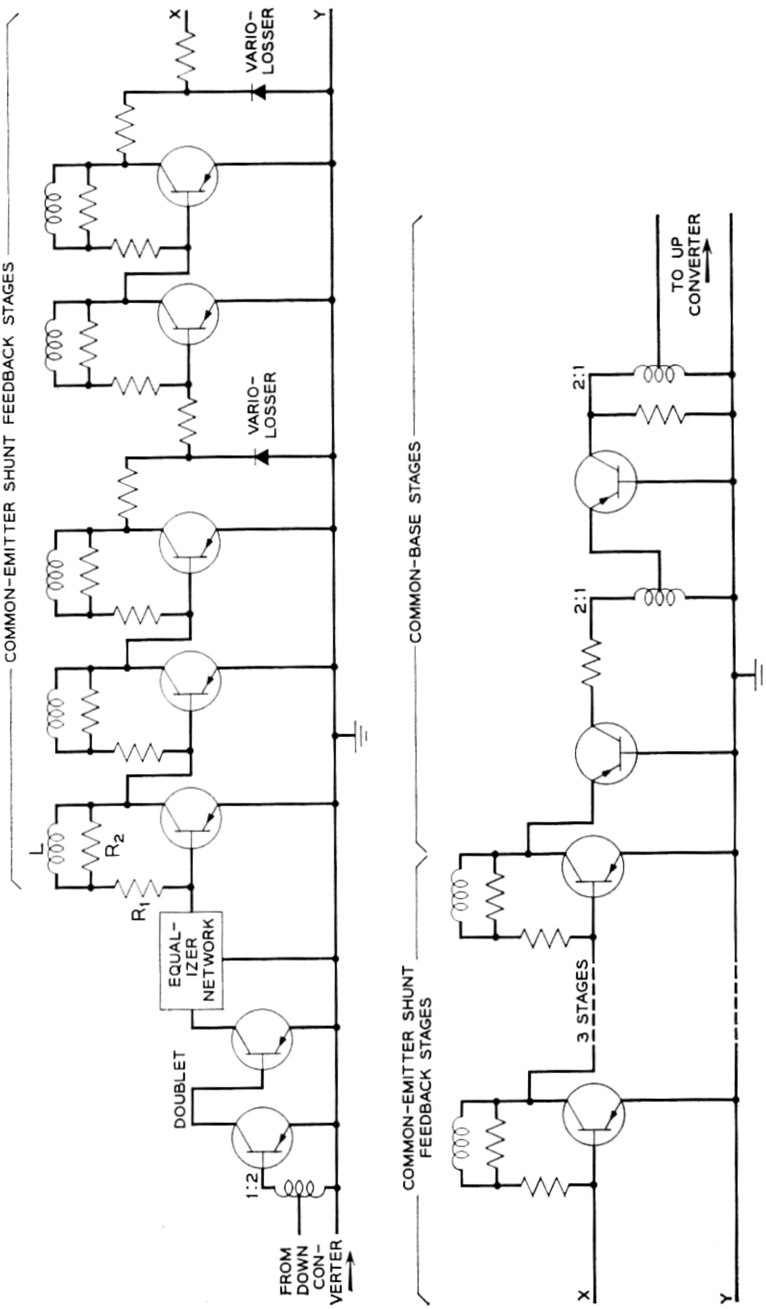

Fig. 6 — Basic circuits in the IF amplifier.

put stage configuration is that the desired undistorted output power should be achieved with minimum expenditure of dc power. For this reason the common-base configuration, with its inherently linear transfer characteristic, is used for the output stage and its driver. Two transformers, each with a 2:1 turns ratio, are used for current step-up between the common-base stages and at the output of the amplifier. Through the use of these transformers, each common-base stage provides about 5.5 db of power gain. The up converter, which is driven through a length of 75-ohm coaxial cable by the IF amplifier, must be driven by a generator with resistive impedance. An output return loss in excess of 20 db is achieved over the IF band. The output stage is capable of delivering up to +6 dbm into a 75-ohm resistive load.

The transmission characteristic of the complete IF amplifier with the variolossers set in the minimum loss condition is shown in Fig. 7. Note that over the 0° to 60°C temperature range the midband gain (nominally 87 db) changes 10 db and the tilt over the band varies from +1.5 db to −2.0 db. Delay slope over the band is less than 7 ns and AM-to-PM conversion is less than 0.4 degree per db. The circuit operates from a negative 16-volt supply and has a current drain of 90 milliamperes.

Conventional high-frequency wiring techniques are applied, and because of the inherent stability and low gain per stage of the configurations used, no interstage shielding is necessary.

3.2.2 *Variolossers*

The variolossers consist of a T-network containing two series resistors and a shunt low-capacitance germanium alloy diode used as a variable

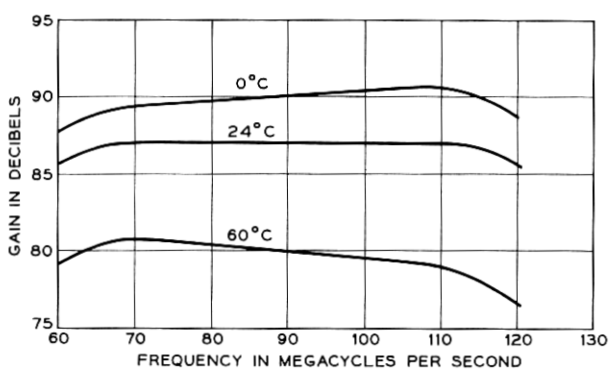

Fig. 7 — Transmission characteristics of the IF amplifier.

element. The resistors are large enough that they essentially mask out the reactive input and output impedances of the adjacent common-emitter shunt-feedback stages. They are small enough, however, to assure a low minimum loss for the variolossers. The loss of the variolossers is controlled by varying the direct current through the shunt diodes.

The placement of the variolossers in the IF amplifier is a compromise. If located too near the input, they cause too severe a degradation of noise figure during stronger input signal conditions. Placement too near the output of the amplifier will result in distortion due to rectification of the IF signal by the variolosser diodes.

The loss range of each variolosser is 15 db, and the minimum loss (relative to directly cascaded common-emitter shunt-feedback stages) is about 5 db. For the full 15-db loss variation at any temperature, the frequency distortion is less than 0.5 db; over the 0° to 60°C temperature range, the frequency distortion is negligible compared with that caused by other parts of the IF amplifier. The control current of 0 to 8 milliamperes flows through the two variolosser diodes. This current, which is a monotonic function of variolosser loss, is telemetered back to earth as an indication of the received signal strength at the satellite.

3.2.3 *AGC Amplifier and Loop Performance*

The AGC amplifier is a direct-coupled amplifier which provides control current for the variolosser diodes proportional to the difference between the dc output from the waveguide monitor and a dc reference voltage. The amplifier, which uses diffused-base silicon npn transistors, is shown diagrammatically in Fig. 8. In addition to providing high gain

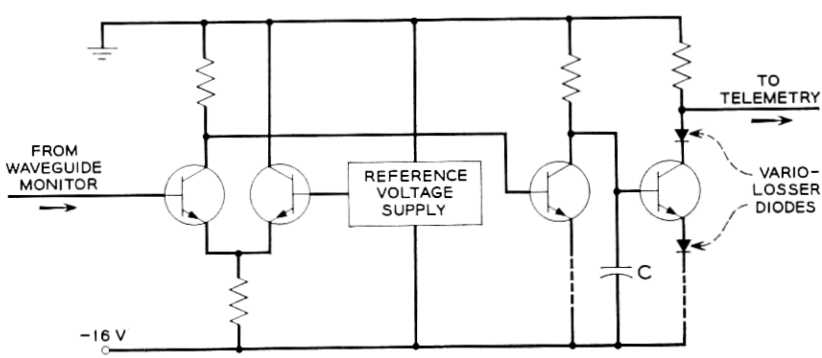

Fig. 8 — The AGC amplifier.

with low drift, the amplifier also has an adjustable feature which programs the reference voltage in accordance with variations in power supply voltage. Since the TWT uses the same power supply, the net effect of this feature is a compensation of the power supply dependence of the TWT overload characteristic through variation of TWT input level; see Fig. 2.

The differential input stage is followed by a common-emitter stage. One variolosser diode is driven from the emitter of the output stage and the other from the collector of that same stage; thus approximately equal currents flow through each variolosser diode. This diode-driving technique and the use of high-frequency transistors in the AGC amplifier (f_T = 250 mc) results in the frequency response of the AGC loop being controlled almost exclusively by capacitor C.

As has been mentioned previously, the AGC loop includes not only the IF amplifier, but also the up converter. Thus, the input level to the TWT is held relatively constant for variation in received signal strength, down converter loss, IF amplifier gain, and up converter gain. The dependence of TWT input level on received signal level is shown in Fig. 9. For received signal levels ranging between −55 and −72 dbm over a temperature range of 0 to 60°C, the input level to the TWT tube varies less than ±0.8 db.

Since the satellite is spinning and because the radiation pattern of the receiving antenna has ripples, the dynamic regulation of the AGC system must be effective for frequencies as high as 500 cps. The dynamic regulation of the AGC system is shown in Fig. 10 for several received signal levels.

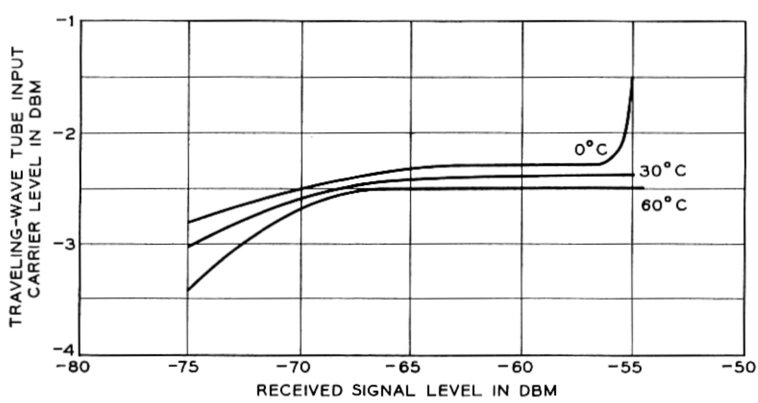

Fig. 9 — Tightness of the AGC system.

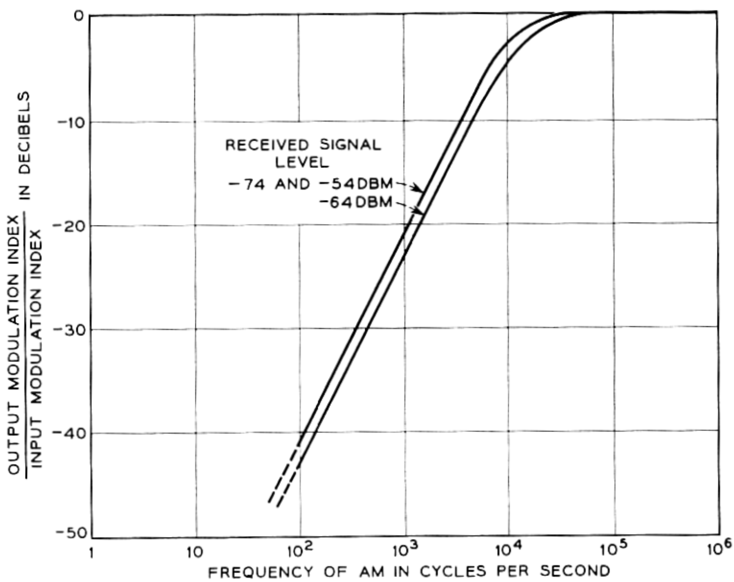

Fig. 10 — Dynamic regulation of the AGC system.

3.3 *Up Converter*

The output of the IF amplifier is connected through a short coaxial lead to the up-converter assembly, where the signal first passes through a high-pass filter. The over-all transmission characteristic of the filter is flat over the frequency band 65 mc to 115 mc, but has a minimum of 40-db rejection at frequencies below 50 mc. The up converter is a balanced diode modulator which was designed for good temperature stability and for low susceptibility to rapid fluctuations in the pump power level. To meet the special requirements imposed on the design of the converter, the diodes are operated in forward conduction and a conversion gain of about 2 db is obtained.

The up converter is designed around a hybrid junction; the output arm is waveguide, while the other three arms are coaxial, as shown in Fig. 11. Two of the arms containing the diodes also house matching transformers and second-harmonic rejection filters. The IF output leads are fed through hollow center conductors within the crystal arms and then through the center of a shorted coaxial stub. These leads are parallel coupled through capacitors to the input network and high-pass filter circuit mounted in the shielding can on the side of the up converter. DC bias voltages for the diodes are developed across self-bias resistors;

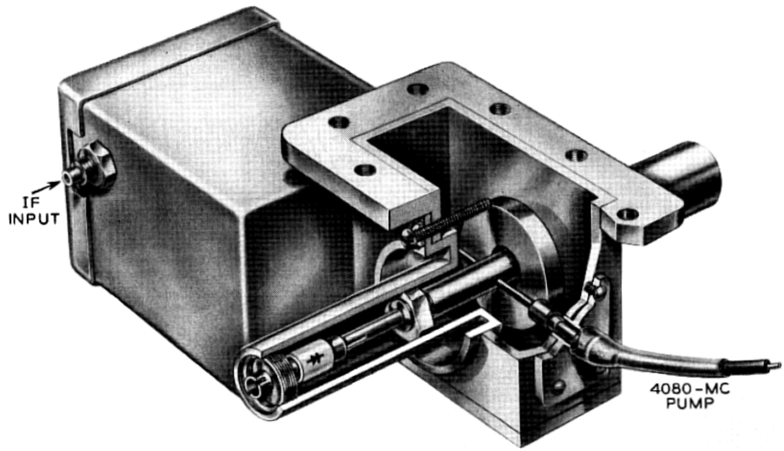

Fig. 11 — Internal structure of the up converter.

these voltages serve the secondary function of indicating the level of the pump signal for telemetry purposes. The waveguide filter on the output of the up converter is critically positioned to reflect the image frequency in correct phase to reinforce the modulator signal output power. Since the circuitry is adjusted to have a flat transmission characteristic, the return loss of the IF input to the up converter varies from 7 to 18 db over the 65- to 115-mc band.

The return loss at the microwave ports is in excess of 20 db. An important characteristic of the up converter is the relationship between the pump input level and the signal output level shown in Fig. 12. This relationship is vital to the AM stability of the circuit, a problem that will be discussed later in this article. Equally important is the absence of the pump signal at the output. Means are provided to adjust the balance of the converter to obtain a minimum loss of 30 db between the pump input and the signal output arm at normal operating levels.

3.4 *Waveguide Monitor*

The 4170-mc signal from the up converter passes through a filter and a monitor which consists of two similar diode mounts inserted one-quarter wavelength apart in a waveguide. Each diode, as shown in Fig. 13, is supported between two chucks, one on the end of the center conductor, the other in the center of a four-arm spider mounted in a die-

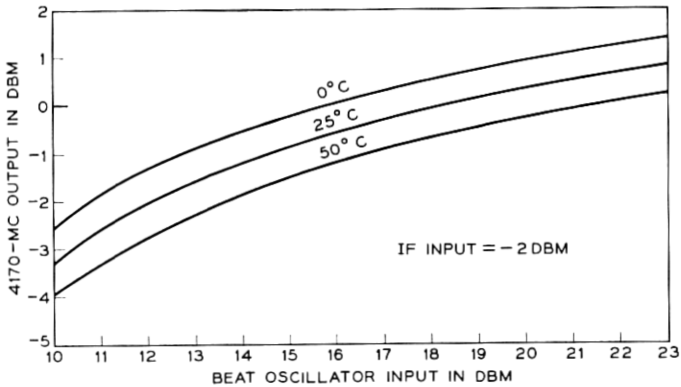

Fig. 12 — Operating characteristics of the up converter.

lectric support within the waveguide. The dc voltage developed across the diode appears between the center conductor and the outside conductor of the crystal mount, both parts being insulated from the waveguide itself. On the center conductor, a bucket-type choke determines the RF susceptance in series with the crystal.

The diode mounts are tested separately. To accommodate variations of the individual crystals, the position of the center conductor choke is varied until the impedance measured at the waveguide input and referred to the plane of the diode is found to be purely resistive. After the two diode mounts have been sealed and have passed their acceptance tests, they are mounted in the waveguide and load resistors and video-frequency decoupling components are added before foaming and final testing. No tuning or adjustment of the entire assembly is necessary.

The output voltage used for AGC purposes is derived by connecting the two diodes in series. Separation of the mounts by a quarter-wavelength along the waveguide ensures a design that is not too sensitive to the nature of the output termination. This eliminates the need for a directional coupler in this part of the circuit, thus simplifying the electronics package and reducing its size. The input-output characteristic of the monitor is shown in Fig. 14, where it will be seen that changes with temperature are relatively small. The monitor has a 20-db return loss into the waveguide input; its insertion loss is 1.5 db.

3.5 *Combining Network*

The function of the combining network is to provide a common connection to the TWT for both the communication and microwave beacon

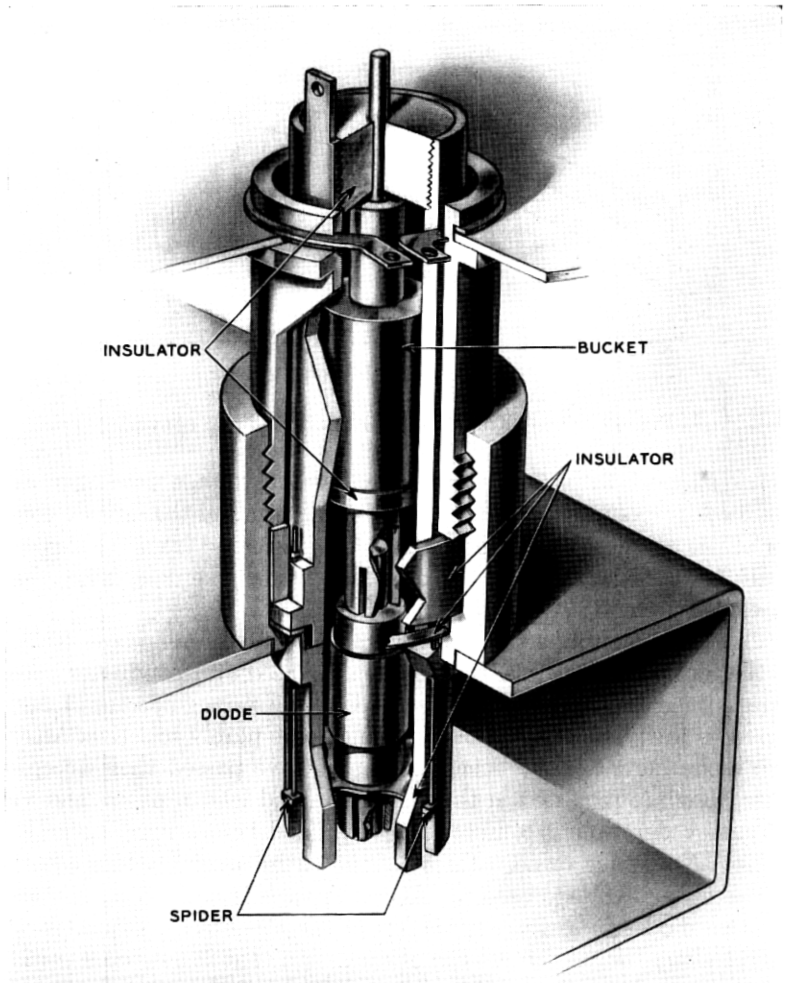

Fig. 13 — Internal structure of the monitor.

signals and at the same time provide isolation between these two inputs. This network consists of two 3-db couplers of the Riblet type,[7] two identical bandpass waveguide filters tuned to 4080 mc, and a matched termination; see Fig. 15.

The 4080-mc beacon signal input to filter B from coupler B lags the beacon signal input to filter A by 90° because of the nature of the coupler. The relative phase of the two signals is unaltered as the signals go

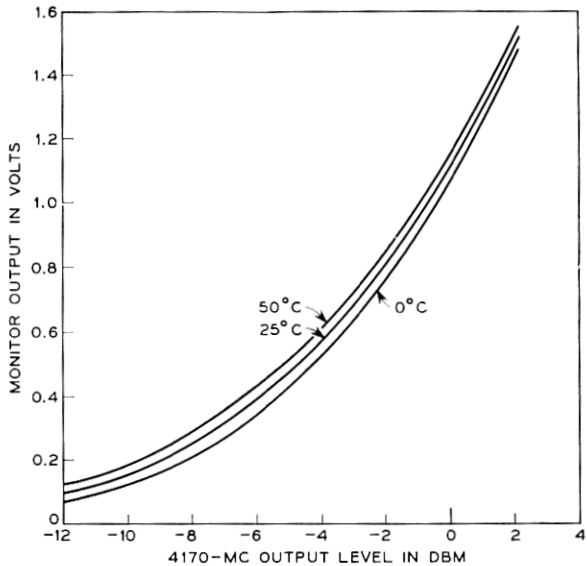

Fig. 14 — Voltage-power relations in the monitor.

through the identical two-section filters. At the output port, the signal from filter A is delayed 90° by coupler A, so at this point it is in phase with the signal from filter B. At the communications input port, the two equal-magnitude 4080-mc signals are 180° out-of-phase because the signal from filter B undergoes a 90° lag in going through coupler A, and it already lagged the output of filter A by 90°. Because of the cancellation of two 4080-mc signals at the 4170-mc input port, the only loss in the microwave beacon signal in this unit is due to the small ohmic losses.

The input signal at 4170 mc enters coupler A, and the nature of the coupler makes the input signal to filter B lag the input signal to filter

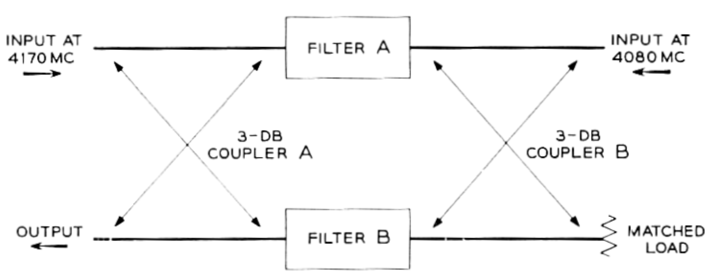

Fig. 15 — The combining network.

A by 90°. The relative phase of these signals is unchanged, as they are almost completely reflected from the filters which are tuned to 4080 mc. Thus, the phase relationship of the signals entering coupler A from the filter side is such that the two signals at the 4170-mc input port are 180° out of phase and the two signals at the output port are in phase. The isolation between the two input ports is greater than 30 db; the insertion loss at 4170 mc is 0.2 db; and the insertion loss at 4080 mc is 0.3 db.

3.6 *Separation Network*

The separation network, shown in Fig. 16, is similar to the combining network in that it comprises two 3-db couplers, two identical waveguide filters, and a matched termination. Because its filters are tuned to 4170 mc, signals near this frequency go through the separation network in the same manner that the signal at 4080 mc goes through the combining network. The filters are relatively broadband, so twenty per cent of the signal at 4080 mc leaks through the filters to the transmitting antenna, and eighty per cent is reflected from the filters. The behavior of the separation network to the reflected 4080-mc signal is similar to the behavior of the combining network to the 4170-mc signal. The 4080-mc signal level at the output port connected to the antenna is 7 db below the input signal. The insertion loss of the network in the band 4145 to 4195 mc is less than 0.2 db.

3.7 *Traveling-Wave Amplifier*

The type M4041 TWT is the only vacuum tube in the satellite. Since this tube is described in a separate paper,[8] only the circuit aspects are discussed here. The purpose of the tube is to provide power amplification for both the communications and microwave beacon signals. The operating characteristics of the tube, including the combining net-

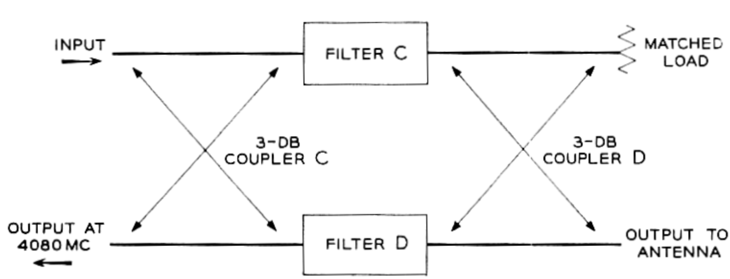

Fig. 16 — The separation network.

work, the separation network, and filter 5, are shown in Fig. 2. These filters are included in the circuit associated with Fig. 2 so the impedances seen by the tube will closely approximate those seen under actual operating conditions. The operating point was chosen well below the saturation power level of 36.5 dbm for four reasons.

First, operation in the near-linear region of the tube reduces the effect of amplitude changes in the communications signal causing amplitude changes in the microwave beacon signal. The circuit instability associated with this condition is discussed later in this paper.

Second, near-linear operation reduces the phase jitter of the microwave beacon signal. Stability of the beacon is important because both the precision tracker and the vernier autotrack systems phase lock on the signal. Even when the output load on the tube and the impedance of the source feeding the tube are perfectly matched, the power output is frequency sensitive. This means that when the communications signal is frequency modulated, the microwave beacon is amplitude and phase modulated. When the load impedance (the antenna) causes a reflection which is frequency sensitive, the above effect is made worse. Both types of modulation are decreased by operating below saturation. From the standpoint of the phase-locked tracking systems on the ground, the amplitude modulation is much less serious than the phase modulation.

The third reason for operating the TWT below saturation is the frequency characteristics of the up and down converters, the IF amplifier, and the monitor. None of these units has a perfectly flat frequency response, so an FM communications signal causes the instantaneous power drive to the tube to vary. This causes phase modulation (through AM to PM conversion) of the beacon and communications signals.

Finally, near-linear operation is desirable to reduce intermodulation of signals when the satellite is used for two-way communications experiments. This is true regardless of any effects on the microwave beacon.

The M4041 tube has very low AM-to-PM conversion when operated below saturation, so the first of the four reasons determines the operating point. With the operating point shown in Fig. 2, the net phase jitter or deviation of the microwave beacon as measured with the precision tracker[9] was less than the $\pm 5°$ "noise level" in the measuring equipment.

3.8 *Waveguide Limiter*

Because the level of the 4080-mc signal from the TWT varies with the level of 4170-mc signal, a waveguide limiter is used to stabilize the level of the 4080-mc signal at the input to the BO modulator. The limiter comprises a half-height, half-wavelength waveguide cavity with a carefully oriented yttrium-iron-garnet (YIG)[10] single-crystal sphere supported in a dielectric material close to the side wall. The cavity is formed from two inductive irises; the reduced waveguide height improves the ratio of the volume of the YIG sphere to that of the cavity and appreciably reduces the size and weight of the permanent magnet required to bias the YIG sphere.

The sphere is highly polished to reduce the insertion loss to less than 0.75 db for operation below the limiting threshold. The power limiter is of the subsidiary resonance type. With the correct external magnetic field, above a critical power level a subsidiary resonance appears in the YIG crystal which is caused by the generation of spin waves at one-half the applied microwave frequency. With increasing power input, the power out of the device remains essentially constant because the excess RF energy goes into the generation of the spin waves or is reflected from the limiter.

The magnet is designed to produce a field of about 1200 gauss across a 0.4-inch gap. It consists of two truncated cones of magnetic material and two soft-iron pole pieces. The outer case, which is used as a magnetic return path, is designed to keep the external magnetic field to a minimum.

3.9 *BO Modulator*

The 6300-mc signal used as a receiver local oscillator is generated in the BO modulator shown in Fig. 17. It is a balanced upper-sideband up

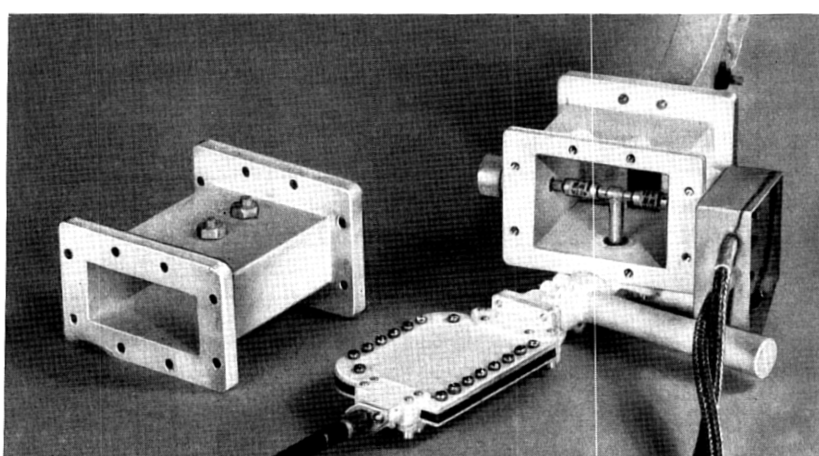

Fig. 17 — The BO modulator.

converter using gold-bonded varactor diodes which are mounted in line across a waveguide cavity. The 2220-mc input is connected to the center of the two diodes. Thus the diodes together with the center conductor of the 2220-mc coaxial input form a "T" inside and across the waveguide cavity. A two-section strip-line filter in the coaxial input provides decoupling to the higher frequencies present; the balanced structure is chosen to avoid generation of the 6300-mc signal across the 4080-mc input port, and all other necessary decoupling is provided by the natural cutoff frequencies of the waveguides used. No attempt was made to broadband the modulator since this was not necessary; the simplest possible matching device was provided in each transmission line external to the cavity. Adjustable tuning screws were used in the waveguide connections, while coaxial stub tuners were used at the 2220-mc input.

To ensure maximum stability, the modulator was operated somewhat conservatively, and the efficiency of the device was low. Due to filter losses there is a loss between the lower frequency input and the 6300-mc output of approximately 1.0 db.

3.10 *The Microwave Carrier Supply*

The local oscillator signals for the up and down converters are obtained by frequency multiplication following crystal-controlled oscillators, as shown in Fig. 18. These oscillators and the accompanying multipliers constitute the microwave carrier supply. The exact crystal frequencies are 17.342600 and 15.936440 mc. These frequencies are multiplied to provide nominal frequencies of 2220 and 4080 mc. As noted previously, the 4080-mc signal is further amplified by the TWT and used for the microwave beacon; it is also used as the pump for the up converter and mixed with the 2220-mc signal to provide the 6300-mc supply for the down converter.

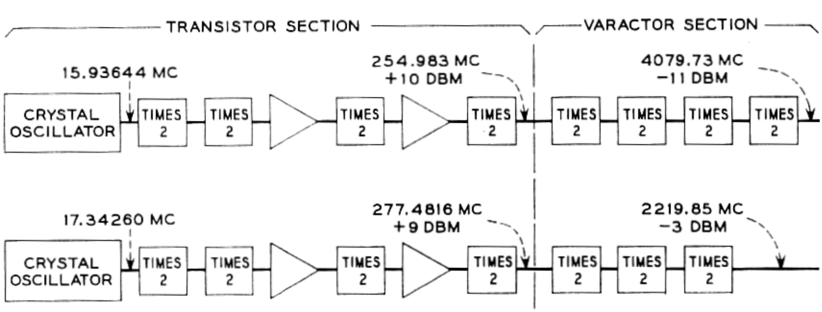

Fig. 18 — The microwave carrier supply.

A more detailed description of the components of the microwave carrier supply follows.

3.10.1 *Transistor Section*

The third overtone AT-cut crystals selected for use in the satellite had a frequency variation of only ±2 parts per million over a 0 to 50°C temperature range, so neither temperature stabilization nor temperature compensation was necessary. Good short-term stability was obtained by maintaining an RF crystal current of 6 milliamperes.

Since frequency multiplication decreases the ratio of signal power to noise power, it is necessary to start the multiplication process with an extremely pure signal. To obtain the spectral purity, extensive shielding, decoupling, and filtering were necessary to isolate the two transistor sections from each other, from the power supply, and from other oscillators in the satellite.

The transistor frequency doublers are class C amplifiers with a collector tank tuned to the second harmonic. The first and second multipliers are operated common-emitter, the third and fourth common-base. The efficiency of the third and fourth multipliers is improved by the inclusion of a full-wave rectifier in the input circuit.

3.10.2 *Varactor Octuplers*

Design of the varactor octuplers (250–2000 mc) is common for the two strings, only the tuning being different. The octupler consists of three doubler stages, each using lumped elements; early experimental work showed frequency doubling to be the most efficient means of multiplication.

The basic doubler circuit is shown in Fig. 19. The series resonant circuit in the output is resonant at frequency f, while the input trap is resonant at $2f$. The shunt input inductor is chosen to match the diode

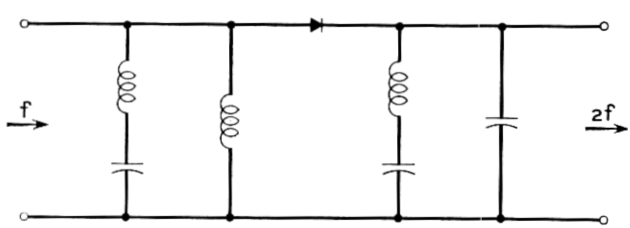

Fig. 19 — Basic varactor frequency doubler circuit.

impedance: the output capacitor is tuned for maximum output. Cascading of these stages requires additional tuning elements.

The first two stages are placed at zero bias by RF chokes connected to ground; the third stage is self-biased. Varactor diodes are Western Electric Co. units chosen for stability and high Q.

Shunt resistors were added around the diode in each stage to improve stability with changing input power level. The addition of these resistors also made tuning less critical and eliminated a problem of parametric oscillation. Critical cable lengths at both the input and output of each octupler were experimentally determined.

The octupler, before encapsulation, is shown in Fig. 20. Connections are made to the diodes through a cup welded to the small end of the diode and a shell screwed to the other. Tuning is accomplished with variable quartz capacitors and air inductors of 14-gauge tinned copper wire.

3.10.3 *The 2040- to 4080-mc Doubler*

This doubler, shown in exploded form in Fig. 21, uses the same principle as the varactor doubler described above. The diode is placed in a coaxial cable which terminates in a 4-gc waveguide. This waveguide effectively reflects the 2-gc energy. The input trap is provided by a quarter-wave sliding transformer.

A short length of rigid coaxial line matches the loop and waveguide to the diode. Output power at 4080 mc is set by rotating the output loop

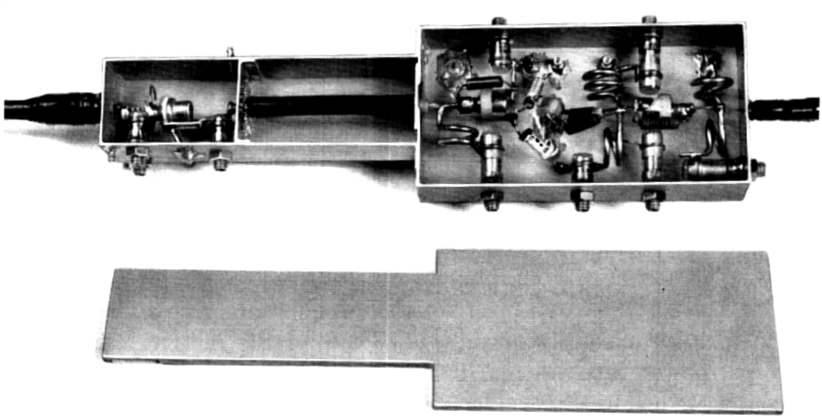

Fig. 20 — The varactor octupler.

Fig. 21 — The 2- to 4-gc doubler.

in the waveguide. The spring shown in the photograph presses the chuck and diodes against the center conductor of the rigid coaxial cable. The capacitor wire which slides inside the insulating sleeve is selected to tune the circuit. The decoupling afforded by this capacitor results in self-biasing of the diode.

3.10.4 *Over-All Design Considerations*

The microwave carrier supply must be self-starting, must be free of spurious oscillation and must maintain proper output level over the temperature range. The above conditions require special attention to tuning procedures, especially for the varactor multipliers.

As mentioned above, relative amplification of noise and spurious sidebands in each multiplier stage required close attention to suppression of these unwanted signals in the early transistor stages. All inherent noise sidebands at 2000 and 4000 mc are at least 40 db below the desired signal.

The crystal-oscillator frequency is initially adjusted to within ± 1 part per million of the desired frequency, and aging in two years is expected to be less than one part per million.

3.11 *Microwave Filters*

Because space in the electronics canister is so limited, all the waveguide filters utilize direct-coupled cavities. There are four single waveguide filters, two dual waveguide filters, one coaxial low-pass filter, one strip-line filter, and a YIG limiter which is also a filter. All data given in this section are for room temperature.

Filter 1 is a two-section filter which is designed to give a 20-db insertion loss at the image frequency, 6210 mc, of the receiver. This filter has a half-power bandwidth of 100 mc and an insertion loss of 0.2 db at the center frequency of 6390 mc.

Filter 2 is also a two-section filter. It is built in a gradual 90° H-plane bend. The insertion loss at the center frequency of 6300 mc is 1.2 db, and the half-power bandwidth is 16 mc. The filter was made narrow band to give high insertion losses except at 6300 mc. The need for the high insertion losses is discussed in the stability section of this paper.

Filter 3 is a three-section filter connected between the up converter and the monitor; its purpose is to pass the signal at (4080 + 90) mc and reject the signal at (4080 − 90) mc. It also must provide about 20 db insertion loss at 4080 mc so leakage of the pump signal through the up converter will not contribute to the output of the microwave monitor. This filter has a 90-mc bandwidth and an insertion loss of 0.15 db at the 4170-mc center frequency.

Filter 4 is the YIG limiter. The YIG crystal is placed in a resonant cavity, which below the threshold has loaded Q of about 200.

Filter 5 is a five-section direct-coupled filter having a center frequency of 4080 mc and a half-power bandwidth of 20 mc; the insertion loss at 4080 mc is less than 1 db. This filter was made as narrow band as space allowed to give high insertion losses except at 4080 mc. The high insertion loss is needed to ensure stability in some of the closed circuits in Fig. 1, which are discussed in the stability section of this paper. In order to position the large filter (15-inch length) in a canister of reasonable size, it was necessary to build the filter with a 90° E-plane bend in one of the cavities. It was found that by measuring the electrical length of the well-matched 90° mitered bend, the distance between the inductive posts separated by the bend could be calculated, and the resulting filter had no measurable difference from the "same" filter made in straight waveguide. The 90° mitered bend was placed as near the center of one of the five cavities as possible.

Filter 6 is a two-section strip-line filter that has a half-power bandwidth of 100 mc centered at 2220 mc. The purpose of this filter is to prevent the pump signal and harmonics and beat frequencies of the two input signals to the BO modulator from going back into the microwave carrier supply. The filter has excellent temperature stability and has an insertion loss of 0.7 db at 2220 mc.

The dual filters used in the combining and separating networks are described in previous sections of this paper.

IV. CIRCUIT STABILITY

As stated previously, the circuit stability problem is complicated by the feedback paths which are inseparably connected with the reflex circuit of Fig. 1. Three feedback paths created by the reflex circuit, the coupling between the satellite output and input, and miscellaneous couplings are described in this section. The loops are shown in Fig. 22.

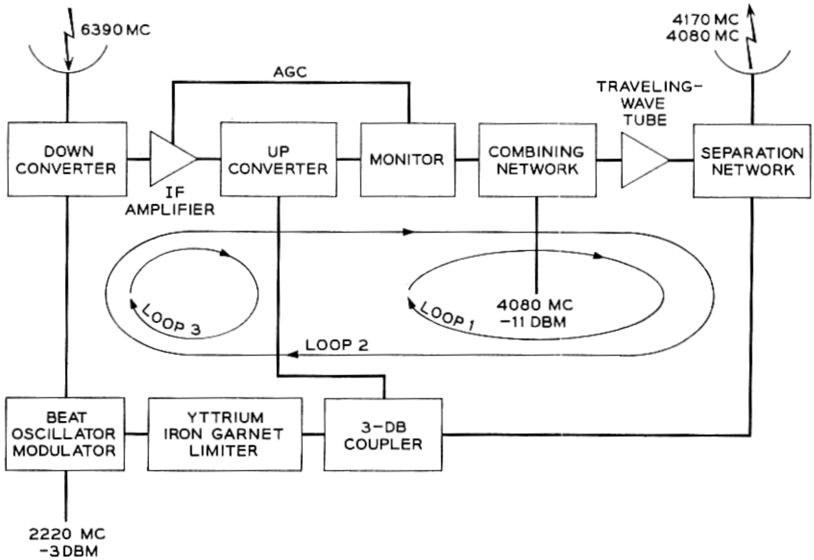

Fig. 22 — Simplified block diagram showing the feedback paths.

4.1 *Feedback Path One*

The circuit loop created by this feedback path from the separation network to the up converter includes the monitor, TWT and combining network. Two types of oscillation are possible in this loop.

The first is a straightforward RF oscillation. The TWT will act as an amplifier over a very wide frequency range; therefore the filters in this circuit were designed so at all frequencies there is margin against oscillation of at least 40 db at room temperature. This margin is made high because at least 30 db of this margin is due to the isolation between the pump input and signal output ports of the up converter, measured when it is new and the two diodes are balanced, and the isolation will be greatly reduced if one diode degrades. Also, the relatively narrow-band filters can change the margin by 5 db at the temperature extremes of 0 and 50°C. At frequencies high enough to allow higher-order modes to propagate in the waveguide, the waveguide filters give small insertion losses, so the low-pass coaxial filter was put in the circuit to provide the proper margin against oscillation at the higher frequencies. The coaxial filter has another use explained later in this article.

A second possible oscillation in this loop manifests itself in the form of amplitude modulation (AM) of existing signals. This is explained by showing that an AM signal applied to the input of the open loop causes a similar AM signal at the output of the loop. If the amplitude and phase of the modulation envelope are proper, this circuit can oscillate in the AM mode; the phase of the carrier signal is not critical. Assume that the loop is opened at the input to the TWT and that an AM wave at 4170 mc is applied to the input. The data in Fig. 2 show that an increase in amplitude of the signal at 4170 mc results in a decrease in the output at 4080 mc. Thus, the envelope of the AM output at 4080 mc is 180° out-of-phase with the envelope of the input AM wave at 4170 mc. The amplitude-modulated wave from the output of the TWT goes through the separation network and filter 5 into the up converter. Unless the sidebands of the AM wave are attenuated more than the carrier, the output of the filter has the same general shape as the input to the filter. The curves in Fig. 12 show that if the pump signal to the up converter is amplitude modulated, the output signal at 4170 mc is also amplitude modulated, even though the modulation index of the output is much less than that of the input. If there is a phase delay in this loop corresponding to 180° at the modulation frequency, the AM feedback is positive; and if the modulation index of the signal at the output of the open loop is the same as or greater than the modulation index of the signal at the input to the loop, the circuit will oscillate in the AM mode when the loop is closed. It should be emphasized that the AM oscillation occurs only when there is a communications signal going through the TWT. The AM response of the various subassemblies in loop 1 are discussed in the next few paragraphs.

When the 4170-mc signal going into the TWT is amplitude modulated, both 4170- and 4080-mc signals from the tube are amplitude modulated. The degree of modulation of the beacon depends on the degree of modulation of the input signal and the operating point of the tube. The AM conversion gain (G_{cam}) of the TWT is defined at the ratio of the modulation index (m_{out}) of the beacon output to the modulation index (m_{in}) of of the input signal frequency. In decibels this is expressed as

$$G_{cam} = 20 \log_{10} \frac{m_{out}}{m_{in}} .$$

Since the ordinate and abscissa of Fig. 2 use the same scale, the TWT will have G_{cam} greater than 0 db when the magnitude of the slope of the beacon-output signal-input curve is greater than 45°. The operating point of the TWT is chosen so that the AM gain of the tube is less than 0 db and the power output at the signal frequency is near 35 dbm. Having the operating point below saturation is desirable, from the standpoints of AM gain and intermodulation of signals, but it is undesirable from a standpoint of obtaining maximum power output.

By selecting the proper biasing resistors for the diodes in the up converter, the output of the up converter can be made relatively independent of the 4080-mc pump signal (see Fig. 12), so this unit can be made to have very small (-18 db) AM conversion gain and thus provide good margin against AM instability for this loop.

At very low modulating frequencies, the feedback in this loop is negative because there is a 180° phase shift in the TWT and no phase shift in the up converter. However, at higher modulating frequencies, the phase delay in the various elements in this loop causes phase shifts of several hundred degrees, so the feedback at some frequencies is positive. The calculated delay for all the networks in the loop is approximately 75 ns; about 36 ns of this is due to the narrow-band filter 5.

If an AM signal with a modulating frequency f_b in loop 1 is delayed by T seconds, the envelope of the wave is shifted by

$$\varphi = 2\pi f_b T \text{ radians.}$$

When this phase angle φ is some odd multiple of 180°, the feedback will be positive because there is a constant 180° phase shift of the modulation envelope introduced in the TWT. Using these relationships and a 75-ns circuit delay, the lowest frequency f_b calculated to give positive feedback is 6.7 mc. When the circuit gain was adjusted to have more than 0 db (by changing the drive to the TWT), the output wave was amplitude modulated at 6.1 mc.

If the input wave to filter 5, which has a half-power bandwidth of 20 mc, is amplitude modulated at 6.1 mc, the sidebands and carrier all experience about the same insertion loss; so the AM gain of the filter is about 0 db. If, however, the modulating frequency is 20 mc, the next modulating frequency that gives positive feedback, the sidebands are attenuated 11 db more than the carrier; so the filter gives an AM gain of -11 db. Even if space were available to reduce the bandwidth of this filter, it would do no good because the filter delay, which contributes an appreciable part of the circuit delay, would increase and the frequency of oscillation would decrease, so the filter would have little effect on AM gain at the lower frequency. The remaining filters in this loop are quite broadband, so they have essentially 0-db AM gain.

The monitor circuit consists of two diodes inserted in the waveguide. The insertion loss caused by these diodes depends upon the resistors used in the self-biasing circuit; the smaller these biasing resistors, the higher the insertion loss. By putting selected capacitors across the biasing resistors, the monitor can be made to partially clip the peaks of the AM wave and have little effect on the troughs. By this means, the AM gain of the monitor is reduced to -3 db at $f_b = 6$ mc.

The final open-loop AM gain of the circuit was made as low as -18 db when the TWT voltages were normal. A decrease in tube supply voltages of 3 per cent causes the tube characteristics to change, and this can reduce this margin to -13 db.

If the 6390-mc received signal power is large enough to extend above the upper limit of the AGC circuit, the 4170-mc signal drive to the TWT is increased an amount depending on the input signal, and the AM conversion gain of the tube will be increased; see Fig. 2. This increase in signal drive to the TWT decreases the beacon output and the 4080-mc pump to the up converter and thus increases its AM conversion gain. An experiment on the ground has shown that if the 6390-mc input is increased to about -35 dbm, the AM oscillation will start, but does not damage the circuit. This high input level is not possible under normal orbital operating conditions.

4.2 *Feedback Path Two*

The local oscillator signal for the down converter is obtained by combining in the BO modulator the 4080-mc signal from the TWT with a 2220-mc signal from the microwave carrier supply. This method of obtaining the local oscillator signal creates feedback path two, which consists of the down converter, the IF amplifier, the up converter, the monitor, the TWT, the YIG limiter, the BO modulator, and several passive filters. Here, as in loop 1, two types of oscillation can exist, straight RF oscillation and AM oscillation.

In loop 2 a signal at frequency f, expressed in mc, in the IF amplifier will cause a frequency of $(4080 + f)$* mc in the output of the up converter. This signal is amplified by the TWT and is fed back through the separation network to the BO modulator through the YIG limiter. Even if the nonlinearities of the TWT and YIG limiter are neglected, this signal at $(4080 + f)$ mc plus the normal 4080-mc signal is combined in the BO modulator with the 2220-mc signal from the microwave carrier supply to give a frequency of $(6300 + f)$ mc. Going into the down converter is the "weak" signal at $(6300 + f)$ mc in addition to the strong local oscillator signal at 6300 mc. These signals are combined in the down converter to give an output frequency f. Since this frequency is the same as the assumed input frequency in the IF amplifier, this feedback loop can oscillate if the gain is large enough, even though the signal changes frequency three times in the loop. When no 6390-mc signal

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* A frequency of $(4080 - f)$ mc is also present, but it is attenuated more than the higher frequency.

is applied to the down converter, the gain of the IF amplifier is about 87 db over a wide frequency range, and the gain of the TWT is greater than 42 db, so the total gain of the amplifiers in this loop is very high. As was the case in loop 1, the margin against oscillation was made 40 db or more because high insertion losses were obtained with two balanced modulators, the BO modulator and the down converter, and an unbalance in any one could change the insertion loss appreciably. In order to get the required margin for this loop, in which the combined gain of the amplifiers is at least 129 db, and still keep the microwave package small, two filters in this loop were built in waveguide bends: filter 5 has a 90° mitered E-plane bend, and filter 2 has a gradual 90° H-plane bend. Also to improve the margin, a high-pass filter was put at the input to the up converter.

The above discussion of loop 2 did not discuss the possibility of nonlinearity of the TWT. When the nonlinearity of the tube is considered and a frequency f occurs in the IF amplifier, there are output frequencies from the TWT of $m(4080 + f) \pm n(4080)$ mc, where m and n are integers. Many of these frequency components pass through the 4-gc filters because the frequencies are such that the waveguide can propagate higher-order modes. These frequencies can combine in the nonlinear limiter to produce frequencies of $4080 \pm f$, going into the BO modulator. For example, the 3×4080 component can combine in the limiter with the $4080 + (4080 + f)$ components to give a frequency of $(4080 - f)$ mc. This signal at $(4080 - f)$ mc combines with 2220-mc component to give a signal at $(6300 - f)$ mc which is fed into the down converter along with the normal 6300-mc signal. These two signals in the down converter give the original f frequency in the IF amplifier, so oscillation occurs if there is sufficient gain. There are many combinations of high frequencies which create this problem; only one example has been given. A three-section low-pass coaxial filter having an upper usable frequency of 4500 mc preceding the YIG limiter provides an insertion loss greater than 60 db above 6000 mc, and this is enough to eliminate oscillation of the type due to nonlinearities in the TWT and other elements.

The same AM problem exists in loop 2 as exists in loop 1, since the TWT is common to both loops. The YIG limiter does not eliminate the AM oscillation in either loop. Positive feedback occurs in loop 2 when the modulating frequency is about 2.5 mc, because the delay in this loop is approximately 200 ns. Four circuits are involved in loop 2 which were not in loop 1: (a) the BO modulator, (b) the YIG limiter, (c) the IF amplifier, and (d) the down converter.

The BO modulator was found to have an AM conversion gain of approximately 0 db when the average pump input is +14 dbm (normal level) and the modulating frequency is below 2 mc. The AM gain decreases to about −3 db when the pump is increased to +16 dbm and increases to +2 db when the pump is decreased to +12 dbm. Decrease in AM gain with increase in modulating frequency is very small in the trouble region below 5 mc.

Characteristics of the YIG limiter show it has excellent limiting characteristics for slow variation in signal level; but unfortunately, when the envelope of the input is amplitude modulated at rates higher than 100 kc, the limiter has little effect on the peak-to-peak variations. Since the average value of the signal is limited, the limiter actually has AM gain. Under normal conditions the limiter has an AM gain of 3 db.

For purposes of AM analysis, the IF amplifier can be considered linear and thus has an AM gain of 0 db. At frequencies below a few thousand cycles, the AGC circuit reduces amplitude modulation, but at 20 kc and above it has virtually no effect; see Fig. 10.

The down converter alone has an AM conversion gain of about −15 db when the modulating frequency is less than 10 mc, and even less when the frequency is higher. However, in early tests on the down converter and IF amplifier together, the AM conversion gain was found to be greater than 0 db. The reason for the increase in gain is as follows. When an AM signal is used as the local oscillator for the down converter, the output will contain, among other signals, a signal at the modulation frequency. If the low-frequency circuitry preceding the first IF amplifier does not give sufficient attenuation to this modulating frequency, it can overload the first two stages and cause the output IF frequency to be amplitude modulated. When the front end of the IF amplifier was modified to properly attenuate the frequencies below 15 mc, the AM conversion gain of the down converter and IF amplifier was consistent with the value predicted from the gains of the units measured separately.

4.3 *Feedback Path Three*

As shown in Fig. 22, feedback path or loop 3 consists of the down converter, the IF amplifier, the up converter, the 3-db coupler and YIG limiter, and the BO modulator.

RF oscillation in this loop is very similar to the RF oscillation described in loop 2. Consider this loop opened at the IF amplifier, where an input signal at a frequency f in mc is applied. The sum and difference frequencies $(4080 + f)$ and $(4080 - f)$ mc, originating in the up converter, leak down the pump arm through the directional coupler and into the BO modulator via the YIG limiter. In this modulator, the two frequencies combine with the 2220 mc to give $(6300 + f)$ and $(6300 - f)$ mc signals into the down converter. These signals combine with the normal 6300-mc signal in the down converter to give a signal f in the IF amplifier. Since this frequency is the same as the assumed input, the circuit can theoretically oscillate. Loop 3 has less gain than loop 2, and the conversion insertion loss from the IF arm to the pump input arm is very large (30 db). Conversion insertion loss is calculated using the ratio of the magnitude of the $(4080 \pm f)$ mc signal in the pump arm to the magnitude of the input signal at f mc. Loop 3 was analyzed after loop 2, and when the margin against RF oscillation in loop 2 was made 40 db, the resulting margin in loop 3 was greater than 50 db.

The AM problem of loops 1 and 2 is no problem in loop 3 because the AM conversion gain between the IF input and pump ports of the up converter is extremely low, and the YIG limiter and BO modulator give a combined AM gain of only a few db.

4.4 *Output-Input Coupling*

The TWT produces noise power over a wide range of frequencies including the satellite input and image frequency bands of 6390 and 6210 mc. Since the cutoff frequency for the TE_{20} mode in the waveguide used to make the filters in the separation network is 6300 mc, their insertion loss at 6390 mc is small. When the output of the repeater, less antennas, is connected to the input through an attenuator, the noise output from the repeater is 3 db higher when the attenuator setting is 27 db than when the attenuator is "infinite." The measured isolation between the transmitting and receiving antennas over the frequency range 6150 to 6450 mc is greater than 75 db in the temperature range -100 to $+100°$F, so the noise output of the TWT that is coupled into the down converter is negligible.

The coupling between the antennas at 4170 mc is not a problem because the input waveguide is below cutoff at this frequency and provides enough insertion loss to prevent the high power from the TWT from having any effect on the down-converter crystals.

4.5 *Miscellaneous Couplings*

Fig. 23 shows an end view of the cylindrical part of the electronics canister after all the electronics subassemblies have been mounted in the canister, but before it has been foamed and the metal domes have been welded in place. After the domes are in place, the canister becomes a big echo box, and reducing the couplings between units to acceptable levels becomes very difficult. The power supply contains one flip-flop circuit operating at a fundamental frequency of 2.5 kc and another operating between 25 and 50 kc. Also, there are four crystal oscillators that are used to obtain the microwave carrier supply signals at 2220 and 4080 mc, the VHF beacon signal at 136 mc, and the local oscillator signal at 128 mc for the command receiver. The frequencies of the crystal oscillators (15.93644, 17.34260, 17.00625, and 31.9750 mc) and their harmonics, plus all harmonics from the flip-flop circuits, create a noise spectrum that can cause trouble in the IF amplifier and the converters. The problems associated with the myriad couplings in the canister were solved with extensive shielding, filtering, and patience.

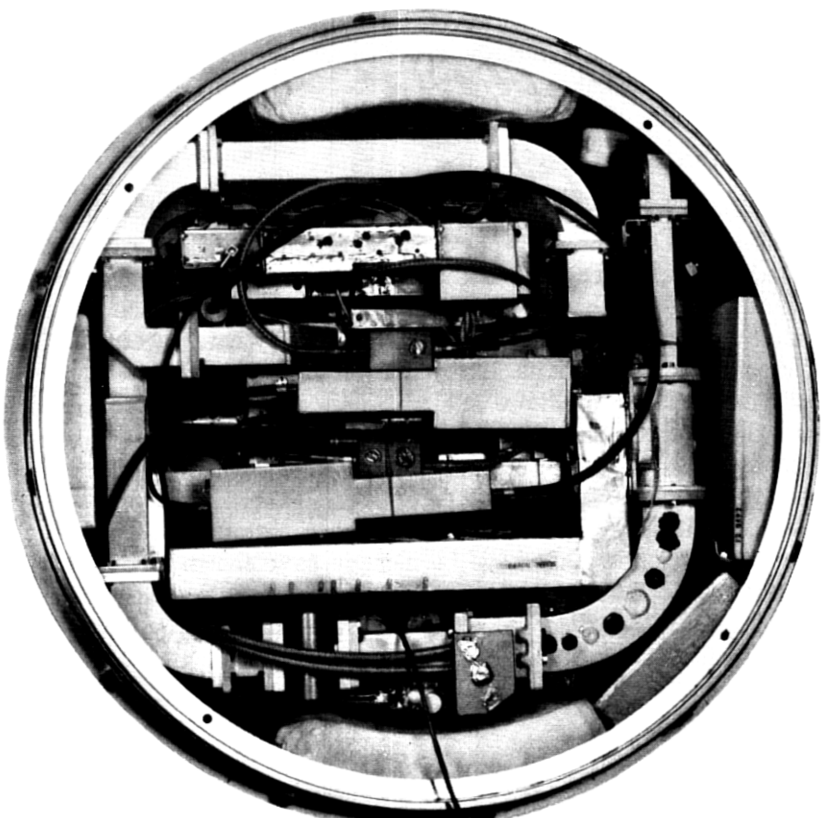

Fig. 23 — Electronic parts in the canister.

4.6 *Over-All Noise Performance*

The frequency modulation noise due to the satellite repeater is an important factor in the operation of the system as a whole. Consideration of this noise contribution is complicated by the complex nature of the repeater circuit because of the dual use of the traveling-wave tube. The predominant noise source is at the input to the IF amplifier. Operated as it is with a low level 6300-mc signal, the down converter does not add significantly to the noise, but there is considerable loss of signal power through the device, and the noise figure of the down converter and the IF amplifier together is as much as 12.5 db. Noise contributions due to other parts of the repeater, including the microwave carrier supply, increase this figure.

The noise figure of the satellite as measured before launch by a noise lamp is 13.5 db ± 1 db.* However, the noise does not have a flat spectrum over the band of interest and this number applies in the region where the noise is flat.

V. CONCLUSIONS

The Telstar communications circuit is a broadband nondemodulating repeater. To obtain a high power efficiency the traveling-wave tube is caused to amplify two signals, a fact that somewhat complicates the circuit. Consequently, several stability problems exist which are overcome by the provision of frequency filters and by careful control of the nonlinearity of the circuits. Every portion of the circuit had to be specially designed for satellite use. Even where components were available with the required electrical performance, it proved necessary to seek reduced power consumption and to meet unusually stringent size and weight restrictions.

The circuit has given an exceptionally good performance. None of the complete assemblies that have been made have shown any significant variations of operating levels over the course of six months. Although no failure in this part of the satellite is expected, in many parts of the circuit sufficient margin has been built into the circuit that in the event of such a failure the circuit will continue to operate with reduced performance.

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* Measurements made of the system noise spectrum using narrow-band analyzers indicate a satellite noise figure in the flat region of 15 db ± 2 db. If the noise spectrum around the carrier is integrated over a 20-mc band, then an equivalent noise figure of 16.5 ± 2 db is obtained.

VI. ACKNOWLEDGMENTS

The circuits described in this paper represents the combined efforts of so many key personnel that a complete listing is not practical. The work was done under the direction of R. H. Shennum, who suggested the use of the reflex circuit and most of the circuit arrangement.

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