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A Dual-Channel, VHF-UHF Transmitter for Gemini Experiment D-14

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ABSTRACT

The objective of Gemini experiment D-14 was to measure the electron content of the ionosphere below the Gemini spacecraft by means of a dual-frequency Faraday rotation experiment utilizing two spacecraft-borne transmitters operating in the vhf-uhf bands. The Model NT-1 dual-channel solid-state transmitter was designed for this experiment. The transmitter produces, simultaneously, four watts of rf energy at 133.9 Mc and two watts of 401.7 Mc. This transmitter is a highly ruggedized unit capable of operation under the severe vibration and temperature requirements of spacecraft environment.

The spacecraft-borne transmitters supplied at least one watt of cw rf energy at each frequency for simultaneous radiation from a colinear dipole antenna. This power output was necessary in order to obtain the desired signal-to-noise ratios at the ground-based receiving stations for accurate measurements.

PROBLEM STATUS

This is an interim report on one phase of this problem; work continues on other phases.

AUTHORIZATION

NRL Problem R07-13
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A DUAL-CHANNEL, VHF-UHF TRANSMITTER FOR GEMINI EXPERIMENT D-14

INTRODUCTION

The DOD Gemini D-14 experiment was concerned with a measurement of the electron content of the lower ionosphere by means of a dual-frequency Faraday rotation experiment. This experiment, conducted as a part of two Gemini spacecraft missions, utilized vhf and uhf signals radiated simultaneously from a transmitter aboard the spacecraft. The vhf signal was utilized as a vernier, while the uhf signal was utilized to avoid ambiguities in measurement. The Gemini spacecraft was selected as the transmitter platform for two reasons: (a) the attitude of the spacecraft with respect to the earth can be controlled in such a way as to maintain a specified plane of polarization to which Faraday rotation angles may be referenced, and (b) only the inhomogeneities in the electron content of the ionosphere below the spacecraft will influence data to be analyzed.

The responsibility of the Naval Research Laboratory concerning the Gemini D-14 Experiment is threefold; namely, the design and development of the vhf-uhf signal source and antenna system installed aboard the spacecraft; the construction, instrumentation, and manning of the ground-based stations located in Antigua and Hawaii; and the analysis of data received during the orbital passes of the spacecraft over the ground stations.

This report concerns the design and production of the solid-state, dual-channel transmitter which was installed aboard the spacecraft to supply the two cw signals required to carry out the proposed Faraday rotation experiment.

GENERAL DESCRIPTION

The Model NT-1 dual-channel transmitter is an all-solid-state generator designed to produce two harmonically related signals and to meet the electrical specifications set forth by NRL as well as the mechanical and environmental specifications set forth by the McDonnell Aircraft Corp. [1,2].

A summary of these specifications is given in the following list.

TRANSMITTER SPECIFICATIONS

Electrical	
Output frequencies	133.9 Mc, 401.7 Mc ±0.005 percent
Output power	One watt per channel
Output impedance	50 ohms
Primary dc voltage	20 to 30 volts
Operating dc voltage	24 volts
DC input power	48 watts maximum
Radio-frequency interference (RFI)	As per MIL-I-26600

TRANSMITTER SPECIFICATIONS — Continued

Mechanical	
Orientation	Any
Volume	Up to 504 cu in.
Dimensions	Restricted to 6 × 7 in. base with up to 12 in. of height
Weight	Six pounds
Environmental Capacity	
Vibration-launch	Random Gaussian vibrations of up to $0.06g^2$/cps from 20 to 2000 cps
Vibration-orbit	Random Gaussian vibrations of $0.008g^2$/cps from 20 to 500 cps
Acceleration	Up to 7 g within 300 seconds
Temperature	Operation over an environmental range of -60° to 160°F, with coldplate temperature of 40° to 120°F
Acoustical noise	An overall sound-pressure level of 155 db directed toward the three most sensitive axes of unit.
Atmospheric pressure	Environmental pressure of 10^{-8} psi
Humidity	As per MIL/E/5272

The transmitter unit generates rf energy at 133.9 Mc and 401.7 Mc radiated, simultaneously, from a vertically polarized colinear dipole antenna [3]. To obtain accurate measurements of Faraday rotation angles, it is desired to have a received signal-to-noise ratio in excess of 25 db for spacecraft ranges varying from 150 to 1000 nautical miles. The output power level specified in the preceding list is derived from the expression of signal-to-noise ratio for a one-way radio transmission as given by Blake [4]:

$$S/N = \frac{P_t G_t G_r \lambda^2 F^2}{(4\pi)^2 R^2 L_t L_r K T_n B_n} \tag{1}$$

where

P_t = transmitter power output

G_t = transmitter antenna gain

G_r = receiving antenna gain

λ = wavelength

F = propagation factor (assumed to be one for free space)

R = range from transmitter to receiver

L_t = transmitter transmission-line and antenna loss factor

L_r = power losses between receiving antenna and receiver input

K = Boltzmann's constant

T_n = receiving system noise temperature

B_n = receiver noise bandwidth.

The parameters of the spacecraft-borne transmitter and the ground-based receiving systems are given in the accompanying list along with the solutions of Eq. (1) for ranges of 150 and 1000 naut mi.

TRANSMITTER AND RECEIVER PARAMETERS

Transmitter		
Output frequencies	133.9 Mc	401.7 Mc
Output power	1 watt	1 watt
Antenna gain	2 db	4 db
L_t	2.5 db	3.2 db
Receivers		
Antenna gain	20 db	30 db
Noise figure	4 db	6 db
Noise bandwidth	40 kc	40 kc
System noise temperature	1952 deg	1569 deg
L_r	0.3 db	1.2 db
S/N Calculations		
R = 150 naut mi	45.2 db	47.9 db
R = 1000 naut mi	28.5 db	31.3 db

It can be seen, from the tabulated results of signal-to-noise ratio, that the selection of one watt power output from each transmitter channel is adequate. In order to assure greater frequency and power-output stability, the transmitter power supply was designed as an integral part of the unit.

The restriction on the dimensions of the transmitter chassis base was dictated by the limited space available in the equipment-adapter section of the spacecraft and existing coldplate facilities. By way of explanation, the coldplate is a heat-exchange platform upon which all electronic equipment in the equipment-adapter section will be mounted. The temperature of the coldplate can vary over a range of 40° to 120°F, depending upon a number of spacecraft operating conditions.

A block diagram relating the transmitter unit to its associated equipments in the spacecraft is shown in Fig. 1, where the transmitter is shown in solid lines; dashed lines represent the diplexer, which combines the rf outputs, and the antenna.

CIRCUIT DESCRIPTION

The primary objective of this transmitter design (Fig. 2) was to obtain the desired power output while utilizing the least number of stages consistent with good frequency and output-power stability. The basic approach employed in this project was to generate moderately high power at a low frequency and then multiply to the desired output frequencies. It was possible to utilize lumped-constant circuitry throughout, since the highest frequency of interest was in the low uhf region, and circuit losses could be kept to a minimum through the use of high-Q coils and tuning capacitors. The two output frequencies were derived from a single crystal-controlled oscillator, operating at 33.475 Mc, the output of which was equally divided between the vhf and uhf channels and multiplied to the desired 133.9-Mc, and 401.7-Mc frequencies.

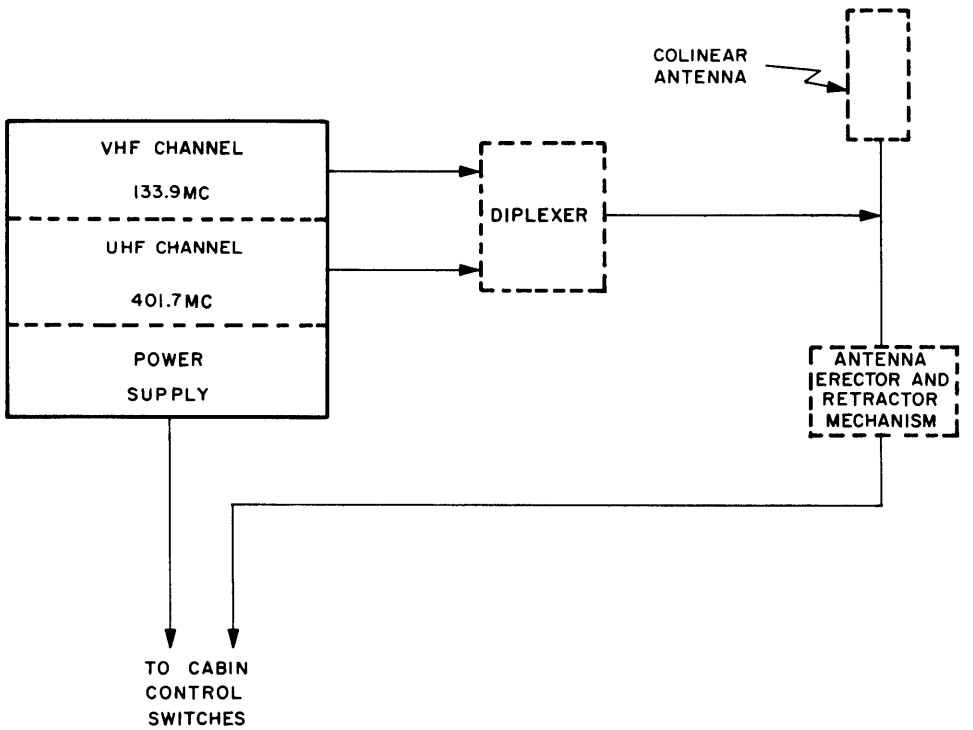


Fig. 1 - Block diagram of spacecraft installation

Oscillator

The oscillator circuit utilized a type 2N2219 silicon transistor in a crystal-controlled, common-emitter Pierce configuration. Positive bias was supplied to the base through a voltage-divider network, and an RC network in the emitter controlled thermal-runaway tendencies. Temperature compensation of the circuit was obtained by shunting the tuning capacitor with a negative-coefficient temperature-compensating capacitor. The oscillator coil was tapped in such a manner that a high feedback voltage was available at the high side of the inductance. The approximately 250-mw oscillator output was equally divided between the two multiplier channels through the use of two bifilar-wound coils link-coupled to the tank circuit.

VHF Multiplying Channel

The vhf channel, utilizing transistorized circuitry throughout, provided a multiplication factor of four. The oscillator frequency was multiplied in two doubler stages to 66.95 Mc and 133.9 Mc respectively. The common-emitter configuration was selected for use in the multiplying stages for the following reasons: (a) maximum available power gain could be obtained at vhf without neutralization, provided the f_{max} of the transistor was well above the operating frequency, and (b) high stability is possible for a given amount of gain, because the power gain is composed of both voltage and current gain, and the internal feedback from output to input is negative [5]. The type 2N3118 first doubler and the type 2N3137 second doubler were selected on the basis of their high value of f_{max}, 190 and 420 Mc respectively, the output power capacity, and the free-air power-dissipation characteristics. In order to obtain the nonlinear characteristics necessary for frequency multiplication, the transistors were biased for Class C operation and adjusted for highest harmonic output. Problems incurred in driving the final power-amplifier stages led to the selection of the 2N3137 as the second doubler stage due to its superior power-gain characteristics.

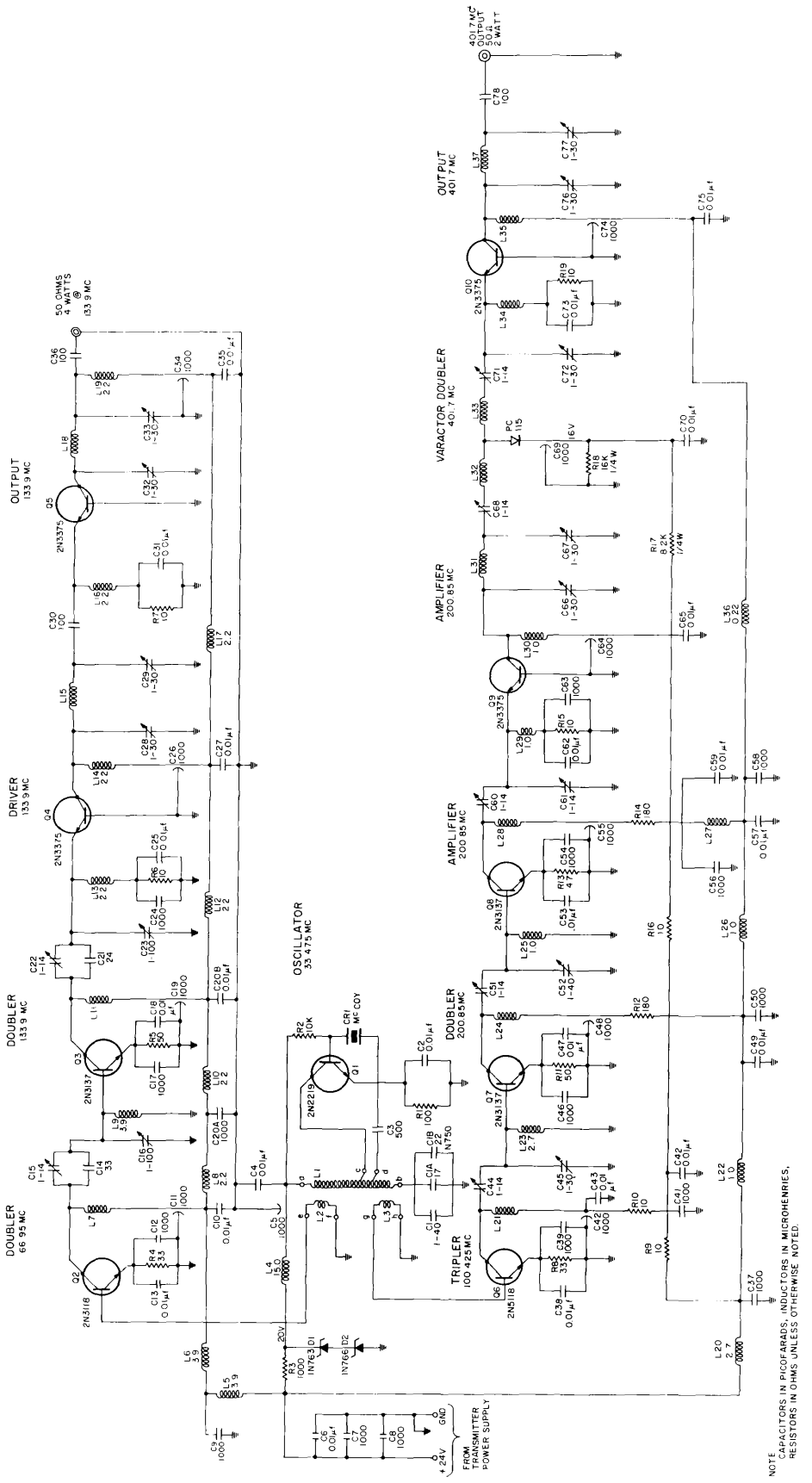


Fig. 2 - Schematic diagram of rf circuitry

In order to obtain efficient operation throughout the multiplier-amplifier chain, the input impedance of each transistor stage must be transformed to the correct value of load for the collector circuit of the preceding stage. A capacity-coupled network was chosen to perform this function in the multiplier stages because it is probably the optimum power-matching network with respect to reliability, simplicity, and size. The equations utilized to design the capacity-coupled network (Fig. 3) are given by Minton [6]:

$$R_1 = \frac{V_c^{\,2}}{2P_o} \text{ (for Class C)} \tag{2}$$

$$X_L = \frac{R_1}{Q_L} \tag{3}$$

$$X_{C2} = \frac{R_2}{\left[\left(\frac{R_2}{R_1}(Q_L^{\,2}+1)\right)\right]^{1/2}} \tag{4}$$

$$X_{C1} = \frac{Q_L R_1}{(Q_L^{\,2}+1)}\left(1 - \frac{R_2}{Q_L X_{C2}}\right) \tag{5}$$

where V_c is the dc collector supply voltage, P_o is the required power output, R_1 is the collector load impedance, R_2 is the input impedance of the next stage, and Q_L is the operating Q of the circuit. Parallel resonance of the circuit is obtained by L and the parallel combinations of $C_1 + C_2$ and the capacities associated with the transistors.

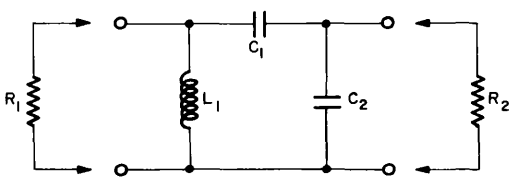


Fig. 3 - Schematic diagram of capacitive-coupled transformer network

For a sample calculation, consider the interstage network of the first doubler circuit in the vhf channel. The circuit parameters for this case are given below:

V_c = 20 volts	R_2 = 70 ohms
P_o = 0.2 watt	Q_L = 10
R_1 = 1000 ohms	$\omega_o = 4.2 \times 10^8$

Substituting these values into Eqs. (2) through (5) yields the following circuit values:

X_L = 100 ohms	L = 0.25 μH
X_{C1} = 75 ohms	C_1 = 32 pf
X_{C2} = 26 ohms	C_2 = 85 pf

The final output power level was obtained through the use of two 2N3375 transistor amplifier stages. Extensive breadboard investigation established that the 2N3375 operated more efficiently as a power amplifier when utilized in the common-base configuration. Conventional pi-matching networks (Fig. 4) were used to provide impedance matching between amplifier stages and the antenna load impedance. Again utilizing Minton's equations:

$$R_1 = \frac{V_c^{\,2}}{2P_o} \tag{6}$$

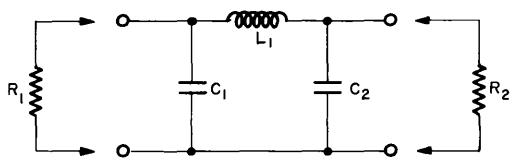
$$X_{C1} = \frac{R_1}{Q_L} \tag{7}$$

$$X_{c2} = \frac{R_2}{\left(\frac{R_2}{R_1}(Q_L{}^2 + 1) - 1\right)^{1/2}} \tag{8}$$

$$X_L = \frac{Q_L R_1}{Q_L{}^2 + 1}\left[1 + \frac{R_2}{Q_L X_{C2}}\right] \tag{9}$$

In power amplifiers, the operating Q is usually between 7 and 10. Lower values result in more efficient transfer of power to the load but do not attenuate harmonics as well as the higher values. A value of 8 was used in the design of these amplifier stages.

Fig. 4 - Schematic diagram of conventional pi-matching network



UHF Multiplying Channel

The uhf channel was composed of a combination transistor-varactor multiplying chain which provided a multiplication factor of twelve. The oscillator frequency was tripled to 100.425 Mc and then doubled to 200.85 Mc. Two transistor amplifier stages were used to raise the 200.85-Mc signal to a level sufficiently high to drive a varactor multiplier stage. In this stage the 200.85-Mc signal was doubled to 401.7 Mc.

A common-emitter circuit configuration was utilized with the 2N3118 frequency-tripler stage as well as the 2N3137 frequency-doubler and amplifier stages. A capacitive transformer-type impedance-matching network was used to provide ac coupling between each of these stages. The second 200-Mc power amplifier utilized a 2N3375 transistor in a common-base configuration with a conventional pi-network coupling to the varactor doubler stage. The choice of transistor, circuit configuration, and impedance-matching network was made on the same basis as that of the vhf channel. The power losses due to frequency multiplication combined with the power gain of the amplifier stages resulted in a signal level of approximately 800 mw to the input of the varactor doubler stage.

Several factors must be considered in the design of a varactor multiplier stage. For this particular application, the varactor must be capable of efficient operation with relatively low input power levels, in the order of hundreds of milliwatts, and at a high frequency which dictates a low spreading resistance and a low capacitance. Ideally, a shunt-type configuration generates only the second harmonic, while the series-type circuit generates all harmonics. As a result, filter requirements for the shunt-type circuit are less stringent, and power losses in the voltage variable capacitor are smaller; consequently, the efficiency is higher [7]. Hence, the shunt configuration was selected for use in the actual circuit design. The preliminary circuit values for the varactor multiplier stage were obtained through the solution of equations presented in Luettgenau's analysis of varactor diode frequency multipliers [7]. These equations are presented in the following paragraph, together with a sample computation for the actual circuit.

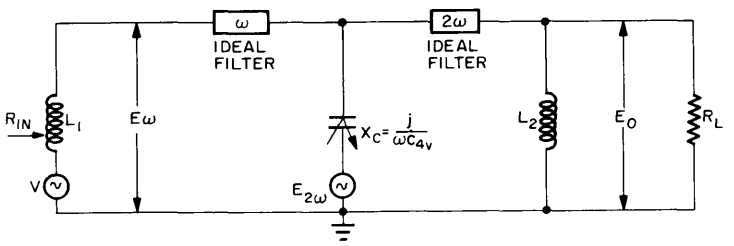


Fig. 5 - Schematic diagram of ideal varactor multiplier circuit

The circuit of Fig. 5 will be used to demonstrate the design of the varactor doubler stage. According to Luettgenau, the power output is equal to the power input, provided that lossless frequency conversion is assumed. That is:

$$P_{out} = P_{in} = \frac{V_B^{3/2}\,\omega\,C_{4V}}{16\sqrt{2}} \tag{10}$$

where V_B is the avalanche breakdown voltage, ω is the angular frequency, and C_{4V} is the small-signal diode capacitance at 4 volts negative bias. The value of output load impedance, R_L, to which the varactor diode should be matched is:

$$R_L = \frac{V_B^{1/2}}{2\sqrt{2}\,C_{4V}\,\omega}\,. \tag{11}$$

The power losses in the shunt configuration of the varactor diode are due to the flow of fundamental and harmonic currents through the high-frequency equivalent resistance R_s of the diode. This power loss is given by:

$$P_{Loss} = \frac{6.36\,V_B\,\omega^2\,C_{4V} \times 10^{-9}}{Q}\,. \tag{12}$$

The circuit efficiency is:

$$\eta = \left(1 - \frac{1.44 \times 10^{-7}\,\omega}{V_B^{1/2}\,Q}\right) \times 100\%\,. \tag{13}$$

The PSI High Q Varicap PC-115-10 was selected for use in the varactor multiplier circuit, since it fulfilled the frequency and power requirements set forth in the preceding paragraph. The ideal filters (Fig. 5) were replaced with high Q series-resonant circuits (Fig. 6), which were tuned to 200.85 Mc and 401.7 Mc respectively. The PC-115 varactor diode has the following parameters:

V_B = 106 volts
C_{4V} = 10 pf at 4 vdc bias
Q = 100 (measured with 4 v bias at 50 Mc)

The preliminary values obtained from the design equations are as follows:

P_{in}	$= P_{out}$ = 613 mw	η	= 82.5 percent
R_{in}	$= R_{out}$ = 36 ohms	L_1	= 0.16 μH
P_{Loss}	= 107 mw	L_2	= 0.03 μH

$\omega = 12.61 \times 10^8$ cps

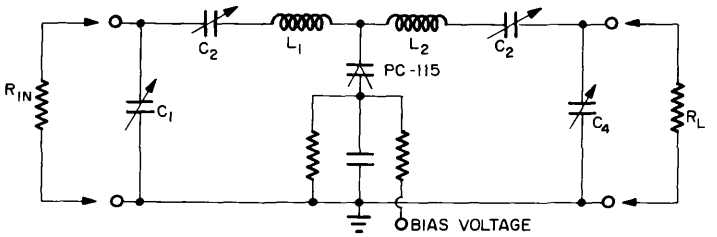


Fig. 6 - Schematic diagram of actual varactor multiplier circuit

Because of the nonlinearity of a varactor multiplier circuit, the exact value of input and output resistances to be matched is unknown. Therefore, the input and output terminals were shunted with variable capacitive reactances, C_1 and C_4 of Fig. 6, to match the impedance of the diode. In this manner the reactances could be adjusted over a wide range of impedances to obtain optimum power output and efficiency. A 16-volt fixed bias on the diode was dictated by temperature considerations. Temperature curves of the PC-115 showed that the diode capacity remained relatively constant from -40° to 140°C for this bias value, which is an important feature in this circuit, since frequency change with temperature must be avoided. The bias voltage was obtained from the transistor collector supply through a voltage-divider network. The efficiency of the varactor doubler circuit is in the order of 70 percent.

The final output-power stage, utilizing a type 2N3375 transistor in common-base configuration, supplied rf energy to the 50-ohm load through a pi-matching network.

Power Supply

The primary dc power supplied to the Model NT-1 Transmitter unit aboard the Gemini spacecraft may vary over a range of 20 to 30 volts, with transients of ±100 volts, and with the negative dc return ungrounded. It was desired that the collector supply voltage of the transmitter be regulated at 24 volts and returned to ground at the transmitter chassis (Fig. 7).

The ungrounded primary voltages were filtered through high-Q toroid networks, to suppress the voltage transients, and fed to a voltage regulator utilizing 2N699 and 2N1725 transistors. The 18-volt regulated output was applied to a common-emitter, transformer-coupled oscillator which utilized two 2N1725 transistors in a push-pull configuration. The frequency of the oscillator, 1000 cps, was selected as the most efficient rate for the type of core used in the transformer. The starting circuit consisted of a low-impedance bleeder network which biased the 2N1725 transistors into conduction before oscillations began. The voltage-divider values were selected to insure starting at full load and under low-temperature conditions. This circuit was feasible because of the relatively low power output and, therefore, low dissipation in the bleeder resistors. The secondary of the transformer was connected to a full-wave bridge rectifier, which consisted of four type 1N1538 diodes, one side grounded.

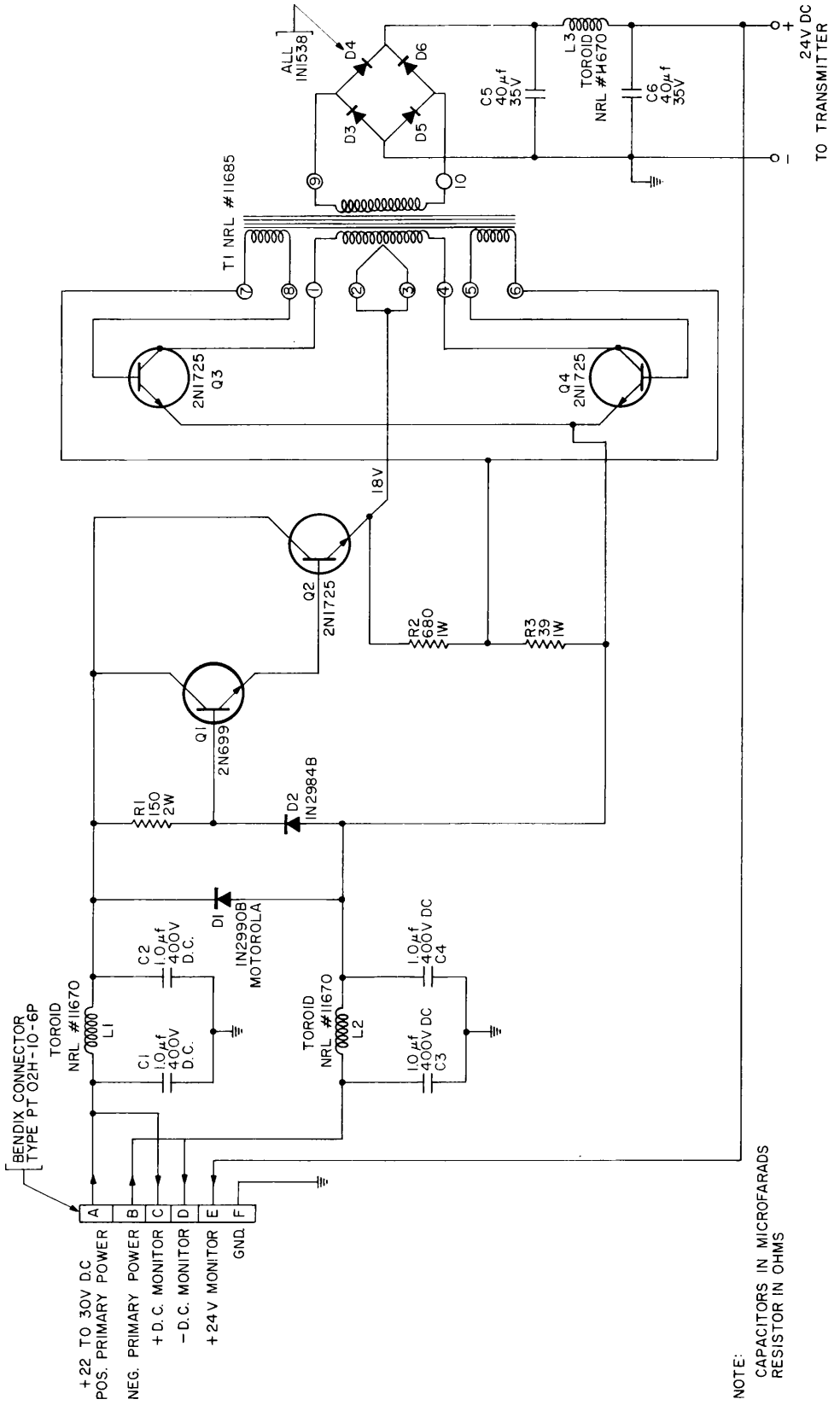


Fig. 7 - Schematic diagram of power supply

ELECTRICAL PERFORMANCE

The final model of the transmitter and power supply was subjected to a series of tests in order to evaluate their electrical performance under the environmental conditions set forth in the preceding list "Transmitter Specifications." The results of frequency and power-output stability tests on the final transmitter model proved to be very satisfactory over the temperature range of -60° to 160°F. During the low-temperature portions of the test procedure, the transmitter unit was mounted on a simulated coldplate, the temperature of which was maintained at 40°F.

The 24-volt regulated dc output from the power supply remained relatively constant within ±0.1 volt over a range of 20 to 32 volts input variation and to within ±0.5 volt over the environmental temperature range. The total current drain of the power supply and rf circuitry was 1.6 amperes, which, at 24 volts, resulted in a power drain of 38.4 watts from the primary dc supply—well under the limit given in the list of specifications.

Problems Encountered

The operation of the vhf channel presented only minor design problems, such as selection of the proper circuit components and circuit configurations which would survive the temperature extremes.

The development of the uhf channel required substantially more attention, particularly with the 200 to 400 Mc doubler circuit. The principal area of difficulty with this circuit was that of obtaining proper operation over the low-temperature regions of -40° to -60°F. Several types of transistors were investigated in the circuit, utilizing both the common-emitter and common-base configurations. Although the 2N3137 and 2N3375 gave satisfactory performance over the higher temperature levels, a consistently unstable condition existed below -20°F which was attributed to a loss in gain and regeneration between the multiplier and 400-Mc driver stages. Experimentation with several different bias and temperature-compensating networks failed to solve the problem. The 2N2616, with its higher cutoff frequency and better gain characteristics, gave the most satisfactory performance over the entire temperature range, although the heat-transfer benefits from a simulated coldplate were necessary at the -60°F level. In order to simulate actual flight conditions, the uhf channel was attached to the coldplate and stabilized at -60°F for a 24-hour period before the collector voltage was applied.

At this point, a rather disturbing observation was made; three minutes elapsed before the channel output level came up to full power, with a snap-on action very similar to that of a locked oscillator. The cause of the delayed channel operation was pinpointed to the final doubler stage, since proper operation of the oscillator and multiplier stages preceding the final doubler was observed at the time the collector voltage was applied. The snap-on action was attributed to a gain reduction which, in turn, resulted in a reduced dc current drawn by the 2N2616 Class C transistor stage. The warm-up time was, therefore, increased because of lower dissipation in the transistor junction. Due to these deficiencies found with transistor multipliers in the 200 to 400 Mc region, the decision was made to change to a varactor type multiplier design. A 200-Mc power amplifier, utilizing a 2N3375, was employed to drive the varactor stage, and the 400-Mc transistor driver stage was eliminated. No changes were necessary to the remaining circuitry.

The most important technique used to achieve stable operation of the uhf channel over the temperature extremes was that of cold-tuning the capacitive reactance matching the varactor impedance to that of the final power-amplifier stage. The overall circuit was first adjusted for optimum performance at room temperature, and then, after the temperature of the circuit had stabilized to the environmental temperature of -35°F, the varactor

output matching reactance was adjusted for maximum power output and a clean spectrum. After this adjustment, the temperature could be lowered to -60°F or raised to 160°F without degrading the electrical performance of the channel.

Several other techniques, evolved from breadboard testing, which contributed to the stability of operation were as follows: (a) temperature compensation of the oscillator stage; (b) referencing the oscillator collector supply to a 20-volt Zener diode to maintain the transistor operating point at a constant level; (c) maintaining the collector supply voltage of the 2N3137 transistors at 20 volts to prevent destructive avalanching, which frequently occurred at 24 volts; and (d) using high-mu ferrite beads to improve decoupling between stages.

Final Achievements

After careful alignment of the final circuitry, power outputs of four watts from the vhf channel and two watts from the uhf channel were measured across a 50-ohm load impedance. The combined rf output level from both channels, measured at the diplexer output, was approximately 5.5 watts. The specification required a long-term power-output stability of +0, -3 db, where long-term stability was defined with respect to the operating time of ten minutes per orbit. Environmental tests indicated that the output power was maintained essentially constant for twice this period of operation and over the required temperature range of -60° to 160°F.

The long-term frequency variation remained well within the ±0.005 percent limits imposed by the Transmitter Specifications. The data for the frequency-stability curves of Fig. 8 were recorded after the unit was energized following a 24-hour stabilization period at each of the temperature extremes. At the high-temperature extreme, 160°F, the frequency of the vhf channel varied less than 25 cps, while that of the uhf channel varied +0, -125 cps. Following the prolonged exposure to the low-temperature extreme of -60°F, the frequency variation was somewhat greater due to transistor warm-up time. As a matter of interest, immediately following the completion of the actual low-temperature tests, the transmitter was permitted to continue operating with the cold-plate turned off. It was found that the vhf channel would operate indefinitely without the benefit of the coldplate, but the output level from the uhf channel began to decrease after a relatively short time. The test was repeated at an environmental temperature of -30°F, and the uhf channel operated steadily. On the other hand, if the uhf channel were already in operation at room temperature, it would continue to perform satisfactorily as the environmental temperature was lowered to and sustained at -60°F.

MECHANICAL DESIGN

The Model NT-1 dual-channel transmitter was constructed in two separate aluminum chassis, each of which was seven inches in length, six inches in width, and two inches in depth, assembled together to form a single unit. When mated together, one inverted on top of the other (Fig. 9), the unit was capable of withstanding the severe environmental requirements imposed by space flight. The upper chassis, or deck, contained the rf circuitry of the two transmitter channels, whereas the lower chassis, or deck, comprised the transmitter power supply.

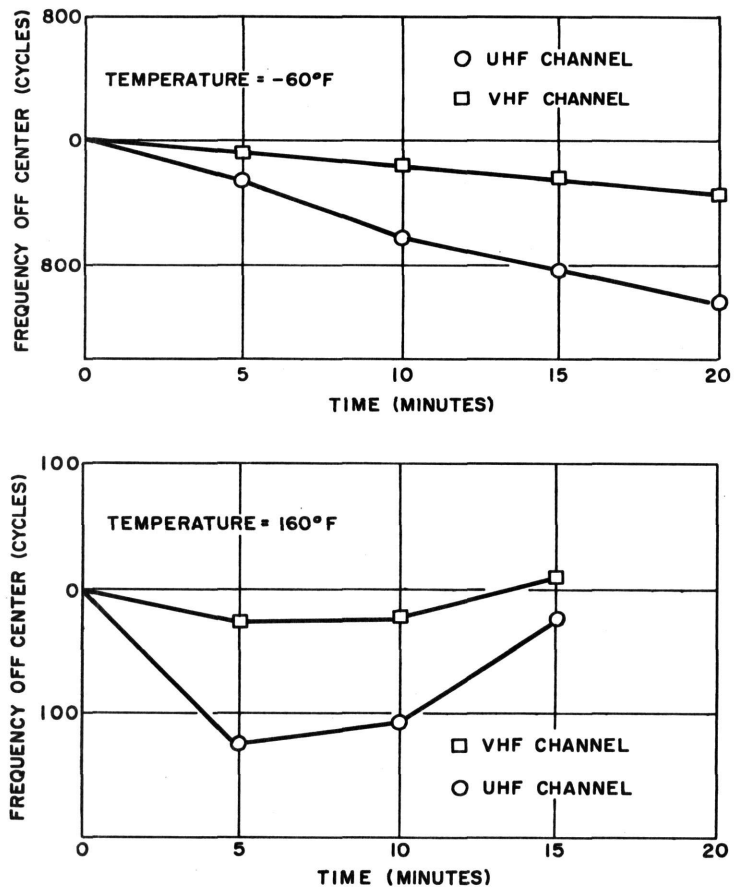


Fig. 8 - Curves showing long-term frequency stability

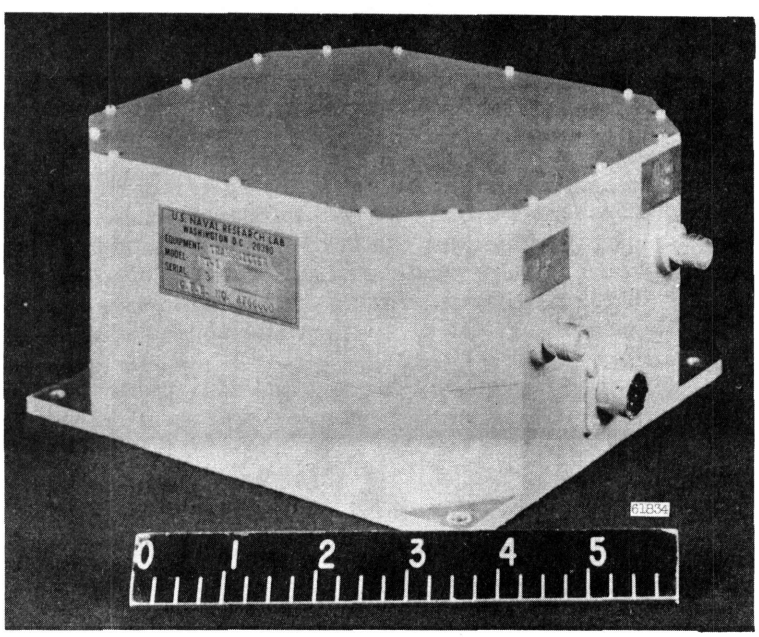


Fig. 9 - Assembled transmitter unit

Packaging

The rf circuitry was arranged in individual compartments milled from solid aluminum stock (Fig. 10), which provided adequate rf stage isolation as well as rugged construction. RF energy was distributed to each compartment through low-loss Teflon feed-through components. A thin aluminum plate was secured in place between the two decks, which completed the shielding for the rf stages as well as for transmitter and power supply. The upper side of the transmitter chassis (Fig. 11) was milled to a depth of 0.625 in. to provide space for the transistor decoupling networks and access to the tuning controls.

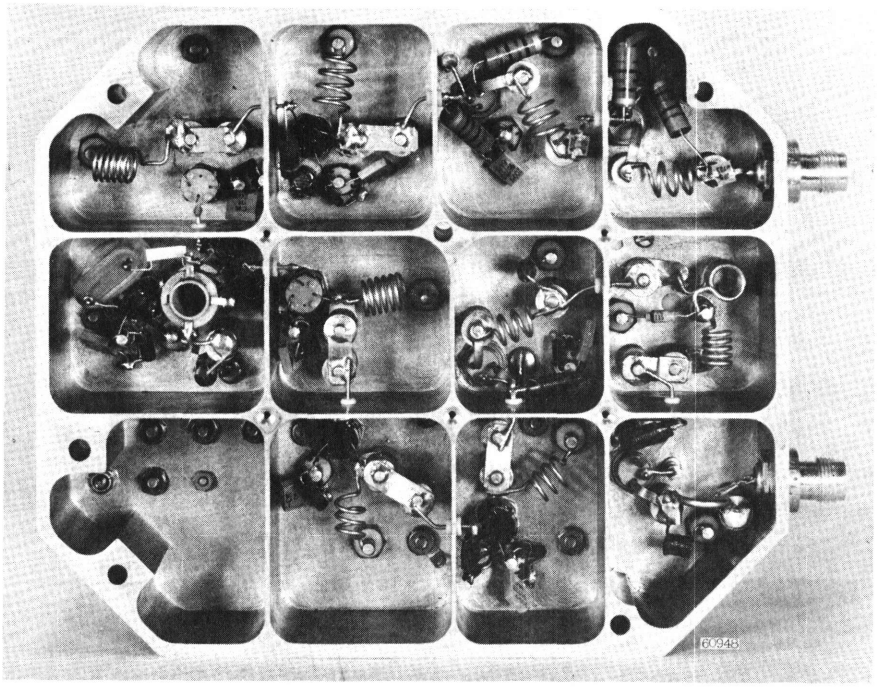


Fig. 10 - RF circuitry deck

The power-supply circuitry was also arranged in a compartment milled from solid aluminum stock (Fig. 12). The bottom plate of this chassis had a thickness of 0.375 in. to provide good heat-sink properties for the large-wattage, stud-type power transistors used in the power supply. The outer surface of the bottom plate was machined to a flatness tolerance of ±0.015 in. in order to provide uniform contact and, hence, good heat exchange with the coldplate on which it will be mounted in the spacecraft.

The two chassis of the transmitter were held together by five stainless steel bolts, each of which was provided with helicoil inserts to prevent thread wear in the aluminum and to provide a locking mechanism for the bolts. In order to facilitate heat transfer between the upper and lower chassis, a layer of silver-loaded epoxy was applied to the inner surfaces between the two decks of the unit. An internal connection between the power-supply chassis and the rf chassis provided voltage to the transistor-collector decoupling networks.

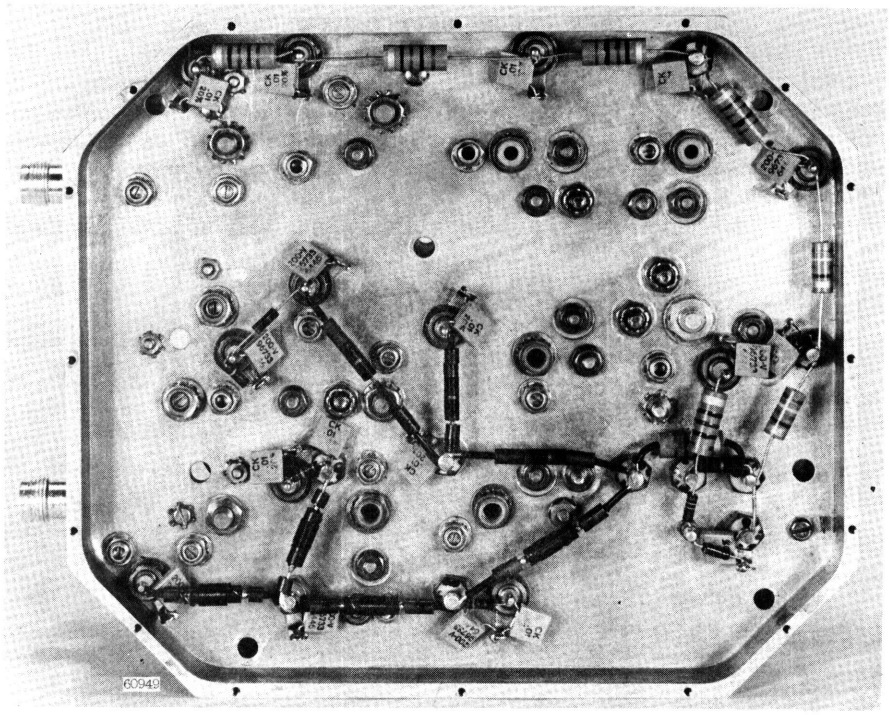


Fig. 11 - RF tuning deck

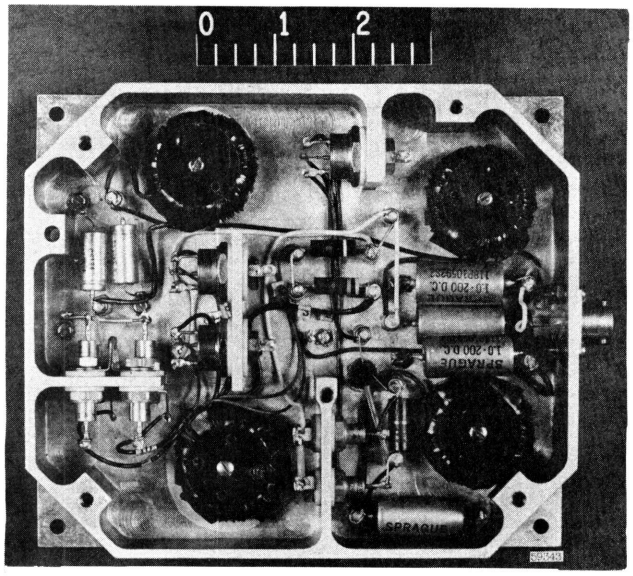
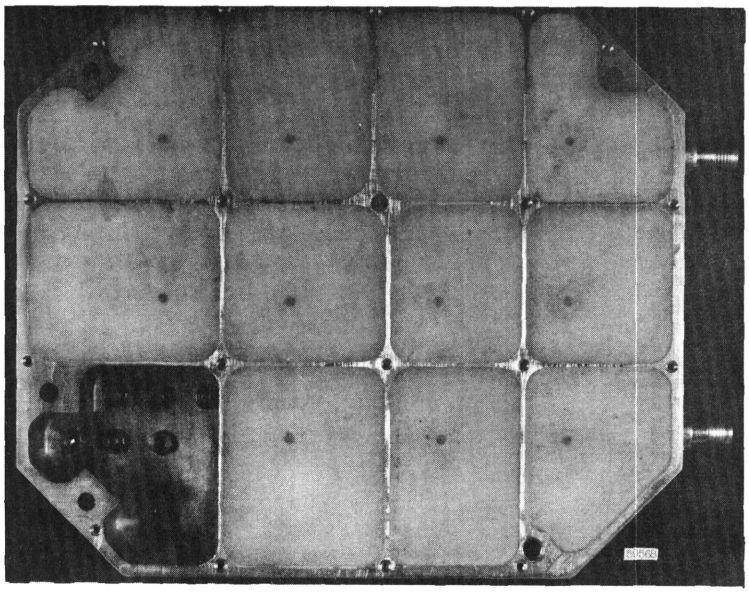


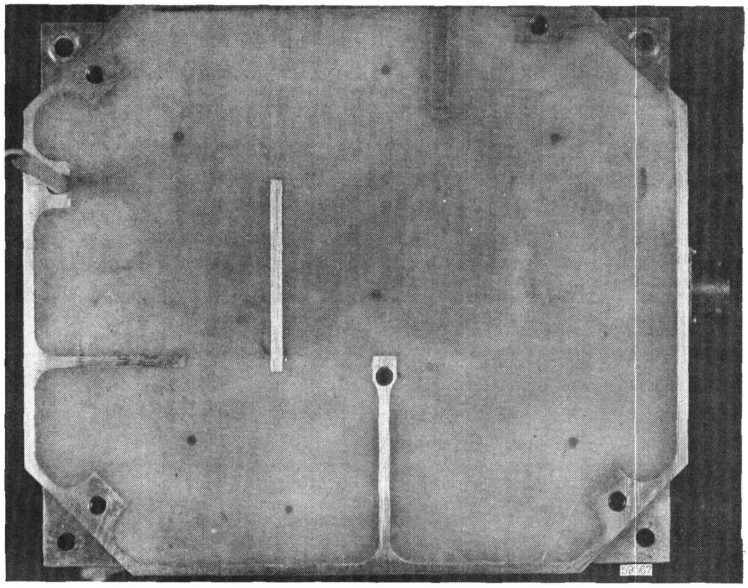
Fig. 12 - Power-supply deck

Ruggedization

Several special features were added to the mechanical fabrication in order to ensure adequate protection from the severe effects of space flight. The interior of each chassis was completely encapsulated with a foam epoxy (Fig. 13), and the decoupling networks were covered with a silicone resin to provide protection against random vibrations and acceleration. As a precaution against outgassing during exposure to low atmospheric pressure, hermetically sealed connectors and JFD Seal-Cap tuning capacitors were used throughout. These features also provided excellent protection against prolonged exposure to high humidity.



a. Power supply



b. RF section

Fig. 13 - Foam-encapsulated chassis

ENVIRONMENTAL TEST RESULTS

During the actual flights of the Gemini 8 and 9 spacecraft the transmitter unit was energized only during that portion of the orbit which was in the line of sight of the ground-based receiving antennas. Hence, the environmental tests simulating the launch phase of the flight were conducted with the transmitter deenergized. After the completion of each test, the unit was energized and measurements of frequency and power output were recorded. In addition, a physical inspection of the chassis was made in order to detect any mechanical failures which might have occurred during the testing procedure. In the case of the high- and low-temperature environmental tests, the unit was stabilized at temperatures of 160° and -60°F for periods of 24 hours before being energized and data recorded.

The transmitter unit was subjected to the random vibration levels given in Fig. 14, which were applied along each of the three principal axes of the unit for a period of 15 minutes per axis. During the orbital phase of the vibration testing, a period of five minutes per axis, the unit remained in continuous operation. The acceleration environmental tests were conducted by means of a centrifuge whose motion was increased in a linear manner to produce forces of from 1 to 7.5 g in a total average time of 250 seconds per transmitter axis. The measurements made following each phase of the above testing procedure revealed no degradation in electrical performance and no mechanical failures of any kind.

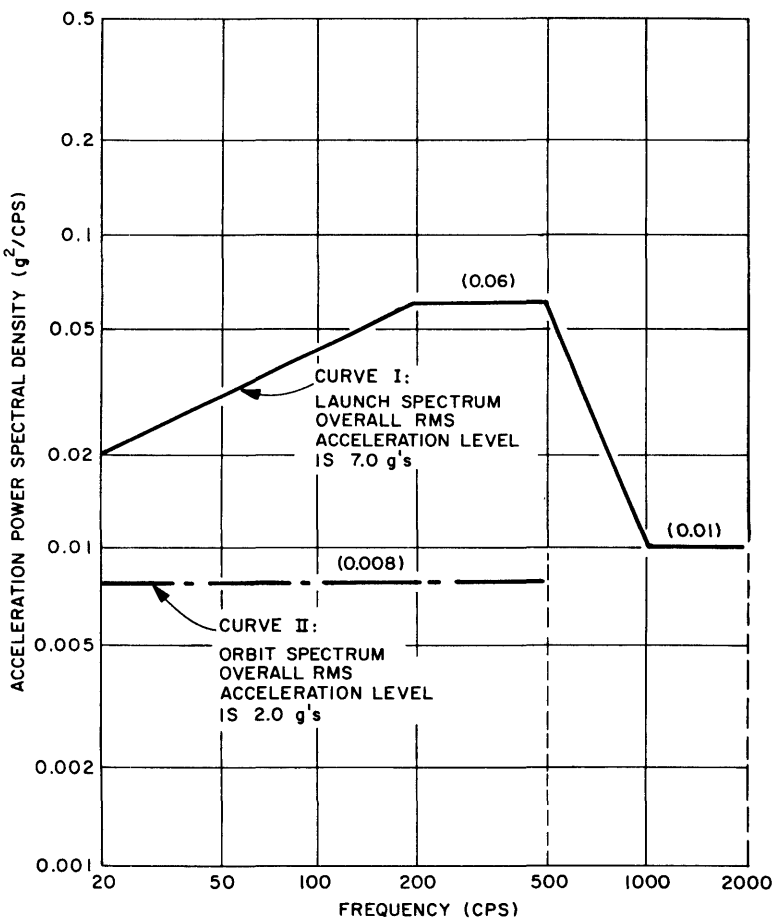


Fig. 14 - Graph of random vibration requirements [1]

The prolonged exposure to the temperature extremes indicated in Transmitter Specifications resulted in no degradation of transmitter performance. Upon energizing the unit, both channels immediately produced the desired output levels. Curves of frequency variation versus temperature (Fig. 15) indicate a frequency change of ±7 kc at 401.7 Mc and ±2 kc at 133.9 Mc over the full temperature cycle, or less than ±0.002 percent.

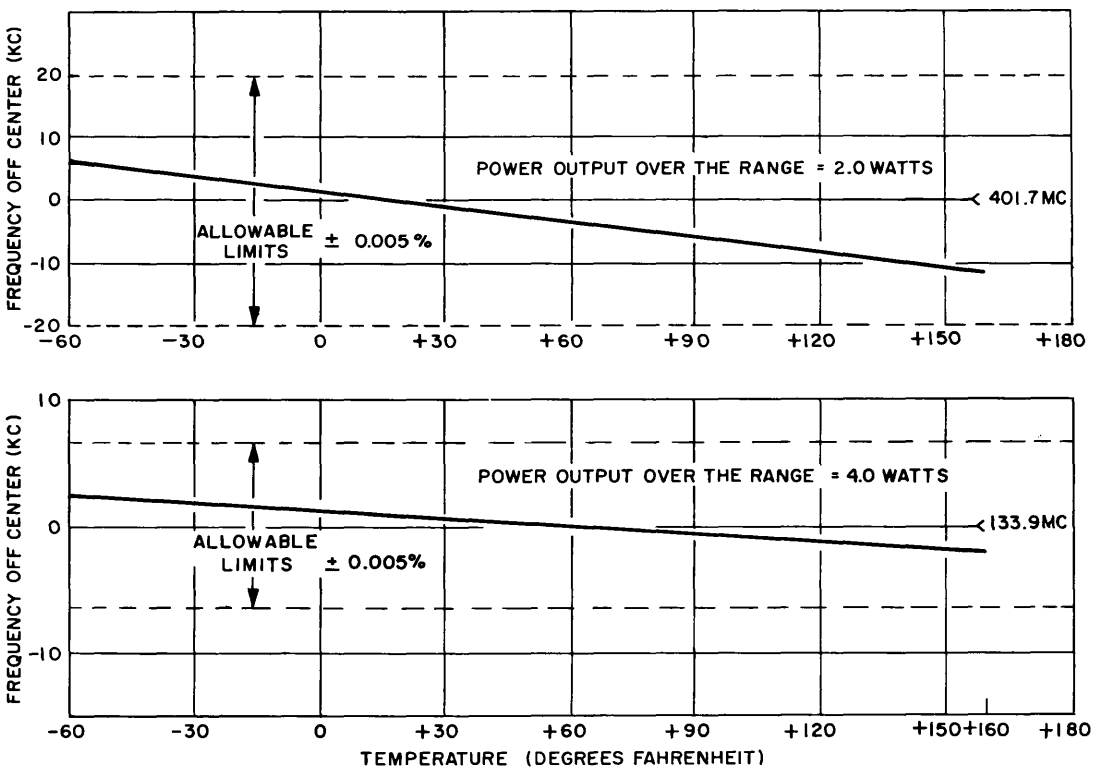


Fig. 15 - Curve of frequency variation with temperature

The acoustical noise environmental test was conducted in three parts, with the octave band sound-pressure levels being directed toward each of the three most sensitive mutually perpendicular axes of the unit for a period of ten minutes per orientation. The most sensitive axes are defined as those having the least amount of external structure between the sensitive items and the noise source. The overall sound-pressure level of the octave band spectral content (Fig. 16) was 156 db. Upon completion of these tests, the transmitter unit operated in a normal manner, and no mechanical failures were noted during the physical inspection.

During the RFI tests, no spurious output signals of a level greater than 80 db below the two fundamental rf outputs reached the antenna. Typical attenuation levels of spurious outputs are shown in Table 1.

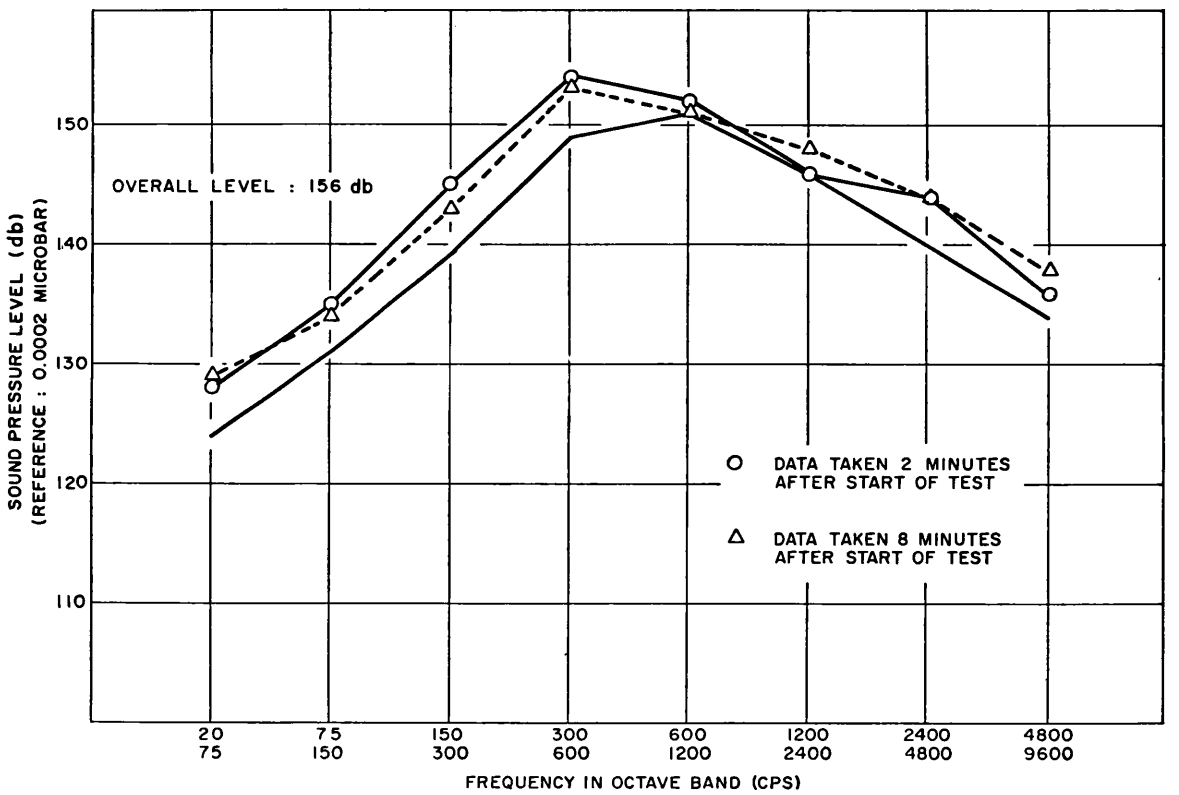


Fig. 16 - Graph of acoustical noise requirements [1]

Table 1 - Attenuation of Spurious Output

Spurious Radiated Frequencies (Mc)	Level Below Principal Frequencies (db)	
	133.9-Mc Channel	401.7-Mc Channel
33.475	107	103
66.95	80	-
100.425	-	85
200.85	-	80

All other RFI tests showed that the radiated and conducted interference levels were below maximum levels set forth in Mil-I-26600. Results of these tests indicated that better RFI shielding should be designed into future equipment, since some of the results were close to tolerance maximums.

The test specification for atmospheric pressure of 10^{-8} psi could not be achieved. A pressure level of 2×10^{-7} psi was the best that the environmental chamber would allow. After a period of 48 hours at this pressure level, the electrical performance was checked and found to be normal. There were some outgassing products, such as water and nitrogen, but these were of negligible amounts.

Humidity tests were conducted on a 24-hour basis for ten days. Physical inspection of the unit upon completion of this test revealed no corrosive effects on any of the connectors or the top and sides of the chassis. The bottom did reveal some minor corrosive action, probably due to scratching of the iridited surface from too much handling. The unit was put into operation within one-half hour of completion of the test and performed in a normal manner.

CONCLUSIONS

The qualification model has fulfilled all of the specifications listed in Transmitter Specifications. The major problems offered by the environmental specifications, such as small physical dimensions, vibration, low atmospheric pressure, and temperature extremes have been overcome with relatively little sacrifice in electrical performance. Through the experience gained during the development of the dual-channel transmitter, and recent high-frequency transistor development, several techniques for the improvement of the electrical and mechanical design are indicated.

1. Replace the T0-5 type transistors with the recently developed stud-mounted transistors, which have better high-frequency gain and power output for lower input power levels. This step should also improve heat dissipation.

2. More efficient harmonic-rejection circuitry would be designed into the unit if, at some future time, separate antennas were to be used for each frequency.

3. Closer attention must be paid to the RFI problem. RF shielding gaskets could be used at the coaxial connectors and inside the chassis divider plate and top cover plate.

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13. ABSTRACT

The objective of Gemini experiment D-14 was to measure the electron content of the ionosphere below the Gemini spacecraft by means of a dual-frequency Faraday rotation experiment utilizing two spacecraft-borne transmitters operating in the vhf-uhf bands. The Model NT-1 dual-channel solid-state transmitter was designed for this experiment. The transmitter produces, simultaneously, four watts of rf energy at 133.9 Mc and two watts at 401.7 Mc. This transmitter is a highly ruggedized unit capable of operation under the severe vibration and temperature requirements of spacecraft environment.

The spacecraft-borne transmitters supplied at least one watt of cw rf energy at each frequency for simultaneous radiation from a colinear dipole antenna. This power output was necessary in order to obtain the desired signal-to-noise ratios at the ground-based receiving stations for accurate measurements.

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Gemini						