Microwave Generator

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The TD-3 microwave generator furnishes all of the local oscillator power used in a TD-3 transmitter-receiver bay. It delivers an output power of 0.5W at a frequency which can be adjusted to any one of 17 frequencies. In designing the generator special emphasis was placed on reliability, ease of tuning, and on obtaining a carrier with high spectral purity and frequency stability. This paper describes the generator's circuit configuration and operation, and discusses important design considerations and techniques.

I. INTRODUCTION

The microwave generator is a highly frequency-stable, low-noise source of microwave power which is the local oscillator supply for the TD-3 microwave transmitters and receivers. It delivers 0.5W output power and is adjustable to any one of 17 different frequencies: 3780, 3800 . . . 4100 MHz.*

To achieve great reliability, minimum maintenance, and low power consumption, it was decided to use only solid state devices. Various devices were considered: varactors, step recovery diodes, and transistors; however, at design time, varactors were the only solid state devices that appeared able to operate at the required frequencies and provide the power levels and reliability demanded.

Once varactors were chosen it was possible to determine the over-all circuit configuration. A crystal oscillator operating in the range of 125 MHz† was needed to obtain the required frequency stability. Then a chain, consisting of a cascade of several varactor multipliers, was used to obtain efficient frequency multiplication up to 4000 MHz. A power amplifier had to be inserted between the oscillator and the

* This equipment is manufactured for Bell System use only.

† In general 125, 500, 1000, and 4000 MHz are used to denote any frequency between 118.125 and 128.125 MHz and their fourth, eighth, or thirty-second multiple, respectively.

multiplier chain to drive the passive, lossy multipliers with sufficient input power to obtain the desired output power at 4000 MHz.

A significant aspect in the development of this generator was the need for a compromise between the usual need for efficiently obtaining a high level carrier, and the need to generate a carrier of high spectral purity; that is, with extremely low noise sidebands. These two requirements led to a rather extensive use of experimental design techniques, because available theory precluded exact design procedures, particularly with the multiplier chain. Instead, the design was based on simple principles which were experimentally refined. For example, even the expected efficiencies of the multipliers could not easily be calculated because the required power levels necessitate driving the varactors into the forward charge storage region, the voltage-charge relationships of the varactors do not follow a simple law, and the series resistance of the varactors is a complicated nonlinear function of charge.[1] In addition, a noise model was not obtainable. Therefore, although it would have been possible to extend Burckhardt's[2] work regarding efficiencies to this situation, an experimental approach was deemed the most satisfactory solution.

II. DESIGN OBJECTIVES

Since the microwave generator furnishes all of the local oscillator microwave power to operate a TD-3 repeater, and since failure would cause loss of a radio channel, high reliability is a prime objective. It is attained by using solid state devices which should insure a generator failure rate of less than 8.7×10^{-3} failures per repeater year.

All performance objectives discussed in this section should be met over the normal operating temperature range of 75 ± 10°F. However, the generator must continue to operate (possibly with degraded performance) from 40 to 140°F to assure system operation in case the air-conditioning malfunctions.

The other performance objectives are:

Output Power: 26.5 dBm ±1 dB.

Frequency: The generator should be able to be tuned to each of the 17 frequencies (3780, 3800 ··· 4100 MHz) by choosing the proper crystal and by tuning the oscillator, the amplifier, and the multiplier chain.

Frequency Stability: The inherent stability should be ±1 part in 10^6 between maintenance intervals.

Noise: The noise objective was set to provide less than 1 dB contribution from the microwave generator to the repeater noise figure. This objective, expressed in carrier-to-noise-per-cycle ratio, is shown on Fig. 1 as a function of frequency increment Δf away from f_c.*

DC Requirements: The generator should be designed to operate from the -19V repeater power supply.

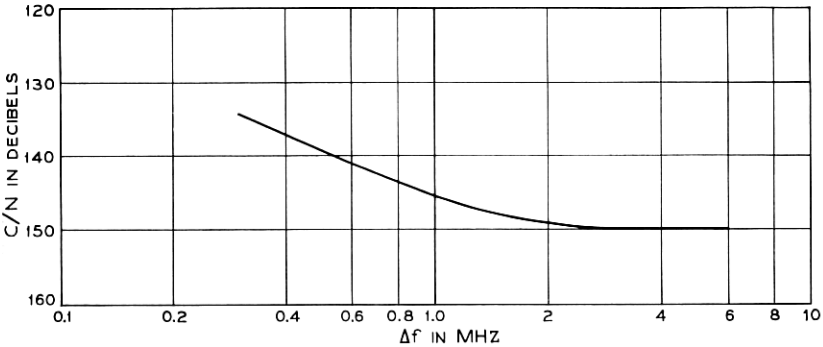

Fig. 1 — C/N objective for the output carrier. C = carrier output power in dBm; N = total noise power in a 1 Hz band in dBm; C/N in dB = $C - N$; Δf = frequency increment away from f_c; f_c = any one of the 17 output frequencies (3780, 3800 . . . 4100 MHz).

Miscellaneous: **Instant start:** The generator should operate on the application of dc power; that is, with a properly tuned generator, removal and restoral of dc power should not require retuning of any circuits to obtain the original output. **Monitor points:** Test points should connect to the main metering panel (20 μA, 5900Ω meter resistance) in order to monitor the condition of the generator without removing it from service. There should be a frequency monitoring point. **Output port:** The output port should be a coaxial N-type connector.

All of these objectives were met by the final design. The performance characteristics are discussed in detail in Section V.

III. GENERAL DESIGN

3.1 *Block Diagram*

Figure 2 is a block diagram of the microwave generator. It consists of two major parts: an oscillator-amplifier and a frequency multiplier chain. The oscillator-amplifier consists of a crystal oscillator and a power amplifier, and the multiplier chain consists of a first quadrupler,

* f_c is any one of the 17 carrier frequencies just mentioned.

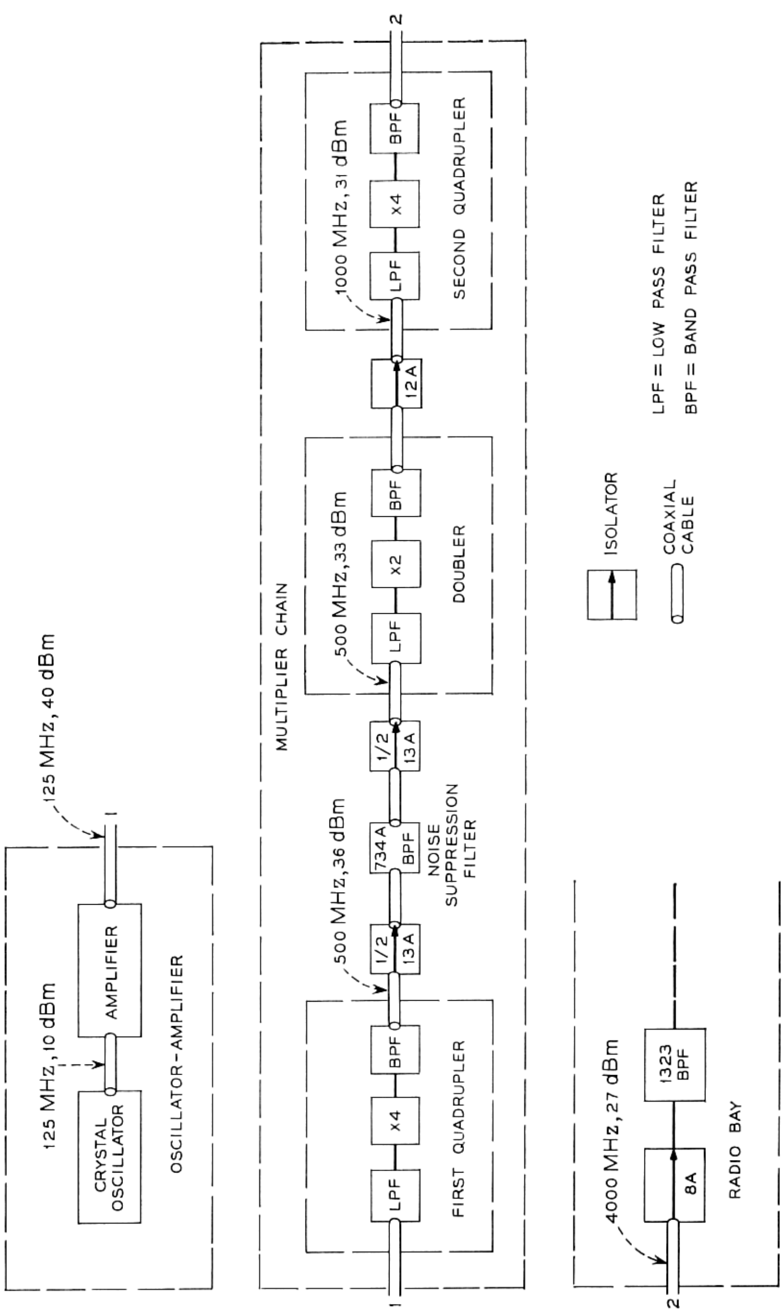

Fig. 2 — Block diagram of the microwave generator.

a noise suppression filter, a doubler and a second quadrupler. To facilitate development and testing, all of the units are designed to operate between 50Ω impedances.

The oscillator is a crystal-controlled transistor oscillator which operates between 118.125 and 128.125 MHz by choice of crystal and proper tuning. It delivers at least 10 dBm of power to the amplifier whose input impedance is 50Ω ($\geq$20 dB return loss). There is a point for monitoring the transistor's emitter current.

The power amplifier has four transistor stages: two class A stages and two class C stages. It is tunable over the same frequency range as the oscillator and can deliver at least 40 dBm of power to the first quadrupler. The output power can be varied from 38.5 to 40 dBm by adjusting an interstage network. This range of output levels is required since the generator must have an output power of about 26.5 dBm and the multiplier chain exhibits a loss ranging from 12 to 13.5 dB. There are points for monitoring frequency, output power level, and individual transistor currents.

The first quadrupler and the doubler are lumped element varactor multipliers; the second quadrupler is a distributed element varactor multiplier. All multipliers are self-biased and are shunt multipliers in order to facilitate better heat sinking. The output level of the chain is 26.5 dBm with input levels ranging from 38.5 to 40 dBm.

A waveguide bandpass filter is used at the output of the chain to eliminate spurious tones at frequencies ± 125, ± 250 ± 375 $\cdots$ MHz removed from the desired 4000 MHz output. This filter is located in the radio bay (see Fig. 2).

3.2 *Physical Characteristics*

Figure 3 shows the generator, which measures 21 by 11 by 7 inches. Figure 4 shows the oscillator-amplifier part, and Fig. 5 shows the three multipliers. As Fig. 2 indicates, the components are connected by coaxial cable. The use of this flexible 50Ω cable (RG214/U or RG98/U), simplifies equipment layout and permits the easy disconnection needed for tuning the chain.

3.3 *Multiplier Chain Design*

3.3.1 *General*

For the general layout of the multiplier chain, three empirical observations were most important. They concerned interaction, noise, and impedance.

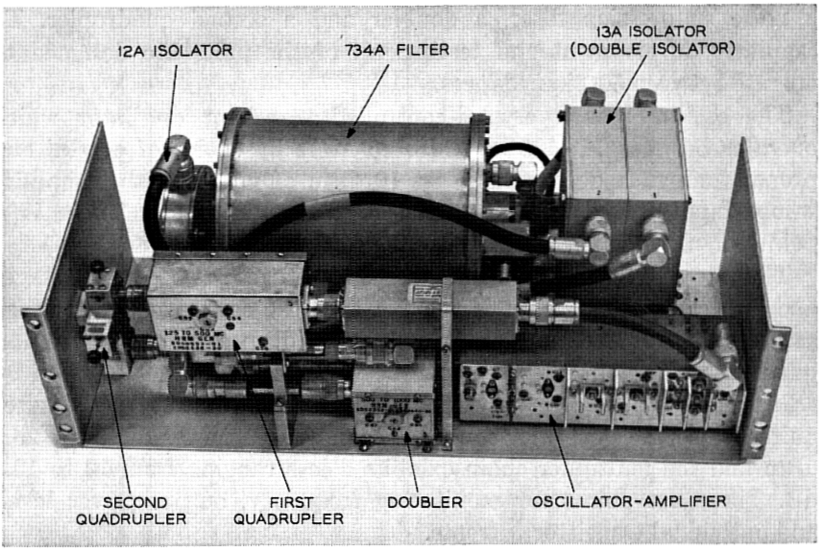

Fig. 3 — Complete generator.

It was found that in cascading multipliers it is important to prevent interaction among the multipliers at undesired harmonic frequencies generated in the chain. Therefore, each multiplier was imbedded between filters which pass only the desired frequency at the input and the output.

It was verified that all multipliers behave like ideal multipliers of a multiplication factor m. Therefore, they amplify the FM correlated part of the noise power spectrum of the input carrier by a factor, expressed in dB, of 20 log m.[3] However, an additional noise component which existed even when the input carrier was essentially free of noise was observed in the noise power spectrum of the output carrier. It was concluded that this additional noise component originates in the multiplier.

When incorporating a properly operating multiplier in a chain it was found that, in order to keep the FM correlated noise originating in the multiplier consistently low and to ensure, when tuning the multiplier in the chain, that this condition remains coincident with that of maximum carrier output power, the multiplier must be imbedded between operating impedances which are not reactive at the carrier frequency or at frequencies close enough so that the bandwidths of the impedances facing the multiplier diode are appreciably reduced.

Therefore, a multiplier is not suitable as a load impedance for another multiplier, since it is reactive at the carrier frequency as long as it does not operate and could, therefore, cause starting and tuning problems. A narrow band filter is also not suitable either at the input or output of a multiplier. When placed at the output the effective bandwidth of the multiplier output circuit becomes much narrower than that of its input circuit. As Ref. 4 leads one to expect, this was found to produce instabilities even under perfectly tuned conditions. When placed at the input of a multiplier, a narrow band filter causes instabilities when the multiplier or the filter are being tuned, or when temperature and humidity changes cause the filter to become slightly detuned during operation.

The observation of multiplier interaction led to the inclusion of a low pass filter at the input of each multiplier and a band pass filter at the output of each multiplier (see Fig. 2). The filters are directly connected to the multipliers to avoid long line effects at the rejected

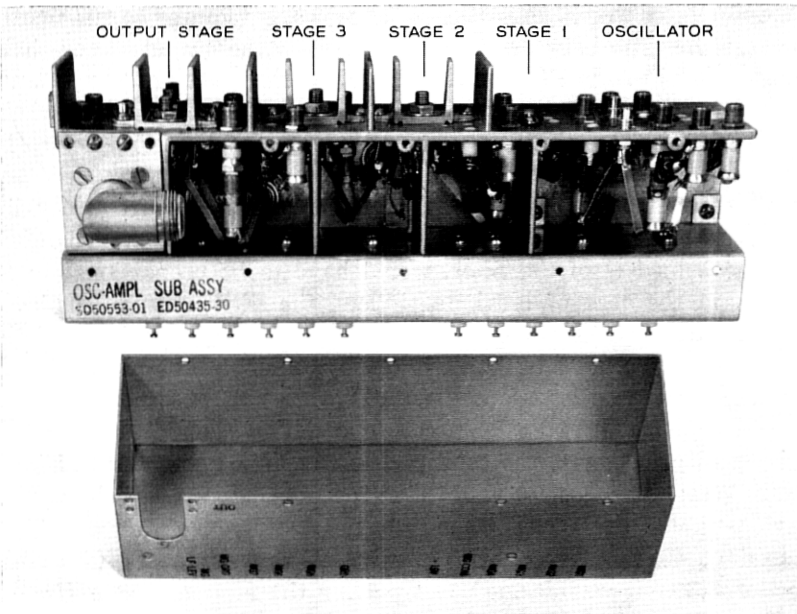

Fig. 4 — Oscillator-amplifier.

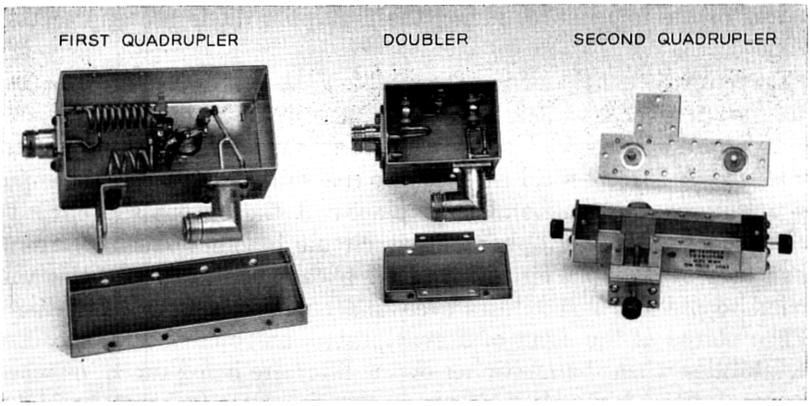

Fig. 5 — The three frequency multipliers.

frequencies. All are commercially available filters designed to operate between 50Ω impedances.

The internally generated noise determined the optimum placement of a noise suppression filter within the chain. This filter is a narrow bandpass filter which, as Section 3.3.2 shows, is required to sufficiently increase the ratio between the carrier level at the output of the chain and the level of the FM correlated part of the noise power spectrum around the carrier at the output of the chain. If the multiplier chain consisted of ideal multipliers which only amplified the input noise, the filter could be placed at 125, 500, 1000, or 4000 MHz, provided that in all cases it furnished the same discrimination at a frequency increment Δf away from the respective carrier frequency f_c. Therefore Q_L, the loaded Q of the filter, would have to be proportional to f_c. Hence, in the case of ideal multipliers, the filter should be placed at 125 MHz because: (*i*) assuming that Q_I, the intrinsic Q of the resonator, is proportional to $1/(f)^{1/2}$, the midband insertion loss of the filter would be lowest, and (*ii*) the variation in insertion loss at f_c resulting from a given temperature variation would be smallest.

In reality, however, the multiplier chain does not consist of ideal multipliers; some noise originates in the multipliers themselves. It is obvious that if this contribution were dominating, the filter should be placed at 4000 MHz. After a number of measurements (described Section 3.3.2) it was decided that placing the filter at 500 MHz was the best compromise for this chain.

The impedance phenomena led to the inclusion of three isolators (realized by terminated circulators). One each was put at the input and the output of the noise suppression filter, and one between the doubler and the second quadrupler. An additional isolator is actually connected to the chain at its output in the radio bay to guarantee the proper load impedance for the second quadrupler (see Fig. 2). It was found that an isolator is not required between the amplifier and the first quadrupler (in accordance with Ref. 4).

3.3.2 *Noise Suppression Filter*

Starting with the C/N objective (Fig. 1) at the generator output, the maximum permissible noise levels N_{IN} at the various multiplier inputs and at the input to the amplifier can be calculated under the assumption that in each case all transmission components between the respective input and the output of the generator contain no noise sources. This has been done in Fig. 6 for $\Delta f = 2$ MHz using measured values for the carrier levels and the C/N degradations of the multipliers and amplifier.

Notice that the measured C/N degradations are 13 dB for the quadruplers and 7 dB for the doubler; that is, 1 dB higher than the theoretical value of 6 dB per frequency doubling. The increment frequency $\Delta f = 2$ MHz was chosen for the calculations in Fig. 6 because (*i*) the C/N at the output of the oscillator, as pointed out in Section 5.2.3, is essentially constant in the band from 0.3 to 10 MHz, and (*ii*) the C/N objective (Fig. 1) decreases at approximately 6 dB per octave for frequencies below 2 MHz. Consequently, a chain containing a one-cavity noise suppression filter which meets the C/N objective at $\Delta f = 2$ MHz will meet the C/N objective for all Δf.

If follows from Fig. 6 that very severe requirements would have to be placed on the individual units. For example, the oscillator must

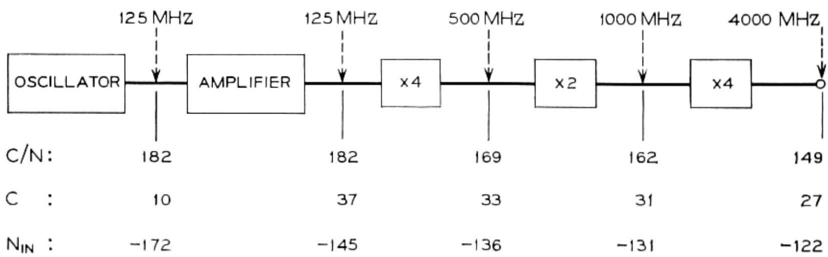

Fig. 6 — Maximum permissible N at the input of the various multipliers and the amplifier (without noise suppression filter). C/N = maximum permissible C/N in dB; C = carrier level in dBm; N_{IN} = maximum permissible N in dBm. Carrier levels and C/N degradations are based on measurements. $\Delta f = 2$ MHz.

have noise sidebands which are only 2 dB above thermal noise (−174 dBm) and the amplifier and multipliers must have no internal noise sources. It was therefore concluded that a noise suppression filter is required. Although it is generally most desirable to place the noise suppression filter at the lowest frequency point, that is, 125 MHz, it is not possible because of the internal multiplier noise.

The noise contributed by the multipliers ranges from 25 to 35 dB above thermal noise when referred to the input of the respective multiplier. These noise contributions were calculated from measurements made on the multipliers with calibrated noise sources and filters of known characteristics at the input of each multiplier stage (125, 500, and 1000 MHz). Also, the noise contributed by the amplifier is 11 dB above thermal noise.

Figure 7 shows the effect of these internal noise sources separately and combined. It can be seen that filtering the noise at the 125 MHz point would not improve the output C/N because the −139 dBm of internal noise contribution of the first quadrupler alone already results in −130 dBm at the input of the doubler which is 6 dB more than the −136 dBm requirement at this point. It can also be seen that the internal noise contributed by the doubler is sufficiently below that required at the doubler input to meet the output objective, and filtering would be effective at this point. Although the filtering could be done at higher frequencies, it was decided to place the filter at the 500 MHz point because of the arguments presented in Section 3.3.1.

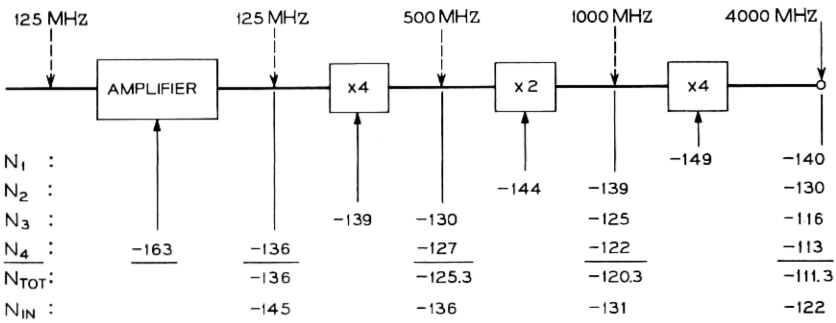

Fig. 7 — Effect of internal noise contributions. N in dBm contributed by the internal noise of: N_1, the second quadrupler; N_2, the doubler; N_3, the first quadrupler; N_4, the amplifier. N_{TOT} is the total N contributed by the internal noise sources, and N_{IN} is the maximum permissible noise in dBm. C/N degradations are based on measurements. $\Delta f = 2$ MHz.

Regarding the filter design, it was decided that a temperature compensated coaxial resonator could be developed which would have a 3 dB bandwidth of 300 kHz. This filter would provide about 20 dB of loss at $\Delta f = 2$ MHz.

Figure 8 shows the maximum permissible noise levels at the input of each of the units with the noise suppression filter located at 500 MHz. It can be seen that the N_{IN} for the amplifier and first quadrupler inputs are now much more reasonable than those in Fig. 6.

3.4 *Generator Tuning*

The entire generator can be tuned to any one of the 17 different frequencies by selecting the appropriate crystal and tuning the various resonant circuits. The major criterion for correct tuning is maximum power output. That is, the oscillator resonant circuits are tuned for maximum power into a 50Ω load.

For tuning the amplifier, swept frequency techniques have proven to be very useful. Therefore maximum power output coupled with a smooth resonant response is the criterion for proper tuning. The input return loss of the amplifier is maximized at the frequency of interest and the interstage tuning controls are then adjusted for smooth response and maximum power at the desired frequency.

Swept tuning techniques were also used in the development of the multiplier chain. However, once the final circuit configuration was obtained, the adjustment for maximum power resulted in the proper response. The multiplier chain is tuned successively for the first quadrupler, the noise suppression filter, the doubler, and the last quadrupler.

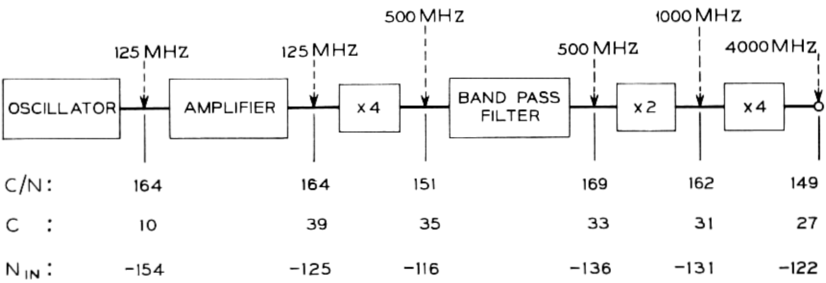

Fig. 8 — Maximum permissible N at the input of the various multipliers and the amplifier (with noise suppression filter). C/N = maximum permissible C/N in dB; C = carrier level in dBm; N_{in} = maximum permissible N in dBm. Carrier levels and C/N degradations are based on measurements. $\Delta f = 2$ MHz.

IV. DETAILED ELECTRICAL DESIGN

4.1 *Oscillator*

The oscillator shown schematically in Fig. 9 is a modified Butler circuit[5] which was suggested by W. L. Smith of the Allentown Laboratory. The transistor is a Western Electric Company 45A transistor which is described elsewhere in the issue.[6] The transistor operates as an amplifier which drives a tuned collector load comprised of L3 in parallel with the total capacitive reactance seen by the collector. Part of this reactance, that from C2 and C3, is used to derive the feedback path which permits energy in the collector circuit to be fed back to the emitter through a series resonant crystal. The crystal is a Western Electric Co. 108 AB crystal which operates in the thickness mode at its fifth overtone. Another part of the collector reactance, from C4 and C5, is used to match the collector impedance to 50 ohms at the output. The resonant circuit (L1, C1) connected to the base is used as a fine tuning adjustment to pull the crystal frequency slightly. Inductor L2 is used to tune out the parasitic capacitance of the crystal. The necessary biasing resistors are not shown.

The oscillator delivers 10 dBm at its output. In order to achieve the ±1 ppm frequency stability required by the generator only 2 mW of power can be dissipated in the crystal; hence the maximum output power was limited to 10 mW (10 dBm). If it were not for this requirement, the output power of the oscillator could have been increased, which, as experiments indicated, would have resulted in a better carrier to noise ratio.

4.2 *Amplifier*

The amplifier, shown schematically in Fig. 10, consists of two class A stages followed by two class C stages. The amplifier gain is

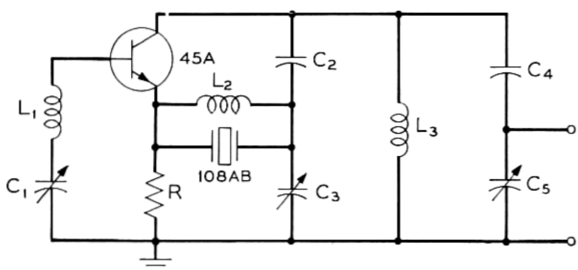

Fig. 9 — Oscillator circuit.

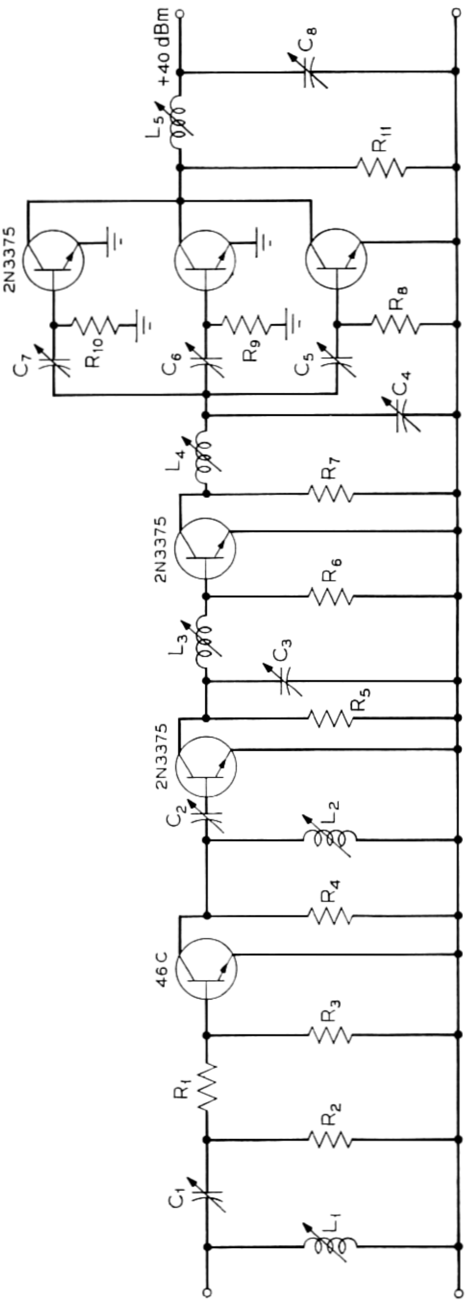

Fig. 10 — Amplifier circuit.

30 dB; that is, it can deliver 40 dBm into a 50Ω load when driven with a power of 10 dBm. It has an input return loss greater than 20 dB referred to 50Ω.

The first stage of the amplifier has an inherent gain of about 13 dB, but because of a lossy input network the net gain is about 10 dB and the output power is approximately 20 dBm. The second stage provides about 8 dB of gain and provides a level of 28 dBm to the input of the first class C stage. The first class C stage provides 6 dB of gain, thus delivering 34 dBm to the composite output stage. The output stage provides another 6 dB of gain which permits the amplifier to deliver 40 dBm to the multiplier chain.

The combined amplifier and oscillator operates at approximately 35 to 40 per cent efficiency; that is, it requires 25 to 30 W dc input power to obtain 10 W of 125 MHz power at the output.

The oscillator requires the amplifier to have a constant resistive input impedance in order to avoid interaction effects which generate noise, produce instabilities, and result in starting problems. It was assumed, and experimentally verified, that an input return loss of the amplifier of greater than 20 dB would be sufficient to provide satisfactory operation.

Since noise performance is very important, it was decided to operate the first stage at as low a noise figure and as high a gain as practicable. Common emitter biaising of a Western Electric Co. 46C transistor was chosen. The input impedance problem then consisted of matching the 50Ω oscillator output into the very low impedance of the first amplifier stage.

Because the input impedance of this stage depends considerably on transistor parameters, second stage input impedance, and tuning frequency, it was decided to buffer the input of the stage with a resistive pad (R1, R2, and R3). The transistor impedance of approximately $Z = (10 + j50)\Omega$ was thereby transformed to an impedance in the range of $Z = (25 + j10)\Omega$ at the pad input. It was then a relatively easy task to design a reactive transformer (L1 and C1) to transform from the pad impedance to 50Ω at the amplifier input.

Although more complicated lossless networks were investigated, this network gave a good compromise among noise performance, stability, sensitivity to transistor variations, ease of tuning, and ability to tune the oscillator and amplifier independently.

The interstage network (L2, C2) is designed to match the collector impedance of the first stage to the base impedance of the second stage. In order to simplify this network and to avoid stability problems as a function of tuning, the interstage has a resistor, R4, which, in effect, lowers the interstage Q and places an upper bound on the collector load impedance at high frequencies; that is, at frequencies above the band of interest but below the f_t of the transistors. The networks (L3–C3, L4–C4) and resistors (R5, R7, R11) used between the remaining stages were designed in a similar manner.

Class A operation was chosen for the second stage. This arrangement minimizes the impedance variations which are reflected back to the input of the first stage, thereby minimizing the oscillator matching problem discussed previously. This stage uses an RCA 2N3375 npn transistor.

The third and fourth stages operate in the class C mode and use RCA 2N3375 npn transistors. In order to realize the desired power output with a power source of only -19 volts, it was necessary to parallel three transistors in the output stage. It is desirable for each of the output transistors to operate at approximately the same level. The input impedance to each transistor is very low; even small variations in wiring inductance among the three inputs would be sufficient to result in unequal input levels. Feedback techniques to increase the input impedances would improve this but would be too wasteful of output power. Therefore, it was decided to independently feed each of the bases through tunable capacitors (C5, C6, and C7).

The output network (L5, C8) is designed to match the collector impedance of the output stage to 50Ω. In addition, this network provides some rejection to the harmonics of the output frequency.

4.3 *Multiplier Chain*

Figure 11 shows the detailed circuit of the first quadrupler. It consists of the varactor diode, a shunt resistor for self-bias, an input resonant circuit (L1, C1) tuned to 125 MHz, an output resonant circuit (L3, C3) tuned to 500 MHz, and an idler resonator (L2, C2) tuned to 250 MHz. The input is connected to a tap on L1, and the output is connected to a tap on L3, to provide 50Ω input and output impedances.

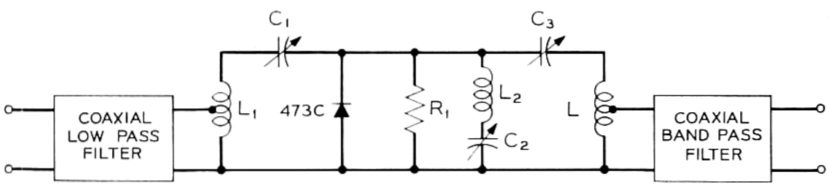

Fig. 11 — Circuit of the first quadrupler.

A low pass filter (LPF) is placed at the input to pass 125 MHz and to reject all other harmonics. A band pass filter (BPF) is placed at the output to pass 500 MHz and to reject all other harmonics. With an input power level of about 39 dBm from the oscillator amplifier the output power level is approximately 35 dBm. The varactor diode is a Western Electric Co. 473C diode, described in Ref. 1. With a typical breakdown voltage of 85V, a typical series resistance 0.25Ω, and a typical capacity at zero bias of 53 picofarads, this graded junction diode is driven into the forward charge storage region by the input power of approximately 8W.

The first quadrupler is followed by a coaxial isolator realized as a terminated circulator which provides approximately 20 dB of isolation (see Fig. 2). The forward loss is typically 0.25 dB. The next component in the chain is a Western Electric Co. 734A noise suppression filter (Fig. 2). It is a tunable temperature-compensated single cavity coaxial band pass filter with approximately 2 dB of loss at midband frequency and a bandwidth between the 3 dB points of approximately 300 kHz. The filter is followed by a second coaxial isolator realized as a terminated circulator. Both of these circulators, the one at the input and the one at the output of the 734A filter, are physically integrated into one package coded as a Western Electric Co. 13A circulator (see Fig. 2).

Figure 12 shows the detailed circuit of the doubler. It consists of the varactor diode, a shunt resistor for self-bias, an input resonant circuit (C4, L4) tuned to 500 MHz, and an output resonant circuit (L5, C5) tuned to 1000 MHz. The input is connected to a tap on L4 and the output is connected to a tap on L5 to provide 50Ω input and output impedances.

A low pass filter is placed at the input to pass 500 MHz and to reject all other harmonics. A band pass filter is placed at the output to pass 1000 MHz and to reject all other harmonics. With an input power level of about 33 dBm, the output power level is approximately 31 dBm. The varactor diode is a Western Electric Co. 473B diode, described in Ref. 1. Also, this diode is driven into the forward charge storage region with an input power of approximately 2 W.

As shown in Fig. 2, the doubler is followed by another coaxial isolator realized as a terminated Western Electric Co. 12A circulator providing about 20 dB of isolation. The forward loss of this isolator is approximately 0.2 dB.

Figure 13 shows the detailed circuit of the second quadrupler. This quadrupler, which uses transmission line elements, consists of the

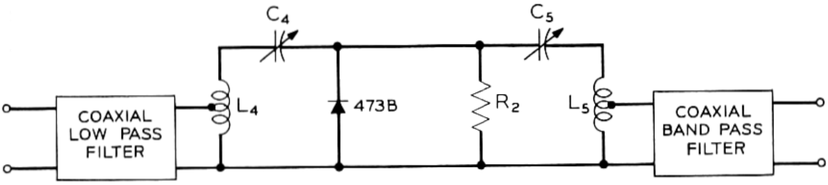

Fig. 12 — Doubler circuit.

varactor diode, a shunt resistor for self-bias, an input resonant circuit consisting of C6 and a shorted piece of coaxial transmission line (L6) tuned to 1000 MHz, an output resonant circuit consisting of C8 and a shorted piece of coaxial transmission line (L8) tuned to 4000 MHz, and an idler resonant circuit consisting of C7 and a shorted piece of coaxial transmission line (L7) tuned to 2000 MHz.

The input is tapped along L6, and the output is tapped along L8 for impedance matching into 50Ω. A low pass filter (LPF 1) is placed at the input to pass 1000 MHz and to reject all other harmonics. In order to insure the proper phase of the 2000 MHz idler signal reflected by this filter, a suitable length of 50Ω line (L) is inserted between the filter and the multiplier input.

A band pass filter, comprised of a low pass filter (LPF 2) and a high pass filter is connected to the output to pass 4000 MHz and to reject all other harmonics. With an input power level of about 31 dBm the output power level is approximately 26.5 dBm. The varactor diode is a Western Electric Co. 473A diode described in Ref. 1. This diode, too, is driven into the forward charge storage region with an input power of approximately 1.3 W.

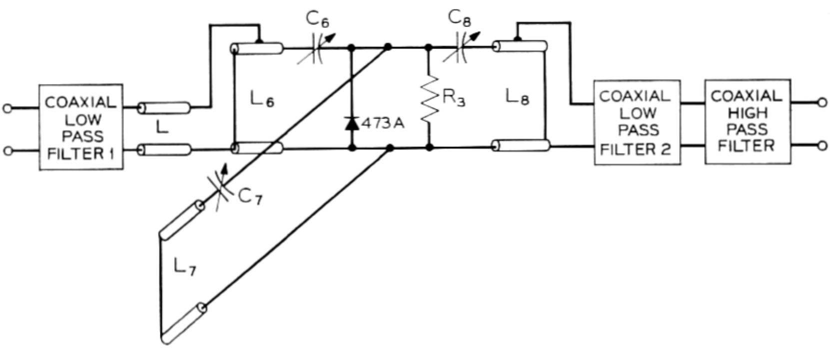

Fig. 13 — Circuit of the second quadrupler.

Since it was recognized that in all three multipliers the varactor diodes were driven into the forward charge storage region and therefore available design theory[7] could not be applied, the multipliers were designed by using simple circuit principles and making improvements experimentally. The filters used at the input and output of each multiplier are commercial units designed to operate between 50Ω impedances.

V. PERFORMANCE

5.1 *Reliability*

In the basic TD-3 System design it was established that a TD-3 microwave repeater could be permitted to have a maximum of 0.12 failure per year for the active devices based on the over-all TD-3 reliability requirements. About 20 per cent of this figure was allocated for the solid state devices and about 80 per cent for the traveling wave tube. Since there are approximately 200 solid state devices in TD-3 this corresponds to about 13.7 fits* per device (2740 fits per bay).

Because this was considered too few for the power devices in the generator, the generator power devices were allocated 100 fit performance objectives and all others were allocated 10. Since only seven devices fall into the high power category the total number of fits per bay is virtually unchanged ($7 \times 100 + 193 \times 10 = 2630$ fits per bay). The seven power devices and the three other devices result in 730 fits permitted for the generator's active devices.

The other passive devices will contribute an additional 270 fits since there are 135 components which should have 2 fit performance or better. This results in 1000 fit performance predicted for the generator which corresponds to 0.0087 failure per generator year.

Up to now there is nothing quantitative to report on the generator failure rate. There have been approximately 30 generator years logged in the laboratory and about 100 generator years logged in the field with no device failures. However, there has been one generator failure because of a defective capacitor. Nevertheless, based on the device reliability described in Refs. 1 and 6, and the reliability objectives specified for the passive components, it appears that the objectives will be met.

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* A fit is a failure per 10^9 hours.

5.2 *Noise Performance*

As indicated in Section II, one of the main objectives is to obtain satisfactory noise performance between 0.3 and 6 MHz from f_c where f_c can be any one of the 17 frequencies between 3780 and 4100 MHz. Therefore, the noise very close to the carrier was of no interest and was not investigated.*

The carrier to noise ratio was measured at various points in the generator. The direct measurements were made at the oscillator output (125 MHz), the amplifier output (125 MHz) and the multiplier chain output (4000 MHz).

5.2.1 *Measurement Method*

The basic technique used to measure the carrier to noise ratio consists of measuring the carrier power with a standard power meter and then using the setup shown in Fig. 14 to remove the carrier and measure the noise on a calibrated spectrum analyzer.

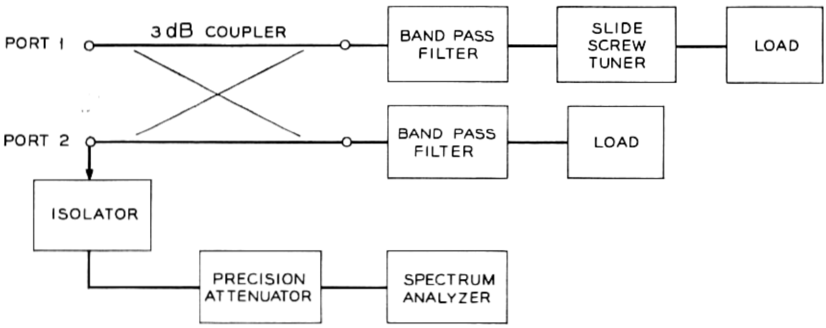

Fig. 14 — Noise measurement setup.

The equipment in Fig. 14 between ports 1 and 2 is a 3 dB coupler, two identical very narrow bandpass filters, and one fixed and one variable termination (fixed termination and slide screw tuner). By tuning both filters to f_c and adjusting the variable load, the insertion loss between port 1 and 2 can be made to behave like a narrow bandstop filter with infinite loss at f_c. The 3 dB points were ± 0.1 MHz for the 125 MHz bridge and ± 0.25 MHz for the 4000 MHz bridge.

The remainder of the setup, consisting of isolator, calibrated at-

* For TV transmission, this statement is inaccurate. However, systems tests involving transmission of TV signals disclosed no impairment from microwave generator noise.

tenuator and spectrum analyzer, merely serves as an indicator. The purpose of the narrow bandstop filter (port 1 to port 2) is to prevent overloading the spectrum analyzer with the carrier power while passing the noise sidebands with virtually no attenuation. By using a calibrated attenuator, the noise at port 1 at any frequency Δf away from f_c can be compared with a known white noise at port 1 (noise lamp). Then by measuring the carrier level at port 1, the carrier to noise ratio at port 1 can be computed.

5.2.2 *Over-all Noise Performance*

Figure 15 shows the typical carrier to noise ratio for a complete microwave generator as a function of Δf about any 4000 MHz f_c as measured with the 4000 MHz bridge described above. The noise performance was measured for several generators at 40°, 75°, and 140°F, and it was found that temperature has virtually no influence.

5.2.3 *Oscillator-Amplifier Noise Performance*

The noise performance of the oscillator and amplifier was evaluated with the 125 MHz noise measurement bridge described in Section 5.2.1. The C/N at the oscillator output was 166 dB and was flat in the band between ±0.3 and ±6 MHz from f_c. It was observed that the C/N could be improved, if necessary, by increasing the carrier power output of the oscillator. This was not done, however, mainly because it was not necessary (a C/N of 164 dB would be adequate according to Fig. 8), but also because it was desirable to limit the power dissipated in the crystal to preserve the desired frequency stability.

The C/N at the amplifier output is shown in Fig. 16. The noise

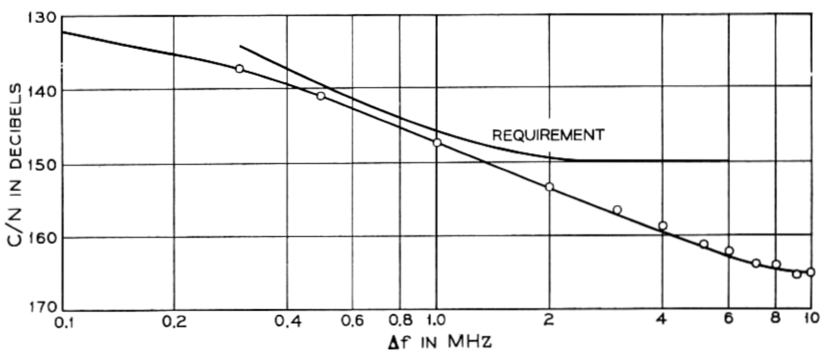

Fig. 15 — Measured C/N—Performance at the output of the generator (4000 MHz.

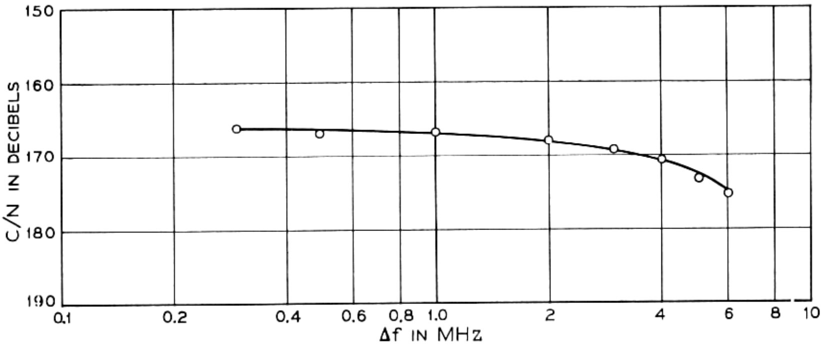

Fig. 16—Measured C/N—Performance at the output of the amplifier (125 MHz.

figure at the input to the amplifier's first transistor is approximately 8 dB, but, because the input network is lossy, the noise level at the amplifier input is approximately 11 dB. This corresponds to $N = -163$ dBm. Based on the oscillator output C/N, the noise contributed from the oscillator is -156 dBm. Therefore, the addition of the lossy input network did not contribute appreciably (0.6 dB) to the output C/N of the amplifier.

The improvement of C/N at frequencies between 1 and 6 MHz from f_c results from the band shaping in the amplifier. Again, this is not significant at this point because further improvements of the amplifier C/N would not necessarily realize an improved C/N at the generator output, since the internal noise of the multipliers would soon become controlling.

5.3 *Output Power and Frequency Stability*

Figure 17 shows the variation of the generator output power and frequency as a function of temperature. The design objectives were met over the 75 ± 10°F temperature range. In fact, the output power level stays within the requirement over the extended range of 40 to 140°F. However, the frequency changes by about 5 ppm over the extended range. Both of these variations are acceptable from a system standpoint.

VI. ACKNOWLEDGMENTS

The development of this generator was successful because of the work and cooperation of many of our colleagues at Bell Telephone Laboratories. The early development was done by H. W. Andrews,

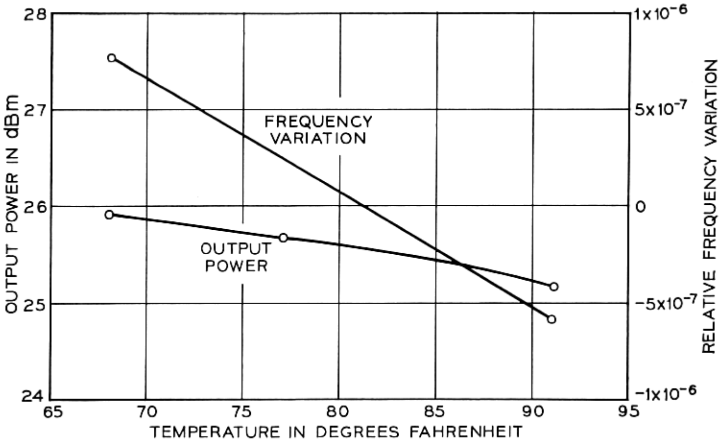

Fig. 17 — Measured variation of the output power and output frequency of the generator as a function of temperature.

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