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WT4 Millimeter Waveguide System:

Regenerative Repeaters

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The electronic circuitry and physical structure of 12 two-phase repeaters made for use in a field evaluation test are described. The repeaters consist of four kinds of circuitry; millimeter-wave circuits covering the 40 to 110 GHz range, IF circuits at 1.2 to 1.6 GHz, pulse circuits for the 274 Mb/sec pulse stream, and power supply circuits. All repeaters have been successfully operated over the trial repeater link in northern New Jersey. The results of this repeater development effort are the basis for an estimate of the performance obtainable in future four-phase repeater performance. This estimate has been used in the system repeater spacing calculations.

I. INTRODUCTION

A repeater design has been developed for the WT4 system which can regenerate a 274 megabit per second binary phase-modulated bit stream on mm-wave carriers in the 40 to 110 GHz range. These repeaters have sufficient gain to provide repeater spacing well in excess of the 45 km objective at an error rate of 1 in 10^9 bits when used with the transmission medium described in this issue of the B.S.T.J. This error rate is attained with a signal to noise ratio which is only a few decibels above the theoretical minimum. Twelve field evaluation test models have been constructed with three models each near 40, 54, 80, and 108 GHz. The models all have identical IF, baseband, and power supply circuitry, but differ in the construction and tuning of the frequency-sensitive millimeter-wave and equalizer circuits. There are four apparatus cases composing each repeater. These are the receiver, the line equalizer, the transmitter, and the power supply. The block diagram of Fig. 1 may be used to trace the signal flow through the repeater.

The modulated signal spectrum enters the receiver via a rectangular waveguide port, through a waveguide isolator into the mixer where the signal is converted to the IF signal centered at 1371 MHz. The IMPATT local oscillator is controlled at 1371 MHz below the incoming millimeter-wave spectrum by an AFC loop. The first IF amplifier has a maximum gain of 50 dB and a noise figure of 6 dB. The signal then leaves the receiver via a 50-ohm coaxial cable to the line equalizer which equalizes the differential delay slope caused by the transmission medium. Returning to the receiver case, the signal passes through the receiver equalizer, a bandpass filter which controls the width of the signal and noise spectra, more IF gain, and into the signal and AFC detector circuits. An error signal is fed back to the AFC circuit which varies the current through the local oscillator IMPATT to hold the IF signal centered at 1371

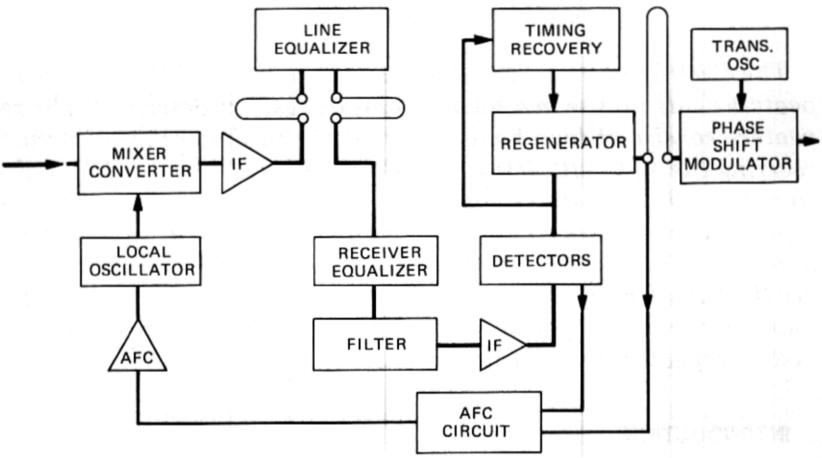

Fig. 1—WT4 repeater block diagram.

MHz ±1 MHz. The binary pulse stream from the signal detector is fed to the timing recovery circuit and the decision circuit in the regenerator. The regenerated pulses leave the receiver case via a coaxial cable (up to 30 feet long) over to the transmitter case. In the transmitter a high-power, cavity-stabilized IMPATT oscillator feeds a phase-shift modulator where a PIN diode reverses the phase of the millimeter-wave signal whenever a binary one is present on the incoming pulse stream. The output from the modulator then leaves the transmitter via a rectangular waveguide to a channel-dropping filter where it combines with all other transmitted spectra at that station and proceeds towards the next repeater. A short description of each of the major circuit blocks in the repeater will be given.

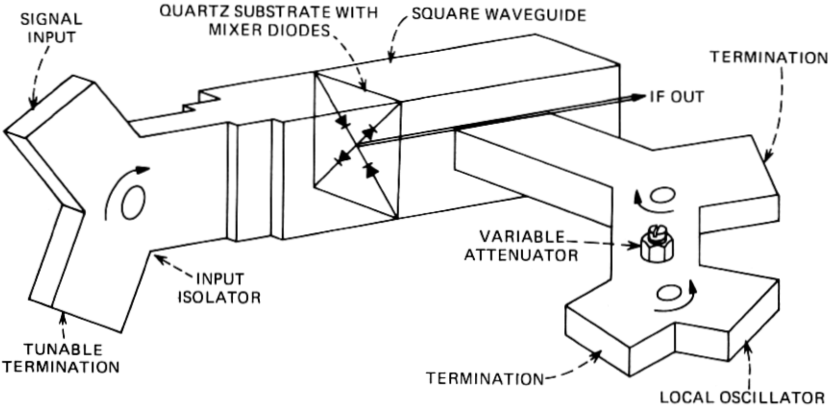

Fig. 2—Down-converter assembly.

II. DOWN CONVERTER

The down converter developed for the WT4 system is shown schematically in Fig. 2. It conists of an input isolator used primarily to reduce local oscillator leakage, a mixer circuit to shift the incoming millimeter-wave spectrum down to the IF passband centered at 1371 MHz, and an isolator-attenuator used to set the local oscillator drive level and prevent interaction between the mixer and the local oscillator (LO). The isolators are terminated circulators. The input isolator uses a single circulator and the isolator attenuator uses two terminated circulators.

The general format of the RF circulator designs is a waveguide, *H* plane, *Y* junction, with a simple cylinder of ferrite mounted on axis against one broad wall and a quarter-wave transformer with a central coaxial stub tuner on axis on the opposing wall. Adjustment of the stub penetration and the biasing field permits a peaking of the circulator characteristics at any frequency in a relatively wide tuning band. The typical instantaneous bandwidth characteristics are:

Return loss > 30 dB; BW 1 GHz
Return loss > 20 dB; BW 3 GHz
Insertion loss in 30 dB return loss band:
  <0.2 dB at 40 GHz
  <0.5 dB at 110 GHz

The isolator is required to provide low insertion loss, good input and output matches, and 20 dB of isolation over the signal band (~ 500 MHz) while providing 40 dB of isolation at the LO frequency, which is 1.37 GHz below the signal frequency. To meet these requirements the circulator is peaked at the signal frequency and is terminated in a tunable termination adjusted to conjugately match the circulator impedance at the LO frequency.

The isolator-attenuator between the receiver local oscillator and the mixer prevents pulling of the oscillator by the mixer and permits adjustment of the LO power to the mixer. It provides isolation in excess of 60 dB, a 20-dB attenuation range, and matches independent of attenuator setting. The insertion loss ranges from <0.5 dB at 40 GHz to <1.0 dB at 110 GHz.

Early in the development of the WT4 repeater a beam-lead mixer diode was selected for use in the down converter. Both theoretical studies and practical experience indicated that much greater lifetime should be attained with a beam-lead device than with wafer-type millimeter-wave diodes. It was also decided to try to avoid the use of resonant circuits to separate the signal and local oscillator frequencies. Since even an IF of 1371 MHz puts local oscillator and signal at 110 GHz fairly close together, sharp filter cutoffs are required and this may mean high transmission loss.

A double-balanced mixer configuration where the local oscillator and signal are in orthogonal polarizations in a square waveguide appeared to provide a straightforward development path. As shown in Figure 2, four beam-lead diodes are bonded to a gold thick-film pattern on a quartz substrate. The quartz substrate is clamped between two square waveguides so that the diodes run along the diagonals of the square. On one side of the substrate, the square waveguide is stepped down to rectangular waveguide for the signal input port. The square waveguide on the other side of the substrate has the IF lead brought out and also has a right-angle transition for the local oscillator waveguide port. Impedance matching elements are inserted in both the signal path and the local oscillator waveguide port. With about 1 milliampere diode bias, the IF impedance for the four diodes comes out fairly close to 50 ohms; however, an impedance matching transformation can be provided in the IF lead. Large gold pads on the quartz substrate provide return to ground for the IF currents on the outer beam leads. A sheet of 0.01-mm Mylar insulates these pads and each diode is provided with an adjustable dc bias current. Seven different basic designs (using seven different sizes of square waveguide) have been developed to cover the 40 to 110 GHz frequency range. Each design can be adjusted to cover any channel in one of the seven subbands[1] into which the WT4 spectrum is divided.

The major factor controlling the performance of the down converter is the quality of the mixer diodes used in the circuit. The important number for the WT4 system is the receiver noise figure, however, this number depends not only on the down-converter performance, but also on the IF amplifier noise figure and the matching network and/or isolator that is used between the down converter and IF amplifier.

$$\mathrm{NF} = L_c(\mathrm{NR} + \mathrm{NFI} - 1)$$

Where NF is the total single-sideband receiver noise figure

L_c is the down-converter conversion loss

NFI is the IF amplifier noise figure including any matching network or isolator

NR is the noise ratio of the mixer diodes

For the field evaluation test, the total receiver noise figure ranged from 12.2 dB at 40 GHz to 18.5 dB at 108 GHz. Laboratory tests with improved mixer diodes and a 3.5 dB noise figure for the IF amplifier, have yielded circuits which will produce repeaters with 7.5 dB noise figure at 40 GHz and 13.5 dB noise figure at 110 GHz.

III. IMPATT oscillators

One of the basic technological advances which permitted the development of a millimeter-wave repeater during the 1970s was the availability of IMPATT diodes operating up to 110 GHz. It was decided early in the repeater development that self-controlled oscillators would be employed in the repeater rather than have a low-frequency (about 100 MHz) crystal oscillator control the millimeter-wave frequencies. This implied that the receiver would have some form of AFC in order to track whatever drifts occurred in the transmitter or local oscillator frequency. With the omission of the many multiplier stages required to get to the millimeter-wave frequency range, the millimeter-wave circuitry occupies a relatively small percentage of the total volume in the receiver or transmitter case. The important parameters in the performance of the millimeter-wave oscillators are the frequency stability, the power output, the noise output, and the lifetime of the diodes.

The frequency stability of an IMPATT oscillator is very closely related to the physical and electrical structure in which the diode is embedded. For the transmitter, maximum frequency stability and minimum noise is required; both of these characteristics improve as the oscillator circuit Q is made larger. The major limitation on increasing Q is the increase in loss in the frequency-controlling resonant cavity. The WT4 transmitting oscillator designs have been planned to limit the resonant cavity loss to about 1 dB. This should yield an oscillator Q which is large enough to provide a transmitter frequency which drifts less than ±25 MHz from all causes and also limit the oscillator output noise to an rms frequency fluctuation of less than 1000 Hz for a noise bandwidth of 1 kHz. The structure chosen for this design consists basically of three parts. First is a diode mount to place the diode package into a dominant-mode rectangular waveguide and provide a contact for dc bias of the diode, second is a damping resistor in the waveguide to stabilize the circuit and inhibit oscillations at undesired frequencies of oscillation. Third is a resonant cavity to provide accurate control of frequency and furnish high-Q,

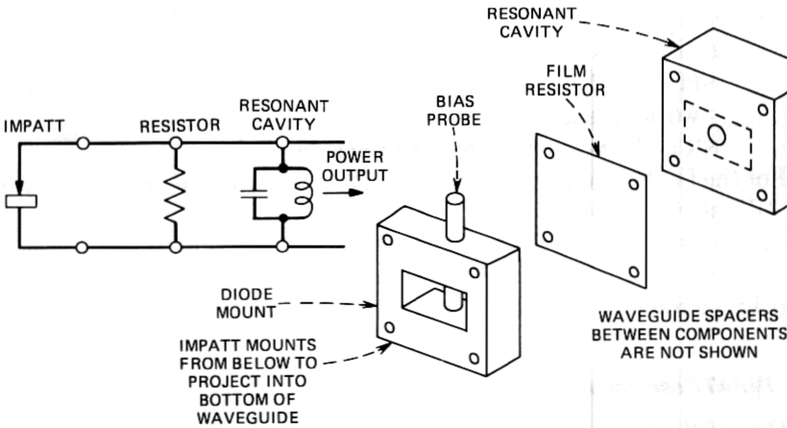

Fig. 3—Transmitting IMPATT oscillator.

low-noise performance. Figure 3 shows a physical and electrical schematic of the transmitting oscillator structure. Table I shows the electrical performance of the 12 models. All objectives were met except for the failure rate and slightly excessive frequency drift in a few cases. There have been five IMPATT device failures in about 2 years of field evaluation test operation. This rate is too high, however, and a new design of hermetically sealed devices has recently been made available. This improved device should provide a transmitting IMPATT with a FIT rate of 600 and also improve the frequency stability.

The local oscillator is a different design because of the AFC requirement that changes in bias current must produce a frequency shift of ±100 MHz or more. The resonant cavity for the local oscillator is formed as part of the dc bias line for the IMPATT. A movable coaxial short is part of the bias line, and the position of this short is the primary frequency-determining element in the oscillator. There is also a movable waveguide

Table I — Field evaluation trial transmitting oscillators

Frequency, GHz	Power, mW	FM noise, rms deviation, Hz/$\sqrt{\text{kHz}}$	Number of failures	Total drift, MHz	Months operating
40.235	145	238	0	+21	12
40.760	145	157	0	− 3	16
41.285	142	174	0	−27	12
53.910	75	147	0	+ 2	20
55.085	86	250	0	+ 6	20
55.610	83	217	0	+15	20
80.465	44	400	1	+14	11
80.990	43	234	0	−12	17
81.515	43	535	0	− 6	19
107.665	23	153	0	+38	18
108.190	26	230	1	+ 3	12
108.715	26	209	0	+ 6	14

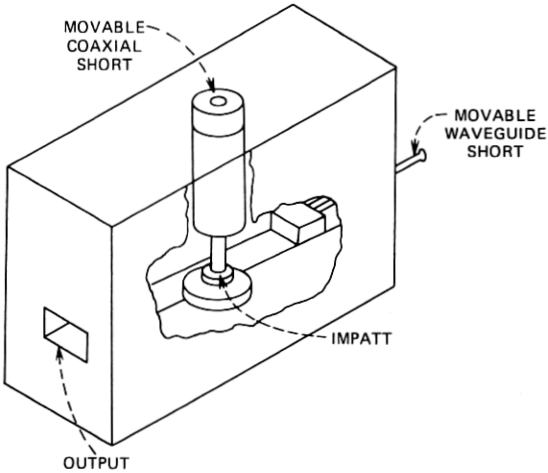

Fig. 4—Local oscillator.

short behind the diode, and this is varied to change the coupling from the diode to the waveguide. There are three basic designs to cover the 40 to 110 GHZ range for either the transmitting or local oscillators with different size bias probes available to provide complete frequency coverage and to accommodate diode variations. The local oscillator structure is shown on Fig. 4 and the field evaluation test performance is listed in Table II.

IV. IF AMPLIFERS AND VARIOLOSSERS

Several basic design preferences were established: (*i*) thin-film microstrip technology because the ground plane on the back of the substrates allows valid testing and trimming with enclosure covers removed;

Table II — Field evaluation trial local oscillators

Frequency, GHz	Power, mW	Modulation sensitivity, MHz/mA	FM noise, rms deviation, Hz/$\sqrt{\text{kHz}}$	Number of failures
38.868	20	24	1250	0
39.389	16	9	1000	0
39.914	16	13	800	0
52.538	31	11	1120	0
53.714	28	13	700	0
54.239	10	10	1000	0
79.096	32	10	880	0
79.620	31	10	880	1
80.145	32	13	640	0
106.295	35	30	1250	1
106.819	35	16	1150	0
107.344	20	17	800	0

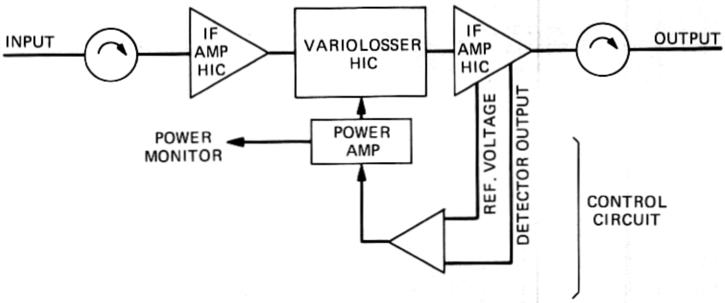

Fig. 5—Field evaluation test IF amplifier.

(*ii*) appliqued capacitors because the deposited capacitor technology is still evolving at these high frequencies; (*iii*) single string design for minimum transistor count with the resulting low input and output return losses masked by isolators; (*ii*) trim capability in order to accommodate any reasonable transistor parameter spread with one substrate design; (*v*) relatively little stress on miniaturization since the size of many repeater components excluded strong miniaturization in any event. Within these preferences, the design continued earlier work at Bell Laboratories.

The repeater requires a maximum IF gain of 100 dB in the band 1.15–1.60 GHz and an Automatic Gain Control (AGC) range of 40 dB. This gain is provided in two assemblies which contain two amplifer substrates each with 26 dB of gain and one variolosser substrate with a loss range from 2 dB to 22 dB. See Fig. 5. Lumped-element isolators provide 20 dB input and output return losses for the assembly.

4.1 IF amplifier substrates

A new beam-lead microwave transistor, developed at Bell Laboratories, has a maximum available gain of 11 dB at 1.6 GHz. Any resistance or inductance in the external circuit from emitters to ground must be kept to a minimum to avoid reducing the gain and the stability. Three transistors per substrate are sufficient for 26 dB of total gain. The gain rolls off at 6 dB per octave per transistor in the 1–2 GHz range; therefore the basic design approach is to use interstage networks (Fig. 6) to match the devices as tightly as possible at the top end of the band for highest gain and, in a controlled manner, mismatch the devices toward the low end of the band to obtain gain flatness. Input and output return losses are improved to the 10 dB level by two-element matching networks (Fig. 7). The 25-ohm resistors in the interstage networks serve to dissipate power at the lower frequencies where the devices are systematically mismatched. The design has great trim flexibility to meet batch-to-batch

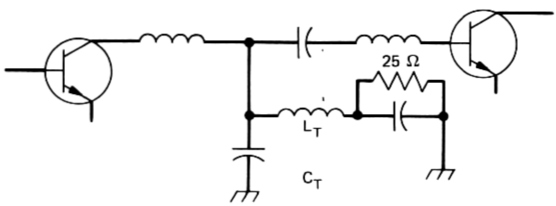

Fig. 6—Interstage network.

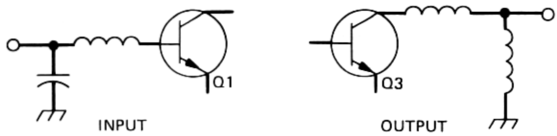

Fig. 7—Input/output matching networks.

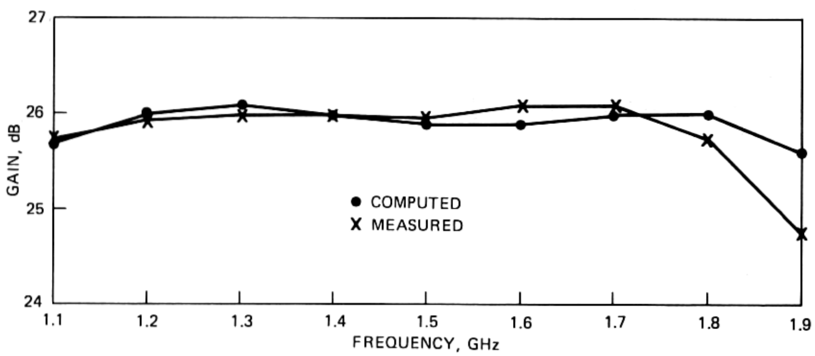

Fig. 8—Insertion gain vs. frequency.

variation in the transistor. All fixed capacitors are selectable; all inductors are 90-ohm transmission lines of varying length formed by laser cutting of rungs in a ladder-like layout; three capacitors are variable trimmer capacitors. The S parameters for a batch of similar transistors are measured and a computer program used to select a typical or nominal transistor for that batch. An optimization program is then used to select all fixed capacitors and determine all laser cuts for the inductors. After that, all transistors within the batch can be tuned with the variable capacitors to give the required flat gain. Figure 8 is an example; it also shows the close agreement between equivalent circuit computation and actual measurement.

4.2 IF variolosser substrate

The variolosser interconnects the two amplifier substrates. Its basic configuration is shown on Fig. 9. By having identical variable attenuator

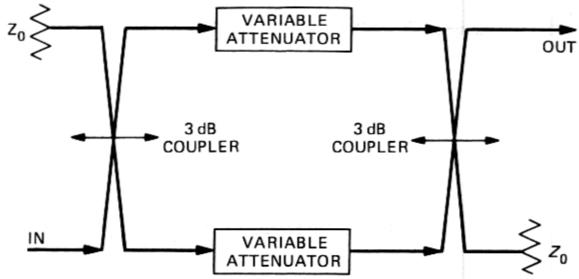

Fig. 9—Basic variolosser configuration.

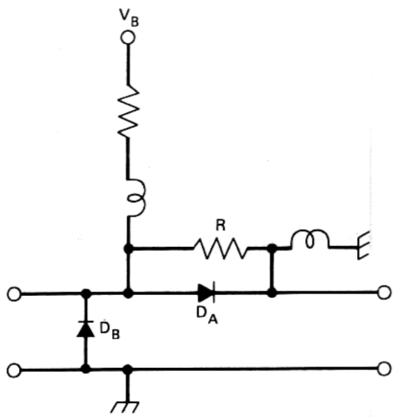

Fig. 10—Schematic of variable attenuator.

networks in both sides of the 3 dB couplers, good input and output return losses are assured even when the impedance of the attenuators deviates markedly from the 50-ohm characteristic impedance of the microstrip transmission lines and couplers. The schematic of the variable attenuator is also shown on Fig. 10. When the control voltage, V_B, is positive, D_A is forward-biased with low resistance and D_B is back-biased with high resistance resulting in little loss (approximately 2 dB) in transmission through the variolossers. As V_B is made less positive and, eventually, negative, D_B becomes forward-biased and D_A back-biased with a resulting high attenuation in the variolosser. The inductors (spiral structures in thin film) serve to isolate the IF paths from the bias circuitry. The final design exhibited about 1.5 dB of change in loss slope as the attenuation at midband changed from 2.0 to 22.5 dB; most of the change occurs in the last few dB of the attenuation range.

The driver for the variolosser gets its control voltage from the AGC detector on the amplifier substrate. The circuit can be adjusted to maintain a level of −12 dBm at the output of the first IF amplifier as-

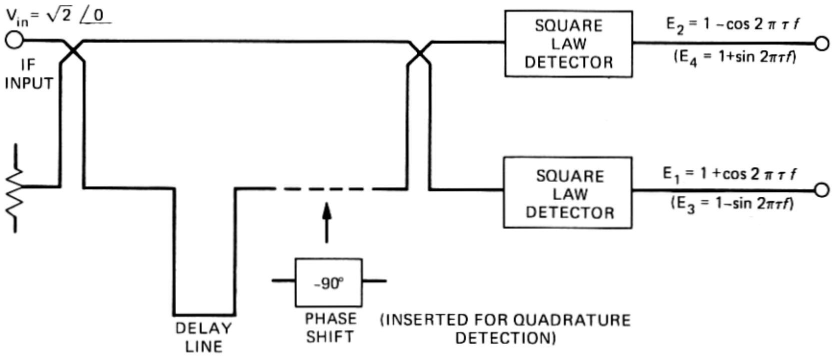

Fig. 11—Delay-line detector.

sembly and a level of +7 dBm at the output of the second IF amplifier assembly.

V. DETECTORS

Since two-phase differentially coherent phase-shift keying (DC PSK) can be demodulated with good performance by a very simple delay-line detector without the need for carrier recovery, it was decided to employ such a scheme in the initial WT4 system. Figure 11 shows the simple delay-line detector. In order to compare the phases of the successive pulses of IF signal, the delay line must be one time slot, $T = 3.65$ ns, in length, In addition, the delay line must be an integral number of half wavelengths at the IF to obtain maximum differential output voltage. For an integer of 10, this yields an IF of 1370.88 MHz. Thus the delay line is the IF-determining element.

Analysis shows that the differential output voltage from the detector, $D_s = E_i - E_2$, may be written:

$$D_s = -k \cdot A(t) \cdot A(t - T) \cos \left[\Delta\omega \cdot T + \phi(t) - \phi(t - T)\right]$$

where $A(t)$ and $\phi(t)$ are amplitude and phase, respectively, of the IF signal, $\Delta\omega$ is any deviation from correct IF center frequency, and k is a detector constant. At the sampling times, t_K, $A(t_K) = A(t_K - T) = A$ and $[\phi(t_K) - \phi(t_K - T)]$ is π for a binary "one" or 0 for a "zero." Therefore (for $\Delta\omega$ small):

$$\text{Binary "one": } D_s(1) = k \cdot A^2 \cdot \cos(\Delta\omega \cdot T + \pi) \approx +k \cdot A^2$$

$$\text{Binary "zero": } D_s(0) = k \cdot A^2 \cdot \cos(\Delta\omega \cdot T) \approx -k \cdot A^2$$

These are the desired type of demodulated outputs. However, $\Delta\omega$ must be kept small; a drift of 68.5 MHz, i.e., 0.1 percent at 75 GHz, makes $D_s(1) = D_s(0) = 0$. Thus automatic frequency control (AFC) is required.

The AFC detector differs from the signal detector only by having an additional $\pi/2$ phase shift in series with the delay line. The output, $D_Q = E_3 - E_4$, may be written:

$$D_Q = -k \cdot A(t) \cdot A(t-T) \cdot \sin\,[\Delta\omega \cdot T + \phi(t) - \phi(t-T)]$$

At the sampling times (for small $\Delta\omega$):

$$\text{Binary "one": } D_Q(1) = -k \cdot A^2 \cdot \sin\,(\Delta\omega \cdot T + \pi) \simeq +k \cdot A^2 \cdot \Delta\omega \cdot T$$

$$\text{Binary "zero": } D_Q(0) = k \cdot A^2 \cdot \sin\,(\Delta\omega T) \approx -k \cdot A^2 \cdot \Delta\omega \cdot T$$

Thus we get an output of magnitude proportional to the frequency error; however, the sign is positive or negative depending on the bit received and the signal averages to zero for a reasonable string of bits. If the product $D_{\text{AFC}} = D_S \cdot D_Q$ (at the sampling times) is formed, then:

$$\text{Binary "one": } D_{\text{AFC}}(1) = D_S(1) \cdot D_Q(1) = k^2 \cdot A^4 \cdot \Delta\omega \cdot T$$

$$\text{Binary "zero": } D_{\text{AFC}}(0) = D_S(0) \cdot D_Q(0) = k^2 \cdot A^4 \cdot \Delta\omega \cdot T$$

D_{AFC} consists of pulses of proper polarity and magnitude, occurring at the sampling times, and may be averaged to form a dc control voltage for the AFC circuit. This circuit is referred to as the "fast" AFC detector.

Unfortunately, the same operation may take place at frequencies which are half a baud frequency, 137.088 MHz, above and below the proper IF frequency. This can result in false lock-ups during turn-on of the repeater. The output from a simple discriminator, the "slow" detector, similar to Fig. 11 but with a short 5/4 λ IF delay line is used to override the output from the "fast" detector at these false operating points.

VI. AFC LOOP

The automatic frequency control (AFC) loop primarily responds to the "fast" detector which has a high sensitivity. In order to bring the local oscillator close to the correct null of the "fast" detector, however, the "slow" detector takes over control of the loop whenever the IF is more than 30 MHz in error. The slow detector has zero crossings separated by 600 MHz, but the local oscillator is set to vary by no more than ±150 MHz. Figure 12 shows a simplified block diagram of the AFC loop. When the slow detector and its instrumentation amplifier have an output larger than the breakdown voltage of the zener diodes, then it dominates the loop and the fast detector has no effect. However, when the IF is within 30 MHz of its correct value, the slow loop voltage does not breakdown the dead band diodes and the fast detector is in control. The fast detector has a sensitivity of about 1 mV per MHz and the local oscillator about

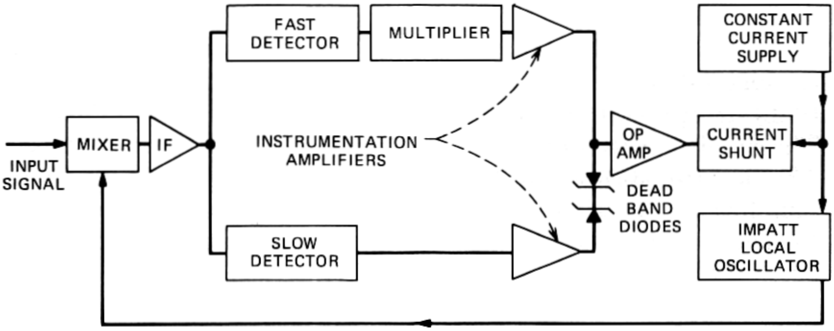

Fig. 12—AFC loop.

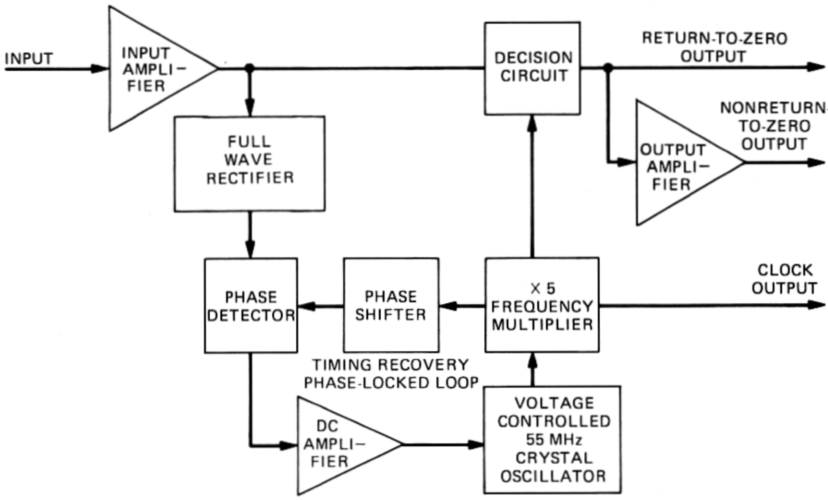

Fig. 13—Block diagram of WT4 regenerator.

10 MHz per mA. The gains of the amplifiers and the current-shunting transistor are adjusted to obtain a loop gain of greater than 100. All millimeter-wave oscillator frequency drifts are therefore reduced by a factor of 100 in the IF band. However, the AFC loop does not reduce any errors or drifts that occur in the fast detector zero crossing.

VII. REGENERATOR

An overall block diagram of the WT4 regenerator is shown in Fig. 13. The regenerator performs two basic functions, bit timing recovery and bit-by-bit binary decisions. The regenerator accepts a differential input signal of ±130 mV and amplifies this by 20 dB with a wideband amplifier before splitting the signal into two paths, one toward the decision circuit and one toward the timing recovery circuit.

The main signal path is toward the decision circuit where the binary decision is made as to where a "1" or a "0" was transmitted. Because of the high rate of transmission, 274.176 Mb/s, an emitter coupled logic (ECL) structure with microwave transistors was adopted for this circuit. The circuit consists of three stages of emitter-coupled transistor pairs. The first two stages provide amplification and the third stage is the sampling and decision-making stage. The third stage is shown in Fig. 14. The amplified baseband signal, V_{in}, to be sampled is applied to the bases of the transistors Q1 and Q2 in differential form.

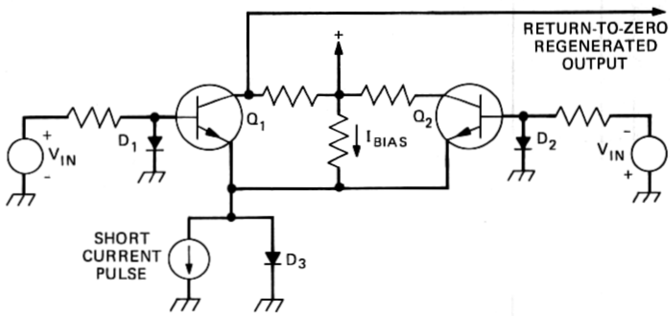

Fig. 14—Third stage of decision circuit.

If V_{in} is positive, then D1 will be forward-biased and D2 will be back-biased. A bias current flowing into D3 keeps it forward-biased. Therefore, under these conditions the biasing of diodes D1, D2, and D3 prevents either transistor from becoming active since the base-to-emitter voltages are either zero or negative. When a short pulse of current of sufficient magnitude is drawn from the node connecting the emitters of Q1 and Q2, the bias current flowing in D3 will be diverted and the additional current will be drawn from Q1, turning it on and producing a short voltage pulse at its collector. This voltage pulse is the return-to-zero (RZ) regenerated output signal. If V_{in} had been negative, then Q2 would have been turned on and no output at the collector of Q1 would have been produced. The RZ output of the regenerator is used to drive the phase shift modulator in the transmitter.

The time at which the short current pulse appears in the regenerator and, therefore, the time at which the baseband signal is sampled, is determined by the timing recovery circuit. The timing recovery circuit is a phase-locked loop (PLL) which is locked to the 274.176 MHz tone generated by full-wave-rectifying the incoming nonreturn-to-zero baseband signal. The PLL for the WT4 regeneration is a second-order loop to keep both phase offsets and phase jitter small. However, the second-order loop has a closed-loop characteristic which exhibits gain over some frequency band. In order to avoid any large gain enhancement in a long chain of repeaters, the loop has been designed to be highly overdamped.

The regenerator by itself requires a baseband signal-to-noise ratio of 16.0 dB in order to achieve a 10^{-9} error rate. This is only 0.4 dB above the theoretical minimum of 15.6 dB. The timing recovery circuit has a hold-in range of ±40 kHz, operates over a 30 dB range of input signal levels, and produces less than 0.5 degree of phase jitter for a single repeater hop.

VIII. PHASE SHIFT MODULATOR

Each PSM is adjusted for operation at one of the 124 channel frequencies. It responds to each "1" in the bit stream from the regenerator by providing a differential phase shift of 180 deg ±5 deg. The switching time is approximately 0.5 nanosecond, and the timing jitter is less than 0.2 nanosecond. The insertion loss ranges approximately from 1.5 dB at 40 GHz to 3 dB at 110 GHz and differs in the two states by less than 0.25 dB. The PSM is designed to accommodate power levels up to 250 mW.

The PSM is of the basic type described by Clemetson, et al.[2] It consists of a PIN diode switch connected to one port of a Y-junction circulator and a terminated circulator used as an input isolator, all in a common integrated housing. The switch consists of the PIN diode[3] coupled to a standard rectangular waveguide. The coupling is through a short section of coaxial line parallel to the E plane of the guide. In a very loose approximation, the diode and its bias lead are seen as a short circuit on the waveguide for one bias state and as an open circuit for the other. In addition to the phase being reversed, the insertion loss of the modulator in the two bias states must be adjusted for a balance. The loss balance is achieved by adjustment of the length of coaxial line between the diode and the waveguide wall and between the choke filter and the waveguide wall.

The PSM bias driver switches the PIN diode between forward- and reverse-bias states in approximately 0.5 nanosecond. The forward current is 10 mA and the reverse-bias voltage is 10 volts. The driver responds to the pulse input from the regenerator. The driver consists of an input buffer amplifier, a flip-flop, an output amplifier, special circuits for adjusting the timing of the PLM output and temperature-compensated dc bias circuits. The pulse circuit and dc bias circuit are fabricated on two separate alumina substrates. Each substrate is approximately 1.1 × 1.3 × 0.025 inches. The circuits use thin-film tantalum nitride resistors, plated up gold conductors, beam-leaded transistors and diodes, and chip capacitors. The circuit cards are mounted on an aluminum drawer which

Table III — Power requirements for the WT4 repeater

INPUT:			
Nominal	−24		
Minimum	−20		
Maximum	−27		
Maximum transient	−30		
OUTPUTS:			
Constant voltage			
	No. 1	No. 2	No. 3
Voltage	−10	+10	−16
Regulation	±0.5%	±.5%	±.5%
Maximum load current	2.4 A	0.3 A	0.4 A
Dynamic load variation	±10%*	70–270 mA	195–370 mA
Ripple	<15 mV P-P	<15mV P-P	<15mV P-P
Constant current (two supplies)			
Current range		60–130 mA	
Voltage		15–50 V	
Regulation		±0.5 mA	
Ripple		<.05 mA P-P	
Noise		No tone > 4μA	
Output capacitance		<50 pF	
Maximum instantaneous current due to step change of input voltage		1.30 × set current or 120 mA, whichever is greater	

* At nonprotection switching repeaters, 1.25 A ± 10%. At protection switching repeaters, 2.10 A ± 10%.

in turn is mounted in the RF housing of the PSM. The average dissipation of the driver excited by a pseudorandom word is 3.8 watts.

IX. POWER SUPPLY

Power at the voltages required by the electronics is obtained via dc-dc converters. Constant-current and constant-voltage regulators are provided as necessary to power the repeater transmitter and receivers. Table III gives the power requirements for the field evaluation test repeaters. The power supply is mounted in its own housing with the current regulators located in the transmitter and receiver housings close to the IMPATT diodes which they feed. The converter portion of the supply uses a closed-loop frequency-controlled ferroresonant transformer whose fundamental switching frequency is 150 Hz. The choice of a low-frequency converter eases the problem of suppressing high-frequency noise. The voltage regulators were designed with a minimum input-output voltage differential to minimize wasted power. Regulator efficiencies of 82 percent and converter efficiencies of 68 percent were attained for the field evaluation test.

X. REPEATER PHYSICAL DESIGN

The WT4 field evaluation test repeater shown in Fig. 15 consists of four parts including a transmitter, receiver, line equalizer, and a dc–dc con-

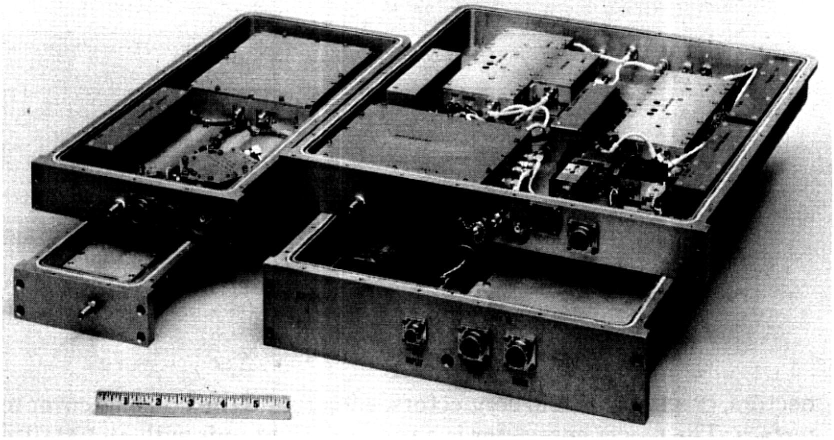

Fig. 15—Field evaluation test repeater.

verter. In an effort to reduce the loss within the channel multiplexer, the repeater was designed to make it as thin as possible to permit minimizing the distance between adjacent channel diplexer ports. To facilitate back-to-back testing of receivers and transmitters at the end of a span, separate housing are provided for receiver and transmitter circuits. The "line" equalizer circuits are closely associated with the transmission medium and the channelization networks, hence the fourth housing. Each of the four repeater housings consists of a cast aluminum structure resembling a frying pan with a flat aluminum cover. Subassemblies within each housing are bolted to the 0.3-inch-thick bottom plate allowing conductive heat transfer to the mounting surface of the housing which contacts a water-cooled channel attached to the repeater frame.

The repeater housings and covers form pressure-tight and RFI-shielded enclosures for the circuits. Dry nitrogen gas at 0.2 psig fills the housing thereby protecting a number of high-frequency devices which have not been encapsulated. The nitrogen is fed from the channel diplexer via the receiver and transmitter waveguide ports (Fig. 16). The line equalizer obtains dry nitrogen through tubing connected to the receiver. The receiver, transmitter and line equalizer have gas valves mounted on the rear of the housings to permit initial factory pressurization and purging during installation. The dc–dc converter has no shielding or pressurization requirement.

The RFI shielding is necessary because the intermediate frequency is in a prominent military radar band which could cause interference with the repeater signal. To maintain both RFI shielding and pressure sealing, dual-purpose seals are employed at the waveguide ports, power con-

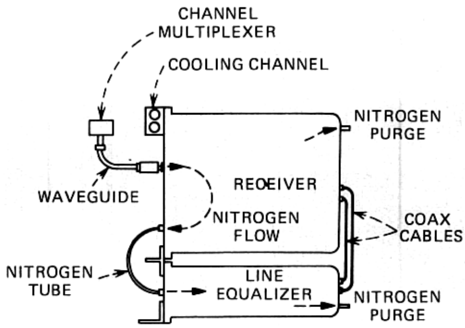

Fig. 16—Receiver and line equalizer.

nectors, IF and baseband connectors, and around the housing-cover interface. The power connector is a pressure tight unit with an RFI filter in series with each pin. Each lead in the power cable is shielded to minimize RFI pick-up. The IF and baseband connections to the repeater housings are pressure-tight and naturally shielded by the coaxial outer conducter.

10.1 Thermal considerations

Each WT4 field evaluation trial repeater dissipates 50 watts. The shielding requirement eliminated the option of removing the heat from critical circuits by convective means. The thermal question is further aggravated by the requirement that the IMPATT diode mounting stud be maintained at a temperature below 50°C. If a forced-air system were employed to maintain the required temperatures, the repeater frames would have been congested with large ducts, the audible noise level would be intolerable, and the system would not grow gracefully from a few repeaters to the full complement of 124. The best choice for the WT4 system was to remove the heat via a chilled water system. As shown in Fig. 17, heat is conducted from the IMPATT diode stud to the oscillator body and then to the repeater housing which is bolted directly to a channel filled with flowing cool (15°C) water. A companion article, "The Repeater Station," this issue, describes the features and reliability aspects of the chilled water system.[5]

10.2 IMPATT diode bias connections

One of the most serious mechanical problems was the design of reliable IMPATT diode bias contacts for the repeater. A bias rod is in contact with a quartz stand-off attached to the diode stud. A gold strap is bonded to both the stand-off and the diode. An attempt was made to minimize the mass in contact with the stand-off while providing a spring which could apply a force of between 60 and 100 grams over a 0.75-mm range of de-

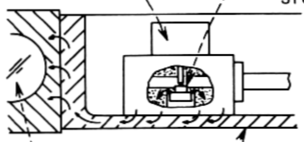

Fig. 17—Thermal path for receiving oscillator.

flection. In addition the natural frequency of the system had to be as high as possible to permit meeting stringent handling and shipping requirements. To minimize RF leakage, a close spacing (0.001 inch) between the bias line and oscillator walls was maintained. Insulation was formed by spraying a thin hard acroloid coating on the bias lines. It was necessary to develop separate designs (Fig. 18) for the transmitting and receiving oscillators.

10.3 Hybrid-integrated circuits

The WT4 field evaluation test repeater contains 21 codes of thin-film hybrid integrated circuits (HICs). The HICs that make up the baseband unit utilize standard bilevel thin-film technology. A number of IF circuits such as signal and AFC detectors, equalizers, and the IF filter require close tolerance ceramic thickness and dielectric properties. In addition, special metal and line delineation systems were developed especially for WT4 to meet the exacting line width requirements.

10.4 Circuit packaging

All WT4 thin-film substrates are housed in aluminum housings (Fig. 19). Spring clips are attached to each substrate and they are held in position when the housing cover is secured. Circuit interconnections are provided by gold straps bonded to gold lands on the top of the substrate.

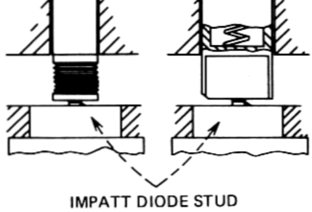

Fig. 18—Bias rods for WT4 oscillators.

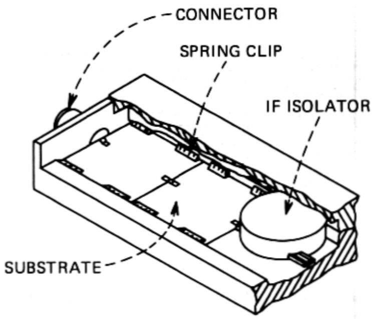

Fig. 19—Mounting of WT4 HICS.

Circuit ground continuity is provided by gold straps bonded to the ground planes on the bottom of the substrates. The metal housings serve only to support and shield the circuits. Filtered feed-throughs are provided for dc and monitoring connections. Circuit connections to the substrates are made through 50-ohm miniature coaxial connectors that clearly define both electrical and mechanical boundaries for each circuit package.

10.5 Repeater to channel diplexer connection

Figure 16 illustrates how a receiver or transmitter interfaces with the channelization network. Connections are made via a variable attenuator, a waveguide bend, and a short length of rectangular waveguide. For WR10 and WR15 these interconnection assemblies are flexible enough to permit small WG-WG misalignments. This is not the case for the WR22 assembly and a special short length of flexible WR22 waveguide is provided.

XI. SUMMARY

Twelve repeaters were manufactured at Bell Laboratories during 1974 and 1975 covering the frequency range 40 to 108 GHz. These repeaters were installed at the ends of the 14-km waveguide line in northern New Jersey and utilized in the systems test as described in the companion paper in this issue of the B.S.T.J. It is planned to continue operating the system test at least through 1977. While there are some known circuit changes which would improve repeater performance, the design described in this paper can be used to manufacture two-phase repeaters for a waveguide transmission system with very good performance.

Reference 4 gives the measured available gain* for the twelve field evaluation test repeaters. It ranged from 78.2 dB at 40 GHz to 64.6 dB at 109 GHz for these two-phase repeaters. Subsequent to fabrication of

* Available gain is defined for any channel as the dB ratio between the receiver input power required to attain an error rate of 1 in 10^9 and the transmitter output power.

the field-evaluation test repeaters, laboratory experiments have shown that considerable circuit performance improvement can be obtained, particularly in the IMPATT oscillators and down-converters. Based on these improved circuits, it is estimated that four-phase repeaters can be built to yield available gain ranging from 75 dB at 40 GHz to 58 dB at 110 GHz.

It is these projected four-phase repeater available gain numbers that have been used in system installation plans for repeater spacing.

XII. ACKNOWLEDGMENTS

The design of a millimeter-wave regenerative repeater is a mixture of a wide variety of electronic technologies and depends on the contributions of a large number of engineers. The authors would like to thank all those who contributed to the development of the repeaters.

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The power supply development was carried out by Herb Stocker, and we are especially appreciative of his cooperation in the WT4 development.

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