A High-Capacity Digital Communication System Using TE_{01} Transmission in Circular Waveguide

THOMAS A. ABELE, SENIOR MEMBER, IEEE, D. A. ALSBERG, SENIOR MEMBER, IEEE, AND P. T. HUTCHISON, MEMBER, IEEE

(Invited Paper)

Abstract—**In this paper, we describe the characteristics and several parts of a high-capacity digital communications system using the low-loss circular waveguide.**

I. INTRODUCTION

AS EARLY AS 1940, engineers were predicting that the special loss characteristics of the TE_{01} mode in circular guide would make it suitable as the transmission medium for long-distance communication systems. Research and development on communication systems built around the use of the TE_{01} mode have been active for a number of years [1]. At the present time, China, England, France, Germany, Italy, Japan, Russia, and the United States are engaged in work on such systems. The different systems use waveguides varying from 40 mm to 70 mm in diameter. The bandwidths of the different systems vary widely, but all fall within the limits of 35–120 GHz. Because of the large available bandwidth and the somewhat limited power available at these frequencies, digital systems have been given most attention. In this paper we shall describe the characteristics of the system called WT4 being developed at Bell Laboratories, the characteristics of the waveguide medium, the channelizing networks, and the regenerative electronic repeaters.

II. GENERAL FEATURES OF THE SYSTEM

The waveguide medium which is described in detail in Section III is a circular waveguide 60-mm inside diameter and supported in a protective steel sheath by roller spring supports. The guide and sheath are nitrogen pressurized and buried at 4-ft minimum depth. A system bandwidth of 40–110 GHz allows us to have 120 broad-band channels plus adequate guard bands. Sixty of these channels in the lower half of the band provide communication in one direction and sixty channels in the upper half of the band provide communication in the opposite direction. Three channels in each direction are used for protection switching and maintenance. Each broad-band channel carries a 2-level 274-Mbit/s bit stream or 4032 digitally encoded voice circuits. Thus, the overall capacity of a fully loaded system is almost 230 000 two-way voice circuits. Repeater stations which house regenerative electronic repeaters are aboveground buildings with a maximum spacing of 25 mi. Commercial and back-up power are furnished at each repeater station. Signals associated with protection-switching, command and telemetry, and order-wire systems are carried over specified broad-band communication channels in the waveguide in dedicated time slots in the 274-Mbit/s bit stream.

The Fig. 1 block-diagram representation of the multiplexing networks shows how the overall band is split into 7 subbands each with a bandwidth less than 15 percent by band diplexers which operate in the TE_{01} mode. Each subband is further divided by tandem connected channel diplexers which are connected to the repeaters. Low-pass filters shown are required to control harmonics because the system bandwidth is greater than one octave.

Fig. 2 shows for a 25-mi repeater span how much loss is caused by the waveguide medium and the diplexers and how much excess power we have at each frequency. The allowable loss curve is based on a transmitter output power that varies from 105 mW at 40 GHz to 13 mW at 110 GHz and a receiver noise figure that varies from 11 dB at 40 GHz to 13 dB at 110 GHz. The allowable loss curve is also based on having an adequate signal-to-noise ratio to have an error rate no worse than 10^{-9} under conditions of significant impairments.

In the following sections, we describe the characteristics of the waveguide medium, the channelizing networks or diplexers, and the repeaters. Because of space limitations, we shall not describe the protection-switching system, the fault-location system, the command and telemetry system, the order-wire system, the equalization test set, or the power supplies.

III. THE WAVEGUIDE TRANSMISSION MEDIUM

It is well known that the loss of the TE_{01} mode in waveguide, which is a perfect right circular cylinder, can be made insignificantly small by using an overmoded guide. However, in the real world, the presence of manufacturing imperfections in the guide and the necessity of having non-straight waveguide runs cross country introduce extra

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T. A. Abele is with Bell Laboratories, North Andover, Mass.
D. A. Alsberg is with Bell Laboratories, Murray Hill, N. J. 07904.
P. T. Hutchison is with Bell Laboratories, Holmdel, N. J. 07733.

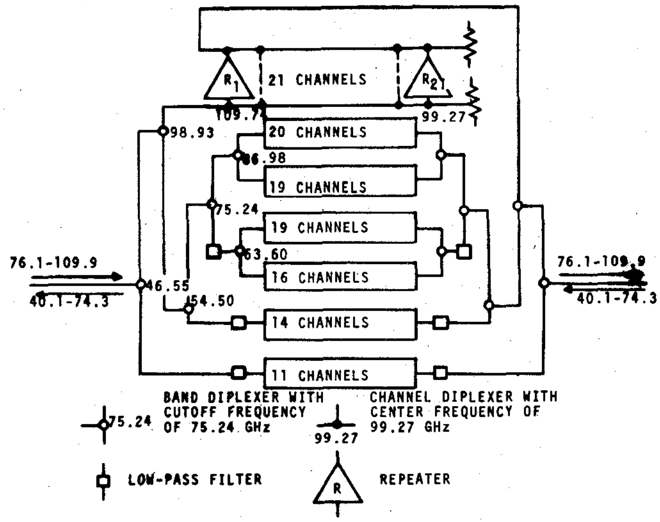

Fig. 1. Multiplexing plan.

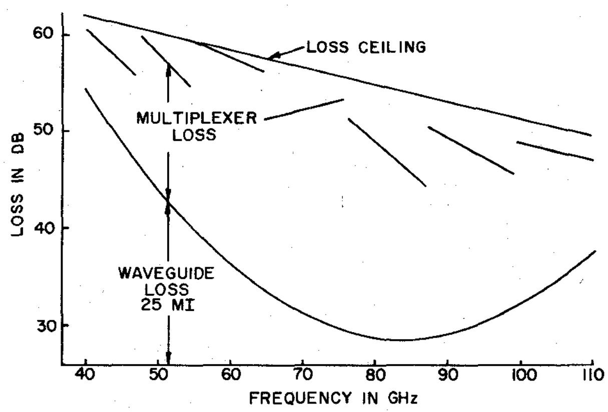

Fig. 2. Transmission-loss utilization.

loss due to mode conversion. To control this loss and realize a viable system, a number of key problems are being solved. These are as follows.

1) Understanding and controlling the manufacturing processes which introduce deviations from the desired circular cylindrical-waveguide geometry. These deviations cause mode conversion and reconversion, resulting in increased signal loss and transmission distortion since the waveguide can support a large number of undesired modes.

2) Minimizing the signal loss due to route bends. Route bends are needed to accommodate property lines and terrain features.

3) Developing installation techniques which are economical and reliable and which do not appreciably increase the waveguide mechanical distortions.

Mechanical gauges have been developed which can map the interior surface of waveguide tubing to the desired detail with a radial resolution of 1 μm. Using a computer, these measurements are processed to estimate electrical performance and to identify the manufacturing process adjustments needed to remove tubing distortions which

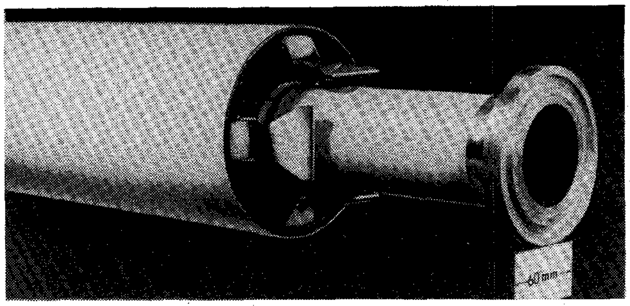

Fig. 3. View of sheath, waveguide flange, and roller spring support.

cause degradation in transmission performance. As a result of this work, there has been a continuous improvement in waveguide-tubing geometry, with the mode-conversion component of waveguide loss having been reduced to the extent that the waveguide diameter was increased from the original 51-mm design to 60 mm. This change substantially decreased the total loss at 40 GHz while increasing the loss slightly at 110 GHz. When other system losses are considered, this arrangement tends to make the performances of the individual channels more nearly equal.

The major causes of TE_{01} loss in route bends is coupling to the TM_{11} mode which is degenerate with the TE_{01} mode in a waveguide with perfectly conducting walls. A thin dielectric lining (180 μm) applied to the 60-mm waveguide removes the degeneracy between the TM_{11} and TE_{01} mode [2]. A recent result which has been demonstrated both theoretically [3] and experimentally is that the liner also reduces the loss of the TM_{11} mode at high frequencies, thereby further reducing the loss penalty for negotiating route bends.

The geometric distortions associated with installing the waveguide underground are minimized by housing the waveguide in a 140-mm-diam steel pipe casing (or sheath) and suspending the guide on roller spring supports as shown in Fig. 3. The combination of sheath and waveguide stiffness and the roller spring supports act to reduce the transfer of trench irregularities to the waveguide.

To eliminate mode-conversion effects associated with expansion joints and to increase reliability, expansion joints are not used in the medium. Instead, the guide is anchored firmly at repeater-station entrances and is permitted to experience either tensile or compressive loading under ambient temperature variations. The roller supports are spaced close enough to prevent buckling of the guide under compression loading.

Although mode-conversion effects have been greatly reduced, additional TE_{12}- and TM_{11}-mode filtration is still required to control transmission distortion. Therefore, filters of the well-known helix type [4], 9 m in length, are installed at approximately 800-m intervals.

Field installation utilizes waveguide modules consisting of 9-m lengths of waveguide with roller spring supports. These modules are welded together in the field and are inserted at discrete points along the right-of-way

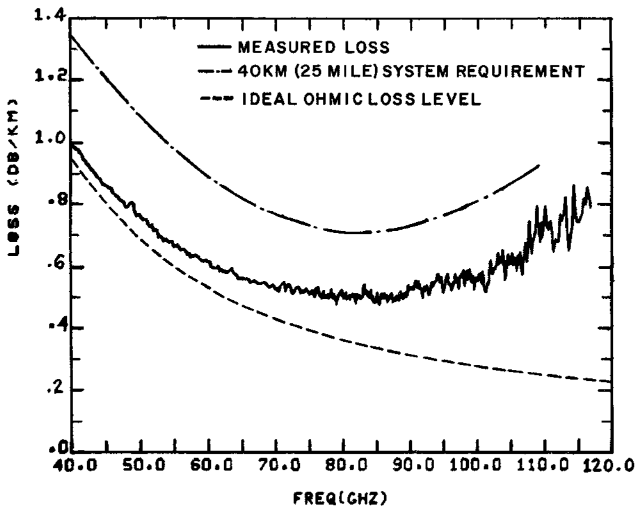

Fig. 4. Loss of 60-mm-diam waveguide with a copper wall and with 165-μm polyethylene liner. Measured waveguide loss: 20-km round trip.

into a previously buried pipe to form a continuous waveguide string. Insertion distances up to 1 mi are contemplated. Welding provides the gas tightness and reliability required by the medium. The waveguide is filled with dry nitrogen to avoid the molecular absorption bands of oxygen and water vapor which fall within the WT4 transmission band.

Fig. 4 shows the measured 20-km round-trip waveguide loss on the first 10-km waveguide section installed in New Jersey going south from the Netcong main repeater-station terminal. This section contains 5 $\sim$ 45° and 5 $\sim$ 90° route bends with an average radius of curvature in the order of 100 m and, in the first kilometer, the route traverses some very steep and rocky terrain. The average mode-filter spacing is 840 m. Also shown for reference are the ideal theoretical ohmic loss of the dielectric-lined guide (180-μm polyethylene) with a perfectly smooth intrinsic copper lining and the waveguide loss design limits (from Fig. 2) for a 40-km (25-mi) repeater spacing, including margin for future system expansion from a 2- to 4-level bit-stream operation. The ripple structure shown in the measured data is primarily due to random addition of various mode-conversion components and therefore decreases as the square root of the distance increases. In extrapolating the measured 20-km waveguide round-trip performance to the 40-km design repeater spacing, the ripple structure shown would be expected to decrease by about the $\sqrt{2}$.

IV. DIPLEXERS

A. Band Diplexers

Of the various ways known for dividing up the very broad frequency spectrum of the WT4 system into smaller subbands, we have chosen the Mach–Zehnder interferometer type described in [5]. As may be seen from the sketch shown in Fig. 5, it consists of two hybrids, two elbows, four helical tapers, two high-pass filters, and a matched load. All ports are in 50.8-mm-diam circular waveguide employing the TE_{01} mode. As a result of the frequency selectivity of the high-pass filter, a signal entering at port 1 is split into two frequency bands, one below the cutoff frequency of the high-pass filter, appearing at port 2, and one above the cutoff frequency of the filter, appearing at port 3. There are six codes of band diplexers, each with a different cutoff frequency as shown in Fig. 1. The hybrids are optimized, in thickness and dielectric constant of their dielectric sheet, for best overall band-diplexer performance. There are different hybrids in the various band diplexers and in some cases two different hybrids in a single band diplexer. An elbow is an integral part of a hybrid; both are machined from aluminum blocks. The dielectric materials used are aluminum oxide, mullite, and beryllium oxide. An air bypass is provided via a hole in the dielectric sheet. The helical tapers are the same for all band diplexers. They are designed according to [6] for mode conversion to the TE_{02} mode not to exceed 40 dB. The high-pass filters are designed essentially according to [7] and made from copper. The matched load consists of a tapered piece of a loaded, epoxy-based material in a 50.8-mm-diam waveguide with a short at the end. Fig. 6 shows typical performance characteristics of a band diplexer.

B. Channel Multiplexers

The subband signal available at the output of a band diplexer is first passed through a helical taper identical to the ones used in the band diplexers and then through a transducer to a semicircular guide still employing the TE_{01} mode but small enough in size so that the TE_{02} mode cannot propagate in this guide within the frequency limits of the particular subband. The channels, each 475 MHz wide, are then split from this guide by a channel multi-

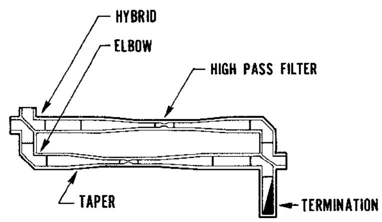

Fig. 5. Cross-section view of a band diplexer.

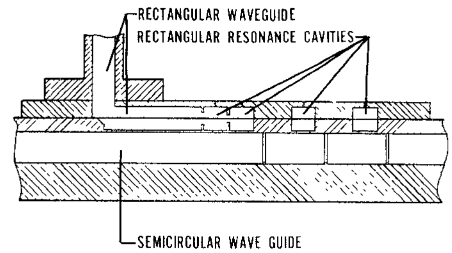

Fig. 7. Cross-section view of a channel diplexer.

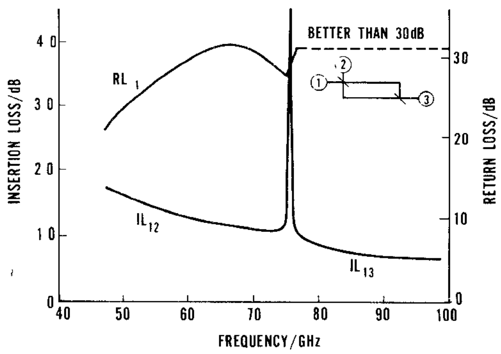

Fig. 6. Typical band-diplexer performance.

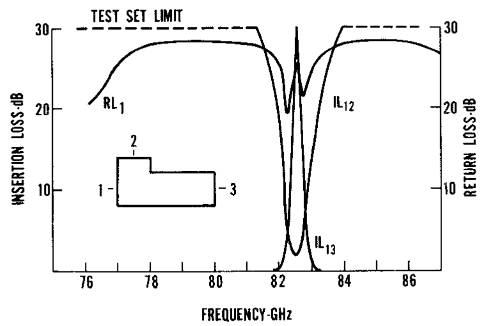

Fig. 8. Typical channel-diplexer response.

plexer, which consists of a cascade (11–21 depending upon the subband) of channel diplexers terminated in a matched load at the end.

The cross-section view of a channel diplexer is shown in Fig. 7. It is a 4-cavity waveguide diplexer similar to the one described in [8], but the second bandpass cavity is coupled to the first one in a different manner. The disturbances created by the three coupling apertures are compensated by three metallic rings in the circular portion of the main waveguide.

Since there are 120 channels, there are 120 codes of channel diplexers with center frequencies ranging from 40.235 to 109.765 GHz. Channels are separated by 525 MHz except in frequency regions that include guard space between subbands. The dimensions of these diplexers have been determined by making maximum use of computer-aided analysis of the cavities, apertures, and obstacles and by resorting to experimental techniques only where necessary. Fig. 8 shows typical performance characteristics of a channel diplexer.

The transducer from circular to semicircular guide needed at the input of each multiplexer must obviously be different for each subband. Hence there are seven codes. The transducers are made from copper using an electroformed outer shell and a machined insert. Measured insertion losses are typically 0.4 dB at 100 GHz.

C. Low-Pass Filter

This filter is a simple varying impedance transmission-line filter consisting of a cascade of spaced dielectric disks in 50.8-mm circular guide as shown in Fig. 9. Since this is an oversized guide, the filter provides the same frequency

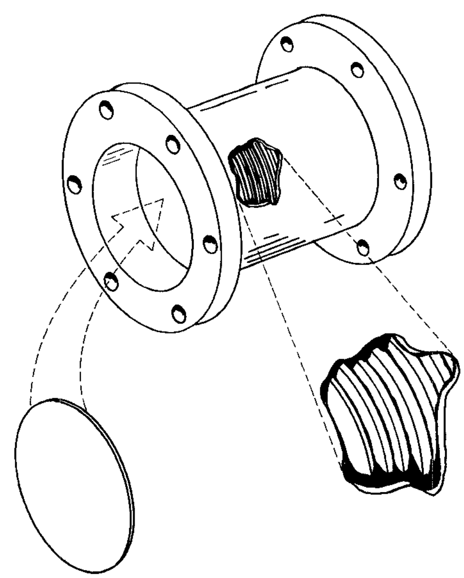

Fig. 9. Low-pass filter.

selectivity characteristics to about the first ten modes; TE_{11} is the first mode and TE_{01} is the third mode. Two codes are required that have cutoff frequencies of 73.5 and 75.2 GHz. The dielectric material employed is fused silica in thicknesses ranging from 5 to 20 mils. An air bypass is provided outside of the filter. Fig. 10 shows a typical performance characteristic.

V. REPEATER

A. General

A regenerative repeater is made in four parts: a receiver, a line equalizer, a transmitter, and a power supply. Fig. 11

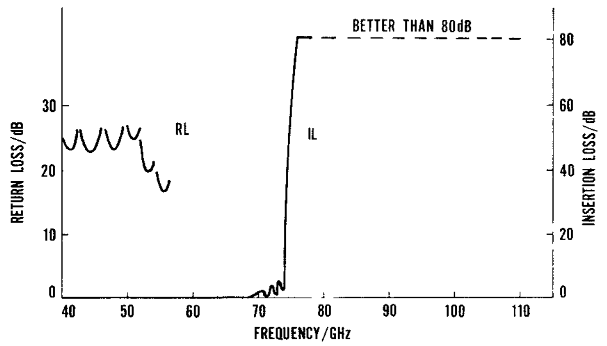

Fig. 10. Low-pass filter response.

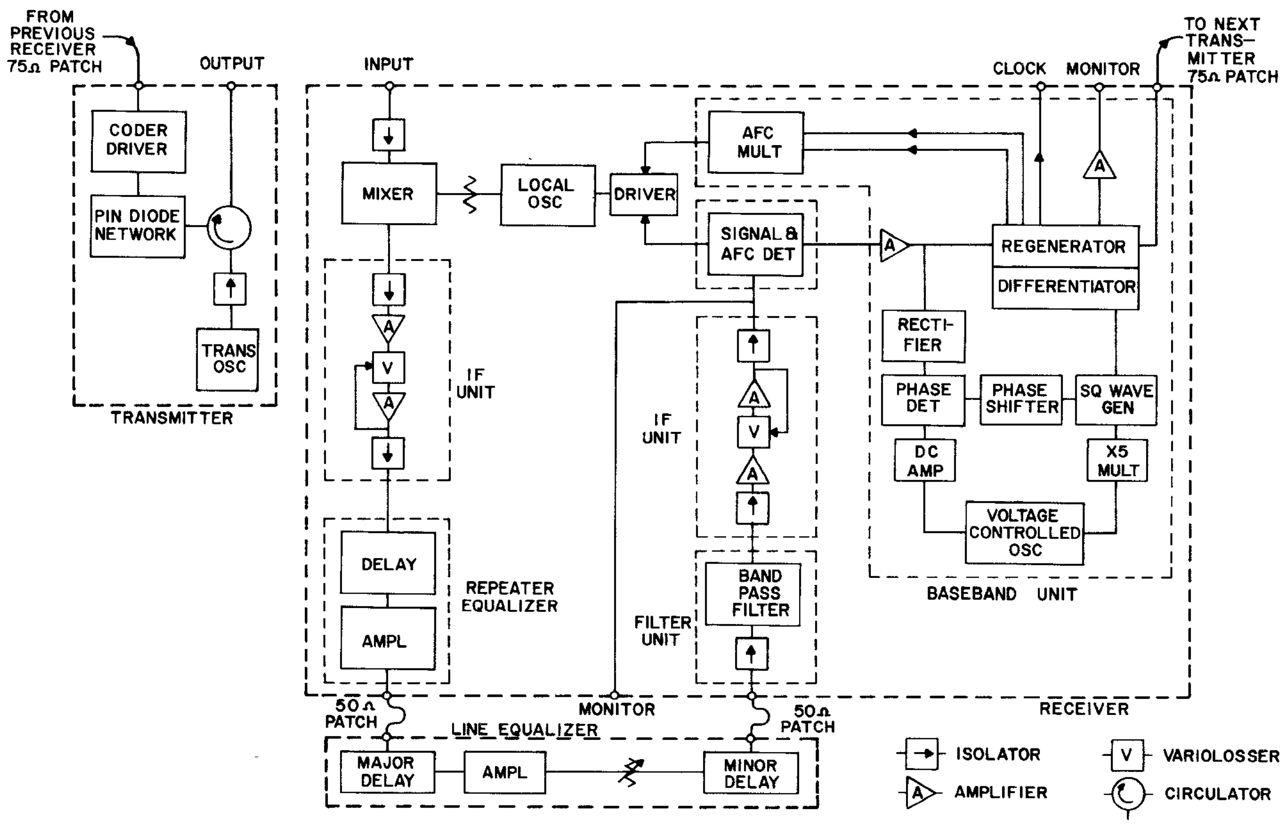

Fig. 11. Regenerative repeater.

shows a block diagram of a repeater less the power supply. We shall describe the repeater by following a signal from the transmitter through the next receiver. Since many parts of this heterodyne repeater are conventional, emphasis will be placed on the parts that are new.

A 274.176-Mbit/s 2-level bit stream from multiplex equipment or from a previous receiver is fed into a coder in the transmitter. The coder is a bistable multivibrator which responds only to negative pulses. Changing the state of the multivibrator changes the conducting state of a millimeter-wave p-i-n switching diode, which is fed by a silicon IMPATT oscillator through a circulator. The p-i-n diode and its associated network are adjusted so the phase difference between the reflected signals in the conducting and nonconducting states of the diode differ by 180°.

Millimeter-wave output power from the transmitter then goes through the 25 mi of waveguide and the associated diplexer networks to a receiver at the following repeater station.

As Fig. 11 shows, the incoming signal goes through an isolator which keeps local-oscillator leakage into diplexer networks to a low level. It is then shifted to a center IF of 1371 MHz in a double-balanced orthomode mixer using beam-lead diodes. After the signal is amplified in a hybrid-integrated-circuit IF amplifier, it goes into a repeater equalizer which is factory adjusted to compensate over a 400-MHz band for transmission distortions caused by circuits in the transmitter–receiver pair. The line equalizer which compensates for the transmission distortion caused by the diplexer networks and the 25 mi of waveguide must of course be adjusted in the field after the repeater has been equalized. If the line and repeater equalizers were com-

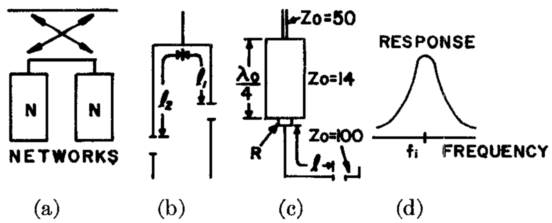

Fig. 12. Single-section equalizer. (a) Configuration. (b) Delay. (c) Amplitude. (d) Response.

bined, it would be necessary to take an equalization test set into the field each time a faulty repeater was replaced. The line equalizer is passive so its failure should be a very rare occurrence. Details of the equalizer are covered in Section V-B.

The IF filter is a maximally flat 4-section filter with a noise bandwidth 1.35 times the baud frequency f_b. Computer studies showed that the shape of the filter required to meet the Nyquist criteria for no intersymbol interference has rather sharp peaks near the edges of the passband. This filter was not used because it has a noise bandwidth greater than 1.35 f_b and its delay distortion is large. Therefore, we use a simple filter which has a bandwidth that is a compromise between minimum noise bandwidth and minimum intersymbol interference. After more IF amplification, the signal goes to a differentially coherent detector. This type of detector is used since it is simpler than the coherent detector and the signal-to-noise (S/N) penalty for its use in a 2-phase system is very small. Details of the signal and AFC detectors are covered in Section V-C.

All IF circuits in the repeater are microstrip hybrid integrated circuits with beam-lead semiconductors and with some beam-lead capacitors, values between 10 and 3000 pF, and some chip capacitors. The proper IF amplifier response is obtained by adjusting variable capacitors and laser trimmed inductances. The relatively poor return loss of the single-ended IF amplifiers can be tolerated because lumped-element isolators are used where needed.

The detected signal is then fed through a baseband amplifier into the regenerator and timing recovery circuits. The 274-Mbit/s bit stream is made up of 6 time-division-multiplexed signals which have been individually scrambled and a few housekeeping bits which have not been scrambled. This means the average signal level is zero, so we do not have to use direct coupling in our baseband circuits and the AFC circuit is conventional. The scrambling also ensures our being able to recover the timing signal without having to penalize the system by using timing bursts. The timing-recovery circuit shown in Fig. 11 is a conventional phase-locked loop. It was convenient to design the regenerator circuit to furnish negative pulses at two output ports, one output corresponds to positive pulses in the input bit stream and the other output corresponds to negative pusles in the input bit stream. Only one line is needed between the receiver and the transmitter since the coder in the transmitter reacts only to negative input pulses. Regenerator output pulses are approximately 1.1-V peak, into 75 Ω, and 0.7 ns wide at the 10-percent height points. The same general construction that is used in the IF circuits is also used in the baseband circuits, i.e., hybrid integrated circuits in microstrip form.

Each repeater requires approximately 50-W dc power. We estimate that the mean time to failure for a repeater will be 15 years.

B. Equalizers

The repeater delay and amplitude equalizers are each made up of three independent "bump" equalizers, shown in Fig. 12, that can be adjusted so that the overall response, which is the sum of the three responses, is a good approximation to a wide variety of desired responses. The networks connected to the 3-dB coupler must be essentially identical to get a good input and output return loss to minimize interaction between sections of the 3-section equalizer. The response of a single network, i.e., delay or loss, will have a predictable shape similar to that shown in Fig. 12 (d). Fig. 12 (b) shows the type network that is used in a delay equalizer. The sum of l_1 and l_2 is made to be a half-wavelength at some frequency f_i and the ratio of l_1 to l_2 is used to control the height of the response. Fig. 13 is an example of how well this approach works for a repeater equalizer.

To obtain amplitude or loss shaping, we use the network shown in Fig. 12 (c). It has a high-impedance (Z_0 = 100 Ω) line connected to a low-impedance line (Z_0 = 14 Ω) through a series resistance that is laser trimmed to some value between 10 and 50 Ω. The lines from the coupler to the 14-Ω sections have 50-Ω impedance. The low-impedance line is one-quarter wavelength at midband. A loss peak can be obtained at a certain frequency f_i by making the 100-Ω line one-quarter wavelength from the resistor to the open circuit. The magnitude of the loss is controlled by laser adjustment of the series resistance. Thus three of these bump loss equalizers are used to approximate the required loss curve. Since amplitude equalizers have some delay distortion and delay equalizers have some amplitude shaping, the actual equalization is an iterative process and is easily handled by the computer.

The line equalization is more difficult because significant delay distortion is introduced by the 25 mi of waveguide. Calculations show that over a 400-MHz band this delay is essentially linear and has a value varying from 30.5 ns at 40 GHz to 1.4 ns at 110 GHz. The line equalizer must also

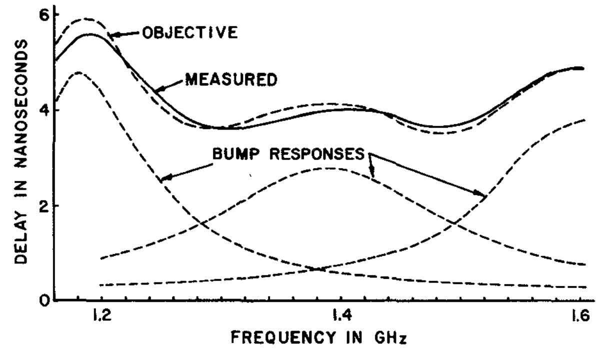

Fig. 13. Delay-equalizer characteristics.

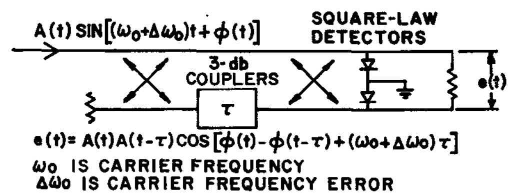

Fig. 14. Differential phase detector.

compensate for delay due to shaping in the millimeter-wave diplexers and ripples caused by mode conversion plus reconversion in the guide and in the diplexers. We will have to control the magnitudes of the fast ripples due to conversion and reconversion since it is impractical to equalize them. In order to minimize the number of codes of line-delay equalizers, we have chosen to use a family of eight fixed linear delay equalizers to compensate for most of the large linear delay. These equalizers are folded tape meander lines [9] having linear delays of 0, 4, 8, 12, 16, 20, 24, and 28 ns over a 400-MHz band. One of these major delay equalizers and one minor delay equalizer will be used to equalize each channel. The minor delay is a 3-bump network like the repeater delay equalizer. It must provide a linear delay component of no more than $3\frac{1}{2}$ ns over a 400-MHz band plus the delay ripples required. Thus a major and a minor equalizer together provide any linear delay to an accuracy of $\pm\frac{1}{4}$ ns at a frequency $f_b/2$ from the center IF.

The line amplitude equalizer is very similar to the repeater amplitude equalizer.

C. Signal and AFC Detector

A simplified diagram of a delay-line discriminator or differential phase detector is shown in Fig. 14. To make this circuit a signal detector, the delay τ is adjusted to be 1-Bd period $(1/f_b)$ and the IF center is made an integer multiple of f_b. We have chosen the center frequency to be $5 \times 274.176 = 1370.88$ MHz, so $\omega_0\tau = 5 \times 2\pi$. With these conditions and no carrier frequency error $(\Delta\omega_0 = 0)$, the output signal from Fig. 14 is

$$e(t) = A(t)A(t-\tau)\cos[\phi(t)-\phi(t-\tau)]. \tag{1}$$

For a system with zero switching time of the phase shifter, no band limiting, and no system impairments, $\phi(t) - \phi(t-\tau)$ is always 0 or π and the detector output is always $+1$ or -1. For a practical system, the detector output varies more slowly because of bandwidth restrictions and diode switching times (about $1/5f_b$), but the detector output is still very nearly $+1$ or -1 over periods about $1/3f_b$ and can easily be regenerated to recreate the original bit stream.

To make an AFC detector, we adjust the delay τ such that $\omega_0\tau = 10\pi + \pi/2$, the detector output becomes

$$e(t) = -A(t)A(t-\tau)\{\sin[\phi(t)-\phi(t-\tau)]\cos\Delta\omega_0\tau + \cos[\phi(t)-\phi(t-\tau)]\sin\Delta\omega_0\tau\}. \tag{2}$$

In an idealized system $\sin[\phi(t)-\phi(t-\tau)] = 0$, so an AFC error signal can be obtained by multiplying the cos $[\phi(t)-\phi(t-\tau)]$ by $+1$ when $\phi(t)-\phi(t-\tau) = 0$ and by -1 when $\phi(t)-\phi(t-\tau) = \pi$, since $A(t)A(t-\tau)$ is always positive or zero [10]. The multiplying signal of $+1$ and -1 is already available; it is the output of the signal regenerator which by design is $+1$ when $\phi(t) - \phi(t-\tau) = 0$ and -1 when $\phi(t)-\phi(t-\tau) = \pi$. The regenerator output pulses are very short, so the output of the AFC multiplier is a series of short pulses given by

$$\text{error} = F(t)\sin\Delta\omega_0\tau. \tag{3}$$

The function $F(t)$ is the Fourier-series expression for a series of positive narrow pulses with a fundamental frequency f_b. In a practical system with transmission distortions, the AFC multiplier output is essentially the same as for an ideal system, since the regenerator output pulses occur when $\cos [\phi(t) - \phi(t - \tau)]$ is almost $+1$ or -1.

VI. STATUS OF DEVELOPMENT

Personnel from Bell Laboratories and the Western Electric Company Engineering Research Center are working together on process development of the waveguide. This has resulted in precision waveguide with manufacturing irregularities far better than we anticipated. This reduces the mode-conversion losses so much that we were able to increase the diameter from 51 to 60 mm and thus decrease the heat loss in the waveguide. The Engineering Research Center has made over 10 000 m of waveguide that has excellent loss characteristics.

We have made and tested a band-diplexer array made up of a complete set of band diplexers and low-pass filters. These tests showed no trapped modes and allowed us to determine what lengths of helix waveguide are required between band diplexers to absorb the TE_{21} and TM_{21} modes generated in the hybrids. Satisfactory models of the circular to semicircular transducers and the 100-GHz channel diplexers have been completed.

We have completed four repeaters and are building eight more for use in a field evaluation. There will be three of these third-generation repeaters operating near each of the following frequencies: 40, 55, 80, and 110 GHz. Our best repeaters operated back to back through a precision attenuator have measured 10^{-9} error rates at signal-to-noise ratios within 0.2 dB of the calculated value and 2 dB of the ideal value.

In late 1974, we plan to start a one-waveguide span Field Evaluation Test to evaluate the characteristics of the waveguide medium, the waveguide installation procedures, and the behavior of the electronics. With twelve repeaters operating at the different frequencies mentioned previously, we can simulate a twelve-hop system. This evaluation will continue into 1976.

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