Experimental Transversal Equalizer for TD-2 Radio Relay System

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To determine the effect of improved equalization on the performance of TD-2 radio relay systems, an experimental adjustable transversal equalizer has been developed. The equalizer is based on the echo principle as used in transversal filters, and operates in the 60- to 80-mc frequency band. Seven pairs of adjustable leading and lagging echo terms provide flexibility for simultaneous gain and delay equalization. Directional couplers are used for tapping and controlling the echo voltages. Field experiments have shown that system equalization can be improved appreciably by the use of such equalizers.

INTRODUCTION

The TD-2 radio relay system[1] employs frequency modulation to transmit multichannel telephony or television. The frequency modulated signal requires a transmission system whose gain and envelope delay are constant over the frequency band, 20 mc wide, used to transmit the signal. Deviations from constant gain or delay result in non-linear distortion of the demodulated signal. For television this results in distorted images, and for multichannel telephone transmission this introduces cross modulation among the voice channels.

Basic equalization is provided in each repeater. In addition, certain fixed equalizers have been used on a mop-up basis. However, there remains some residual distortion of random shape. This paper discusses the design of an adjustable equalizer to compensate for this distortion.

I. BASIC EQUALIZATION

The transmission path through a TD-2 repeater consists of an intermediate frequency portion covering the 60- to 80-mc band and radio frequency channels 20-mc wide in the 3,700- to 4,200-mc band. At each repeater the RF channels are separated by wave guide filters and each is demodulated to the IF frequency for amplification and equalization. The outgoing signal is then modulated back to an RF channel, at a different frequency from the incoming signal to reduce interference. The RF and IF portions of the repeater were designed so that the combination would have flat gain to within the closest practicable limits over the 20-mc band, and a number of adjustments are provided in the IF amplifier to maintain this flatness under field conditions. The unequalized repeater, however, has an envelope delay distortion characteristic shown in Fig. 1 which is approximately parabolic. To minimize this distortion, each repeater contains a 315A equalizer, which has approximately the inverse of the delay distortion of a typical repeater.

II. SUPPLEMENTARY EQUALIZATION

In a TD-2 system consisting of many repeaters in tandem, both gain and delay distortions may accumulate to the point where additional equalization is necessary. If the pass band of the repeaters shifts slightly in frequency, due to changes in temperature or adjustment, the difference between the repeater delay and the delay of its equalizer will result in delay distortion which has approximately a linear slope with frequency. This may be corrected at main repeater stations by combina-

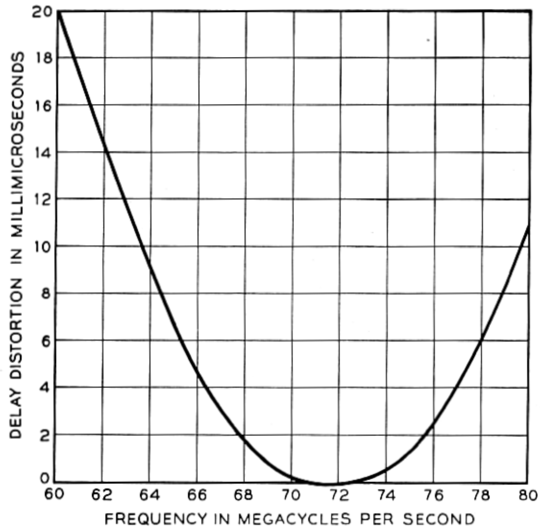

Fig. 1 — Over-all delay distortion of a typical microwave repeater, TD-2 system.

tions of delay slope equalizers. The characteristics of the 319A, B and C equalizers provided for this purpose are shown in Fig. 2. Each equalizer consists of two bridged-T all-pass sections. There is also some variation in bandwidth of the TD-2 repeaters, resulting in part from the fact that the waveguide filters used in the higher frequency radio channels are somewhat broader than those in the lower frequency channels. This variation in bandwidth results in delay distortion which has a parabolic shape with frequency. Use of a larger or smaller number of the basic 315A equalizers corrects this.

Over long circuits, small distortions in the gain shape of the TD-2

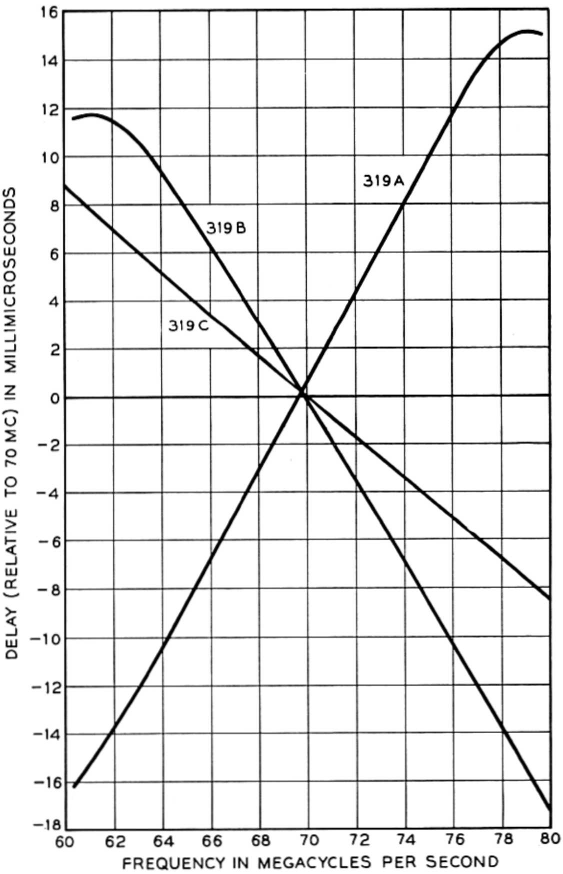

Fig. 2 — Delay characteristics of 319 type equalizers. Combinations of these are used at main stations to equalize delay slope of the system.

equipment produce a cumulative gain-frequency distortion which is noticeable in television circuits. Present practice is to correct for this by standard video equalizers after the FM signal has been demodulated to baseband. In connection with the experimental equalizing program to be described, parabolic gain equalizers operating on the FM signal before demodulation were used.

III. RESIDUAL DISTORTION

After correction of the known shapes discussed above, there remains a certain residual gain and delay distortion which results from a random summation of many minor sources. The shape of this distortion is not predictable, but its statistics are known. Examination of typical delay versus frequency characteristics have shown that these may be reasonably well approximated by six cosine terms: a 40-mc fundamental and the next five harmonics. Similar gain terms are needed. However, the gain and delay distortion, when examined within the 20-mc band of interest, do not have a minimum phase relationship. This is to be expected because of the presence in the system of the delay equalizers, which are non-minimum-phase networks, and of amplifiers with compression.

The magnitude of the residual distortion is small enough so that transcontinental TD-2 circuits provide television and telephone transmission of commercial quality. Some effects, such as cross modulation, are sufficiently marginal so that improvement would be desirable. To determine whether this could be achieved by improved gain and delay equalization, the development of an experimental adjustable equalizer was undertaken. The considerations outlined show that such an equalizer should approximate the desired characteristics with independent gain and delay terms of the harmonically related cosine type. Equalization to reduce cross modulation in telephone channels and differential phase in color television must be performed before demodulation of the FM signal to base band. The equalizer was, therefore, built to operate in the 60- to 80-mc IF band.

IV. TRANSVERSAL EQUALIZER

One method of obtaining independent control of the loss and delay characteristics of a network has been achieved in the transversal filter.[2] Equalizers have been designed on this principle for the equalization of television circuits.[3, 4] This type of equalizer, referred to here as a transversal equalizer, provides a flexible means of synthesizing any loss characteristic and any delay characteristic limited only by the number of harmonics that are provided and the range of each.

Basically, the transversal equalizer consists of a delay line with equally spaced taps, with a means for independently controlling the amount of signal fed through each of the taps to a summing circuit, as shown schematically on Fig. 3. The input signal is fed into one end of the delay line which is terminated at the other end. The center tap is fed to the output and forms the main transmission path.

The operation of the equalizer can best be described using the "time domain" analysis based on the theory of paired echos.[5] Portions of the signal tapped off the "leading" or first half of the delay line will not be delayed as much as the main signal and will introduce leading "echos". Similarly, lagging echos can be obtained from the taps on the lagging or second half of the delay line. Combinations of both types of echos, either positive or negative as required, can be added to cancel out, to a first approximation, distortion present in the input signal.

This analysis can also be carried out in the frequency domain. To obtain a family of cosine loss versus frequency characteristics without any appreciable delay characteristic, equal leading and lagging echos of the same polarity are added to the main signal in the summing circuit. To obtain a corresponding family of cosine delay versus frequency characteristics without loss distortion, leading and lagging echos equal

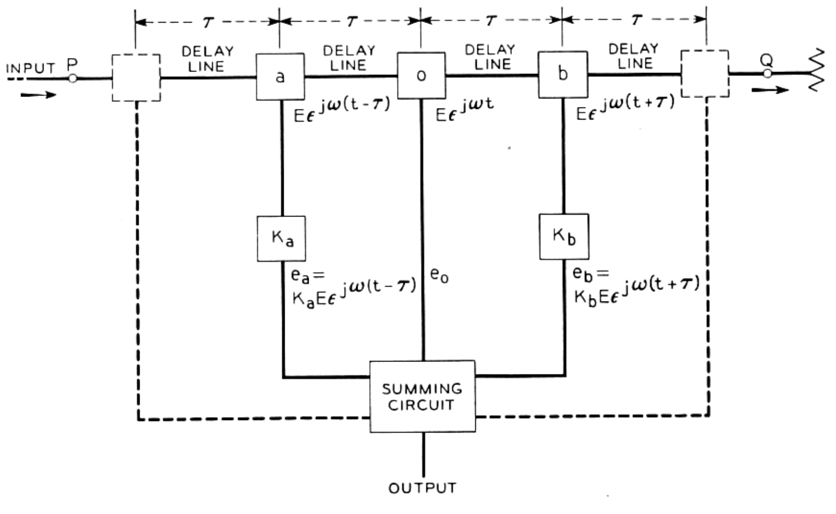

Fig. 3 — Block schematic of transversal equalizer.

in magnitude but of opposite polarity are added in the summing circuit to the main signal.

To achieve a practical equalizer for operation over the 60- to 80-mc band requires the following components: delay line or delay networks, means for tapping off small portions of the signal controlled both in amount and polarity, and a suitable summing circuit.

V. DERIVATION OF THE EQUALIZER CIRCUIT

A brief analysis of the operation of the equalizer will be given at this point as a basis for discussion of the method of tapping the signal and controlling the amplitude and polarity of the tapped portion.

Fig. 3 shows the basic delay line PQ as well as the means used for producing the main signal and a single pair of leading and lagging echos. The tap labeled "o" in the center of the line produces the main signal. The tap "a", being closer to the input, produces a signal which leads the main signal by time τ. The tap "b" produces a signal which lags the main signal by the same amount. The boxes "K_a" and "K_b" control the amplitude and polarity of the leading and lagging signals which are to be combined with the main signal to produce one term of the desired equalization characteristic. It will be shown that these three signals will provide one cosine gain term and one cosine delay term, both having the same period, but being independently controllable as to amplitude and polarity.

We will choose as our reference point for phase the main output signal, $e_0 = E\varepsilon^{j\omega t}$. The output from tap "$a$" is then

$$E\varepsilon^{j\omega(t-\tau)}.$$

After passing through box "K_a", this becomes

$$e_a = K_a E\varepsilon^{j\omega(t-\tau)}.$$

Similarly,

$$e_b = K_b E\varepsilon^{j\omega(t+\tau)}.$$

Here the terms K_a and K_b are of the form

$$K = \pm\varepsilon^{-\alpha}$$

where α is the attentuation in nepers of the box K. Note that $|K|$ is less than unity, assuming the box represents a passive network. Combining these two signals with the main signal, we have

$$e_r = e_0 + e_a + e_b = E\varepsilon^{j\omega t}[1 + K_a\varepsilon^{-j\omega\tau} + K_b\varepsilon^{j\omega\tau}]. \tag{1}$$

Now it will be shown that, by the adjustment of the two parameters K_a and K_b, it is possible to realize independent control of a cosine gain term and a cosine delay term.

Since, in general, $K_a \neq K_b$, let us define

$$K_a = K_g + K_p$$

and

$$K_b = K_g - K_p . \tag{2}$$

Then

$$e_r = Ee^{j\omega t}[1 + K_g(e^{-j\omega\tau} + e^{j\omega\tau}) + K_p(e^{-j\omega\tau} - e^{j\omega\tau})]. \tag{3}$$

Substituting the trigonometric form:

$$e_r = Ee^{j\omega t}[1 + 2K_g \cos \omega\tau + j\, 2K_p \sin \omega\varepsilon]. \tag{4}$$

Note that for K_a equal to K_b, the sine phase term is zero and that for K_a equal to $-K_b$ the cosine gain term is zero. Similarly, by proper proportioning of K_a and K_b, K_g and K_p may be assigned any desired values.

If we normalize (3) by setting $e_0 = E\varepsilon^{j\omega t} = 1$, the expression in brackets can yield two vector diagrams which are useful in explaining the functioning of the equalizer. To obtain the diagram shown in Fig. 4(a), we have set $K_p = 0$. We then have a unit vector, representing the main signal, a leading echo $K_g\varepsilon^{-j\omega\tau}$, and a lagging echo $K_g\varepsilon^{j\omega\tau}$. The

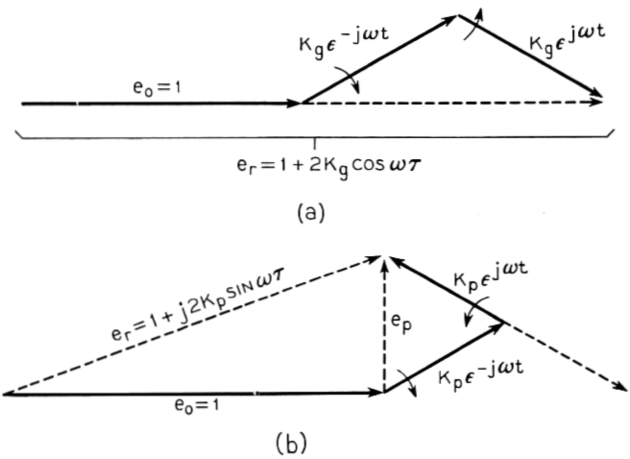

Fig. 4 — Vector diagrams of paired echos. (a) Equal echos of same polarity produce magnitude change without phase change. (b) Equal echos of opposite polarity produce a change in phase shift with a minor change in magnitude.

vector representing the leading echo rotates clockwise with respect to the main signal when the frequency increases, whereas the vector representing the lagging echo rotates counterclockwise by the same amount, and the resultant thus varies in magnitude but not in phase. The magnitude of the resultant is given, for this case, by the first two terms in parentheses in (4).

If, on the other hand, if we set $K_g = 0$, we have the three vectors shown in Fig. 4(b), identical with those in Fig. 4(a) except that the polarity of the lagging echo has been reversed. In this case, the two echos produce a resultant, e_p, which is in quadrature with the main signal. For small echos, e_r is thus shifted in phase from the main signal, with substantially no change in magnitude. The resultant in this case is given by the first and third terms in parentheses in (4). This gives a sinusoidal variation in the phase of the resultant. Since envelope delay is defined as $d\beta/d\omega$, where β is the phase shift through the circuit in question, the sinusoidal phase ripple will be seen to yield, after differentiation, a cosine delay ripple.

The period of the ripple can be seen from the above expressions to depend on τ, the delay between the leading and the main tap, and between the main tap and the lagging tap. Other pairs of echos, each pair symmetrically disposed about the main tap, but with different values for τ, will give transmission ripples of different periods. To provide a series of orthogonal terms, the values of τ must be integral multiples of a common value, normally that required to produce 180° phase shift across the band of interest.

A complete equalizer must, of course, sum up the various echos and the main signal, taking care that the delay between the tap and the summing point is the same for each echo and the main signal, that parasitic losses such as losses in cabling are the same for each path through the equalizer, and that any frequency characteristic in the tapping device or other parts of the equalizer is properly equalized out so that the over-all equalizer introduces a minimum of distortion of its own.

VI. DIRECTIONAL COUPLER

To reduce incidental distortion, it is desirable that the device used to tap the delay line for the main signal and the echos introduce substantially no discontinuity in the main line. The device chosen for this purpose is a directional coupler. It is shown symbolically in Fig. 5. The directional coupler is a four port device having properties similar to a hybrid coil. Power entering one port divides (not necessarily equally) between two other ports, but none of it reaches the fourth, or conjugate port. In Fig. 5, the power entering at 1 divides between 2 and 4, that entering at 2 divides between 1 and 3, that entering at 3 divides between 2 and 4, and that entering at 4 divides between 1 and 3. Directional couplers inherently provide an impedance match at all four ports. Thus, such a coupler sets up no reflections in the main line. Its insertion loss in this line may be kept small by having nearly all the power entering at 1 come out at 2; then only a small fraction is diverted to 4. Coaxial directional couplers have been discussed in the literature[6, 7, 8] and will not be dealt with in detail here.

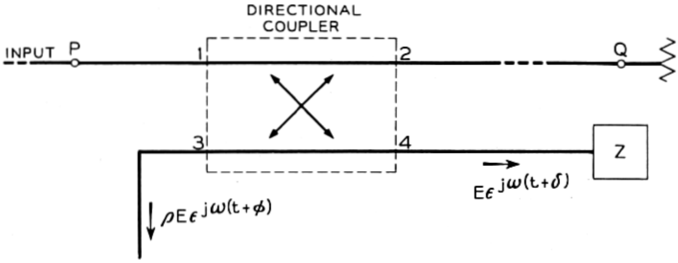

Fig. 5 — Diagram of directional coupler. Input signal divides between Ports 2 and 4 with no output at Port 3. Termination Z at Port 4 reflects some signal to Port 3, proportional to the reflection coefficient, ρ.

The coupler used here (J68333C) is one originally developed to measure reflections on IF transmission lines in the TD-2 system. The directivity of a coupler is defined as the coupling loss between main line and branch line in the undesired direction less the loss in the desired direction (loss from 1 to 3 less the loss from 1 to 4, for example). In the J68333C coupler, the directivity can be adjusted to exceed 45 db over the band of interest. This can be done by adjusting two screws, shown on model in Fig. 6, to obtain the optimum spacing between the coupling elements. The loss between the main line and the branch line in the desired direction is about 23 db at mid-band (70 mc), and decreases 6 db per octave with increasing frequency. The loss along one of the coupled lines (1 to 2 or 3 to 4) is very small.

Use has been made of the directional properties of the coupler in providing a simple means of controlling the amplitude and polarity of the tapped signal. Referring to Fig. 5, and keeping in mind the properties of the coupler, it will be noted that a small portion of the input signal appears at Port 4 of the coupler, but none at Port 3. If the impedance Z

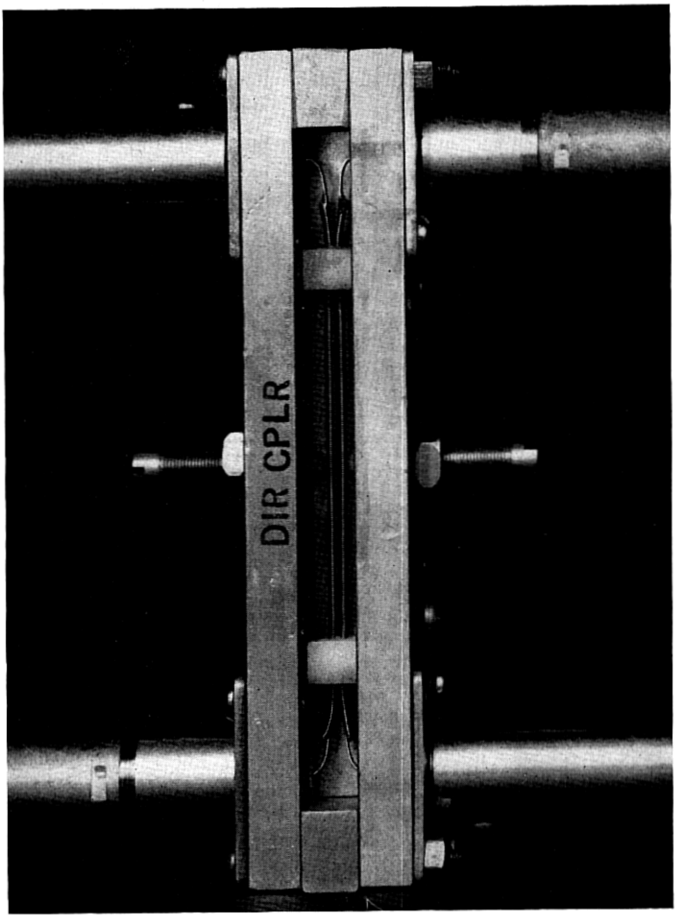

FIG. 6 — J-68333C directional coupler with cover plate removed to show coupling elements. Optimum spacing for maximum directivity can be obtained by adjusting screws.

matches the impedance seen looking into Port 4, all of the small signal will be absorbed in Z. If Z is an adjustable resistance, then a controllable portion of the small signal can be reflected back into Port 4, whence most of it will come out Port 3. A small portion of the reflected signal will emerge at Port 1, headed toward the input. This portion will be attenuated by twice the coupling loss between the main and the branch line plus the return loss of the reflection at Z, and can be made negligible. The interactions caused by the reflected signal on the main delay line entering previous couplers are also reduced by twice the coupler loss.

The coupler in Fig. 5 represents any one of the taps on the line PQ in Fig. 3. The signal emerging from Port 4 can be written as $E\varepsilon^{j\omega(t+\delta)}$, where δ is a time delay dependent on the location of the tap on the line PQ. The signal emerging from Port 3 is then $\rho E\varepsilon^{j\omega(t+\delta)}$, where ρ is the voltage reflection coefficient at Z, and is given by

$$\rho = \frac{R - R_0}{R + R_0}.$$

Here R is the value of resistance used to provide the impedance Z, and R_0 is the impedance seen looking into Port 4. An examination of the signal from Port 3 shows that it is the same as the signals e_a or e_b in Fig. 3, with the reflection coefficient ρ substituted for variable K_a or K_b in Fig. 3. Thus, it is seen that we may use the reflection at Z, variable by controlling the value of R, to perform the function of the box K in Fig. 3. Neglecting parasitic losses we may then write:

$$K = \rho = \frac{R - R_0}{R + R_0}$$

and

$$R = R_0 \frac{1 + K}{1 - K}. \tag{5}$$

This gives us the value of R to use for any desired value of K for any of the taps which derive echos, assuming the summing circuit has equal attenuation in all paths. In the case of the main central tap, the signal from Port 4 of the coupler is seen to be the same as the main signal e_0 in Fig. 3, and is used as such directly.

VII. METHOD OF ADJUSTMENT

The detailed design of a manually adjustable equalizer is materially influenced by the method to be used in the field for determining the setting of its controls. The present equalizer with 14 independent controls would present a complex problem of field adjustment unless special procedures were developed to simplify the adjustment. To adjust the equalizer, the radio circuit being equalized must be taken out of commercial service, so any reasonable measures to simplify the adjustment or reduce the time required are justified.

Two methods appeared to be feasible at the time the development was started. One would be to use existing gain and delay sweep test circuits. These present a visual display of the circuit gain or delay distortion versus frequency. These displays are not available simultaneously with present equipment. To adjust the equalizer controls using this equipment, it must be possible to adjust either gain or delay without affecting the other. Thus, all the gain terms can be adjusted in succession using the gain display. Then the procedure is repeated with the delay display, adjusting the delay terms. Since a combination of leading and lagging echos in equal amounts is required for this procedure, an arrangement of the controls to facilitate this is required. One way to achieve this is to use stepped rheostats with the steps proportioned to introduce equal amplitude changes in the echo voltage. With this arrangement, gain changes can be introduced by rotating the two switches corresponding to a pair of echos in the same direction an equal number of steps. Delay changes can be obtained by similar rotation in opposite directions. A further refinement consisting of mechanically ganging the controls is possible but this was not done on these experimental models.

The second method of adjustment would be to develop a special test set similar to the one developed for the L3 system.[9] This could produce a meter reading proportional to the amount of gain and delay distortion present in the circuit. Successive controls could then be adjusted for minimum meter readings. Experience with the L3 system cosine equalizers has shown the desirability of continuously adjustable controls for such a method.

To test both methods under field-trial conditions, two versions of the equalizer were built — one with stepped rheostats and one with continuously adjustable rheostats.

VIII. COAXIAL RHEOSTAT

Since there were no available continuously adjustable rheostats satisfactory for operation at 70 mc, a special rheostat was developed for

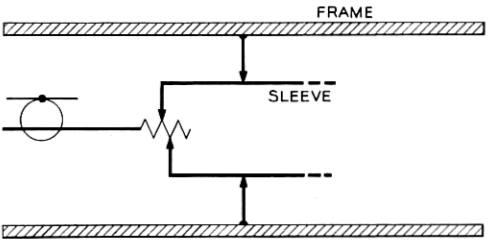

Fig. 7 — Schematic of coaxial rheostat. Moveable sleeve changes position of inner contacts touching ceramic rod, changing resistance. Fixed outer contacts maintain constant path length to frame.

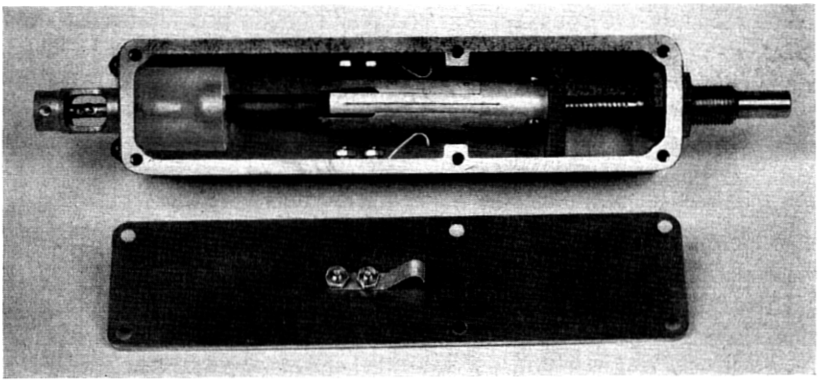

Fig. 8 — Model of coaxial rheostat with cover removed.

this purpose. It employs a ceramic rod, $\frac{1}{4}$ inch in diameter, coated with a pyrolytic carbon film, as a center member of a coaxial structure. A metal sleeve which is moved longitudinally by a lead screw carries sliding contacts along the rod. These parts are supported inside a rectangular housing which forms the outer conductor. A second set of fixed contacts attached to the rectangular housing makes contact with the sleeve. This arrangement maintains a substantially constant length of path from the input end of the rod to the housing, which forms the ground, independent of the position of the sleeve. The schematic of the rheostat is shown in Fig. 7. A model of the rheostat is shown in Fig. 8.

To obtain uniform adjustment of amplitude in decibels, a resistance that varies exponentially with length or with rotation of the lead screw is required. Such a resistance characteristic is realized by varying the thickness of the carbon film along the rod. This produced a total resistance which varied from 20 ohms at the low setting to about 350 ohms at the high resistance setting. After an initial wearing-in period of 1,000 cycles of moving the contacts over their full travel, the resistance was changed less than 1 per cent by another 9,000 cycles. This amount of wear is estimated to be greater than that encountered in twenty years of normal operation.

The housing and rod were dimensioned to form a 75-ohm transmission line. Measurements of the impedance at the input connector, made at frequencies from 60 to 80 mc, showed that this impedance can be approximated by a resistor terminating 6.7 cm of 75-ohm coaxial cable. For the 75-ohm setting, the reflection coefficient of the rheostat is less than 2 per cent across this frequency band.

One model of the equalizer was completely equipped with these rheostats. By allowing for the equivalent length of cable within the rheostat, an essentially pure resistive termination was obtained.

IX. OTHER COMPONENTS

Other components required for the equalizer included stepped rheostats, delay line or delay networks, a suitable summing network, and a loss equalizer.

The stepped-switch rheostats were made from standard switch parts with eleven positions. Deposited carbon resistors, 205D, were used for the steps. The mid-position corresponded to the circuit impedance level, 75 ohms. The other steps were arranged to provide equal increments of echo amplitude measured in decibels. With careful control of lead lengths, special shielding and a coaxial cable connector, satisfactory control of the return loss of this rheostat was obtained.

Resistance pads were added to the switch assemblies associated with each of the echo terms. The loss of each pad was determined so that the corresponding term would have the desired maximum amplitude. In addition, the losses of the pads associated with the leading echo terms were increased to compensate for the midband loss of the delay line between the leading coupler and the corresponding lagging coupler. This insured that the two echos would have equal amplitudes.

The delay required between taps in the delay line is 0.025 microsecond, corresponding to a change in phase shift of 180° from 60 to 80 mc. In order for the equalizer cosine characteristics to have maxima at the band edges, the total phase shift at 60 and 80 mc must be successive integral multiples of 180°. Since the phase shift of coaxial patch cable is closely linear and proportional to length, it could be used for the delay line. Lumped-element delay networks consisting of two or more all-pass sections are a feasible alternative and would reduce the over-all size and weight. In view of the additional development effort involved to produce these and the experimental nature of this equalizer, it was decided to use coaxial patch cord. The type selected, 728A cable, has a polyethylene dielectric and is tested during production for return loss in the 50- to 95-mc band. The length required for each section is about 15.6 feet. This much cable has a loss of about 0.3 db at 70 mc.

It was originally proposed to use a series of directional couplers for summing the echo voltages with the main signal. This would provide additional isolation between terms. However, tests on a preliminary model indicated this isolation was not required in this application. In-

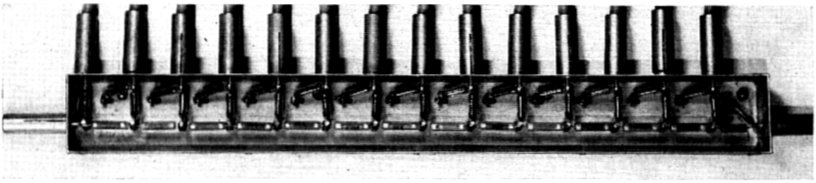

Fig. 9. — Summing network with cover removed. Main signal input is at right, echo signal inputs on top, and output is at left..

stead, a resistance summing network was developed using deposited carbon resistors. An L-pad is used in each echo path and a series resistor is added to the main path to preserve the 75-ohm impedance level, introducing a main path loss of about 0.4 db per tap. A model with cover removed is shown on Fig. 9. The main signal is introduced at one end of the structure, the echo voltages are connected along the side and the sum is taken off the other end. The return loss measured at any of the connectors with the others terminated was of the order of 40 db over the 60- to 80-mc band.

An attentuation equalizer is required to make the transmission through the main path constant. This path consists of about 108 feet of 728 cable, the straight-through loss of six couplers, and the coupling loss of the main coupler. The net distortion over the band is a slope of about 1.5 db and is corrected for by a constant resistance equalizer. Return losses exceeding 34 db were obtained over the frequency band.

X. ASSEMBLY

All the components were mounted on the rear of a standard relay rack panel. The rheostat controls are arranged on the front of the panel as shown on Fig. 10. This is a front view of the completed equalizer. The controls for the leading echo terms are on the left and for the lagging echo terms on the right. They are arranged vertically in numerical order with the first terms (shortest time separation from the main signal) at the top.

The rear of the panel is shown on Fig. 11. The directional couplers are mounted horizontally in two vertical columns. The cables forming the delay line sections are terminated in a plug and a jack and these are inserted in successive couplers, from the second port of one to the first port of the next. The third port of each coupler is connected through a short cable to its corresponding rheostat assembly. The fourth port of each coupler is connected to the summing network. An exception is

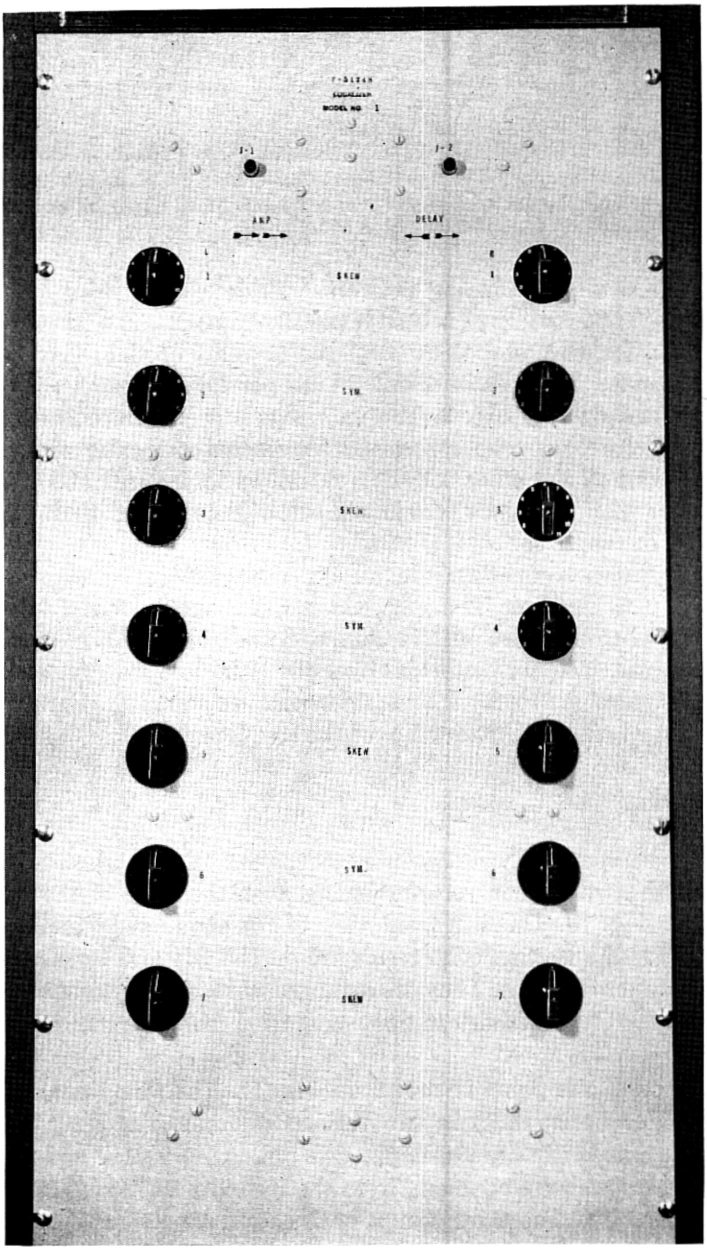

Fig. 10 — Front view of equalizer, showing rheostat controls. Leading echo controls are on left, lagging ones on right. First harmonic controls are at top.

the middle coupler, the fourth port of which is terminated in 75 ohms and the third port connected to the summing network.

The envelope delay in the cables connecting each coupler to the rheostat and to the summing network appears as delay for the particular echo path. Since these delays are not negligible compared to the 25 millimicrosecond delay between echos, the cable lengths were controlled so that the same amount of additional delay was introduced into each path including the main path.

XI. ADJUSTMENT AND PERFORMANCE

After the equalizer was assembled, the length of each of the cables connecting the couplers to the summing network was adjusted so that

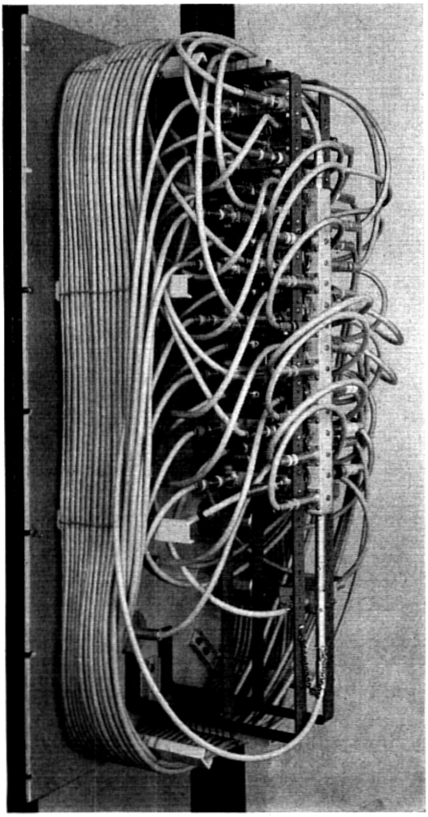

Fig. 11 — Equalizer with cover removed. Cables wound outside frame are the delay line sections. Summing network and directional couplers are in center. Rheostat cases are mounted on panel with coaxial connection in rear.

the zeros of the cosine shape occurred at the proper frequencies, as observed when the associated rheostat was set at maximum and minimum positions, with all other rheostats set at midrange, or no-echo, setting.

Some reflections were present in the main signal path as evidenced by ripples in the gain characteristic when all rheostats were set at midrange, corresponding to the "flat" loss condition. These reflections were reduced to some extent by minor readjustments of the balancing screws on the directional couplers. The over-all flat gain characteristic obtained after these adjustments is shown on Fig. 12.

This figure also shows the seven gain characteristics obtained when each pair of rheostats is set for maximum gain. The markers on the reference trace correspond to the band edges. A sharp gain bump resulting

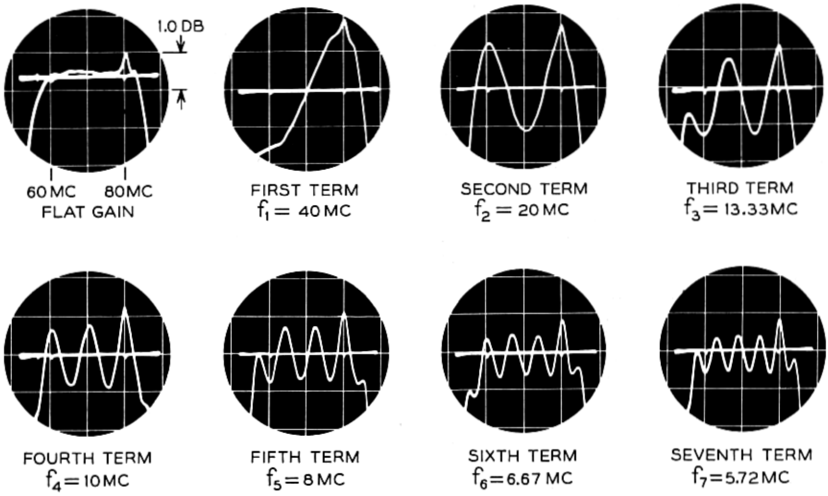

Fig. 12 — Measured gain characteristics of equalizer.

Fig. 13 — Gain change introduced by changing delay terms from zero to maximum. Left, normal case. Middle, third harmonic term at maximum. Right, fifth harmonic term at maximum.

from reflections on the delay line occurs just above 80 mc and distorts each characteristic near this frequency. The delay characteristics obtained closely resemble the corresponding gain characteristics. The delay characteristic with all rheostats on midrange was flat to within about ± 1.5 millimicroseconds.

Another measure of the performance is the amount of interaction between gain and delay characteristics. As shown on Fig. 13 the gain changes less than 0.2 db when the third or fifth harmonic delay terms are set at their maximum values of 11 and 12 millimicroseconds, respectively. Similarly, when a pair of rheostats are set for the maximum gain characteristic, the effect on delay is of the order of two millimicroseconds. These results, which are typical, indicate the interaction effect is of the order of 20 per cent using one neper of gain distortion ripple as equivalent to a ripple of one radian of phase shift amplitude. The effect of this interaction on the field use of the equalizer is to require a second round of equalization to correct for interactions after the gross distortions in a circuit have been equalized.

XII. FIELD EXPERIMENTS

Models of this equalizer were installed in two channels of the TD-2 system between Denver, Colorado, and Omaha, Nebraska, early in 1956. This route is about 500 miles long and includes 18 microwave links. One equalizer was installed at the center of the route and a second one at the receiving end. The results of a typical set of characteristics obtained are shown in Figs. 14 and 15. The first shows the delay characteristic of the whole channel measured at the receiving intermediate

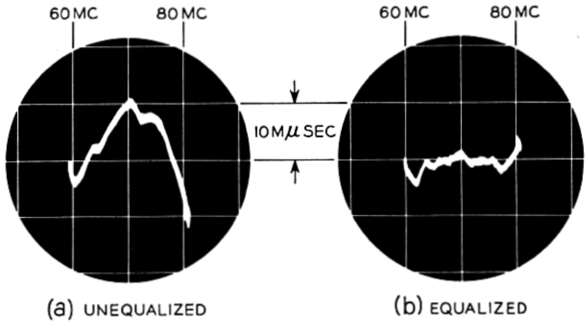

Fig. 14 — Measurements on Denver-Omaha route. Envelope delay distortion measured at intermediate frequency point. Left, unequalized; right with equalizer adjusted.

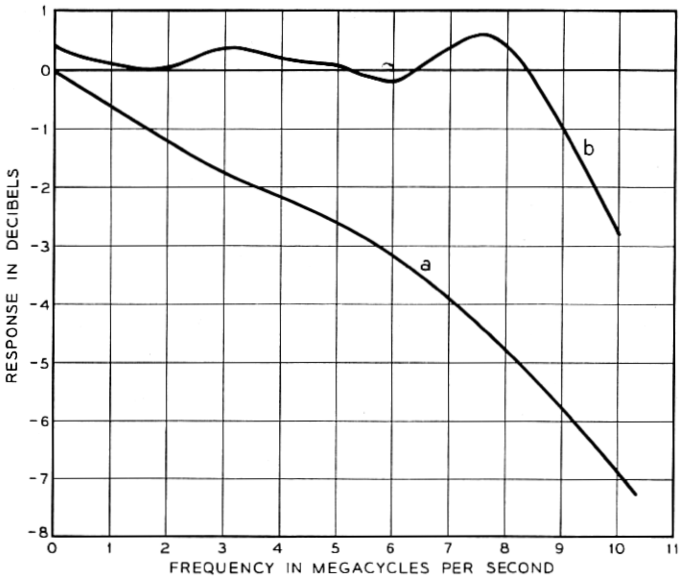

Fig. 15 — Measurement on Denver-Omaha route. Transmission characteristic measured at video (a) unequalized, (b) with equalizers adjusted.

frequency point. The "unequalized" characteristic is the circuit delay distortion immediately after standard line-up procedures without video equalization. The "equalized" characteristic shows the same circuit distortion corrected by the addition of the transversal equalizer. The first two delay terms of this equalizer were supplemented by the use of 319 type (linear slope) and 315A (parabolic) equalizers. The second loss term was supplemented by the use of experimental fixed parabolic loss equalizers. An attempt was made to obtain the best performance over the center 10 mc of the band, corresponding to the first order modulation band.

The gain distortion was adjusted on a demodulated video basis as this was the only type of sweep gain circuit available. As shown in Fig. 15, the unequalized circuit had more than 0.5 db per mc of slope up to 10 mc. The addition of the equalizer produced a flat band to about 8.5 mc.

The effect of this equalization on cross modulation, measured by simulating the message load with a flat band of noise, is shown in Fig. 16, for two sample channels. In these curves, normal drive represents the noise load whose power is 12 db below the power in a sine wave giving 4-mc peak deviation of the TD-2 carrier. Channel A showed 5-db im-

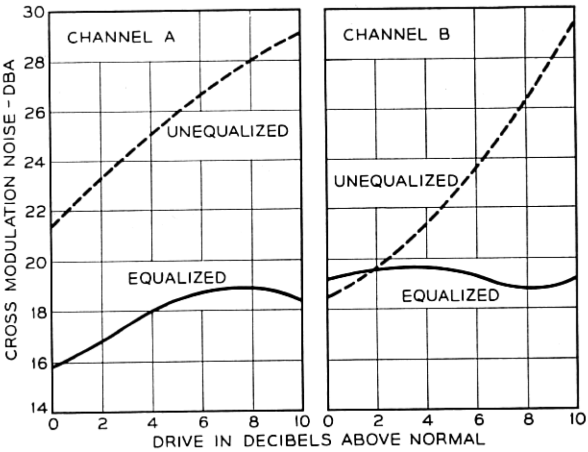

Fig. 16 — Effect of improved equalization on cross modulation noise, Denver-Omaha route. Measurements on two channels, referred to − 9 db transmission level.

provement at normal drive, and a somewhat greater improvement at higher drives. Channel B, however, showed a slight degradation at normal drive, which becomes a 9- or 10-db improvement at high drive. In the case of the latter channel, the improvement at normal drive was limited by the presence of a delay ripple, due to waveguide echoes, which was of too short a period to be equalized by the present equalizer.

The realization of such improvements in a working system is limited by such waveguide echoes, which are not stable enough for ready equalization, as well as by other instabilities in the transmission characteristic which have been attributed to antenna and air path effects.

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