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RADIO FACSIMILE BY SUB-CARRIER FREQUENCY MODULATION

BY

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Summary—*Progress of the radio art in turn permits advances in radio-photo methods. Of several under investigation, the use of frequency modulation of an audio sub-carrier is described, together with the transmitting and receiving circuits. The method minimizes the effect of selective fading and permits direct linear recording of picture tonal values at increased operating speeds. Effect of multipath variations on texture of recording is indicated.*

Commercial use on the New York-London circuit is pointed out.

HISTORICAL

THE history of radio and wire facsimile transmission is one of long standing and covers a vast amount of detailed effort on the part of an army of scientific workers. It has been exceptionally well outlined by J. L. Callahan in a recent bibliographic article[1].

As a result of these efforts, a time modulation method was first put into use in 1924[2] and, with modifications and improvements, attained world-wide acceptance as a practical method of radio transmission of facsimile copy. The method, as perfected,[3] is predicated on the idea of synthesizing picture tone values by recording dots of varying duration, but of uniform frequency of occurrence.

This method has some characteristics which it would be desirable to improve.

(a) When signal conditions are sub-normal multipath and selective fading produce streaks in the recorded copy.
(b) The screen pattern produced on the received copy is not entirely satisfactory when rescreened for enlargement or reduction by newspapers.
(c) Some of the fine detail in the picture is lost.
(d) The time of transmission must be relatively long in order to minimize the effect of multipath variations on the recorded copy.

It was apparent even before the development of the time modulation systems, that a method utilizing a linear modulation controlled directly from the phototube output would constitute an improvement on methods involving synthesizing of the picture tone values.

At that early date (1924) it was feasible to use only telegraphic keying for radio transmission of the picture signals. Early application to short-wave radio circuits still imposed the same restrictions. However, the short-wave radio art has progressed in the last ten years to such a degree as to warrant a reconsideration of some of the previously considered basic limitations. Notably, the development of good quality, short-wave radio, point-to-point, program circuits has opened the field of all the various forms of telephonic and tone modulation to facsimile. Probably the two greatest contributing factors in this have been the

Fig. 1—Photograph transmitted from San Francisco to New York by radio, using sub-carrier frequency modulation.

use of directive antennas to increase the effective signal-to-noise ratios, and the use of three-receiver diversity reception[4] to reduce vastly the effect of fading.

In spite of these advances, fading—and selective fading in particular—detracts from the expected efficiency of systems utilizing simple amplitude modulation for facsimile work. Other schemes have been proposed, such as the use of a constant amplitude pilot tone to operate the automatic gain control of the receiver; the use of amplitude modulation applied to several harmonically related sub-carriers; and the use of frequency or phase modulation of the radio-frequency carrier, or frequency or phase modulation of a sub-carrier tone. Due to the selective fading encountered on long-distance short-wave radio circuits,

Fig. 2—Commercial radiophoto picture transmitted from London to New York, using sub-carrier frequency modulation. Radio propagation conditions good—1939. (Courtesy Associated Press)

the frequency modulation of a sub-carrier tone seemed to be the most promising method.

The most practical line of development to follow appeared to be the use of frequency modulation of a sub-carrier tone, which, in turn, may be applied as amplitude modulation to a short-wave radio telephone circuit. This provided a means for utilizing the advances in the art and the available facilities for point-to-point program handling above cited. It also offers great opportunities for use on wire-line networks, such as are in extensive operation both in the United States and throughout the world.

Development of Sub-Carrier Frequency Modulation

Some early work on this type of system was done by R. R. Beal and J. W. Cox of R.C.A. Communications, Inc. at San Francisco soon after the commencement of commercial transmission of facsimile matter. No published account of this work has been made.

Fig. 3—Commercial radiophoto picture transmitted from London to New York using sub-carrier frequency modulation. Radio propagation conditions poor—1939. (Courtesy National Broadcasting Company)

In 1934 Kunert and Stahl of Berlin[5] reported on such a proposal in much the same vein as above stated, although they did not then report any actual experience on long-distance circuits.

A program was also undertaken at New York looking toward the development of a practical facsimile system utilizing the sub-carrier frequency modulation principles for commercial use on the international radio circuits of R.C.A. Communications, Inc. Tests on this system were run on the overland circuit from San Francisco to New York, and, as a result, the inherent feasibility of the scheme was then conclusively confirmed. Figure 1 is illustrative of the results obtained on these tests.

As the next step in the development, the tests were transferred to the London-New York circuit, with the full collaboration of Cable & Wireless, Ltd. The results warranted undertaking commercial service.

Therefore, facsimile service wherein the sub-carrier sytem of frequency modulation was used, was made available to public service on May 15th, 1939 by Cable & Wireless, Ltd. and R.C.A. Communications, Inc. acting cooperatively.

Figure 2 shows a typical commercial picture transmitted from London to New York when radio circuit conditions were good. Figure 3 indicates the efficiency of the system when radio conditions were poor and the noise level high. It will be seen that the noise has not too seriously affected the recording and the picture is completely usable.

The system has resulted in the realization of several advantages,—

(a) The speed of operation has been increased threefold. Facsimile matter is now sent at 60 revolutions per minute with a line advance of 120 lines per inch.

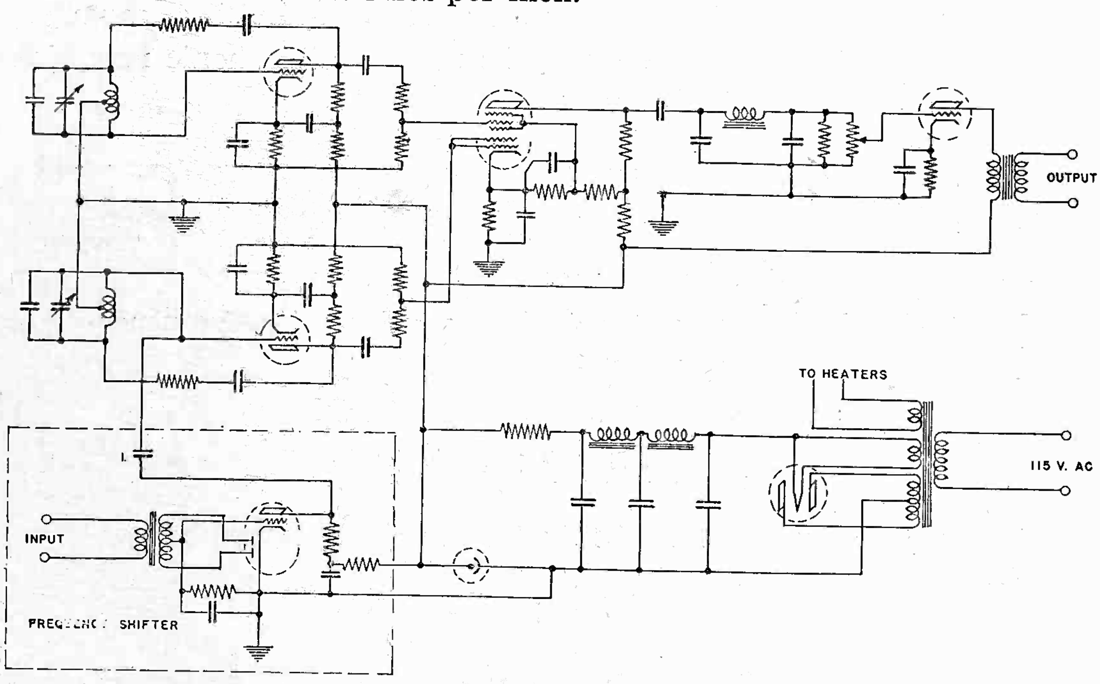

Fig. 4—Schematic diagram of sub-carrier frequency modulator.

(b) Linear amplitude recording is obtained, and results in much improved detail, fidelity of tonal values and elimination of the screen or dot pattern.

(c) Usable pictures may be obtained through poorer signal conditions.

(d) Minimizing of streaks caused by multipath or selective fading.

(e) Uses the existing standard methods and equipment for amplitude-modulated radio transmitters and receivers.

(f) Makes possible direct connection to wire-line systems by means of an additional modulator at the receiving office of the radio circuit.

System Used by R.C.A. Communications, Inc.

The transmitting plant of R.C.A. Communications, Inc., although primarily designed for telegraphic use, is provided with a number of portable amplitude modulators. By this means the majority of the transmitters can be adjusted for telephonic service in less than half an hour. Thus, a considerable choice of transmitter powers and frequencies is available to this service. Because harmonic distortion of the sub-carrier tone is not of major importance, essentially full 100 per cent modulation of the transmitter by the sub-carrier can be utilized. Connection between the Central Office and the transmitting station is obtained over the control lines used for the point-to-point program service.

At the receiving station three directive antennas are fed into three receivers connected for diversity action.[4] The frequency-modulated sub-carrier from these receivers is sent to the Central Office over program-quality control lines. Again, the receiving plant is so arranged that a majority of the receivers may be used for telephonic reception as well as for telegraphic work.

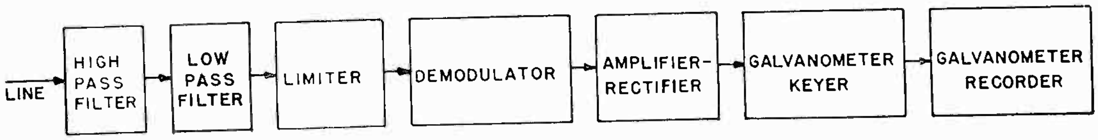

Fig. 5—Block diagram of terminal-office, sub-carrier, frequency-modulation, receiving system.

Terminal Apparatus

The equipment used at the Central Offices is described below:

Transmitting Converter

The ultimate in design of a transmitting arrangement for frequency modulation may be some form of a special scanning head having a frequency-modulated output. There are available several schemes for accomplishing this end. However, it is equally possible, in view of the flexibility of operation of the sub-carrier frequency-modulation system, to utilize existing scanners and other available terminal equipment.

With this thought in mind, the investigation was directed toward a system which would linearly produce a tone of constant amplitude and of a varying frequency, from a tone of constant frequency and varying amplitude. Also, it was desirable to include a means for adjusting the output frequency of the converter over a reasonably wide range without appreciably altering the percentage of frequency change for a given input amplitude variation.

The solution to the problem was found in a set-up arranged as a beat-frequency oscillator. With such a system, the amount of frequency shift automatically remains constant as the average output frequency is varied. Also, since the desired change in frequency is a very small percentage of the r-f oscillator frequency, the amplitude is constant over a wide range of output frequency.

Figure 4 shows the complete schematic diagram of the frequency-modulation converter comprising the beat-frequency oscillator, the frequency shifting unit, and the regulated power supply.

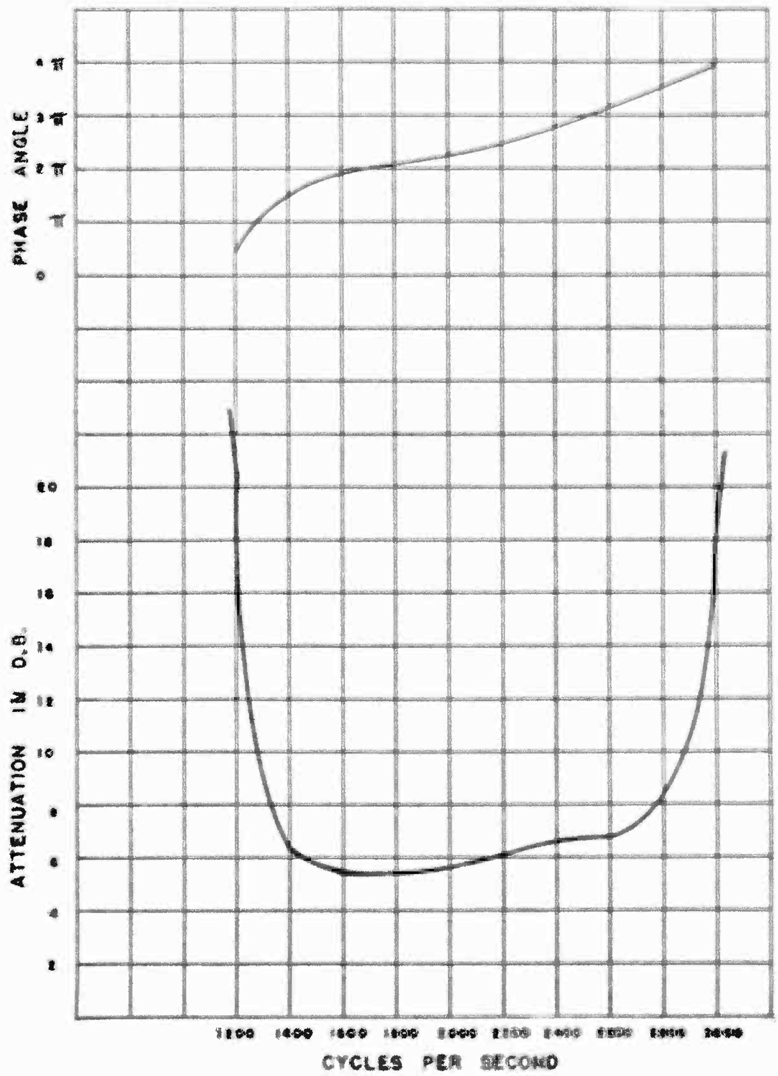

Fig. 6—Frequency and phase characteristics of band-pass filter set-up used for reception.

The beat-frequency oscillator is quite conventional. It consists of two resistance-stabilized oscillators operating at approximately 75 kilocycles, a mixer tube, a low-pass filter, and an output amplifier.

The frequency-modulator schematic is separated from the rest of the drawing by dashed lines. It consists of a signal rectifier, a time-constant circuit for smoothing out the ripple in the rectified signal, and a d-c amplifier. The d-c amplifier acts as a variable resistance between condenser (1) and ground. This combination is effectively connected between the grid circuit of one of the oscillators and the center tap of the oscillator coil. The tuning effect of condenser (1) upon the oscillator tank coil is dependent upon the resistance of the tube, which, in turn, is varied by the input signal. Plate potential is applied to the frequency-shifter tube to prevent distortion of the oscillator waveform due to plate rectification. The plate power supply is regulated to minimize amplitude and frequency variations which might be caused by line-voltage fluctuations. The frequency change will be as linear with respect to input amplitude as is the plate resistance of the d-c amplifier with relation to its control-grid potential. The frequency range required in the present frequency-modulation work is readily obtainable without operating beyond the linear characteristic of the tube.

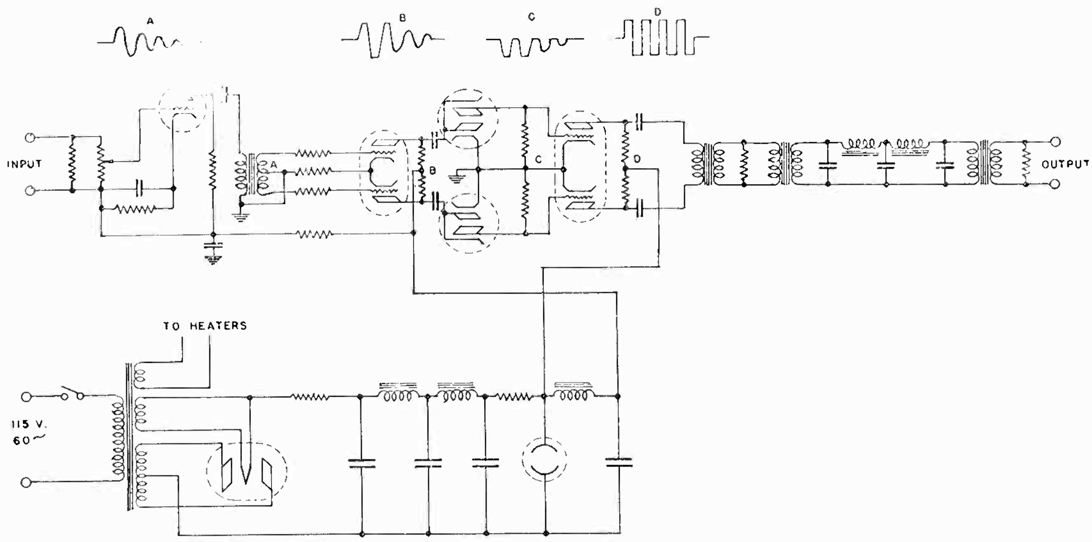

Fig. 7—Schematic diagram of cut-off-type limiter, also showing waveforms appearing in different parts of the circuit.

Receiving System

The general layout of the terminal receiving system used for frequency-modulation work is shown in Figure 5. The signal passes through a band-pass filter, amplitude limiter, demodulator, amplifier-rectifier, galvanometer keyer, and galvanometer recorder.

The band-pass filter eliminates frequencies above 3,000 cycles and below 1,200 cycles. This prevents the second harmonic of the sub-carrier frequency and spurious low frequencies from operating the limiter. The characteristic frequency-response curve of this filter is shown in Figure 6. The phase characteristic is also shown.

Push-Pull Cut-Off Limiter

A schematic diagram of the push-pull cut-off limiter is shown in Figure 8. The signal is first amplified in a single-ended amplifier stage followed by a push-pull amplifier. The push-pull amplifier operates as a conventional amplifier for low-signal levels, but acts as a positive peak limiter when the grid input is sufficiently high to cause grid current to flow. This limiting action is obtained by the insertion of a high value of resistance in series with each grid. The output of the push-pull stage is coupled to the grids of the push-pull limiter stage through condensers and rectifiers. The rectifiers are arranged to pass the negative half only of each cycle to the grids of the cut-off limiter.

The push-pull limiter tube is operating at zero bias and therefore each triode section is drawing its maximum plate current. When a signal is applied, the negative half-cycle affects one grid, reducing the plate current of that triode section to zero. The positive half of the cycle is blocked off the other grid by the rectifier. When the cycle reverses, the first grid returns to zero, and the opposite grid receives

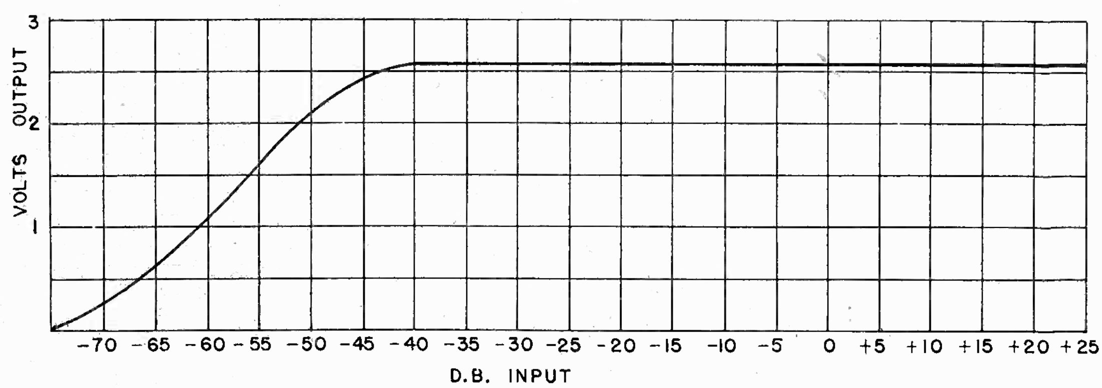

Fig. 8—Input-output characteristics of the cutoff type limiter.

a negative pulse. Thus, the tubes operate from zero to cut-off on alternate half cycles. The resultant output is a square waveform of a constant amplitude.

The waveforms appearing at different points in the circuit are indicated on Figure 7. The input is assumed to be a sine wave of diminishing amplitude in order to illustrate the limiting action. The flat characteristic of this limiter is shown in Figure 8.

DEMODULATION SYSTEMS

(a) *Low-Pass Filter.* The most simple device for frequency-shift demodulation is a low-pass filter as shown in Figure 9. The cut-off frequency is adjusted so that the desired frequency shift falls along the linear portion of the cut-off characteristic.

This filter is designed to have a sloping characteristic of substantial linearity over a range of from 1,500 to 2,300 cycles. (This permits satisfactory operation of the frequency-modulation receiving system without a critical adjustment of the frequency modulator at the transmitting end.) The operating curve of this filter is shown in Figure 9.

(b) *Pulse Integrator.* The pulse integrator was designed to produce an amplitude-modulated output tone from a frequency-shifted input tone, and to operate in this manner over a very wide range of frequencies. The output tone is supplied locally and may be a very high frequency if desired.

A schematic diagram of the pulse integrator appears in Figure 10. The following stages are shown in sequence: a push-pull signal ampli-

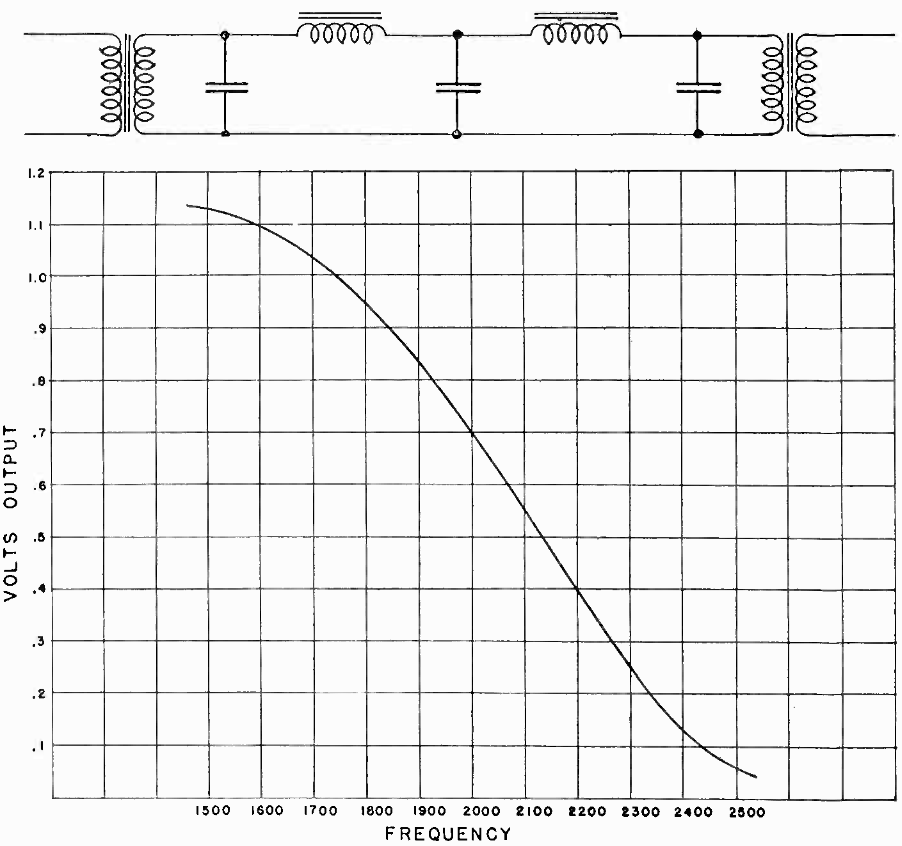

Fig. 9—Demodulation filter and its frequency characteristics.

fier, a signal-controlled gas-triode oscillator, a full-wave rectifier, a low-pass filter, and a balanced modulator.

The operation of the circuit is as follows: The signal is amplified in the push-pull amplifier and applied to the grids of the thyratron tubes for control. The thyratron tubes will trigger back and forth following the postive peaks of the input tone cycles. Energy from the thyratron cathode circuits is passed on through condensers C and C′ to the plates of a full-wave rectifier and thence through a low-pass filter to the control grids of the balanced modulator.

The time constant of C-R and C′-R′ is selected to pass only a short pulse from the thyratron cathode circuits to the rectifier. These pulses are of constant duration and amplitude irrespective of the frequency of the input tone. It therefore follows that the amplitude of the filtered potential appearing across R_T will be greatest for the highest frequency, by virtue of more pulses occurring during a given limit of time. The output of the low-pass filter is applied to the grids of the balanced modulator stage to control directly the amplitude of the modulator output.

The filter type of demodulator has the advantages of simplicity and reliability. The pulse integrator type, on the other hand, has the advantage of great flexibility.

One characteristic of the sub-carrier frequency modulation is that the contrast ratio after demodulation is relatively low. For this reason it is desirable to use some method of volume expansion in the recording

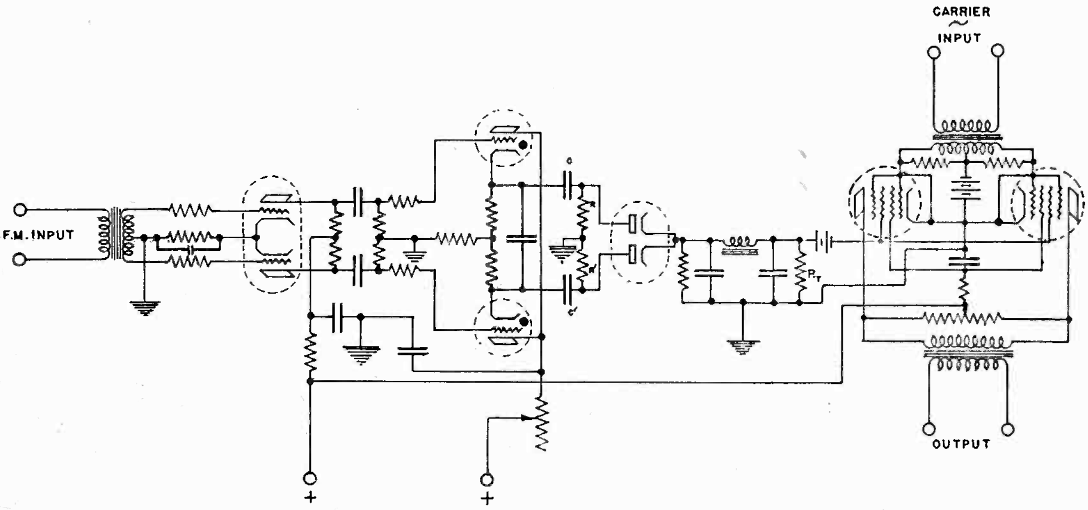

Fig. 10—Schematic diagram of the pulse demodulator.

system to obtain the necessary contrast ratio for satisfactory recording.

A schematic illustration of the keyer developed by R.C.A. Communications, Inc. is shown in Figure 11A. This device is analogous to a bridge circuit with a vacuum-tube amplifier in one arm, a variable resistor in the opposite arm, and fixed resistors in the two remaining arms, as shown in Figure 11B.

The desired operating characteristics and magnitude of output for any given input may be obtained by adjustment of the variable arm and the grid bias of the amplifier tube. In practice, the grid bias of the tube is adjusted to provide linear operation. The magnitude of the output is then controlled by the level of the input signal and the null adjustment.

With the proper grid bias adjustment of the keyer, an output of —20 to +20 milliamperes, or any lesser value, may be obtained from an input signal having a contrast ratio even as low as 4 to 5, and the linearity of output compared to input will be constant for any of these adjustments.

In order to permit ease of operating adjustment and to assure a linear recording characteristic, the standard R.C.A. Communications radiophoto galvanometer recorders have been equipped with linear

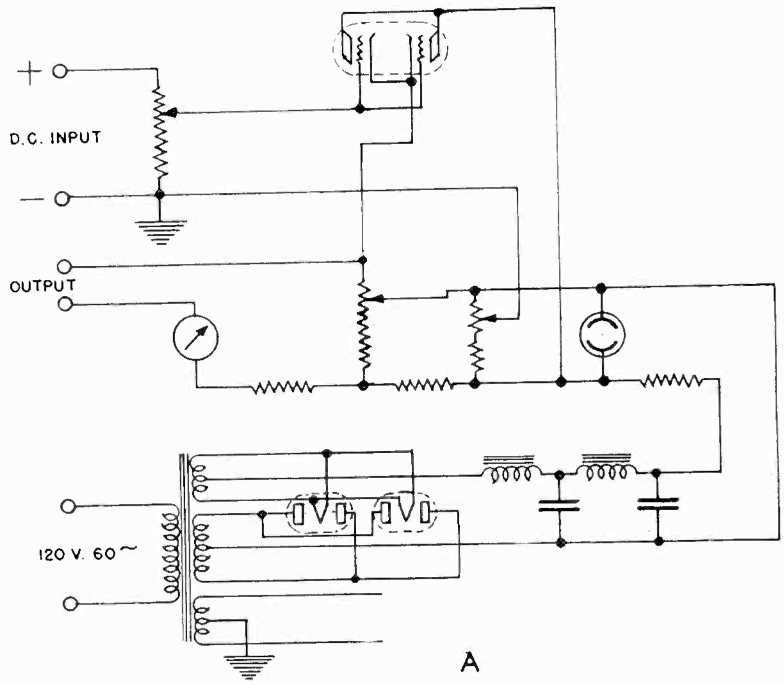

Fig. 11A—Schematic diagram of the galvanometer keyer.

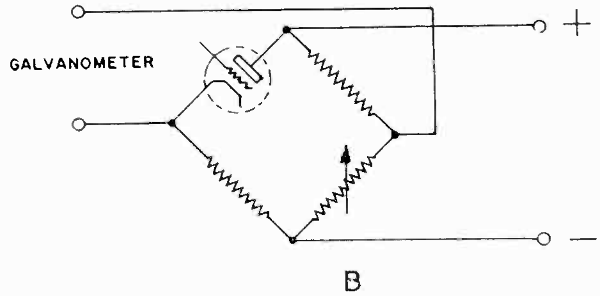

Fig. 11B—Equivalent circuit of galvanometer keyer.

penumbra arrangements, such as described by G. L. Dimmick for RCA photophone recording.[6]

Propagation Phenomena

Laboratory investigation, as well as circuit tests, disclosed that the classical advantages of frequency modulation are obtained in this system. In particular, the "improvement threshold"[7] phenomena results in minimizing the effect of noise on the record when the peak signal is more than 6 db above the peak noise. In this application, it is the ratios of the peak values that are of primary interest.

The effectiveness of this "threshold" effect is evidenced by the repeated useful recordings received from London with very low signal

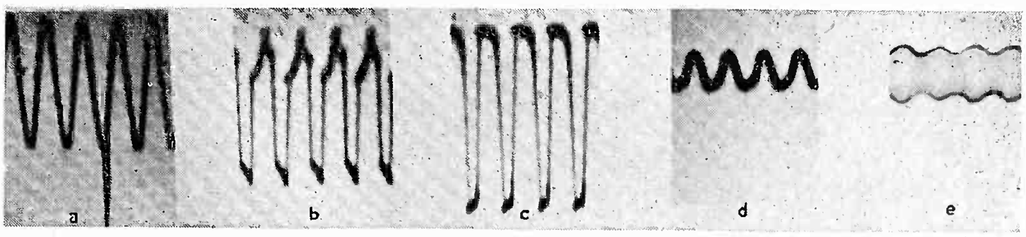

Fig. 12—Oscilloscope records showing effect of noise impulse occurring in phase with the incoming signal.

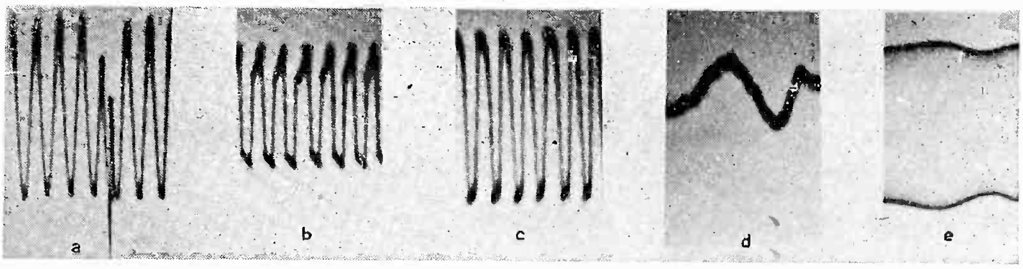

Fig. 13—Oscilloscope records showing effect of noise impulse arriving out of phase with the incoming signal, but passed through a band pass filter before limiting.

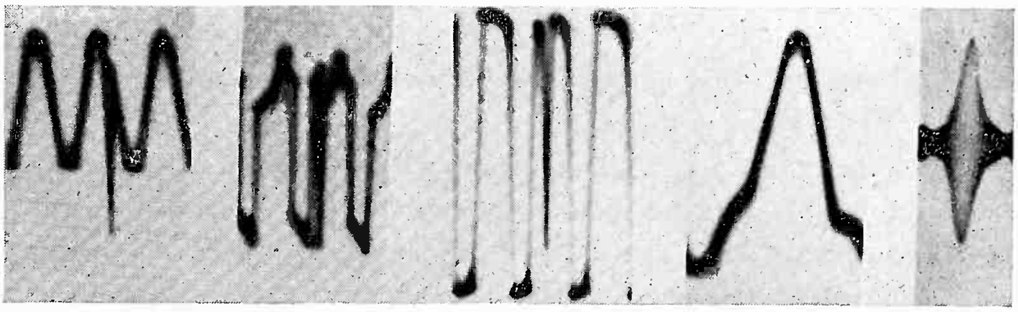

Fig. 14—Oscilloscope records showing the effect of an out-of-phase noise pulse when no band pass filter is used prior to limiting.

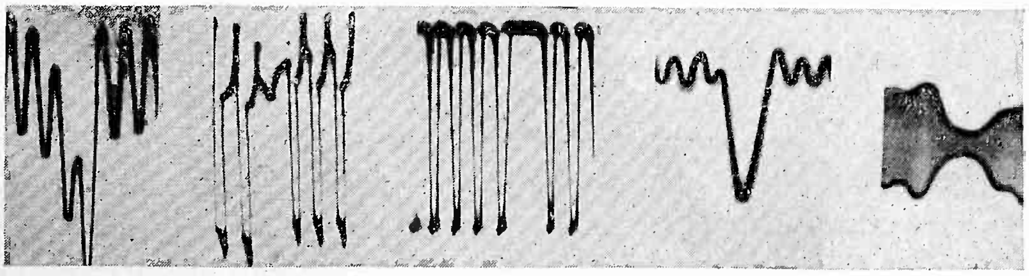

Fig. 15—Oscilloscope records showing the effect of a noise impulse of long duration.

levels at the receivers. At such times, telephone communication was impossible and even telegraph operation was possible only at a very slow manual rate.

Figures 12, 13, 14, and 15 are oscilloscope records of impulse noise as it affects the frequency, demodulated output. In all of these Figures (a) is the noise superposed on the signal at the input to the system, (b) is the output of the limiter feeding into the inductive load of the "pulse integrator" described above, (c) is the plate current of the gas triode of the integrator, (d) is the rectified and filtered signal applied to the grids of the tone-modulator stage of this unit, and (e) is the amplitude-modulated output fed to the recording system. Figure 12 shows that when a short impulse is in phase with the signal, no effect is produced in the output. Figure 13 shows that a band-pass filter ahead of the limiter converts the impulse into an instantaneous phase shift of the limiter output and will produce a minor disturbance by elongating one-half cycle passed by the limiter. Figure 14 shows the large disturbance created by an out-of-phase impulse of sufficient amplitude to cause an extra crossing of the zero axis at the limiter. This, of course, results in an instantaneous frequency considerably higher than normal and is so interpreted by the demodulator. Figure 16 indicates that impulses of duration greater than one cycle of the signal results in an instantaneous lower frequency and is so interpreted by the demodulator. It will be noted that for impulses shorter than one-half cycle of the sub-carrier, the phase of the impulse with respect to signal is of importance as well as its amplitude.

The selective fading can usually be greatly minimized by sufficiently rigid limiting over a wide range of signal variation. Even so, the sub-carrier signal will at times drop below the noise level at the Central Office and result in occasional streaks of noise being recorded. This usually happens when selective fading causes the momentary complete cancellation of the r-f carrier. It is frequently accompanied by a relatively strong audio tone of twice the frequency of the original sub-carrier, and carrying the complete frequency-modulated variations of the latter.

Another interesting phenomena caused by propagation is the shifting of the occurrence time or "phase" of the recorded sharp changes from one tonal value to another. This is most noticeable on printing or other black-and-white copy having narrow vertical lines closely spaced. It has been previously pointed out[8] that the fading encountered between the various "wave-bundles" comprising the complete multipath signal will cause the effective arrival of the signal to vary in time accordingly. The limiter, of course, produces an output whose basic frequency is, in general, dictated by the input component having the greatest amplitude. Now, the new frequency due to change in picture tonal value is arriving over a number of paths having different delays. If the instan-

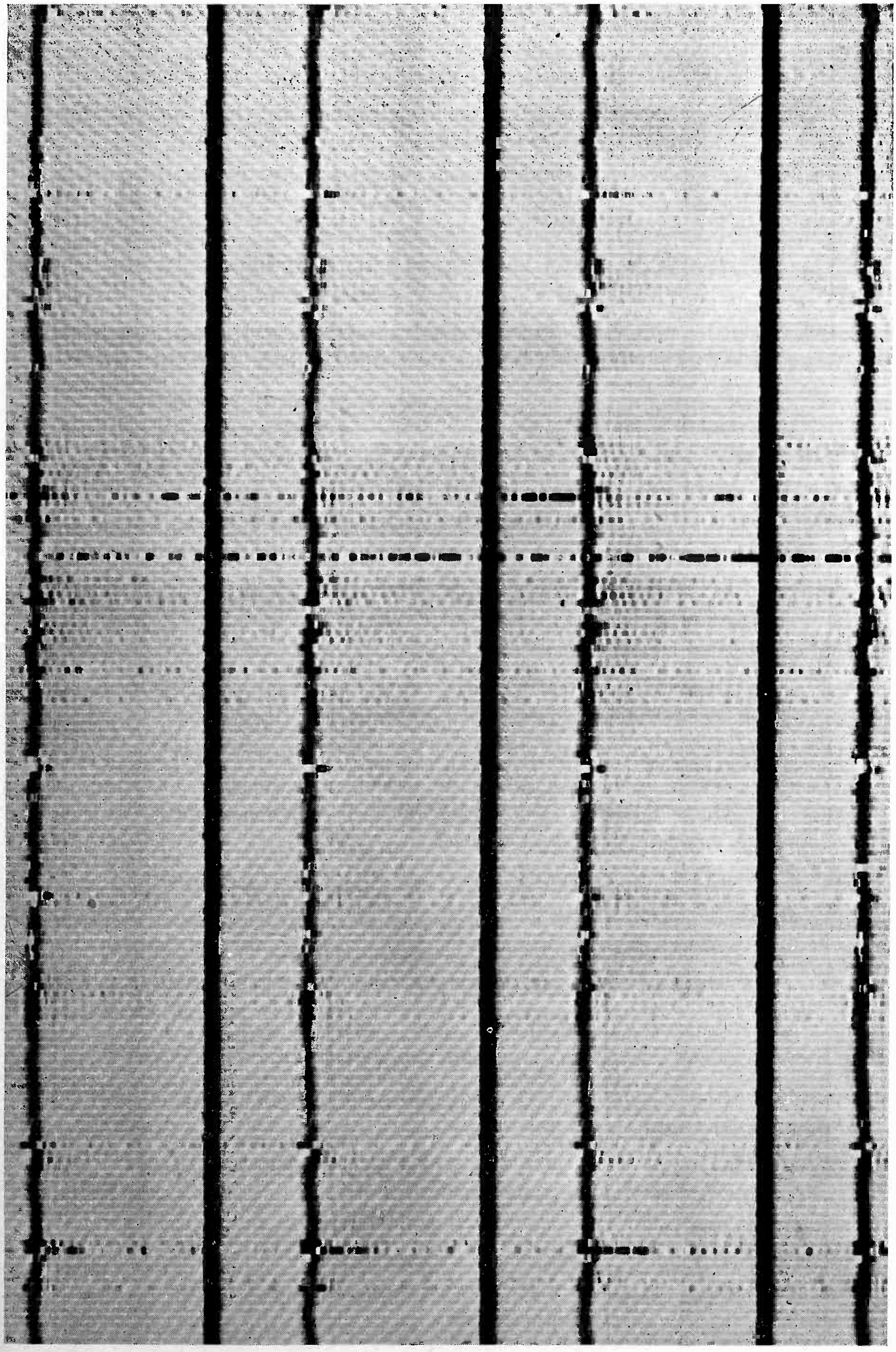

Fig. 16—Enlargement of recordings of 2 millisecond pulses transmitted from New York to San Francisco and retransmitted to New York. The outgoing and incoming pulses were recorded simultaneously.

taneous fading condition is such that the shorter paths of least delay have low amplitudes, the resultant of the new frequency will remain too low to assume control of the limiter over the old frequency. This condition will continue until the energy carried by the longer paths is sufficient and of the proper phase to build up the resultant of the new frequency to a higher value than the old frequency. On the other hand, the fading conditions may shift to where the shortest path carries enough energy of the new frequency to permit it to overcome the old frequency immediately. The result is a variation in the occurrence time of the change in recorded tonal values. When applied to the scheme under discussion, this appears under certain conditions to cause a wavering or raggedness of what would otherwise be straight, clean-cut vertical lines.

A second aspect of this same effect is that multipath generally causes an elongation of a signal which is greater than the variation of the effective first-arrival time above cited. Occasionally, the amplitude of this elongated signal remains above that of the arriving new signal or frequency, and thus retains control of the limiter instead of permitting the limiter to follow this new frequency. The net result at the recorder is an effective delay of the change from one tonal value to another.

A third possible source of this raggedness—which may be named "gollywobble"—may reside in the phase modulation effect of noise on the instantaneous peak signal when the signal-to-noise ratio is on the order of one or less.[7] Experiences on the long-distance circuits indicate that this condition is considerably worse when the radio circuit conditions are poor.

Figure 16 is an enlargement of about five times of a recording of a transmission from New York to San Francisco and return, using an automatic relay at San Francisco. The radio conditions were quite poor in both directions. The narrow solid lines are recordings of the 2 millisecond black bars transmitted from New York at a rate of 20 cycles. The black corresponds to a sub-carrier frequency of 1,600 cycles and the white background to a frequency of 2,000 cycles. The other vertical lines are the recorded signal back from San Francisco. The "gollywobble" is quite apparent as are the occasional streaks when the selective fading causes the signal to fall below the noise level.

Circuit tests to date have shown that when the above phenomena occurs it results in a record, which, although not fully satisfactory, does not destroy complete readability of the 8-point type which is the accepted standard of definition in radiophoto operations.

The effect, naturally, is not desirable, but it is less apparent in a system of this character than in many other systems, and it must be recalled also that this is not the normal operating occurrence. It is feasible to suppose that the phenomena is also affected by the vertical directivity of the antennas—as this will in great degree control the number of paths effective at the receiver.

Choice of Operating Circuit Standards

It is obvious that noise should be kept to a minimum. Therefore, as narrow a band of frequency as possible should be delimited by the selective circuits of the receivers and the band-pass filter at the Central Office.

On the other hand, it is desirable to get as nearly full 100 per cent amplitude variation recovered from the frequency demodulator as possible. This means as great a frequency swing as possible.

Further, the high and low range of the frequency swing may approach, but not equal, a ratio of 2 to 1 in order that the harmonics of the sub-carrier inserted by selective fading may be filtered out prior to applying the signal to the limiter.

Again, the lowest frequency of the sub-carrier that may be used is desirable in order to maintain the frequency components of the r-f carrier as close thereto as possible to minimize the effect of the selective fading and to reduce the extent of the r-f sidebands. This lower frequency limit is determined by the smallest picture detail to be transmitted. It must be such that the narrowest vertical line or space will contain at least one full cycle of the sub-carrier. Theoretically, it should contain even more cycles, but it has been found in practice that, generally, one complete cycle will suffice.

All the above factors are at present being subjected to careful quantitative analysis, but an empirical determination has been incorporated in the equipment described herein.

Dual Harmonic Channels

It was stated that at such times as the r-f carrier completely disappears, due to selective fading, there is often present a strong second harmonic bearing the desired frequency modulation. This is due to the upper and lower r-f sidebands beating together to cause a double frequency resultant. This component is also present whenever the carrier has faded to the extent that the sidebands cause effective overmodulation, viz., when the instantaneous sum is greater than the amplitude of the r-f carrier.

It would seem feasible to utilize this component of double frequency, or "second harmonic" of the fundamental sub-carrier component, to fill in the gaps on the recording caused by the disappearance of this fundamental component as above cited. This has been done by using band-pass filters to separate the two components in the central office. The fundamental is then doubled in frequency to match the second harmonic doubled by the propagation phenomena. This is selectively fed through a differentially operated electronic switch and thence to the limiter, demodulator, rectifier, and recording system. The equipment used is described below.

The general layout of the receiving system used for the frequency-modulation tests incorporating this feature is shown in block diagram form in Figure 17. Referring to this diagram, it is seen that the signal is first passed through 3,000-cycle low-pass and 500-cycle high-pass

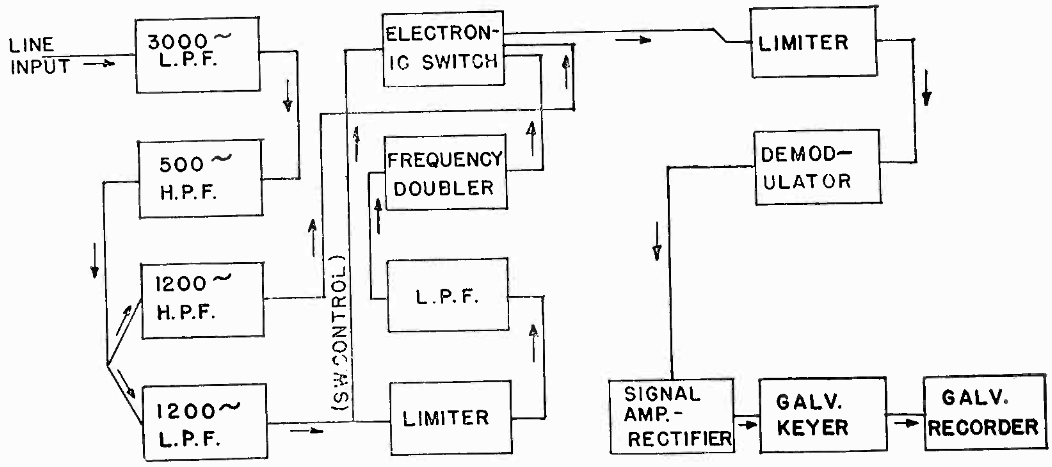

Fig. 17—Block diagram of the dual channel system.

filters for the purpose of limiting the band to only the useful range, thereby improving the signal-to-noise ratio. Following this high-low combination are 1,200-cycle high-pass and low-pass filters for separating the fundamental and harmonic frequencies of the sub-carrier. This is assuming a sub-carrier frequency varying between 800 and 1,000 cycles.

The fundamental frequency passes through the 1,200-cycle low-pass filter, a limiter, another low-pass filter, and on through the frequency doubler to the electronic switch.

The second harmonic of the sub-carrier frequency passes through the 1,200-cycle high-pass filter and directly to the electronic switch. The differential electronic switch is arranged to cut in the harmonic channel automatically when the signal in the fundamental channel fades below a predetermined level, provided the harmonic channel contains a signal of higher level than the fundamental channel. In this manner, the fundamental channel maintains control at all times (even though

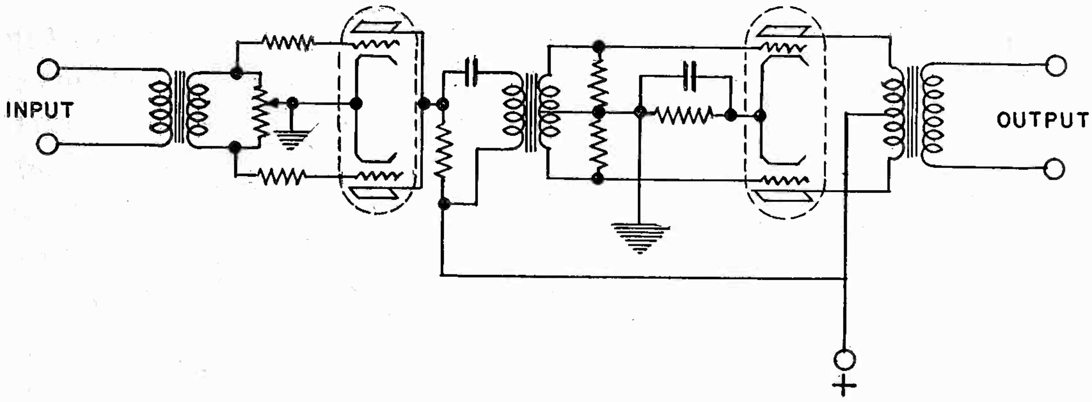

Fig. 18—Schematic diagram of the frequency doubler.

it may be very weak) unless the level of the harmonic channel is great enough to be useful.

The output of the differential electronic switch passes through a limiter and then through the frequency demodulator, as indicated in Figure 17. The output of the demodulator is next amplified and rectified. The output of the signal amplifier-rectifier operates the galvanometer keyer.

A schematic diagram of the frequency doubler is shown in Figure 18. It consists of a zero biased push-pull amplifier stage, with the plates of the two triode units connected in parallel. Grid resistors prevent the signal from driving the grids excessively positive. There is nothing, however, to prevent the grids from swinging negative beyond the point of plate-current cut-off, which they do. The result is similar to that of a full-wave rectifier, except that the output is of sinusoidal waveform, since the operation is not unlike that of a Class "B" amplifier. This type of frequency doubler is very satisfactory for application where it is desirable to provide an undistorted output of constant amplitude irrespective of input frequency or of sudden changes therein.

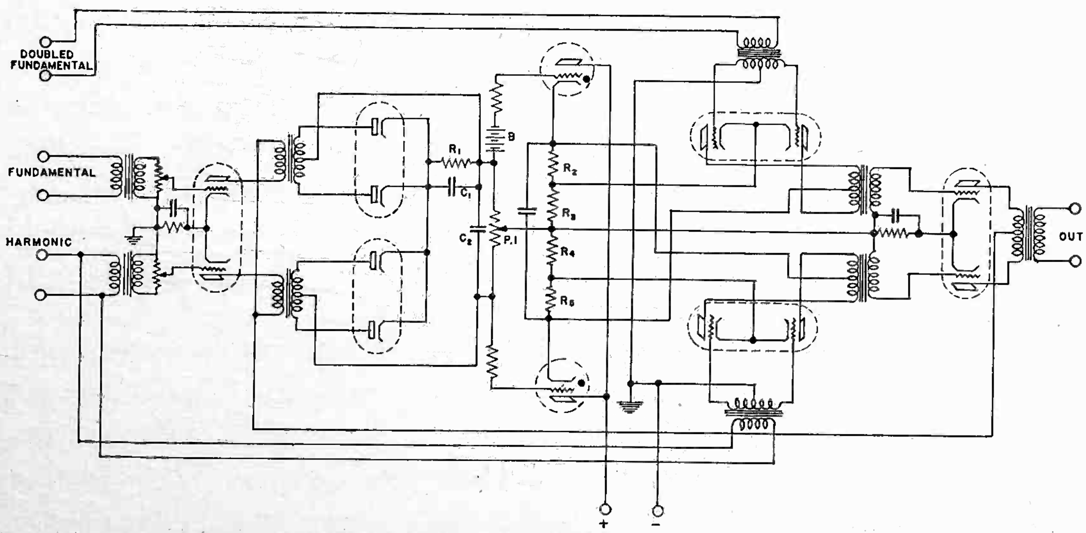

Fig. 19—Schematic diagram of the differential electronic switch.

This result could not readily be obtained with a frequency doubling scheme incorporating tuned circuits for the elimination of harmonics of the output frequency, due to the "flywheel" effect of such circuits.

The differential electronic switch used in connection with the sub-carrier, frequency-modulation, receiving system is arranged to transfer the control from the fundamental to the second harmonic of the sub-carrier frequency at times when the fundamental has faded below a usable level and when the level of the second-harmonic frequency is sufficiently high for satisfactory recording. The switch will not operate unless the harmonic channel is usable, even though no signal appears in the fundamental channel.

A schematic diagram of the switching unit is shown in Figure 19. There are three signal inputs as indicated. These are the harmonic input, the fundamental input, and the doubled fundamental input. The doubled fundamental input is normally amplified and passed on through to the output circuit.

The fundamental and second-harmonic frequencies are amplified and rectified separately, then smoothed out by the time-constant arrangement shown in the diagram as R_1-P_1 and C_1-C_2. In order to accomplish this switching, the harmonic channel must contain sufficient energy to overcome the IR drop across R_1 produced by the rectified fundamental frequency. In addition to this, it must overcome the thyratron keyer bias shown as battery "B" in the drawing. The threshold at which the switch operates is variable by the adjustment of P_1 and bias battery "B". The differential between the two signals is controlled by the adjustment of the volume controls of the fundamental and harmonic input amplifiers.

The operation of the electronic switch and keyed amplifiers is conventional. The mixer stage in the output circuit consists of a push-pull amplifier with one grid connected to each channel, and each of the two plates connected to one-half of the primary of the output transformer.

Any noise which will pass through both channels simultaneously, or any fade-out of both channels will not cause the switch to operate. The operation is based solely on a differential between the signal level contained in the two channels. The optimum adjustment of the differential between the two signals and the threshold point of the switch will depend somewhat upon propagation conditions.

TESTS

Local Tests

(a) *Linearity.* The linearity of the frequency-modulation system is unusually good. An overall check showed the exact curve of the demodulation filter, and disclosed substantially complete linearity when the pulse integrator was employed in place of the demodulation filter.

(b) *Signal-to-Noise Ratio.* The ratio of minimum peak-signal-to-peak noise for satisfactory picture recordings is about 1 to 1.

Radio Circuit Tests

Figures 20 and 21 show a comparison between the new and the previous system on the circuit from London to New York. Figure 1 is

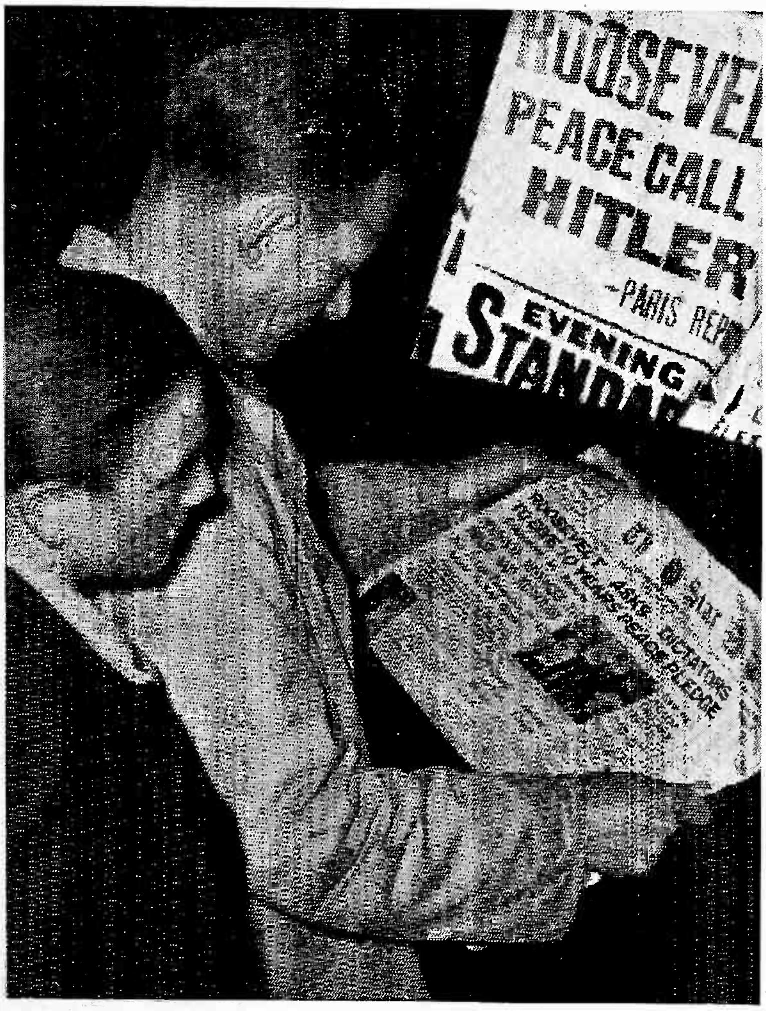

Fig. 20—Commercial radiophoto picture transmitted from London to New York by the CFVD method. (Courtesy International Newsphoto)

illustrative of the use of the second harmonic generated by selective fading. It was transmitted at 40 revolutions per minute and 120 lines per inch from San Francisco to New York using 800-1,000-cycle frequency shift, and received by the dual channel set-up. The frequency was doubled in the receiving circuits, as previously explained in the text.

(a) *R-F Band Width Requirements.* The r-f band width required for the transmission of sub-carrier frequency modulation appears to be comparable to the requirements for the transmission of an amplitude-modulated sub-carrier, having a modulating frequency equivalent to the highest used in the first case.

(b) *Occurrence of Selective Fading.* Selective fading occurs at random and to varying extents. In general, the rigid limiting employed

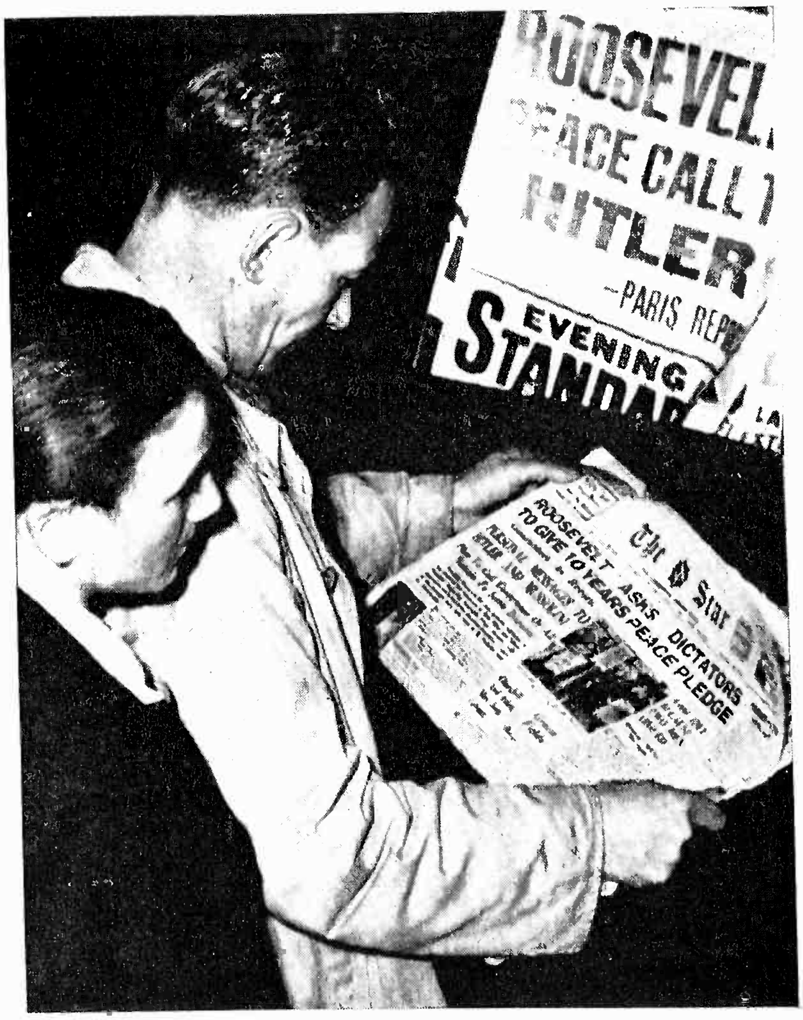

Fig. 21—Commercial radiophoto picture shown in Fig. 20 transmitted from London to New York by the sub-carrier frequency modulation system.
(Courtesy International Newsphoto)

in the frequency-modulation system effectively minimizes this phenomena on the recorded picture.

(c) *Effect of Diathermy or Noise.* Interference from diathermy or noise will mar the picture to some extent when the peak ratio of such interference to the minimum peak-signal amplitude becomes greater than 1 to 1. The effect is no worse than the same type of interference in any other form of system heretofore used.

Conclusions

Progress of the radio art has urged the improvement of methods for transmitting facsimile matter on long-distance radio circuits. Several means are under intensive investigation by R.C.A. Communications, Inc. and but one of them is reported upon herein.

The reported method of sub-carrier frequency modulation is a sufficient improvement to have warranted its use in commercial operations even at this early date in its development.

Some of the unique effects produced in the recordings by the r-f propagation phenomena are pointed out and feasible explanations offered. However, intensive quantitative study of all phases of the problem is being continued.

The increased fineness in texture and definition of the recordings has an importance of the first magnitude. In addition, the possibility of more readily working through sub-normal radio conditions, and the increased speed of operations is advantageous.

A future development to even overcome more fully the effects of selective fading by utilizing the second-harmonic signal generated thereby is a distinct possibility.

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PROGRAMMING THE TELEVISION MOBILE UNIT

By

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THE engineers and technicians of the RCA laboratories have developed a new medium of communication in television, and how best to use this new means of seeing is the major problem of the television program department. The mobile unit is an especially vital unit of the system. With cameras capable of being taken to the scene of an event as it is happening, the whole world opens up as potential program material for television, and the wide scope of possibilities has presented many new problems that we are attempting to solve in the pick-up of mobile unit programs today.

Among the first major demonstrations of outside pick-ups, was one given in Washington, D. C., from January 25th to February 2nd, 1939. The schedule called for four ten-minute programs every hour, four hours a day, on five days. Since the point of pick-up was out of doors, with no protection from the weather, it was obvious immediately that any set program routine, with rehearsed performers, was impossible. The question of program material was answered by having the audience supply its own programs. After witnessing a demonstration in the viewing rooms, members of the audience were transported by automobile to the scene of the pick-up somewhat over one mile distant, where they were interviewed before the television camera, thus providing entertainment for the viewers at the next demonstration.

From the program point of view, as well as the engineering, these fourteen odd hours of continuous television programs, with one camera, in sunshine, rain, sleet and bitterly cold winds, proved that the job could be done, even under adverse conditions. More than ten thousand persons witnessed these demonstrations and their reactions were highly favorable. The demonstrations revealed, however, some of the difficulties to be expected in future operations.

To do a first-class job, from a program point of view, on events happening out of the studio, the necessity of more than one camera was made apparent. While one ten-minute program unit was intensely interesting to the viewer who was looking at a television program for the first time, longer units definitely lacked interest value. A single camera was unable to provide the high-lights, the detail, and the overall picture, which were necessary to sustain interest for any long period of operation. Additional camera equipment was undoubtedly needed.

One of the high-lights of the five-day operation period in Washington occurred when the camera was pointed overhead to televise a blimp and a large transport plane, both plainly visible, with the Washington monument looming up in the distance. Another was the pick-up of actual field manoeuvers by a troop of cavalry and a battalion of field artillery from Fort Meyers, on January 31st, 1939.

Fig. 1—NBC television mobile unit, consisting of the "pick-up bus" and the "transmitter bus", in action.

These simple attempts at something other than close-up interviews demonstrated the necessity of closer co-ordination between the camera man at the point of pick-up, and the director and control engineers at the viewing screen in the truck. Men at the control point in the truck had no way of knowing what the cameraman was trying to pick up. For, while these two points were in telephonic communication, the cameraman could not tell the operators in the truck what his plans were, since the nearby microphone would pick up this conversation also. This showed the need of viewing monitors at the point of pick-up, with control of the program in the hands of men stationed where they could see what the camera was focused on, as well as what might be picked up next. With the addition of this equipment, the outside pick-ups of the future will have more motivation and greater entertainment value than those of the first few months.

Despite the originally limited facilities, however, much has been accomplished since the start of regular program service on April 30th of this year. The pick-up of President Roosevelt at the Federal Government Building at the opening ceremonies of the New York World's Fair was a ranking first in the field of television pick-ups in America. The battery of movie cameras present delivered its picture story to America a day or two later. Every television viewer saw the events when and as they happened. The opening ceremonies, the waiting crowds, the

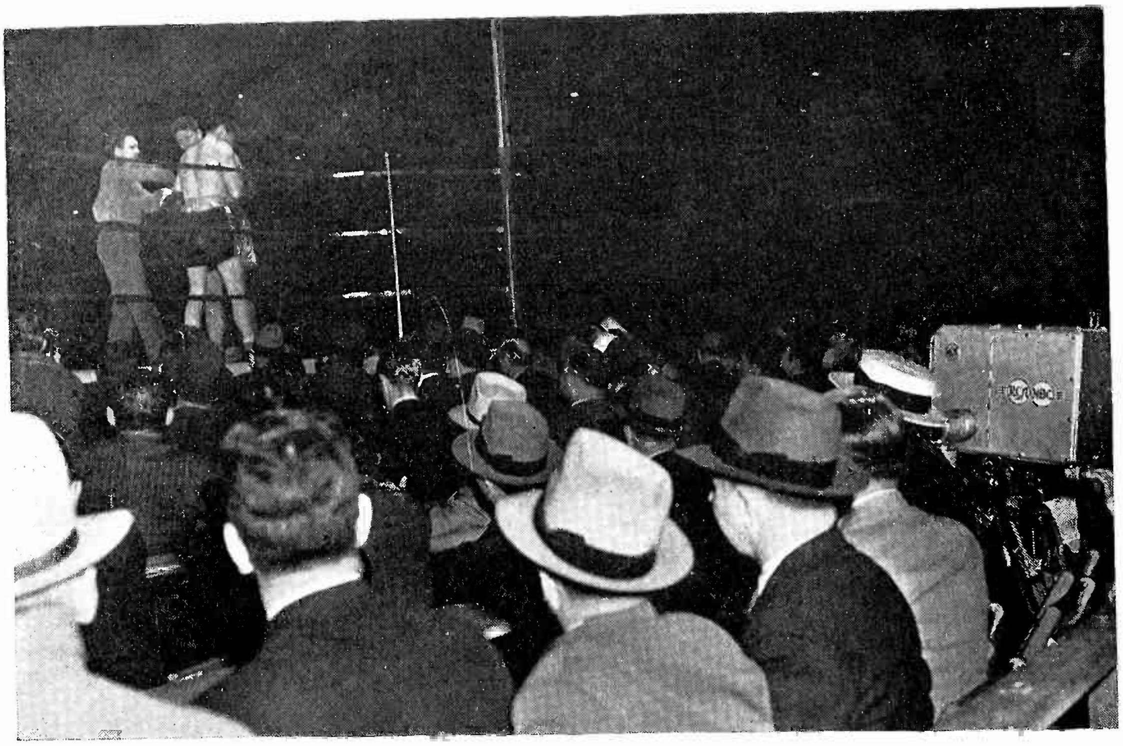

Fig. 2—NBC television camera pick-up of Baer-Nova fight.

arrival of the participating groups, and the opening addresses, were plainly seen and heard by the television audiences.

Upon the inauguration of a regular program schedule on April 30th, the mobile unit was put into regular program service. It has since furnished not only interesting program material, but has performed very reliably. During May and June the mobile unit delivered over twenty hours of television programs. The wide variety of subjects covered was fairly representative of what the public may expect in the future.

The first major outdoor event was the Columbia-Princeton baseball game on May 17th, 1939. This game was followed play-by-play, and when we compare this program to the crude play-by-play boards which have furnished the public some visual representation in baseball games in past years, this pick-up in spite of its limitations, is readily seen to be a long stride forward in sport reporting.

The Six-day Bicycle Race at Madison Square Garden in May, 1939 made two television "firsts." It was the first television broadcast of a sporting event held indoors. Although the size of the track presented a serious handicap to proper pick-up with one camera, all riders were visible and a good view of the race as a whole was delivered. This broadcast was also the first high-definition television program ever sent over regular telephone wires in the United States. The picture impulses were carried from Madison Square Garden to the RCA Building

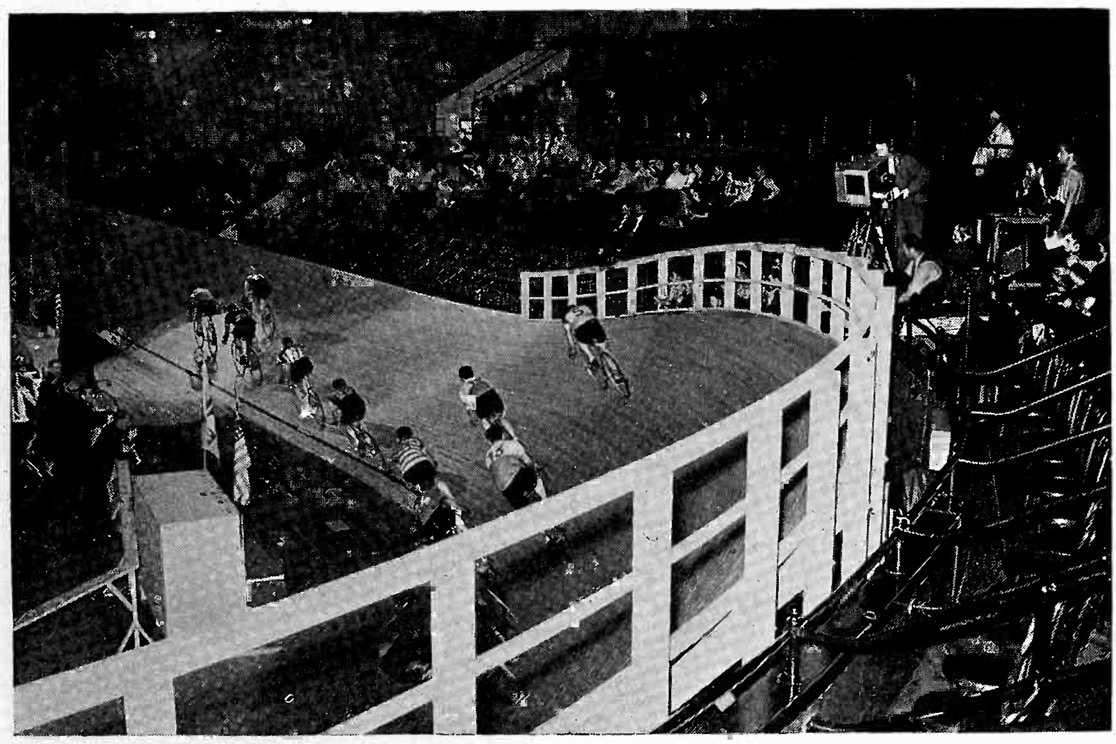

Fig. 3—NBC television camera pick-up of bike race.

over ordinary wires furnished by the New York Telephone Company and provided with special terminal equipment.

The Memorial Day Parade picked up on Riverside Drive not only showed each group unit of the parade, but individual marchers could be recognized easily. In the future many unhappy hours of transportation and waiting will be unnecessary in order to enjoy parades. The camera crew will do the traveling and the viewers may wait in comfort in their homes. The pick-up of the King George VI and Queen Elizabeth parade, during their visit to the United States in June, 1939, was a notable proof of this. Thousands of people, anxious for a sight of Their Majesties, arose at dawn to wait for hours along the line of march for one brief glimpse. Television viewers saw the King and Queen of England as they left the Federal Building at the World's Fair without moving from chairs in their homes.

The ICAAAA Track Meet on Randall's Island, May 26 and 27, 1939, offered another opportunity for program experimentation which was definitely successful. The camera was set up near the finish line and the viewer followed the events with little effort. Again however, the need for additional equipment was felt. While the races which finished in one location—the sprints in particular—were well within the range of one camera, the start of the races could not be covered. They took place too far away to have pictorial value. This was true also of the pole vault, javelin and other field events. Horse races, polo and football will require, not one or two cameras, but three or four, if the audience

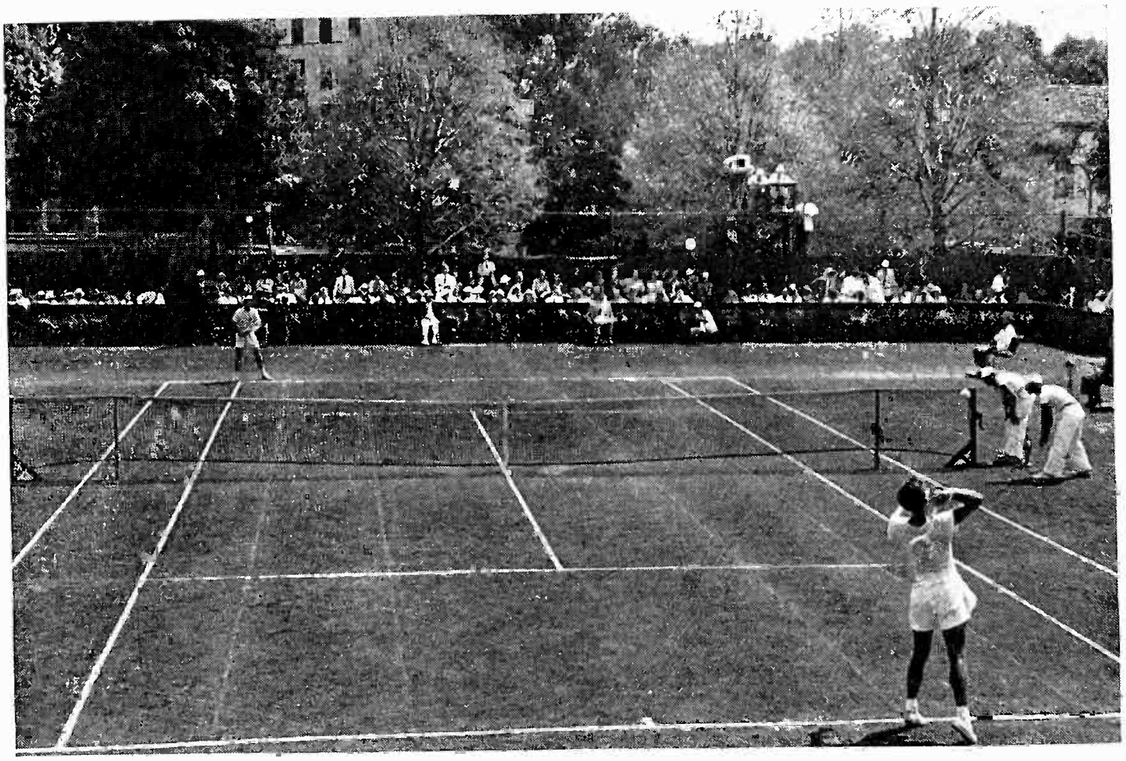

Fig. 4—NBC television camera pick-up of Eastern Grass Court Tennis Tournament, Rye 1939.

is to see everything. To give complete pictorial coverage, cameras must be installed at every point where vital action takes place.

The Baer-Nova fight on June 1st brought television into its own. Boxing contests are ideal sporting events for this new medium. Everything tends to aid the cameraman at a boxing contest—a restricted field of action, concentrated lighting, and but two central figures to follow. Television covered the Baer-Nova fight ably and well.

One thing brought out in the boxing program that is of interest from an audience point of view, was the difference in home reaction between radio and television. During the broadcast reception was switched from visual picture to the sound broadcast. Every viewer objected strenuously, and all realized the enormous addition made by sight to the interest of the event.

This program demonstrated that television announcers must develop an entirely new microphone technique. The audience is able to see the same thing the announcer sees at the same time he sees it; it is quite obvious that any blow-by-blow description becomes unnecessary, and frequently annoying. In describing any event by sound alone, the radio announcer tries to give his audience a word picture of the action taking place before him. In television the audience sees that action. The announcer should function therefore, as a commentator on things of interest outside the camera's field of vision, or as an expert, clarifying and interpreting the televised event. Sound holds secondary interest in television broadcasts and this is particularly true in mobile unit pick-up.

The New York World's Fair has been the scene of many interesting pick-ups. The exhibit at the RCA Building has shown thousands of visitors a brief glimpse of television. As program features there have been broadcasts from the Federal Building and the Court of Peace. The first telecast beauty contest was held when a "Television Girl" was selected from among over 300 young women employed at the Fair. The Parade of the Centaurs, Sun Valley, Jungle Land, and the operation of drilling for oil, were all brought to the viewer by the means of the mobile unit.

Almost every step in television programming involves the discovery of new methods of procedure and operation, and this indicates the magnitude of the problem. There are no men in this country trained in television mobile-unit operation except those employed by the National Broadcasting Company. There is at this time only one television mobile unit. The engineers and program directors who are operating this unit have no guiding precedents. That they are learning rapidly was amply demonstrated in the broadcast of the double header baseball game between the Brooklyn Dodgers and the Cincinnati Reds from Ebbets Field on August 26, 1939. The favorable audience reaction was a definite proof of progress.

The life saving demonstration televised at Atlantic City for members of the National Association of Broadcasters was an excellent program job. Swimming meets at Manhattan Beach and Astoria Pool further demonstrated that the field staff can handle a wide variety of events.

A basic fact to be considered in programming television, and one which affects all "live talent" subject matter whether in or out of the studio, is that for the first time in the history of the entertainment world—or for that matter, in the lives of the people of America—the viewing audience is able to see, to look in on events as they happen. Every motion picture that has ever been shown is a visual record of something that has already taken place, and every news reel is a record of past history before it is seen by the public. With television it is different. The viewer before a television screen sees history in the making. He is as much a spectator of the event taking place as is the

Fig. 5—NBC television camera pick-up of Brooklyn-Cincinnati baseball game.

camera man making the pick-up. In mobile unit pick-ups no one knows the answer to the question "what is going to happen next?" Anything may happen, and if the director and camera man are alert, every viewer will see whatever does happen at the very moment it happens. Our first incident of this kind happened on November 14th, 1938, when we were conducting a test pick-up of Rikers Island. The test was proceeding as scheduled with the camera focused on the island about a quarter of a mile away, when without warning smoke began to issue from one of the buildings on the Island, and television made its first fire scoop. Viewers saw the fire before anyone else could be notified through other means of communication. The television viewer in effect was at the scene. Obviously this success from a program point of view was accidental but every viewer saw history being made. In their own homes, they saw the fire start, saw the fire boats arrive, saw the fire subdued, and that is what outside television pick-ups can do.

Television broadcasting stations can not guarantee the thrill of a fire, but if a fire does occur within camera range the viewer will see it The television camera becomes the eyes of the distant viewer, the device that enables him to look in on a world beyond his horizon. The horizon extension which is afforded by the mobile unit is not great today perhaps, but who can say what it may be within a few years? The country, perhaps much of the whole world, may be brought to our windows.

It seems to be quite true that distance lends enchantment even to familiar and common scenes. The simple pick-up of a television camera pointed out of a window at a busy Metropolitan street is interesting. Life is seen as it happens—commonplace things it is true, but they have a new interest when one can sit at home and view them. Television broadcasts of Times Square, Columbus Circle, City Hall Park, Radio City, Coney Island and Wall Street, with ordinary people going about their daily routines, will probably become frequent television pick-ups in the near future.

The television mobile unit will enable Mr. and Mrs. America to look in on outstanding events in this country of ours—not as viewers of happenings in the past, but as participating spectators of events as they occur. Mr. and Mrs. America will *see* history being made.

ULTRA-HIGH-FREQUENCY PROPAGATION FORMULAS

By

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Summary—*Propagation formulas applicable to the usual problems of ultra-high-frequency propagation within "optical" distances are given. The constants are so chosen that heights and distances are in terms of feet and miles, respectively. In the appendix are given equations for solving problems where the constants are such that the usual approximations do not apply. A tabulation of typical ground constants is given.*

IN RADIO circuits operated at frequencies above 50 Mc, probably a majority will have geometrical properties similar to Figure 1. For such circuits the field strength received at R for a given amount of power radiated at T can be determined by calculations based on the relationships of physical optics. If we apply certain limitations in the constants of Figure 1, the calculations may be made according to relatively simple formulas.

LIMITATIONS

This discussion is based on the assumption that the radio circuit constants fall within the following limitations:

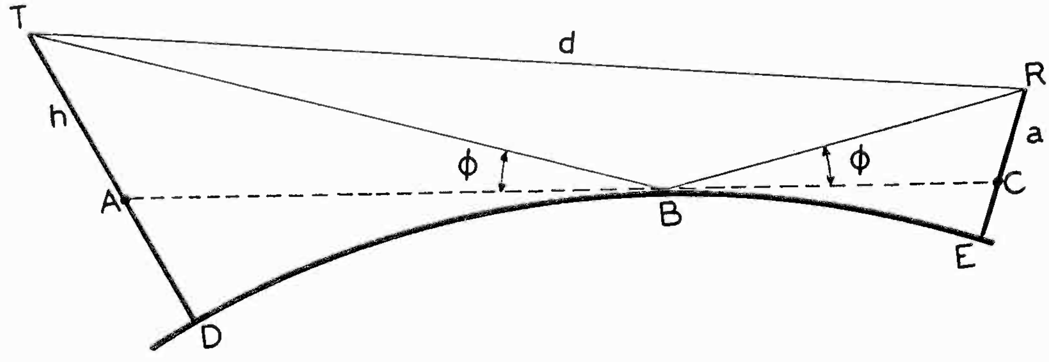

Fig. 1

1. The signal received at R (Figure 1) consists of a direct component arriving along line TR and a reflected component arriving along path TBR.

2. The coefficient of reflection at B is 100 per cent, and the phase change at reflection is 180 degrees. For horizontal polarization these requirements are generally satisfied in practical cases where reflection is from a surface of earth or water at frequencies above 50 Mc. For vertical polarization these requirements are satisfied for reflection from earth or fresh water at angles ϕ below about ½ degree at frequencies

above 50 Mc, but more comprehensive formulas must generally be used for cases of vertical polarization where reflection is from salt water or where the angle ϕ exceeds about ½ degree.

3. The path length *TR* is shorter than the reflected path *TBR* by less than one-sixth wavelength. This condition is satisfied when:

$$\frac{haF}{d} < .433 \times 10^6 \qquad (1)$$

where $h = TA$ in Figure 1, expressed in feet.
$a = RC$ in Figure 1, expressed in feet.
$F =$ frequency in Megacycles per second (Mc).
$d =$ distance *TR* in Figure 1 expressed in miles.

The values h and a in Figure 1 may be determined if heights *TD* and *RE* and distance d are known. The position of *B* is determined by the requirement that angle *TBA* must equal angle *RBC*. Knowing distances *BD* and *BE*, heights *AD* and *CE* may be calculated from the relationship:

$$AD = \frac{BD^2}{2} \text{ and } \qquad CE = \frac{BE^2}{2} \qquad (2)$$

where *AD* and *CE* are in feet, whilst *BD* and *BE* are in miles. This relationship takes into account the average refraction of the earth's atmosphere. The effect of refraction is to cause a slight downward bending of the signal paths, and it has been demonstrated[1] that this effect can be taken into account if we assume the paths of propagation are straight, but that the curvature of the earth's surface is that which would be obtained if the radius were increased by 33 per cent.

4. For the case of vertical polarization the antennas are assumed to be at least two wavelengths above the earth's surface.

PROPAGATION FORMULA

If the above limitations are satisfied, the received field at *R* will be:

$$\varepsilon = \frac{.01052 \sqrt{W}\, h\, a\, F}{d^2} \text{ microvolts per meter} \qquad (3)$$

where $W =$ watts radiated by half-wave dipole at *T*.
h and $a =$ altitudes above reflecting surface in feet.
$F =$ frequency in Mc.
$d =$ distance in miles.

Received Power

Assuming reception is by means of a half-wave dipole with a radiation resistance of 75 ohms, matched to a load circuit, the watts absorbed in the load circuit will be:

$$W_r = \frac{30.4\ \varepsilon^2}{F^2} \times 10^{-12} \text{ watts} \tag{4}$$

Combining (3) and (4) gives:

$$W_r = \frac{3.37\ W\ h^2 a^2}{d^4} \times 10^{-15} \text{ watts} \tag{5}$$

Attenuation

The attenuation in a radio circuit may be expressed in terms of the ratio of power transmitted to power absorbed in the receiver. By converting (5), we obtain the following relationship:

$$\frac{W}{W_r} = \frac{2.97\ d^4 \times 10^{14}}{h^2 a^2} \tag{6}$$

This power ratio may, of course, be converted into decibels.

As previously stated, the above relationships are on the basis of half-wave dipoles at the transmitter and receiver. At ultra-high frequencies it is generally convenient to use antennas which yield a power gain. The effect of this power gain is to reduce the overall attenuation in the circuit. Taking antenna gains into account, the ratio of transmitted power to power absorbed in the receiver becomes:

$$\frac{W}{W_r} = \frac{2.97\ d^4 \times 10^{14}}{h^2 a^2\ G_1 G_2} \tag{7}$$

where G_1 = power gain of transmitting antenna
G_2 = power gain of receiving antenna.

If both antenna gains are equal, this becomes:

$$\frac{\mathrm{W}}{W_r} = \frac{2.97\ d^4 \times 10^{14}}{h^2 a^2\ G^2} \tag{8}$$

For a given area of antenna array, the power gain is generally proportional to the square of the frequency. For parabolic antennas the power gain is approximately:

$$G = 8\ S\ F^2 \times 10^{-6} \tag{9}$$

where S is the area of aperture in square feet.

Substituting this value in (8), we have for the case of equal parabolic antennas at T and R, the relationship:

$$\frac{W}{W_r} = \frac{4.64\, d^4 \times 10^{24}}{h^2 a^2 S^2 F^4} \tag{10}$$

RADIO RELAY

In a chain of radio-relay links over flat country the heights and distances may be approximately equal. On the assumption that they are equal, certain simplifications can be made in accordance with the

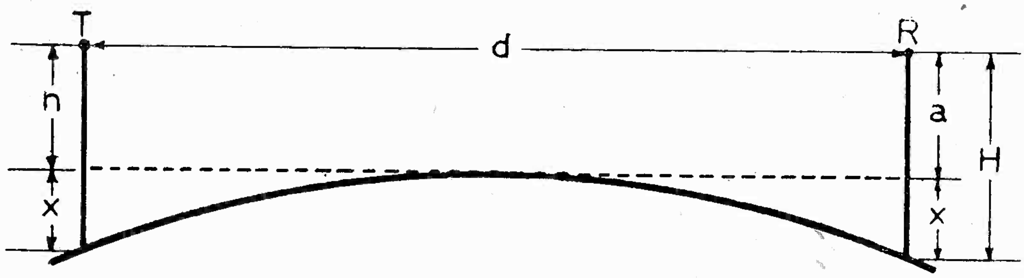

Fig. 2

geometry of Figure 2. Here h and a are equal. Also the tower heights H are equal. The height of the tangent plane may be calculated:

$$X = \frac{d^2}{8} \tag{11}$$

where X is in feet and d is in miles.

Using (11), the product $h \cdot a$ can be expressed as a function of tower height and distance:

$$h \cdot a = \left(H - \frac{d^2}{8}\right)^2 \tag{12}$$

Substituting in (10) gives the relationship:

$$\frac{W}{W_r} = \frac{4.64\, d^4 \times 10^{24}}{\left(H - \frac{d^2}{8}\right)^4 S^2 F^4} \tag{13}$$

where $\frac{W}{W_r}$ is the power gain required at relay point.

d = link distance in miles.
H = tower heights in feet.
S = parabolic antenna aperture area, sq. ft.
F = Frequency, Mc.

Thus it is seen that within the limitations previously cited, the ratio of transmitted to received power, with directive antennas of a given size, is inversely proportional to the fourth power of frequency. This means that much less transmitter power will be required at the higher frequencies to produce a given carrier level at the receiver over circuits of the type being discussed. For instance, 10 watts at 300 Mc should be as effective as 810 watts at 100 Mc.

Appendix I

For situations where the constants do not fall within the limitations defined at the beginning of the above discussion, the received field may still be calculated.

If T and R of Figure 1 lie above the tangent plane through the point of reflection, the received field will be due to two components, the direct and the reflected. The resultant field strength is determined by the vectorial combination of these two components.

The component arriving over the direct path will be:

$$\mathcal{E}_o = \frac{4.35\sqrt{W} \times 10^3}{d} \; \mu v/m \qquad (14)$$

where d is in miles and W is power radiated by a half-wave dipole.

The amplitude of the reflected component may also be determined from (14), substituting for d the total length of the reflected path, and multiplying by the coefficient of reflection.

To combine vectorially the direct and reflected components it is also necessary to know the phase angle between them. This phase angle is made up of two parts, one due to the difference in path lengths, and the other due to the phase of the reflection coefficient. The phase angle introduced by the difference in path lengths may be calculated from the geometry of the case (Figure 1).

The phase angle introduced by the reflection coefficient as well as the magnitude thereof may be calculated according to the following equations:[2]

$$K_v = \frac{\epsilon_o \sin\phi - \sqrt{\epsilon_o - 1 + \sin^2\phi}}{\epsilon_o \sin\phi + \sqrt{\epsilon_o - 1 + \sin^2\phi}} \qquad (15)$$

for vertical polarization and:

$$K_H = \frac{\sin\phi - \sqrt{\epsilon_o - 1 + \sin^2\phi}}{\sin\phi + \sqrt{\epsilon_o - 1 + \sin^2\phi}} \qquad (16)$$

for horizontal polarization, where

$$\epsilon_0 = \epsilon - j\frac{2\sigma}{f} \qquad (17)$$

and where

ϵ = dielectric constant of reflecting medium.
σ = conductivity of reflecting medium in e.s.u.
f = frequency in cycles per second.

It will be noted that where the term $\frac{2\sigma}{f}$ is less than about 10 per cent of ϵ, the value ϵ_0 is a simple number and the phase of the reflection coefficient is either zero or 180°.

The tendency is for the phase of the reflection coefficient to be nearly 180° for all cases of horizontal polarization. For vertical polarization at ultra-high frequencies the phase is generally approximately 180° when the angle ϕ is less than Brewster's angle and approximately zero when ϕ is greater than Brewster's angle.

Brewster's angle is the angle at which cot $\phi = \sqrt{\epsilon}$ when conditions are such that $\frac{2\sigma}{f}$ is inconsequential relative to ϵ.

Whilst considerable data are available relative to earth constants,[3, 4] the tabulation of a few example herein might be helpful:

Type of Ground	*Dielectric Const.* ϵ	*Conductivity* σ, *e.s.u.*
Sea Water, Long Island and Hawaii...	80	$40{,}000 \times 10^6$
Fresh Water, Long Island Pond.......	80	45×10^6
Sandy Soil, Long Island	9	1×10^6
Dry Soil	4	$.1 \times 10^6$
Fertile Farm Land	15–25	$50–150 \times 10^6$

The author would also like to call attention to an excellent set of curves presented by C. R. Burrows giving calculated values of the coefficient of reflection.[5]

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SIMPLE TELEVISION ANTENNAS*

By

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Summary—*The frequency band widths demanded by high-definition television have considerable range when considered in relation to resonant circuits. The transmitting antenna and transmission-line systems must, therefore, meet stringent requirements if multiple images or ghosts in the received picture are to be avoided.*

Before discussing the characteristics of particular antenna systems, the transmitting and receiving antenna problems are considered. The input impedance of a transmission line, even when loaded with a resonant circuit having a Q as low as 2, undergoes considerable variation with frequency within the transmission band. If the television receiver is designed to present a pure resistance to its transmission line, equal to the characteristic impedance of the latter throughout the transmission frequency band, the receiving antenna requirements are not difficult to meet.

The measured impedance-frequency characteristic of a half-wave dipole of large diameter conductors, when compared with that obtained for a similar antenna of small diameter conductors, shows the advantages of the former.

A method of impedance matching has been devised whereby the usual narrowing of the useful frequency band caused by impedance transformation is overcome.

The "folded dipole" antenna and combinations of these units are superior to ordinary dipoles for television purposes. Measurements indicate that ground and other reflecting surfaces considerably affect the impedance-frequency characteristics of antennas.

The use of a type of antenna called the "double cone" or "hour glass" antenna results in a very flat impedance-frequency characteristic at the input terminals of a transmission line over a wide range of frequency. By properly proportioning the dimensions of this antenna its impedance can be made to match the characteristic impedance of all practical open-wire transmission lines. The current and electric field distributions along the surfaces of the conical conductor have been measured. The theory of this antenna is briefly considered.

Curves of the characteristics of the systems discussed are included. For comparison purposes the measurements of line reflection vs. frequency for the several antenna systems considered are shown by a family of curves in a single figure.

The frequency band demanded by high-definition television transmission has a width of a few megacycles. Its ratio to the carrier frequency is of the order of 10 per cent to 15 per cent, a very substantial range when considered with respect to resonant circuits. The transmission lines used in conjunction with television antennas, whether for reception or transmission, rarely have lengths equivalent to less than two or three wave lengths.

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Unless the impedance of a transmitting antenna perfectly matches the characteristic impedance of its associated transmission line at all frequencies within the transmitted band, displaced images or ghosts may appear on the screen of the receiver. Ghosts may also be caused by imperfect matching between the input impedance of the receiver and the characteristic impedance of its transmission line. In addition to the ghost phenomena the pictures may be more or less distorted by poor impedance matching between the transmitter and its transmission line or between the receiving antenna and its transmission line. It would thus appear that the impedances at both terminals of both the transmitter and receiver transmission lines should be perfectly matched throughout the frequency band. Since practical transmitters

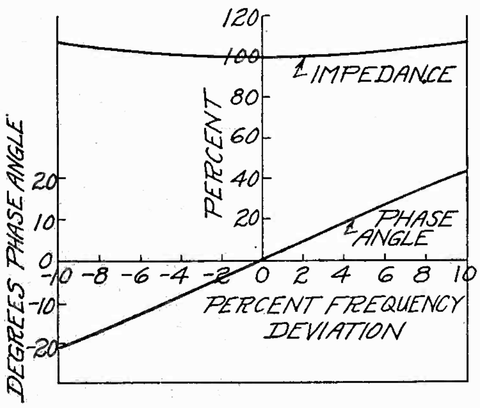

Fig. 1

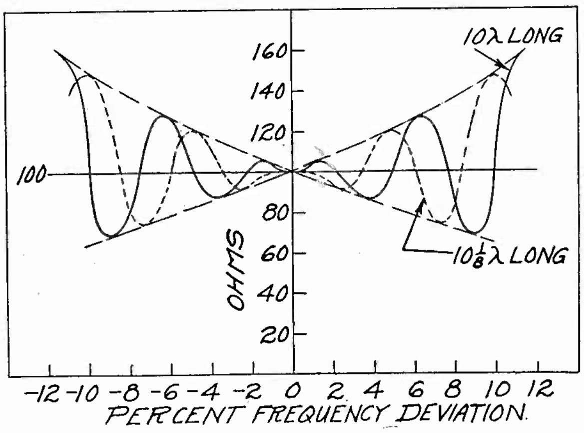

Fig. 2

and receivers are not electrically equivalent in all respects it seems desirable to consider each end of the space circuit as a separate problem before investigating particular antenna systems.

The Transmitting-Antenna Problem

Unless the transmitting antenna presents a pure resistance load to its associated transmission line and in addition this resistance is equal to the characteristic impedance of the line at all frequencies within the transmission band, the input impedance of the system, as seen by the transmitter, will vary with frequency. The phenomena involved and the approximate rules governing them can best be explained by a simple though somewhat idealized combination. Let us assume that a transmission line ten wave lengths long feeds an antenna having an impedance-frequency characteristic like that of a simple series-tuned circuit. Also, let us suppose the circuit to have a resistance of one hundred ohms, a reactance of two hundred ohms, or Q of two, and that the characteristic impedance of the transmission line is one hundred ohms. Figure 1 shows the impedance characteristic of the load circuit by itself. The magnitude of the impedance is nearly constant through a frequency band of 20 per cent of the resonant frequency, and, at first sight the antenna represented by this circuit might appear ideal for television transmission. Figure 2 shows the input impedance of the line terminated with this antenna, which is far from constant.

The large oscillation in the curve of Figure 2 is due to reflection caused by the reactive component of the load impedance since the magnitude of the impedance is nearly constant over a considerable range of frequency. Charts have been developed by the writer by which the magnitude and phase angle of the reflected wave and the complex input impedance of the transmission line may be rapidly determined

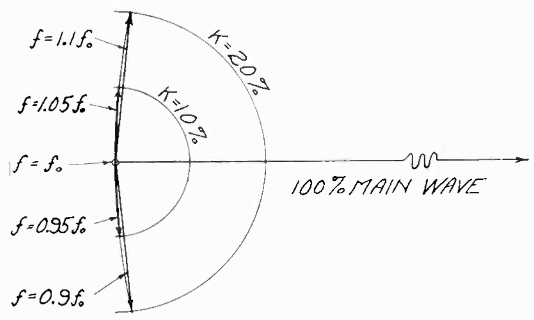

Fig. 3

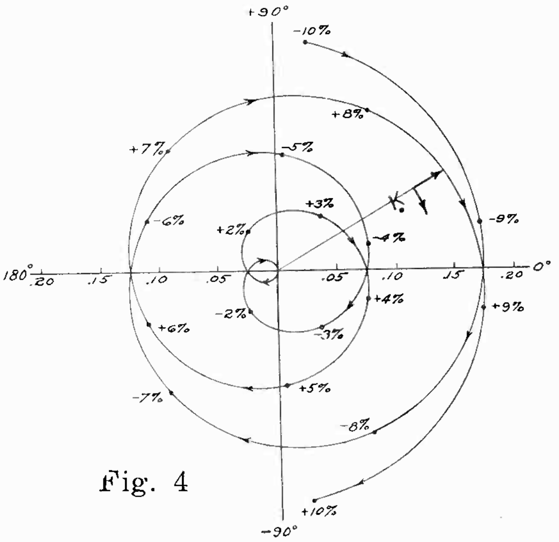

Fig. 4

for all conditions of load impedance.[1] Curves like that of Figure 2 are easily computed from the charts. The dashed curve of Figure 2 shows the effect of increasing the length of the line one-eighth wave length. It will be noted that this curve is symmetrical with respect to the resonant frequency as well as the solid curve for a length of 10 wave lengths. Such symmetry exists for lengths equal to any integral multiple of an eighth wave length. An increase of the length to 10¼ wave lengths results in an impedance curve approximately equivalent to the solid curve turned upside down. In order to avoid confusion only two curves are shown in the figure but the limits of the oscillation of the input impedance for any length of line are the two envelope curves indicated. Figure 3 shows the variation in the reflected wave vector at the load for the conditions assumed, as the frequency is increased from 90 per cent to 110 per cent of the resonant frequency. It will be noted that the reflected-wave vector leads or lags the main-wave vector

[1] Charts for Transmission-Line Measurements and Computations by P. S. Carter, RCA REVIEW, January 1939.

by nearly 90 degrees for frequencies above or below resonance respectively.

Let us consider the relation between the input impedance of the line and the reflection coefficient. Line length in terms of wave length is directly proportional to frequency and therefore a 10 per cent increase in frequency increases the length of our 10 wave length line by one wave length. As we proceed from the antenna toward the transmitter a vector representing the main wave is rotated in a leading direction through an angle of 10 times 360 degrees at the carrier frequency, while the vector representing the reflected wave is rotated in the lagging direction by the same angle. At the input end of the line, the relation between the main wave and reflected-wave vectors is the same as though we had rotated the reflected-wave vector through an angle of 20 times 360 degrees without rotating the main

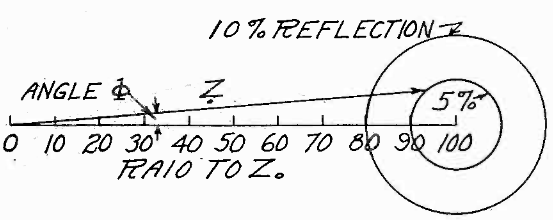

Fig. 5

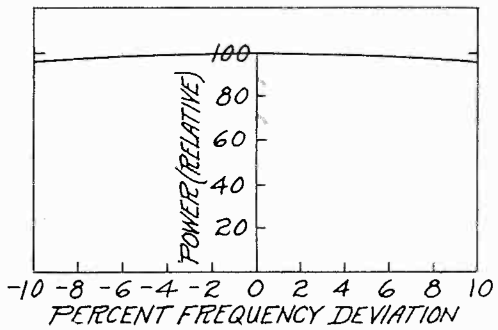

Fig. 6

wave vector. Figure 4 shows the trace of the terminus of the reflected-wave vector at the input terminal of our system when we assume the main vector fixed and the frequency is increased from 90 per cent to 110 per cent of the resonant frequency. The resulting input impedance has already been shown in Figure 2.

If the reflected-wave vector were of a constant magnitude of 5 per cent throughout the band the input impedance would oscillate between 90 and 110 ohms as the frequency varies. The reason for this is that when the frequency is of such a value as to bring the reflected and main-voltage vectors into phase the vectors representing the main and reflected current waves are in phase opposition. The resultant voltage is then 105 per cent of the main-wave voltage and the resultant current 95 per cent of the magnitude of the main-current wave vector. The input impedance is therefore $(105/95) \times 100 = 110$ ohms, approximately. When the frequency is such that the main- and reflected-voltage vectors are in phase opposition the above relations are reversed and the input impedance becomes approximately 90 ohms.

If we agree on a maximum allowable variation in the input terminal impedance, we may lay down tolerance rules. Let us assume the allowable limits of this impedance to be 90 per cent and 110 per cent of the characteristic impedance. The reflection coefficient must then be less than 5 per cent throughout the frequency band. If the magnitude of the antenna impedance remains substantially constant and equal to the characteristic impedance of the line the reactance component, whether capacitive or inductive, must not be greater than 10 per cent of the resistive component. In most practical systems both components vary with frequency and the terminus of the load impedance vector must lie within the circle as shown in Figure 5.

The transmission lines for use with television transmitting installations usually can be designed to have efficiencies of better than 90 per cent; and the consideration of power losses may, therefore, be neglected without seriously affecting the conclusions resulting from the present discussion.

If the transmitter is either very loosely coupled to its transmission line so as to act, in effect, as a constant-voltage generator, or very

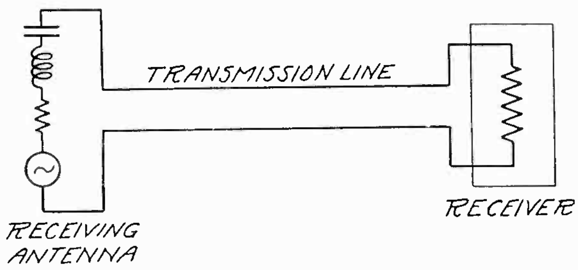

Fig. 7

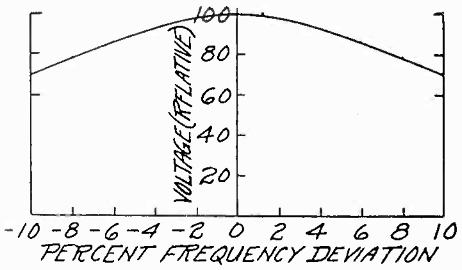

Fig. 8

tightly coupled so that it becomes approximately equivalent to a constant-current generator, the power radiated from the antenna would have the same variations as the impedance curve of Figure 2. The power output under either assumption would be very low. If the transmitter could be represented by a voltage in series with a pure resistance equal to the characteristic impedance of the line the power output of our assumed antenna system would be nearly constant over the frequency band in spite of the wide variations in the input terminal impedance of the transmission line. Under this assumption the power output would then be as shown in Figure 6. However, such an equivalent circuit representation for an actual transmitter is far from justified. An alternator in series with a series-tuned circuit having a low value of resistance relative to the characteristic impedance of the transmission line is a more nearly true equivalent. The variation in radiated power will then be considerable. We, therefore, conclude that in a practical television transmission system the input terminal impedance of the transmission line feeding the antenna should show little variation with frequency within the transmission band.

THE RECEIVING-ANTENNA PROBLEM

Within regions where the field strength is high a television receiver may be so designed as to have a substantially constant input resistance equal to the characteristic impedance of its associated transmission line within the television transmission band. The receiving antenna is equivalent to a generator in series with a complex impedance. Both the resistance and reactive components of this equivalent impedance vary with frequency but as a first approximation the antenna may be represented by an alternator together with a series tuned circuit, the equivalent circuit for the complete system being as shown in Figure 7.

Let us assume the receiver to act as a pure resistance and the

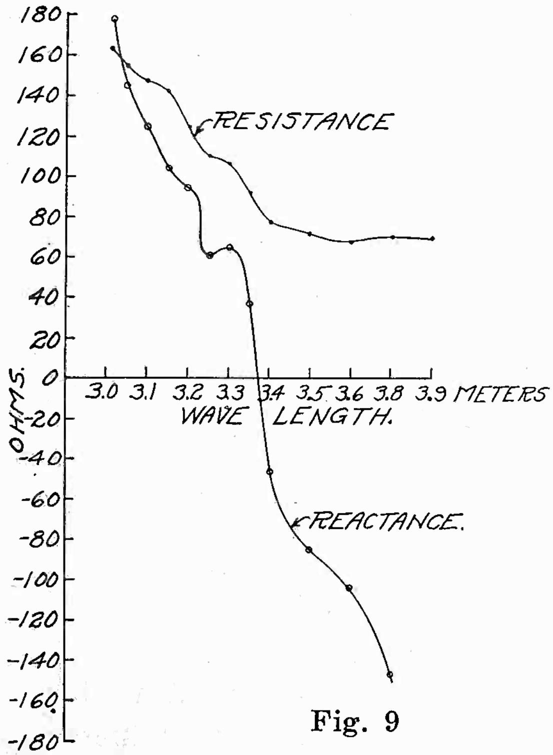

Fig. 9

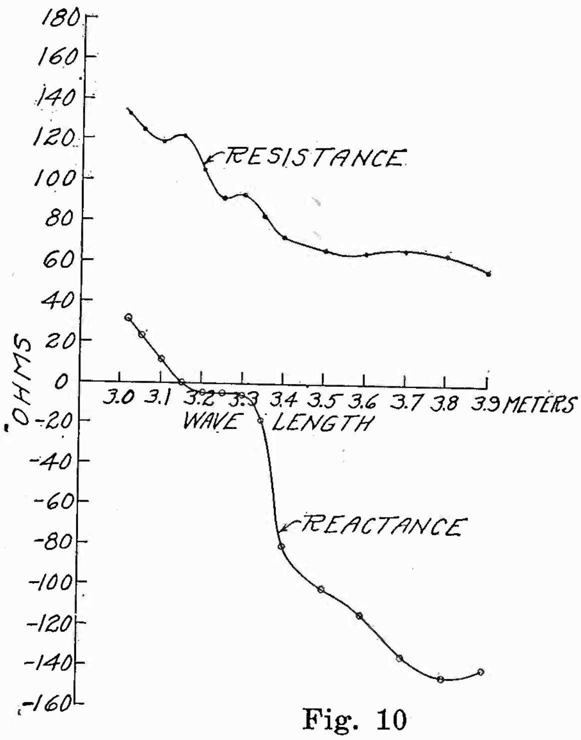

Fig. 10

antenna to be the same as that discussed in the transmitting problem where the equivalent circuit has a Q of 2. If the transmission line is designed to match the resistance of the antenna, the power in the receiver is the same as that shown in Figure 6 for the transmitting condition where the transmitter is an alternator in series with a pure resistance equal to the characteristic impedance of the line. Thus we see that, due to the fact that a receiver may be designed to have a constant resistance over a considerable band, an antenna and transmission line system which is far from satisfactory when used with a transmitter may serve as an excellent receiving system.

Figure 8 shows the voltage developed across the input resistance of a receiver when the receiving antenna has an equivalent Q of 10, and the receiver is assumed to be matched to the transmission line. This curve is true regardless of whether the characteristic impedance of the transmission line matches the antenna resistance or not. However, maximum signal strength is, of course, obtained when these impedances are matched. This figure, therefore, indicates that an antenna with the relatively high Q of 10 may be satisfactory for reception purposes.

For locations where the signal strength is relatively low, receiver amplification cannot well be sacrificed in order to obtain a wide-band constant-resistance input characteristic. A good match between the antenna and transmission line impedances is then desirable for two reasons,—(1) Under these conditions the receiver circuits are quite similar, insofar as the transmission line is concerned, to the transmitter circuits previously discussed. Multiple images may be caused by impulses which, after having been reflected by the receiver, are

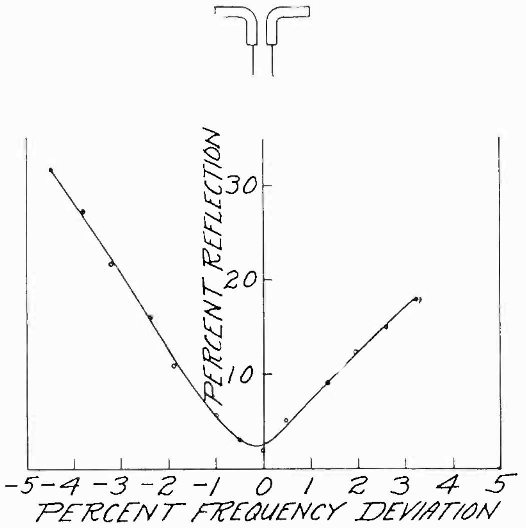

Fig. 11

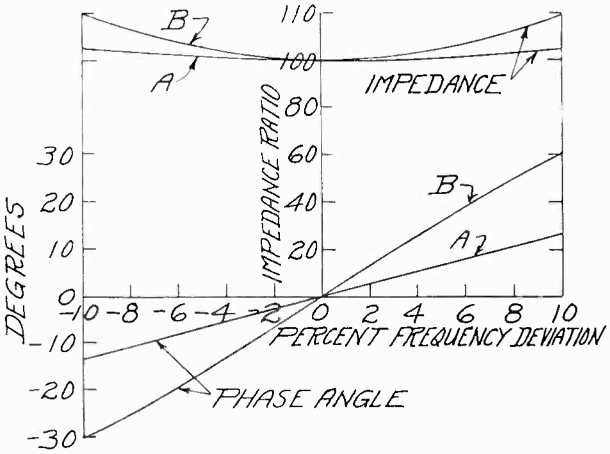

Fig. 12

again reflected at the antenna and returned to the receiver at a later time corresponding to the time required for the wave to make one or more round trips to the antenna, i.e. roughly one-tenth microsecond for each fifty feet of line. (a) Noise generated in the receiver itself becomes comparable with signal strength and the maximum energy should be drained from the antenna to overcome this noise. This can only be accomplished when the line and antenna impedances are equal.

Some transmission lines in use with receivers have efficiencies lower than 10 per cent even when their lengths are not greater than 100 feet. The use of an efficient line together with a proper matching of antenna and line impedances may, in some instances, give an improvement of 100 to 1 in received power or 10 to 1 in received signal voltage.

HALF-WAVE DIPOLE ANTENNAS

Figure 9 shows the impedance vs. frequency characteristic of a simple horizontal half-wave dipole of No. 10 wire when fed by an open, No. 10 wire, transmission line. The height above ground was 180 cm or a little under a half wave length at the resonant frequency. The ripple in these curves is without doubt due to the wave reflected from the side of a nearby sheet-metal building, since the frequency between ripple crests corresponds quite accurately with the time of travel from the antenna to the building and back again.

Figure 10 is a similar characteristic for a dipole of 1½-inch copper pipe positioned at the same height and fed by the same line. By comparing these two figures the effect of the increase in the diameter is seen. The reactance of the pipe antenna is less than 10 per cent of the resistance over a band width of nearly 7 per cent but within this range the resistance changes from 80 to 120 ohms even if we average out the ripples. The larger-diameter antenna is very much superior

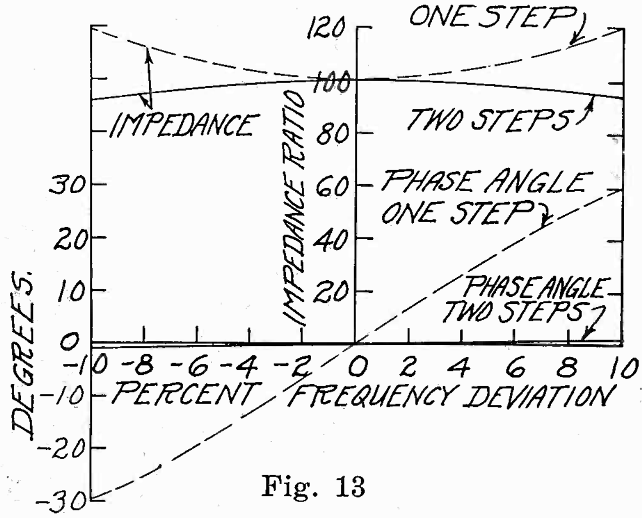

Fig. 13

to the No. 10 wire dipole because of the improved reactance characteristic of the former even though the improvement in the resistance variation is not very great.

The 1½-inch pipe dipole might be satisfactory for some transmitting installations if the band width were not further narrowed by impedance-matching circuits necessary to transform this impedance up to the value of the characteristic impedance of a practical line.

Figure 11 shows the reflection coefficient vs. frequency for a 1½-inch pipe dipole and quarter-wave impedance matching system in which the pipes are bent so that a single section of pipe serves both as a leg of the dipole and as one conductor of the impedance matching circuit. The band width at which the reflection is less than 5 per cent is only 1.5 per cent.

Before proceeding further it would appear well to consider briefly the effect of impedance matching circuits upon the impedance-frequency characteristics of antenna systems. Let us assume a quarter-wave line section loaded with a pure resistance. When this line is used for transforming in the impedance ratio of 2 to 1, the resulting input terminal impedance characteristic is that shown by curves A of Figure 12. When used for an impedance transformation ratio of 16, the characteristic is as shown in curves B of the same figure.

It is apparent from these curves that the detrimental effect of quarter-wave impedance matching circuits is in proportion to the magnitude of the impedance transformation ratio. This law generally holds for all types of impedance matching circuits. It will be observed that the quarter-wave line circuit gives a symmetrical input impedance characteristic. Some types of circuits, such as circuits formed with shunt reactances at particular positions along the transmission line

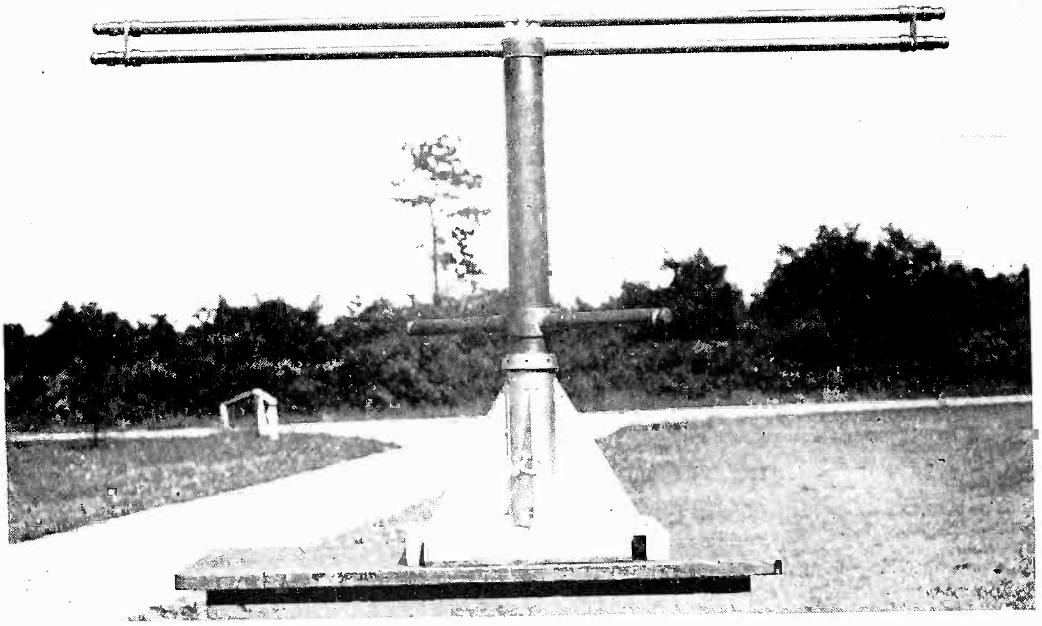

Fig. 14—Dipole for 81-86 Mc frequency sweeping transmitter used in television propagation survey.

produce unsymmetrical impedance-frequency characteristics. We shall not take the time to consider other types of circuits or the impedance matching problem in detail.

The difficulties due to an impedance transformation may be almost entirely overcome by dividing the transformation into two or a greater *even* number of steps. As an illustration let us assume that we use two quarter-wave line sections in series to transform a load impedance from $16R$ to R. If we make the characteristic impedance of the line section near the load $8R$ and the remaining section $2R$ the first section transforms from $16R$ to $4R$ and the second section from $4R$ to R. Figure 13 shows the input impedance characteristic when such a system is used; the curve for the same transformation in one step being shown in broken lines. The reactance present when a single step is used has been practically eliminated by the use of two steps.

The Folded Dipole

Figure 14 is a photograph of a transmitting type of antenna which we call "the folded dipole". It was developed in order to do away with impedance matching circuits and their detrimental effects upon the frequency band width and also to obtain a mechanically strong structure. This particular unit was installed at the 85th floor of the Empire State Building (see Figure 15) and has been in use for several months in order to provide a temporary system while changes were being made in the main antenna and transmission line. The impedance-frequency characteristic of this antenna as measured on its associated transmission line is shown in Figure 16. The transmission line was a pair of concentric lines, each having a characteristic impedance of 120

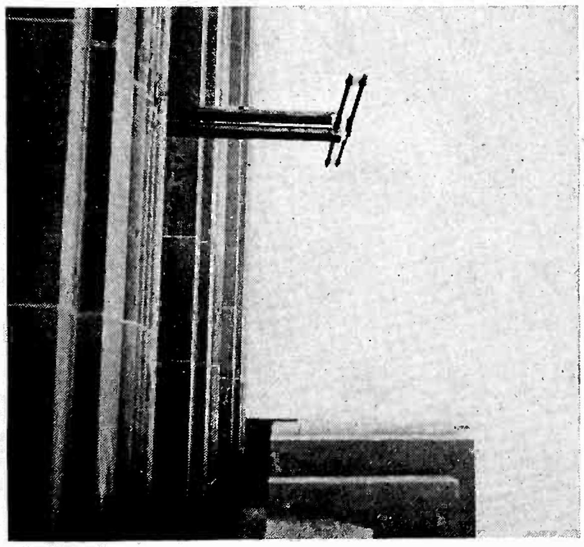

Fig. 15—Folded dipole.

ohms. The antenna perfectly matched the line at 46.5 megacycles, the carrier frequency in use at that time. Other units for survey tests in New York City were located near the top of the tower of the Empire State Building and are shown in Figure 17 and Figure 18.

This type of antenna consists of two closely spaced half-wave dipoles connected together at their ends. One of the dipoles is broken at its center where it is fed from a balanced transmission line. The instantaneous currents in both units are in the same direction in space while both are flowing toward a nodal point at one extremity of the radiating structure. The current distribution does not differ greatly from that for an ordinary half-wave dipole and is approximately sinusoidal.

Since the two radiators are very closely spaced in terms of wave length the radiation pattern is essentially the same as the pattern for an ordinary single half-wave dipole. The total power radiated per total loop current squared, or radiation resistance, is therefore about 73 ohms. However, if the diameters of the two radiators are equal, this same radiated power is equivalent to a radiation resistance with respect to the current in one branch of four times 73 ohms or 292 ohms. The latter value of resistance is that which is seen by the transmission line at its terminals. This type of antenna thus serves the double purpose of a radiator and impedance matching transformer.

When three radiators of equal diameters are arranged in accordance with this method, a transformation ratio of nine is obtained. Any desired ratio of transformation may be obtained by the use of two or more radiators of unequal diameters. In such an arrangement of two units wherein the smaller-diameter cylinder is fed from the

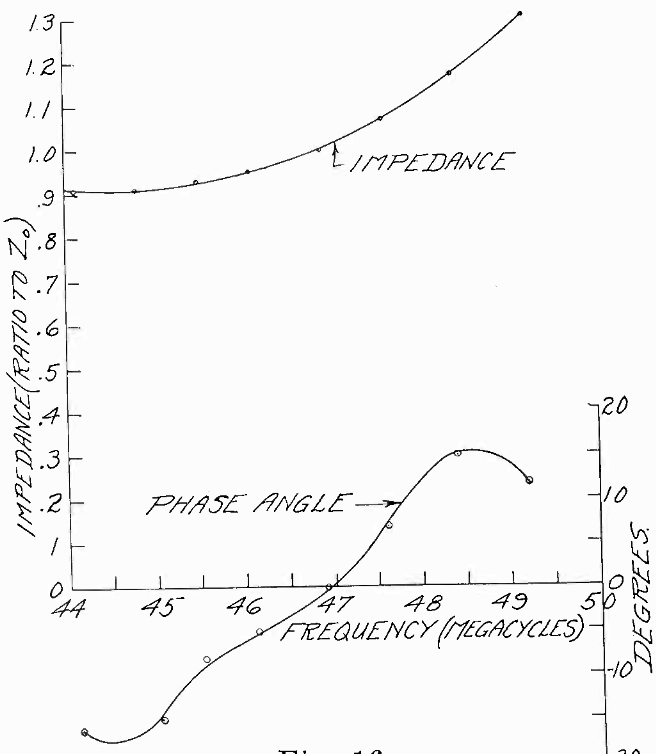

Fig. 16

transmission line the transformation ratio is greater than four since the greater of the two currents flows along the larger pipe.

When a single folded dipole is used with a long transmission line the input impedance-frequency characteristic does not quite meet the most rigid transmitting requirements. However, when two such units are arranged in criss-cross fashion and fed in quarter-phase relation by means of two branch feeder lines differing in length by a quarter-wave length the impedance characteristic presented to the main line at its junction with the branch feeders is quite satisfactory for present band widths. Figure 19 is a photograph of such a unit. The impedance characteristic for such an arrangement is shown in Figure 20. The resulting small reactance component is due to the impedance-inverting property of a quarter-wave line section.

The same advantages insofar as band width alone is concerned may be obtained from an array of four of these units in a square when one pair of units is fed in quarter-phase relation with the other pair in order to provide an antenna for the radiation of horizontally polarized waves. However, unless the square is small in terms of wave length the horizontal radiation pattern is far from circular.

Impedance and Height

The impedance characteristic of an antenna, when located near the ground or other reflecting surfaces, may differ considerably from the characteristic when remote from such surfaces. Figure 21 shows the variation of both resistive and reactive components of the imped-

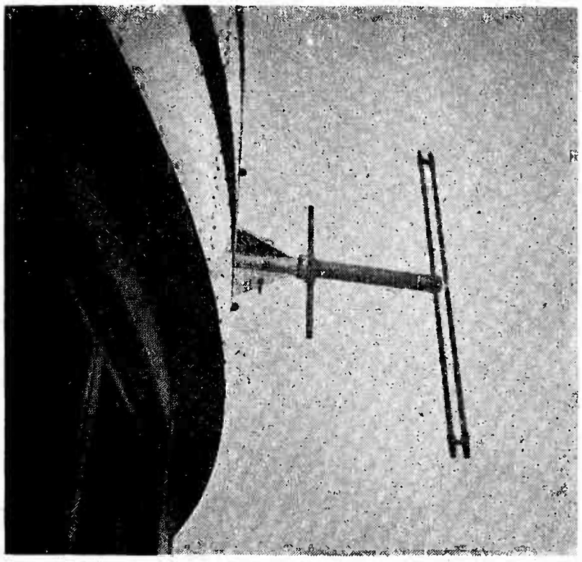

Fig. 17—Folded dipole.

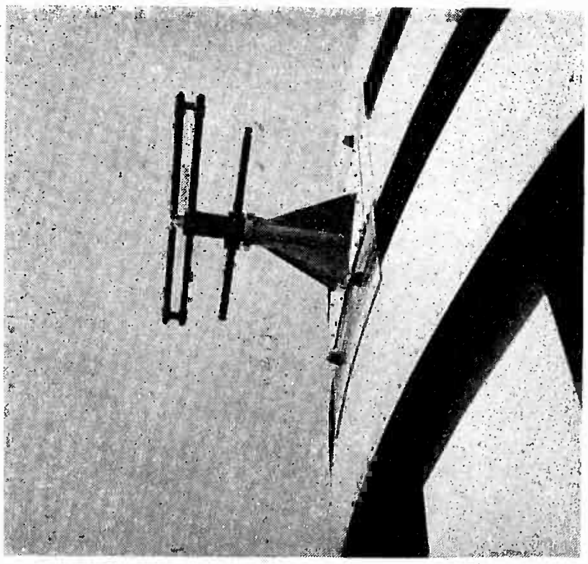

Fig. 18—Folded dipole.

ance of a dipole with the height of the antenna. Data were taken at several different frequencies, but the curve shown corresponds to the frequency at which the average reactance is approximately zero. The measurements were made at Rocky Point, New York, where the soil is mostly quartz sand. The dielectric constant of the soil is around 9 and conductivity is extremely low. In a previous paper[2] the writer has shown a method by which both the resistance and reactance components of the radiation impedance of an antenna above a perfectly conducting plane may be determined. The theoretical curves obtained by this method, assuming a perfectly conducting ground, are shown in dotted lines in Figure 21 for comparison purposes. It will be noted that the variation in impedance for the antenna above the extremely poor ground at Rocky Point is nearly as great as when a perfectly

[2] Circuit Relations in Radiating Systems, Proceedings I.R.E., Vol. 20, No. 6, June 1932.

conducting ground is assumed and that the approximate heights for the maximums may be obtained by adding about 0.1 wave length to the actual height and assuming ideal ground.

The effect of a metal building, at some distance from the antenna, upon the impedance characteristic has already been shown in Figure 9. These two examples of the effect of reflecting surfaces should serve to explain why it is seldom possible to predict the impedance characteristic of an antenna system accurately.

The Double-Cone or Hour-Glass Antenna

Figure 22 is a photograph of an arrangement of two copper cones together with an open two-wire transmission line. The impedance-

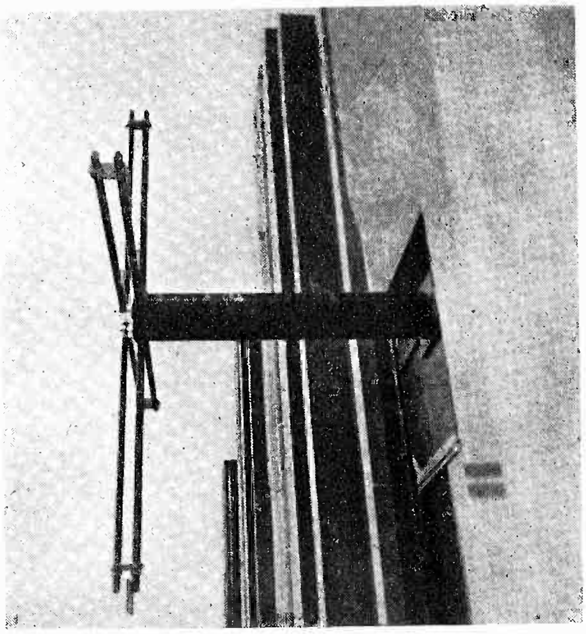

Fig. 19—Folded dipole antenna for study of propagation characteristics of circular polarization.

frequency characteristic is shown in Figure 23. It will be noted that the magnitude of the impedance curve is nearly constant and the phase angle less than 10 degrees within a frequency band of nearly 20 per cent. This characteristic is far superior to that of any of the antennas so far discussed.

The impedance characteristics of a number of double-cone antennas of sheet copper have been measured in order to determine the laws governing the relationship between length, angle of revolution of the cones, and impedance presented to the transmission lines. Tests were also made with several sizes of wire-cage cones. When twelve or more wires were used for the cage with these wires converged into a small solid metal cone to form the apices of the large cones, the impedance characteristics were found to be nearly as good as those for the sheet cones.

Figure 24 shows the effective impedance as a function of angle of revolution of the cones. The length of each cone taken along the surface from apex to rim should be about 0.365 wave length at the mid-frequency regardless of angle of revolution, at least for the impedance values usually encountered in practice.

Figure 25 shows the measured current distribution along the surface of a double-cone antenna. Strictly, the curve represents the product of the magnetic field intensity near the conducting conical surface multiplied by the circumference of the cone at the position of measurement but according to the fundamental laws of Maxwell this product is equivalent to the relative total current flowing along

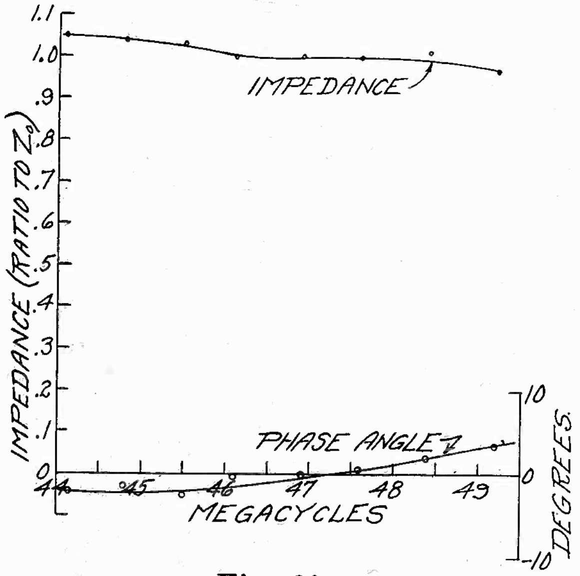

Fig. 20

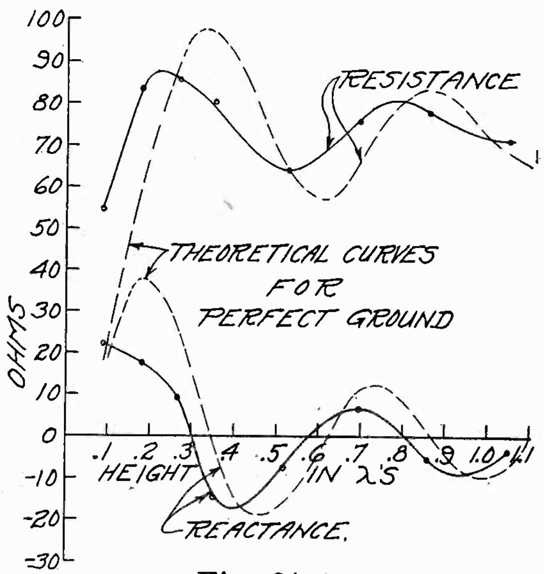

Fig. 21

the surface of the cone. It will be noted that the current distribution departs greatly from the usually assumed sine-wave form and varies little with distance for positions near the apex.

The distribution of the voltages between the surfaces of the conical sheets is also shown in Figure 25. In order to interpret these two distribution curves and obtain an understanding of the nature of the phenomena involved it is necessary to consider the field configurations of the electromagnetic waves in the vicinity of such conducting surfaces.

Radio engineers are familiar with the principles of radio-frequency transmission lines but usually consider the phenomena in terms of inductance and capacitance visualized as coils and condensers. In order to obtain a clear understanding of electromagnetic wave propagation we must take the more fundamental viewpoint of Maxwell, Faraday, and Heaviside. The energy is carried by the space surrounding the conductors while the currents flowing in the surfaces of the conductors of a two-wire transmission line are of secondary importance only. The small portion of energy which penetrates the wires serves no useful purpose and is wasted in heat. In any wave front, which for a transmission line is a plane cutting the wires at right angles, the lines of electric force are arcs of circles terminating on the conductors while the lines of magnetic force are a complete system of circles surrounding one or the other of the conductors. From the inverse-distance relations governing the electric and magnetic field strengths it should be obvious that by far the larger portion of the energy of the electromagnetic wave is carried by the fields in close proximity to the conductors. The conductors thus serve as excellent guides allowing very little energy to escape into free space.

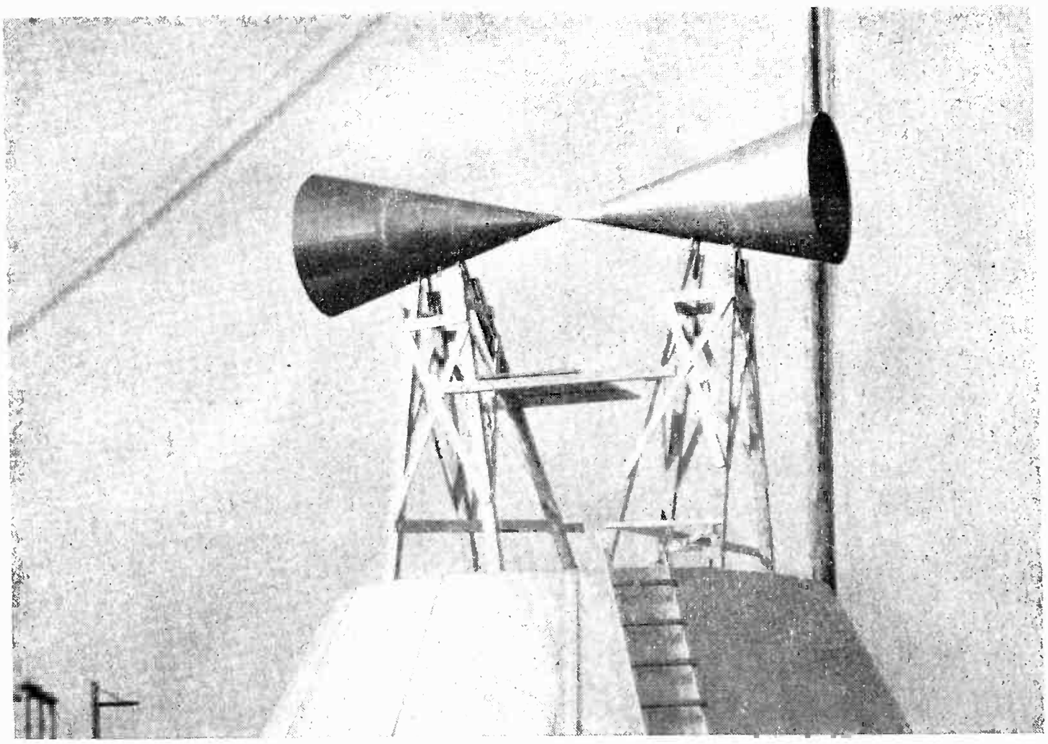

Fig. 22—Experimental double-cone antenna.

The purpose of a television antenna system is the exact opposite from that of a transmission line. The conductors forming the antenna must continue to influence the energy which has been guided to their vicinity by the transmission line but, instead of guiding it to a fixed location, they must help the wave to expand into free space without any sudden disturbances in its electromagnetic field configuration. Such sudden disturbances result in reflected-wave energy which, in general, makes itself known in terms of reactance at the junction of the transmission line and the antenna.

Let us consider the electromagnetic wave propagated between two conical conducting surfaces extending indefinitely and arranged in hour-glass relationship. This wave is the simplest possible type of spherical wave. Figure 26 shows the distribution of electric field at a particular instant of time. These lines simply slide along the conductors as time progresses. The number of lines per inch represent an attempt to show the variation of intensity with radial distance. The variation in thickness is a rough indication of the variation in strength with latitude when the common axis of the two cones is considered as the polar axis for all spherical wave fronts. The lines of magnetic force are latitude lines on any wave front and are not shown in the figure.

Since there are no radial components of electric or magnetic force, this wave is purely transverse and for this reason is much simpler than any type of free space wave. In a consistent system of units such

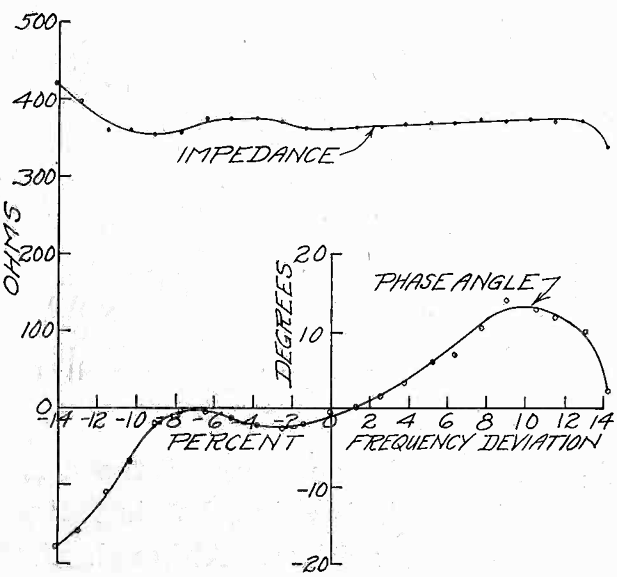

Fig. 23

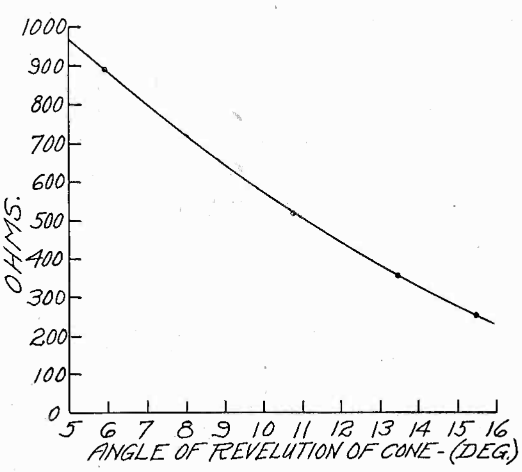

Fig. 24

as the Heaviside-Lorentz system, the electric and magnetic forces are everywhere equal in magnitude. These intensities are inversely proportional to the distance from the apex and inversely proportional to the sine of the co-latitude angle on any spherical wave front. The voltage from one conducting surface to the other in any wave front is a constant independent of distance of travel. The same is true of the total surface current along the conductors. If we define the wave impedance or characteristic impedance in the same manner as for an ordinary transmission line, i.e. as the ratio of voltage to current in a travelling wave, we see that the characteristic impedance of the expanding wave between this system of two cones is a constant, independent of distance. Its value is 120 $\log_{\varepsilon}$ ($\cot \alpha/2$) where α is the angle of revolution of the cones. This constancy of the wave impedance is a very important property of this type of wave.

The preceding statements summarize the results of a mathematical analysis of spherical wave propagation on conical surfaces. Those familiar with harmonic analysis may be interested to know that the conical conductors are responsible for a spherical wave derived from the Legendre function of the second kind. In most treatments of wave functions this solution is not discussed, probably because the corresponding type of wave cannot exist without conical surfaces and the investigators were concerned almost entirely with free space waves which must be derived from the Legendre function of the first kind.

We are now ready to consider the wave phenomena in connection with the actual arrangement of two finite cones. As a wave front expands while the wave travels outwardly from the apex, the energy

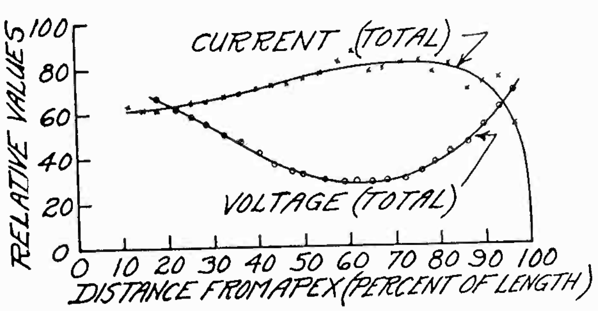

Fig. 25

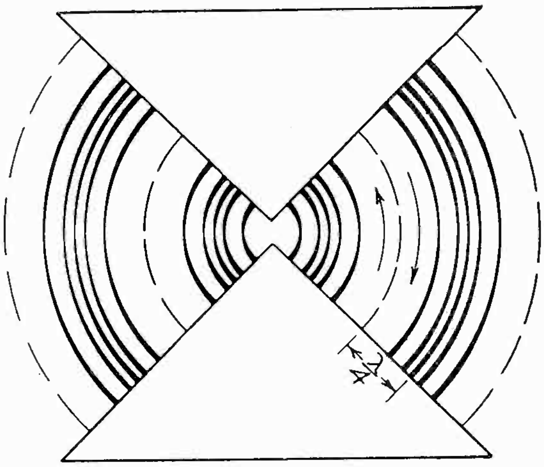

Fig. 26

does not remain heavily concentrated near the surfaces of the conductors as in the case of an ordinary transmission line, but is fairly evenly distributed throughout a wave-front surface. Although the energy immediately adjacent to the conducting surfaces must be completely reflected when the wave front reaches the outer edges of the cones the larger portion of the total energy in the wave front is free to continue to progress into free space. Speaking rather loosely, the wave has by this time expanded sufficiently "to gain a good hold on the surrounding ether." Since the characteristic impedance of this antenna is independent of distance from the apex, the only reflection which can take place on the antenna itself is that which occurs at the rims of the cones. The reflected-wave energy reaching the apices is quite small in comparison with that of the outgoing wave and for this reason the antenna presents a substantially pure resistance to the feeder line over a considerable range of frequency.

Summary and Conclusions

For reception at locations where the signal strength is high an ordinary half-wave dipole of large conductors fed from a convenient type of balanced transmission line should generally prove satisfactory if the transmission line is loaded, at the receiver, with a pure resistance equal to its characteristic impedance. For remote locations a better system is a folded dipole matched to an efficient transmission line.

For transmitting a horizontally polarized wave, a pair of folded dipoles arranged in turnstile fashion and fed in quarter-phase relation,

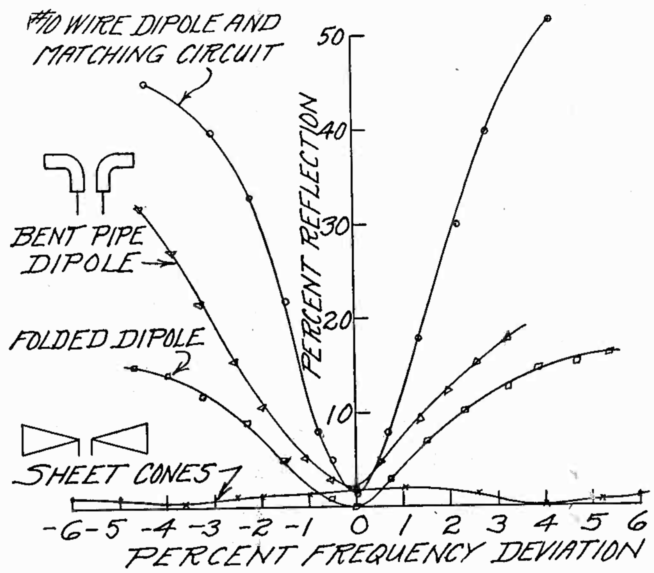

Fig. 27

would seem to satisfy most requirements at the present time. A double-cone antenna, together with an open-wire transmission line, satisfies the most rigid requirements with regard to band width but is not without serious mechanical disadvantages.

For comparison purposes curves of the coefficient of reflection versus frequency as measured on the transmission line feeding several of the various antennas which have been discussed are shown together in Figure 27.

THE ORTHICON, A TELEVISION PICK-UP TUBE

BY

ALBERT ROSE AND HARLEY IAMS

Research and Engineering Department, RCA Manufacturing Company, Inc., Harrison, N. J.

Summary—*Extensive laboratory and field tests have shown that the Iconoscope is capable of transmitting clear, sharp television pictures even under conditions of unfavorable illumination. An analysis of the operation of the tube suggests that improved efficiency and freedom from spurious signals should result from operating the mosaic at the potential of the thermionic cathode, rather than near anode voltage. The beam electrons then approach the target with low velocity, and the number of electrons which land is dependent upon the illumination.*

Several new designs were developed to make sure that the beam of low-velocity electrons was brought to the cathode-potential target in a well-focused condition, that the scanning pattern was undistorted, and that the focus of the beam was not materially altered by the scanning process. A magnetic field perpendicular to the target was found to be useful in focusing and guiding the beam. In some of the tubes, the scanning beam was released by a flying light spot moving over a photocathode. In other tubes it was found more convenient to develop the beam in an electron gun with a thermionic cathode. Special horizontal and vertical deflection systems capable of operating in the presence of a magnetic field were evolved.

The electron gun type of pick-up tube, which has been called an Orthicon, has a maximum signal current output over 300 times the noise in a typical amplifier. The signal is proportional to light intensity. The resolution is sufficient for the transmission of a 441-line picture. Spurious signals are neglible. Within the accuracy of measurement, all the photoemission is converted into video signals. In its present developmental form, the Orthicon gives promise of becoming a useful television pick-up tube.

INTRODUCTION

WITH the beginning of scheduled television broadcasting in New York, it is natural that a great deal of attention should be given to the commercial aspects of the art. Engineers realize, however, that research and development work must continue if future improvements are to be assured. Further investigations have, therefore, been carried on to provide ways to transmit clearer images, with less illumination. This paper will discuss an improved form of pick-up tube resulting from some of these investigations.[1]

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[1] See, also, Albert Rose and Harley Iams, "Television Pick-up Tubes Using Low-Velocity Electron-Beam Scanning," *Proc. I.R.E.*, Vol. 27, No. 9, pp. 547-555, September (1939).

At the present time, the RCA television system uses Iconoscopes to convert the optical image into a sequence of video signals for transmission to the receiver. Several previous publications have described these tubes and explained how they operate,[2,3,4] so that a brief review of their characteristics is sufficient to serve as a basis for a discussion of new pick-up tubes.

Extensive laboratory and field tests have shown that the Iconoscope is capable of transmitting clear, sharp television images, even under conditions of unfavorable illumination[5]. The spectral response is sufficiently like that of the human eye to give a natural appearance to the viewed scene. In the circuits associated with the tube, provision is made to "keystone" the deflection (so as to make the scanning beam move over the mosaic in a rectangular pattern) and to introduce shading signals into the amplifier (to compensate for the "dark spot" signal).

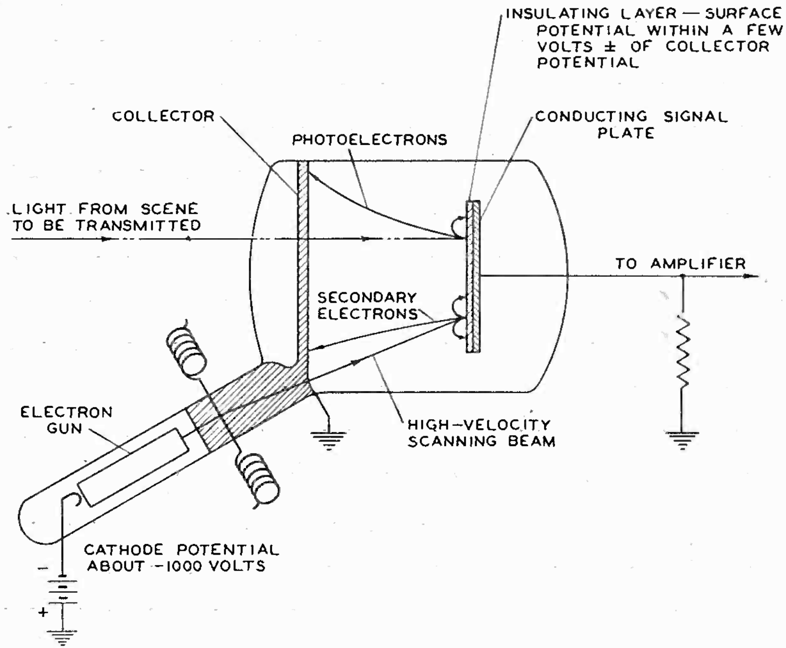

Fig. 1—Schematic diagram of an Iconoscope.

In the course of these tests, several significant discoveries were made. One of these was that the good operating sensitivity of the tube

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[2] V. K. Zworykin, "The Iconoscope—A Modern Version of the Electric Eye," *Proc. I.R.E.*, Vol. 22, No. 1, pp. 16-32, January (1934).

[3] V. K. Zworykin, "Iconoscopes and Kinescopes in Television," RCA Review, Vol. 1, No. 1, pp. 60-84, July (1936).

[4] V. K. Zworykin, G. A. Morton and L. E. Flory, "Theory and Performance of the Iconoscope," *Proc. I.R.E.*, Vol. 25, No. 8, pp. 1071-1092, August (1937).

[5] Harley Iams, R. B. James, and W. H. Hickok, "The Brightness of Outdoor Scenes and Its Relation to Television Transmission," *Proc. I.R.E.*, Vol. 25, No. 8, pp. 1034-1047, August (1937).

is obtained in spite of an operating efficiency only 5 or 10 per cent of that which is theoretically attainable[4]. In other words, during typical operation only about one-third of the photoelectrons which the mosaic emits are drawn away, and only about one-quarter of the stored charge is effective in producing the video signal. This lowered efficiency is connected with the release of secondary electrons from the mosaic by the scanning beam.

The situation is illustrated in Figure 1, which shows the essential parts of an Iconoscope. Light in the optical image focused on the mosaic causes the emission of photoelectrons, leaving a pattern of charges corresponding in intensity to the light and shade of the scene to be transmitted. This pattern of charges is scanned by a beam of electrons, which strike the mosaic at high velocity. On the average, each beam electron releases several secondary electrons. Since the mosaic is an insulated surface, the electron current leaving it must (on the average) be equal to the electron current arriving. Thus, when the tube is in darkness, only as many secondary electrons can escape from the mosaic as there are beam electrons which arrive. The rest of the secondary electrons fall back on the surface near or far from the point of emission. Nonuniformities in the escape and rain of secondary electrons cause the "dark spot" signal.

Many of the numerous secondary electrons are emitted with appreciable velocity, so that the condition of one secondary electron leaving for each beam electron arriving means that the electric field near the mosaic is such as to hinder the escape of secondary electrons. This field also reduces the escape of the photoelectrons, which have lower average emission velocity. When the mosaic is lighted, those photoelectrons which escape contribute a positive charge to the lighted parts of the surface. These charges are partly dissipated by the rain of secondary electrons, but sufficient charge is stored during a frame period to produce a strong signal in an amplifier connected to the signal plate when the beam releases the stored charge.

Low-Velocity Electron Beam Scanning

If one could ignore the immediate practical problems and choose an ideal mode of operation for a television pick-up tube, he might want to provide a field strong enough to draw away all of the photoelectrons which are emitted, and he might prefer to do the scanning without involving secondary emission in the process. These conditions can be met by operating the mosaic, in known fashion, at the potential of the cathode in the electron gun.

Cathode-voltage operation is possible, for the potential of an insulated surface exposed to an electron beam is stable at this voltage[6]. High-vacuum cathode-ray tubes are usually operated so that the beam electrons strike the screen with high velocity, and liberate many secondary electrons. The screen potential then adjusts itself (usually near anode voltage) so that the number of secondary electrons which escape is equal to the number of beam electrons which arrive. However, the other stable condition occurs when the surface is at cathode potential. The beam electrons then approach the target, but are repelled and retire without striking. If the target becomes slightly positive (by photoemission, for example), the beam electrons land without producing appreciable secondary emission and restore the original voltage.

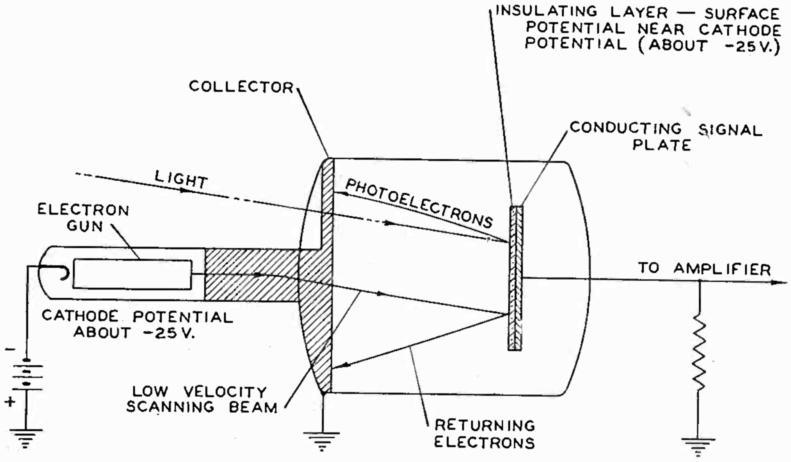


Fig. 2—Pick-up tube with cathode-potential target.

Figure 2 illustrates the operation of a television pick-up tube with its photosensitive target at cathode potential. In the absence of light the beam electrons approach the surface, are slowed to zero velocity, and then are drawn away. There is no signal in an amplifier connected to the signal plate. When light falls on the mosaic, all the photoelectrons are pulled away by the strong electrostatic field between the mosaic and the anode. The charges given the surface are not dissipated by a rain of electrons, but are stored until the scanning beam approaches. When the beam comes near a lighted area, and finds it a few volts positive, electrons land until their negative charge brings the surface to cathode potential again. The velocity with which the beam electrons reach the surface is so low that secondary emission is not involved to any appreciable extent. The signal is simply due to the impulses given the signal plate by the beam electrons, as they arrive at the lighted parts of the target.

[6] A. W. Hull, "The Dynatron," *Proc. I.R.E.*, Vol. 6, No. 1, p. 5, February (1918).

Design Problems

So far, much of the discussion has related to ideals. Some of the problems which must be solved before a pick-up tube can be made to operate in a satisfactory manner with its target at cathode potential are illustrated in Figure 3. When the electron beam is deflected, in usual fashion, at an angle to the axis of the tube, the electrons do not approach the target perpendicularly. The negative voltage needed to keep an electron from landing on the mosaic is only as great as the component of velocity perpendicular to the mosaic, and the electric field is not able to stop motion parallel to the surface. When conditions are as illustrated, the beam charges one part of the target to cathode potential, and other parts slightly more positively. Also, the point of contact of the beam may be elongated into a line. This situation may

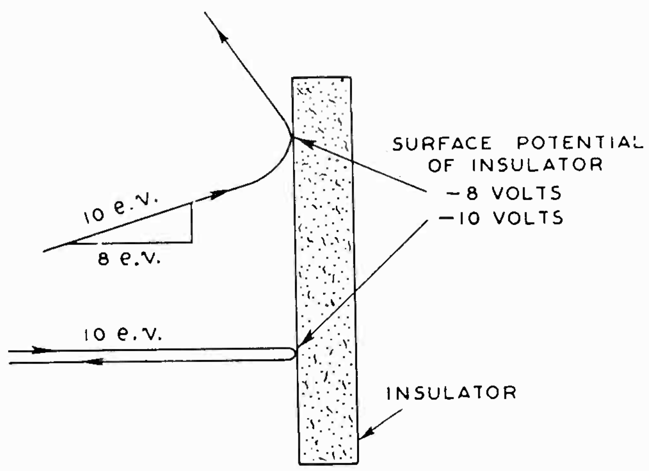


Fig. 3—Electron paths near target.

be expected to cause a loss of resolution at the edges of the picture, and unstable operation when the electrons move with considerable velocity tangent to the surface of the target. These considerations suggest that it would be preferable for the beam always to approach the mosaic nearly perpendicularly, or for the beam to be constrained.

Another important problem is that of providing sufficient beam current (about one microampere) in a beam which retains its small diameter when the electrons are slowed to almost zero velocity and are subject to strong local fields at the mosaic surface.

In the solution of these problems, a magnetic field perpendicular to the mosaic and extending to the source of the electron beam has been found useful. When the field is made sufficiently strong, the beam is focused and refocused many times between the cathode and the target. In a uniform magnetic field, the final size of the scanning spot is substantially the same as that of the source of the beam. The magnetic field can also be used to keep the beam electrons from proceeding very far across the surface of the mosaic at grazing incidence. Electrons which tend to move in this fashion cut across the lines of flux, and move in circular paths. The diameter of these circles can be made as small as is desired by increasing the strength of the magnetic field.

Several types of pick-up tubes based upon these principles have been designed. In some, the scanning beam is developed at a photocathode illuminated by a moving spot of light, while in others the beam originates at a thermionic cathode. The name Orthiconoscope (or Orthicon, for short) has been used to denote these tubes in which the target is operated at cathode potential. (The Greek prefix "orth",

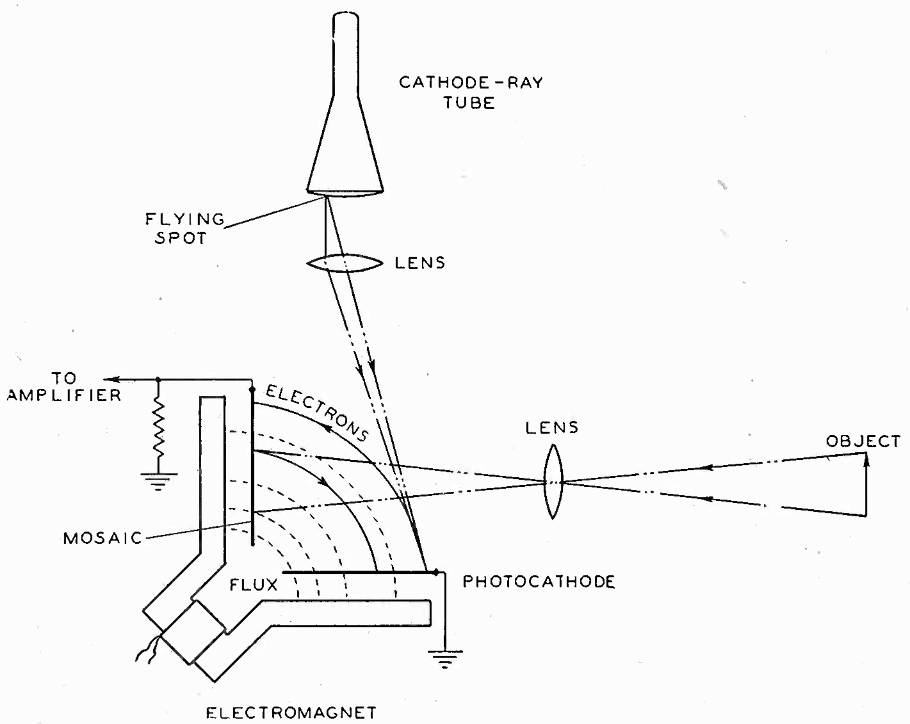


Fig. 4—Pick-up tube using photoelectrons for beam.

meaning straight, is added to the well known term "Iconoscope" to describe the linear relation between light and signal output, which has been observed.)

Tubes with Photoelectric Scanning Beam

One of the tubes which has been built and tested is illustrated schematically in Figure 4. A charge image of the scene to be transmitted is developed by focusing the optical image upon a conventional mosaic. The electron beam which scans the mosaic is produced by photoemission from a conducting photocathode, which is illuminated by a flying light spot focused from the face of a cathode-ray tube with a short-time-lag screen. At any instant, the light from a single spot on the cathode-ray-tube screen is focused on the photocathode. The emitted photoelectrons are guided by the curved lines of flux between the pole faces of an electromagnet, and the beam is also focused by this field. When beam electrons approach a lighted part of the mosaic, they are absorbed and produce signals in the amplifier. From the dark parts of the mosaic the beam electrons are reflected back to the photocathode. Photoemission from the mosaic is collected by the photocathode.

The picture transmitted by this tube is quite sharp, and is free from spurious shading. Streaking, which one might expect because of time lag in the luminescent material, is not apparent. Most of the discharging of charged areas takes place in the first fraction of a

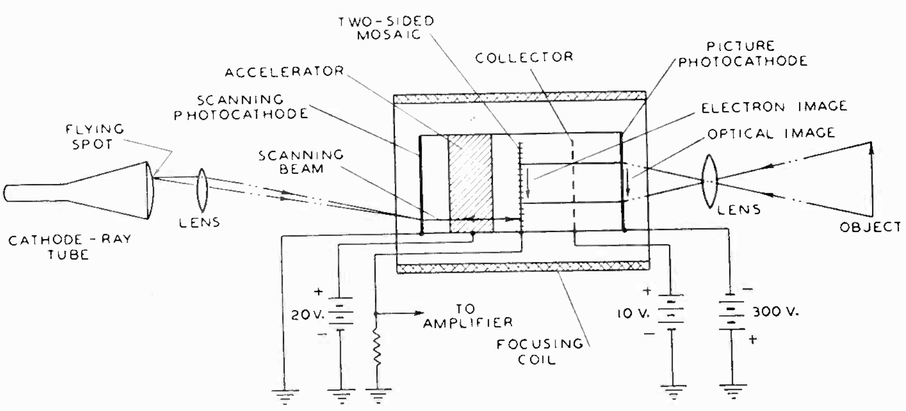


Fig. 5—Pick-up tube using secondary emission amplification.

microsecond, and after that the beam electrons fail to reach the target and their presence is not observed.

Tubes which incorporate secondary emission amplification of the image, in a fashion somewhat comparable with the method used in the Image Iconoscope[7], have also been built and tested. Such an arrangement is shown in Figure 5. In this case, the optical image is focused upon a translucent photocathode and the resulting photoemission is focused upon a two-sided mosaic by means of an axial magnetic field. Secondary electrons released from the mosaic by the high-velocity picture electrons are drawn away to a collecting electrode, giving image amplification. The electron beam which scans the other side of the mosaic is obtained from a flying light spot moving over another translucent photocathode, and is focused by the same axial magnetic field which focuses the electron image. As in the previous tube, the scanning beam restores the mosaic to the potential of the

[7] Harley Iams, G. A. Morton, and V. K. Zworykin, "The Image Iconoscope," *Proc. I.R.E.*, Vol. 27, No. 9, pp. 541-547, Sept. (1939).

scanning cathode. When the tube is operated, the gain in sensitivity due to secondary emission amplification is readily observed.

While a number of tubes which use photoelectric scanning beams have been tested, these two examples are sufficient to indicate the methods that have been used. These methods make possible the scanning of a large mosaic with a well-focused beam of low-velocity electrons, resulting in the transmission of video signals free from spurious signals. Further, an increase in sensitivity through secondary emission amplification may be obtained. However, because the auxiliary apparatus to generate the flying light spot represented a complication, an investigation was made of tubes in which the beam is derived from a thermionic cathode, and is deflected in the presence of a magnetic field.

Tube with a Thermionic Scanning Beam

The most important problem in the design of a pick-up tube, operating with its target at cathode potential, and using an electron gun to

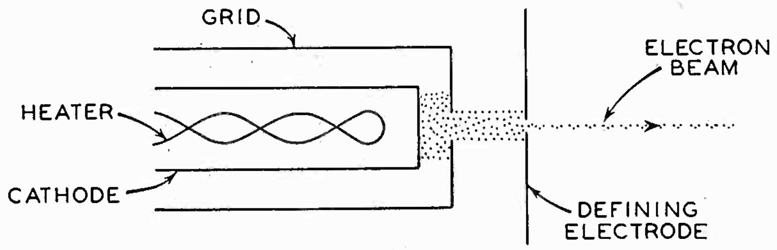


Fig. 6—Electron gun.

generate the scanning beam, is that of deflecting the beam without defocusing it. In the case of the tubes described above, each point on the target had a corresponding point on the photocathode such that the same magnetic line intersected both of them. Photoelectrons generated at one end of the line at the photocathode were focused at the target with substantially one-to-one magnification. A slight enlargement of the scanning spot occurred due to the emission velocities of the photoelectrons transverse to the magnetic field. If, in some way, larger transverse velocities were introduced into the motion of the beam electrons, they would describe helices of larger amplitude around the magnetic lines and result in a larger spot at the target. In the case of a tube using an electron gun, only the central point on the target is normally connected with the cathode by a magnetic line. If electrons from the gun are to reach other points on the target, either they must cross the magnetic lines of the axial field or they must be guided to other points by warping the axial magnetic field. Both of these devices are used in the form of Orthicon to be described. To insure, however, that the deflected spot is not larger than the undeflected spot it is necessary that no significant amount of velocity transverse to the magnetic field, imparted to the beam electrons by the deflection system, be retained by the electrons as they approach the target.

The simple electron gun shown in Figure 6 generates a stream of electrons moving parallel to the axis with a velocity of about a hundred volts. The cross section of the beam is limited to the size of a picture element by the defining aperture in the last electrode. In this state, the beam enters the deflection system and in this state it should emerge except for a displacement from the axis.

The high-speed horizontal deflection is accomplished by a pair of electrostatic deflection plates in combination with the axial magnetic

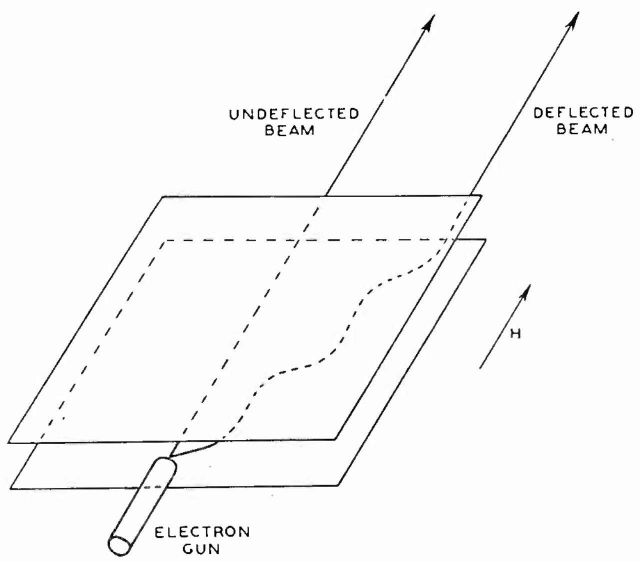


Fig. 7—Path of electron beam in electric and magnetic fields.

field. A somewhat schematic representation of the path of the beam through the plates is shown in Figure 7. The beam is seen to be deflected in a plane parallel with the plates and to diverge from the axis only while it is between the plates. After leaving the plates, the beam continues parallel to the axis. The amplitude of deflection is proportional to the electric field and the transit time of the electrons through the plates, and inversely proportional to the strength of the axial magnetic field. Since the maximum amplitude of deflection is limited to the width of plates, these must be as wide as the target to be scanned. The wiggles in the beam in Figure 7, when viewed from the end of the plates, that is along the axis, appear as a series of cycloids. This is the two dimensional path described by electrons moving in crossed electric and magnetic fields. If the electric field from the deflection plates could be sharply cut off at the entrance and exit edges, the transit time of the beam could be adjusted so that only an integral number of cycloids would be performed by the beam in passing through the plates. In this way, none of the transverse velocity represented by the cycloidal motion would be retained by the beam after it left the plates. While this is a possible arrangement, it has been found that a less critical way of insuring that the emergent beam retains substantially none of the transverse velocity acquired in the plates is to suppress the amplitude of cycloidal motion within the plates. The amplitude of cycloidal motion may be considerably reduced by admitting the beam to the plates through a gradually increasing electric field and similarly letting it leave through a gradually decreasing field. For this reason, the deflection plates are flared out at the entrance and exit ends.

Two significant distinctions are to be noted, in the above account, between electrostatic deflection in the presence of a magnetic field, as

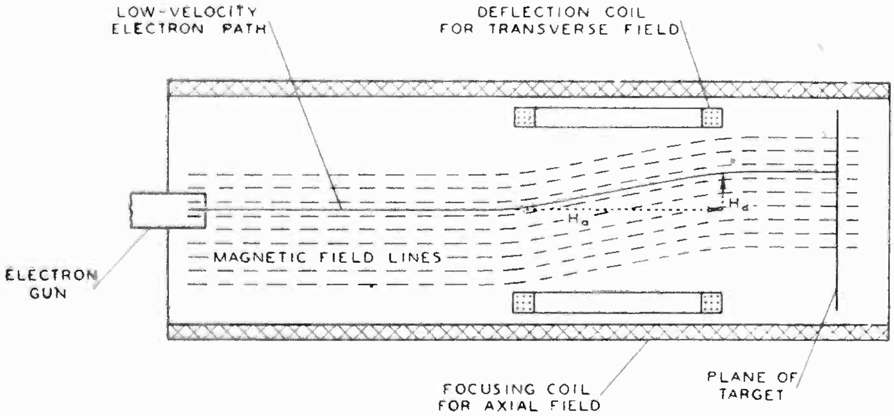


Fig. 8—Path of electron beam in warped magnetic field.

in an Orthicon, and electrostatic deflection in a magnetic field free space, as in the usual cathode-ray tube. First, the plane of deflection, which is perpendicular to the plates in the usual electrostatic deflection, has been rotated through ninety degrees into a plane parallel with the plates, in an Orthicon. Second, while the usual plates impart a transverse velocity to the beam which causes the beam to continue to diverge from the axis after leaving the plates, the plates in an Orthicon cause the beam to diverge from the axis only while the beam is between the plates. The axial magnetic field constrains the beam to motion parallel with the axis after it leaves the plates.

The low-speed vertical deflection is accomplished by a pair of magnetic coils. Here, again, while magnetic coils are used in the usual cathode-ray tube without an axial magnetic field, their action in an Orthicon is essentially different by virtue of this field. Briefly, the axial magnetic field rotates the plane of deflection through ninety degrees and causes the electrons in the beam to move parallel to the axis after leaving the deflection coils. The average path of the beam through a pair of deflection coils immersed in an axial magnetic field is shown in Figure 8. The amplitude of deflection is proportional to the magnitude and axial length of the deflection field and inversely proportional to the magnitude of the axial magnetic field. From Figure 8, it is evident that the separation of the coils must be as large as the height of the target to be scanned. While Figure 8 shows the average path of the beam to follow the magnetic lines closely, the actual path contains a helical motion imparted to the electrons due to their passing through a curved magnetic field. For small deflections, the amplitude of this helical motion is of the order of, or less than, the helical motion of the electrons due to their emission velocities. The magnetic coils may therefore be considered as a means of displacing the beam from

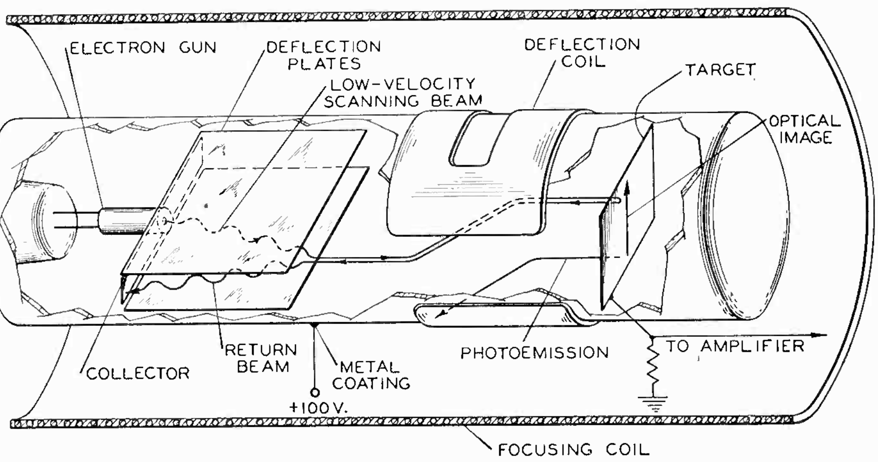


Fig. 9—Schematic diagram of an Orthicon.

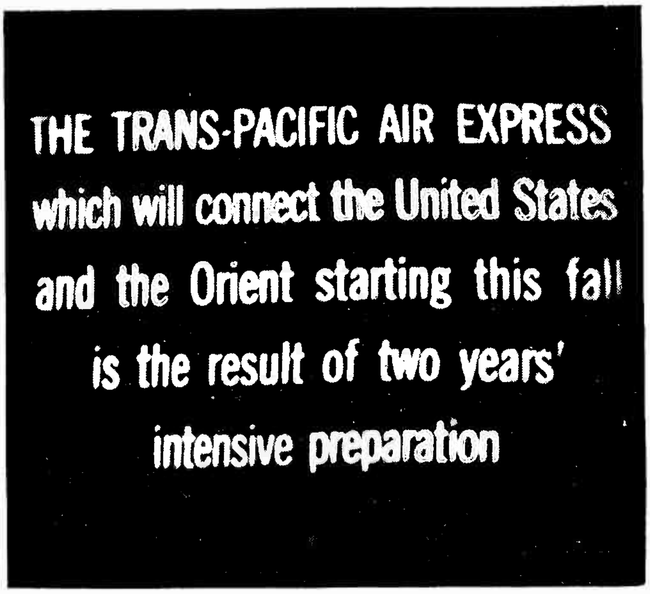
the axis, without contributing any significant transverse velocity to the beam electrons.

The beam, as it leaves the deflection system and approaches the target, is in substantially the same condition as when it left the gun, except that it may be displaced from the axis. Since the target surface is at cathode potential, the beam passes through an electric retarding field sufficient to slow it to near zero velocity at the target. The target is a thin mica sheet, the side facing the beam being covered with a mosaic of photosensitive elements and the side away from the beam being coated with a translucent conducting film of metal known as the signal plate. The optical picture is focused on the photosensitive side through the translucent signal plate. The photoelectrons are drawn away from the zero-potential surface to various positive electrodes near the target (see Figure 9). As described earlier, if the element approached by the beam has been unlighted, the beam electrons will not land on the target, but will be brought to rest near it and be accelerated back away from the target. If the element has been lighted, enough of the beam electrons will land to replace the photoelectrons that have been drawn away during the previous frame time. In this way, the beam maintains the target at cathode potential and generates the video signal. The electrons which do not land at the target retrace substantially their original path as they return toward the electrostatic plates. In passing back through the plates, the beam diverges from its going path in the direction of the original deflection, as shown in Figure 9. The beam eventually strikes an elongated collector electrode, also shown in Figure 9.

Operating Characteristics

The description of the operation of an Orthicon would lead one to expect a set of simple operating characteristics. This expectation has been borne out by numerous observations on the tubes. Briefly, substantially no signal is transmitted with no light on the target. With an optical picture focused on the target, a signal proportional to the light intensity at each point is transmitted. The maximum signal is limited by the amount of beam current. In some tubes a modulated beam current of one microampere has been observed. This corresponds to a signal current about three hundred times the noise level of a typical television amplifier.

Not only is the transmitted signal proportional to the light on the target, but also the conversion of possible photoemission into signal takes place at substantially 100 per cent efficiency. This requires that the photoemission from the target be saturated throughout the frame time, that the charge be fully stored for that time, and that all of the stored charge be useful in producing a video signal when the scanning beam passes over it. Tests made on Orthicons have shown that the photoemission from the target is saturated when the collector electrodes surrounding the target are more than twenty volts positive with respect to the target. Since these electrodes are usually about plus one hundred volts, a saturated photocurrent is assured under static conditions. During a frame time, the photoemission from the target is swept over the collecting electrodes by the field from the vertical deflecting coils. To test whether the photocurrent was saturated equally throughout the frame time and equally stored, a spot of light was projected on the target once a frame time for about one-tenth of the frame time. The signals, both of the photocurrent and of the discharge process, were observed on an oscilloscope. The storage time was varied from zero to a full frame time by varying the time at which the spot was projected on the target. No variation in either of these signals was observed throughout the full range of storage time, indicating that the photocurrent was equally saturated and the charge equally stored throughout the frame time. A final overall test, which compared the transmitted signal with a known amount of light on a target of known photosensitivity, showed that (within ability to measure) all of the stored charge (or by the above tests, all of the photoemission at the target) was utilized in producing video signal. As a result of this high operating efficiency, Orthicons with a target photosensitivity of one microampere per lumen exhibited approxi-



(Courtesy R. D. Knell)

Fig. 10—Television picture transmitted by an Orthicon.

mately the same operating sensitivity as Iconoscopes with a target photosensitivity of ten microamperes per lumen.

Observations on the resolution of the transmitted picture showed more than four-hundred-line resolution over the entire picture and as high as six-hundred-line resolution in the center. For the particular size of target used, two and one-half inches wide, this means that the scanning beam near zero velocity can resolve elements on the target less than one two-hundredth of an inch apart.

Two representative pictures transmitted by an Orthicon are shown in Figures 10 and 11. No shading-compensation signals were introduced into the television system from which these pictures were taken.

Conclusions

As a result of the tests which have been described, it may be concluded that the operation of a television pick-up tube with its target at cathode potential makes possible (1) the efficient conversion of photoemission into video signals, (2) a large signal output, and (3)

(Courtesy R. D. Knell)

Fig. 11—Television picture transmitted by an Orthicon.

the elimination of spurious signals. While developmental Orthicons incorporating these features have been built and operated, additional work to determine optimum designs is in progress.

POLYSTYRENE APPLIED TO RADIO APPARATUS

BY

R. L. HARVEY AND W. L. CARLSON

RCA Manufacturing Company, Inc., Camden, N. J.

Summary—*Polystyrene is a new thermoplastic material of extremely low-loss character. Its moisture absorption is practically nil. The material being especially adapted to high-speed injection molding makes its use for production of radio parts very practical.*

A typical design of a temperature-compensated, high-frequency inductance is discussed, where the combination of polystyrene coil form and magnetic core is used.

A mixture of polystyrene polymer and monomer, and its use as an impregnating compound for coils and porous materials, are also discussed.

POLYSTYRENE in its pure state is colorless and, when molded, has a hard surface which will retain a polish. The accompanying table shows the general mechanical and electrical properties of polystyrene as compared to the best available grades of two other types of insulating materials which are used extensively in radio apparatus.

	Polystyrene	XXX Type Laminated Phenolic Sheet	Low Loss Steatite Ceramic
Tensile strength—lb./sq. in.	5000-5500	7000	8000
Flexual strength—lb./sq. in.	7000-7500	15000[1]	22000
Compression strength—lb./sq. in.	10000	35000[1]	65000
Specific gravity	1.05-1.07	1.3	2.6
Maximum operating temperature	75°-80° C	125° C	1000° C
Water absorption	0.05%	1.0%	0.02-.05%
Coefficient linear expansion	$70\times10^{-6}/°$ C	$20\times10^{-6}/°$ C	$6.3\times10^{-6}/°$ C
Power factor 1000 kc	0.0002	0.035	0.0006
Dielectric constant 1000 kc	2.6	5	6
[2]Dielectric strength volt/mil	500-525 ($\frac{1}{8}''$ thick)	450 ($\frac{1}{16}''$ thick)	200 ($\frac{1}{4}''$ thick)

[1] Perpendicular to laminations.

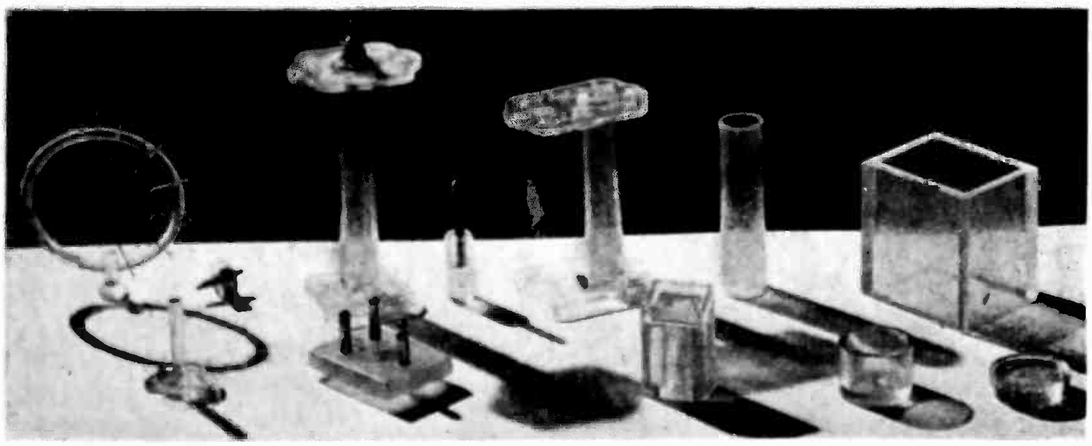
[2] Step-by-step method.

The material is especially well adapted to high-speed injection molding. Indications are that molded polystyrene will hold dimensions better than any other generally used thermoplastic material. There is no shrinkage due to evaporation of plasticizer and very little tendency to cold-flow up to temperatures of 70° C.

Polystyrene is commercially feasible for radio-frequency coil forms, terminal boards with inserts, tube sockets, insulation supports, condenser dielectrics, and other uses in radio receivers and antenna systems where lower losses are required than can be obtained with phenolic and ceramic materials. A miscellaneous assortment of polystyrene molded parts is shown in the accompanying photograph.

Molded polystyrene is limited to apparatus operating at ambient temperatures not exceeding 70° to 80° C. at which point heat distortion starts. This is a limitation for the application of polystyrene to transmitter apparatus. At about 100° C. the material softens perceptibly.

As a specific example of the application of polystyrene to high-frequency circuits, a precision inductance design will be described.



A miscellaneous assortment of polystyrene molded parts.

Referring to Figure 1, it will be noted that the coil is wound on the polystyrene form and is adjustable by the powdered-magnetic core held in position by the brass-screw assembly. By utilizing the difference of linear expansion of the polystyrene and brass stud, the curves A and B of temperature characteristic vs. core position, were obtained, the difference between the curves being the value of X (see diagram).

The interesting feature of this set of curves is that—if the core is worked between the values of 4/16 to 6/16, the temperature characteristic is substantially constant, but the inductance values change sufficiently to allow the core to be used as a trimmer.

In the design of a circuit, one should determine the combined temperature characteristic of the tube, capacitor, switch, wiring, etc., of the circuit (this will usually be a positive value), and select the proper value of X to compensate exactly for this change.

By changing the form factor of the coil, the material of the core stud or the thickness of the coil form, a further variation in the temperature characteristic will be had.

Figure 2 shows how the wall thickness of the coil form affects the temperature characteristic. For this curve, the core was placed at the dip (5/16 position) as shown in Figure 1.

In the design of r-f and i-f transformers as well as oscillator circuits, this type of compensation has been used to produce circuits which are very stable even with severe changes of temperature.

Polystyrene is being commercially employed as a substitute for paper in rolled fixed condensers. For this purpose, it is available in

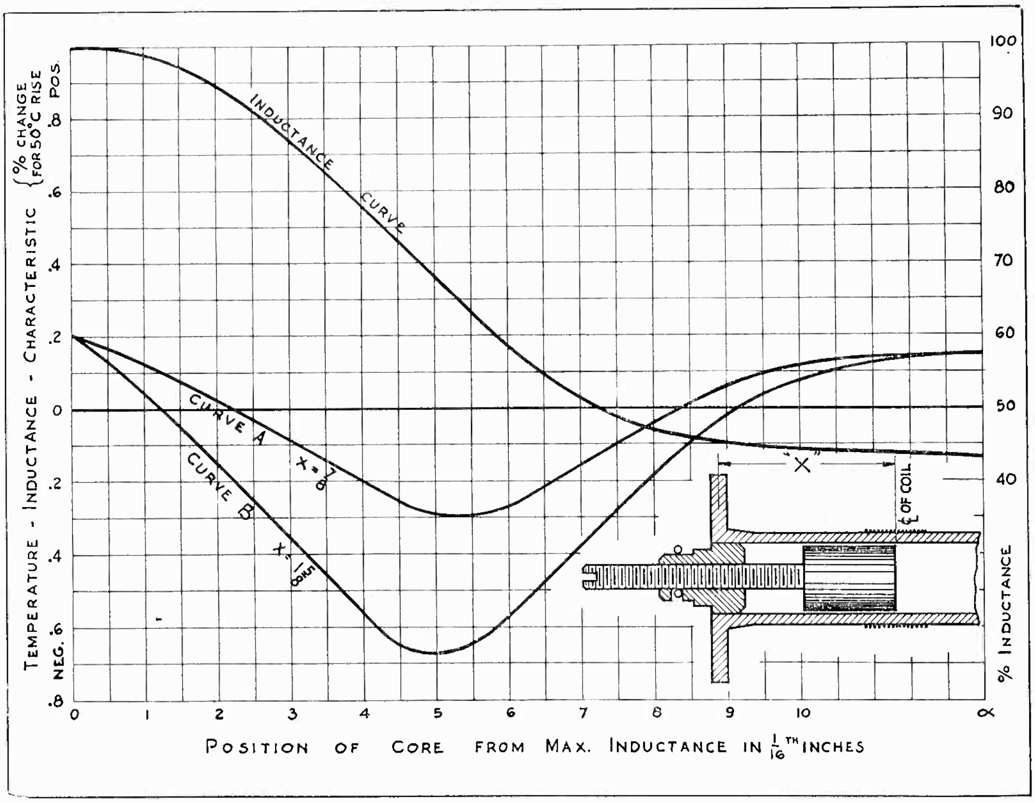


Fig. 1—Diagram of temperature-inductance characteristic in relation to position of core.

sheets down to 1 mil in thickness. It offers attractive possibilities in other forms for other types of condensers.

Polystyrene is styrene, a liquid solidified by the application of heat or light. The arrangement of the styrene molecules is changed during this hardening process called polymerization. For some applications it is preferable to start with styrene in the liquid state and delay complete polymerization until after the article has been treated and molded into its final form. For example, styrene can be used to impregnate multi-layer coils and porous materials. When used in this form, it has the decided advantage over the usual solvent-type solution in that it can be changed to the solid form without evaporation and loss of volume which means that the coil form or porous material is completely filled with polystyrene.

To reduce the brittleness of polystyrene, for some applications it is desirable to compound polystyrene with a low-loss plasticizer and adhesive material. In general this tends to increase the power factor and decrease the flow temperature.

There are a variety of common solvents for polystyrene, the most powerful of which is its own monomer. By combining the polymer

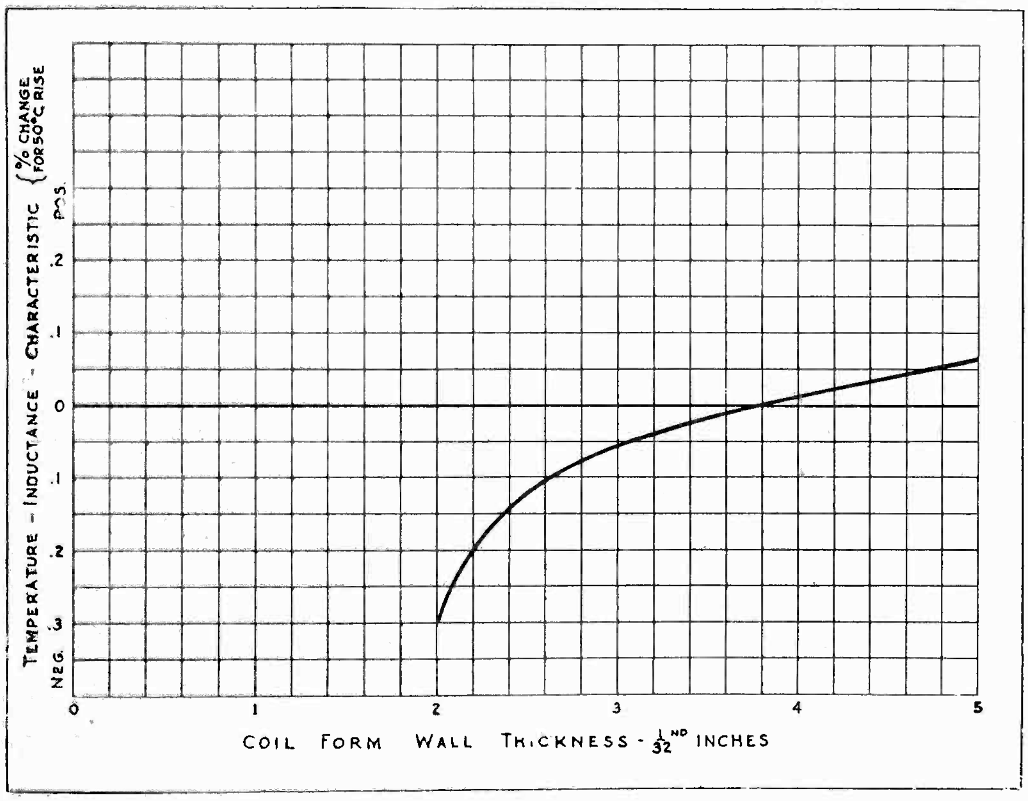


Fig. 2—Temperature-inductance characteristic in relation to coil-form wall thickness.

and monomer, a solution of any viscosity may be obtained, and as pointed out above, can be converted to the solid form by heat or ultra-violet light.

Coloring materials can successfully be added to polystyrene for coding or decorative purposes, but the addition of such materials usually impairs the electrical characteristics.

Filling materials, such as silicon, mica, glass, quartz, etc., add to the mechanical strength, and increase the flow temperature, but such compounds are usually affected by humidity unless completely sealed by styrol.

TELEVISION SIGNAL-FREQUENCY CIRCUIT CONSIDERATIONS

By

Garrard Mountjoy

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Summary—*The relative performances of pre-selector circuits are compared as to gain and selectivity. Antenna circuits having one and two tuned meshes are quite similar in gain characteristics. The double-tuned type has the advantage of much greater image and i-f signal selectivity. All varieties discussed have gain dependent upon the factor* $\frac{1}{\sqrt{\Delta\omega C r_a}} \times$ constant. *It is thus seen that antenna gain is primarily dependent on the square root of three factors, namely, band width, internal resistance of the antenna, and capacitance of the last tuned circuit. The magnitudes of the reactances of the first tuned circuit are immaterial in the case of two-tuned circuit systems. Antenna gain is further influenced, particularly in the case of a single-tuned circuit, by the amount of attenuation at the cut-off frequency. The constant by which the basic gain factors must be multiplied to obtain the gain for a given circuit arrangement varies with the circuit arrangement. The gain characteristics of an r-f stage differ from those of the passive networks and depend upon* $\frac{1}{\Delta\omega}$, *thus showing the greater effect of band width in determining r-f stage gain. A double tuned r-f stage is materially better than a single tuned stage for both gain and selectivity.*

THE signal-frequency circuits of television receivers involve considerations which differ materially from those for the standard sound broadcast band receivers. This article discusses various types of signal-frequency circuits and their characteristics.

Each individual television channel is 6 Mc wide and includes both the sound and picture carriers with their sidebands, the picture signal being single sideband and the sound being double sideband. It is believed that the majority of receivers will have a single oscillator and single converter so that the signal-frequency circuits must pass both sound and picture. This means that these circuits must have a pass band of some 4.5 Mc, the exact band width required depending upon the desired receiver pass band. Where illustrative calculations are used in this article a band width of 4.5 Mc has been assumed, unless otherwise noted. In considering the pass band, the prime consideration is the picture-frequency requirement, as it is not necessary that the signal-frequency circuits have flat response characteristic to include the sound channel. Some attenuation at the sound frequency is permissible provided it does not produce too low a ratio of sound signal to hiss for the service intended.

There are two bands which contain frequencies assigned to television, 44 to 108 Mc, and a television relay band 150 to 300 Mc. It is the former band with which this article deals. In that band there are seven television channels as follows:

1.	44-50 Mc	5.	84-90 Mc
2.	50-56	6.	96-102
3.	66-72	7.	102-108
4.	78-84		

The basic antenna for television is a horizontal doublet one-half wavelength long, connected to the receiver by means of a transmission line. Such an antenna system has been assumed in this paper.

At the frequency for which the antenna is one-half wavelength long, it has a resistance of approximately 75 ohms. If the transmission line is terminated at the receiver end in its characteristic impedance, it acts electrically as though that terminating resistance were located at the antenna and no undesirable reflections exist in the antenna system. Calculations herein will be based on the assumption of an antenna of constant resistance at any considered frequency, and a transmission line of zero loss with termination equal to the antenna resistance. The latter equality is made to permit maximum pre-selector gain for cases where the terminating resistance is developed by reflections of the pre-selector damping resistance. In any applied case, the transmission line will produce attenuation, the amount of which will depend upon the specific line used. This factor may be taken into account in determining the gain possibilities at various frequencies. Also it may be common practice to use a transmission line of slightly higher characteristic impedance than 75 ohms to increase the damping of the resonant antenna. This will only modify slightly the gain formulas derived in succeeding paragraphs. In practice a doublet will exhibit substantially 75 ohms for a considerable frequency range, and some antennas (such as a double doublet) will exhibit constant driving impedance over a very wide frequency spectrum.

Pre-selector design involves consideration of gain, band width, transmission shape, receiver noise level, image ratio, i-f response ratio, and simplicity. The latter consideration limits this discussion to the following types:

(1) one tuned circuit
(2) two coupled tuned circuits
(3) an r-f stage of one tuned circuit preceded by either (1) or (2)
(4) an r-f stage of two tuned circuits preceded by either (1) or (2).

In the succeeding paragraphs relative systems are discussed from an idealized circuit basis. Thus inductors, capacitors, and damping resistors are assumed lumped, and no account is taken of the effects of leads, mal-distribution of capacitance, and other factors encountered in application. Therefore, the presentation is intended to show the possibilities of the several circuits rather than the results of individual applications and chassis layouts. Precautions to restrict lead lengths and generally refined technique in applying these circuits will produce results quite comparable to the ideal cases.

Damping resistors consist of all factors introducing resistance components into the tuned circuits. Chief among these is input conductance of the succeeding tube. This effect may be minimized and made independent of bias variations by use of resistive impedance in the cathode circuit. Resistance of circuit components is usually low compared to the required amount of damping resistance. Advantage may be taken of this in the construction of coils. Since high Q is not required, attention may be given to simplicity of construction and restriction of lead lengths.

One Tuned Circuit

(a) With Secondary Damping

Figure 1(a) illustrates a pre-selector of one tuned circuit L, C, and damping resistor R coupled to an antenna and transmission line through a primary L_p. The latter is center tapped to ground to limit any capacity coupling between L_p and L, thus limiting any pickup on the lead-in itself. The electrical equivalent is shown in Figure 1(b) where:

e_a = induced voltage = 1
r_a = antenna internal resistance
L_p = a negligibly small primary inductance tightly coupled to L to produce M

$\frac{\omega_o}{2\pi}$ = center frequency of the pass band

$\omega_o^2 LC = 1$

$\frac{\omega}{2\pi}$ = any frequency under consideration

$k = \frac{\omega}{\omega_o}$

For proper termination of the transmission line,

$$\frac{\omega_o^2 M^2}{R} = r_a$$

From the derivation shown in the appendix, gain at center frequency is

$$g_o = \frac{\sqrt[4]{\dfrac{1}{P^2} - 1 - \dfrac{\Delta\omega}{\omega_o}}}{\sqrt{2\Delta\omega C r_a}} \tag{1}$$

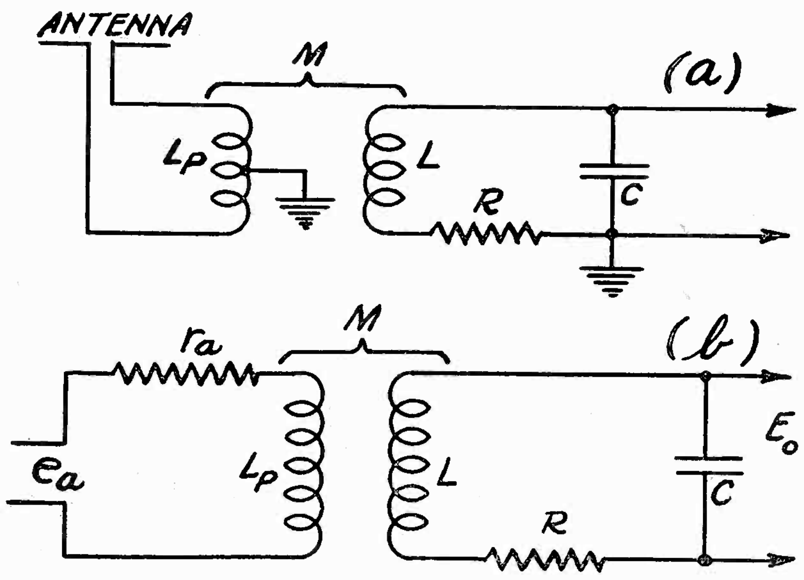

Fig. 1

and in an illustrative case

where $\Delta\omega = 4.5$ Mc $\qquad \omega_o = 45$ Mc $\qquad P = 0.90$

$$g_o = \frac{0.42}{\sqrt{\Delta\omega C r_a}} \tag{2}$$

for values of $C = 18$ $\mu\mu$f, $r_a = 75$ ohms, and $\Delta\omega = 2\pi \times 4.5 \times 10^6$

$$g_o = 2.17$$

Response ratio $\dfrac{g_o}{g}$ for any frequency remote from ω_o is expressed as

$$\frac{g_o}{g} = \frac{1 + k^2}{2} + j\frac{\varphi_o}{\Delta\omega}\sqrt{\frac{1}{P^2} - 1 - \frac{\Delta\omega}{\omega_o}}\left(k - \frac{1}{k}\right) \tag{3}$$

If, for example, an image frequency is considered where:

$$k = 1.5$$

$$\frac{\omega_o}{\Delta\omega} = 10$$
$$P = 0.9$$

then
$$\frac{g_o}{g} = 3.28$$

(b) Unterminated Line

If the termination of the transmission line were not a consideration, the circuit of Figure 1(a) and (b) may be redrawn with R omitted, i.e., no secondary damping factors are assumed to exist. Under these conditions gain at center frequency is

$$g_o = \frac{\sqrt[4]{\frac{1}{P^2} - 1 - \frac{2\Delta\omega}{\omega_o}}}{\sqrt{\Delta\omega C r_a}} \tag{4}$$

In the illustrative case where $\frac{\Delta\omega}{\omega_o} = 0.1$, $P = 0.9$,

$$g_o = \frac{0.43}{\sqrt{\Delta\omega C r_a}} \tag{5}$$

Equation (4) will show more gain than (1) only when P is small or $2\frac{\Delta\omega}{\omega_o}$ is small. At center frequency the impedance across L_p would (theoretically, not practically) be an open circuit and, therefore, would provide no termination for a transmission line.

(c) With Primary Terminating Resistance

Putting a terminating resistor (r_a in magnitude) across L_p will produce the equivalent circuit of Figure 3. A solution similar to the above will show

$$g_o = \frac{\sqrt[4]{\frac{1}{P^2} - 1 - \frac{2\Delta\omega}{\omega_o}}}{\sqrt{2\Delta\omega C r_a}} \tag{6}$$

Equation (6) is seen to have the bad features of both (1) and (4) to wit, the numerator has the factor $2\frac{\Delta\omega}{\omega_o}$ of (4), and the denominator has the factor $\sqrt{2}$ of (1). It is therefore illogical to provide termination in this manner. For large values of P the circuit of Figure 1 is the best of the three described systems.

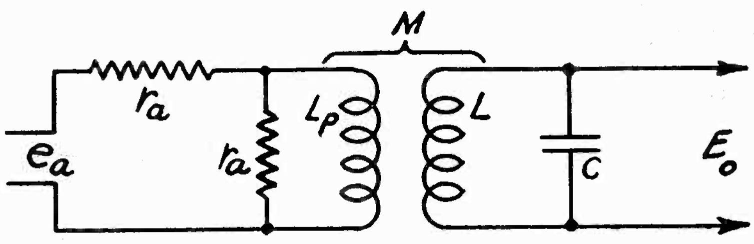

Fig. 3

Two Tuned Coupled Circuits

(a) Line Termination by Resistance Induced in the Primary

Figure 4(a) and (b) shows two coupled circuits and their electrical equivalent. The damping resistor R is here considered in the second tuned circuit, and the first circuit comprising L_1, C_1 is assumed to have negligible resistance compared to the second (a valid assumption). All symbols used have similar significance to those employed in preceding paragraphs

L_1C_1 and L_2C_2 are tuned to the center frequency, i.e.,

$$\omega_o^2 L_1 C_1 = 1$$
$$\omega_o^2 L_2 C_2 = 1$$

M_2 is chosen to produce two cut-off frequencies spaced $\Delta\omega$ apart.

$$\omega_o M_2 = \Delta\omega\sqrt{L_1 L_2}$$

R and M_1 are then chosen to produce a gain critical condition resulting in substantially flat transmission shape over the spectrum $\omega_o - \frac{\Delta\omega}{2}$ to $\omega_o + \frac{\Delta\omega}{2}$, and simultaneously producing the proper terminating impedance across L_p.

The primary tuned impedance looking into L_p is then

$$Z_p = r_a + j\omega_o L_p$$

$$Z_p = r_a + j\frac{1}{K_C^2} \times \frac{\Delta\omega}{\omega_o} r_a \tag{9}$$

(R reflects the equivalent r_a in the primary) in a case where $\frac{\Delta\omega}{\omega_o} = 0.1$; $K_C = 7$

$$Z_p = r_a + j\,0.2 r_a$$
$$Z_p = 1.02\ r_a, \text{ or nearly } r_a$$

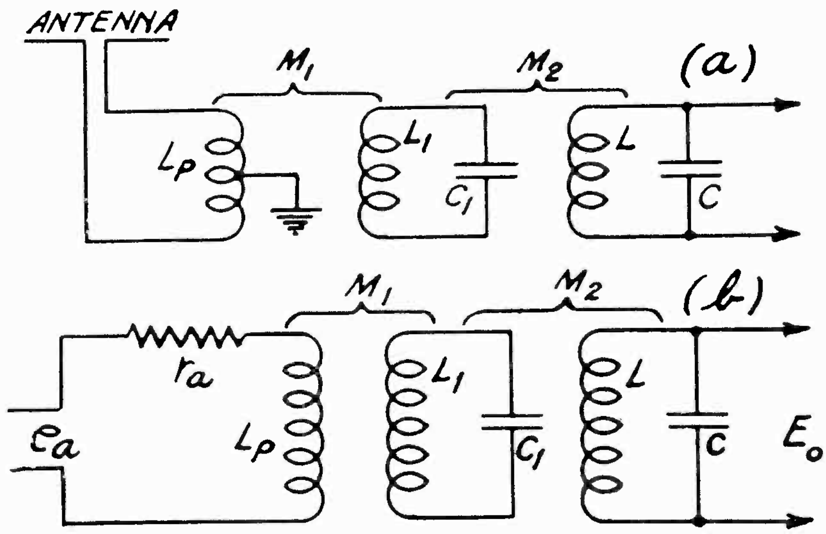

Fig. 5

(b) No Terminating Resistance

Figure 5 illustrates a system identical to Figure 4 except for the absence of R in the second tuned circuit.

In this case the impedance looking into the primary is equal to $j\omega_o L_p$ or is practically a short circuit at the resonant frequency, but has interesting gain and selectivity characteristics.

As in Figure 4, let

$$\omega_o M_2 = \Delta\omega \sqrt{LL_1}$$
$$\omega_o M_1 = \sqrt{r_a \Delta\omega L_1}$$

As derived in the appendix, gain is

$$g = \frac{1}{\sqrt{\Delta\omega C r_a}} \times \frac{1}{k - \left(\frac{\omega_o}{\Delta\omega}\right)^2 \left(k - \frac{2}{k} + \frac{1}{k^3}\right) + j\frac{\omega_o}{\Delta\omega}\left(k^2 - 1\right)} \tag{10}$$

Gain at center frequency ($k = 1$) is

$$g_o = \frac{1}{\sqrt{\Delta\omega C r_a}} \tag{11}$$

This is twice the gain obtained with the circuit of Figure 4 which differs from this only by the secondary damping.

Response at frequencies far removed from center will be largely governed by the factor in Equation (10) of:

$$\left(\frac{\omega_o}{\Delta\omega}\right)^2 \left(k - \frac{2}{k} + \frac{1}{k^3}\right)$$

which is like that found in Equation (7). Hence, selectivity may be considered twice as good as the case of (7) since center frequency response is better by that ratio.

Response at cut-off where $k = \dfrac{\omega_1}{\omega_o} = 1 + \dfrac{\Delta\omega}{2\omega_o}$ may be calculated after assuming $\dfrac{\Delta\omega}{\omega_o}$ to have a significant value.

Let $\dfrac{\Delta\omega}{\omega_o} = 0.1$ (a probable maximum)

Then, $g_1 = 0.975\, g_o$

Such a system, particularly if it were used with a fairly long, low attenuation, transmission line would probably produce intolerable "multi-path" signals due to standing waves (reflections) on the transmission line. Its chief interest is its high theoretical gain.

(c) Termination of the Transmission Line by a Resistor in the Primary Circuit

The circuit of Figure 5 may be modified to include a terminating resistance in series with the low inductance primary. Figure 6 shows this system where $r_1 + r_2$ are made equal to r_a. Dual resistors are used to preserve a line balance.

As shown in the appendix, gain at center frequency is

$$g_o = \frac{1}{\sqrt{2\Delta\omega C_2 r_a}} \tag{12}$$

Gain is, therefore, $\sqrt{2}$ times larger than in Equation (8), and $\dfrac{1}{\sqrt{2}}$ times as large as in Equation (11).

Of course, this assumes that the tuned circuit I and II are resistanceless, which in consideration of the input resistance of the first tube is inexact. Gain with this system will be between the limits of

$$\frac{.5}{\sqrt{\Delta\omega C_2 r_a}} \text{ to } \frac{.707}{\sqrt{\Delta\omega C_2 r_a}}$$

Image and i-f response ratio will be improved almost directly as gain is improved over Equation (8).

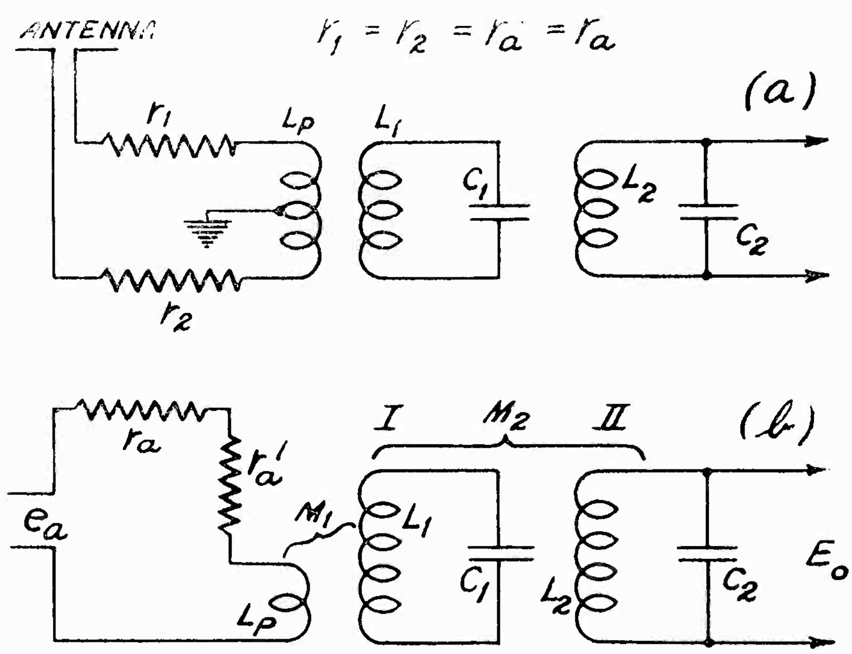

Fig. 6

(d) Tuned Primary, Tuned Secondary

In Figure 7 a system is shown in which a balanced primary is tuned to the center frequency. The reflected resistance of R provides the termination of the transmission line. Gain may be determined after making the following assumptions:

$$\omega_o{}^2 L_1 C_1 = \omega_o{}^2 LC = 1$$

$$\omega_o M = \Delta\omega\sqrt{L_1 L}$$

$$\frac{M}{L_1} = \sqrt{\frac{R}{r_a}}$$

$$R = \Delta\omega L$$

Gain at center frequency may be calculated to be

$$g_o = \frac{1}{2\sqrt{\Delta\omega C r_a}} \tag{13}$$

This is seen to be identical to the gain of the system of Figure 4 with the use of one less inductor. The circuit may be further simplified in cases where the center tapping to ground of L_1 is not necessary to reduce pickup from the lead-in.

Summary of the Discussed Passive Networks

The single tuned circuit system of Figure 1 in which a damping resistor was used to provide reflected terminating resistance for a

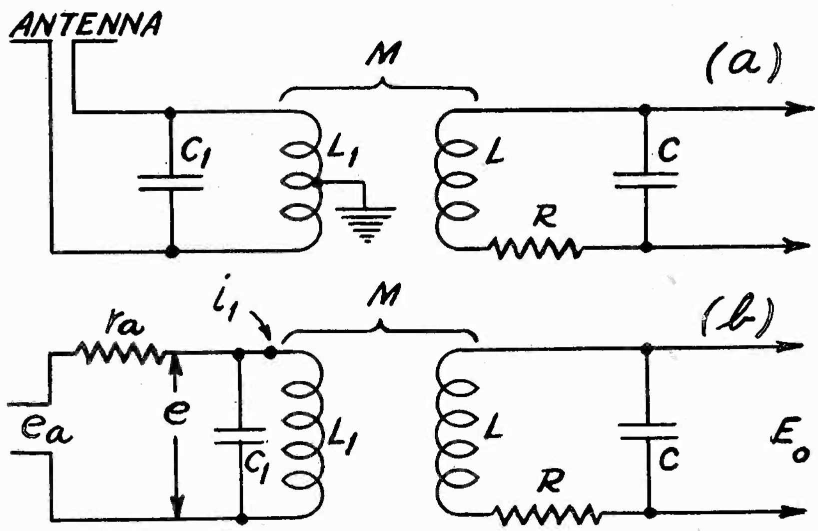

Fig. 7

transmission line of characteristic impedance equal to the internal impedance of the antenna has a gain characteristic equal to

$$g_o = \frac{\sqrt[4]{\frac{1}{P^2} - 1 - \frac{\Delta\omega}{\omega_o}}}{\sqrt{2\Delta\omega C r_a}}$$

and when $P = 0.9$, $\dfrac{\Delta\omega}{\omega_o} = 0.1$,

$$g_o = \frac{0.42}{\sqrt{\Delta\omega C r_a}}$$

Other single-tuned systems are inferior to it except at small (unusable) values of P.

Two tuned circuits critically coupled in the manner shown in Figure 4 or in Figure 7 have gain from Equation (8) equal to

$$g_o = \frac{0.5}{\sqrt{\Delta\omega C r_a}}$$

The system of Figure 5, while not providing proper termination for a transmission line, has the interesting gain characteristic that

$$g_o = \frac{1}{\sqrt{\Delta\omega C r_a}}$$

The system of Figure 6 provides termination and has better gain and selectivity possibilities than Figure 4 or 7. This system, when the coil and tube resistances are negligible, has a gain factor

$$g_o = \frac{0.707}{\sqrt{\Delta\omega C r_a}}$$

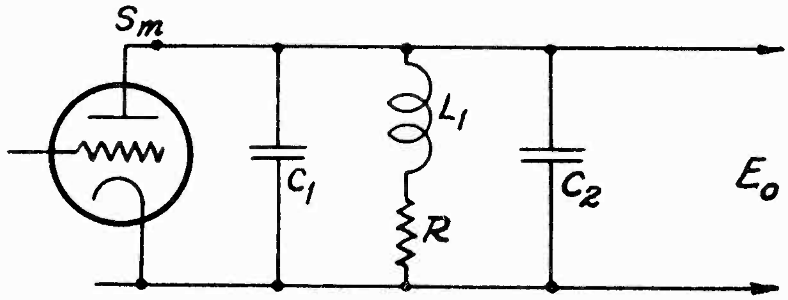

Fig. 8

However, in practice the tube input conductance is not negligible so that the realizable gain lies between that given by (8) and that given by (12).

R-F Stage

The r-f stage circuits treated in this analysis are restricted to two types, namely those having single and double tuned output circuits. These are treated in a manner similar to that used with the preceding antenna circuits, and symbols which have been previously used have identical significance.

(a) One Tuned Circuit

In Figure 8 is shown an r-f stage. C_1 is the capacitance associated with the output of the r-f amplifying tube. C_2 is the capacitance associated with any succeeding tube. C_1 and C_2 combine to act as one capacitance, C_T. When $\omega_o^2 L C_T = 1$, gain as derived in the appendix is

$$g_o = \frac{S_m}{\Delta\omega C_T} \times \sqrt{\frac{1}{P^2} - 1} \qquad (14)$$

and when $P = 0.9$

$$g_o = \frac{0.48\, S_m}{\Delta\omega (C_1 + C_2)}$$

The response ratio for any frequency remote from ω_o is

$$\frac{g_o}{g} = 1 + j\frac{\omega_o}{\Delta\omega}\sqrt{\frac{1}{P^2} - 1}\left(k - \frac{1}{k}\right) \tag{15}$$

This expression is similar to that of Equation (3), showing image ratios very close to those obtained with one tuned antenna circuit.

(b) Two Tuned Circuits

With two tuned circuits such as shown in Figure 9 considerable improvement in gain may be obtained for comparable flat transmis-

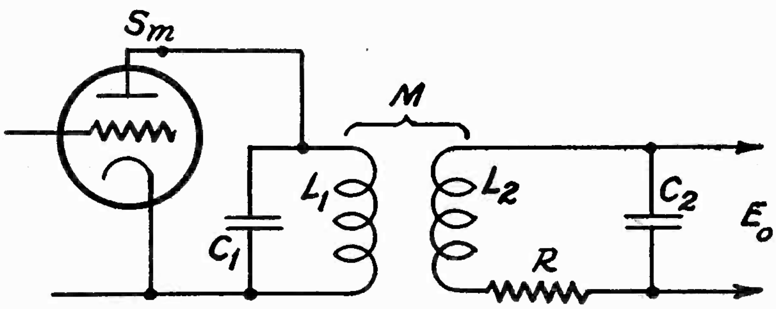

Fig. 9

sion. The elements of Figure 11 are assumed to have the following relationships:

$$\omega_o{}^2 L_1 C_1 = \omega_o{}^2 L_2 C_2 = 1$$
$$\omega_o M = \Delta\omega\sqrt{L_1 L_2}$$
$$R = \Delta\omega L_2$$

Gain at any frequency is

$$g = \frac{S_m}{\Delta\omega\sqrt{C_1 C_2}} \times \frac{1}{j\dfrac{\omega_o}{\Delta\omega}(k^2 - 1) - \left(\dfrac{\omega_o}{\Delta\omega}\right)^2\left(k^3 + \dfrac{1}{k} - 2k\right) + k^3} \tag{16}$$

At center frequency

$$g_o = \frac{S_m}{\Delta\omega\sqrt{C_1 C_2}} \tag{17}$$

At cut-off, when $\dfrac{\omega_o}{\Delta\omega} = 10$, $(k = 1.05)$ gain is

$$g_1 = 0.98\ g_o$$

If $C_1 = C_2 = 20\ \mu\mu\text{f}$

$\Delta\omega = 2\pi \times 4.5 \times 10^6$

$S_m = 6300 \times 10^{-6}$ (with resistive cathode impedance to stabilize input conductance)

then $g_o = 11$

For large values of k the terms

$$-\left(\frac{\omega_o}{\triangle\omega}\right)^2\left(k^3+\frac{1}{k}-2k\right)+k^3$$

will determine the approximate relative gain. If, for example, image ratio were to be determined where

$$k = 1.5$$

$$\text{and} \quad \frac{\omega_o}{\triangle\omega} = 10$$

Then, $$\text{image ratio} = 100$$

This is better than the image ratio obtained from the double tuned antenna circuit of Figure 4, which resulted in an image ratio of 23, and much better than the image ratio of one tuned circuit.

Combining R-F and Antenna Circuits

When single tuned circuit stages are cascaded, the choice of alignment frequency is quite important. Under certain conditions greatest gain for a given departure from flat transmission shape will result when the circuits are aligned at opposite cut-off frequencies. This is illustrated in Figure 10. Curve A depicts the transmission shape of one tuned circuit and curve B shows the accentuated attenuation resulting from the compounding of two such circuits as A. If alignment is made at cut-off frequencies as shown in curves C and D, the overall response will be that of curve E. To approach this flatness of response with circuits adjusted at ω_o would require more damping than is indicated by curve A, with attending loss in gains. When a double tuned circuit is combined with a single tuned circuit, the attenuation of the latter at cut-off may be compensated by double peaks in the former with resulting high overall gain. Less damping in both stages may be used in this case.

Station Selection

Individual channel selection has sufficient advantage over continuous tuning, both in simplicity of operation and electrical performance, to be considered essential in receiver design. Selection may be made by either push buttons or rotating switch. Probably the latter offers the greater possibility of restricting lead length and compacting assembly, in the event that channel selection is obtained by tapped inductors rather than by separate components for each channel.

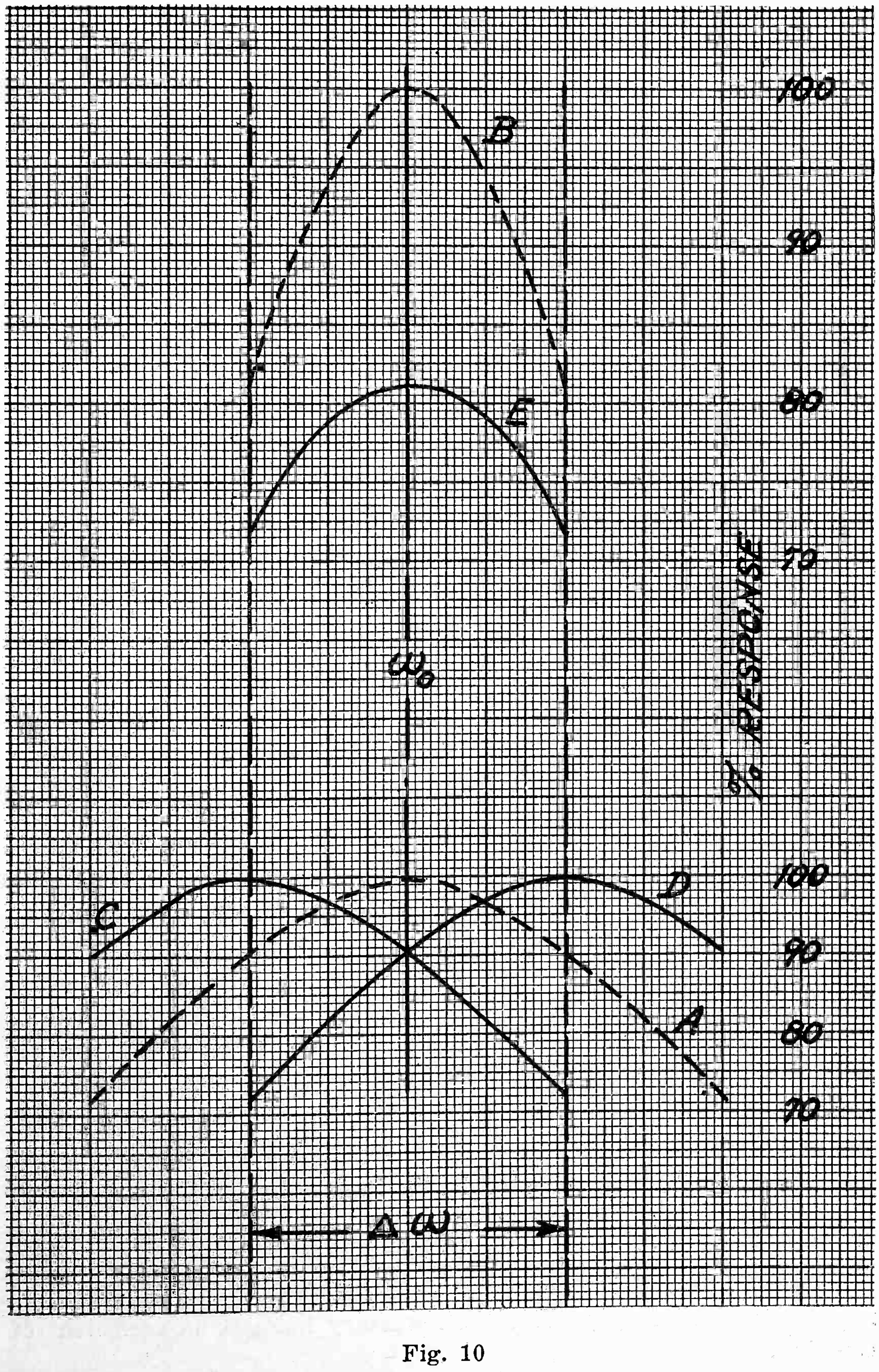

Fig. 10

It is important that nearly optimum coupling be maintained if maximum energy transfer at each channel is to be obtained. This implies that the percentage coupling must be materially reduced at the high-frequency end of the television frequency range. Several methods which produce this effect are described below.

In Figure 11 is shown a system in which the total tank circuit inductances are comprised *of* L'_1, fixed, plus L''_1, tapped, with like treatment of L'_2 plus L''_2. Coupling between L_p and L''_1 may decrease rapidly for high-frequency channels since L'_1 may be made any predetermined portion of the total tuning inductance. Likewise, coupling to L''_2 may be subjected to the same design treatment. If L''_1 and L'_1 are designed to provide optimum coupling at both ends of the tuning spectrum, the deviation from optimum coupling at the middle tuning

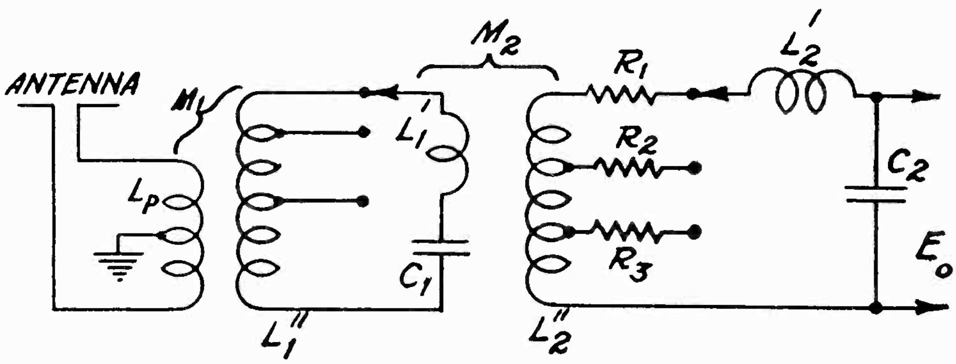

Fig. 11

frequencies will not be great. Further overall exactness may be had by designing L''_1, and L'_1 to provide optimum coupling near the ends of the tuning spectrum, but partly removed toward the center of the spectrum. This will provide two "cross-over" points for optimum coupling.

A second method is to provide reversed capacitive and inductive couplings. This system would not apply well to coupling between L_p and L_1, Figure 4, for instance, since in a system of this type capacitive coupling is reduced to a minimum to insure negligible signal pickup from the lead in. It does apply to coupling between L_1 and L_2, and to circuits where the above mentioned difficulty is not a factor.

A third method is to construct coils in the physical form indicated in Figure 12. For high-frequency tuning L'_1 and L'_2 are used. They may be so spaced as to provide proper coupling. The low-frequency spectrum is assumed tuned by $L'_1 + L''_1$, $L'_2 + L''_2$. The increased nearness of coils may provide the necessary increase in coefficient of coupling.

Internal Receiver Noise

The problems of shot noise and thermal agitation in television receivers are similar to the problems encountered in broadcast receiver design, although accentuated in importance by the much greater modulation band width. Theoretical calculations will assist in determining the chief sources and relative magnitudes of noise, particularly since accurate experimental tests to determine the ratio of thermal agitation to shot noise are complex. The often used method of shorting the grid of the first tube (to remove thermal agitation effects) and observing the change in noise output voltage is not rigorous. The grid load impedance is a part of the feed-back circuit to the grid of the first tube, and removing this impedance introduces factors which may not be neglected. When an unby-passed cathode resistor in the first tube is used, as is generally the case, the experimental determination of sources and magnitudes of noise is further complicated.

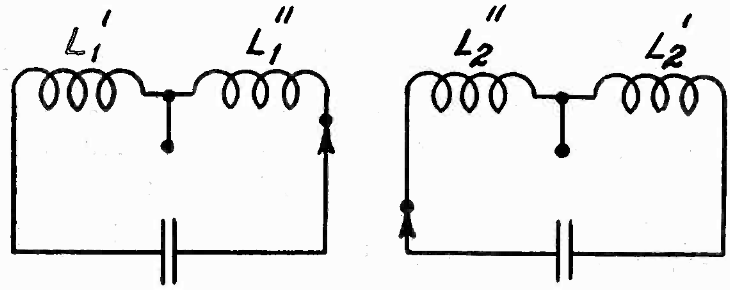

Fig. 12

Determination of television receiver noise involves many considerations, the magnitude and effect of which have not at present been completely evaluated. Particularly is this true of the optical effects of noise which are influenced by such factors as line structure and ratio of high- and low-frequency noise components, etc. The following discussion is therefore limited to an evaluation of the theoretical RMS noise voltages, and the more complex considerations of optical effects are not dealt with.

Thermal agitation for one tuned circuit as shown in Figure 1, with $P = 0.9$ will be used as an example.

The impedance looking back from the terminals of E_o may be expressed as:

$$Z_n = 2r_a g_o^2$$

as derived from Equation (2) for specific values of components

$$g_o = 2.17$$

therefore, $Z_n = 2 \times 75 \times 2.17^2 = 707$ ohms

This resistance component will be maintained (nearly) over the modulation band. In cases where the resistance component varies materially, a compensation for the effective band reduction must be made.

The noise voltage generated in 707 ohms for an assumed band width of 4.5 Mc may be calculated from the expression for thermal agitation.

$$\begin{aligned} \text{Thermal noise voltage} &= 1.27\sqrt{Rgf} \times 10^{-10} \\ &= 7.2\ \mu \text{ volts} \end{aligned} \tag{18}$$

With an antenna gain of 2.17 the equivalent noise voltage in the antenna is

$$E_a = \frac{7.2}{2.17} = 3.32\ \mu\text{v.}$$

An r-f amplifier tube will have a shot-noise voltage which may be expressed as an equivalent resistance in series with the grid. This "resistance" will then generate noise voltage calculable from Equation (18). The 1852 tube has been measured to have the approximate equivalent of 2000 ohms when used at 2 volts bias and with a degenerating cathode resistor to stabilize the input impedance (as is usually done). The use of such a tube as an r-f amplifier will generate a grid voltage which may be calculated as follows:

$$N_g = 1.27\sqrt{R_g f} \times 10^{-10}$$

$$N_g = 1.27\sqrt{2000 \times 4.5 \times 10^6} \times 10^{-10} = 12.2\mu\text{v.}$$

The r-f stage under consideration is preceded by an antenna circuit gain of 2.17 times, hence the antenna noise voltage derived from the r-f stage alone is 5.6 μv.

The total antenna noise voltage is

$$E_a = \sqrt{5.6^2 + 3.32^2} = 6.5\ \mu\text{v.}$$

If the r-f stage gain is high, shot noises generated in succeeding tubes will have small effect on total noise.

The 1852 tube used as a converter may have roughly 6000 ohms equivalent grid noise resistance, when a degenerating cathode resistor is used. The grid noise voltage developed is then $\sqrt{3}$ times higher than the same tube used as an r-f amplifier, or

$$N_g = 21.2\ \mu\text{v.}$$

Used without an r-f stage the equivalent antenna noise voltage would be 9.75 μv and total antenna noise would be

$$E_a = \sqrt{9.75^2 + 3.32^2} = 10.3\ \mu\text{v.}$$

This is 58 per cent more than that obtained with a high gain r-f stage and antenna circuit.

An r-f stage will lose its effectiveness in reducing noise if its gain is small. For example if the r-f stage gain for these considerations were only 1.22, the total r-f, converter, and antenna noise would be equal to that obtained without an r-f stage. If the r-f stage gain were 5, the total antenna noise voltage would be 6.8 μv. The effect of the converter hiss is then negligible.

The use of existing pentagrid types of converters as the first tube with their relatively low conversion gain (usually less than unity for the stage) results in materially reduced signal to noise ratios. Low gain and resulting high noise level will undoubtedly preclude their widespread use.

Interference noises, such as automobile ignition, diathermy, and other electrical disturbances frequently exceed the receiver internal noise level, and preclude the favorable continuous reception of signals of less than several millivolts. Some locations are quite free from these forms of interference, and in such places a receiver of low noise level is quite desirable. Under these conditions the above gain of 58 per cent in signal-to-noise level obtained by the use of an r-f stage would have some utility. However, for locations beset with considerable and sustained interference, the merit of low noise level in the receiver will not be greatly appreciated. Hence, the practicability of an r-f stage for reducing noise will be influenced by the nature and location of the service for which the receiver is intended.

There are other and perhaps more important considerations in determining the desirability of an r-f stage. Preselector selectivity is important in reducing interference from powerful signals adjacent to the desired channel and in reducing image and i-f interference. An r-f stage of one tuned circuit preceded by a single tuned antenna circuit has the disadvantage of possessing less selectivity than that of two coupled antenna circuits. But an r-f stage provides amplification for both picture and sound frequencies simultaneously, thus offering possibilities of economy. Its use in sensitive receivers where the reduction of otherwise necessary i-f gain may permit more complete stabilization of the receiver has possibilities of development. Whether or not an r-f stage should be used in any particular receiver design requires very careful study and test. The overall results are influenced materially by various practical factors in addition to the fundamental theoretical ones, and present detailed construction technique is such that these practical considerations tend to nullify the normal advantages of an r-f stage, from the overall standpoint.

APPENDIX A

Derivation of Equation (1) from Figure 1(b) and from information in the text preceding Equation (1)

$$g = \frac{k\omega_o M}{k\omega_o C \left[r_a R + j\omega_o L r_a \left(k - \frac{1}{k} \right) + k^2 \omega_o^2 M^2 \right]}$$

$$g = \frac{1}{\omega_o C \sqrt{r_a R}\,(1 + k^2) + j \sqrt{\frac{r_a}{R}} \left(k - \frac{1}{k} \right)}$$

Gain at center frequency (where $k = 1$) is

$$g_o = \frac{1}{2\omega_o C \sqrt{r_a R}}$$

The gain at cut-off frequency considering a total frequency pass band of $\Delta\omega$ width has the condition that

$$k = \frac{\omega_o + \frac{\Delta\omega}{2}}{\omega_o} = 1 + \frac{\Delta\omega}{2\omega_o}$$

$$\left(k - \frac{1}{k} \right) = \frac{\Delta\omega}{\omega_o} \quad \text{(very closely)}$$

$$(1 + k^2) = 2 + \frac{\Delta\omega}{\omega_o} \quad \text{(very closely)}$$

and gain at cutoff is

$$g_1 = \frac{1}{\frac{1}{2g_o}\left(2 + \frac{\Delta\omega}{\omega_o} \right) + j \frac{\Delta\omega}{\omega_o} \frac{\sqrt{r_a}}{R}}$$

The relative response at cut-off is $\frac{g_1}{g_o}$ which may be expressed as

$$P = \frac{g_1}{g_o}$$

$$P = \cfrac{1}{\cfrac{g_o\left(2+\cfrac{\Delta\omega}{\omega_o}\right)}{2g_o} + j\cfrac{\Delta\omega}{\omega_o}\sqrt{\cfrac{r_a}{R}}\,g_o}$$

$$P = \cfrac{1}{\left(1+\cfrac{\Delta\omega}{2\omega_o}\right) + j2r_aC\Delta\omega g_o^2}$$

then, taking the absolute value

$$\left\{\left(1+\frac{\Delta\omega}{2\omega_o}\right)^2 + 4r_a^2C^2\Delta\omega^2 g_o^4\right\}^{\frac{1}{2}} = \frac{1}{P}$$

$$g_o = \frac{\sqrt[4]{\dfrac{1}{P^2} - 1 - \dfrac{\Delta\omega}{\omega_o}}}{\sqrt{2\Delta\omega C r_a}} \qquad (1)$$

Values of the numerators of the various gain formulas may be read from the table of Figure 2 for various values of P, the edge frequency response factor.

FIGURE 2

P	$\sqrt{\dfrac{1}{P^2}-1}$	$\sqrt[4]{\dfrac{1}{P^2}-1}$	$\sqrt{\dfrac{1}{P^2}-1-\dfrac{\Delta\omega}{\omega_o}}$	$\sqrt[4]{\dfrac{1}{P^2}-1-\dfrac{\Delta\omega}{\omega_o}}$
0.95	0.329	0.574	0.084	0.29
0.9	0.483	0.695	0.365	0.605
0.85	0.62	0.787	0.53	0.73
0.8	0.75	0.865	0.68	0.825
0.707	1.0	1.0	0.948	0.975

When $\dfrac{\Delta\omega}{\omega_o} = 0.1$ as a probable maximum

Appendix B

Derivation of Equation (4) gain under the conditions imposed is

$$g = \frac{\omega M}{\left\{ r_a + \dfrac{\omega^2 M^2}{j\omega L - j\dfrac{1}{\omega C}} \right\} \left\{ j\omega L - \dfrac{j}{\omega C} \right\} \omega C}$$

$$g = \frac{M}{j r_a C \omega_o L \left(k - \dfrac{1}{k} \right) + k^2 \omega_o^2 M^2 C}$$

At center frequency

$$g_o = \frac{1}{\omega_o^2 M C}$$

At cut-off

$$k = 1 + \frac{\Delta\omega}{2\omega_o}$$

$$g_1 = \frac{1}{\left(1 + \dfrac{\Delta\omega}{\omega_o}\right) \omega_o^2 M C + j \dfrac{r_a \Delta\omega C}{\omega_o^2 M C}}$$

$$P = \frac{g_1}{g_o} = \frac{1}{\left(1 + \dfrac{\Delta\omega}{\omega_o}\right) + j r_a \Delta\omega C g_o^2}$$

$$P^2 = \frac{1}{\left(1 + \dfrac{2\Delta\omega}{\omega_o}\right) + r_a^2 \Delta\omega^2 C^2 g_o^4}$$

$$g_o = \frac{\sqrt[4]{\dfrac{1}{P^2} - 1 - \dfrac{2\Delta\omega}{\omega_o}}}{\sqrt{\Delta\omega C r_a}} \qquad (4)$$

APPENDIX C

Derivation of Equation (7) from the conditions imposed by the text.

$$g = \frac{k^2\omega_o^2 M_1 M_2}{\left[r_a + \cfrac{k^2\omega_o^2 M_1^2}{j\omega_o L_1\left(k - \cfrac{1}{k}\right) + \cfrac{k^2\omega_o^2 M_2^2}{j\omega_o L_2\left(k - \cfrac{1}{k}\right) + R}} \right]}$$

times
$$\frac{1}{\left[j\omega_o L_1\left(k - \cfrac{1}{k}\right) + \cfrac{k^2\omega_o^2 M_2^2}{j\omega_o L_2\left(k - \cfrac{1}{k}\right) + R} \right]}$$

times
$$\frac{1}{\left(j\omega_o L_2 \left[k - \cfrac{1}{k} \right] + R \right)(k\omega_o C_2)}$$

which reduces to:

$$g = \frac{1}{\sqrt{\Delta\omega C_2 r_a}} \times \frac{1}{2k - \left(\cfrac{\omega_o}{\Delta\omega}\right)^2 \left(k - \cfrac{2}{k} + \cfrac{1}{k^3}\right) + j\cfrac{\omega_o}{\Delta\omega}\left(k^2 - \cfrac{1}{k^2}\right)} \qquad (7)$$

APPENDIX D

Derivation of Equation (10) from the conditions imposed by the text.

$$g = \frac{k^2\omega_o^2 M_1 M_2}{\left\{ r_a + \cfrac{k^2\omega_o^2 M_1^2}{j\omega_o L_1\left(k - \cfrac{1}{k}\right) + \cfrac{k^2\omega_o^2 M_2^2}{j\omega L\left(k - \cfrac{1}{k}\right)}} \right\}}$$

times
$$\frac{1}{\left\{ j\omega_o L_1 \left(k - \frac{1}{k} \right) + \frac{k^2 \omega_o^2 M_2^2}{j\omega_o L \left(k - \frac{1}{k} \right)} \right\}}$$

times
$$\frac{1}{j\omega_o L \left(k - \frac{1}{k} \right) k\omega_o C}$$

$$g = \frac{1}{\sqrt{\Delta\omega C r_a}} \times \frac{1}{k - \left(\frac{\omega_o}{\Delta\omega} \right)^2 \left(k - \frac{2}{k} + \frac{1}{k^3} \right) + j \frac{\omega_o}{\Delta\omega} (k^2 - 1)} \tag{10}$$

APPENDIX E

Derivation of Equation (12)

$$g = \frac{1}{\sqrt{(r_a + r'_a) \Delta\omega C_2}}$$

times
$$\frac{1}{k - \left(\frac{\omega_o}{\Delta\omega} \right)^2 \left(k - \frac{2}{k} + \frac{1}{k^3} \right) + j \frac{\omega_o}{\Delta\omega} (k^2 - 1)}$$

$$g = \frac{1}{\sqrt{2 r_a \Delta\omega C_2}}$$

times
$$\frac{1}{k - \left(\frac{\omega_o}{\Delta\omega} \right)^2 \left(k - \frac{2}{k} + \frac{1}{k^3} \right) + j \frac{\omega_o}{\Delta\omega} (k^2 - 1)}$$

At center frequency
$$g_o = \frac{1}{\sqrt{2 r_a \Delta\omega C_2}} \tag{12}$$

APPENDIX F

Derivation of Equation (14)

Gain may be expressed as follows:

$$g = \frac{S_m}{R\omega_o^2 C_T^2 + j\omega_o C_T \left(k - \frac{1}{k} \right)}$$

Gain at ω_o (center frequency) is then

$$g_o = \frac{S_m}{R\omega_o^2 C_T^2}$$

At cut-off $k = 1 + \dfrac{\Delta\omega}{2\omega_o}$ gain is

$$g_1 = \frac{S_m}{R\omega_o^2 C_T^2 + j\Delta\omega C_T} \qquad \text{(about)}$$

The response factor P, is then

$$P = \frac{g_1}{g_o} = \frac{S_m R\omega_o^2 C_T^2}{(R\omega_o^2 C_T^2 + j\Delta\omega C_T)\, S_m}$$

$$P = \frac{1}{1 + \dfrac{j\Delta\omega C_T S_m}{R\omega_o^2 C_T^2 S_m}}$$

$$P = \frac{1}{1 + \dfrac{j\Delta\omega C_T g_o}{S_m}}$$

Then taking the absolute value

$$\frac{1}{P^2} = 1 + \frac{\Delta\omega^2 C_T^2 g_o^2}{S_m^2}$$

$$g_o = \frac{S_m}{\Delta\omega C_T} \times \sqrt{\frac{1}{P^2} - 1} \qquad (14)$$

A table of the radical $\sqrt{\dfrac{1}{P^2} - 1}$ is shown in Figure 2 for various values of P. If, for example, P is taken as 0.9 then

$$g_o = \frac{.48\, S_m}{\Delta\omega (C_1 + C_2)} \qquad (15)$$

APPENDIX G

Derivation of Equation (16)

Gain may be derived as follows:

Voltage across $C_1 = E$ (produced by IV on grid)

$$E = \frac{S_m \omega_o{}^2 L_1{}^2}{j\omega_o L_1\left(k - \dfrac{1}{k}\right) + \dfrac{k^2 \Delta\omega^2 L_1 L_2}{R + j\omega_o L_2\left(k - \dfrac{1}{k}\right)}}$$

$$E = \frac{S_m}{j\omega_o C_1\left(k - \dfrac{1}{k}\right) + \dfrac{k^2 \Delta\omega^2 C_1}{\Delta\omega + j\omega_o\left(k - \dfrac{1}{k}\right)}}$$

The current through L_1 is:

$$i_1 = \frac{E}{jk\omega_o L_1}$$

and gain may be expressed as

$$g = \frac{E\Delta\omega\sqrt{L_2}}{\left[\Delta\omega + j\omega_o\left(k - \dfrac{1}{k}\right)\right] k\sqrt{L_1}}$$

$$g = \frac{S_m}{jk\dfrac{\omega_o}{\Delta\omega}\sqrt{C_1 C_2}\left(k - \dfrac{1}{k}\right)}$$

times
$$\frac{1}{\left[\Delta\omega + j\omega_o\left(k - \dfrac{1}{k}\right)\right] + k^3\Delta\omega\sqrt{C_1 C_2}}$$

$$g = \frac{S_m}{\Delta\omega\sqrt{C_1 C_2}} \times \frac{1}{\mathrm{j}\dfrac{\omega_o}{\Delta\omega}(k^2 - 1) - \left(\dfrac{\omega_o}{\Delta\omega}\right)^2\left(k^3 + \dfrac{1}{k} - 2k\right) + k^3} \qquad (16)$$

SOUND INSULATION CHARACTERISTICS FOR IDEAL PARTITIONS*

BY

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VARIOUS methods of specifying the practical performance of partitions and other building units have been proposed and used from time to time, all tending, for reason of simplicity, to the use of a single number expression.

Let us consider these methods in the order of their appearance. Early measurements on the effectiveness of a partition consisted in observing the sound intensity on the two sides of the partition and expressing the sound reduction as the ratio of these two intensities. This ratio was later expressed in decibels. Because the measurement was made at a single frequency, usually 500 c.p.s., a single number expression existed per se. When, later, measurements were extended to include a number of frequencies, the old criterion of specifying the reduction at 500 c.p.s. was still adhered to by some observers, while others used the average value of the results at all the several frequencies used. A yet different method of determining an average figure grew from an attempt to procure uniformity among the different observers by drawing the average value over the arbitrarily selected range 250 c.p.s. to 2000 c.p.s.

Early observers did not always differentiate between transmission loss and reduction factor, which fact led to some uncertainty in the interpretation of the data. The transmittivity, τ, is defined as the ratio of the intensity in the beam transmitted by the partition to the intensity in the beam incident upon the partition. Transmission loss expresses this ratio in decibels as

$$\text{T.L.} = 10 \log_{10} \frac{I_i}{I_t} = 10 \log_{10} \frac{1}{\tau}$$

The reduction factor, k, is the ratio of the sound energy density in the transmitting room to that in the receiving room, and is related to the transmission loss by the expression

$$\text{R.F. (decibels)} = \text{T.L.} + 10 \log_{10} \frac{a}{A}$$

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where A is the area of the test partition and a is the total absorption in the receiving room. The receiving room conditions therefore appear as a correction to be applied to the reduction factor in order to obtain the transmission loss. In the laboratory this factor may vary over a considerable range, either side of zero depending upon whether a dead room or a reverberant room is used as the receiving room. Knudsen[1] performed an invaluable piece of work in analyzing all these various data and expressing the results uniformly in reduction factors.

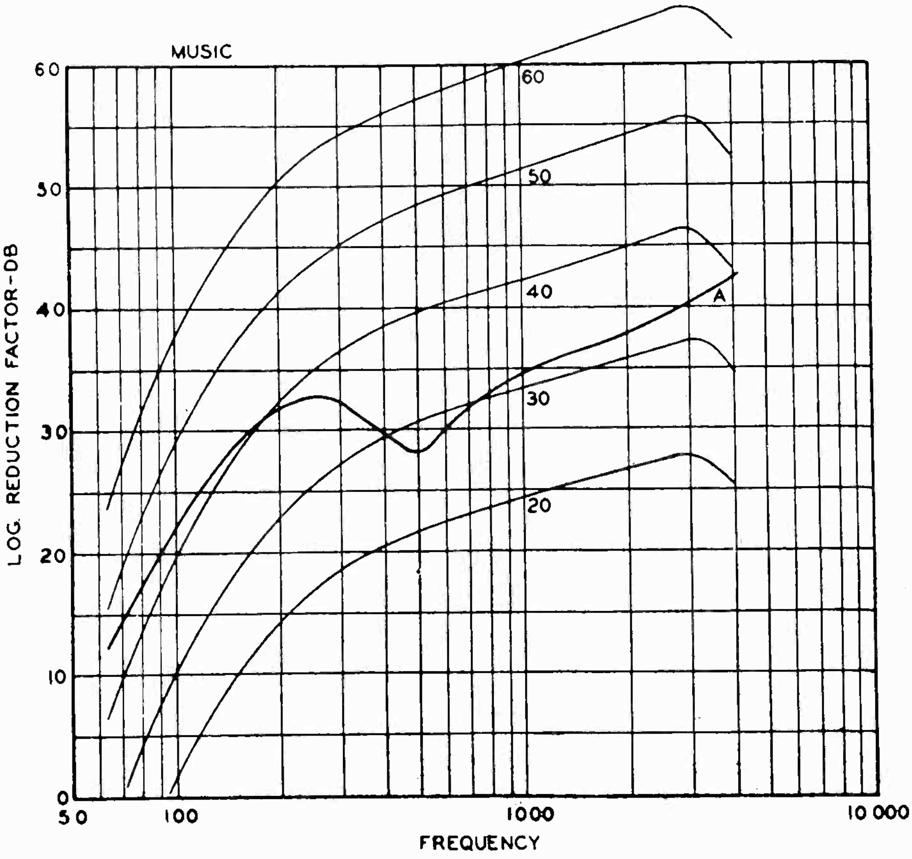

Fig. 1—Contours of uniform loudness reduction for music.

The writer has long objected to the use of an arbitrary average value of transmission loss on the grounds that any average, no matter how determined, does not present a true picture of the action of a partition. Considering now only the physical side of the problem, the establishment of an average value from the transmission characteristic does not reveal important features such as a pronounced weakness at some particular point within the usable range, or a characteristic in which too much attenuation is provided in regions where it is not required and too little where it is important, with the result that a good average is obtained upon an unsatisfactory structure. The situation is analogous to the attempt to specify the attenuation of an electrical network as an average of the losses introduced at certain frequencies. A true specification for performance should consider the use to which the unit is to be put, the type of conditions under which it must function, and a statement of the minimum rather than of the average performance to be expected. In addition to the physical considerations, the mechanism of hearing must be brought into the picture so that some idea of the effect on the loudness of the sound may be obtained.

Ross[2] has proposed what he termed a "Minimum Loudness Reduction" factor obtained by plotting the transmission loss-frequency characteristic of the partition on a background of Uniform Loudness Reduction Contours. Figure 1 shows such a set of Uniform Loudness Reduction Contours with the reduction factor of a partition, *A*, drawn on it. These contours show the reduction in intensity required to give the reduction in loudness indicated, and were established after a consideration of the frequency distribution of speech and music, the characteristic of the human ear, and normally encountered noise levels. It is seen that at 65 c.p.s. the partition has a maximum loudness reduction factor of 46 db, and that the factor rapidly drops off to a minimum value of 27 db at 500 c.p.s. Observing the performance of the partition in this manner has the advantage that the magnitude of the weakness and its location in the frequency spectrum are readily apparent. A glance at the figure conveys the information that sounds in the region of 500 c.p.s. will generally be first heard as the intensity of the sound on the other side is increased. This effect is actually experienced in the hearing of a very limited, and therefore very annoying, portion of a neighbor's radio program transmitted through the partition. The minimum loudness reduction figure of 27 db at 500 c.p.s. might be used as a statement of merit of the partition.

Recently Morreau[3] has very carefully analyzed the problem after considering the response of the human ear, the loudness of complex tones and receiving room conditions and comes to the conclusion that "a figure which may be derived by taking the arithmetic mean of a representative number of sound reduction factors over the frequency range 200-2000 c.p.s. does, however, give an approximation to the minimum reduction in equivalent loudness afforded, under practical conditions, by many walls to many sounds." The validity of this conclusion is restricted by the assumption of a mass controlled structure having an attenuation that increases at the rate of 5 db per octave, and by the adoption of 1000 c.p.s. as the highest frequency considered, although, in effect, the value at 700 c.p.s. is taken as the mean.

In general, partitions do not always exhibit a mass reaction, which produces an attenuation that rises monotonicly with frequency, but sometimes show an irregularity as in the case of partition *A*. Irregularities are especially likely to occur in the characteristics of special partitions of high insulating value which consist of one light face resiliently mounted to a heavier structure. In this connection, it might well be said here that measurements of transmission loss should be made on partitions of a size encountered in practice, say $8' \times 16'$, rather than on small specimens, in order that the test partition possess the same mechanical constants as its counterpart in the field. In addition, so that any irregularities in the characteristic are not passed over, the frequency interval between determinations should not be less than

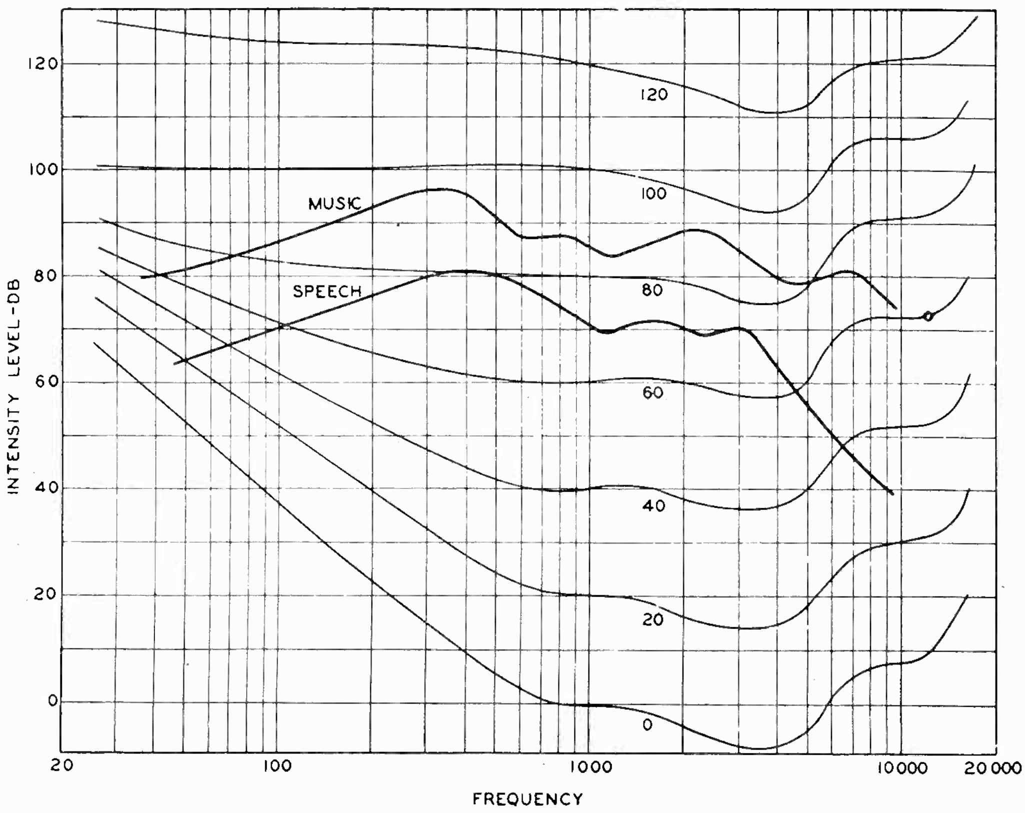

Fig. 2—Spectral distribution of peak acoustic powers in speech and symphony music, showing upper limit of lower 75 per cent of peaks.

$\sqrt[3]{2}$ (three points per octave) or, what is substantially the same, $\sqrt[10]{10}$ (ten points per logarithmic cycle).

Morreau's analysis also shows that substantially the same conclusions can be drawn whether the loudness of pure tones is used in the analysis or whether the loudness of complex tones is taken into consideration.

A different approach to the problem of accurately specifying the effectiveness of a sound insulating structure is to consider the loudness level to which the partition will reduce normally encountered sounds. Figure 2 shows the spectral distribution of energy in representative speech and symphony music, as determined by Sivian[4] and by Sivian, Dunn, and White[5], drawn on a background of Fletcher-Munson equal loudness contours. The data are presented as the upper limit of the lower 75 per cent of the peak powers measured. Several characteristics are readily apparent. Peak powers of music have intensity maxima at 350 c.p.s. and 2300 c.p.s., while maxima for speech occur at 400 c.p.s. and 1600-3000 c.p.s. Speech, naturally, has a much lower level than music and contains less high frequency components.

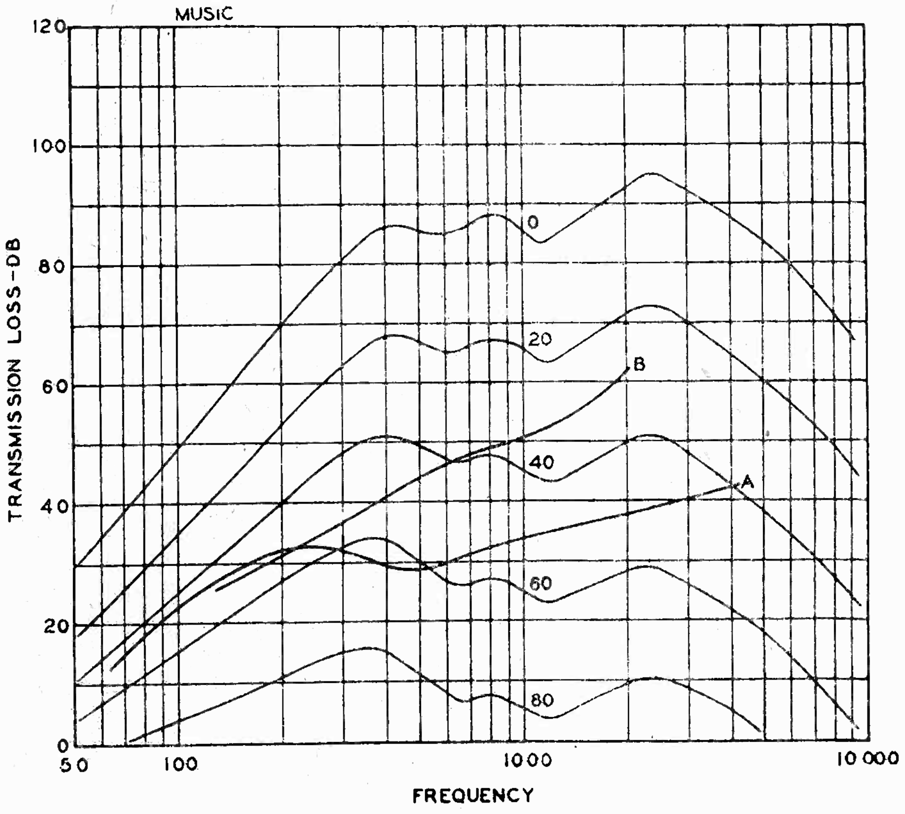

Fig. 3—Contours of transmission loss necessary to reduce symphony music to loudness level indicated.
Partition A—3″ Pyrobar plastered both sides.
Partition B—Double 2″ solid gypsum tile with 2″ air space, unplastered.

Figure 3 shows a family of transmission loss contours for music, each curve specifying the attenuation characteristic necessary in a partition if it is to reduce symphony music to the loudness level indicated on the curve. Figure 4 shows the same contours for speech. The shape of these curves can be considered to be the shape of the transmission characteristic of an ideal partition, which is a partition that has no "un-used" insulation, or one which reduces all components of the incident sound to the same loudness level. Less attenuation is required at low frequencies both because of convergence of the equal loudness contours in that region and because of the lower original energy content in that region. By the same token, insulation in this region is very critical and small differences in T.L. will result in large variations in loudness reduction. This is another condition that the use of average values does not bring out. The same is true to a less degree for the high frequencies. The important point to observe, however, is that the secondary maximum for music peak powers occurs at very nearly the frequency at which the acuity of hearing is greatest, and therefore maximum sound insulation is required for the region of 2000-3000 c.p.s.

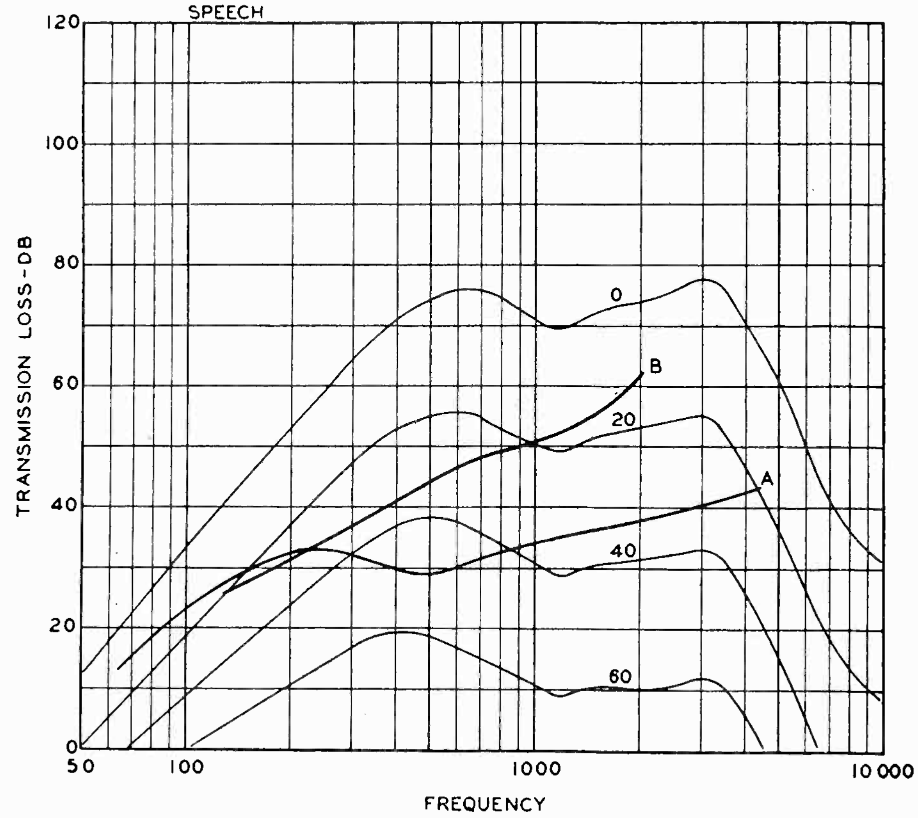

Fig. 4—Contours of transmission loss necessary to reduce declamatory speech to loudness level indicated.
Partition A—3″ Pyrobar plastered both sides.
Partition B—Double 2″ solid gypsum tile with 2″ air space, unplastered.

The attenuation characteristics of two partitions, *A* and *B*, are drawn on these contours and it is seen that they do not conform to the ideal shape. Partition *A* is 3″ Pyrobar plastered on both sides. Partition *B* is double 2″ solid gypsum tile, separated 2″, unbridged and structurally isolated and unplastered. On the basis of overall average, *A* has a T.L. of 31 db and *B* of 44 db. On the basis of the average from 250-2000 c.p.s., *A* has T.L. of 33 db and *B* of 48 db. On a minimum loudness reduction basis (music), *A* possesses a MLR of 27 db at 500 c.p.s. and *B* a MLR of 39 db at 220 c.p.s. Under the proposed plan of specification, *A* is said to be able to reduce symphony music produced on one side to a maximum loudness level of 63 db on the other side, the weakest point occurring at 500 c.p.s. Likewise *B* results in a maximum loudness level of 54 db at 270 c.p.s. For speech, the final maximum levels are 48 db at 500 c.p.s. for *A* and 35 db at 360 c.p.s. for *B*. These values are tabulated in Table I.

If the two partitions are merely to be compared, *A* can be said to be from 9 to 18 db better than *B* since the results from all methods yield differences within this range. If, however, a statement of absolute performance is required, the first two methods state only that certain mean values exist and give no further information. On the other hand, the Minimum Loudness Reduction method specifies the minimum reduction in loudness to be expected, as well as the frequency at which failure of the partition begins. The last method shows the maximum loudness to be expected on the one side of the partition, and the frequency where it occurs, if certain sounds originate on the other. In other words, the last two methods have something definite to say concerning a partition, what can be expected of it, and where its weak point is.

TABLE I

Summary of Results of Various Methods of Specifying Performance of Sound Insulating Structures

	Average	*Average 250-2000 c.p.s.*	*Minimum Loudness Reduction*		*Maximum Loudness Level*	
			Music	*Speech*	*Music*	*Speech*
Partition A 3″ Pyrobar plastered both sides	31 db	33 db	27 db at 500 c.p.s.	31 db at 500 c.p.s.	63 db at 500 c.p.s.	48 db at 500 c.p.s.
Partition B double 2″ solid gypsum tile with 2″ air space unplastered	44 db	48 db	39 db at 220 c.p.s.	49 db at 800 c.p.s.	54 db at 270 c.p.s.	35 db at 360 c.p.s.

The attenuation characteristics of the partitions have been drawn directly in T.L. since, in contrast to laboratory conditions, conditions in the field are usually such that the area of the partition is numerically of the same order of magnitude as the total sound absorption within the room and the correction between T.L. and reduction factor becomes negligibly small.

It is necessary to compare the maximum loudness of the transmitted sound with the existing noise level in order to determine whether or not the transmitted sound is audible, and therefore the frequency distribution of the ambient noise should be considered. In general, especially in instances of quiet apartments, studios, etc., applications in which sound insulation is really seriously considered, the noise energy is distributed fairly evenly throughout the frequency spectrum, being slightly higher in the mid-range than at either end. The noise level in offices tends to contain more high frequency components. For an approximation, for this purpose, the noise levels encountered can be considered to be uniformly distributed throughout the spectrum and a direct comparison can therefore be made.

The method of reduction to uniform loudness is most readily applied in evaluating sound insulation requirements for studios, etc., because in this instance the original sound is dealt with directly. This application is also one in which maximum sound insulation is required. In apartments, a 90 db level is not likely to be experienced, any original music arising probably having a level 15-30 db below this and insulation must be provided then only against incident sound whose level is much lower than shown by Figure 2. Moreover, the sound may have a slightly different distribution because of the characteristic of the reproducing system. It seems that the best method of attacking this aspect of the problem is to prepare similar contours based on the reproduction of the same sound at mean levels of say 20 db and 40 db lower, and judging the partition's performance on the set of contours appropriate to the conditions under which the partition will be used in the field. Three different levels could be agreed upon and standardized similar to the procedure used in adopting the A, B and C levels and characteristics in sound level meters. Also performance values could be agreed upon for several different grades of insulation.

It is frequently necessary to insulate against sounds other than speech and music, such as, for example, noise from rotating and vibrating machinery, traffic noise, etc. Since complete data on the average distribution of energy in these sounds are not available, contours similar to those developed for speech and music have not been prepared. However, in instances in which insulation from certain particular sounds is required, an accurate survey should be made to determine the energy-frequency distribution of the sound and a set of such contours prepared on this basis. Then, with receiving room conditions known, the reduction factors of the partitions can be drawn on the contours and the merit of the structure ascertained under true field conditions. In general, it can be said that more insulation in the low frequency region will be required if noises of other than speech and music types are dealt with.

With the growing consciousness of the desirability of sound insulation and its increasing use in new buildings, the economic factor involved in the selection of the type of structure becomes of increasing importance. Formerly, it was decided that insulation was or was not required, and if it was, nothing was overlooked in the way of massive and elaborate construction, etc. to accomplish this end without too much regard to the cost. Nowadays the insulation of a partition must be accurately known before it can be decided whether or not its cost is justified for the application at hand. Some method similar to the one described must be used to estimate this performance.

Conclusions

The present method of evaluating the effectiveness of sound insulative structures on the basis of mean values of T.L. is inaccurate and unscientific in that it does not predict the performance of the unit in practice. In actual practice, we are interested primarily in what loudness level will result in the insulated room rather than in any physical characteristics of the partition. The transmission characteristic of "ideal" partitions, or partitions which reduce the incident sound to a uniform loudness level, has been derived from a consideration of the intensity and frequency distribution of typically encountered speech and music and of the mechanism of hearing, and it has been shown that a family of these characteristics can be used as a means of predicting the insulation furnished by a particular structure and its manner of failure. The manner of extension of this method to ascertain the merit of a partition for different types of sounds is given. It is suggested that a further study of the subject be undertaken toward the end of specifying certain procedure to follow and methods to use in evaluating the performance of partitions.

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THE APPLICATION OF THE TENSOR CONCEPT TO THE COMPLETE ANALYSIS OF LUMPED, ACTIVE, LINEAR NETWORKS

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(Continued from July issue)

IV. Illustrative Example of Method

The general theory outlined above can perhaps be clarified by considering in some detail the simple two-mesh network shown in Figure 2 in which two driving forces E_1 and E_2 are present.

The tensor $R_{\mu\nu}$ thus has the following three components $R_{11} = R_1 + R_{12}$, $R_{12} = R_{12}$, $R_{22} = R_2 + R_{12}$.

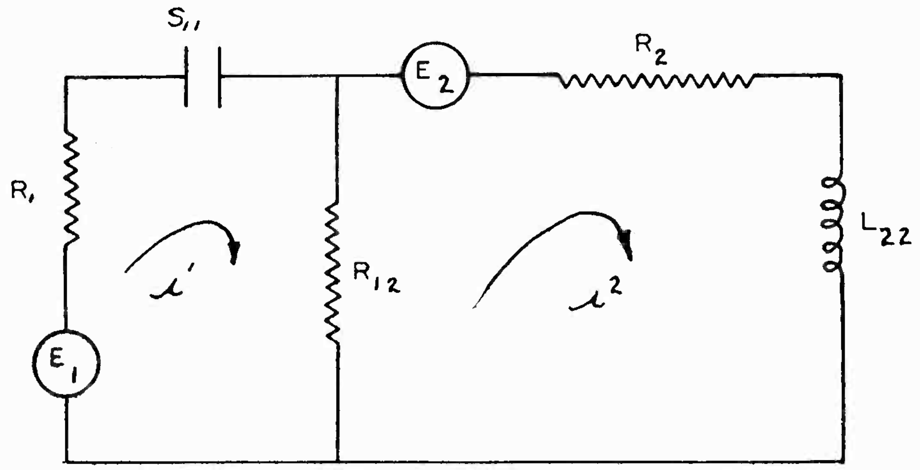

Fig. 2—Illustrative example comprising two meshes.

The tensors $S_{\mu\nu}$ and $L_{\mu\nu}$ have but one component each S_{11}, L_{22}.

The tensor i^ν has two components i^1, i^2.

The tensor q^ν has one component q^1.

The tensor E_μ has two components, E_1 and E_2.

The tensor $Z_{\mu\nu}$ has three components,

$$Z_{11} = R_{11} + \frac{S_{11}}{p}; \; Z_{12} = R_{12} = Z_{21}; \; Z_{22} = R_{22} + pL_{22}.$$

Hence $Z = Z_{11}Z_{22} - Z_{12}{}^2$.

$$A^{11} = (\text{cofactor of } Z_{11})/Z = Z_{22}/Z$$
$$A^{12} = (\text{cofactor of } Z_{12})/Z = -Z_{21}/Z$$
$$A^{22} = (\text{cofactor of } Z_{22})/Z = Z_{11}/Z$$

If the current in the second mesh is considered, then $\nu = 2$, and the integral equation (4)

$$\int_0^\infty i^\nu e^{-pt}dt = A^{\mu\nu}\left[\int_0^\infty E_\mu e^{-pt}dt + L_{\mu\sigma}(i^\sigma)_0 - \frac{S_{\mu\sigma}}{p}(q^\sigma)_0\right] \quad (4)$$

becomes, for μ, $\sigma = 1, 2$

$$\int_0^\infty i^2 e^{-pt}dt = A^{\mu 2}\left[\int_0^\infty E_\mu e^{-pt}dt + L_{\mu 1}(i^1)_0 + L_{\mu 2}(i^2)_0 - \frac{S_{\mu 1}}{p}(q^1)_0 - \frac{S_{\mu 2}}{p}(q^2)_0\right]$$

$$= A^{12}\left[\int_0^\infty E_1 e^{-pt}dt + L_{11}(i^1)_0 + L_{12}(i^2)_0 - \frac{S_{11}}{p}(q^1)_0 - \frac{S_{12}}{p}(q^2)_0\right]$$

$$+ A^{22}\left[\int_0^\infty E_2 e^{-pt}dt + L_{21}(i^1)_0 + L_{22}(i^2)_0 - \frac{S_{21}}{p}(q^1)_0 - \frac{S_{22}}{p}(q^2)_0\right]$$

which, for the network shown in Figure 2 reduces to

$$\int_0^\infty i^2 e^{-pt}dt = A^{12}\left[\int_0^\infty E_1 e^{-pt}dt - \frac{S_{11}}{p}(q^1)_0\right] + A^{22}\left[\int_0^\infty E_2 e^{-pt}dt + L_{22}(i^2)_0\right]$$

Let $E_1 = E_2 = 1$ and let them both be applied at time $t = 0$, then

$$\int_0^\infty 1 e^{-pt}dt = 1/p$$

so

$$\int_0^\infty i^2 e^{-pt}dt = A^{12}\left[\frac{1}{p} - \frac{S_{11}}{p}(q^1)_0\right] + A^{22}\left[\frac{1}{p} + L_{22}(i^2)_0\right]$$

$$= \frac{A^{12}}{p}\left[1 - S_{11}(q^1)_0\right] + A^{22}\left[\frac{1}{p} + L_{22}(i^2)_0\right]$$

Also,

$$Z = \left(R_{11} + \frac{S_{11}}{p}\right)\left(R_{22} + pL_{22}\right) - R_{12}{}^2 = \frac{a}{p}\left[p^2 + bp + c\right]$$

where

$a = L_{22}R_{11}$; $b = (R_{11}R_{22} + S_{11}L_{22} - R_{12}{}^2)/L_{22}R_{11}$; $c = S_{11}R_{22}/L_{22}R_{11}$

Then $$A^{22} = (pR_{11} + S_{11})/a(p^2 + bp + c)$$

and $$A^{12} = (-pR_{12})/a(p^2 + bp + c)$$

Hence

$$\int_0^\infty i^2 e^{-pt}dt = -\frac{R_{12}[1 - S_{11}(q^1)_0]}{a(p^2+bp+c)} + \frac{(S_{11}+pR_{11})L_{22}(i^2)_0}{a(p^2+bp+c)} + \frac{(S_{11}+pR_1)}{ap(p^2+bp+c)}$$

$$= \frac{1}{a}\left[\frac{(S_{11}+pR_{11})}{p(p-p_1)(p-p_2)} + \frac{R_{12}S_{11}(q^1)_0 + (S_{11}+pR_{11})L_{22}(i^2)_0 - R_{12}}{(p-p_1)(p-p_2)}\right]$$

$$= A^2(p)$$

where $$p_1, p_2 = -b/2 \pm \tfrac{1}{2}(b^2-4c)^{1/2}$$

By the Mellin Inversion Theorem and Complex Integration

$$i^2 = \frac{1}{2\pi j}\sum \text{residues } A^2(p)e^{pt}$$

Since all the poles of $A^2(p)$ of this example are simple, the residue at a pole p_1 is simply $\lim_{p\to p_1}(p-p_1)A^2(p)e^{pt}$.

Residue at pole $p=0$ which yields the steady state is $S_{11}/ap_1p_2\cdot$ and the residues at p_1 are

$$\frac{1}{a}\left[\frac{(S_{11}+p_1R_{11})}{p_1(p_1-p_2)}e^{p_1t} + \frac{\{R_{12}S_{11}(q^1)_0 + (S_{11}+p_1R_{11})L_{22}(i^2)_0 - R_{12}\}}{p_1-p_2}e^{p_1t}\right]$$

and the residues at p_2 are

$$\frac{1}{a}\left[\frac{(S_{11}+p_2R_{11})}{p_2(p_2-p_1)}e^{p_2t} + \frac{\{R_{12}S_{11}(q^1)_0 + (S_{11}+p_2R_{11})L_{22}(i^2)_0 - R_{12}\}}{p_2-p_1}e^{p_2t}\right].$$

Hence,

$$i^2 = \frac{1}{a}\left[\frac{S_{11}}{p_1p_2} + \frac{(S_{11}+p_1R_{11})}{p_1(p_1-p_2)}e^{p_1t} + \frac{(S_{11}+p_2R_{11})}{p_2(p_2-p_1)}e^{p_2t}\right.$$

$$+ \frac{\{R_{12}S_{11}(q^1)_0 + (S_{11}+p_1R_{11})L_{22}(i^2)_0 - R_{12}\}}{p_1-p_2}e^{p_1t}$$

$$\left. + \frac{\{R_{12}S_{11}(q^1)_0 + (S_{11}+p_2R_{11})L_{22}(i^2)_0 - R_{12}\}}{p_2-p_1}e^{p_2t}\right]$$

As a check consider the limiting values of i^2 at $t = 0$ and $t = \infty$.

When $t = 0$

$$i^2 = \frac{1}{a}\left[\frac{S_{11}}{p_1p_2} + \frac{S_{11}}{p_1 - p_2}\left(\frac{1}{p_1} - \frac{1}{p_2}\right) + \frac{R_{11}(p_1 - p_2)L_{22}(i^2)_0}{p_1 - p_2}\right]$$

$$= \frac{1}{a} R_{11}L_{22}(i^2)_0 = (i^2)_0$$

When $t = \infty$ since $p_{1,2}$ have negative real parts $e^{p_{1,2}t} = 0$.

Then
$$i^2 = S_{11}/ap_1p_2$$

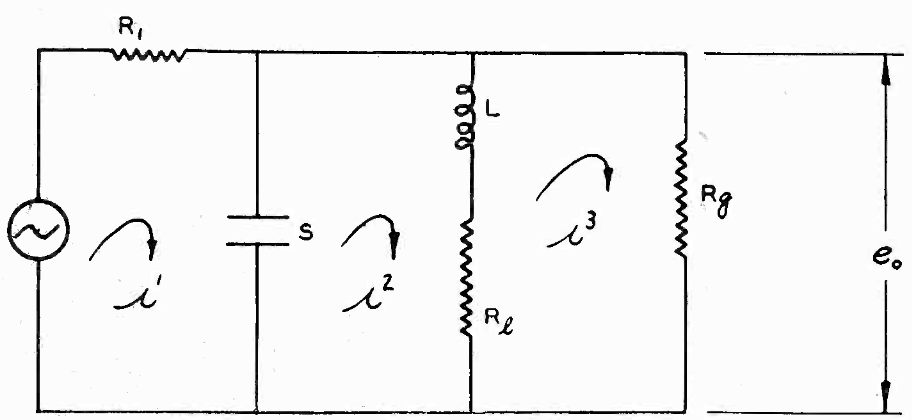

Fig. 3—High-frequency equivalent circuit of usual compensated video-amplifier stage.

Since $p_1, p_2 = -b/2 \pm (b^2/4 - c)^{1/2}$; $p_1p_2 = c = S_{11}R_{22}/L_{22}R_{11}$.

Thus,
$$i^2 = \frac{1}{a}\frac{S_{11}}{c} = \frac{1}{R_{22}}.$$

V. Application of Method to Usual Compensated Video Amplifier Stage

As an application of the general analysis given above, the high-frequency response of the ordinary compensated video-amplifier stage to a Heaviside unit e.m.f. has been calculated. The high-frequency equivalent circuit of the amplifier stage is shown in Figure 3 where R_1 = plate resistance, L = "peaking" coil, R_1 = load resistance, R_g = grid leak, and the tube and wiring capacitance are lumped together in the stiffness constant S.

Since the output voltage e_0 is desired it is necessary to solve for i^3 (see Figure 3), the third mesh current with voltage applied in the first mesh. Assume no initial charges and currents, and since there is only one e.m.f. the final current will be given by Equation (13) or

$$i^3(t) = \frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} \mu A^{13} \left[\int_0^\infty E_1 e^{-pt} dt \right] e^{pt} dp \tag{20}$$

For a constant e.m.f. applied at time $t = 0$, Equation (20) further reduces to

$$i^3(t) = \frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} \mu \frac{A^{13}}{p} e^{pt} dp \tag{21}$$

Upon evaluation of A^{13} for the circuit shown in Figure 3 for $E_1 = \text{constant} = 1$ the integrand of (21) becomes

$$\frac{A^{13}}{p} e^{pt} = \frac{(a_1 + a_2 p) e^{pt}}{p(a_3 + a_4 p + p^2)} = \frac{(a_1 + a_2 p) e^{pt}}{p(p - p_1)(p - p_2)} \tag{22}$$

where p_1 and p_2 are the roots of $a_3 + a_4 p + p^2$, i.e., p_1, p_2

$$= -\frac{a_4}{2} \pm \frac{1}{2} (a_4{}^2 - 4a_3)^{1/2}$$

where $$a_1 = R_l S / R_1 R_g L;\ a_2 = S / R_1 R_g$$

$$a_3 = \frac{R_l}{L} \left(\frac{1}{R_g} + \frac{1}{R_l} + \frac{1}{R_1} \right);\ a_4 = S \left(\frac{R_l}{L} + \frac{S}{R_g} + \frac{S}{R_1} \right)$$

By the general discussion given above $i^3(t)$ can then be obtained by finding the sum of the residues of (22).

Upon completing this process the final voltage e_0 is found to be

$$e_0 = R_g i^3 = \frac{\mu R_l R_g}{R} \left[1 - e^{-\alpha t/2} \left\{ \cosh \beta t - \frac{1}{\beta} \left(\frac{SR}{R_1 R_l R_g} - \frac{\alpha}{2} \right) \sinh \beta t \right\} \right] \tag{23}$$

where $$R = R_1 R_l + R_1 R_g + R_l R_g$$

$$\frac{\alpha}{2} = \frac{R_1 R_l R_g + SL(R_1 + R_g)}{2 R_1 R_g L}$$

$$\beta = \left\{ \left(\frac{\alpha}{2} \right)^2 - \frac{SR}{R_1 L R_g} \right\}^{\frac{1}{2}}$$

In the case of a pentode amplifier stage, R_1 is large and Equation (23) reduces to

$$\frac{\text{Gain}}{g_m} = \frac{e_0}{g_m e_g} = \frac{R_l R_g}{R_l + R_g} \left[1 - e^{-\alpha t/2} \left\{ \cosh \beta t - \frac{1}{\beta} \left[\frac{S(R_l + R_g)}{R_l R_g} - \frac{\alpha}{2} \right] \sinh \beta t \right\} \right] \qquad (24)$$

where

$$\frac{\alpha}{2} = \frac{R_l R_g + SL}{2LR_g} \; ; \; \beta = \left\{ \left(\frac{\alpha}{2} \right)^2 - \frac{S(R_l + R_g)}{LR_g} \right\}^{\frac{1}{2}}$$

For the case where $R_g >> R_l$ Equation (24) reduces to

$$\frac{\text{Gain}}{g_m} = \frac{e_0}{g_m e_g} = R_l \left[1 - e^{-\alpha t/2} \left\{ \cosh \beta t - \frac{1}{\beta} \left[\frac{S}{R_l} - \frac{\alpha}{2} \right] \sinh \beta t \right\} \right] \qquad (25)$$

where

$$\frac{\alpha}{2} = \frac{R_l}{2L} \; ; \; \beta = \left\{ \left(\frac{\alpha}{2} \right)^2 - \frac{S}{L} \right\}^{\frac{1}{2}}$$

Equation (25) may be transformed into a more useful form by the substitution $R_l = kL\omega_0 = S/\omega_0$. Equation (25) then becomes

$$\frac{\text{Gain}}{g_m R_l} = \left[1 - e^{-k\omega_0 t/2} \left\{ \cosh \left(\sqrt{\frac{k^2}{4} - k} \right) \omega_0 t - \frac{1 - k/2}{\sqrt{\frac{k^2}{4} - k}} \sinh \left(\sqrt{\frac{k^2}{4} - k} \right) \omega_0 t \right\} \right] \qquad (26)$$

Figure 4 shows the response of a single video-amplifier stage for various k values. The k value giving optimum response is seen to be approximately 2.

It may be of some interest merely to state here the expression for the gain of the customary video-amplifier stage for a suddenly impressed sinusoidal voltage. For a single pentode amplifier stage the gain is given by

$$\frac{\text{Gain}}{g_m} = \frac{e_0}{g_m e_g} = \frac{\omega S e^{-\frac{\alpha}{2}t}}{L\left\{\left[\beta^2 + \frac{\alpha^2}{4} + \omega^2\right]^2 - \beta^2\alpha^2\right\}}$$

$$\left[\left\{L\left(\beta^2 + \frac{\alpha^2}{4} + \omega^2\right) + \alpha\left(R_l - L\frac{\alpha}{2}\right)\right\}\cosh \beta t\right.$$

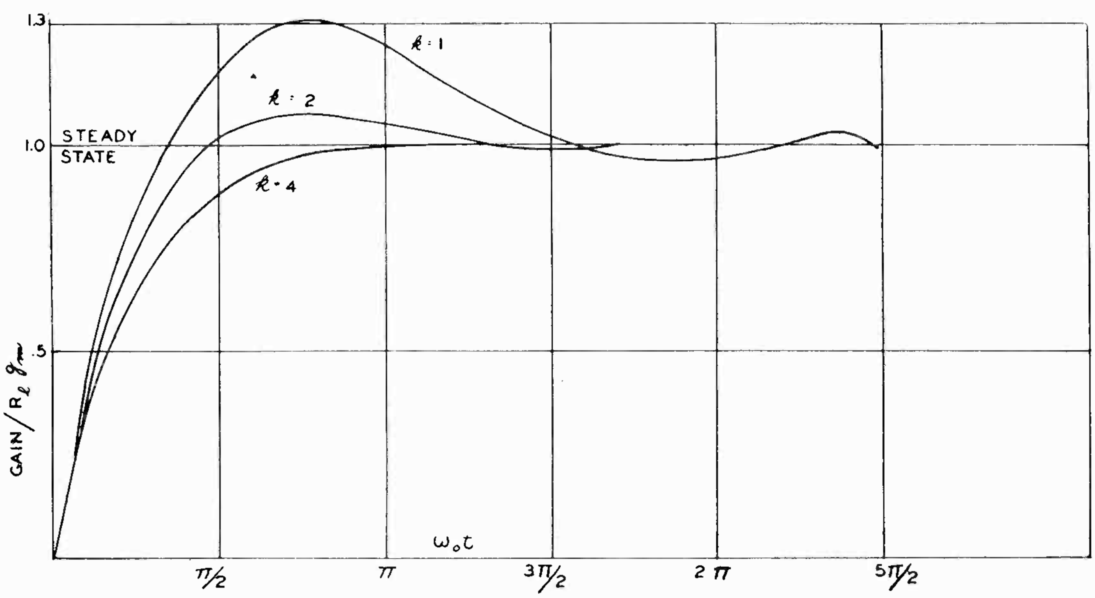

Fig. 4—Response of compensated video-amplifier stage to unit function.

$$\left. + \left\{\frac{1}{\beta}\left(R_l - L\frac{\alpha}{2}\right)\left(\beta^2 + \frac{\alpha^2}{4} + \omega^2\right) + L\alpha\beta\right\}\sinh \beta t\right]$$

$$+ \frac{S}{L\left\{\left[\frac{\alpha^2}{4} - \omega^2 - \beta^2\right]^2 + \alpha^2\omega^2\right\}}\left[\left\{L\omega\left(\frac{\alpha^2}{4} - \omega^2 - \beta^2\right) - R_l\alpha\omega\right\}\cos \omega t\right.$$

$$\left. + \left\{R_l\left(\frac{\alpha^2}{4} - \omega^2 - \beta^2\right) + L\alpha\omega^2\right\}\sin \omega t\right] \qquad (27)$$

where $\alpha/2 = R_l/2L$; $\beta = \{(\alpha/2)^2 - S/L\}^{1/2}$; $R_g >> R_l$ and ω equals 2π times the frequency of the impressed voltage.

VI. High-Frequency Response for n Identical Compensated Video-Amplifier Stages

Another case of practical interest is the complete solution for an n-stage amplifier where each stage can be represented by the equivalent circuit shown in Figure 3. The results obtained for a single compensated stage are immediately of use here since for n identical stages the final voltage can be obtained by summing the residues of the function $(a_1 + a_2 p)^n e^{pt}/p(p-p_1)^n(p-p_2)^n$. The validity of this procedure has already been established in the general theory. If the above function is resolved into partial fractions, after considerable

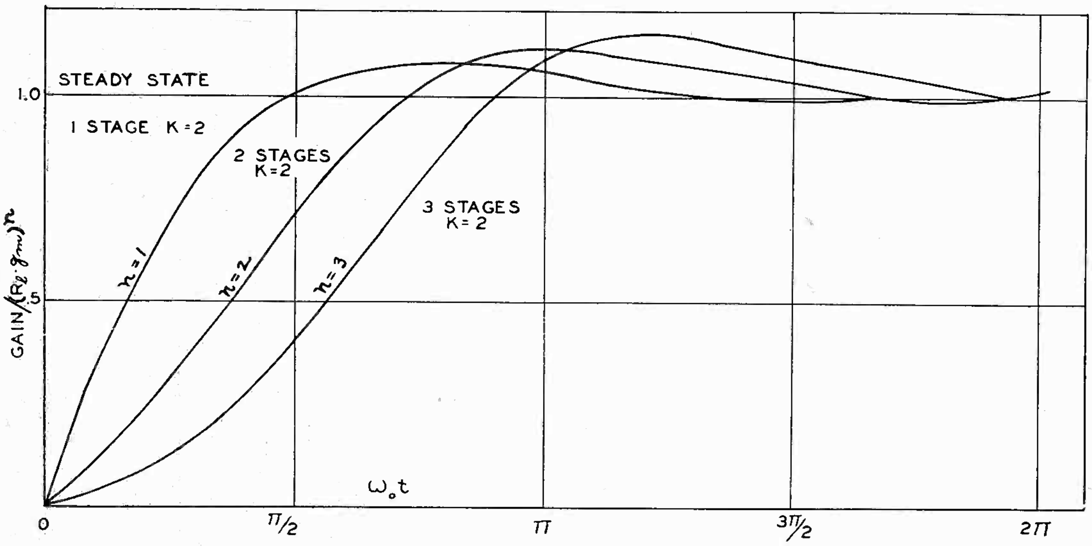

Fig. 5—Response of ordinary compensated video-amplifier stages.

reduction the following convenient expression for the output voltage for n pentode stages of amplification is obtained,

$$e_0 = (g_m e_g)^n R_l^n \left[1 + \frac{2}{(n-1)!} e^{-k\omega_0 t/2} \right.$$

$$\sum_{r=0,\ m=0}^{r=l,\ m=n-(l+1)} (-1)^{n-1+r-m} (\omega_0 t)^r nC_m \frac{[2n-m-(l+2)]!}{r![n-(l+1)-m]!} \cdot$$

$$\frac{(\sqrt{k})^{n-m+r-l-1}}{(2\gamma)^{2n-(l+1)-m}} \left[\sin\left[(m+l+1-2n)\ \pi/2 - (r-l-1)\theta\right] \right.$$

$$\cos(n-m)\theta \sin \gamma\omega_0 t$$

$$\left. - j \cos\left[(m+l+1-2n)\ \pi/2 + (n-m-r+l+1)\theta\right] \sin \gamma\omega_0 t \right] \quad (28)$$

where $\qquad l = 0, 1, 2, \cdots\cdots n-1;\ \theta = \tan^{-1} 2\gamma/k$

$$\gamma = (k - k^2/4)^{1/2};\ R_l = kL\omega_0 = S/\omega_0$$

Figure 5 gives the response as calculated from (28) for three video-amplifier stages where the approximate optimum value $k = 2$ for a single stage was chosen.

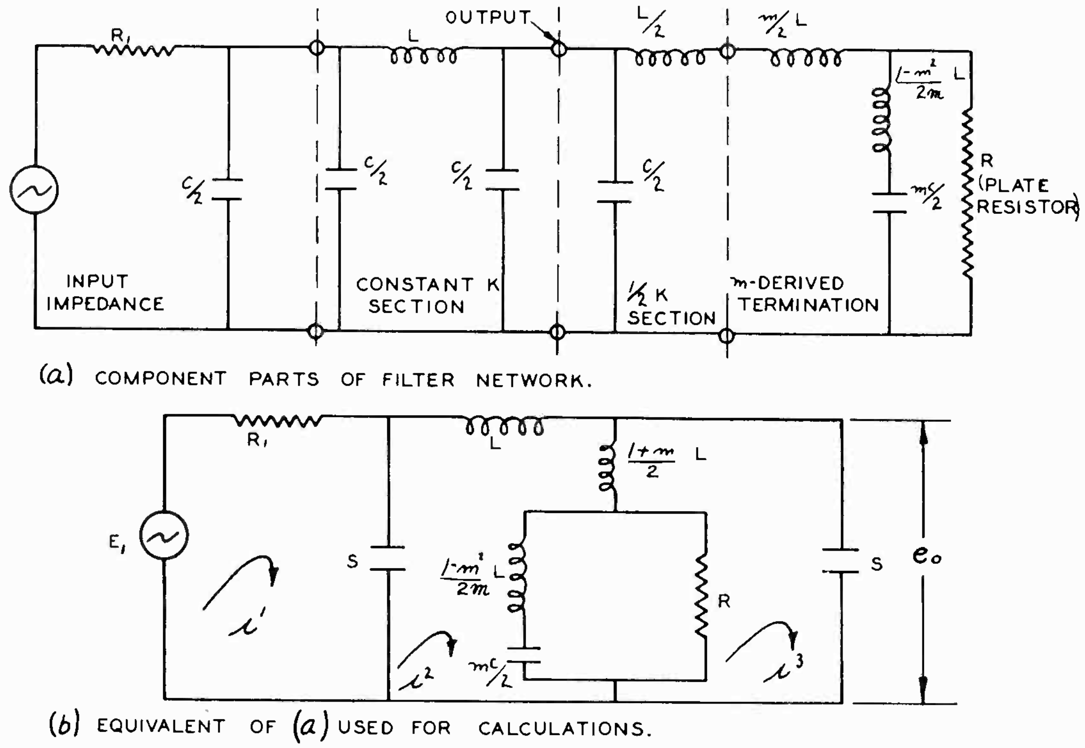

Fig. 6—High-frequency equivalent circuit for video-amplifier stage employing low-pass filter network.

VII. High-Frequency Response of Several Video Amplifiers Using Low-Pass Filter Networks

Recently, considerable attention has been given to video-amplifier circuits which are low-pass filter networks in which additional capacitance is shunted across the filter input by the output tube capacitance.[7] The voltage which is applied to the grid of the succeeding amplifier stage is taken from an intermediate point in the filter in order that the input capacitance of this stage is included as part of the total shunt capacitance of the filter section. One of the two arrangements which are considered in this paper is shown in Figure 6 which is a conventional low-pass filter with additional shunt capacitance across the input

[7] See, for example, W. S. Percival, British Patent No. 475490 and H. A. Wheeler, "Wide Band Amplifiers for Television," paper given at I. R. E. Convention, June 18, 1938, New York City.

and with m-derived termination to provide a uniform characteristic impedance.

From the last form of the equivalent circuit shown in Figure 6, the desired current i^3 for a Heaviside unit voltage in the first mesh will be given by (see Equation 13)

$$i^3(t) = \frac{1}{2\pi j}\int_{c-j\infty}^{c+j\infty} \frac{A^{13}e^{pt}}{p}\,dp \tag{29}$$

where no initial currents and charges have been considered. Therefore $i^3(t)$ will be given by the sum of the residues of $A^{13}e^{pt}/p$. It is more convenient to express results in terms of gain e_0/e_g where e_0, the output voltage, is $Si^3(t)/p$ (see Figure 6). The equation for the output voltage becomes $e_0 = Si^3(t)/p = \dfrac{1}{2\pi j}\displaystyle\int_{c-j\infty}^{c+j\infty} \frac{SA^{13}e^{pt}}{p}\,dp.$

After evaluation of A^{13} the problem reduces further to the following

$$\frac{\text{Gain}}{g_m R} = \frac{e_0}{g_m e_g}\,\frac{1}{R} =$$

$$\sum \text{residues} \frac{(a_3p^3 + a_2p^2 + a_1p + 1)e^{pt}}{p(b_6p^6 + b_5p^5 + b_4p^4 + b_3p^3 + b_2p^2 + b_1p + 1)} \tag{30}$$

where

$a_3 = (1+m)(1-m^2)L^2/8RS;\; a_2 = (1+m)L/4S;$

$a_1 = (1+m)L/2R;\; b_6 = (1+m)(1-\mathrm{m}^2)L^3/8S^3;$

$b_5 = (1+m)L^2R/4S^3;\; b_4 = (4+3m-2m^2-m^3)L^2/4S^2;$

$b_3 = (3+2m)LR/2S^2;\; b_2 = (9+4m-m^2)L/4S;\; b_1 = (4+m)R/2S$

From low-pass filter requirements

$$R = \sqrt{\frac{LS}{2}}\;;\quad \sqrt{\frac{L}{S}} = \frac{\sqrt{2}}{\omega_c} \tag{31}$$

where $\omega_c = 2\pi$ times the cut-off frequency of the filter. Substituting the relations (31) into (30) and letting $P = p/\omega_c$, it is found that Equation (30) reduces to

$$\frac{\text{Gain}}{g_m}\cdot\frac{\omega_c}{R} = \sum \text{residues} \frac{(1 + a_1P + a_2P^2 + a_3P^3)\,e^{\omega_c Pt}}{P(1 + b_1P + b_2P^2 + b_3P^3 + b_4P^4 + b_5P^5 + b_6P^6)} \tag{32}$$

where

$$a_1 = (1+m);\ a_2 = a_1/2;\ a_3 = a_2(1-m^2);$$
$$b_1 = (4+m)/2;\ b_2 = (9+4m-m^2)/2;\ b_3 = 3+2m;$$
$$b_4 = (4+3m-2m^2-m^3);\ b_5 = a_1;\ b_6 = (1+m)(1-m^2)$$

For a uniform impedance characteristic $m = 0.6$, substituting this value in (32) and solving numerically for the roots of the denominator of the above expression, summing the residues, one obtains

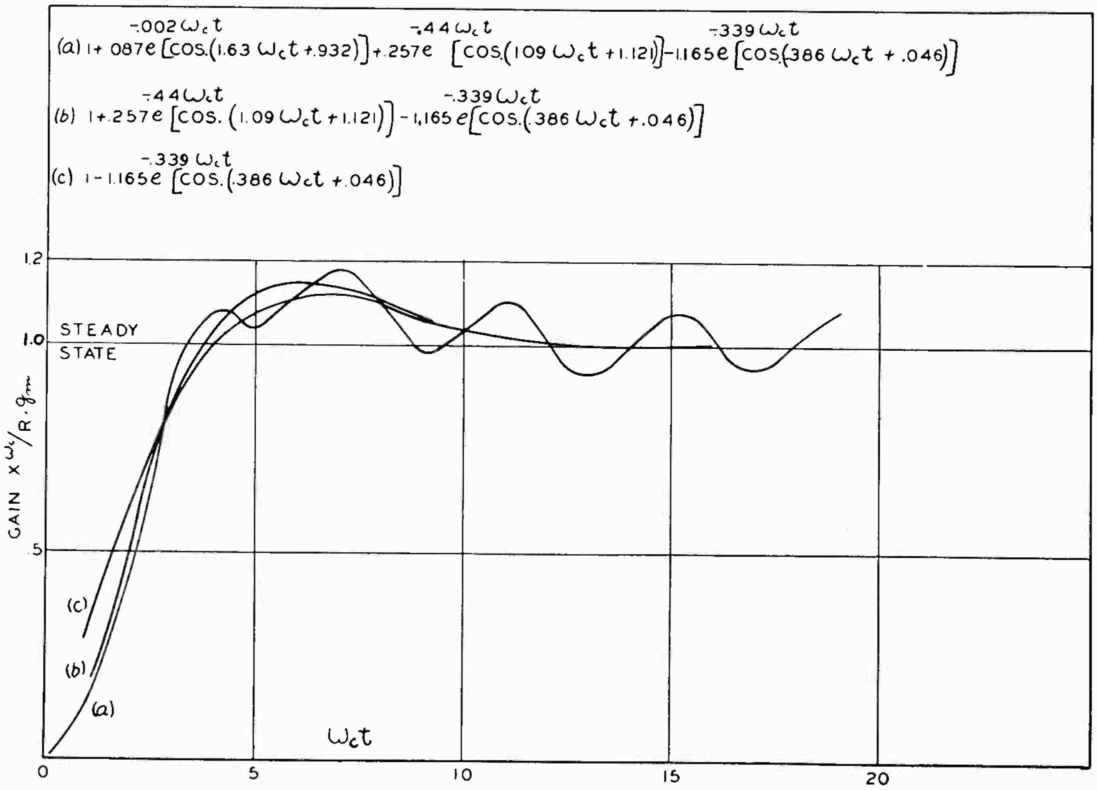

Fig. 7—Response of video-amplifier stage with low-pass filter as coupling network.

$$\frac{\text{Gain}}{g_m} \times \frac{\omega_c}{R} = 1 + .087e^{-.002\omega_c t}\{\cos(1.63\,\omega_c t + .932)\}$$
$$+ .257e^{-.44\omega_c t}\{\cos(1.09\,\omega_c t + 1.121)\} \tag{33}$$
$$- 1.165e^{-.339\omega_c t}\{\cos(.386\,\omega_c t + .046)\}$$

Figure 7 represents the conclusions given by Equation (33). Curve (a) shows the response of such an amplifier to a unit function. However, in a television system, an ideal unit function is never applied. The approximation to the unit function which is applied does not contain frequencies above ω_c, the cut-off frequency. Therefore, roughly the response to a sudden change of brightness will correspond to a response in the output of the amplifier given by curve (c) where the terms containing frequencies above ω_c have been omitted. This curve gives approximately an 11 per cent "overswing." Curve (b) is the case where the term involving the frequency 1.63 ω_c has been omitted. Note that curves (b) and (c) practically coincide. From these curves it is evident that noise signal components 1.63 the cut-off frequency of the amplifier will cause a transient of long duration unless limited by some other circuit in the channel.

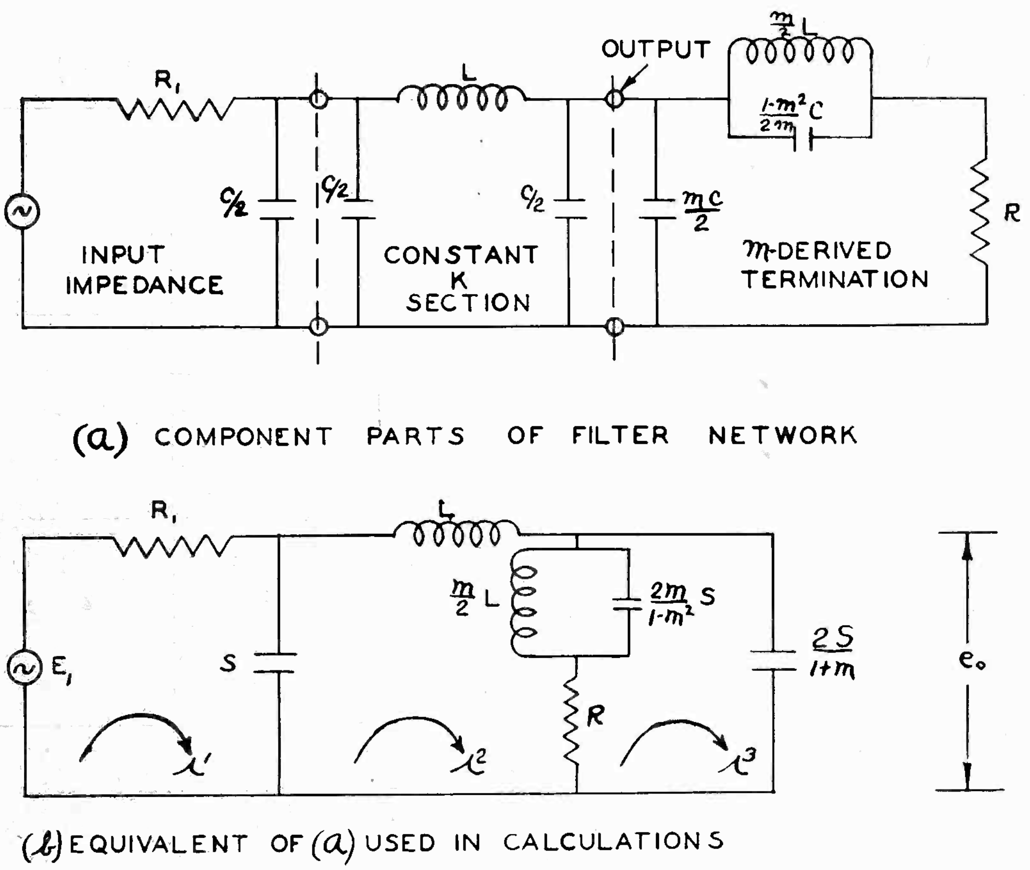

Fig. 8—Low-pass filter network for video-amplifier stage.

The other low-pass filter network considered here is shown in Figure 8. If the steps outlined in the discussion immediately above are carried through there finally results

$$\frac{\text{Gain}}{g_m}\,\frac{\omega_c}{R} = \sum \text{residues} \frac{(a_2P^2 + a_1P + 1)\,e^{\omega_c Pt}}{P\,(b_5P^5 + b_4P^4 + b_3P^3 + b_2P^2 + b_1P^1 + 1)} \tag{34}$$

where

$$a_2 = (1 - m^2)/2;\ a_1 = m;\ b_5 = (1 + m - m^2 - m^3)/2;$$

$$b_4 = 1 + m;\ b_3 = (7 + 5m - 3m^2 - m^3)/4;$$

$$b_2 = (5 + 3m)/2;\ b_1 = (3 + m)/2$$

For $m = 0.6$, the final solution of (34) can be written as

$$\frac{\text{Gain}}{g_m} \cdot \frac{\omega_c}{R} = 1 - .038e^{-2.313\omega_c t} + .266e^{-.093\omega_c t} \cos(1.3\,\omega_c t + .876)$$
$$- 1.16e^{-.313\omega_c t} \sin(.632\,\omega_c t + 1.349) \qquad (35)$$

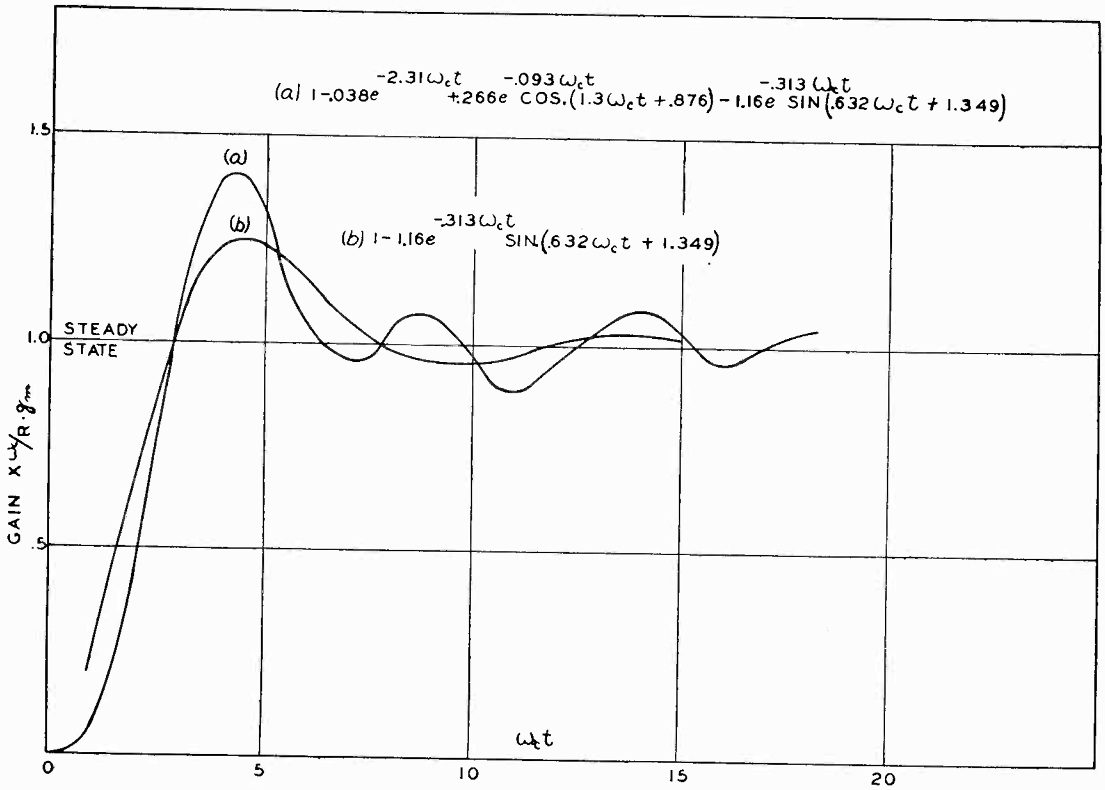

Fig. 9—Response of video-amplifier stage with low-pass filter as coupling network.

Curve (a) of Figure 9 shows the complete Equation (35) which contains only two natural frequencies whereas the previous filter circuit contained three natural frequencies. Curve (b) Figure 9 shows Equation (35) with third term omitted. Upon comparing curves 9 (a) and (b) it is apparent that the third term of Equation (35) aids greatly in producing 40 per cent "overswing"; further, the transient due to this term persists in considerable amplitude for an appreciable time. It is interesting to note that even with the omission of the third term there still remains a 25 per cent "overswing."

From this analysis it would thus seem that the low-pass filter of Figure 6 is superior to that of Figure 8.

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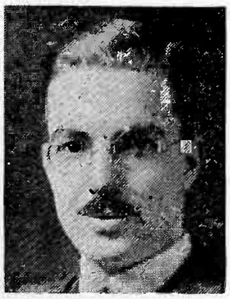

JAMES N. WHITAKER is a native of Santa Rosa, California, and was graduated from the High School of that city. He served in the United States Navy from 1923 to 1927. Upon his discharge he joined the engineering department of the Radio Corporation of America, changing in 1928 to the engineering department of R.C.A. Communications, Inc., where he since has been engaged.

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